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CCIR

XVIIth PLENARY ASSEMBLY
DÜSSELDORF, 1990

INTERNATIONAL TELECOMMUNICATION UNION

REPORTS OF THE CCIR, 1990

(ALSO DECISIONS)

ANNEX TO VOLUME I

SPECTRUM UTILIZATION AND MONITORING

CCIR INTERNATIONAL RADIO CONSULTATIVE COMMITTEE

ARCHIVES U.I.T. GENÈVE

Geneva, 1990

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(Study Group 1)

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SECTION 1A: SPECTRUM ENGINEERING AND COMPUTER-AIDED PRINCIPLES AND TECHNIQUES

REPORT 1108

SPECTRUM MANAGEMENT

(Question 68/1)

(1990)

1. General

Spectrum management is the combination of administrative (legal, economic), scientific and technical procedures necessary to ensure the efficient operation of radiocommunication equipment and services without causing harmful interference. There are national and international aspects to spectrum management and this Report emphasizes the national aspects. This Report and its references provide a guide to more detailed information on this subject contained in CCIR texts. Two documents, an IFRB/CCIR Booklet on National Spectrum Management and a CCIR Handbook on Spectrum Management and Computer-Aided Techniques, have been prepared as a result of WARC-79, Resolution No. 7 and Recommendation 31, respectively. This Report summarizes some of the findings in these documents dealing with the functions, organizational structure and techniques of national spectrum management.

In order to implement the provisions of the Constitution, the Radio Regulations, and bilateral or multilateral treaties as appropriate with neighboring countries, there is a need for an identifiable spectrum management authority in each country having the necessary powers and appropriate supporting units. Each country should establish a spectrum management authority according to its particular requirements and available resources and adopt for it the organizational structure best suited for carrying out its tasks.

Recognizing that the international and national functions of spectrum management are very closely interlinked, most countries have found it appropriate to set up a single authority whose responsibilities cover both aspects. This authority is the "administration" (defined in RR3) for the country concerned.

2. The Basic Functions to be performed by a National Spectrum Management Unit. (Report 841, Section 2.2)

The basic functions of a spectrum management unit in support of the national authority can be identified as:

- long-term spectrum management policy and planning;
- establishment of a table of national frequency allocations and national regulations;
- assignment of frequencies;
- licensing and authorization of the use of radiocommunication equipment;

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- collecting fees where appropriate;
- information exchange and liaison with radio users;
- national and international coordination;
- establishment of technical standards and engineering support facilities;
- record keeping;
- inspection of installations;
- monitoring the use of the spectrum;

The specific tasks arising from carrying out these basic functions are:

a) planning, coordinating, regulating and administering the use of radio frequencies within the country;

b) establishing regulations, technical parameters and standards governing the use of each frequency band or specific frequency by stations of different services, having regard to current international regulations and agreements;

c) optimizing the use of the radio spectrum and the geostationary-satellite orbit, ensuring the harmonious operation of the different services which use them;

d) allocating frequency bands in accordance with international regulations and the national priorities and assigning specific frequencies as appropriate;

e) authorizing the installation and operation of radio stations and radio operators, assigning call signs, granting appropriate licences and collecting fees if appropriate;

f) updating all information on authorized radiocommunication systems such as frequencies, the location of stations, powers, call signs, etc., and their notification to the IFRB if necessary;

g) appropriate notification of information required for inclusion in the publications and service documents of the ITU;

h) representing the administration as appropriate, establishing relations with communications industries and the public, coordinating and issuing technical opinions concerning the use of frequencies in national and international forums;

i) measuring the technical parameters of the emissions of radiocommunication stations as appropriate;

j) conducting inspections of radiocommunication stations to check that they meet the technical standards and parameters for which their equipment and operation were authorized;

k) participating, so far as the use of frequencies is concerned, in the development plans and projects of all radiocommunication services, ensuring that those plans are in accordance with current international and national regulations;

l) resolving radio frequency interference problems;

m) preparing for participation in international conferences convened by the ITU, participating in such conferences convened by the ITU, _____ _____ and implementing any decisions adopted;

n) conducting negotiations in connection with spectrum management and related problems with other countries and international organizations;

o) providing a national technical forum for work relating to the Study Groups and Working Groups of the CCIR, preparing for the participation of specialists in meetings of the CCIR and participating therein;

p) constituting the national body for relations with international organizations other than the ITU on technical, regulatory and administrative matters, technical cooperation and other subjects related to utilization of the radio frequency spectrum and the geostationary-satellite orbit.

3. Organizational Structure (Report 841, § 2.3)

A national spectrum management unit capable of carrying out the tasks listed in section 2 may be organized in many different ways according to the requirements and resources of the country concerned: there is therefore no single ideal arrangement. However, Figure 1 is an example of the sub-units in a fully developed national frequency management unit. The role of each sub-unit is discussed in sections 3.1-3.6 below.

3.1 Planning and Policy

The Planning and Policy sub-unit is typically responsible for the ongoing development of national plans and policies relating to the use of the radio frequency spectrum, taking into account advances in technology as well as social, economic and political realities.

National telecommunication policy is commonly associated with regulation development because of their close relationship. Accordingly it might also be a primary function of this sub-unit to conduct studies to determine existing and future telecommunication needs of the country and to develop policies to ensure that the best combination of radio, **fixed network and cable systems is employed in meeting the identified needs.**

The resulting national radio regulatory policies and regulations must be fully documented and must be in line with the Radio Regulations of the ITU and particularly the Table of Frequency Allocations.

3.2 International Coordination

World and regional administrative radio conferences of the ITU and other types of conferences and meetings can require a great deal of advance preparation depending on the level of participation desired by the administration and the issues under consideration. In order to participate effectively at such gatherings an administration may set up working groups to develop national positions on the key issues of interest, and this may be done under a coordination sub-unit, which would work closely with the Planning and Policy sub-unit, the Engineering Support sub-unit and users as necessary.

This sub-unit could also be responsible for the development of bilateral agreements with neighboring countries to settle policy or operational issues for the purpose of coordinating the establishment of communication systems and other items of mutual interest in the radio field.

3.3 Assignment, licensing and liaison functions

The assignment, licensing and liaison functions could be organized into three sections.

3.3.1 Frequency Assignment Section

The frequency assignment section is responsible for conducting electromagnetic compatibility analyses, and for assigning appropriate frequencies to radiocommunication systems. It may also initiate requests to the Inspection and Monitoring sub-unit for the monitoring of specific frequencies or frequency bands to facilitate the engineering analysis of required assignments.

This section is also responsible for coordinating all proposed assignments with regard to existing assignments on a national, regional or international basis. It notifies frequency assignments to the IFRB as required under the Radio Regulations. It also carries out coordination and related actions on request to protect the country's radiocommunication systems from interference or when information on assignments notified by other administrations appears in the Weekly Circular published by the IFRB.

3.3.2 Licensing Section

The licensing section plays a major role in any well-structured spectrum management unit. This section applies the national legislation, regulations, policies and procedures governing radiocommunications. It also exercises control over the operation of stations and the use of frequencies by:

- examination of license applications and related documents to determine the licensing eligibility of the applicant from a legal and regulatory point of view and the technical acceptability of the radio equipment proposed;
- granting authorization to entities which may not require a license, such as government agencies;
- assigning call signs to individual stations;
- issuing licenses and collecting fees, if appropriate;
- renewing and cancelling licenses as appropriate;
- conducting examinations of operator competence and issuing operator certificates;

3.3.3 Information and liaison section

The information and liaison section is responsible for providing the interface between the spectrum management unit and radio users composed of the communications industry, government users where appropriate and general public. It disseminates information on the policies, rules and practices of the administration and provides mechanisms for feed back to evaluate the results of these policies, rules and practices. It maintains media relations, presents public notices, conducts meetings and acts as an intermediary in resolving interference problems between users, drawing on the support of inspection, monitoring and investigation units.

3.4 Inspection and Monitoring

The Inspection and Monitoring sub-unit may be organized in two sections, which according to the size and requirements of the country concerned may be centralized or dispersed.

3.4.1 Inspection Section

The inspection section is provided with the inspection equipment and mobility needed to enforce the national and international statutory and regulatory requirements of frequency management. It works closely with the monitoring section and the Assignment and Licensing sub-unit in collecting information.

The functions of this section may include the following:

- to investigate interference complaints;
- to investigate radiation from industrial, scientific and medical appliances;
- to investigate illegal operations, and operations not in keeping with the terms of radio station licenses;
- to collect information for prosecution cases and to assist law enforcement agencies in carrying out the process of the law; and
- to ensure that radio station operators comply with national and international statutory and regulatory requirements.

3.4.2 Monitoring Section (Reports 371, 1103, 1115 and 1116)

In the management of the frequency spectrum, the monitoring section has two basic responsibilities. One is to provide statistical information of a technical and operational nature on spectrum occupancy. The second is to obtain information for regulatory and enforcement purposes on the operation of individual radio stations. The monitoring section also performs an important function in establishing the locations and identities of stations causing interference.

According to its resources a country may wish to participate in the international monitoring system which exists to assist, to the extent practicable, in the implementation of the Radio Regulations, in particular to ensure efficient and economical use of the spectrum, and to help in the prompt elimination of harmful interference (see Article 20 of the Radio Regulations, and the CCIR Handbook for Monitoring Stations).

3.5 Engineering Support

The role of the Engineering Support sub-unit is vital and central to the development of the full national frequency management unit because the management of the radio frequency spectrum is becoming increasingly dependent on technical support. The functions of the Engineering Support sub-unit can be carried out by two sections, as follows.

3.5.1 Standards and Specifications Section

The standards and specifications section is responsible for the development of procedures for the approval of radio equipment, and radio standards specifications (documents which set forth

the minimum performance standards required for the type approval or radio transmitters and receivers and other equipment).

3.5.2 Spectrum Engineering Section (Report 841, § 3)

The spectrum engineering section is an engineering support facility which typically provides the following services to frequency management operations as a whole:

- laboratory testing of transmitting and receiving equipment in keeping with prescribed type-approval procedures;
- maintenance and calibration of laboratory test equipment and other equipment used by the Inspection and Monitoring sub-unit;
- acceptance evaluation of equipment being purchased for inspection and monitoring purposes; and
- outfitting of special-purpose vehicles and calibration of equipment to be fitted in such vehicles.

3.6 Computer Support (Recommendations 667 and 668; Reports 841, § 5 and 1110)

The extent to which computer support facilities are available to and used by the frequency management authority depends on the resources, priorities, and particular requirements of the country concerned. In the early stages of its introduction, computer support may be limited to, say, licensing records or the more complex engineering calculations; ultimately the Computer Support sub-unit may assume responsibility for the development, provision and maintenance of support facilities for nearly all spectrum management activities, including record keeping, forecasting, and financial management related to licensing.

The CCIR Handbook on Spectrum Management and Computer-Aided Techniques addresses the use of computers in this context in considerable depth.

4. Spectrum Management Techniques (Report 841)

Sections 2 and 3 have identified the basic functions and tasks resulting from these functions, and a possible organizational structure for a national spectrum management unit. Several spectrum management techniques are important to the success of interference free national operations. These techniques include standards, planning, frequency assignments and international registration.

4.1 Standards (Recommendations 239, 326, 328, 329, 331 and 332)

The main purpose of developing standards is to specify in legal or administrative texts the conditions to be met by transmitting and receiving equipment and any other apparatus which may produce interference, with a view of achieving better use of the spectrum. Some of these conditions may take the form of restrictions on the technical characteristics covered by provisions in the Radio Regulations, e.g. frequency tolerances, spurious emission tolerance levels, etc. Each administration preserves the freedom to adopt values for these restrictions different from those laid down in the Regulations, in light of its needs and specific possibilities, provided that those values are in no way less restrictive than the ones laid down in the Regulations.

Standardization is essential for proper use of the spectrum. For example, if man-made noise is not regulated, it may become extremely strong in large metropolitan centers where radio stations tend to be set up, and operations will be compelled to use a fairly high transmitter power to ensure satisfactory reception in competition with this noise. In so doing, they will increase the probability of interference to other users.

4.2 Planning (Reports 976 and 1111)

The primary goals of this technique are to determine and periodically update existing and future national requirements for various radiocommunication services. The Table of Frequency Allocations in the Radio Regulations should be the starting document for national spectrum planning. ITU administrative radio conferences allocate frequency bands to services. An administration may deem it advisable to adopt for its administrative purposes more detailed provisions than those in the Table of Frequency Allocations, but they must, however, remain within the framework of the Table's provisions. Thus, for a band allocated in the table with equality of rights to services A and B, an administration may decide to divide the band into two parts, one part allocated exclusively to service A and the other to service B. In this field, i.e., the use of frequency bands, the spectrum management authority must coordinate the needs of its various users.

After understanding users requirements it is essential for an administration to develop a national frequency plan for its radio services which could include sub categories. Such a case is with the land mobile service, where it is necessary to establish radio user categories. User categories are normally business or activity dependent. One category might be for public safety which encompasses police, fire fighting and ambulance services. Other categories may be for taxis and public mobile telephone systems. The categories may have different priorities for access to the spectrum. Thus certain frequencies will be reserved for certain categories. The spectrum management authority should be responsible for such planning.

4.3 Assignment of Frequencies (Reports 842 and 1113)

The technique of assigning frequencies is to ensure that no harmful interference is created from proposed or requested assignments to existing assignments on a national and sometimes international basis. It will include the analysis of requirements for proposed radio services together with any relevant studies, and the assignment of the frequencies in accordance with the national plan. It may include related actions necessary to protect the country's radiocommunication systems from potential interference from another country's assignments published in the latest edition of the International Frequency List (IFL), List I, and in the Weekly Circulars published thereafter by the IFRB.

Once a frequency has been assigned to a transmitting or a receiving station, all the technical and operating data indicating the spectrum, and geographical area occupied by this assignment should be entered in the national frequency register. This register not only serves as a reference when subsequently selecting other usable frequencies but also provides the basic material for taking effective measures required to adapt national planning to the real requirements of the various users. The greatest care should be taken in compiling the national register and keeping it up to date; it must have room for recording a sufficient number of assignments and or all the information needed for the clear and complete description of each. If the size of the register is greater than approximately 5000 records, it may be very useful to employ modern computer processing and recording techniques.

4.4 International Registration

International registration of an administration's frequency assignments provides international recognition and protection for the stations' operations. It is in the best interest of an administration to register all its frequency assignments which it feels need protection from other international users. There are certain circumstances when an administration is obligated under the Radio Regulations to notify its use of the spectrum.

In accordance with the provisions of the Radio Regulations, any frequency assignment, with certain specified exceptions, shall be notified to the IFRB if:

a) the use of the frequency is capable of causing harmful interference to any service of another administration;

b) the frequency is to be used for international radiocommunication; or

c) it is desired to obtain international recognition of the use of the frequency.

The procedures governing notification and registration of frequency assignments in the *Master International Frequency Register* (MIFR) may be broadly subdivided into the acts of coordination, notification, examination and registration. Chapter IV of the Radio Regulations provides the specific procedures to be followed.

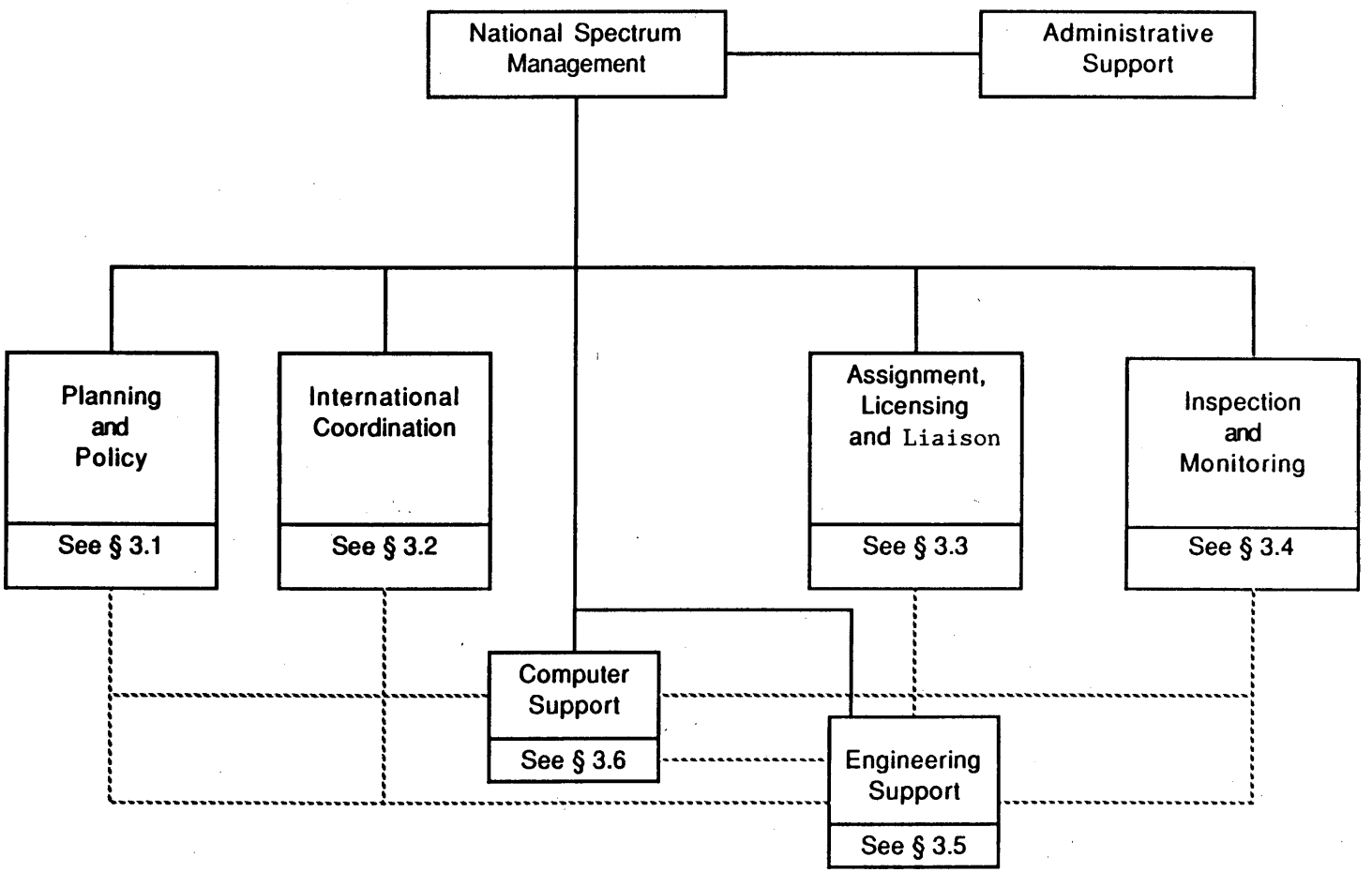


FIGURE 1
NATIONAL FREQUENCY MANAGEMENT FUNCTIONS

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REPORT 1103 *

SETTING LIMITS ON EMC PARAMETERS FOR MOBILE SERVICE COMMUNICATION EQUIPMENT BASED ON AN ANALYSIS OF THE INTERFERENCE ENVIRONMENT

(Question 18/1 and 44/1)

(1990)

1. Introduction

A large number of individual EMC parameters have been developed in the past and are now used to characterize the receptivity of receivers to interfering signals present in the non-fundamental reception channels and which also characterize the non-fundamental radiation of transmitters. These parameters include receiver susceptibility to spurious, blocking, intermodulation and reciprocal mixing channels [Winn, R., 1974, Rohde, U., 1975] and also transmitter out-of-band and spurious radiation.

The present system for limitating EMC parameters is directed towards minimizing the EMC parameters of receivers and transmitters as far as possible. At the same time there is no clear link between the requirements for the EMC parameters and the operational conditions of equipment, i.e. the interference environment (IE) at the receiving point. It is obvious that the EMC parameters of receivers and transmitters should not be higher or lower than a level which will guarantee a given reception quality of the wanted signal within a typical IE. Moreover, the present system makes it difficult to compare different types of receivers and transmitters, and it does not allow any quantitative evaluation to be made of the improvement in the EMC of radio installations when grouped together if one of the EMC parameters of the transmitters or receivers is modified.

The present report proposes one of the various possible approaches to the problem of achieving optimum limitation of the EMC parameters of receivers and transmitters which will guarantee a given quality of reception in conditions of a typical IE. The proposed method makes it possible to make a qualitative evaluation of improvements in the quality of reception (total probability of reception being suppressed) if the EMC parameters of receivers and transmitters are improved, and also to optimize the EMC parameters in relation to the criterion of the cost of radio equipment.

2. EMC parameters of a group of radio installations

In attempting to solve the problems of the limitation of EMC parameters of receivers and transmitters, it is necessary to define a typical IE in which radio installations operate and to take into account all the EMC parameters of the receivers. The following parameters are proposed to characterize the IE, or the EMC of a group of radio installations: the probability of the reception of a

* This Report should be brought to the attention of Study Group 8.

signal being suppressed by the ith interference mode q_i; and the cumulative distribution function F_i (P_{ui}), where P_{ui} is the amount by which the power of the interfering signal at the receiver input caused by the ith mode interference (P_{Ri}) exceeding an admissible value: $P_{ui} = P_{Ri} - P_{RiN}$.

The probability of a reception channel being suppressed q_i is equal to the ratio of the number of channels suppressed by the ith mode of interference for which $P_{ui} > 0$, to the total number of channels used by the radio installations in the group. By statistically processing the values of P_{ui} for all radio installations in the group, it is possible to obtain the cumulative distribution function F_i (P_{ui}) of the P_{ui} values.

The proposed parameters for the EMC of a group q_i and F_i (P_{ui}) may be obtained by measuring the levels of the interfering signals at the input of the receivers P_{Ri} and comparing them with the admissible levels of the interfering signals P_{RiN}. Measurements of this sort were carried out over a long period for broadcasting systems in the LF, MF and HF bands [Bank, M.U., 1973] and also for professional communications in the HF band [Sosin, B., 1971]. For mobile communication systems operating with considerable intervals, it is better, in defining the EMC parameters for a group, to use a computerized EMC mathematical modelling method. The interference modes (i) to be taken into account may include: 1) the effect of the out-of-band radiation of the transmitter on the fundamental receiving channel of the radio receiving installation; 2) the effect of the harmonic radiations of the transmitter on the fundamental receiving channel; 3) the effect of the fundamental radiation of the transmitter on the spurious receiving channel; 4) the effect of harmonic radiation on a spurious receiving channel; 5) the effect of reciprocal mixing or blocking; 6) third order intermodulation interference with two interfering signals, etc.

The total probability for reception to be suppressed through at least one of the fundamental or non-fundamental receiving channels in accordance with the law governing the addition of probabilities of elementary random independent consistent events is:

$$q_\Sigma = 1 - \prod_{i=1}^{n_i} (1 - q_i) \qquad (1)$$

where n_i - number of interference modes taken into account.

Thus, real-life measurements or mathematical modelling can be used to obtain the initial data q_i, q_Σ, F_i required to limit the EMC parameters of receivers and transmitters which will guarantee a given quality of reception $q_{\Sigma N}$ having regard to the IE.

3. Limitation and effective improvement of the EMC parameters of receivers and transmitters

A decrease in the susceptibility of a receiver for a given ith mode of interference, and also a reduction in the level of a transmitter's out-of-band or harmonic radiation by K dB will give rise to a decrease in the level of excess interference at the receiver input P_{ui} by a value of ΔP_{ui} and, consequently, a reduction in the probability of reception being suppressed q_i to a value of q_{iK}. The value of q_{iK} can be determined using a distribution function for the selected value of K:

$$q_{iK} = q_i \, [1 - F_i(K)]. \qquad (2)$$

Thus, by successively changing each of the EMC parameters of the receivers and transmitters by one with the same value for K, it is possible to determine a new value for the probability of reception being suppressed q_{iK} and, on the basis of formula (1), the value of q_Σ. By comparing the values of q_i and q_{iK}, we can then determine the parameters in which an improvement will produce a substantial modification in q_{iK} in comparison with q_i and also those parameters which have little influence on q_{iK}. It is obviously advisable to perfect those parameters which substantially reduce q_{iK} and thus improve the EMC of the group q_Σ.

For the purpose of limiting those EMC parameters of radio receiving and radio transmitting installations that will guarantee a given quality of reception $q_{\Sigma N}$ (i.e. a given total probability of reception being suppressed) in a typical IE, the following condition has to be satisfied:

$$q_\Sigma \leq q_{\Sigma N}.$$

This problem can be solved by different requirements for q_i. The choice of a specific requirement is defined on the basis of technical feasibility and economic considerations. In order to optimize the cost of receivers and transmitters, at the same time guaranteeing a given quality of reception, the condition has to be satisfied:

$$C_\Sigma = \sum_{i=1}^{m} C_i(K_i) = \min \quad \text{where } q_\Sigma \leq q_{\Sigma N},$$

where $C_i(K_i)$ is the cost of improving the ith parameter by K_i;

and m is the number of EMC parameters which are improved.

4. Example of the use of the method

In order to apply the proposed limitation method to radio installations in the land mobile service, a computerized mathematical model was constructed for a group of base stations in the mobile services operating in the bands 40, 160 and 330 MHz (j = 1,3). The automated modelling system is described in [Vasilev, A.V., Kaganer, M.B., Rubinshtein, G.R., 1982]. In modelling the EMC of the group of stations, the following initial data were fed into the computer: the working frequencies of the receivers and transmitters; the territorial location and height of the antennas; and the parameters of the models of the receiving-transmitting equipment and the propagation environment.

The output information comprised the P_{ui} level and the frequencies of the suppressed receiving channel and the interfering transmitter. The results of the calculations of q_i for the six modes of interference indicated in § 2 above (i = 1,6) and three bands (j = 1,3) are given in the table below (Alter, L.S., 1986).

Mode of interference	q_i for frequency band, MHz			q_{ik} for frequency band, MHz	
	40	160	330	40	160
1	0.106	0.106	0.013	0.037	0.072
2	-	0.030	0.047	-	0.012
3	0.008	0.030	-	0.007	0.015
5	0.052	0.016	-	0.031	0.011
6	0.025	0.025	-	0.022	0.022
q_Σ	0.180	0.191	0.060	-	-

The cumulative distribution functions are shown in Figure 1; where curves 1 and 2 correspond to the 40 and 160 MHz bands, curve 3 shows function F_3 for 40 MHz. The value of q_4 is negligibly small (q_4 = 0.0004).

Using the cumulative functions obtained for K = 10 dB on the basis of formula (2), a calculation was made of the values of q_{iK}; these are shown on the right hand side of the table. Then, using the values of q_{ik} in accordance with formula (1), a calculation was made of the values of q_Σ for the equipment with improved EMC parameters. Thus, an improvement of K = 10 dB for only one of the following EMC parameters : level of transmitter out-of-band radiation (i = 1), susceptibility in the reciprocal mixing channel (i = 5), intermodulation (i = 6), spurious channel (i = 3), with all the other parameters remaining unchanged, reduced the probability of reception being suppressed $q_{\Sigma K}$ to 0.117, 0.162, 0.178 and 0.179, respectively, in the improved equipment in the 40 MHz band. It follows that the maximum gain in comparison with the existing probability of q_Σ = 0.180 is obtained if the levels of out-of-band radiation and the susceptibility to interference in the reciprocal mixing channel are reduced. Reductions in susceptibility to intermodulation interference and in the spurious reception channels produce only slight decreases in the probability of reception being suppressed.

Similarly, in improved equipment in the 160 MHz band a K = 10 dB change in one of the following EMC parameters : level of out-of-band radiation (i = 1). susceptibility in the spurious receiving channels (i = 3), level of harmonic radiations in the receiver (i = 2), susceptibility in the reciprocal mixing channel (i = 5), intermodulation channel (i = 6), with all the other parameters remaining unchanged, results in a decrease in the probability of the channel being suppressed $q_{\Sigma K}$ to 0.161, 0.177, 0.179, 0.187 and 0.189, respectively. It will be seen that the maximum gain in comparison with the existing probability of q_Σ = 0.191 is obtained with a reduction of the out-of-band radiation levels.

To obtain a value of $q_{\Sigma H}$ = 0.15 under the conditions considered, it is necessary to reduce the level of the out-of-band radiation of a transmitter operating in the 40 MHz band by K = 5 dB, and in the 160 MHz band by K = 10 dB.

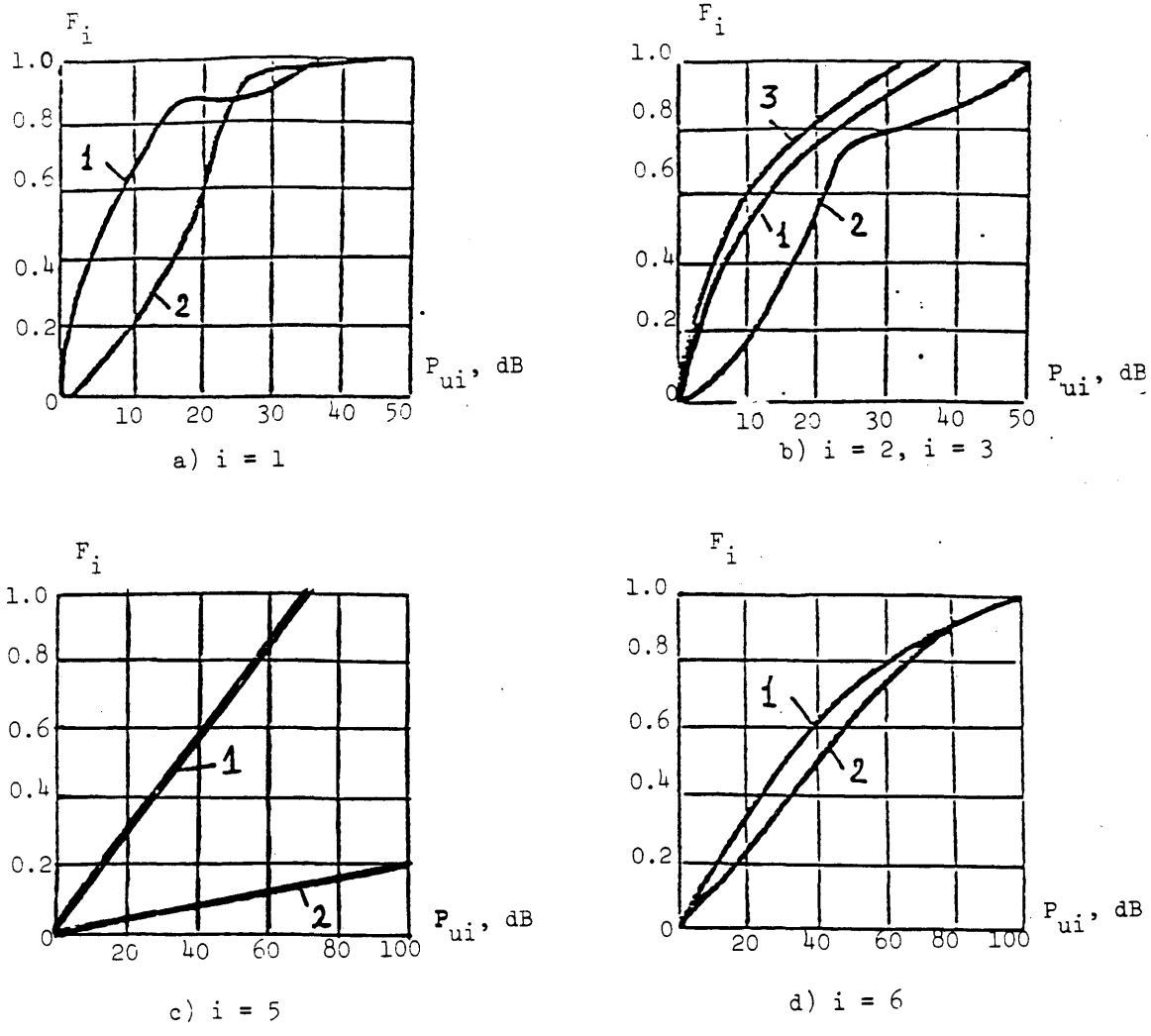


FIGURE 1

The cumulative distribution functions $F_i(P_{ui})$, where P_{ui} - the amount by which the power of i_{th} mode interference exceeds an admissible value.

Curves 1,3 correspond to the 40 MHz band
Curve 2 corresponds to the 160 MHz band.

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REPORT 526-1 *

MEASURES OF VOICE TRANSMISSION PERFORMANCE APPLICABLE FOR ELECTROMAGNETIC COMPATIBILITY ANALYSIS

(Question 44/1)

(1974-1978)

1. General

This Report discusses the measurement of baseband output signals (desired and undesired) to evaluate performance degradation of voice-modulated radio systems. Such measures are important as a means of quantitative assessment of interference effects in spectrum utilization and may be used in presenting "signal to interference" protection ratios as in Report 525. Objective measures are valuable for computing interference effects as discussed in Report 520.

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* The Director of the CCIR is asked to communicate this Report to the Director of the CCITT for comments by the appropriate CCITT Study Group.

The "complete" mathematical modelling of a system's performance is the objective of a prediction analysis. However, there does not exist one mathematical operation for analyzing all types of system performance and the best that can be accomplished is to use the measures that are most appropriate to a particular system (i.e., mean-square measures, probability measures, etc.). The basic difficulty is to determine what exact type of evaluation should be associated with interference degradation. Although considerable research has been conducted on performance degradation evaluation, the desired outputs for receiving systems still reduce to a few basic types. In particular, for voice systems, Articulation Score (the percentage of words correctly received) is still used as the main intelligibility standard. For digital systems, the probability of detection and probability of false alarm are desired. For analogue signals, the mean-square error (or the r.m.s. error) is usually desired.

The following discussion will examine the performance measures of articulation score, articulation index, CORODIM and minimum interference thresholds for voice systems.

2. Articulation score

The basic measure of the intelligibility of a voice system is in terms of the percentage of words correctly understood over a channel perturbed by interference. This intelligibility indication has been designated as an articulation score (AS) and its measurement is usually conducted with specific types of words or syllables as well as specific system parameters [Robertson, 1962]. In an attempt to define the main voice parameters that are involved, experiments have been conducted by varying (at audio frequencies) the word content, bandwidth, audio (S/N) ratio and the types of talkers and listeners that are involved. Through these experiments, articulation scores have been obtained as functions of the above variables and, as one would expect, the scores increased with increasing bandwidth, number of syllables in the words, speaker-listener familiarity and audio (S/N) ratio.

If the receiving system is subjected to a range of distortion or masking conditions, the articulation score may then be determined as a function of the interfering condition. Fig. 1 presents typical articulation score curves for different phonetically balanced (PB) word groupings in which the interference was white noise of various bandwidths [Robertson, 1962]. White noise, which contains a continuous uniform spectrum, is one of the most effective maskers of speech and is often used in speech intelligibility studies as a standard or reference interference.

The articulation testing procedure is not simple nor has it always been standardized. Because it deals with the performance of human beings, the tests can yield variable results in individual cases when proper statistical safeguards are not taken. It is generally necessary to use a number of listeners so as to obtain statistically meaningful results. Proper conduct of the test is tedious and time consuming. The situation is aggravated by the necessity for training the listeners to an efficiency level where the improvement resulting from repeated exposure to most word lists no longer occurs. The test procedures, the material used, and the techniques employed to measure the average power of the desired and undesired signals vary among investigators.

In spite of such shortcomings, the tests provide the most valid objective method available for evaluating the intelligibility of speech communication components or systems. When the articulation score tests are carefully organized, the scores are repeatable 68% of the time within a 2 dB data spread.

The articulation score test is generally used as the basic standard of intelligibility. However, degradation measurements involving a large number of parameter variations, signal-to-noise, undesired signal pulse width, undesired signal pulse rate, etc., can be required for an interference investigation. Because of the time required to run articulation score tests with these parameter variations, it is desired to use an automated measure. A number of techniques that can be automated will be subsequently discussed.

3. Articulation index

The best predictor of voice intelligibility for a channel interfered with by noise and/or interference is the percentage of words which are correctly perceived by a group of trained listeners (the articulation score testing procedure). The previous discussion delineated the difficulties of making AS tests. In order to be able to circumvent these difficulties, the Articulation Index (AI) procedure was developed and validated for white noise interference. In order to validate the AI procedure for different types of receivers interfered with by various types of undesired signals, calibration curves relating AI to simultaneously measured AS scores were examined.

Thompson [1966 and 1967] discusses this problem and shows the possible variation of AI versus AS with the AI measured by means of the Voice Intelligibility Analysis Set (VIAS) procedure for non-pulsed interfering signals. Hatch *et al.*, [1970] discussed the same problem for pulsed signals interfering with AM receivers. These reports indicate that, for the continuous modulation and pulsed interference cases, the relationship between AS and AI was not constant.

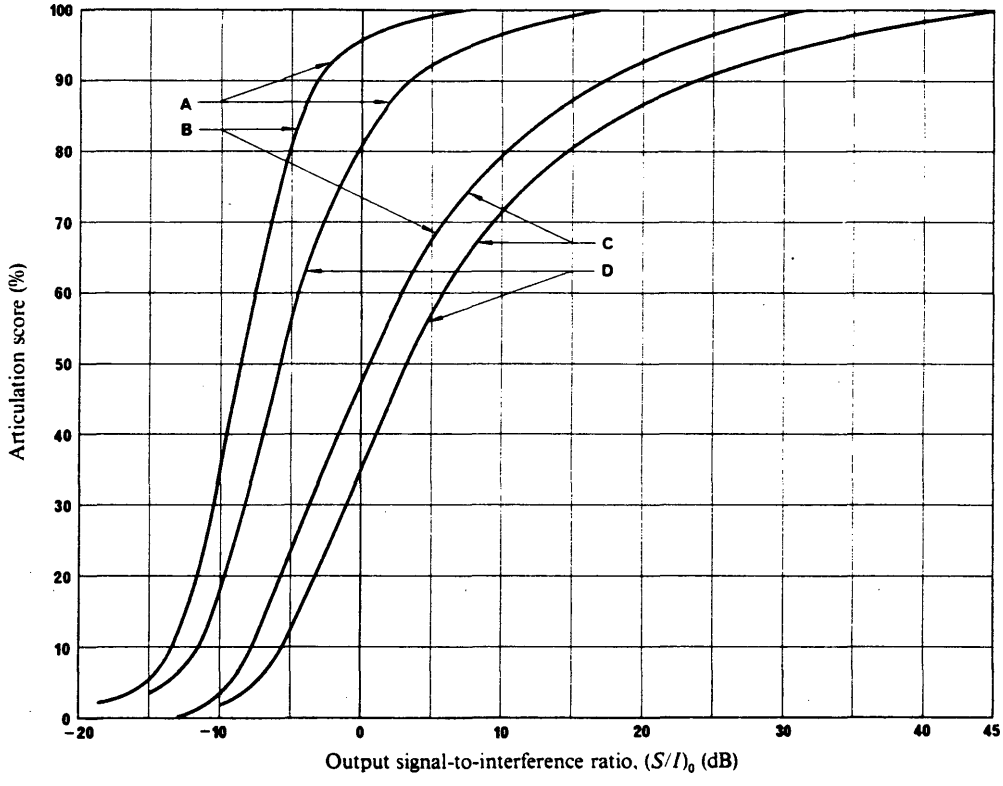


FIGURE 1 — *Articulation score as a function of noise interference*

A: phonetically balanced word groupings: 50

B: white-noise bandwidth $\geq$ 6 kHz

C: phonetically balanced word groupings: 1000

D: white-noise bandwidth $\approx$ 3 kHz

Experience has shown that the smallest value of AI which consistently provides correct information transfer in the presence of normal language configuration is 0.7 on a scale ranging from 0 to 1, and the lowest acceptable value of AI for a useful link is 0.3 [Kryter and Licklider, 1963]. For the narrow-band AM, FM and SSB interference cases, the Thompson reports show that the range of variation in the AS scores for the typical AI criteria of 0.7 and 0.3 are 8% and 26% respectively. The AI criterion of 0.7, therefore, results in a reasonably constant standard in terms of AS intelligibility scores. The AI criterion of 0.3, however, can result in values of AS between 11% and 37% for the cases of AM, FM and SSB modulation with the same types of interference. Therefore, the type of desired modulation and interference would have to be specified in order to accurately determine a lower acceptability threshold.

For the pulsed interference cases, the AS score is approximately independent of AI score [Hatch *et al.*, 1970] because of the large amount of redundancy in a voice signal and the low duty cycle of the pulsed interference. As scores typically vary between about 98% and 95% for AI scores from 0.7 to 0.3, respectively. Because of the large variation in AI scores for a relatively small change in AS scores, an AI value of 0.7 was chosen to describe the lower threshold for pulsed interference [Hatch *et al.*, 1970].

There are a number of approaches that provide a measure of the effects of undesired signals on speech communications systems by calculation and/or instrumentation of a criterion measure in each of a number of bands in the speech frequency spectrum. The articulation index (AI) approach is relatively well-known. Others are the formant intelligibility and pattern correspondence index (PCI) approaches.

All these methods operate on the short-term power spectrum to obtain a performance measure of speech. Basically, the procedures stem from the original work of French and Steinberg that led to the concept of articulation index (AI) [French and Steinberg, 1947]. The results of their work, essentially, determined that one can divide the speech spectrum into n unequal contiguous bands which contribute equally to intelligibility (in terms of AS). The method ideally assumes there are negligible effects on intelligibility due to the speech sounds from one band masking, or in some way affecting, sound components of another band. Effects of noise and other factors (interference, distortion) prevent these bands from making their full contribution to intelligibility. The intensity of speech varies according to the band. For these and other reasons, a weighting factor must be included for each band in recognition of the fact that some bands do not make the maximum possible contribution to speech intelligibility. The weighting factors vary for each band according to the ratio of the speech energy (in that band) to the hearing threshold. When the speech energy level in the band is 30 dB or more above the threshold level, it contributes its maximum value and hence has a unit weighting factor. When the speech energy level is between 0 and 30 dB above the threshold, the band's contribution is proportional to its energy level, in decibels. When the energy level is below the threshold, there is no contribution to intelligibility and the weighting factor vanishes. These weighting factors are additive and the sum can be used with empirical curves to determine the corresponding articulation score.

The French and Steinberg method is, however, still fairly complex and simpler methods have been developed. Another procedure, the tonal method, asserts that the intelligibility of speech depends, not on the absolute magnitude of speech and undesired signal intensities, but rather on the amount by which the speech exceeds the auditory threshold level for a particular type of noise. This perception level is determined in each of twenty equally contributing bands covering 100 to 10 000 Hz for standardized speech and for particular undesired signals. The tonal method, "formant intelligibility" [Tkachenko, 1955] has the property of additivity, such that the overall intelligibility is the sum of the contributions from each band.

The formant intelligibility process is readily automated by feeding pure tones from an artificial voice source, one at a time, to each of the n channels. Listeners then measure the excess noise in each band by attenuating the standard test signal until it is barely audible. The formant intelligibility can then be related to syllabic intelligibility by empirically obtained curves. The importance of this method is that it eliminates most of the variabilities associated with the transmission process and eliminates the articulation scoring procedure. It does not, however, eliminate the listener as the final subjective evaluator.

Other methods have been developed which measure the effects of the undesired signals without subjective listener evaluation. Two of these have led to the development of testing machines based upon the assumption that speech intelligibility resides principally in the short term spectrum [Licklider *et al.*, 1959; Thompson, 1967].

One machine measures a number called the pattern correspondence index (PCI) [Licklider *et al.*, 1959]. This number is an index of the correlation between a speech spectrum without interference and the same speech pattern with interference. The PCI is actually obtained by taking the average spectral difference between recorded sentences without interference and the transmitted sentences with interference. The PCI, theoretically, has a monotonic relationship to articulation score and should be calibrated for white noise interference. It is postulated that the curve for white noise is universally applicable as a function of signal-to-noise, independent of the type of interference. If curves of articulation score as a function of noise are available for the particular undesired-signal case being investigated, a direct translation between PCI and AS can be made. This machine uses an autocorrelation measure of the desired signal and the corrupt output. Therefore, except for possible mechanical deficiencies, this approach is adequate or inadequate depending on the effectiveness of the autocorrelation measure for the particular interference being considered.

The other machine, produced to measure voice intelligibility mechanically, is called the Voice Intelligibility Analysis Set (VIAS) [Thompson, 1967; Fitts, 1963]. This device also operates on the principle, previously described, of dividing the spectrum into a number (fourteen) of unequal continuous bandwidths and measuring the desired-to-undesired signal ratio relative to the hearing threshold. The width of each band is selected such that all bands contribute equally to intelligibility. The sum of the contributions from each band is then averaged over all fourteen bands to produce the composite articulation index. The fourteen VIAS frequency bands are shown in Fig. 2 and the calculation of the articulation index is depicted graphically in Fig. 3. To perform this basic calculation, a synthetic desired speech signal, which consists of a triangle-modulated 950 Hz tone, is transmitted over the test channel and is then measured by the recording portion of the device, in order to determine representative speech levels in the 14 bands. The average power (over a period of 17 s) of the undesired signal in the 14 bands is then measured and from knowledge of the average desired signal in that band, the desired-to-undesired signal ratio is computed. The articulation index is then computed by summing the contribution from each of the 14 bands. VIAS incorporates empirically-derived correction factors to account for the upward spread of masking. This is the phenomenon in which interference at a low frequency masks a higher frequency portion of the voice spectrum. A correction must also be inserted manually for the receiver's frequency characteristics, which are determined by measurement of the system. Thompson [1967] discusses this correction factor in detail. The important difference between the articulation index machines and the tonal method is the simplification to one test signal and the elimination of the subjective evaluation. The VIAS method implies that interfering effects are independent and, consequently, additive. This last statement is especially critical since the use of a number based on this technique, even for the simplest use (i.e., that of system comparison and not performance measurement), requires validation for situations in which the undesired signals (interference) are not additive.

Another device that automatically calculates the articulation index, in a slightly different manner, is the Speech Communications Index Meter (SCIM) [Hudson and Limburg, 1964]. The basic difference between this device and VIAS is the manner in which the synthetic voice signal is generated. In the SCIM machine, a noise spectrum is transmitted that has been frequency-filtered or shaped to correspond to the average voice spectrum. This signal is then filtered into nine frequency bands and used to compute the desired-to-undesired signal ratios. Ideally, therefore, this system has an advantage over VIAS in that the actual synthetic signal power in band n is used rather than an extrapolation of that signal. The SCIM machine also takes into account the upward spread of the masking effect.

The latest version of the automated articulation index calculator is the PSI/COMP machine. The performance of this machine should be very similar to the SCIM machine, since it employs the same basic signal processing.

The American National Standards Institute (ANSI) has specified a procedure for calculating AI [ANSI, 1969]. The major difference in this procedure and the previous machine scoring procedures is that the 20 band method of calculating AI has been used.

Another concept, called CORODIM (Correlation of the Recognition of Degradation with Intelligibility Measurements), has also been developed [Strohmeyer *et al.*, 1967]. This technique is similar to the previous methods in that the baseband power spectrum is again used as a basic measure. It differs from other automatic intelligibility measuring techniques in that it transmits a test signal composed of speech-like sounds representative of phonemic consonants. The degradation manifests itself as an "effective noise spectrum" which is measured and matched to one of a library of reference noise spectra. For each reference spectrum, data are stored relating phonemic recognition probability to speech-to-noise ratio. Thus, by means of the spectrum matching operation and a measurement of signal-to-noise ratio, each component sound of the test signal is assigned a probability of recognition. These values are weighted by phoneme probability of occurrence factors, summed and normalized to obtain a score representative of word intelligibility based on either initial or final consonant recognition. CORODIM evaluates scores for both the initial and final consonants and takes their product for the overall word intelligibility score.

The scores obtainable for CORODIM are directly comparable to listener panels [Strohmeyer *et al.*, 1967]. If sufficient audio spectra have been pretested, the AS results from CORODIM should also be generally applicable for other (but not necessarily all) interfering signals. This technique, therefore, has an important theoretical advantage over all previous automated scoring techniques.

CORODIM has potential use in future voice degradation problems. In particular, it is only necessary to couple the CORODIM process with the simulated receiver output described in Report 520 to obtain simulated articulation scores.

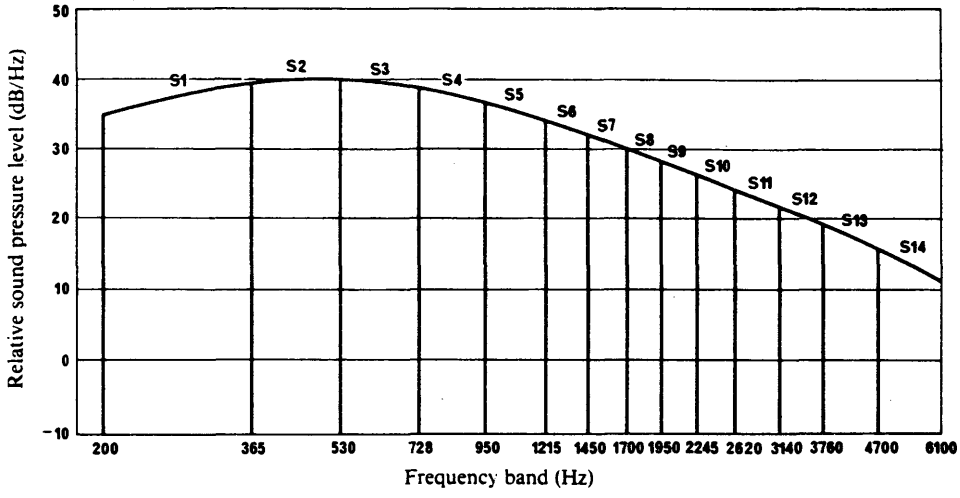
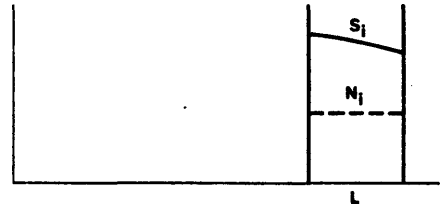


FIGURE 2 — *Long term speech spectrum and associated AI bands*

Note. — The curve represents the voice spectrum, white the S values represent the voice power in each AI band.



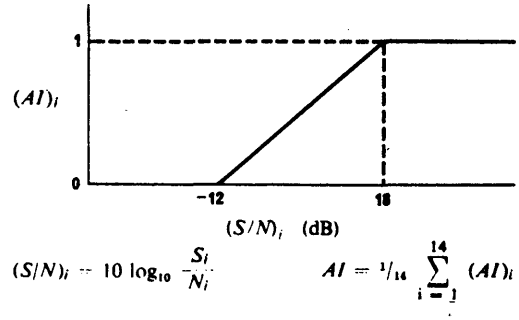


FIGURE 3 – *Theoretical calculation of articulation index score*

S : Desired signal N : Noise

4. Minimum interference thresholds

The degradation threshold, or minimum interference threshold (MINIT), although not a performance threshold measure, is useful in assessing the effects of interference for voice transmissions. It is the level at which the interference is first detected in the audio output. Since this level is obtained through a subjective evaluation, there is an inherent variability due to the human observer and also one due to the manner in which the threshold is defined to the observer. In particular, the threshold level can be determined by decreasing or increasing the interference level relative to a fixed desired signal level. In the first case, the test begins with very noticeable interference and stops when the interference is just perceptible. In the second case, the interference is increased until the subject records that the interference was first heard. The first method is more repeatable than the second, although care must be taken to ensure that the level recorded is indeed the last level that can be heard. This is easily implemented by allowing the subject to adjust the interference level above and below the threshold level to definitely determine that the interference was or was not heard.

The test can also be made without the presence of a desired signal. This type of test would be used for high fidelity or television sound systems where the presence of interference during the time the desired signal intelligence is absent may be unacceptable. A lower threshold interference level would be required for this case than if the desired signal were present, since the desired signal aids in masking the presence of the interference.

The validity of this type of measurement is shown by [Egan, 1944]. Two separate listening crews were used to determine the threshold of perceptibility (MINIT) for speech masked by noise. One crew contained three experienced listeners and the other contained eight inexperienced listeners. The signal (speech)-to-noise ratio (S/N) was then adjusted by each listener until he obtained the threshold of perceptibility. With the exception of one individual in the crew of eight, the maximum variation in the S/N ratio required by each listener was 3 dB. The average difference in S/N between the two crews was less than 1 dB.

The MINIT can be obtained from an objective procedure by measuring the interference level which produces approximately a 1 dB change in the desired signal-to-interference plus noise ratio. This has been documented in the literature as a "Just Noticeable Difference" [Miller, 1947]. The procedure can also be automated by taking the simulated receiver output described in Report 520 and obtaining the input level of interference which corresponds to a 1 dB change in the simulated output desired-to-undesired signal ratio.

Measurements have shown that the MINIT is a function of the interference-to-noise ratio. For a specified signal-to-noise ratio, the MINIT therefore also corresponds to a signal-to-interference ratio. In the linear portion of the receiver, a change in the desired signal level results in a corresponding change in the signal-to-noise and linearly changes the signal-to-interference although the interference-to-noise remains constant.

The MINIT is a threshold that can be used as the boundary between a region of negligible interference and a region of permissible interference and can be used in frequency coordination problems (see [CCIR, 1974-78]).

5. Speech quality

The concept of speech quality encompasses the total auditory impression of speech on a listener. In addition to intelligibility, speech quality includes factors such as loudness, naturalness and clarity, speaker identifiability, timbre and rhythmic character, systematic amplitude or time distortions, and many others. Fidelity of speech is usually measured by subjective methods. Subjective methods as opposed to objective methods are most often used when intelligibility in terms of articulation score is almost 100%. A Special Committee [IEEE, 1969] was organized in 1963 to study and to recommend a practice for *subjective* speech quality measurements. After six years of study, the committee concluded that a single method should not be recommended. However, they did narrow their selection to three *subjective* methods: the Isopreference Method, the Relative Preference Method, and the Category Judgement Method. The three methods all require a trained listening group, who must compare and choose between a test signal and a reference signal. These methods are difficult to implement because of the number of people employed and the amount of time needed to administer a test.

6. Psophometer method

The disturbing effects of additive noise and/or interference on a voice signal have also been evaluated by the psophometer method [CCITT, 1980; Belger and von Rautenfeld, 1965]. A psophometer is a power meter which contains a baseband filter with a frequency response that simulates the typical frequency response of an electro-acoustic transducer (such as an earphone or a loud speaker) and that of human ear. When the disturbing effects of additive noise and/or interfering signals are a major problem of voice quality degradation, hearing tests for assessment of subjective preference quality can be substituted by measurements of the ratio of voice signal power to psophometrically weighted noise and/or interference power. In order to effectively use this method for the general spectrum utilization problem, it is, however, necessary to subjectively evaluate various power ratio

criteria for different types of demodulated baseband AM, FM, SSB, Pulsed, etc. interfering signals – that is, as one example, it is necessary to determine subjectively what psophometrically weighted output signal-to-interference ratio corresponds to "good commercial quality" for FM interference being demodulated by a FM receiver with a desired FM signal present. This type of evaluation has only been accomplished for white Gaussian noise.

7. Conclusions

Several measures of baseband signals (desired and undesired) are described which permit evaluation of voice transmission degradation due to interference and/or noise. Those most easily computerized are the articulation index and the MINIT methods. These methods have been used to calculate voice degradation in the output of the Receiver Performance Model described in Report 520 and documented [ECAC, 1975].

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REPORT 1099

A PROCEDURE FOR ANALYZING THE ELECTROMAGNETIC COMPATIBILITY OF FREQUENCY-HOPPING SPREAD SPECTRUM RADIOS

(Study Programme 18B/1)

(1990)

1. Introduction

A procedure is presented for analyzing the electromagnetic compatibility (EMC) between a victim receiver and a frequency hopping radio system. The victim receiver may be either a frequency hopping or a fixed frequency system. The procedure addresses the intersite EMC issue and is a useful tool for evaluating frequency management strategies. The procedure presented accommodates combinations of radio equipment types, frequency hopset configurations, duty cycles, and radio station locations. Frequency hopset configurations can include either fully shared, partially shared, or non shared frequency resources. The net station location area can either be totally or partially coincident with the area occupied by fixed-frequency radio nets.

2. Assumptions

The procedures are based on the following assumptions:

- Only one net member of a net transmits at any time.
- All of the transmitters of a net have the same characteristics, i.e., same power output, antenna gain, etc; however they may have different duty cycles.
- The basic transmission loss is independent of the bearing to the interference source and depends only on the terrain roughness of the area, frequency, antenna height, and distance.
- The receiver noise is attributable to both internal and man-made or external.
- The basic transmission losses and the minimum required distance separations between the victim receiver and interferring transmitters to prevent frequency hopping dwell erasure need to be computed only for a few frequencies (1 to 3 (low, mid, and high portion of the band), due to the difference in propagation loss from the lower band edge to the upper band edge of the spectrum range under analysis, e.g. 30-50 MHz) rather than for each frequency used by the victim.

- The emission spectra are attributable to transmitter noise as well as the baseband signal.

- The receiver is treated as a passive bandpass filter that operates as a linear device; spurious responses are not considered nor is there a floor on the receiver selectivity.

- The hopping sequences are treated as being random, and a channel may be occupied by one or more transmitters at any instant. This situation will not occur in an electromagnetic environment where all of the frequency hopping transmitters are operating in synchronism (i.e. with respect to the same time base reference system) and with the same transmitter and receiver technical parameters.

- The basic measure of performance is the probability of dwell erasure, i.e. the information in the desired signal is obscured or erased by the reception of an undesired emission during a dwell time. The performance of the receiver during a dwell time is either satisfactory (no erasure) or unsatisfactory (erasure occurs due to interference from one or more environmental transmitters).

- The procedure is limited to analyzing cochannel and adjacent channel interference.

3. Input parameters for the analysis procedure

3.1 Equipment characteristics of the environmental transmitters:

- Power (dBm)

- Antenna gain (dBi)

- Antenna height (meters)

- Bounds on emission spectrum

3.2 Equipment characteristics of victim receiver:

- Noise figure (dB)

- Antenna gain (dBi)

- Antenna height (meters)

- 3 dB bandwidth of selectivity curve

- The minimum required signal-to-(interference-plus-noise) (S/(N+I)) ratio to prevent the information in the desired signal from being obscured or erased.

3.3 The area boundaries for each net in which the exact locations of the members cannot be specified. See Figure 1. The Figure shows that both discrete and random locations can be specified. Random locations for mobile systems need only be specified by an area of location.

3.4 Frequency assignments for:

- Fixed frequency transmitters.
- Hopset used by the frequency hopping net of which the victim receiver is a member.
- Each hopset used by frequency hopping nets of which the victim receiver is not a member.

3.5 Net duty cycles:

- Duty cycle of each net (fraction of time the net is active).
- Fraction of time the net base station is the active member when appropriate.

3.6 Propagation path loss curves for the antenna heights and frequencies involved.

4. Analytical Procedure

The analytical procedure includes four processes. See Figure 2 for a flow chart of the procedure. The procedure calculates the probability P(>0) that one or more environmental transmitters will cause dwell erasure in the victim receiver. Both the long term (compared to typical message lengths) and worse case short term probabilities are computed.

Process 1 determines the probability that the channels in a hopset will be offset by at least j channels.

Process 2 determines the minimum distance separations between a given type of interfering transmitter and the victim receiver required to prevent dwell erasure.

Process 3 determines the probability that the active member (transmitter on) of a net is within the distance interval required to cause dwell erasure if the offset is less than the channel separation required.

Process 4 uses the results of Process 1 and 3 to determine the long term and worse case short term probabilities that one or more of the nets will cause dwell erasure in the victim receiver.

5. Process details

5.1 The hopset probabilities calculated in Process 1 are denoted by $/\bar{A}m(k)/_j$, which is the probability of a channel in the m-th hopset, Hm (made up of $N\bar{m}$ channels), being offset by at least j channels from the receiver when the receiver occupies the kth channel in the band. Figure 3 is a flow diagram of this process, which utilizes the following expression:

$$\left[\bar{A}m(k)\right]_j = \begin{cases} Hv(k), & \text{for } j=0 \\ Hv(k)\left[1 - Hm(K)/Nm\right] & \text{for } j=1 \\ Hv(k)\left[\bar{1}-1/Nm \quad Hm(k+u)\right], & \text{for } j=2,3,.. \end{cases}$$

5.2 Figure 4 is a flow diagram of Process 2. The required input parameters are listed above. The noise power in the victim receiver passband is obtained by:

$$Nr = -174 + NF + 10 \text{ Log } Br, \quad \text{in dBm}$$

where

Br= 3 dB bandwidth, Hz
NF= 10 Log [$\bar{a}$ntilog (NFr/10 + antilog (NFe/10) -1], in dBm

NFr= receiver noise figure, in dB
NFe= external noise figure, in dB

Average levels of man-made noise in residential areas, (Hagn, 1980)

$$NFe = 76.8 - 27.7 \text{ Log } F \text{ (MHz)}$$

In a business area NFe is about 5 dB greater.

The desired signal-to-noise ratio can be calculated as follows:

$$S/N = Pt + Gt + Gr - Lp - Ls - Nr, \quad \text{in dB}$$

where Ls are the equipment system losses, Lp is the propagation loss, and the other parameters are defined above.

The desired S/(N+I) is the threshold level where the victim receiver will not suffer dwell erasure.

The interference margin, Mi, is defined by:

$$Mi = S/N - S/(N+I), \quad \text{in dB}$$

The interference to noise threshold, $(I/N)_t$, is obtained by:

$$(I/N)_t = 10 \text{ Log } (10^{Mi/10} - 1)$$

The mean value of the I/N in the bandwidth of the receiver is given by:

$$I/N = Pt + Gt + Gr - Nr - OTR - OFR - Lp$$

where OTR is the on-tune rejection and OFR is the off-frequency rejection, both in dB. (See Report 654 and Kuebler and Cameron, 1979.) The other terms are defined above.

5.3 Figure 5 is a flow diagram of Process 3 which calculates the zones required by minimum distance separations. Figure 6 depicts the zones. On this Figure, the $(Rz)_j$ define the region bounded by the circles of radii $(Dz)_j$, centered on the victim receiver. Net duty cycle or net on time percentages are also included in this process.

5.4 Figure 7 is a flow diagram of Process 4. The input parameters are the hopset probabilities (from process 1), the zone probabilities (from process 3), and the net hopset assignments. The probability that the members of the zth net will be offset by a sufficient number of channels so that dwell erasure will not occur when the victim receiver is occupying the kth channel is denoted by qz(k) and is obtained by:

$$qz(k) = (1-DCz)+DCz \sum_{j=o}^{J} (nz)_j \, \underline{/}\bar{A}z(k)\underline{\rangle}_j$$

where DCz is the duty cycle of the zth net and $\underline{/}\bar{A}z(k)\underline{\rangle}_j$ is the hopset probability of the z-th net. The probability that none of the nets will cause dwell erasure when the receiver is occupying the kth channel is denoted by q(k). It is obtained by:

$$q(k) = \prod_{z=1}^{Z} qz(k)$$

The probability of none of the nets causing dwell erasure, P(0), at any time is:

$$P(0) = 1/Nv \sum_{k=1}^{K} q(k)$$

The probability of one or more nets causing dwell erasure, P(>0), is

$$P(>0) = 1- P(0)$$

P(>0) is the fraction of time that one or more of the environmental transmitters can individually cause dwell erase.

The other part of Process 4 is the calculation of the worse case short term probability of interference. This involves the determination of which zone contains the interference source and the probability of dwell erasure from that source during the time that source is transmitting. Whether the intermittant dwell erasures would cause unacceptable degradation in a communication link or net may depend on how many dwell erasures occur during some particular interval of time, e.g. during a word or a message segment. This could be described as the number of dwell erasures that occur during a specified number of consecutive dwells. The probability of at least m dwell erasures occurring during an interval of n consecutive dwells, P(>=m), can be calculated using the binomial distribution:

$$P(>=m) = \sum_{i=m}^{n} \binom{n}{i} \underline{/}\bar{P}(>0)\underline{\rangle}^{i} \, \underline{/}\bar{P}(0)\underline{\rangle}^{n-i}$$

where

$$\binom{n}{i} = \frac{n!}{i!\,(n-i)!}$$

6. Conclusion

A procedure has been outlined to calculate the possible interference between fixed frequency and frequency hopping mobile radio systems. The procedure can be calculated manually or be automated. The procedure can be used to determine the possibility for sharing of the frequency resource or the exclusions needed if interference is too great to tolerate. A complete numerical example is provided in Annex .

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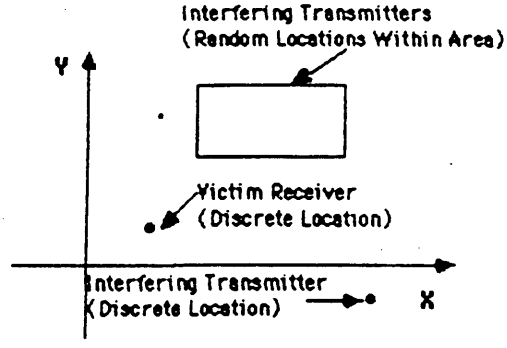


Fig. 1 - Transmitter Deployment Area

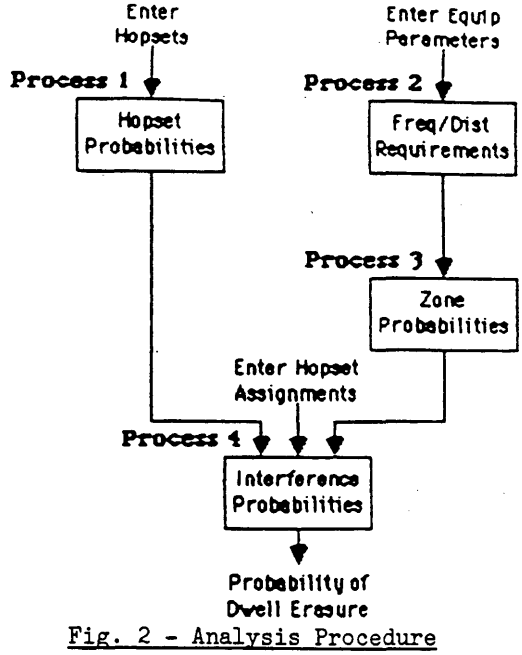


Fig. 2 - Analysis Procedure

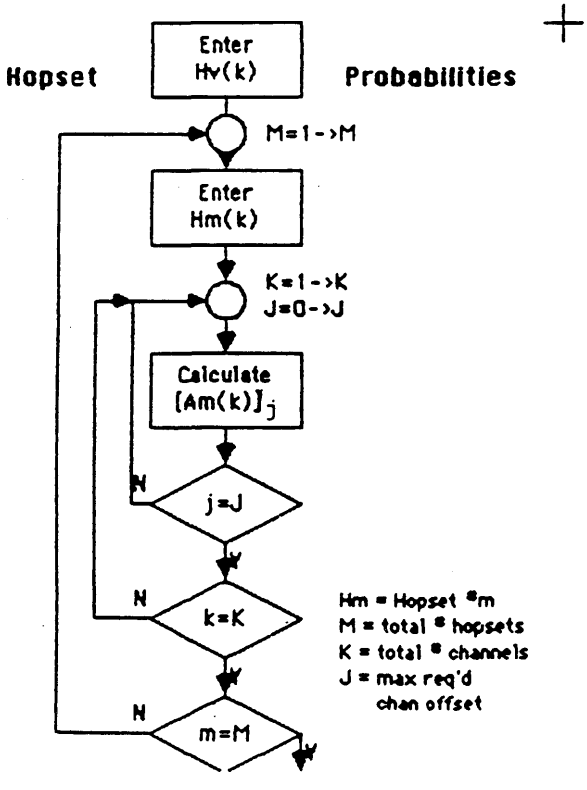


Fig. 3 - Process 1

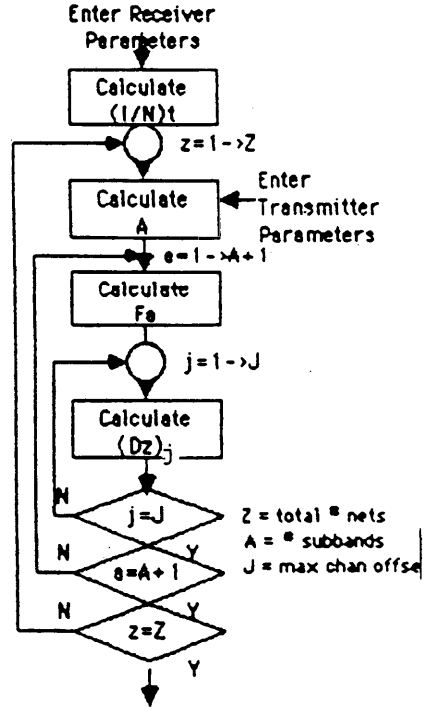


Fig. 4 - Process 2

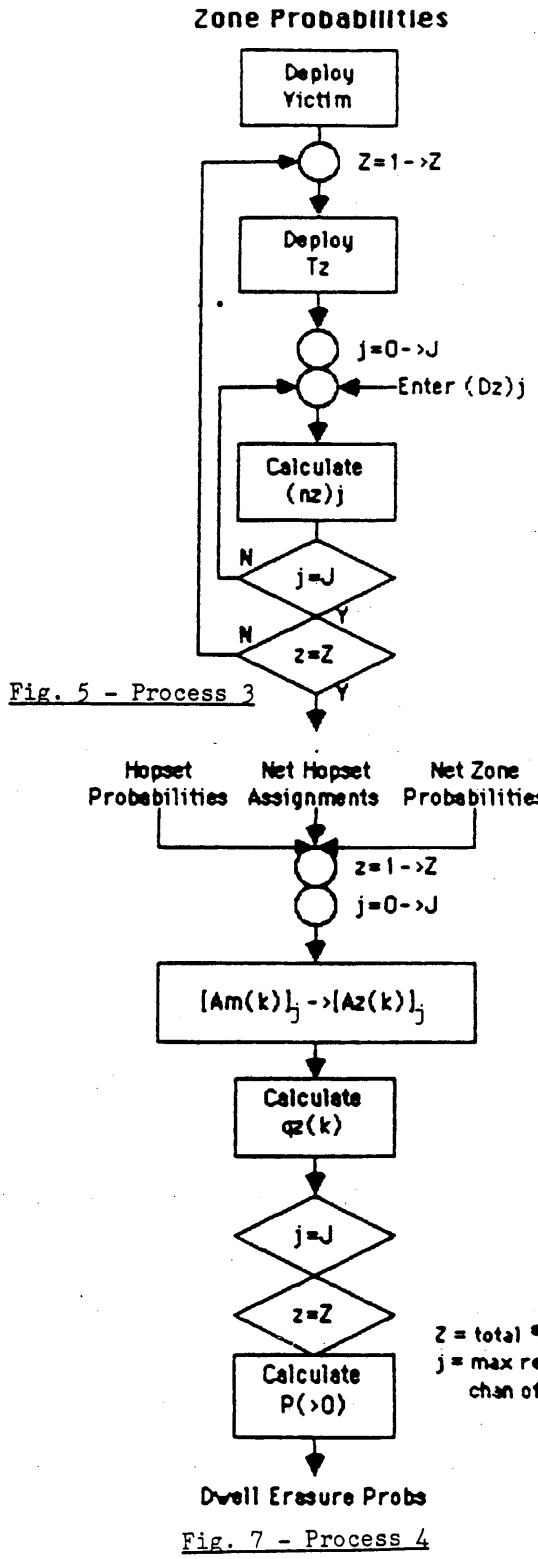


Fig. 5 - Process 3

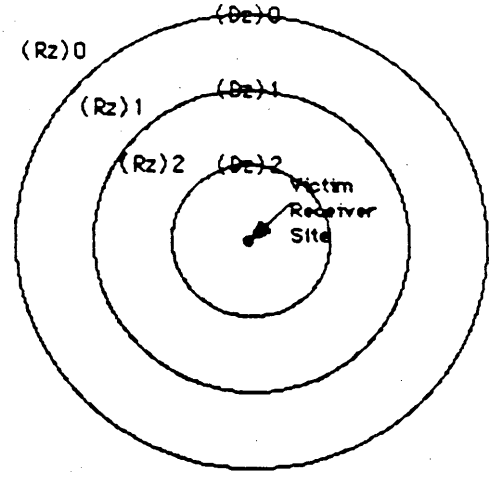


Fig. 6 - Boundary Zones

Fig. 7 - Process 4

ANNEX

A Numerical Example

For this numerical example, the victim receiver is subjected to interference from three transmitters. The location of the victim receiver relative to the geographic boundaries of the three emitters is shown in Figure 8. All of the radio systems are frequency hoppers, the emission is MSK at 15 kbps, the channel spacing is 25 kHz, the I/N is set at $(I/N)_t$, the OTR is zero because both the desired and interfering signals have the same bandwidth, and the OFR is assumed to be:

$$OFR(j) = \begin{cases} 0, & j=0 \\ 40 \log (jFs), & j=1,2,....,6 \end{cases}$$

where Fs is the RF channel spacing, kHz

The victim receiver uses hopset Hv, net number 1 uses hopset H1, and nets 2 and 3 use hopset H2. The hopset probabilities are calculated using Process 1 and are shown in Tables I and II.

For the purpose of this numerical example, the probability of dwell erasure is calculated for three conditions, that is desired link distances of 5, 10 and 20 KM. Each net has a duty cycle of 25%. The basic transmission loss falls off at 40 dB per decade of distance. Dwell erasure occurs when the S/I is assumed to be less than 5 dB.

The condition required to ensure that S/I => 5 dB is:

$$(Li + OFR) = Ls + 5$$

$$OFR - 5 => (Ls - Li)$$

where Li and Ls are the basic transmission losses between the victim receiver and an interfering transmitter and the desired transmitter respectively. This simple relationship is valid in this numerical example because all of the radio systems are identical.

The minimum required distance separations to satisfy the above expression is

$$Dj = \text{antilog } [(\log Ds) - ((OFR(j) - 5)/40)]$$

where Ds is the distance between the desired transmitter and the victim receiver. The required distance separations are listed in Table III.

The zone probabilities obtained using Process 3 are tabulated in Tables IV, V and VI. Using Process 4, the probability of dwell erasure P(>0) that occurs in the victim receiver was calculated for several conditions and the results are listed in Table VII.

The value of P(>0) increases as the length of the desired link, Ds, increases. Net No. 2, which is the most isolated from the victim receiver, produces the smallest probability of dwell erasure. Net No. 3, which is the least isolated, produces the largest probability of dwell erasure.

When the mean I/N is set equal to the $(I/N)_t$, the threshold will be exceeded with a probability of 50%. A lower probability can be obtained when the required path loss is sufficiently large so that $I/N=(I/N)_t - x\sigma$, where x is a confidence factor and σ is the standard deviation of the I/N. For example, if the I/N (in dB) has a normal distribution and $x=1.64$, the I/N would exceed $(I/N)_t$ with a probability of only 5%. The statistical factors are treated extensively in [Hagn, 1980].

TABLE I

Hopset Probabilities $(A1(k))_j$ for hopsets Hv and H1

j \ k	1	3	5	7
0	1	1	1	1
1	1	1	4/5	1
2	4/5	3/5	3/5	4/5
3	4/5	2/5	3/5	3/5
4	3/5	2/5	1/5	1/5
5	2/5	2/5	1/5	1/5

TABLE II

Hopset Probabilites $(A2(k))_j$ for hopsets Hv and H2

j \ k	1	3	5	7
0	1	1	1	1
1	3/4	1	1	1
2	3/4	3/4	2/4	3/4
3	3/4	2/4	2/4	2/4
4	2/4	1/4	2/4	1/4
5	2/4	1/4	0	1/4

TABLE III

Required Distance Separations
(Obtained with Process 2)

j	OFR(j)	Dj Ds=5 km	Ds=10 km	Ds=20 km
0	0 dB	6.67 km	13.33 km	26.60 km
1	28.0	1.33	2.66	5.31
2	40.0	0.67	1.33	2.66
3	47.0	0.45	0.89	1.78
4	52.0	0.33	0.67	1.33
5	55.9	0.27	0.53	1.06

TABLE IV

Zone Probabilities When Desired Link is 5 km

j	(n1)j	(n2)j	(n3)j
0	0.889	1.000	0.403
1	0.106	-----	0.569
2	0.003	-----	0.021
3	-----	-----	-----
4	-----	-----	-----
5	-----	-----	-----

TABLE V

Zone Probabilities When Desired Link is 10 km

j	(n1)j	(n2)j	(n3)j
0	0.558	0.702	0.000
1	0.424	0.298	0.885
2	0.014	-----	0.089
3	0.002	-----	0.014
4	-----	-----	0.005
5	-----	-----	-----

TABLE VI

Zone Probabilities When Desired Link is 20 km

j	(n1)j	(n2)j	(n3)j
0	0.000	0.000	0.000
1	0.930	1.000	0.476
2	0.053	-----	0.413
3	0.010	-----	0.061
4	0.004	-----	0.022
5	0.002	-----	0.010

TABLE VII

Probability of Dwell Erasure in Victim Receiver
(Due to individual nets and to all nets)

Distance to Desired Transmitter	Net #1 P(>0)	Net #2 P(>0)	Net #3 P(>0)	All Nets P(>0)
5 km	0.0019	0.0000	0.0112	0.0137
10	0.0067	0.0047	0.0249	0.0362
20	0.0181	0.0156	0.0562	0.0876

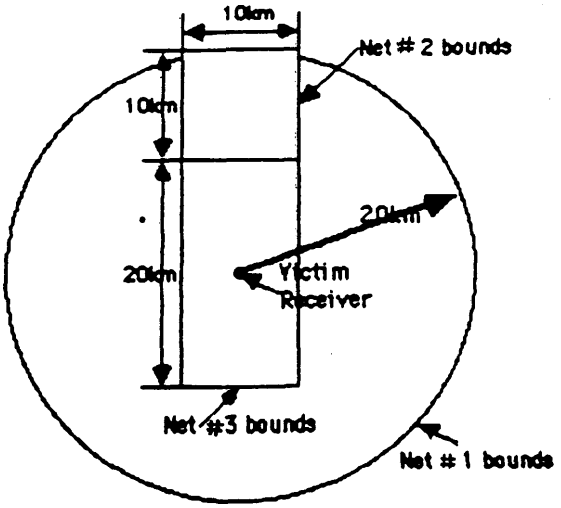


Fig. 8 - Net Boundaries - Example

REPORT 181-4

FREQUENCY TOLERANCE OF TRANSMITTERS

(Question 54/1)

(1963-1966-1974-1978-1982)

1. Introduction

Question 54/1 invites the CCIR to study further frequency tolerances with a view to achieving a more economical use of the radio-frequency spectrum; to predict what the ultimate values of these tolerances might be under currently known conditions of operation; to report upon the possibilities of achieving these ultimate values consistent with technical and economic considerations; and to indicate which of the currently specified tolerances have already achieved these ultimate values.

The greatest difficulty in adopting improved tolerances is the economic problem created by the large number of transmitters in operation and which were manufactured in accordance with existing tolerances. For this reason, and in accordance with Question 54/1, Table I has been prepared for certain categories of stations where it can be foreseen that improved tolerances are desirable and feasible for new equipment manufactured in the near future. The intention is to provide guidance for the manufacturer of new equipment so that, at some future date, improved tolerances may be adopted without serious economic injury, because of large amounts of equipment not able to meet the new tolerances.

2. Factors affecting frequency tolerances

The first consideration, with respect to the efficient use of the radio-frequency spectrum, is that the frequency space lost because of instability should be a small part of the necessary bandwidth used for communication. For purposes of illustration, the figure of ± 1% of the representative bandwidth has been used to provide a guide to the value of frequency tolerances which may be acceptable from the standpoint of spectrum economy in each case.

Reduction in the frequency bandwidth, lost by instability, is not the only criterion with respect to conservation of radio spectrum. For example, in A3E broadcasting and in other forms of A3E emission, the tolerance should be small enough to reduce common channel interference caused by the beat note between off-frequency carriers.

Because of the increased congestion in the LF, MF and HF broadcasting bands the deviations in carrier frequency in accordance with Table I are not likely to ensure the optimum use of the available frequency spectrum. Once the power ratio of wanted-to-unwanted co-channel signals is sufficiently small, small frequency differences can produce annoying amplitude variations before the modulation of the unwanted signal itself causes serious degradation. Especially cyclic variations of constant frequency can cause serious degradation, when the frequency difference is of the order of a few hertz.

In order to avoid such annoying amplitude variations it may be necessary to accept in the future a maximum frequency difference between carrier frequencies of about 0.2 Hz for instance in situations where strong co-channel interference is predominant and should be reduced or eliminated. Equipment of the required stability is available at little if any extra cost.

In radiotelephone single-sideband classes of emission, by a number of stations on a single frequency, the tolerance should be small enough to permit the suppression of the carrier and to provide good voice intelligibility without the readjustment of receivers.

There are certain categories of stations which should not be required to meet a strict tolerance for operational and administrative reasons. An example is mobile radar systems, where the administrative problem of rigid frequency assignments is now unnecessary and, from an operational standpoint, interference is reduced by permitting normal manufacturing tolerances to cause a distribution within the assigned bands.

The advantages of tighter frequency tolerances for transmitters cannot, in all cases, be fully realized unless corresponding improvements are made in receivers.

3. Form of the Table

Table I includes categories of stations with respect to which it is believed possible to make recommendations at the current state of development of technique. As studies continue under Question 54/1, it is hoped that additional categories of stations in additional frequency ranges may be added.

TABLE I

Frequency bands and categories of stations	Representative value of necessary bandwidth of emission (kHz)	Tolerance achievable now, or in the near future (Hz)	Ultimate tolerance (Hz)
(1)	(2)	(3)	(4)
Band 535-1605 kHz			
Broadcasting stations	10	10 (3)	10 (3)
Band 1605-4000 kHz			
1. Fixed stations:			
A3E	6	60	60
H3E-J3E	3	10	10
R3E-B8E	3-6	10	10
2. Land stations:			
A3E	6	20	10
H3E-R3E-J3E	3	20	10
3. Mobile stations:			
(a) Ship stations: H3E-R3E-J3E	3	100	20
(b) Land mobile stations: H3E-J3E	3	100	20
(c) Aircraft stations: H3E-J3E	3	20	20
4. Broadcasting stations	10	10 (3)	10 (3)
Band 4-29·7 MHz			
1. Fixed stations:			
(a) Telephone network with several stations on a single frequency: R3E-J3E	3	10	10
(b) Other fixed stations	1·7 to 12	10	3
2. Land stations:			
(a) Coast stations:			
H3E-R3E-J3E	3	10	10
R7B	3	3	3 (1)
A1A-A1B	0·1	100	100
Other than A1A-A1B	1·7	17	17
(b) Aeronautical stations: H3E-J3E	3	10	10
(c) Base stations: H3E-J3E	3	20	10
3. Mobile stations:			
(a) Ship stations: H3E-R3E-J3E	3	50	20
F1B	0·2	10	10 (1)
(b) Aircraft stations: H3E-J3E	3	20	20
(c) Land mobile stations: H3E-J3E	3	50	20
4. Broadcasting stations	10	10 (3)	10 (3)
Band 29·7-108 MHz		(kHz)	(kHz)
1. Land stations (50 MHz)	16	1	
2. Mobile stations (50 MHz)	16	1	
3. Broadcasting FM stations	200	2	2
4. Television stations	6000	2·5 Hz (2)	2·5 Hz (2)

TABLE I *(continued)*

Frequency bands and categories of stations	Representative value of necessary bandwidth of emission (kHz)	Tolerance achievable now, or in the near future (Hz)	Ultimate tolerance (Hz)
(1)	(2)	(3)	(4)
Band 108-470 MHz		(kHz)	(kHz)
1. Land stations:			
(a) Coast stations (156 MHz)	16	1	
(b) Base stations (470 MHz)	16	1·5	0·36
2. Mobile stations:			
(a) Ship stations (156 MHz)	16	1·6	
(b) Land mobile stations (470 MHz)	16	2·5	
3. Television stations	6000	2·5 Hz (2)	2·5 Hz (2)

(1) This tolerance is required for reception without automatic frequency control, for narrow-band frequency-modulation telegraph channels.

(2) The tolerance of the sound-channel carrier with respect to the visual carrier frequency is 1 kHz.

(3) It is suggested that administrations avoid carrier frequency differences of a few hertz, which cause degradations similar to periodic fading. This could be avoided if the frequency tolerance were 0.1 Hz, a tolerance which would also be suitable for single-sideband emissions.

The purpose of the columns is discussed below:

(1) The frequency bands, category of station and class of emission.

(2) A value of necessary bandwidth regarded as representative.

(3) Tolerances which can be attained now or in the near future, taking into account economic and environmental factors. Such tolerances should not be adopted by an Administrative Radio Conference until new equipment, meeting these tolerances, has replaced a major portion of existing equipment.

(4) Ultimate tolerances which will be equal to, or less than ± 1%, of the representative necessary bandwidth except in unusual circumstances. The choice of these values should take into account system advantages which would be available because of strict tolerances, for example, in A3E broadcasting and the J3E telephony service (see § 2 above). It is not necessary that the tolerances shown be obtainable in the foreseeable future.

REPORT 838-1 *

SPURIOUS EMISSIONS FROM TRANSMITTERS

(Question 55/1)

(1982-1986)

1. Introduction

In considering the transmitter performance that is achievable using the best modern techniques in design and construction, the four types of unwanted emission most liable to be troublesome are examined in turn.

In the case of stations comprising several transmitters operating with neighbouring frequencies and a common antenna or several antennas placed close to each other, additional spurious emissions can be produced. These spurious emissions are caused by the generation of intermodulation products between different fundamental emissions.

Output circuits of most transmitters operating in the high-frequency band are based upon the classical Π-Γ configuration consisting of a Π network followed by an inverted Γ section. This arrangement provides effective tuning and impedance transformation. Elaboration of the basic circuit is often necessary, because the physical size of components used in high-power circuits can result in circuit configurations having undesirable responses which are prone to support unwanted oscillations.

2. Harmonic emissions

2.1 *Harmonic emissions from transmitters*

Harmonics generated in drive equipment are not usually a problem in conventional transmitters due to the suppression provided by the tuned amplifier stages. Where, however, a major proportion of the amplification is provided by wideband amplifiers, the performance of the drive equipment must be adequate to satisfy the overall equipment performance requirements.

The major cause of harmonic emissions from transmitters is the waveform distortion in radio-frequency power amplifiers. Also it frequently happens that the anode and output circuits can combine to produce a resonant condition at a harmonic frequency, and if this occurs the harmonic can be accentuated to an unacceptable level. The elimination or the suppression of such resonances is one of the major design considerations, as alterations in equipment layout tend to shift the frequency at which resonance occurs rather than remove it, unless resistance absorbers in some form are used.

Care in design and equipment layout is necessary to avoid potentially troublesome circuit resonances at the harmonic frequency, which can occur if the reactance of the anode tuning capacitor, in association with its connections, becomes inductive and resonates with the reactance of the valve output capacitance. Equal and opposite reactances can also occur in output coupling and matching networks and form a high impedance circuit at the harmonic frequency, which is then poorly attenuated.

In general with attention to circuit layout in initial design, harmonics within the HF band can be reduced to about -60 dB relative to the wanted emission, using simple Π-Γ anode output circuits, but in high-power transmitters the addition of tuned filters may be necessary to reduce some harmonics to less than the present maximum permitted level of 50 mW.

Especially for short-wave transmitters, the influence of mismatch at the harmonic frequencies on the attenuation of these filters should be taken into account. The design of such filters should be based on a standing wave ratio of at least 10 for the harmonics (see Doc. [CCIR, 1963-66a]).

Although the limits for spurious emission given in Recommendation 329 are adequate for the protection of receivers at large distances from the transmitter, they may not be sufficiently low to avoid interference to other local services (e.g. television and mobile services). Spurious emission outside the high-frequency band can be dealt with in several ways. Low-pass filters may be fitted at the output of the transmitter to attenuate all harmonics above about 40 MHz (see Doc. [CCIR, 1963-66b]). By this means additional suppression up to 30 dB is achievable at reasonable cost. Account should be taken of the possible influence on the necessary values of transmitter tuning

* This Report, which is the result of a merging of Reports 326, 532 and 838 (Reports 326 and 532 are hereby cancelled) should be brought to the attention of Study Group 8.

parameters caused by inserting such filters (see Doc. [CCIR, 1963-66c]). In special cases a further suppression of harmonics at discrete frequencies may be necessary. This may be achieved by the use of resonant $\lambda/4$ open-circuit transmission lines in either coaxial or parallel-twin form (see Doc. [CCIR, 1963-66d]).

2.2 *Radiated field strength*

Studies have been carried out in the USSR on ground-wave fields set up on fundamental and harmonic frequencies of a short-wave transmitting antenna and feeder system. The studies included both theoretical calculations and measurements under actual operating conditions [CCIR, 1970-74a].

The calculations have been made on the basis of formulae derived from [Norton, 1937 and Aizenberg, 1962] for maximum values of the vertical field component E_z and the horizontal component, E_φ. The formulae used for the calculations are given in [Rubinshtein, 1968 and 1971]. They are also contained, together with a description of the experimental method applied to verify the theoretical values, in Doc. [CCIR, 1970-74b].

Three types of short-wave transmitting antenna and feeder systems have been investigated (see Table I).

The calculated values of field strength are given in Table I. They have been calculated for a distance of 5 km from the antenna, for three types of soil (moist, of average moisture and dry), assuming a transmitter power of 1 W for the fundamental and each harmonic frequency, at the input of the antenna-feeder system, and for receiving point heights of 2-20 m above the ground.

For the purpose of measurements, the transmitting antennas were fed by either two-wire or four-wire feeders in the push-pull mode, transformation being made from four-wire feeder to two-wire connection at the transmitter and at the antenna in the appropriate case.

There was satisfactory agreement between the theoretical and the measured ground-wave field strength with deviations not exceeding 10 dB.

In only one case, when the spurious radiation power is supplied from the transmitter to the system input, with a feeder length of more than 500 m, in a push-push mode, the feeder being virtually the only radiator, the maximum values of the vertical field strength component E_z may exceed those given in Table I and reach values up to 5 mV/m with 1 W power of the harmonics at the input of the feeder line. This field strength value may occur in a narrow sector of 15° to 25° symmetrically located relative to the feeder axis. If the feeder is not straight, the radiation pattern will not correspond to that described above and the corresponding values of field strength can be taken from Table I.

3. Parasitic emissions

The frequencies of parasitic oscillations are essentially unrelated to those of the input signals to the transmitter. No general rules on the methods of suppressing these oscillations can be given, because parasitic oscillations bear no relation to the normal operation of the circuit, and each case has to be treated on its merits. There are, however, four major causes of such oscillations:

- ineffective screening between high- and low-level circuits;
- insufficient care in wiring between different sections of the transmitter. As far as possible all cabling and control wiring should be kept out of compartments in which radio-frequency circuits are mounted and all leads entering the radio-frequency compartments should be filtered;
- circuits introduced to cancel an oscillation at one frequency may produce a positive feed-back path at some other frequency, generally higher than the fundamental, e.g. neutralizing circuits in grounded cathode amplifiers;
- transistor amplifiers have voltage-dependent capacitances; if the circuits are insufficiently decoupled at low frequency, oscillation may be present at high frequency, unrelated to the fundamental.

In general, push-pull circuits are more prone to this type of oscillation, because there are more paths inherent in the circuits.

TABLE I – *Maximum calculated values of field-strength components generated on the fundamental and harmonic frequencies of short-wave transmitting antenna-feeder systems at 5000 m distance*

Field-strength components	Harmonic No.	Height of reception point above ground (m)	DPH				RHO/4/1/65				HR/4/4/–			
			φ°	Moist soil ε = 20 σ = 10^{-1} S/m	Soil of average moisture ε = 8 σ = 10^{-2} S/m	Dry soil ε = 3 σ = 10^{-3} S/m	φ°	Moist soil ε = 20 σ = 10^{-1} S/m	Soil of average moisture ε = 8 σ = 10^{-2} S/m	Dry soil ε = 3 σ = 10^{-3} S/m	φ°	Moist soil ε = 20 σ = 10^{-1} S/m	Soil of average moisture ε = 8 σ = 10^{-2} S/m	Dry soil ε = 3 σ = 10^{-3} S/m
Horizontal component E_φ (μV/m)	Fundamental	2	90	7.0	10	20	0	50	65	105	90	90	100	130
		20	90	60	65	75	0	435	450	500	90	785	800	800
	2	2	50	16	20	30	0 10	50	60	70	20	95	100	110
		20	50	145	150	160	0 10	445	450	450	20	850	850	850
	3	2	33	8.5	10	13	5	75	80	90	20 70	175	180	185
		20	33	88	90	90	5	660	660	660	20 70	1750	1750	1750
	4	2	53	27	30	36	18	190	200	220	90	295	300	300
		20	53	250	260	260	18	1700	1700	1700	90	2900	2900	2900
	5	2	42	60	65	70	18	240	250	260	35	300	300	300
		20	42	580	580	580	18	2400	2400	2400	35	3000	3000	3000
Vertical component E_z (μV/m)	Fundamental	20	0	75	30	10	8	325	100	110	0	85	25	25
	2	20	0	140	35	35	30	250	250	270	0	70	70	77
	3	20	0	45	25	25	20	420	500	525	0	130	150	160
	4	20	0	60	65	65	18	450	520	550	0	120	130	135
	5	20	0	45	50	55	35	650	720	720	37	90	95	100

Notes to Table I:

Note 1. – DPH: wideband horizontal folded dipole;

RHO/4/1/65: dual rhombic antenna 65° being half of the obtuse angle of the rhombus, 4 being the rhombus side-length and 1 being the height above ground given in wavelengths corresponding to the optimum frequency of the antenna;

HR/4/4/–: array of dipole antennas, fed in phase, 4 dipoles on each row and 4 dipoles in each stick, aperiodic net reflector.

Note 2. – The maximum field strength values given in the Table are valid for any relationship between the push-pull and push-push components of spurious radiation in a total power of 1 W (apart from the case referred to in the text).

Note 3. – The field strength values quoted for the DPH and RHO/4/1/65 antennas are valid for the entire working frequency range of such antennas (i.e. for fundamental frequencies in the range of 3-30 MHz).

Note 4. – The field strength values given for the HR/4/4/– antennas are valid for the entire working wavelength of 0.41 l, where l is the length of the dipole arm.

Note 5. – Symbols:

$\varphi°$: azimuthal angle relative to the antenna axes where the maximum values of the components E_ϑ and E_φ occur, the axis of the HR/4/4/– antenna being a horizontal straight line in the antenna plane (i.e. parallel to the antenna dipoles), the axis of the RHO/4/1/65 being the large diagonal of the rhombus, the axis of the DPH antenna being a horizontal straight line passing along the axis of the dipole.

ε: dielectric constant of the soil;

σ: conductivity of the soil.

4. Other spurious emissions

Other spurious emissions can arise from a number of causes and generally comprise unwanted components which bear no harmonic relationship to the carrier or modulation frequencies. Those bearing a relationship to the carrier are produced in the frequency-determining circuits of the transmitter, particularly in modulators and frequency synthesizers. They are not produced in final-frequency amplifier stages and they can be either in or out of band. Those bearing a relationship to the modulating frequencies are inter-modulation products and are caused by non-linearity in amplifier stages.

In some transmitters the radiated signal (F_r) is produced by modulating an oscillation at a conversion frequency (F_a) with the signal carrying the intelligence at a sub-carrier frequency (F_b). The modulating process produces a number of components of the form $pF_a \pm qF_b$, p and q being integers. The wanted signal is the component for which $p = q = 1$. The modulator must be designed so that the levels of higher order components are kept to low values.

A case of particular difficulty occurs in practice when

$$F_r = \pm F_b\left(\frac{p - q}{p - 1}\right)$$

usually when F_r is a multiple of F_b, since the unwanted component cannot be removed by tuning the modulator. However, by carefully selecting the levels at which the modulator operates, the unwanted component can be reduced to 60 dB below the level of the wanted component. The production of even-order components is largely avoided by using a balanced modulator. Sufficient selectivity should follow the modulator to avoid the formation of intermodulation products at the grid of the following amplifier.

5. Intermodulation products – co-sited transmitters

A preliminary examination is made below the mechanism that causes the generation of third-order transmitter intermodulation products and of the dominant mechanisms for various types of equipment and services. Further study is needed to consider higher order intermodulation products.

Report 524 describes a model that can be used to evaluate the mutual interference at a single site which employs a large number of transmitting and receiving equipments that are co-located and can produce transmitter intermodulation.

5.1 *Cause of the generation of intermodulation products*

Intermodulation products are generated when two or more signals are applied to a non-linear device.

The frequency of a typical intermodulation product may be expressed as follows:

$$f_i = C_1 \cdot f_1 + C_2 \cdot f_2 + \ldots + C_n \cdot f_n$$

where the sum $| C_1 | + | C_2 | + \ldots + | C_n |$ is the order of the product and $C_1, C_2 \ldots C_n$ are positive or negative integers or zero and $f_1, f_2 \ldots f_n$ are the frequencies of the signals applied to the non-linear device.

The most important intermodulation products are those of the third and higher odd order for transmitters having tuned output stages and antennas, since these products are close to the frequency spectrum of the wanted signals. The frequencies of the third order products are equal to $(2f_1 - f_2)$ and $(2f_2 - f_1)$, where f_1 and f_2 are the fundamental frequencies of the two co-sited transmitters. Higher order products may need to be considered when a large number of transmitters are co-sited. The relationship between the number of high order products and the number of transmitters is given by [Wass, 1948].

In general, intermodulation products can be generated in:

- non-linear elements in structures close to the transmitter,
- the non-linear output stage of transmitters with the following equipment arrangements:
  - transmitter antennas in close proximity of each other,
  - transmitters sharing a common antenna or power amplifier.

Such spurious emissions from a transmitter communal site can cause local or world-wide interference to receivers depending on the service. For bi-directional radio communication services, the receivers for return transmissions, which may be located at, or close to, the transmitter communal site, may suffer severe interference. For HF, VHF and UHF mobile services, receivers and transmitters are often co-sited. For LF and MF services, the receivers and transmitters are usually on separate sites.

5.2 *Transmitters with their antennas close to each other*

Mutual coupling between antennas causes unwanted signals to be fed back into each transmitter power output stage. The output stage of a power amplifier can present an effective non-linear impedance to unwanted signals fed back into the transmitter and hence intermodulation products can be generated and re-radiated. The magnitude of the resultant radiated spurious signal depends primarily on the:

- power of the interfering transmitter;
- antenna coupling loss. This is the ratio of the power emitted from one transmitter to the power of that emission at the output of victim transmitter which may produce the unwanted intermodulation product;
- conversion loss: this is the ratio of the power of the interfering signal from an external source and the intermodulation product, both measured at the output of the victim transmitter. The frequency selectivity of the transmitter output and antenna are not included;
- frequency selectivity at the transmitter output circuits and antenna.

5.2.1 *Antenna coupling loss*

The antenna coupling loss depends on the type of antennas employed. The following two paragraphs consider specific services and further study is required to consider other services.

(a) *Antenna coupling for VHF and UHF mobile communication services*

For VHF and UHF mobile communication services, the antenna coupling loss is generally of the order of 15 to 30 dB, where the frequency of the interfering signal is within ± 5% of the frequency of the wanted signal. Report 524 derives a relationship between antenna coupling loss and antenna separation and orientation. In general, the coupling loss between closely spaced antennas employing vertical polarization is greater for vertical separation than for horizontal separation for similar spacing between antennas [Sachs, 1980]. Filters and isolators can be used (see Report 739) to increase the coupling loss between transmitters.

(b) *Antenna coupling for LF, MF and HF services*

Table II shows that for LF and HF antennas the coupling loss between antennas is usually in excess of 20 dB. The coupling loss is dependent on the ground spacing between antennas which is normally restricted because of the limited land area available [Sosin and Butt, 1976]. The design aim is to ensure that the desired antenna radiation patterns are not distorted by mutual coupling by more than 2 dB. This usually requires the coupling loss to be at least 25 dB.

TABLE II – *Typical coupling loss between antennas at LF, MF and HF*

Frequency band	Type of antenna	Typical coupling loss (1) (dB)
40 kHz-150 kHz	Umbrella, "T"	20
140 kHz-30 MHz	Vertical radiator	25
2 MHz-30 MHz	Wide-band omni-directional employing vertical polarization	30
2 MHz-10 MHz	High angle radiators	25
2 MHz-30 MHz	Log periodic	20

(1) The values of coupling loss between like antennas are based on measurements carried out at a number of antenna sites.

5.2.2 *Conversion loss (for third order products)*

The conversion loss of a transmitter output stage is determined by the broadband non-linear function of the output stage and the degree of isolation between the non-linear function and the load. This conversion loss does not take into account the frequency selectivity of the output stage of the transmitter.

For land mobile radio transmitters which use solid-state power amplifiers, the conversion loss is in the range of 3 to 20 dB. The reason for such low conversion loss is that the output transistor operates in a saturated mode and the final collector exhibits non-linear reactive and resistive effects.

For HF communication transmitters designed for SSB, the conversion loss of a high power valve amplifier is in the order of 50 dB. The reason for this high conversion loss is that a tetrode valve is normally used, operated in a linear mode.

For LF, MF and HF broadcasting transmitters, the conversion loss of a typical high power valve amplifier may be as low as 10 dB. The reason for this low conversion loss is that the valve anode operates in a saturated mode resulting in a mixing effect which is not as efficient as for the case of transistors considered above.

5.2.3 *Frequency selectivity of transmitter output and antenna*

The frequency selectivity of transmitter output stages is determined by the output circuit Q-factor. This will vary according to the type of service provided by the transmitter and the frequency band. The antenna does not necessarily improve the transmitter effective selectivity. The selectivity of the victim transmitter output circuit operates on interfering signals entering the transmitter output port and upon generated spurious signals reradiated subsequently.

Transmitters with a high output Q-factor will restrict radiated intermodulation products primarily to within the service bandwidth although some interference may be caused to nearby services operating in the adjacent frequency bands.

For LF transmitters operating at frequencies below about 100 kHz the tuning of the transmitter output circuit and antenna is usually sufficient to severely attenuate intermodulation products. The reason for this is that the restriction on the size of practical LF antennas results in the electrical length of the antennas being a small fraction of a wavelength. Consequently, the Q-factor of a complete anode/antenna tuning circuit becomes very high.

5.2.4 *Interference range of intermodulation products*

The interference range is determined primarily by the following:

- effective radiated power of the intermodulation emission,
- receiver minimum usable signal level and protection ratio requirement,
- propagation characteristics.

(a) *Effective radiated power (e.r.p.) of intermodulation emission*

This may be readily calculated if the level of the intermodulation signal(s) is known relative to the carrier of the wanted signal.

Tables III, IV and V show the level of the third-order intermodulation products below the carrier for various types of transmitters.

(b) *Receiver characteristics*

The interference range is determined by the distance at which the spurious signals from transmitter sites are of sufficient magnitude to seriously degrade the quality of the wanted signal at the receiver output. The maximum permissible magnitude of an interfering signal is specified by the receiver minimum usable signal and protection ratio requirements.

(c) *Propagation characteristics*

The propagation characteristics appropriate to the frequency band and antenna may be determined by consulting the relevant Recommendations 368, 370, 435 and 533. The interference range for HF sky-wave propagation and for LF communications is likely to greatly exceed the interference range for MF ground-wave propagation used for broadcasting services. Tables III, IV and V are examples of the likely effective interference range for various types of transmitters.

VHF and UHF television transmitters generally have combining networks with minimum cross-coupling and high selectivity, and administrations have encountered few problems involving the generation of intermodulation products.

TABLE III — *Example of calculation of interference range of third order intermodulation products — co-sited HF communication transmitters*

Frequency (MHz)	Frequency separation (MHz)	Transmitter output		Intermodulation emission (dBW) (2)	Noise degradation factor (dB) (3)	Effective noise at receiver input (dBW) (4)	Required path attenuation (dB) (5)	Interference range		
		Q-factor	Attenuation (dB) (1)					Free space (km)	Ground wave (km) (6)	Sky wave (km) (7)
2	0.1	10	10	−50	42	−127	77	85	25	
2	0.2	10	18	−58	42	−127	69	34	12	(7)
2	0.5	10	32	−72	42	−127	55	7	<5	
10	0.1	10	1	−41	30	−139	98	190	8	
10	0.2	10	3	−43	30	−139	96	150	5	(7)
10	0.5	10	10	−50	30	−139	89	67	<5	
10	1.0	10	18	−58	30	−139	81	27	<5	
10	2.0	10	28	−68	30	−139	71	9	<5	
30	0.2	15	1	−41	15	−154	113	356	10	
30	0.5	15	4	−44	15	−154	110	251	7	
30	1.0	15	10	−50	15	−154	104	126	<5	(7)
30	2.0	15	19	−59	15	−154	95	45	<5	

(1) This is the attenuation of the incoming unwanted signal from the adjacent transmitter plus the attenuation of the produced intermodulation product.

(2) These values of spurious emission apply for the following transmitter parameters: e.r.p. = 1 kW, antenna coupling loss = 20 dB, conversion loss = 50 dB (linear power amplifier designed for SSB operation).

(3) Atmospheric, environmental and galactic noise above thermal noise at quiet areas and times.

(4) The effective noise input is given by the thermal noise at the receiver input for a 3 kHz bandwidth (−169 dBW) plus the appropriate noise degradation factor. It is assumed that intermodulation products fall into the communication band of frequencies.

(5) It has been assumed that the maximum permissible power of spurious signals at the receiver input should not exceed the effective noise power.

(6) These expected interference ranges are typical under the conditions of (3). When a propagation mode other than the ground wave predominates, Recommendations 368 and 370 should be consulted to determine the interference range.

(7) Sky-wave propagation loss exceeds the free-space loss so that interference in the sky-wave path can be neglected.

TABLE IV — *Example of calculation of interference range of third order intermodulation products — co-sited HF broadcasting transmitters*

Frequency (MHz)	Frequency separation (MHz)	Transmitter output		Intermodulation emission (dBW) (2)	Noise degradation factor (dB) (3)	Effective noise at receiver input (dBW) (4)	Required path attenuation (dB) (5)	Interference range			
		factor Q	Attenuation (dB) (1)					Free space (km)	Ground wave (km) (6)	Sky wave (km × 1000) (6) and (7)	
										Night	Day
2.4	0.1	10	8	9	40	−129	138	79 000	180	22	0
2.4	0.2	10	16	1	40	−129	130	32 000	120	9	0
2.4	0.3	10	29	−12	40	−129	117	7 000	80	2	0
9.6	0.1	10	1	16	30	−139	155	140 000	100	22	18
9.6	0.2	10	3	14	30	−139	153	111 000	85	22	15
9.6	0.5	10	10	7	30	−139	146	50 000	65	22	7
9.6	1.0	10	19	−2	30	−139	137	17 600	55	8	2
9.6	2.0	10	29	−12	30	−139	127	5 600	30	3	1
26	0.2	15	1	16	15	−154	170	290 000	55	0	22
26	0.5	15	5	12	15	−154	166	183 000	45	0	22
26	1.0	15	11	6	15	−154	160	92 000	35	0	22
26	2.0	15	21	−4	15	−154	150	29 000	25	0	11

(1) This is the attenuation of the incoming signal from the adjacent transmitter plus the attenuation of the produced intermodulation product.

(2) These values of spurious emission apply for the following transmitter parameters: e.r.p. = 50 kW, antenna coupling loss = 20 dB, conversion loss = 10 dB (anode modulated-saturated).

(3) Atmospheric, environmental and galactic noise above thermal noise, at quiet areas and times (Report 322).

(4) The effective noise input is given by the thermal noise at the receiver input for a 3 kHz bandwidth (−169 dBW) plus the appropriate noise degradation factor. It is assumed that intermodulation products are in the communication band of frequencies.

(5) It has been assumed that the maximum permissible power of spurious signals at the receiver input should not exceed the effective noise power.

(6) These expected interference ranges are typical under the conditions of (3). When a propagation mode other than the ground wave predominates, Recommendations 368 and 370 should be consulted to determine the interference range.

(7) Typical sky-wave ranges have been determined from Report 252.

TABLE V — *Example of calculation of interference range of third order intermodulation products — co-sited UHF mobile communication transmitters (450-470 MHz)*

Frequency separation (MHz)	Attenuation (dB) (1)	Intermodulation emission (dBW) (2)	Minimum usable field strength (dB (μV/m)) (3)	Protection ratio (dB)	Interference range (km) (4)
< 0.25	0	−18	20	17	5
5	2	−23	20	17	4

(1) This is the attenuation of the incoming signal from the adjacent transmitter plus the attenuation of the produced intermodulation product.

(2) These values of spurious emission apply for the following transmitter assumed parameters: e.r.p. = 10 W, antenna coupling loss = 20 dB, transmitter conversion loss = 8 dB. e.r.p. of intermodulation product type $(2f_1-f_2)$ and $(2f_2-f_1)$ is 10 dBW − 20 dB − 8 dB = −18 dBW.

(3) Assume receiver minimum usable field strength is +20 dB (μV/m) and let the protection ratio required be 17 dB. Receiver antenna has 0 dB gain relative to a half-wave dipole and is 1.5 m above the ground.

(4) The maximum permissible spurious emission field strength at the receiver is +3 dB (μV/m), the transmitter effective antenna height = 30 m and the median interference range may be found from Report 567.

5.3 *Generation of intermodulation product emissions by transmitters feeding a common antenna or power amplifier*

Co-sited transmitters are connected to a common antenna by a combining unit subsequent to power amplification, or by combination prior to high power amplification.

5.3.1 *Transmitters sharing a common antenna*

It may be possible for energy transfer between transmitters to occur via the combining unit.

This cross-coupling via the combining unit may be regarded as equivalent to antenna coupling loss. Hence, § 5.2 applies, by replacing values for antenna coupling loss by the cross-coupling loss.

5.3.2 *Transmitter sharing a common power amplifier*

Intermodulation products are likely to be generated in the non-linearity present in a high power amplifier.

Further study is required into the general problem of co-sited transmitters operating in this mode.

5.4 *The radiation of intermodulation product spurious signals by non-linear elements close to antennas*

5.4.1 *From non-linear elements in structures close to antennas*

Although not strictly a question of transmitter design, the possibility of unwanted harmonics and intermodulation components by the excitation of conductors containing non-linear contacts in antennas and elsewhere in the vicinity of the transmitters should not be overlooked (see Doc. [CCIR, 1963-66d]). It is shown that, in at least one case of a multi-transmitter installation, the level of spurious emission can increase by up to 30 dB due to this cause (see Doc. [CCIR, 1963-66e]). Non-linear elements can be formed at metal-to-metal junctions in antenna masts and feeders and these elements may be closely coupled to nearby antennas [Watson, 1980]. Some non-linear elements are caused by the unavoidable use of dissimilar metals and corrosion.

Corrosion is an ever-present threat especially at coastal sites or in areas of atmospheric pollution. Further study is necessary in order to determine typical coupling and conversion loss so that the interference potential of spurious emissions generated in this way can be assessed. Careful attention to the bonding of all joints in metallic structures and antennas is the only way to avoid this undesirable effect.

5.4.2 *From receivers co-sited with transmitters*

Signals, generated by the co-sited transmitters, at the receiver input can cause the receiver to generate intermodulation products which are re-radiated.

The generation of intermodulation products depends primarily on the following:

- cross-coupling loss between transmitting antennas and the receiving antenna,
- receive antenna frequency selectivity,
- receiver input stage conversion loss,
- receiver input stage frequency selectivity.

The mechanism of the generation of intermodulation spurious signals is similar to that considered in § 5.2.1 to 5.2.3 inclusive.

6. Screening and earthing

Because all transmitters generate harmonic and other unwanted frequencies, an efficient screening enclosure (comprising walls, floor and roof) is necessary to minimize their radiation. The presence of exterior circulating radio-frequency currents in members and panels forming the enclosure must be avoided and good electrical contact between all parts is essential. The use of closely-spaced bridging contacts is advisable for all removable panels and doors, also wedge-action catches are preferred.

Openings and slots in transmitter enclosures can radiate strongly and in a highly directional manner at frequencies for which they are resonant. This can be avoided by placing a suitable screen behind the openings.

Conductors carrying power supplies to the transmitter and control and other cables connected to remote equipment may need to be screened and have filters inserted where they pass through the walls of the enclosure. Filters provided in radio-frequency output feeders should be individually screened and mounted in the wall of the screening enclosure.

Efficient earthing of the transmitter framework and enclosure is desirable to minimize radiation due to the existence of unavoidable radio-frequency currents in the transmitter enclosure and in any case is essential for the safety of personnel. Where balanced output feeders are employed, earth currents should be minimized by paying attention to the symmetry of the feeder system. The design of coaxial output feeder terminations should be such as to avoid the production of voltages on the outer surface of the feeder. Where changes from the unbalanced to the balanced mode of operation occur, the balance point of the transformer or other device used should be connected to an efficient low-reactance earth.

It is common practice to make radio-frequency connections between units within the transmitter by coaxial cables. These cables should be terminated as directly as possible at the generator and load terminals of the equipment, to avoid the introduction of undesirable paths (via frameworks, etc.), which can become efficient resonators at harmonic frequencies and often contribute to severe degradation in crosstalk performance of coaxial cables. Coupling between feeder cables or equipment casing is not usually significant provided good engineering practices have been observed.

7. Conclusion

The levels of spurious emissions from transmitters are largely controlled by various features of transmitter output coupling network design and generally can be reduced to the extent required by the Radio Regulations. However, additional suppression is frequently necessary, to avoid interference to other services operating outside the HF band.

Other unwanted emissions in the HF band can also arise from the frequency generation and frequency conversion equipment and also from intermodulation products caused by non-linearity of amplifier stages and of conductors external to the transmitter. These spurious emissions can be kept to low values only by taking account of the design factors involved, a number of which have been referred to above.

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[1963-66]: **a.** I/46 (USSR); **b.** I/9 (United Kingdom); **c.** I/45 (USSR); **d.** I/3 (Federal Republic of Germany); **e.** I/8 (United Kingdom).

[1966-69]: I/21 (USSR).

[1970-74]: **a.** 1/48 (USSR); **b.** 1/153 (USSR).

REPORT 839

PROCEDURE FOR CALCULATING THE AMPLITUDE OF TRANSMITTER INTERMODULATION (TIM) PRODUCTS

(Questions 44/1 and 55/1)

(1982)

1. Introduction

Reports 524 and 838 describe transmitter intermodulation analysis techniques which should be considered in a co-site analysis. Interference mechanisms treated include adjacent signal effects (which, in turn, includes the effects of broad-band transmitter noise), receiver spurious responses, receiver intermodulation, and transmitter (narrow-band) spurious emissions. All of these mechanisms, involving antenna-to-antenna coupling, should be considered when evaluating the performance of co-sited transmitters and receivers.

This Report describes a modified version of the transmitter intermodulation model (TIM) presented in Report 524, based on recent findings [Maiuzzo and Mackouse, 1981]. Some numerical estimates of typical values will be given. It should be noted that relatively large uncertainties can be anticipated for transmitter intermodulation calculations.

2. Description of model

Two-signal TIM effects require formulations of frequency and power relationships. The frequency relationship is:

$$f_r = \pm mf_a \pm nf_i \qquad (1)$$

where:

f_r: frequency of the receiver of interest,

f_a: frequency of the affected transmitter,

f_i: frequency of the interfering transmitter,

m and n: positive integers greater than zero.

The order of the mix is $m + n$.

The power relationship is:

$$P_R = n[P_I - \beta_{ai}] - \beta_{ar} - K_{(m),n} \quad (2)$$

where:

P_R: power level (dBm) of the TIM product, at frequency f_r, at the output of the affected transmitter;

P_I: power level (dBm) of the interfering power level at frequency f_i, at the affected transmitter output terminal. This power level may be obtained by subtracting coupling loss between the transmitters from the power level of the interfering transmitters (see Report 524, equation (1) or practical values given in Report 838);

β_{ai}: attenuation of the interfering power (dB) due to the output and antenna circuits of the interfered-with transmitter at the frequency f_i;

β_{ar}: attenuation of the generated intermodulation products (dB) due to the output and antenna circuits of the interfered-with transmitter at the frequency f_r.

$K_{(m),n}$ is a conversion loss coefficient which is different from the term used in Report 524. It can be shown that this factor is relatively independent of m [Maiuzzo and Mackouse, 1981]. Hence, $K_{(1),1}$ is approximately equal to $K_{(2),1}$, $K_{(3),1}$, etc. The bracket noted for the first subscript is meant to imply the independence of m; the number is needed to account for the frequency relationship.

The attenuation terms β_{ai} and β_{ar} are functions of Q (the circuit quality factor), N (the number of tuned stages, normally one for transmitters) and frequencies of interest. The formulation is:

$$\beta = 10\,N \log \left[1 + Q^2 \left[\frac{f}{f_0} - \frac{f_0}{f}\right]^2\right] \quad (3)$$

f_0: tuned frequency,

f: either f_i or f_r, as appropriate.

Table I provides a summary of Q values, based on a survey of approximately 60 transmitters of low and medium powers situated at a variety of locations. As can be seen, Q tends to increase with operating frequency. As indicated above, N is almost always equal to one.

TABLE I — *Range of Q values*

Frequency range (MHz)	Average value (1)	Range of Q values
2-30	10	3-25
30-100	15	5-30
100-200	25	15-30
200-400	45	20-60

(1) When considering high power transmitters, the average Q values shown in Table I would be expected to be 10, 10, 15 and 20 for the frequency ranges 2-30, 30-100, 100-200, and 200-400 MHz respectively.

The range of $K_{(2),1}$ values is 3-40; 18 is an estimated mean value. In effect, these values will be usable for all values of $K_{(m),1}$.

A preliminary survey indicates that the following approximate relationships may be applicable:

For vacuum tube devices:

$$K_{(m),n} \approx K_{(m),1} + 40\,(n - 1) \quad \text{(dB)} \quad (4)$$

For solid-state devices:

$$K_{(m),n} \approx K_{(m),1} + 30\,(n - 1) \quad \text{(dB)} \quad (5)$$

It is likely that additional research will provide more precise estimates. For example, it is likely that class of operation (e.g., amplifier classes, A, B or C) will have some effect on the degree of non-linearity that is produced.

It should also be noted that tuned circuits exhibit an effective attenuation "floor", i.e., a single tuned stage will rarely provide more than 40 dB attenuation. However, when TIM products are produced outside the transmitter's tuning range, additional losses, due for example to mismatch effects, can be anticipated. The same situation will prevail with regard to an out-of-band interfering signal.

3. Conclusions

Given suitable coupling loss models for co-site and inter-site distances (see Report 524), the preceding material can be used to estimate typical or worst-case IM product levels. It should, however, be noted that receiver intermodulation (RIM) effects can, in many cases, overshadow TIM effects. Both types of interference mechanisms as well as the other major co-site phenomena noted above should be considered in co-site interference evaluations.

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MAIUZZO, M. and MACKOUSE, E. [18-20 August, 1981] Transmitter intermodulation product amplitudes. Proc. IEEE International Symposium on Electromagnetic Compatibility, Boulder, CO, USA.

REPORT 533-1 *

SENSITIVITY OF RADIO RECEIVERS FOR CLASS OF EMISSION F3E

(Question 57/1)

(1974-1978)

1. Introduction

Two methods that can be used to obtain the reference output for sensitivity and other measurements (i.e. selectivity, spurious response, etc.) of a receiver are:

- The "SINAD method", which uses the ratio (Signal + Noise + Distortion)/(Noise + Distortion) (SND/ND) as reference magnitude. The value of both the numerator (SND) and the denominator (ND) is determined in the presence of modulation. The desired signal (S) is removed in the denominator by means of a distortion analyser notch filter.
- The "signal plus noise plus distortion to noise method", which uses the ratio (Signal + Noise + + Distortion)/(Noise). The ratio is usually abbreviated as SND/N_0 where N_0 is the noise determined in the absence of modulation. **

A modification of the above methods can be obtained by using a filter at the receiver output in order to obtain a standard filtered output signal. For the particular case of a receiver connected to a telephone network, a psophometric filter is typically used to represent the subjective effects of noise on a voice signal through a typical telephone handset. Therefore, a special method of measuring output criteria for a voice system could be used which combines either of the above two methods with a psophometric filter. ***

It should be noted that presently Recommendation 331, § 11, and Recommendation 332, § 4 and 6, and also the draft IEC Standard recommend that the SINAD method with no special filtering be employed for measurements of single-channel FM receivers.

The following compares various test results using the SINAD method and the SND/N_0 method with and without a psophometric filter.

The SINAD and SND/N_0 ratios were determined with and without the psophometric filter as shown in Fig. 1.

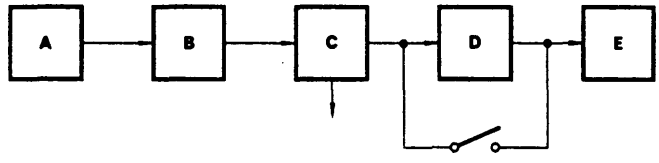


FIGURE 1

A: audio-frequency generator
B: signal generator
C: test item
D: psophometer weighting network
E: distortion meter

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* This Report is brought to the attention of Study Group 8.

** The spectral characteristics of the noise in the absence of modulation N_0 are not the same as the spectral characteristics of the noise N in the presence of modulation.

*** According to CCITT Recommendation P.53, Volume V (*Yellow Book*).

2. Comparison of two methods of measuring sensitivity

[CCIR, 1970-74] gives the results of a comparison between the SND/N_0 method and the SINAD method of measurement for a typical 160 MHz FM radio telephone receiver used in conjunction with the public telephone system in Switzerland. This comparison is given both with and without the psophometric filter for the SND/N_0 and the SINAD method. Figure 2 summarizes this comparison and indicates that for signal-to-noise ratios up to $\approx$ 20 dB, the two methods of measurement give identical results if both ratios are determined either with or without a psophometric filter. Above 20 dB, the SINAD method asymptotes to a level governed by the residual distortion of the receiver while the SND/N_0 method (the modulation on-off method) should asymptote at the maximum quieting level of the receiver. In Fig. 2, no asymptote for the SND/N_0 method has been reached.

Figure 2 also shows the difference between the two methods with and without the psophometric filter. The figure indicates that for this particular receiver, the sensitivity based on a SND/N_0 ratio of 20 dB, measured with psophometric filter, and the sensitivity based on a SINAD ratio of 12 dB, measured without psophometric filter, are equal.

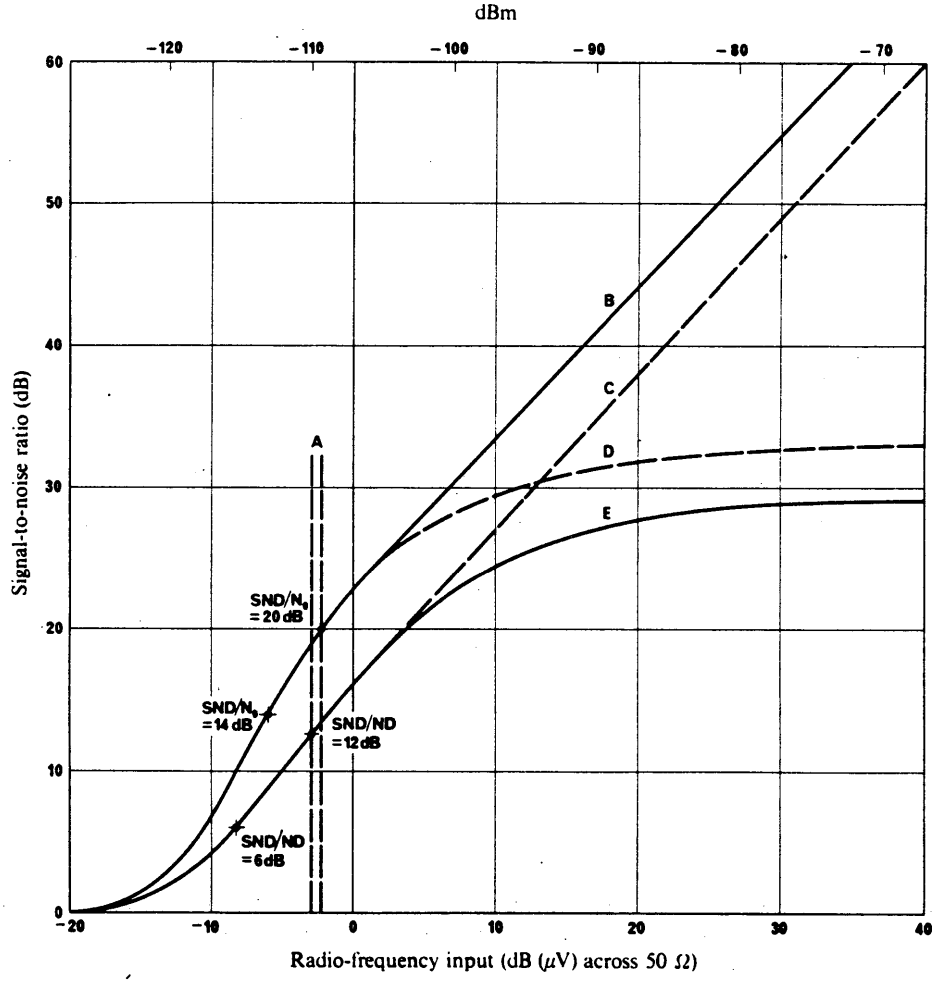


FIGURE 2 — *Audio-frequency signal-to-noise ratio as function of the receiver input voltage in a typical 160 MHz radio telephone set for 25 kHz channel spacing (measured by the SINAD and the SND/N_0 methods with a modulation frequency of 1000 Hz and a frequency deviation of 3 kHz).*

A: receiver reference sensitivity for the two criteria

B: SND/N_0 with psophometric filter

C: SND/N_0 without psophometric filter

D: SINAD with psophometric filter

E: SINAD without psophometric filter

Additional measurements made on several high quality 159 MHz receivers are described in [CCIR, 1974-78a]. The results are shown in Fig. 3. This figure shows considerable difference between the two methods of measurements. The SND/N_0 method predicts the sensitivity at a signal-to-noise ratio of 12 dB as 4 dB better than the SINAD method. The addition of a psophometric filter would still leave the results different.

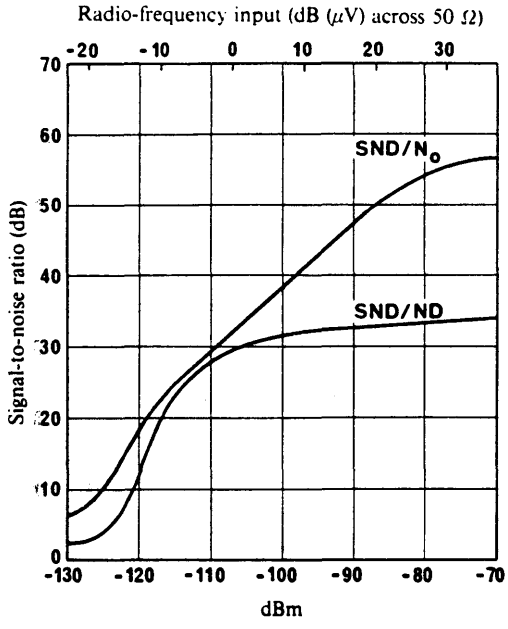


FIGURE 3 — *Audio-frequency signal-to-noise ratio as function of the receiver input voltage in a typical 159 MHz radio telephone set for 25 kHz channel spacing (measured by the SINAD and the SND/N_0 methods without a psophometric filter and with a modulation frequency of 1000 Hz and a frequency deviation of 3 kHz)*

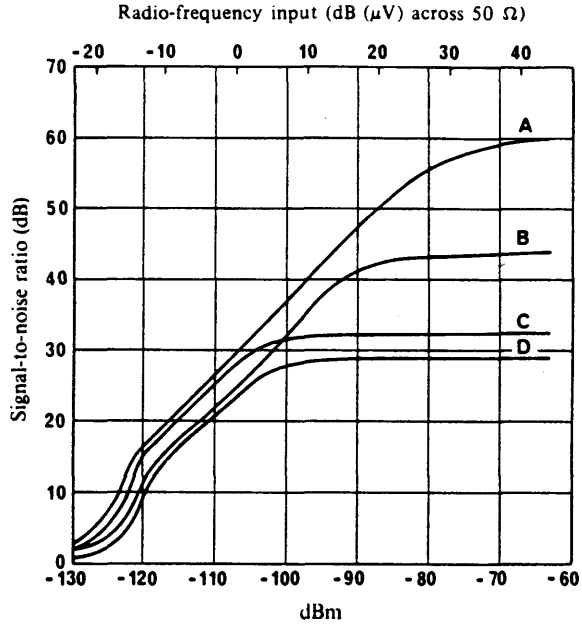


FIGURE 4 — *Audio-frequency signal-to-noise ratio as function of the receiver input voltage in a typical 400 MHz radio telephone set for 12.5 kHz channel spacing (measured by the SINAD and the SND/N_0 method with a modulation frequency of 1000 Hz and a frequency deviation of 1.5 kHz).*

A: SND/N_0 with psophometric filter

B: SND/N_0 without psophometric filter

C: SINAD with psophometric filter

D: SINAD without psophometric filter

Measurements on a typical 400 MHz radio telephone set designed for 12.5 kHz channel separation with the SINAD and the SND/N_0 method either with or without a weighting network have also been made in Japan [CCIR, 1974-78b]. The results are shown in Fig. 4.

For signal-to-noise ratios up to approximately 20 dB almost the same results were obtained by both methods. This is in accordance with the results of the Swiss investigations (Fig. 2). However, the difference between the measured sensitivity levels with or without a weighting network were somewhat smaller than shown in Fig. 2. This may be caused by different characteristics (e.g. IF-bandwidth; distortion characteristics) of the receivers measured.

In general, it can be concluded that for:

- low signal levels, the SND/N_0 method and the SINAD method either with or without the psophometric filter may or may not give the same results depending upon the distortion and noise characteristics of the receiver being tested;
- high signal levels, the two methods with or without filters give different results.

3. Summary

3.1 The two types of signal-to-noise ratios commonly used for measuring the sensitivity of receivers for class of emission F3E for use in the land and maritime mobile services are:

$$SND/ND \text{ (or SINAD)} = 12 \text{ dB}$$

measured in the presence of modulation with a test signal rejection filter, and

$$SND/N_0 = 20 \text{ dB}$$

measured with the psophometric weighting network according to CCITT Recommendation P.53, Volume V (Yellow Book). This last network takes into account the hearing mean sensitivity curve and the average characteristics of a subscriber telephone set. In both cases, the measurement of sensitivity needs to be made with the actual baseband filter, if any, used in the receiver.

3.2 Tests have shown that the results of sensitivity measurements made by using the 12 dB SND/ND criterion may or may not agree with those obtained by using the 20 dB SND/N_0 criterion with the psophometric filter.

3.3 It is noted that IEC were required to recommend only one method for establishing the reference sensitivity and that they have adopted the method using the 12 dB SND/ND criterion in conjunction with a low-pass filter having a cut-off frequency of approximately three times the maximum modulating frequency.

3.4 Although the adoption of a single method facilitates the comparison of different receivers, it is noted that for certain applications it may be desirable to make the measurements with the aid of a suitable weighting network. If, for example, the receiver is intended to be connected to the public telephone network, it seems appropriate to make the measurement by using the 20 dB, SND/N_0 criterion with the psophometric filter mentioned in § 3.1 above.

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[1970-74]: 1/56 (Switzerland).

[1974-78]: **a.** 1/44 (USA); **b.** 1/142 (Japan).

REPORT 521-2

RECEIVER RF NON-LINEAR MODELLING

(Question 44/1)

(1974-1978-1986)

1. General

This Report describes the basic analysis required to model the receiver radio-frequency non-linear effects of cross-modulation, intermodulation desensitization, etc. Such analysis is frequently required to investigate spectrum utilization problems. Validation of the methods shows that good accuracy can be obtained from the analysis when an adequate description of the non-linearities is obtained.

2. Non-linear canonical models of receivers

A communication (or radar, telemetry, etc.) receiver may be subdivided into a number of stages such as amplifiers, frequency convertors, demodulators, etc. Some of these stages are designed to provide nominally linear signal processing operations, e.g. radio-frequency and intermediate-frequency amplifiers (linear filtering operations) and frequency convertors (frequency translation). Others are designed to provide certain desired non-linear operations, e.g. a frequency discriminator (which extracts the instantaneous frequency of the signal) and a hard limiter (which removes amplitude variations). The modelling of a complete communication receiver can be carried out by modelling both nominally linear and non-linear functional blocks [Graham and Ehrman, 1972 and 1973; Bello, 1972; Ehrman and Ahmed Meer, 1972; O'Donnell, 1972].

Canonical models can be developed for a non-linear receiving system where the structure of the models are independent of the system, but where certain parameters in the models are dependent upon the non-linear system. With these canonical models it is possible to characterize such familiar non-linear receiver effects as desensitization, cross-modulation, intermodulation distortion and spurious response for general classes of interfering signals (e.g. narrow-band, wide-band, multiple signal).

Each stage of a receiver may be represented by a circuit, with each component represented by a (generally) non-linear circuit model. Determination of the response of circuits of this type involves the solution of simultaneous non-linear differential equations. Such detailed modelling can be very complex and quite unnecessary to describe transformation of input signals into output signals. The input-output behaviour can be characterized by canonical models whose complexity can be adjusted to include constraints on the input signal, such as bandwidth and power, and on the signal processing, such as finite memory and bandwidth. The parameters of the canonical model can be determined either by measurement or by computer-aided circuit analysis.

The generality of non-linear receiving systems makes their characterization difficult. When applicable, the most tractable representation of input-output relations is by way of Volterra functionals [Volterra, 1930; Brilliant, 1958; George, 1959] and Fourier transforms of Volterra functionals. These are called, respectively, non-linear impulse responses and non-linear transfer functions. The canonical models are interpretations of expansions of non-linear impulse responses and transfer functions. (These may be regarded as generalizations to non-linear systems of the canonical models for linear time-varying systems derived by Kailath [1959] and Bello [1963].)

The derivation [Graham and Ehrman, 1973; Bello, 1972] of the canonical models involves three basic steps. The first step is the representation of a receiver stage in terms of a parallel combination of (simpler) elementary blocks arranged in order of increasing non-linearity (e.g. the Volterra functional expansion). The second step replaces each of the ordered blocks by a set of equivalent functional blocks whose inputs are the low-pass complex representations of the multiple-input band-limited signals. The third step involves the representation of the multiple-input functional blocks.

Two methods for representing the functional blocks are basic, although others are possible. The first is the tapped delay line model. The second is the frequency power series, useful when the frequency dependencies of the non-linear system transfer functions change slowly over the bandwidth of the individual input signals.

The Volterra functional series converges rapidly for a small input signal level. There are cases where strong signals and more than one input port must be considered. For example, in the modelling of a second input port, the local oscillator input must be considered. To handle such cases, the Volterra expansion may be generalized to multiple input ports where some, but not all, of the input ports have "strong" excitations. From this generalized expansion non-linear canonical models follow by analogy as in the single-input port case.

3. Summary of Volterra series relationships

In this section, important input-output relationships for Volterra functional expansions are summarized.

When both input $x(t)$ and output $y(t)$ are represented as time functions, the expansion becomes:

$$y(t) = \sum_{n=1}^{\infty} \int\int \ldots \int h_n(\tau_1, \tau_2, \ldots \tau_n)\, x(t-\tau_1)\, x(t-\tau_2) \ldots x(t-\tau_n)\, d\tau_1 \ldots d\tau_n \quad (1)$$

where $h_n(\tau_1, \tau_2, \ldots \tau_n)$ is called the non-linear impulse response of n'th order. The Fourier transform of the time-domain kernel is called the non-linear transfer function of n'th order. This latter transfer function is given by:

$$H_n(f_1, f_2, \ldots f_n) = \int \ldots \int h_n(\tau_1, \ldots \tau_n)\, e^{-j2\pi(f_1\tau_1 + \ldots f_n\tau_n)}\, d\tau_1 \ldots d\tau_n \quad (2)$$

If $h_n(\tau_1, \ldots \tau_n)$ is expressed as the Fourier transform of $H_n(f_1, \ldots f_n)$ in (1) it is quickly found that:

$$y(t) = \sum_n \int\int \ldots \int H_n(f_1, \ldots f_n)\, X(f_1)\, X(f_2) \ldots X(f_n)\, e^{j2\pi(f_1 + \ldots f_n)t}\, df_1\, df_2 \ldots df_n \quad (3)$$

which relates the output time function to the input spectrum $X(f)$.

Fourier transforming both sides of equation (3) yields the input-spectrum to output-spectrum relationship:

$$Y(f) = \sum_n \int\int \ldots \int H_n(f_1, \ldots f_n)\, X(f_1) \ldots X(f_n)\, \delta(f - f_1 - f_2 \ldots - f_n)\, df_1\, df_2 \ldots df_n \quad (4)$$

where $Y(f)$ is the spectrum of $y(t)$ and $\delta(\cdot)$ is the unit impulse function.

It is readily demonstrated that the system output can be characterized in terms of frequency-dependent n'th-order transfer functions. Non-linear transfer functions determine the small-signal output of a receiver at frequency $f_1 + f_2 + \ldots + f_n$, if the input consists of unit exponential signals applied simultaneously at frequencies $f_1, f_2, \ldots f_n$. Receiver response to arbitrary inputs can then be expressed in terms of the non-linear transfer function by utilizing the frequency-domain form of the Volterra series.

4. An elementary non-linear canonical model

The simplest and frequently most useful non-linear canonical model representation of a receiving system is the leading term of the frequency-power series model. Tabulated in Table I are significant linear and non-linear output responses for a system whose input consists of a desired and two relatively larger undesired signals. Output responses listed are components of the output complex envelope associated with the frequency noted in the second column.

The frequency-dependence of the response is contained implicitly in the non-linear transfer function, while the dependence upon the input signal amplitudes is contained explicitly in the multiplicative complex-envelope factor. For example, the output voltage response across the circuit terminating load impedance at frequency f_1 is:

$$v_0(t) = \mathrm{Re}\left[\overline{V}_0(t)\, e^{\,j2\pi f_1 t}\right] \tag{5}$$

where the complex envelope is:

$$\left.\begin{aligned} V_0(t) = H_1(f_1)\, S_1(t) + \frac{3}{2}\, H_3(f_1, f_1, -f_1)\, S_1^2(t)\, S_1^*(t) + \\ + \frac{3}{2}\, H_3(f_1, f_2, -f_2)\, I_2(t)\, I_2^*(t)\, S_1(t) \end{aligned}\right\} \tag{6}$$

The undesired signal at f_3 is absent. Third-order effects contributing to gain-compression, cross-modulation and desensitization are included in $V_0(t)$. Amplitude-modulation distortion introduced by the non-linear terms can be investigated by studying the magnitude of V_0. Frequency-modulation distortion can be studied by investigating the derivative of the phase of V_0.

TABLE I – *Several signal components for a mildly non-linear system having narrow-band inputs*

Signal component	Frequency	Complex envelope
Inputs:		
Desired	f_1	$S_1 = S_1(t)$
Undesired	f_2	$I_2 = I_2(t)$
Undesired	f_3	$I_3 = I_3(t)$
Outputs:		
First-Order		
Desired	f_1	$H_1(f_1)S_1$
Adjacent or co-channel	f_2	$H_1(f_2)I_2$
Adjacent or co-channel	f_3	$H_1(f_3)I_3$
Second-Order		
Intermodulation	$f_2 + f_3$	$H_2(f_2, f_3)I_2I_3$
Intermodulation	$f_2 - f_3$	$H_2(f_2, -f_3)I_2I_3^*$
Third-Order		
Gain "compression" – desired signal	f_1	$3/4\ H_3(f_1, f_1, -f_1)S_1^2S_1^*$
Gain "compression" – undesired signal	f_2	$3/4\ H_3(f_2, f_2, -f_2)I_2^2I_2^*$
Gain "compression" – undesired signal	f_3	$3/4\ H_3(f_3, f_3, -f_3)I_3^2I_3^*$
Cross-modulation and desensitization	f_1	$3/2\ H_3(f_1, f_2, -f_2)I_2I_2^*S_1$
Cross-modulation and desensitization	f_1	$3/2\ H_3(f_1, f_3, -f_3)I_3I_3^*S_1$
Intermodulation	$2f_2 - f_3$	$3/4\ H_3(f_2, f_2, -f_3)I_2^2I_3^*$
Intermodulation	$2f_3 - f_2$	$3/4\ H_3(-f_2, f_3, f_3)I_3^*I_3$

The approach to receiver modelling involves an investigation of both the receiver and signal structure. The non-linear canonical model needed to model the receiver can then be selected. Transfer functions that parameterize the canonical model are then determined and the receiver output investigated. Transfer functions can be found by direct circuit analysis or by measurement.

5. Small-signal intermodulation distortion

In Fig. 1 are illustrated graphically, in a particularly useful form, equal-strength two-tone intermodulation data for the small-signal or well-behaved, region of a mildly excited non-linear system. On this graph, the first-order, or linear-response, has a slope of 1. The second-order intermodulation component has a slope of 2 while the third-order term has a slope of 3. Also shown are the second and third-order intercept points I_2 and I_3 as well as the output power ordinates O_1, O_2, and O_3 for 0 dBm input power. If the ordinates are known, the small-signal intermodulation distortion curves can be drawn immediately. A simple expression for P_{lim}, the upper limit of input power for which small-signal theory is applicable, is of considerable interest. Particularly simple expressions in terms of non-linear transfer functions are given in Table II for these critical points and limit. The small-signal limit is expressed in terms of single-tone input gain-compression. The expressions given in Table II are applicable to non-linear systems having 50 Ω sources and load impedances.

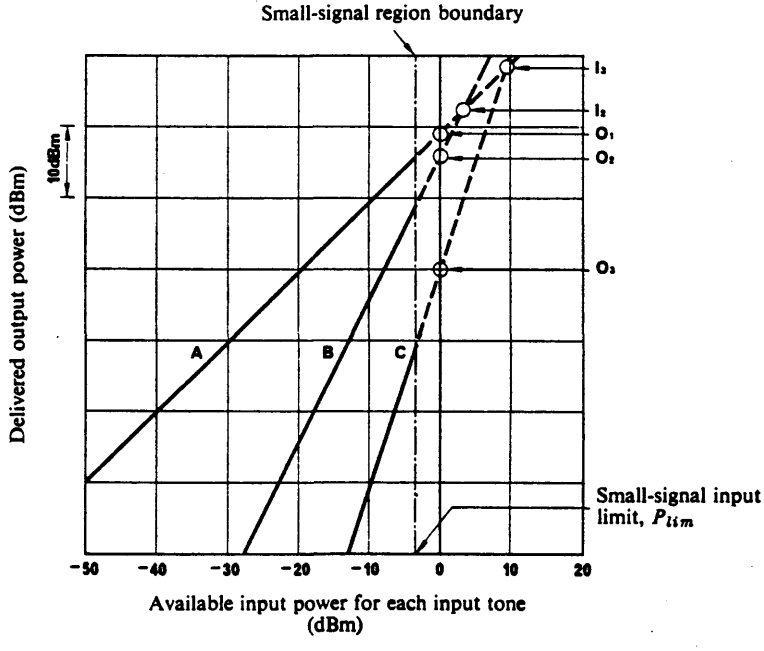


FIGURE 1

Small-signal amplitude responses for two-tone intermodulation distortion

A : first order
B : second order
C : third order

6. Experimental validation of non-linear transfer functions

In this section are presented theoretical predictions and experimental measurements from typical communications receiver circuits, including a variety of theoretical and experimental examples for test circuits and for a VHF receiver. The theoretical result in each case has been obtained by applying the simple expressions found in Tables I and II together with an analysis of a non-linear circuit model to obtain the required non-linear transfer functions [Graham and Ehrman, 1973; Ehrman and Ahmed Meer, 1972; Wiener, 1942; Narayanan, 1967 and 1969]. The extent of the agreement depends to a large extent upon the circuit model employed and upon the accuracy of the parameters associated with the models. Circuits have been modelled to such a degree that passive linear components have been assigned element values and all active and non-linear devices have been replaced by their incremental non-linear equivalent circuits, with parameters values dependent upon operating-point conditions in the physical circuit.

TABLE II – *Significant parameters of two-tone intermodulation non-linear responses for a 50 ohm system*

Zero dBm input ordinates First order: $O_1 = 20 \log_{10} \lvert H_1 \rvert + 6$ dBm Second order: $O_2 = 20 \log_{10} \lvert H_2 \rvert + 2$ dBm Third order: $O_3 = 20 \log_{10} \lvert H_3 \rvert + 4.4$ dBm
Intercepts Second order: $I_2 = 40 \log_{10} \lvert H_1 \rvert - 20 \log_{10} \lvert H_2 \rvert + 10$ dBm Third order: $2I_3 = 60 \log_{10} \lvert H_1 \rvert - 20 \log_{10} \lvert H_3 \rvert + 22.4$ dBm
Small-signal input limit (gain compression) $P_{lim} \cong -10 \log_{10} \left[\frac{1}{x} \text{ Re} \left\{ \frac{H_3}{H_1} \right\} \right] - 4.2$ dBm where x = gain compression in dB

7. Single-stage transistor amplifier

A single-stage untuned transistor amplifier provides the first circuit for comparison of measured and predicted performance. The measured and predicted small-signal insertion gain is shown as a function of frequency in Fig. 2. Agreement is seen to be excellent throughout the entire frequency range. In Fig. 3 a comparison is shown between measured and predicted two-tone intermodulation data for in-band inputs at 2.5 and 3.0 MHz. The theoretical results are in excellent agreement with measurements. Intercepts I_2 and I_3 are also predicted accurately by the simple expressions given in Table II. The input power level of -16.5 dBm has been indicated as the upper-limit for the small-signal region in which the behaviour is regular. The criteria used is 0.3 dB gain-compression for a single-tone input. A more complete comparison of measured and predicted gain-compression for single-tone input is shown in Fig. 4.

In addition to the input power dependence of intermodulation and gain-compression, it has been found that non-linear model prediction of the frequency dependence for both the second and third-order non-linear transfer functions is in excellent agreement with measurement. These results are presented in Figs. 5 and 6. Excellent agreement can be expected between measurement and prediction for carefully modelled circuits. The agreement will be good over a wide range of frequencies, with no degradation of prediction accuracy in regions well beyond the mid-band and cut-off frequencies.

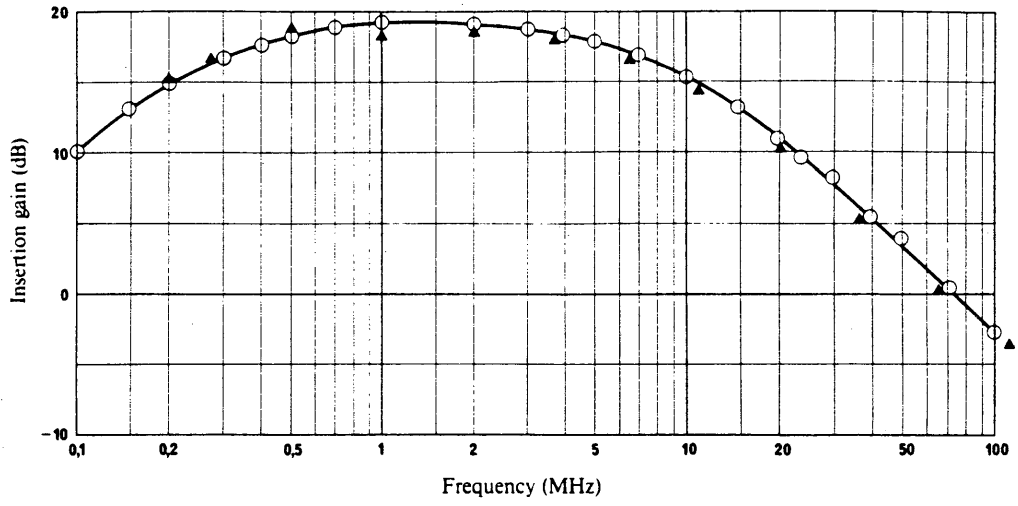


FIGURE 2 — *Comparison of 2N2950 common-emitter amplifier insertion gain*

—○—○— measured

▲ points calculated from the receiver model

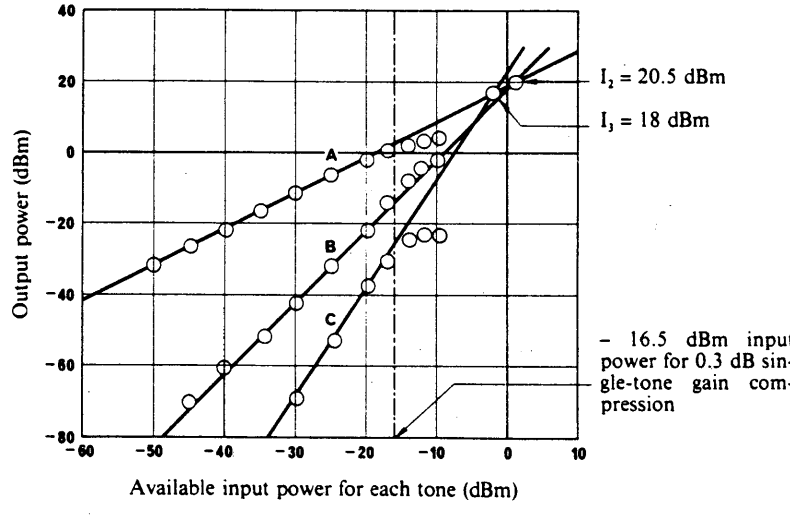


FIGURE 3 — *Two-tone test data for the 2N2950 test amplifier*

V_c = + 10V I_c = 10 mA ——— : theoretical curve ○ : measured points

Curve A: f_1 = 2.5 MHz, f_2 = 3.0 MHz B: $f_2 - f_1$ = 0.5 MHz

C: $2f_2 - f_1$ = 3.5 MHz

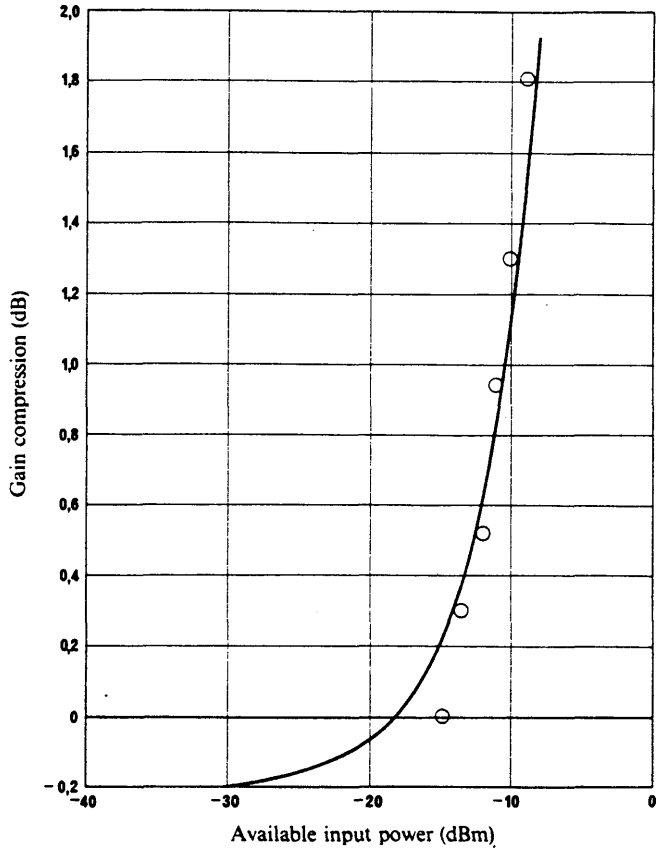


FIGURE 4 — *Gain-compression curve for the 2N2950 test amplifier at 3.0 MHz*

—— : theoretical curve ○ : measured points

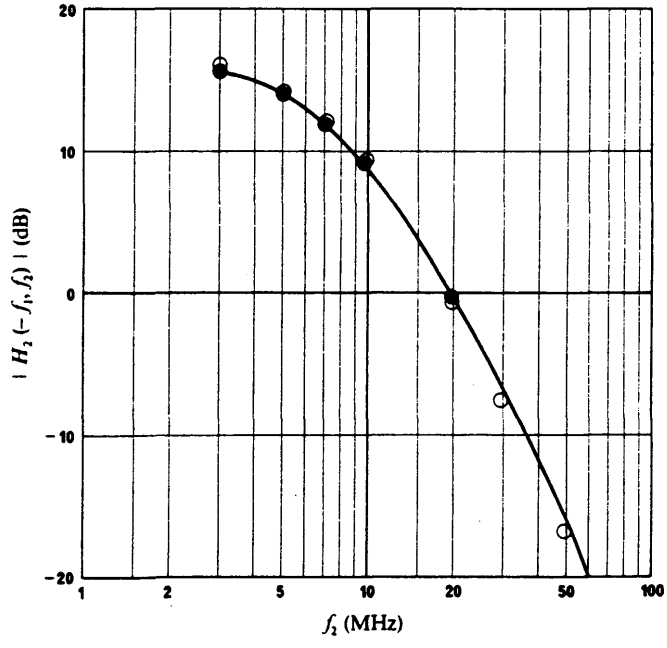


FIGURE 5 — *Second-order non-linear transfer function for the CE 2N2950 amplifier*

$f_1 = f_2 - 0.5$ MHz —— : theoretical curve ○ : measured points

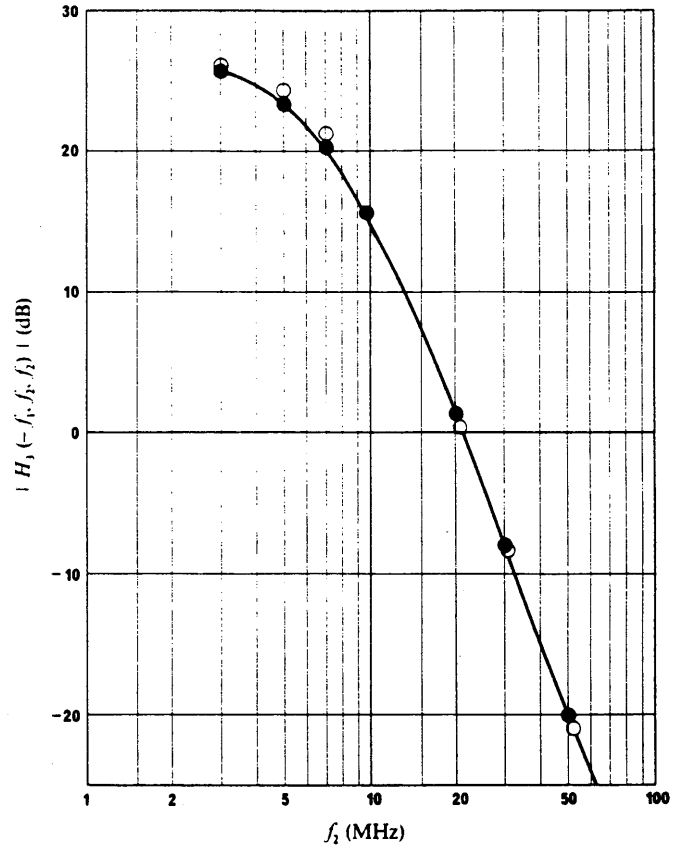


FIGURE 6 — *Third-order non-linear transfer function for the CE 2N2950 amplifier*

$f_1 = f_2 - 0.5$ MHz ——— : theoretical curve O : measured points

8. An example of the application of this method to a VHF communications receiver

A typical communication receiver capable of detecting either AM or FM modulation over the 30 to 100 MHz band was used in the example. The tuner is all solid-state and consists of two stages of RF amplification, one stage of mixing, and one stage of IF amplification. There are five linear interstage networks in the signal path from input to output.

The nominal modelling procedure employs a Non-linear Circuit Analysis Program (NCAP) [Graham and Ehrman, 1973]. Using NCAP, the RF tuner is modelled by entering into the computer program the individual component values of the tuner as specified by its circuit schematic. For those components which are tunable, specific values are obtained by using the computer program to simulate the alignment procedure as suggested by the manufacturer. The transistors in the tuner are modelled using a non-linear T equivalent circuit. Where the non-linear parameters needed for the transistor models are not included in transistor specification sheets, nominal values are obtained experimentally by performing selected measurements on transistors obtained from laboratory stock [Graham and Ehrman, 1973]. The component values obtained by means of the schematic, the alignment procedure, and transistor measurements are used to describe the tuner.

The-second-order non-linear transfer function describes the mixing of the input signal with the local oscillator to produce a signal at the intermediate frequency. Since the local oscillator is 21.4 MHz higher than the tuned frequency, the local oscillator frequency is fixed at 66.4 MHz while the input frequency is swept from 30 MHz to 100 MHz. Assuming unit voltage excitation, the second-order non-linear transfer function yields the magnitude and phase of the IF response at the difference frequency between the local oscillator and signal frequencies.

Figure 7 shows the magnitude of the nominal second-order non-linear transfer function as computed and measured experimentally. Note the desired peak at the tuned frequency of 45 MHz and the secondary peak in the neighbourhood of 90 MHz. The secondary peak results from the image response of the tuner and occurs when the frequency of the input signal equals the sum of the IF frequency plus the local oscillator frequency. For the tuner under consideration this happens at 21.4 + 66.4 = 87.8 MHz. In addition, a null occurs when the input frequency equals the local oscillator frequency of 66.4 MHz resulting in a zero beat.

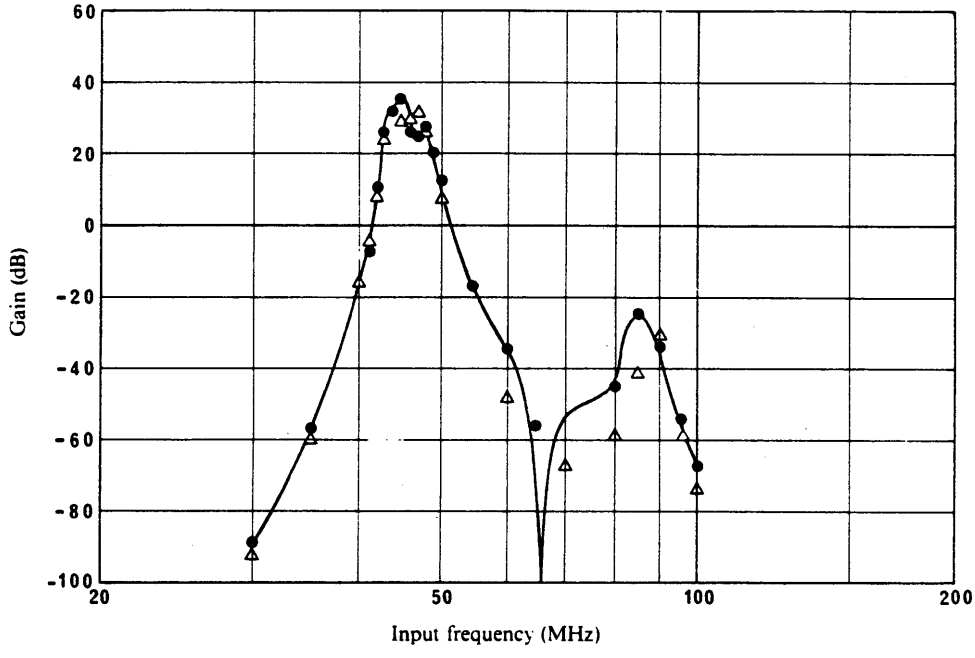


FIGURE 7 — *Nominal vs. measured 2nd order transfer function*

△ Measured

● Predicted (schematic)

The third-order non-linear transfer function is computed using a two-tone input. The two tones, at frequencies f_1 and f_2, are selected such that $f_1 \pm f_2$ is constrained to equal 45 MHz. The magnitude of the transfer function yields the magnitude of the IF response at the constant frequency $f_{LO} - (f_1 \pm f_2) = 21.4$ MHz. This is plotted in Fig. 8 as a function of f_1. The transfer function peaks in the 80 to 100 MHz range because this corresponds to f_2 varying about the tuned frequency of 45 MHz with values ranging from 35 to 55 MHz.

Because of the differences in components and active device parameters, each tuner that is built has its own unique set of non-linear transfer functions. Of interest, therefore, is the spread in performance that might be expected from unit to unit as values are varied from the nominal.

One item of concern is the effect of transistor non-linear parameter variability on the non-linear transfer functions. This is easily examined using the computer program. For example, the results on the second-order non-linear transfer function of varying by ± 10% the parameters that characterize the non-linear base-to-emitter capacitance are shown in Fig. 9. Note that the maximum difference is less than 3 dB.

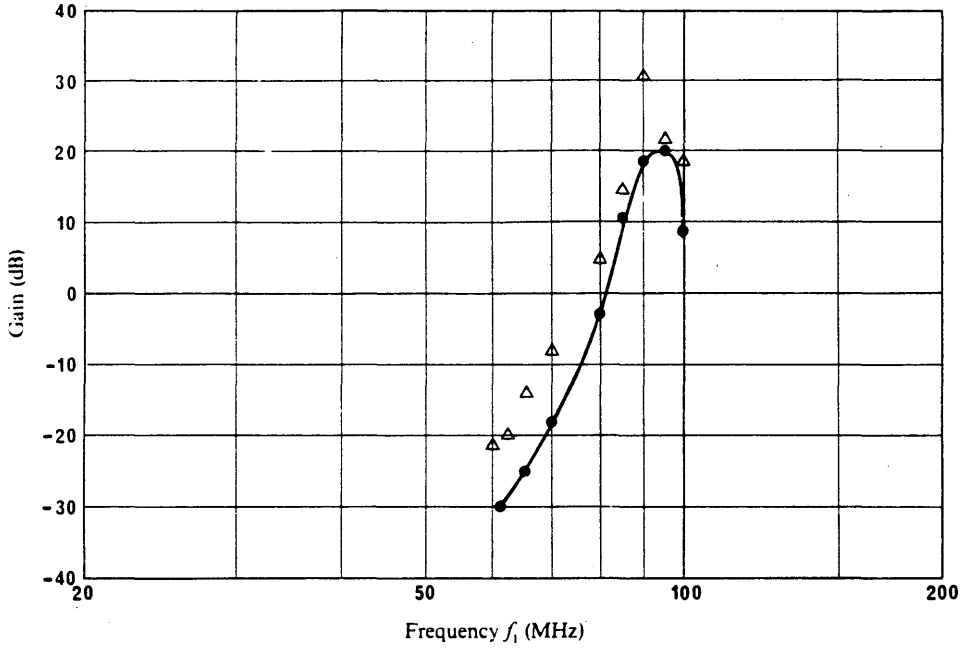


FIGURE 8 — *Nominal vs. measured 3rd order transfer function*

● Measured
△ Predicted (schematic)

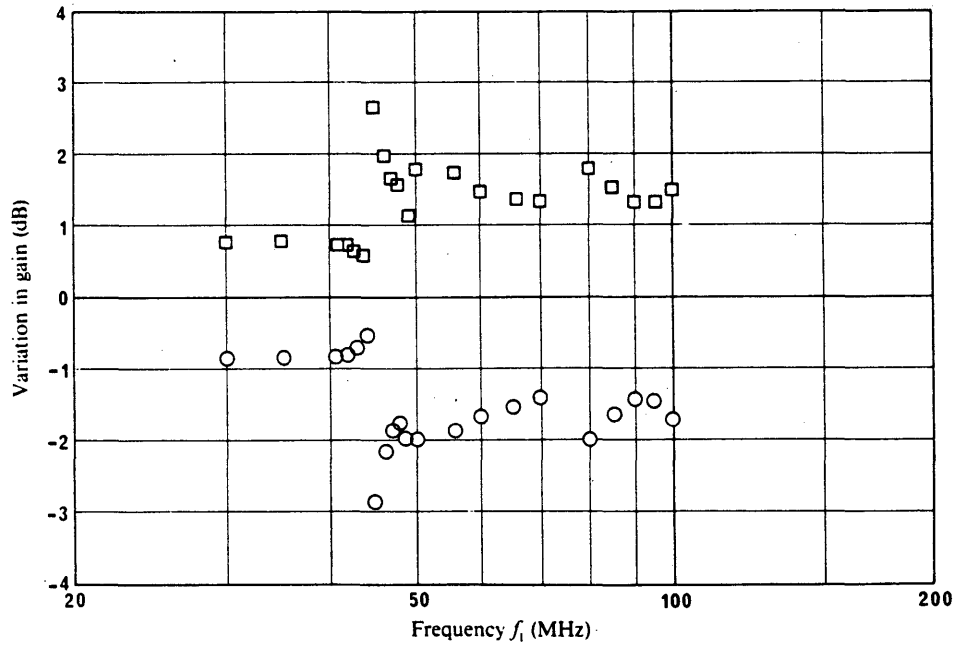


FIGURE 9 — *Variation of 2nd order transfer function for ± 10% change in B-E non-linear capacitance*

□ + 10 %
○ - 10 %

9. Conclusions

This Report defines the analysis requirements for radio-frequency non-linear receiver modelling, and thus is a partial response to Question 44/1. This Report has shown that with an adequate description of the non-linear devices, interference effects can be accurately predicted and models for the investigation of spectrum utilization can be constructed.

The data required for the analysis can be obtained from:

- the receiver's instruction manual, parts list, and circuit schematic, and
- transistor or other amplification device characteristics obtained from either specification sheets or measurements on off-the-shelf devices.

The analysis procedure can also be used to investigate interference effects due to variations in parameters and parasitic elements.

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REPORT 840-1 *

DEGRADATION OF MAXIMUM USABLE SENSITIVITY OF FM AND AM RECEIVERS OPERATING IN NON-LINEAR CONDITIONS

(Question 44/1)

(1982-1986)

1. Sensitivity of receivers

1.1 Recommendation 331 defines the maximum usable sensitivity for various types of receivers operating in the linear mode.

1.2 Under actual operating conditions, the maximum usable sensitivity may be degraded due to non-linear or other effects (e.g. reciprocal mixing) in the receiver input and frequency conversion stages, when an unwanted signal is present at a frequency in the vicinity of the wanted carrier and outside the receiver pass band. This phenomenon is similar to that covered by the concept of adjacent signal selectivity defined in § 6.3 of Recommendation 332.

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* This Report should be brought to the attention of Study Groups 8 and 10.

1.3 The following results of some measurements carried out in Yugoslavia give typical values of the sensitivity degradation of a frequency-modulation receiver operating at a frequency of 160 MHz and with a maximum usable sensitivity of −113 dBm. The sensitivity degradation measurements are given as the receiver input level expressed in decibels above the maximum usable sensitivity level. The sensitivity degradation values represent the increase in the level of wanted signal required to give a $(S + N + D)/(N + D)$ ((signal + noise + distortion)/(noise + distortion)) ratio of 12 dB in the presence of the unwanted signal levels for frequency spacings between the wanted and unwanted signals of 1 MHz and 5 MHz.

Figure 1 illustrates the characteristics of the degradation (D) of the maximum usable sensitivity as a function of the level (U) of the unwanted signal.

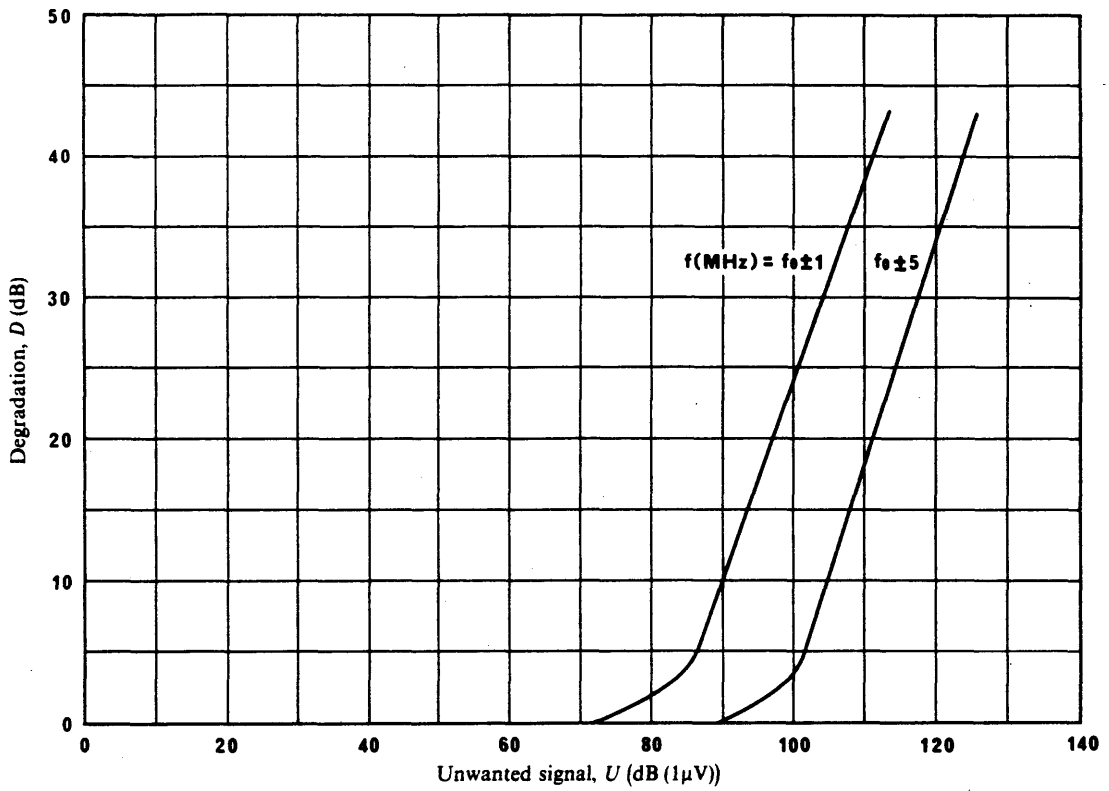


FIGURE 1 — *Typical degradation characteristics*

2. Intermodulation characteristics of frequency-modulation receivers

2.1 A method which can be used to obtain the intermodulation characteristics is described below. The method is based on measurements of the degree of non-linearity of the receiver transfer characteristics. In order to obtain the receiver transfer characteristics, the curves of degradation of maximum usable sensitivity of the receiver are used [Stevančevič, 1980].

2.2 Measurements were carried out in Yugoslavia to determine the degree of non-linearity of a receiver transfer characteristic and its influence on the appearance of the third order products in the form:

$$f_0 = 2f_1 - f_2 \qquad (1)$$

2.3 Generally, intermodulation products are not developed until the working point of the receiver is in the non-linear region of the transfer characteristic. For an increasing input level $U_{(x)}$, the transition from the linear domain into the non-linear domain is continuous but, to a first approximation, point A (Fig. 2) determines the upper limit of the linear domain. Non-linear dependence begins at low levels of the input signals and becomes more pronounced with increased input signals. When transfer characteristic non-linearity is clearly evident, increasing the level of the input signals by n dB causes the magnitude of intermodulation products, $U_{(y)}$, to increase by $K \cdot n$ dB.

2.4 For example, for third order intermodulation products this is given by the equation:

$$\Delta U_{(y)}\ (\text{dB}) = 3 \cdot \Delta U_{(x)} \qquad (\text{dB}) \tag{2}$$

When the non-linearity of the transfer characteristic is small, this equation is not applicable. Therefore, the purpose of the investigation in the domain of small non-linearities of transfer characteristics was to determine the point at which the third order intermodulation products having the form $(2f_1 - f_2)$ would be predominant (point B, Fig. 2).

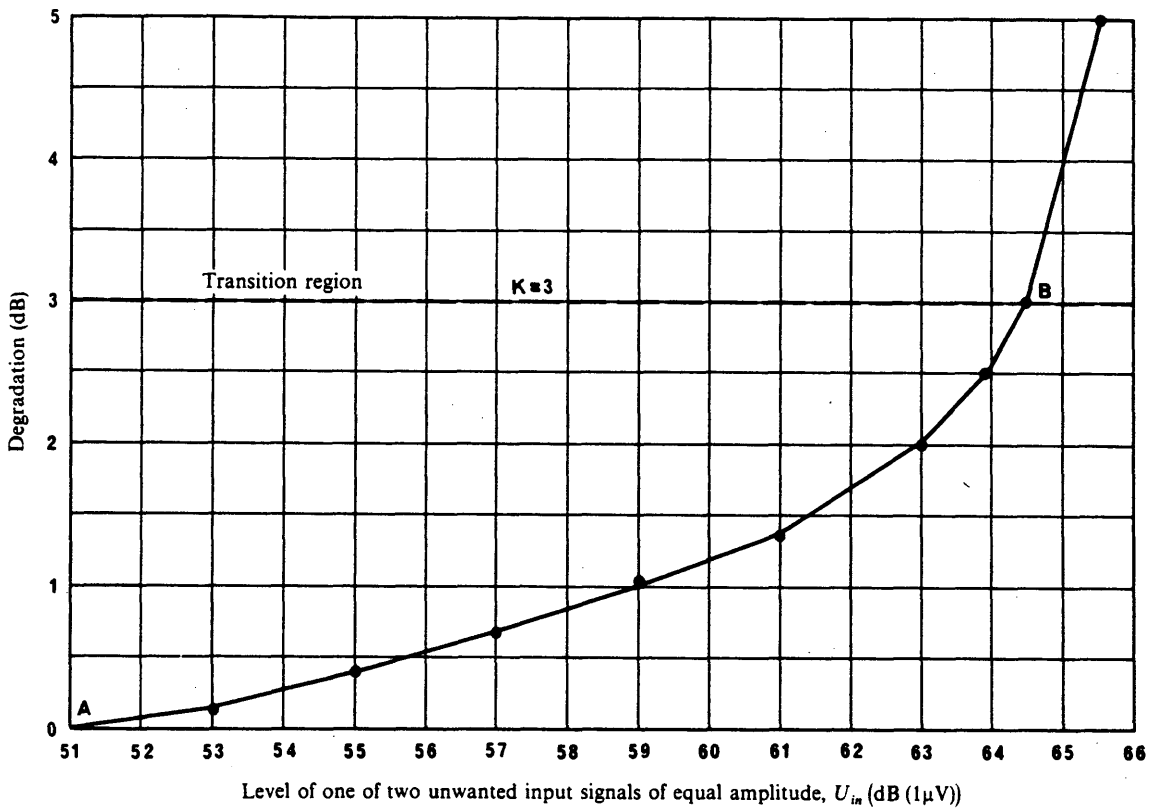


FIGURE 2 — *Degradation as a function of input voltage showing the transition region*

2.5 Analysis of the non-linear domain shows that it is possible to divide it into two sub-domains with a clearly expressed border at the point B where degradation equals 3 dB. In the sub-domain of small non-linearity between points A and B, the degradation is due to a number of effects, so equation (2) does not apply.

Above point B, the third order intermodulation product is predominant. This occurs when the degradation of maximum usable sensitivity is greater than 3 dB.

In cases of high degradation figures (> 18 dB) the fifth order term is predominant and leads to the relation $\Delta U_{(y)} = 5\Delta U_{(x)}$.

2.6 The general form of the transfer characteristic of a receiver, expressed with the aid of degradation curves of maximum usable sensitivity, is shown in Fig. 3.

This statement is valid for any FM receiver.

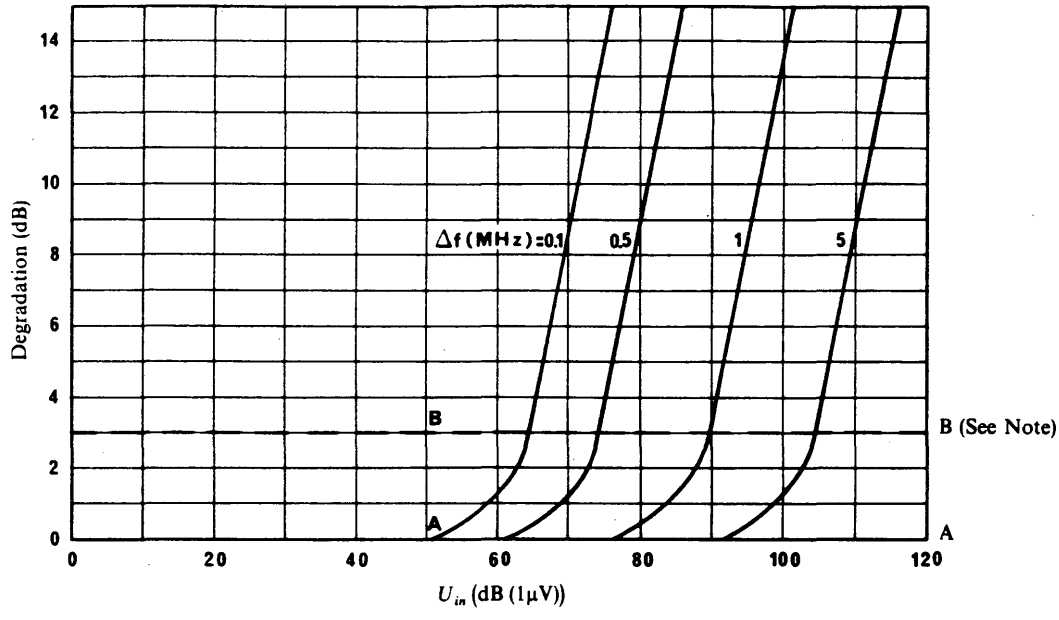


FIGURE 3 — *Degradation for various* Δf

Note. – The points of intersection with the curves correspond to the values of the third order intermodulation product level for $(S+N+D)/(N+D)$ ratio of 12 dB and a wanted signal input level 3 dB above the maximum usable sensitivity in accordance with the IEC Standard for frequency-modulation receivers for the mobile service (IEC Publication 489-3).

These curves indicate that the point of 3 dB degradation is the same for any Δf.

3. Cross-modulation characteristics of AM and FM receivers

3.1 Cross-modulation is caused by unwanted non-linearity effects in AM and FM receivers and therefore directly related to the degradation of maximum usable sensitivity under the condition that there is no reciprocal mixing.

Generally, the cross-modulation causes amplitude cross-modulation in AM receivers but it can also cause phase cross-modulation in FM receivers.

3.2 *Amplitude cross-modulation*

In order to obtain a quantitative criterion of cross-modulation in AM receivers, it should be assumed that two signals are present at the input of an AM receiver: a wanted non-modulated one and an unwanted modulated signal, (U_{un}), the modulation factor of which is m_{un}.

From the equation:

$$y(t) = a_1x(t) + a_2x^2(t) + a_3x^3(t) + \ldots + a_nx^n(t) \tag{3}$$

the cross-modulation factor can be expressed by the relation:

$$m_{CM} \approx 3\,\frac{a_3}{a_1}\,U_{un}^2\,m_{un} \tag{4}$$

if the terms of higher than the third order are negligible.

The conclusion from the relation (4) is that the cross-modulated factor does not depend upon the wanted signal level, however it is directly proportional to the modulation factor of the unwanted signal and to the square of the unwanted signal level.

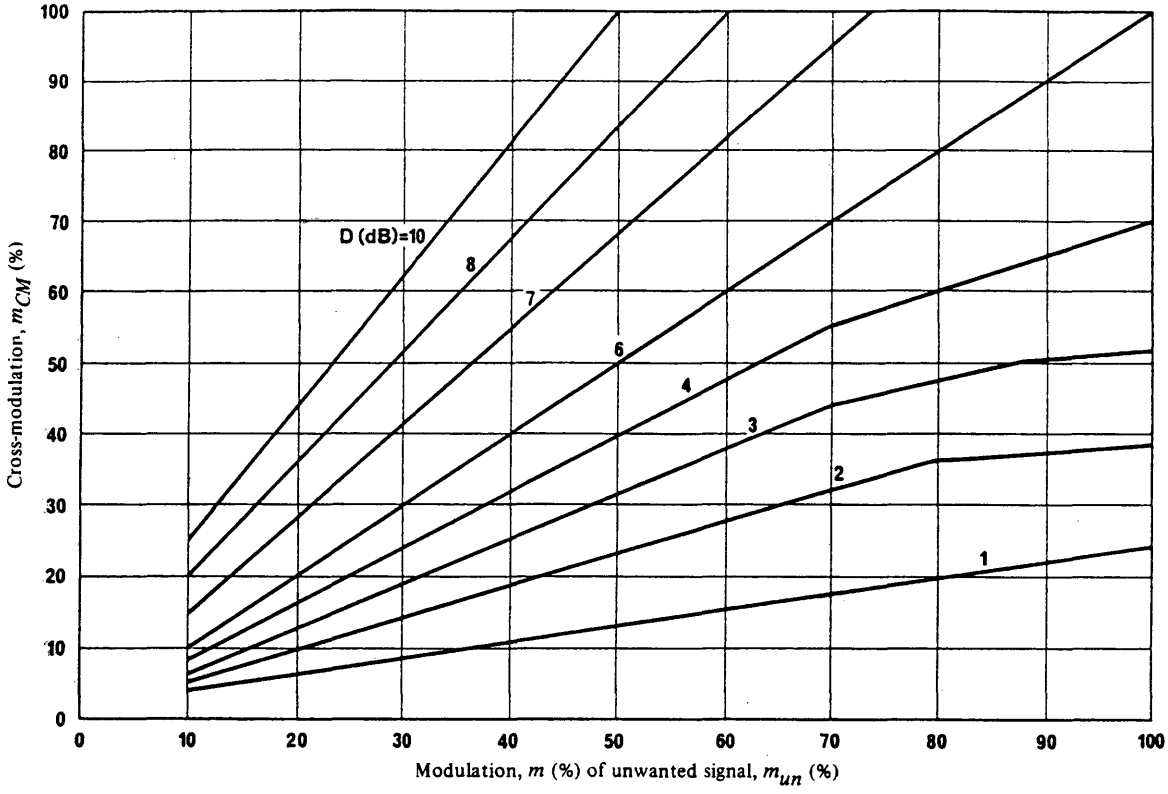


FIGURE 4 – *Cross-modulation, m_{CM} (%) as a function of degradation, D, and modulation levels, m_{un}*

The results of measurement from Fig. 4 show that a linear relation exists between the unwanted modulation factor and the cross-modulation factor (relation (4) is applicable) for values of degradation higher than 6 dB. In this case, the cross-modulation factor is always higher than the modulation factor of the unwanted signal.

The linear parts of the curves of Fig. 4 are generalized curves for all types of AM receivers.

3.3 *Phase cross-modulation*

Phase cross-modulation may be caused in FM receivers by AM emissions at different frequencies. For this case equation (3) and relation (4) are not applicable. Results of measurements are given in Fig. 5. From this figure it is clear that large degradations of the maximum usable receiver's sensitivity cause important cross-modulation frequency deviations.

4. Conclusions

4.1 The effect of a large interfering signal is indicated by the degradation of the maximum usable sensitivity of a receiver.

4.2 Large interfering signals cause intermodulation products which is indicated by the degradation of the maximum usable sensitivity. A 3 dB degradation of maximum usable sensitivity is considered tolerable. When both input signals levels are increased simultaneously, the magnitude of the third order intermodulation products will be proportional to the cube of the amplitude of the variation of the input signals.

The measuring results show that in cases of high degradation figures (> 18 dB) the fifth order term is predominant.

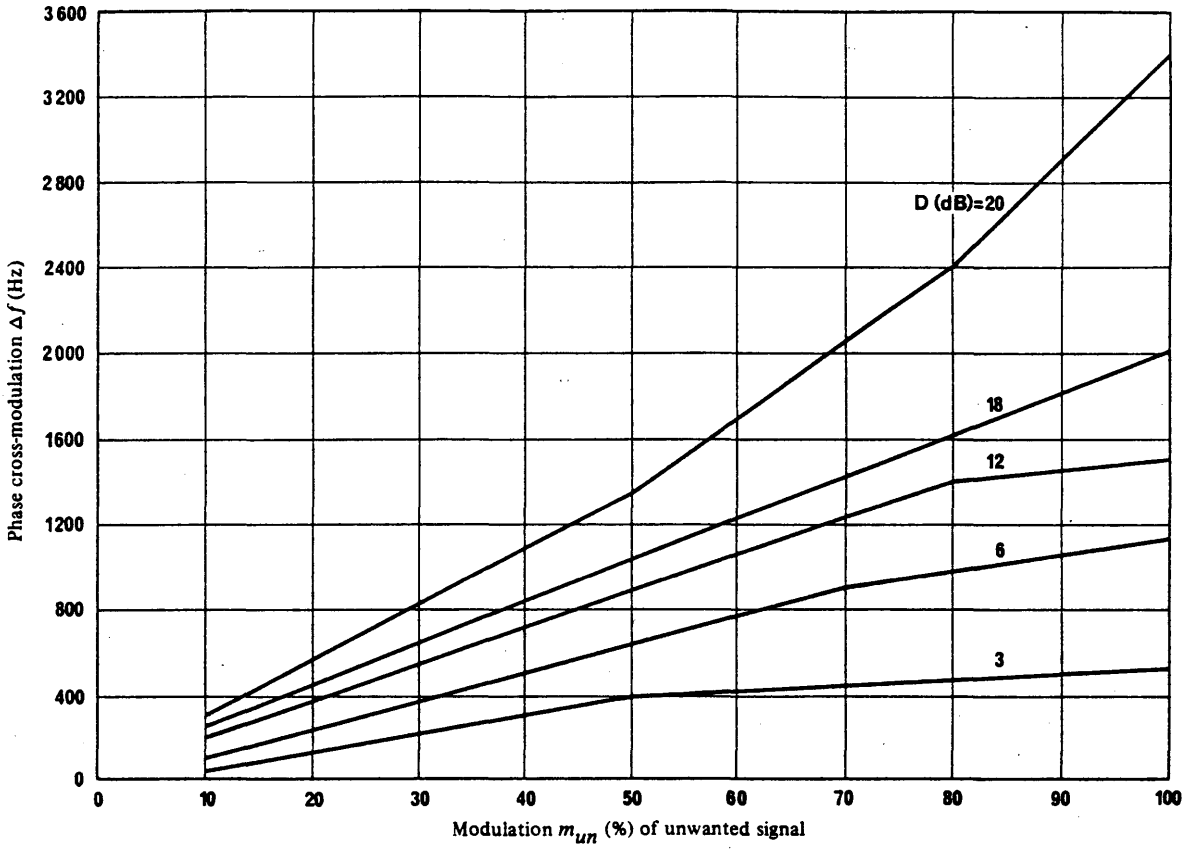


FIGURE 5 – *Phase cross-modulation Δf (Hz) as a function of degradation, D, and modulation levels, m_{un}*

4.3 The relation (4) for cross-modulation is applicable only to AM receivers if the unwanted signal level causes a degradation of 6 dB or more.

4.4 Phase cross-modulation is only important to FM receivers if the unwanted signal level causes a degradation of 18 dB or more.

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REPORT 522-2

A PROCEDURE FOR MODELLING RECEIVER INTERMODULATION CHARACTERISTICS

(Question 44/1)

(1974-1978-1990)

1. Summary

A method for calculating receiver intermodulation (RIM) characteristics, given suitable measured data, is described. The technique is illustrated with a specific example. A generic model, relating the model parameters to receiver circuit characteristics, is suggested. Typical frequency/distance separation criteria, to avoid the phenomena, are supplied. A measurement procedure and its application to equipment standards is also given.

2. Introduction

This Report describes an analytical procedure which can be used to model 2- or 3-signal receiver intermodulation (RIM) characteristics. The model is reported in the literature [Lustgarten, 1970], which briefly describes the Cosite Analytical Model (COSAM)*, of which the receiver interference model is a part and a description is available [Lustgarten and Hughes, 1972] of the results of a model validation exercise.

The receiver is modelled in a functional manner. A tuned circuit is assumed to be followed by a non-linear circuit which is, in turn followed by one or more tuned circuits and a second non-linear circuit.

The tuned circuit characteristics are assumed to be representable by Q (the circuit quality factor) and N, the number of tuned circuits. The use of Q permits consideration of the tuned frequency of the receiver, f_0. The non-linear circuit is assumed to be dependent on the input power level and is characterized by a conversion loss constant, $K_{m,n}$. (The sum of m and n is the order of the interaction; for 3rd order, $m = 2$, $n = 1$). That is, each receiver intermodulation order (e.g., 2nd, 3rd, 5th, etc.) require specification of two conversion loss factors, $K_{m,n}$ and $K'_{m,n}$.

An interfering power level breakpoint, P_{ib}, is used to determine which of the two $K_{m,n}$ values is to be applied. The prediction is statistical, in that the $K_{m,n}$ values are mean estimates with associated values of standard deviations.

The parameter which is predicted is termed P_{ino}; the effective input on-tune power level which would produce the same output as the receiver intermodulation interaction being considered. In COSAM, this value is summed with all of the contributions resulting from other interactions; a desired signal distribution is provided and an output $(S + I + N)/(I + N)$ output ratio calculated.

This Report describes a generic receiver model which appears to apply to a wide class of devices. A set of measured data on a single channel voice communications receiver is supplied and methods for calculating Q, N, $K_{m,n}$ and $K'_{m,n}$ are given.

Given the description of the intermodulation characteristics of a receiver, it may be determined whether it meets some specified standard, whether it will experience interference in a given environment and what measures can be taken to avoid harmful degradation.

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* See Report 524 (Dubrovnik, 1986).

Incidentally, the same general procedure applies to the transmitter intermodulation problem, which is also considered by COSAM.

New § 9 proposes a second version of a receiver model for the calculation of intermodulation interference which takes into account variations in susceptibility as a function of the level of the wanted signal at the input of a FM receiver.

3. Basic intermodulation phenomena

The basic intermodulation phenomena involve the mixing of two or more signals at different frequencies, in a non-linear circuit, which generate a signal at a different frequency.

In general:

$$f_0 = \pm mf_1 \pm nf_2 \pm pf_3 \pm \ldots \quad (1)$$

m,n and p are positive integers. f_0 is the frequency of the intermodulation product; the other f values represent the frequencies of the impinging signals. In this Report, only one form of intermodulation will be discussed, that due to the dominant effect, the 2-signal, 3rd order mix, given by:

$$f_0 = 2f_1 - f_2 \quad (2)$$

Equation (2) may be re-written:

$$|f_1 - f_2| = |f_0 - f_1| = \Delta f \quad (3)$$

The frequency relationships can be visualized as portrayed in Fig. 1.

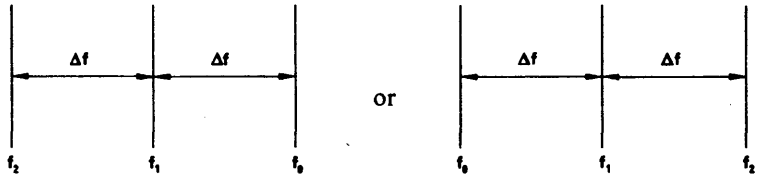


FIGURE 1

Frequency relationships for 2-signal, 3rd order, intermodulation

Given the above, the power relationships involved in the interaction are given by:

$$2P_1 + P_2 - K_{2,1} = P_{im} \quad (4)$$

P_1 and P_2 are input power levels, in dBm, of the signals at frequencies f_1 and f_2, respectively. P_{im} is the output power, in dBm, of the intermodulation product at the frequency f_0. $K_{2,1}$ is the 3rd order conversion loss factor.

4. The COSAM receiver intermodulation sub-model

The power levels P_1 and P_2 noted in equation (4) are assumed to occur at the input to the non-linear circuit. Equation (5), contains the power levels at the receiver input, as well as effective selectivity terms which provide off-frequency rejection estimates, due to passive stages which precede the non-linear circuits.

$$2(P_{i1} - \beta_1) + (P_{i2} - \beta_2) - K_{2,1} = P_{ino} \quad (5)$$

P_{i1} is the input interfering power level, in dBm, at frequency f_1; P_{i2} is the power level at f_2. β_1 and β_2 are off-frequency rejection parameter values, in dB, applying to frequency separation values Δf_1 and Δf_2, respectively. $\Delta f_1 = |f_1 - f_2|$ and $\Delta f_2 = |f_2 - f_0| = 2\Delta f_1$. P_{ino}, in dBm is the equivalent input, on-tune interfering power level (see Report 524, Dubrovnik, 1986).

The β values are calculated by using the standard procedure for computing tuned circuit selectivity characteristics. A graph of these generalized characteristics, [ITT, 1969], is given in Fig. 2. Note that β is a function of the ordinate parameter k, which is equal to $2Q\Delta f/f_0$.

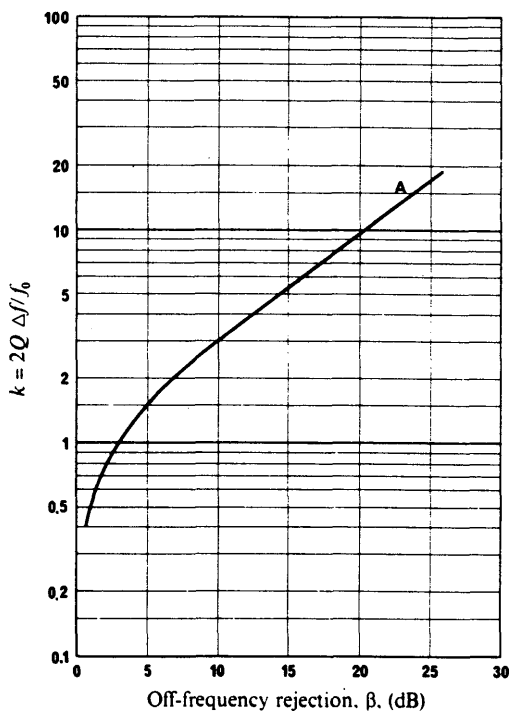


FIGURE 2 — *Tuned circuit selectivity characteristics [ITT, 1969]*

A: slope, 6 dB/octave

The slope of the straight portion of the curve is 6 dB/octave. This is characteristic of a single tuned stage, N, the number of tuned stages, is 1. If $N = 2$, the slope would be 12 dB/octave, etc.

Then, given Q, N, and f_0, one may calculate β for any Δf, using Fig. 2. Similarly, if we can estimate values of N, β_1 and β_2 from measured data and are given f_0, we can calculate Q. Both procedures are illustrated below.

Given the above parameter values and $K_{2,1}$, which can be estimated from measured data, we will have defined the 3rd order intermodulation characteristics of the receiver. As indicated above, more than one value of $K_{2,1}$ is generally necessary to describe the effects of several non-linearities, i.e., the first radio-frequency amplifier and mixer. For the example in this Report, only one value of $K_{2,1}$ is required.

5. Evaluation of measured data

Considerable measured receiver intermodulation data is given by [Efremov and Hagn, 1972]. Figure 3 is extracted from that reference. The points relating to the receiver labelled Manufacturer C are considered. Table I contains these measured values.

The value of P_i is the power level of both input signals, required to produce a standard response. The tuned frequency (f_0) was 42.0 MHz. The "sensitivity" was -123 dBm. Figure 4 is a plot of the data in Table I: a method is available which calculates the "best fit" of the data, together with an associated standard deviation (σ).

Since the Δf value of 0.02 MHz is "small", that is, within the intermediate-frequency bandpass of the receiver, we assume that $\beta_1 = \beta_2 = 0$. Substituting these values in equation (5):

$$3\ P_i - K_{2,1} = P_{ino} \tag{6}$$

Substituting the measured values $P_i = -53$ dBm and $P_{ino} = -123$ dBm, calculation gives $K_{2,1} = -36$.

The value of Q has now to be calculated. Since the slope of the straight portion of the curve in Fig. 4 is 6 dB/octave, it can be assumed that $N = 1$. Re-writing equation (5) as follows:

$$2\beta_1 + \beta_2 = 3\ P_i - K_{2,1} - P_{ino} \tag{7}$$

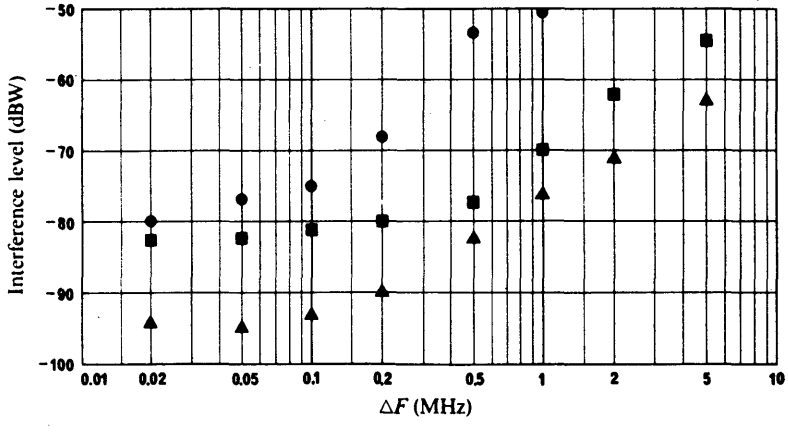


FIGURE 3 — *Measured two-signal third-order interference levels as a function of Δf for reference sensitivity input to a 42 MHz receiver*
[Efremov and Hagn, 1972]

Average desired signal = −153 dBW (Average reference sensitivity)

● Manufacturer A
▲ Manufacturer B
■ Manufacturer C

TABLE I

Measured 3rd order receiver intermodulation data

Δf (MHz)	P_i (dBm)
0·02	−53
0·05	−53
0·1	−52
0·2	−51
0·5	−46
1·0	−40
2·0	−32
5·0	−24

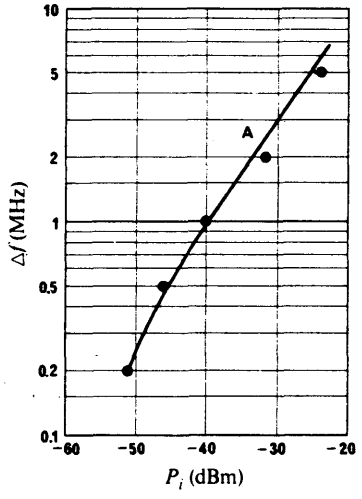


FIGURE 4 — *Plot of 2-signal, 3rd order, receiver intermodulation data*

A: slope, 6 dB/octave

Values of 1 and 2 MHz are selected for Δf_1 and Δf_2 respectively, which are on the straight portion of the curve. This implies that:

$$\beta_2 = \beta_1 + 6 \tag{8}$$

At Δf_1, $P_i = -39$ dBm, the values of $K_{2,1}$ and P_{ino} are given above. We then solve for β_1 and β_2, which are 12 and 18 dB, respectively. Using Fig. 2, we note that, for $\beta_1 = 12$ dB, $k = 2Q\Delta f/f_0 = 4$. Since $\Delta f_1 = 1$ MHz, and $f_0 = 42$ MHz, $Q = 84$.

With the values of $K_{2,1}$, Q and N determined, the procedure will now be reversed. Values of P_i will be predicted and compared with the measured values (rather than the smooth curve), we shall then note the differences and indicate how well the model conforms with the measured data. (The measured value of P_{ino} will be used in the predictions.)

The prediction equation is yet another variation of equation 5, namely:

$$P_i = (2\beta_1 + \beta_2 + K_{2,1} + P_{ino})/3 \tag{9}$$

The β values are taken from Fig. 2. Table II compares the predicted and measured values.

TABLE II

Comparison of measured and predicted data

Δf (MHz)	P_i (dBm meas.)	P_i (dBm pred.)	Δ (dB) (1)
0·02	−53	−53	0
0·05	−53	−53	0
0·1	−52	−52	0
0·2	−51	−50	−1
0·5	−46	−44	−2
1·0	−40	−39	−1
2·0	−32	−33	+1
5·0	−24	−24	0

(1) $\Delta = P_i$ (meas.) − P_i (pred.)

As can be seen, the maximum difference between measured values and the predicted values is 2 dB. The average difference is less than 1 dB. The standard deviation of the differences is approximately 1 dB. Measurement error is at least ± 1 dB. Consequently, it can be concluded that the model fits the data very closely.

6. Generic model

The preceding example is, admittedly, a special case. A more comprehensive illustration of receiver intermodulation characteristics is given in Fig. 5. This graph might be thought of as a generic receiver intermodulation model, since it covers most known situations and provides a physically based hypothesis for each portion of the curve as well as a model for the overall phenomena.

Briefly, at lower input power levels and small frequency separations, the mixer is said to be the dominant circuit and is preceded by several tuned stages. At some input power breakpoint, P_{ib}, the effects of the non-linearity of the first radio-frequency amplifier become dominant, this amplifier being, of course, preceded by fewer stages. Other non-linear stages can be present but their effects, apparently are over-shadowed by those of the first amplifier and the mixer.

At large input power levels, the curve flattens due to the fact that a maximum value of attenuation of the first tuned stage is reached.

The numerical values are for illustrative purposes only.

7. Influence of receiver intermodulation characteristics in an equipment deployment

Ideally, intermodulation-free frequencies would be used in the field [Lustgarten, 1968]. If both interfering transmitters, using power levels of less than 50 dBm, in the 50 MHz region, are more than one mile from the receiver, it is unlikely that serious interference due to receiver-intermodulation would occur. However, if one of the transmitters is "near" the receiver it may be quite difficult to avoid the problem.

A brief illustration will be given, indicating the order of magnitude of the problem. Assume the following: the desired signal is −90 dBm; the acceptable interference threshold (for approximately 90% confidence) is −108 dBm (P_{ino}); the interfering, on-site power level is 0 dBm, corresponding to P_{i1} (at f_1) in equation (5); the interfering transmitter power (at f_2) is 50 dBm; a simple mean-value inter-site coupling loss ($\overline{C}$) model is given by:

$$\overline{C} = 90 + 40 \log d_{km} \quad \text{dB} \tag{10}$$

Equation (10) is an estimate of losses in the 50 MHz range, if relatively low, vertically polarized antennas are used. Then, $P_{i2} = 50 - \overline{C}$; β_1 and β_2 are taken from Fig. 2 ($Q = 84$, $N = 1$, $f_0 = 42$ MHz).

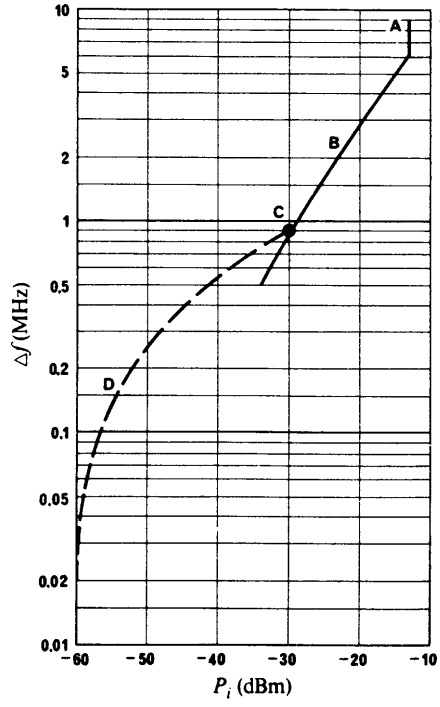


FIGURE 5 — *Generic receiver intermodulation model (numerical values are examples)*

Sensitivity: −120 dBm; f_0: 50 MHz

A: first tuned stage: maximum rejection (30 dB)

B: first radio-frequency amplifier, non-linearity: $K_{2,1} = -10$; $Q = 100$; $N = 1$

C: $P_{ib} = -30$ dBm

D: mixer, non-linearity: $K_{2,1} = -60$; $Q = 100$; $N = 3$

Using the above estimates, Fig. 6 illustrates the frequency/distance separation values required to avoid the interaction. It can be seen that distance separations as great as 50 km may not avoid the problem.

8. Application to equipment standards

Certain receiver intermodulation standards call for a single measurement, using equal input power levels at a "small" frequency separation. As noted, this measurement may test the ability of the mixer to discriminate against intermodulation interactions, but will not test the capability of the first radio-frequency amplifier.

In co-site situations, frequency separations are likely to be relatively large, but the presence of a transmitter, many miles away, can cause serious performance degradation if an intermodulation frequency combination is inadvertently employed.

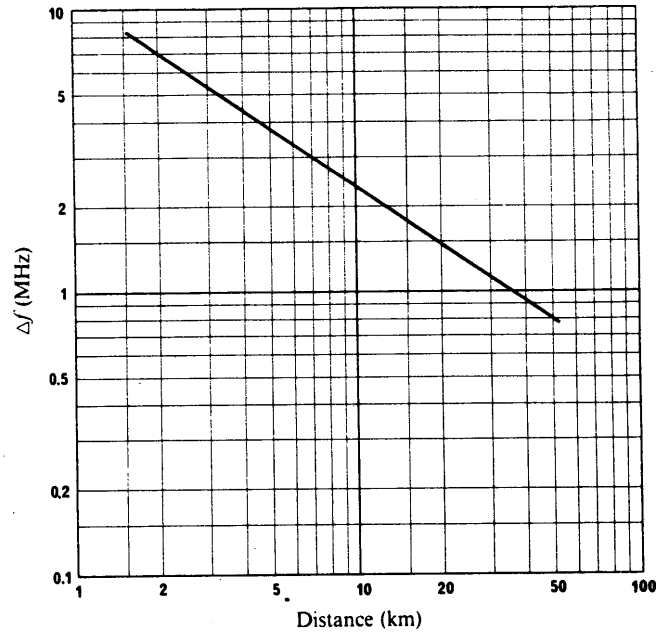


FIGURE 6 — *Sample frequency/distance separation criteria*

Co-sited transmitter (f_1, P_{i1} = 0 dBm); f_0 = 50 MHz

Consequently, it is essential that standard equipment measurement procedures and criteria include effects of large frequency separations, which represent a test of the ability of the input circuit and the first radio-frequency amplifier to discriminate against receiver intermodulation.

A brief outline of the suggested measurement procedure is given below.

The receiver is tuned to a standard test frequency. The frequency f_1 is unmodulated; the frequency f_2 is modulated 30%, with a standard tone, for amplitude modulation and unmodulated for frequency modulation.

Both signals are adjusted for maximum equal output; at least +20 dBm is desirable. The maximum Δf value is determined at which a standard output response is achieved. A 10 dB output $(S + N)/N$ is the value used for modulated signals. If unmodulated frequency modulation signals are used, the standard response is 20 dB of quieting, measured with a true r.m.s. voltmeter.

Given Δf_{max}, the next separation is $(\Delta f_{max})/2$. The equal input power level required to provide a standard response is noted. The process is continued until f_1 is within the measured 60 dB receiver bandwidth. Care is taken to avoid spurious responses.

9. Third-order intermodulation model with two interfering signals

It is known [White, D., 1973] that a receiver is exposed to inadmissible interference when the following condition is met:

$$2P_1 + P_2 \geq 3W,$$

where W is the susceptibility of the receiver to intermodulation interference (dBm).

Theoretical studies [Alter, L.S., 1986] have shown (and this has been verified by experimental work) that the magnitude W is a function of the power level of the wanted signal at the input and of the frequency differences of the interfering signls, as follows:

$$W = \begin{cases} b_1 \ (1 + P_s/P_{so}) & \text{where: } \Delta_1 \leq \Delta_{1.0} \\ b_2 \ (1 + P_s/P_{so})(\Delta_1/f_o) & \text{where: } \Delta_1 > \Delta_{1.0} \end{cases}$$

$$\Delta_1 = |f_1 - f_o|;$$

where:

P_s: power of the wanted signal at receiver input (mW);

b_1, b_2, Δ_1, P_{so}: parameters of the model;

W: susceptibility of the receiver to intermodulation interference (mW).

The variation in receptivity as a function of P_s and Δ_1 (Δ_1: frequency difference of the interfering signal whose frequency is closest to f_o) is shown in Figures 7 and 8.

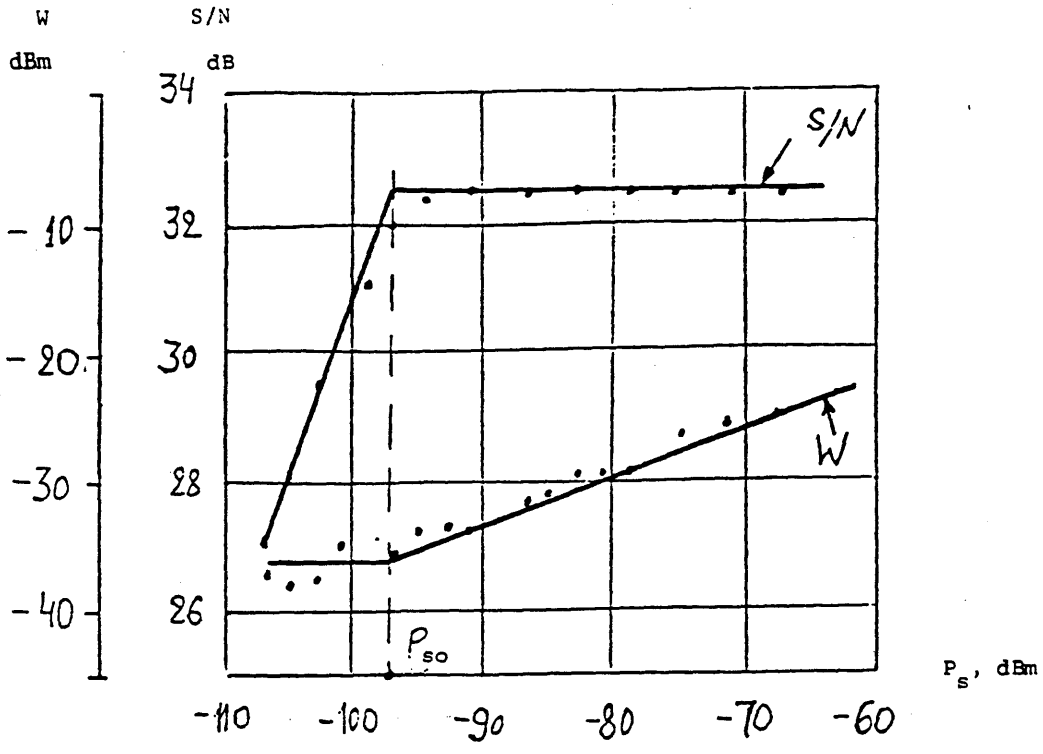


FIGURE 7

Dependence of W and S/N on P_s

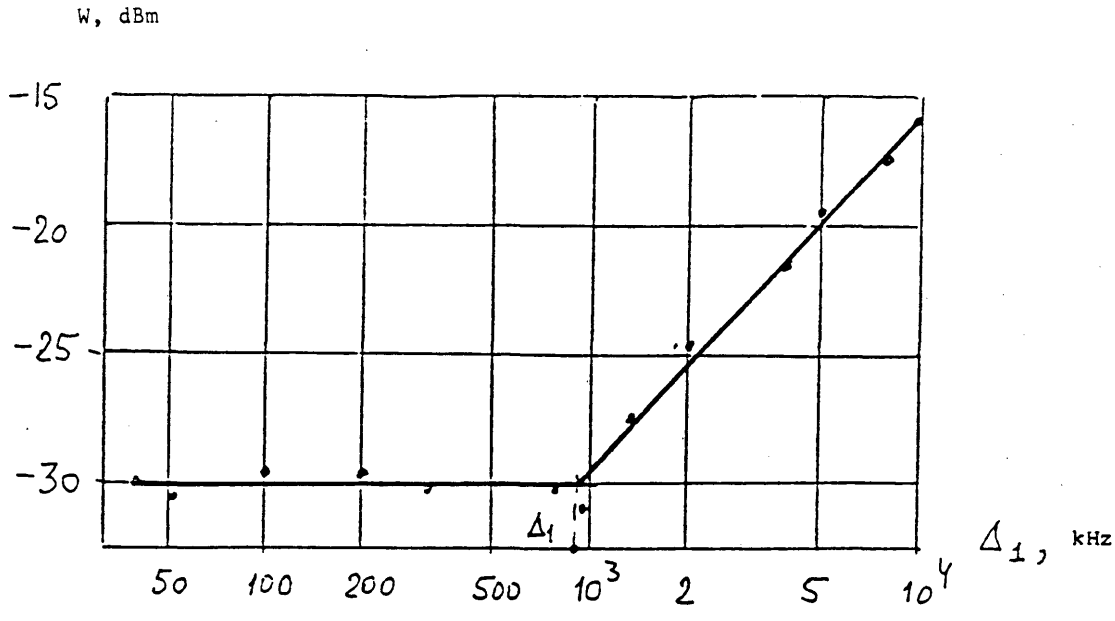


FIGURE 8

Dependence of W on Δ_1

The model has been verified with single channel receivers of various types in the land mobile service in the 40, 160 and 330 MHz bands and the results have shown good agreement with experimental data. The method for measuring the model parameters is described in the Appendix to this Report.

10. Concluding comments

The preceding material suggests that the complex receiver intermodulation phenomena can be adequately modelled by describing circuit performance with relatively few, simple parameters. Given adequate input-output measurements and a reasonable number of test samples, the method can be used to generate a statistical model of the phenomena, or form the basis of a procedure for deriving a meaningful electro-magnetic compatibility (EMC) figure of merit (FOM) for this important interaction. (A committee sponsored by the IEEE EMC group and ECAC has published a report which describes the FOM concept [IEEE/ECAC, 1975].

Recent efforts indicate that reasonable estimates of the RIM non-linear parameter values ($K_{m,n}$) can be made without measurements, given a detailed circuit diagram and knowledge of the basic characteristics of the active components in the receiver (see Report 521). It develops, however, that detailed knowledge of the passive components, particularly the off-tune characteristics of the input circuit, is equally important. In fact, this aspect seems to represent a more difficult problem, since necessary passive parameter values, e.g., radio-frequency transformer characteristics and stray capacitance values, are generally not available in circuit diagrams or manufacturer's literature.

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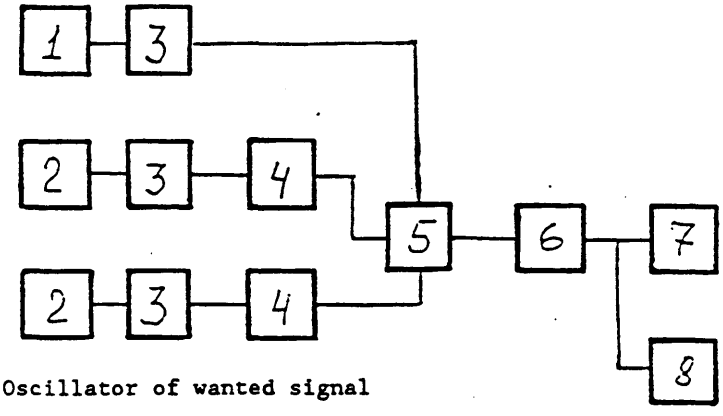
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APPENDIX

Method for measuring model parameters

To determine these parameters, it is necessary, using the SINAD method, to ascertain experimentally the variation in the signal-to-noise ratio S/N at the output of the receiver as a function of the signal level at the input P_s. The points thus obtained should be approximated by two straight lines, the first of which is inclined at an angle of 45° and the second is horizontal. The intersection corresponds to the parameter P_{so} (see Figure 7). The parameter P_{so} has the following physical meaning: when $P_s = P_{so}$ the level of the converted noise of the local oscillator from the band $[f_h - B; f_h + B]$ (f_h - oscillator frequency, B - half passband of main selectivity channel) when exposed to the effect of the wanted signal P_s is equal to the noise in the input stages at the output of the mixer.

It is then necessary to determine the variation in receptivity W as a function of Δ_1 when P_s = const. The points obtained should be approximated by two straight lines (see Figure 8). Point Δ_1 corresponds to half the passband of the preselector. The arrangement of the measuring unit is shown in Figure 9.



1. Oscillator of wanted signal
2. Oscillator of interfering signal
3. Attenuator
4. Band rejection filter
5. Matching device
6. Receiver
7. Load
8. Non-linear distortion meter

FIGURE 9

Test arrangement for measurements

REPORT 974

PROCEDURES FOR ANALYZING SPREAD-SPECTRUM INTERFERENCE TO CONVENTIONAL RECEIVERS

(Study Programme 18B/1)

(1986)

1. Introduction

This Report presents procedures that can be used to quantify the interactive effects between spread-spectrum (SS) interferers and conventional receivers. The procedures extend the use of the frequency-dependent-rejection (FDR) concept for calculating the frequency distance (FD) separation, which is described in Report 654. The FDR and FD are measures that aid in the solution of co-channel frequency sharing and adjacent band interference problems by providing minimum frequency and distance separation criteria between interferer and receiver for acceptable receiver performance.

A detailed description of SS modulation techniques is given in Report 651. The two general types of SS signalling techniques that are of interest here are direct-sequence (DS) and frequency-hopping (FH). The hybrid forms employing both of these methods (FH/DS) and time-hopped DS (TH/DS) modulation are also of interest.

2. General approach

In Report 654, a procedure is given for computing an FD curve, which is based on the inequality:

$$L_b(d) + FDR(\Delta f) \geq P_t + G_t + G_r - I_m \qquad (1)$$

where:

$L_b(d)$: basic transmission loss for a separation distance d between interferer and receiver (dB),

P_t : interferer transmitter average power (dBm),

G_t : gain of interferer antenna in direction of receiver (dBi),

G_r : gain of receiver antenna in direction of interferer (dBi),

I_m : maximum allowed average interference power to prevent unacceptable performance (dBm),

$FDR(\Delta f)$: defined by equation (2) in Report 654 (dB).

Inequality (1) is applicable when the unwanted emission is continuous in time and produces a response waveform that is also continuous in time. When the emission is frequency-hopped or time-hopped, it produces a response waveform that may not be continuous in time, and the performance is dependent on the peak rather than the average interference power. For that condition, inequality (1) is replaced by:

$$L_b(d) + \widehat{FDR}(\Delta f) \geq \hat{P}_t + G_t + G_r - \hat{I}_m \qquad (2)$$

where:

$\hat{P}_t$: interferer transmitter peak power (dBm),

$\hat{I}_m$: maximum allowed peak interference power to prevent unacceptable performance (dBm).

The function $\widehat{FDR}(\Delta f)$, which is used when peak interference levels are involved is obtained by:

$$\widehat{FDR}(\Delta f) = FDR(\Delta f) + \eta_i - \eta_0(\Delta f) \quad (3)$$

The terms η_i, the duty cycle of the unwanted emission waveform at the receiver input, and $\eta_0(\Delta f)$, the duty cycle of the response waveform, are defined in Report 972. Expressions for these two terms given in Table III of that Report are applicable to gated digital signals and can be used also when the responses to the SS emission consist of impulses or bursts.

After all of the terms except $L_b(d)$ have been calculated using the procedures given in the following sections, an FD curve can be obtained using the procedures given in Report 654.

The primary subject of this Report is the procedures that are used to calculate I_m and $\hat{I}_m$ as required for the various types of SS emissions.

3. Response waveform considerations

As explained and illustrated in Report 972, the response waveforms produced by emissions involving digital signalling take a number of different forms that depend on the characteristics of the interfering signal and the selectivity of the receiver [Newhouse, 1978]. The procedure required to compute the peak interference levels and FD for a particular type of SS emission depends on the type of response waveform that is produced.

3.1 *Noise-like response, continuous in time*

When the response is noise-like and continuous in time, which occurs when a receiver is subjected to a DS emission that has a bandwidth greater than the bandwidth of the receiver, the interference can be considered as additional receiver noise, as shown in Report 652. The effects on the receiver performance can be determined using a measure that is appropriate for noise interference.

3.2 *Response continuous in amplitude and time*

A response that is continuous both in amplitude and time is produced by a co-channel DS emission when the receiver bandwidth is greater than the instantaneous emission bandwidth.

3.3 *Impulsive and burst responses*

Some types of SS emissions, namely TH/DS, FH, and FH/DS, usually produce responses that consist of randomly spaced impulses or noise bursts. Also, adjacent channel interference from DS emissions produces randomly spaced impulses when the receiver bandwidth is greater than the emission bandwidth, as explained in Report 972.

Tables given in Report 972 give equations for calculating the input and output waveform duty cycles, η_i and $n_0(\Delta f)$ respectively, for various bandwidth conditions. The fraction of time that the response waveform is non-zero is denoted by DC, which is approximated by:

$$DC = \text{antilog}\,[0.1\ \eta_0(\Delta f)] \quad (4)$$

When the response waveform consists of short bursts and has a low duty cycle, signal-to-peak interference ratios ($S/\hat{I}$) much less than 0 dB can be tolerated. For example, with experienced listeners, the intelligibility as measured by articulation score in an AM voice transmission subjected to radar pulses is not reduced even when ($S/\hat{I}$) = −70 dB if the duty cycle of the incoming pulse is less than 8% [Hatch *et al.*, 1971]. Other observations indicate that to avoid listener annoyance, the duty cycle of a response consisting of noise bursts should not exceed about 5%. Although criteria are available for interference involving pulses that are of the order of 1 μs long [Licklider and Goffard, 1947], criteria that are applicable to the longer burst lengths that are typical of intermittent SS interference have not been published.

Guidelines for determining $\hat{I}_m$ when the interference produces intermittent responses are given in Table I for AM voice receivers, Table II for FM voice receivers, and Table III for conventional digital receivers. Note from the tables that the receivers must be protected from several types of interference mechanisms, namely: listener annoyance, loss of intelligibility, excessive probability of symbol error, impairment of the automatic gain control, and impairment of the synchronization process.

TABLE I – *Guidelines for determining $\hat{I}_m$ for intermittent * interference in AM voice receiver*

Conditions	Requirement	Procedure
$\tau_B \leq 1$ ms and DC $\leq 5\%$	Prevent loss of intelligibility	Use procedures given in § 4.3.2 or 4.4.2
$\tau_B \leq 1$ ms and DC $> 5\%$	Prevent listener annoyance	Treat as if the duty cycle were 100%, use procedures given in Report 654 (1)
$\tau_B > 1$ ms	Prevent impairment to AGC	Treat as if the duty cycle were 100%, use procedures given in Report 654 (1)

τ_B: length of interference bursts (s).

DC: duty cycle of interference response (%).

* Includes responses to TH/DS, FH and FH/DS emissions.

(1) This procedure is sufficient to satisfy the requirement; however, it may not be necessary under all circumstances. A more comprehensive procedure may show that less severe constraints would be adequate to protect the receiver.

TABLE II – *Guidelines for determining $\hat{I}_m$ for intermittent interference in FM voice receivers*

Type of emission	Conditions	Requirement	Procedure
TH/DS	DC $< 5\%$ and $\tau_B > 100$ ms	Prevent loss of intelligibility	Use procedures given in § 4.3.3
TH/DS	DC $> 5\%$	Prevent listener annoyance	Treat as if DC = 100%, use procedures given in Report 654 (1)
FH or FH/DS	$\tau_B < 100$ ms	Prevent listener annoyance	Use procedure given in § 4.4.3
FH or FH/DS	$\tau_B \geq 100$ ms	Prevent listener annoyance or loss of intelligibility	Use procedure given in § 4.4.3

(1) This procedure is sufficient to satisfy the requirement; however, it may not be necessary under all circumstances. A more comprehensive procedure may show that less severe constraints would be adequate to protect this receiver.

TABLE III – *Guidelines for determining $\hat{I}_m$ for intermittent interference in conventional digital receiver*

Conditions	Requirement	Procedures
$\tau_B < T_u$ and $\tau_B < T_s$	Prevent unacceptable probability of symbol error	Use procedures given in § 4.3.1 or 4.4.1
$\tau_B > T_u$	Prevent impairment of AGC	Treat as if the duty cycle were 100%, use procedures given in Report 654 (1)
$\tau_B > T_s$	Prevent impairment to synchronization process	Treat as if the duty cycle were 100%, use procedures given in Report 654 (1)

τ_B: length of interference bursts (s).

T_u: AGC attack time (s).

T_s: synchronization break time (s).

(1) This procedure is sufficient to satisfy the requirement; however, it may not be necessary under all circumstances. A more comprehensive procedure may show that less severe constraints would be adequate to protect the receiver.

4. Procedures for determining $\hat{I}_m$

4.1 *General*

A general expression for obtaining $\hat{I}_m$ is:

$$\hat{I}_m = N_0 + 10 \log(10^{M_I/10} - 1) \quad (5)$$

where N_0, the receiver noise power, is obtained with equation (3) in Report 654, and M_I, the interference margin, is given by:

$$M_I = (S/N)_0 - (S/N)_i \quad (6)$$

where:

$(S/N)_0$: signal-to-noise ratio (dB) when the receiver is not subjected to an unwanted emission,

$(S/N)_i$: signal-to-(noise-plus-interference) ratio (dB) when the receiver is subjected to the SS emission.

In some interference analysis either $(S/N)_0$ and $(S/N)_i$, or M_I may be specified. In other analyses, the performance in the absence of interference and the minimum allowed performance in the presence of interference are specified. For voice transmissions, the articulation score (AS) is the performance measure; AS_0 denotes the AS in the absence of interference and AS_i denotes the AS in the presence of interference. For digital transmissions, the probability of symbol error (P_e) is the performance measure; $P_{e,0}$ denotes P_e in the absence of interference, $P_{e,i}$, denotes P_e in the presence of interference, and $P_{e,I}$ denotes the probability of error of symbols that are coincident in time with an impulse or burst response produced by the SS emission.

AS curves for noise interference are given in Report 526. An alternative measure for voice transmission is the articulation index (AI). AI curves for noise interference are given in Report 652. AS and AI curves for other types of interference are also available in [ECAC, 1975]. Curves of P_e for noise interference are available in Weaver [1962] or other publications. Methods for calculating P_e for co-channel interference (CW-like) are also available in [Juroshek, 1976; Rosenbaum, 1968]. Whether the response will be noise-like or CW-like depends on the characteristics of the emission and the receiver selectivity, as explained and illustrated in Report 972.

4.2 *Procedure used for DS emissions*

A DS emission produces a response that is continuous in time when the emission bandwidth exceeds the receiver bandwidth. For that condition, inequality (1) and the procedures in Report 654 are used to determine I_m and FD. As it is very unlikely that the bandwidth of a DS emission will be less than the bandwidth of an AM-voice or FM-voice receiver, the procedure for that condition is not included here. For a digital receiver with a bandwidth that exceeds the DS emission bandwidth, the procedure given in § 4.3.1 is applicable. In using equation (4) to calculate DC, $\eta_0(\Delta f)$ is obtained from Table III of Report 972 using the equation that is appropriate for a continuous input waveform.

4.3 *Procedure used for TH/DS emissions*

An important characteristic of the response waveforms produced by SS emissions is the fraction of time the amplitude of the response waveform is non-zero. An approximation of that fraction is the output duty-cycle $\eta_0(\Delta f)$ which is a function of τ_g, T_g, B_r, and Δf. Report 972 defines $\eta_0(\Delta f)$ and Table III in that Report lists expressions that can be used to calculate that function for responses produced by TH/DS emissions.

4.3.1 *Digital receivers*

For digital receivers, determination of $\hat{I}_m$ entails first determining $P_{e,I}$ using the relationship:

$$P_{e,I} = \left[P_{e,i}\, DC^{-1}(\Delta f)\right] + \left[P_{e,0}\,(1 - DC^{-1}(\Delta f))\right] \tag{7}$$

After $P_{e,I}$ has been calculated, the corresponding interference level that yields $P_{e,I}$ is determined. When the response is noise-like, a curve or expression for $P_e(S/N)$ is used to determine $(S/N)_i$. When the response is CW-like, i.e. continuous in amplitude, a curve or expression for $P_e(S/N, S/I)$ [Juroshek, 1976] is used to determine $S/\hat{I}$. $\hat{I}_m$ is then obtained by:

$$\hat{I}_m = (S/N)_0 - (S/\hat{I}) + N_0 \tag{8}$$

where $(S/N)_0$ is determined using a curve of $P_e(S/N)$, i.e. $(S/N)_0$ is the signal-to-noise ratio that yields the specified value of $P_{e,0}$. The receiver noise N_0 is obtained using equation (3) of Report 654.

Having determined both $(S/N)_i$ and $(S/N)_0$, M_I is calculated using equation (6), and then $\hat{I}_m$ can be obtained using equation (5).

4.3.2 *AM voice receivers*

For AM voice receivers, $\hat{I}_m$ is determined using the expression:

$$\hat{I}_m = 10 \log \left[\left(\frac{(S/N)_0}{(S/N)_i} - 1\right) \frac{N_0}{DC(\Delta f)}\right] \tag{9}$$

where:

$(S/N)_0$: signal-to-noise ratio that yields AS_0,

$(S/N)_i$: signal-to-(noise plus interference) ratio that yields AS_i.

Curves of the AS as a function of noise interference are given in Report 526.

Note. – None of the ratios or terms in the brackets are in dB.

4.3.3 *FM voice receiver*

When an FM voice receiver is subjected to a TH/DS emission which produces a response with a peak power greater than the desired signal power, the performance is dependent on the duty cycle of the response waveform rather than on the average or peak power. When $\tau_B > 100$ ms, unacceptable degradation will not occur if:

$$(1 - DC)\, 100 > AS_i \tag{10}$$

where:

τ_B : burst length of the response,

AS_i : minimum allowable articulation score.

To prevent listener annoyance, DC should not exceed 0.05 (5%). If DC exceeds 0.05, then the TH/DS emission should be treated as if the emissions were continuous in time which results in $DC = 1.0$. For that case, $\hat{I}_m = I_m$, and the procedures given in Report 654, which applies to emissions that are continuous in time, should be used.

4.4 *Procedure used with FH and FH/DS emissions*

The maximum allowed interference power for FH emissions is determined using a trial and error procedure. The peak power of the interference response, $\hat{I}(\Delta f_j)$, associated with each frequency in the hop set used by the FH emitter is determined by:

$$\hat{I}(\Delta f_j) = \hat{I}_a - \widehat{FDR}(\Delta f_j), \qquad j = 1, 2, \ldots, N_F \tag{11}$$

where $\hat{I}_a$ is a chosen value for the maximum allowed peak interference at the receiver input (at output of receiving antenna). As will be seen from the discussion that follows, the correct value for $\hat{I}_a$ is found by trial and error. $\widehat{FDR}(\Delta f_j)$ is obtained using equation (3), N_F is the total number of frequencies in the hop set, and:

$$\Delta f_j = f_r - f_j \tag{12}$$

where:

f_r: tuned frequency of receiver (Hz),

f_j: jth frequency in the hop set used by the FH emitter (Hz).

4.4.1 *Digital receivers*

When the response is noise-like, the probability of error $P_{e,j}$ for symbols that are coincident in time with an interference burst or impulse associated with the jth frequency in the hop set is a function of $(S/N)_j$, which is obtained by:

$$(S/N)_j = \frac{(S/N)_0}{1 + \dfrac{\hat{I}(\Delta f_j)}{N_0}}, \qquad j = 1, 2, \ldots, N_F \tag{13}$$

where N_0 is determined using equation (3) in Report 654 and then converting from dBm to mW.

Note. – None of the terms of equation (13) are in decibels.

When the response is CW-like, $P_{e,j}$ is a function of $(S/N)_0$ and $S/\hat{I}(\Delta f_j)$ where:

$$S/\hat{I}(\Delta f_j) = (S/N)_0 - \hat{I}(\Delta f_j) + N_0 \tag{14}$$

The expected probability of symbol error P_e for the value selected for $\hat{I}_a$ is given by:

$$P_e = \frac{1}{N_F} \sum_{j=1}^{N_F} \left[P_{e,j}\, \eta_f(\Delta f_j) + P_{e,0}\,(1 - \eta_f(\Delta f_j)) \right] \tag{15}$$

where:

$$\eta_f(\Delta f) = (\tau_0/\tau_f)\, DC(\Delta f) \tag{16}$$

τ_0 is the half-power burst length and τ_f is the frequency dwell time, which is equal to the reciprocal of the hopping rate R_h. As shown in Fig. 1, the transmitter is off during an interval of length σ_0 when the frequency is changed. After P_e has been calculated, it is compared to $P_{e,i}$. If $P_e \neq P_{e,i}$, a different value is chosen for $\hat{I}_a$ and the computations involving equations (11), (13), (14) and (15) are repeated. If $P_e = P_{e,i}$, the value that was chosen for $\hat{I}_a$ and used in equation (11) is the maximum allowed peak interference level at the receiver input.

The receiver performance is acceptable only if:

$$L_b(d) \geq P_t + G_t + G_r - \hat{I}_a \tag{17}$$

Using $L_b(d)$ obtained from above, the required distance separation can be determined using the procedure given in Report 654. To calculate an FD curve, different values are selected for f_r (see equation (12)) and the above process is repeated for each value of f_r.

4.4.2 *AM voice receivers*

For AM voice receivers, the articulation score is dependent on the average interference power I in the receiver bandwidth, for the conditions that were given in Table I. After a value is chosen for $\hat{I}_a$ and equation (11) is used to calculate $\hat{I}(\Delta f_j)$, $j = 1, 2, \ldots, N_F$, the average interference power is obtained by:

$$I = \frac{1}{N_F} \sum_{j=1}^{N_F} \eta_f(\Delta f_j) \cdot \text{antilog} \left[0.1\ \hat{I}(\Delta f_j)\right] \tag{18}$$

where $\eta_f(\Delta f_j)$ is obtained with equation (16). The maximum allowed average interference power I_m is obtained by:

$$I_m = \left(\frac{(S/N)_0}{(S/N)_i} - 1\right) N_0 \tag{19}$$

where N_0 is obtained using equation (3) in Report 654 and converted from dBm to mW. It is assumed here that $AS_i \geq 70\%$; for that condition, the maximum allowed peak power of the response will be less than the clipping level of the audio amplifier.

Note. – None of the terms of equation (19) are in decibels.

If $I_e \neq I_m$, a different value is chosen for $\hat{I}_a$ and the computations involving equations (11) and (18) are repeated. If $I = I_m$, the value chosen for $\hat{I}_a$ is the maximum allowed peak interference at the receiver input. $\hat{I}_a$ is then used in inequality (17) to determine the minimum allowed value for $L_b(d)$. The required distance separation is determined using the procedure given in Report 654. To compute an FD curve, different values are selected for f_r (see equation (12)) and the process given above is repeated for each value of f_r.

4.4.3 *FM voice receiver*

For FM voice receivers subjected to FH or FH/DS emissions, the performance is dependent on the duty cycle of the response rather than the average or peak power when the peak power of the interference response exceeds $S - 3$ dB, where S is the wanted signal.

When $\tau_B \geq 100$ ms, inequality (10) must be satisfied to prevent unacceptable degradation.

When $\tau_B < 100$ ms, DC must not be allowed to exceed 5%, otherwise listener annoyance will occur. That duty cycle requirement is satisfied when:

$$N_c = 0.05\ N_F \cdot \tau_f \tau_0 \tag{20}$$

where N_c is the maximum number of frequencies (within the hop set used by the emitter) for which $\hat{I} \geq S - 3$, $\hat{I}$ is the peak power of the IF response waveform, S is the wanted signal power, N_F is the total number of frequencies in the hop set, τ_f is the dwell time on each frequency, and τ_0 is the effective time the power is at the peak value during the dwell time (see Fig. 1).

The interference waveform will have a duty cycle of 5% or less when the following inequality is satisfied:

$$L_b(d) + \widehat{FDR}(\Delta f_x) > \hat{P}_t + G_t + G_r - \hat{I}_m \tag{21}$$

where:

$$\hat{I}_m = S - 3 \tag{22}$$

$$\Delta f_x = max \left[|\Delta f_h| - \left(\frac{N_F}{2} - N_c\right) \Delta f_c,\ \frac{N_c}{2} \Delta f_c \right] \tag{23}$$

Δf_c: spacing between the frequencies in the hop set (Hz), the spacing is assumed to be uniform,

$\Delta f_h = f_r - f_h$, and f_h is the centre of the frequency-hopping band.

If parameters of the emission and receiver are such that a duty cycle of 5% or less cannot be attained, if for example $N_F < 20$, then the emission should be treated as if it were continuous, i.e. $DC = 100\%$. For that case, $\hat{I}_m = I_m$, and the procedures in Report 654 which apply to emissions that are continuous in time, should be used.

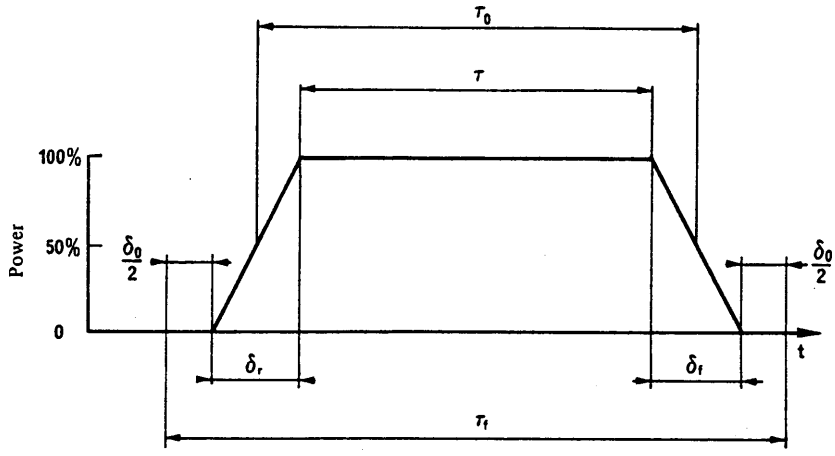


FIGURE 1 – *Envelope of an FH emission*

τ_f: frequency dwell time

δ_r: rise time

δ_f: fall time

δ_0: dead time

5. Conclusions

Procedures have been given for computing the maximum allowed interference power in receiver responses produced by a class of SS emissions. The computed value can be used to compute the frequency and distance separations required between the SS interferer and a conventional receiver. These computations are required for sharing problems involving co-channel and adjacent channel interference. If the computation of required distance separation results in a distance short enough to be considered as a co-site situation, additional co-site interference computations should be made.

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REPORT 1109*

MODEL OF EFFECTS ON RECEIVER NOISE DUE TO AN INTERFERING SIGNAL MIXING WITH THE LOCAL OSCILLATOR NOISE

(Question 44/1)

(1990)

1. Summary

This paper describes a model that explains the effect on receiver noise and sensitivity from the mixing of the local oscillator noise with a strong out-of-band interfering signal. The model has been tested on a number of FM receivers of the mobile service and of VHF radio-relay systems and has been found to provide satisfactory accuracy in the computer solution of problems of electromagnetic compatibility (analysis and prediction of interference, determination of frequency-distance separations which can eliminate interference, etc.). The model addresses only the effects of this contribution to receiver noise and must be used in conjunction with other models, particularly when dealing with adjacent channel or near adjacent channel interference. A method of calculating the quantitative values of the model parameters is described.

2. Model of receiver sensitivity to a single interfering signal

It is well known [Report 704; Recommendation 612; Grebenkemper C.J., 1981; Bogdanovich, B.M., 1984 (p. 39)] that interference can result in an increased noise level at the receiver output. This process may be described as follows for the case of a single interfering signal mixing with the local oscillator noise.

In addition to the main oscillation at the frequency f_{LO}, the local oscillator signal also contains a weaker noise spectrum which occupies a large bandwidth. The noise within the local oscillator signal causes an increase in the level of two noise components in the following intermediate frequency (IF) signal. The first noise source is due to the conversion in the mixer of the local oscillator noise from the frequency band $f_{LO} \pm B$ with the wanted signal, where B is half of the IF stage bandwith (See Figure 1). The second noise source results from the conversion in the mixer of the local oscillator noise from the frequency band $f_{LO} + \Delta f_i \pm B$ with the unwanted signal at f_i, where $f_i = f_o - \Delta f_i$ (See Figure 2).

In a number of cases, for example, FM receivers in the land mobile service, heterodyne noise conversion has no effect, since it is at a level below subsequent processing thresholds. In some cases, however, sensitivity reduction effects are noted. An analysis of heterodyne noise conversion is contained in the Annex. Using a partly linear approximation of the heterodyne and preselector noise-frequency curve, Alter, L. Sh. (1984) obtained an equation for P_n, the permissible level at the receiver input of a single interference signal that mixes with the local oscillator noise:

$$P_n = 10 \log (1 + p_s/p_{co}) + E_1 + u_1 \log \Delta f_i \text{ with } f_i \leq d \quad (1)$$

$$P_n = 10 \log (1 + p_s/p_{co}) + E_2 + u_2 \log \Delta f_i \text{ with } f_i > d \quad (2)$$

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* This Report should be brought to the attention of Study Group 3.

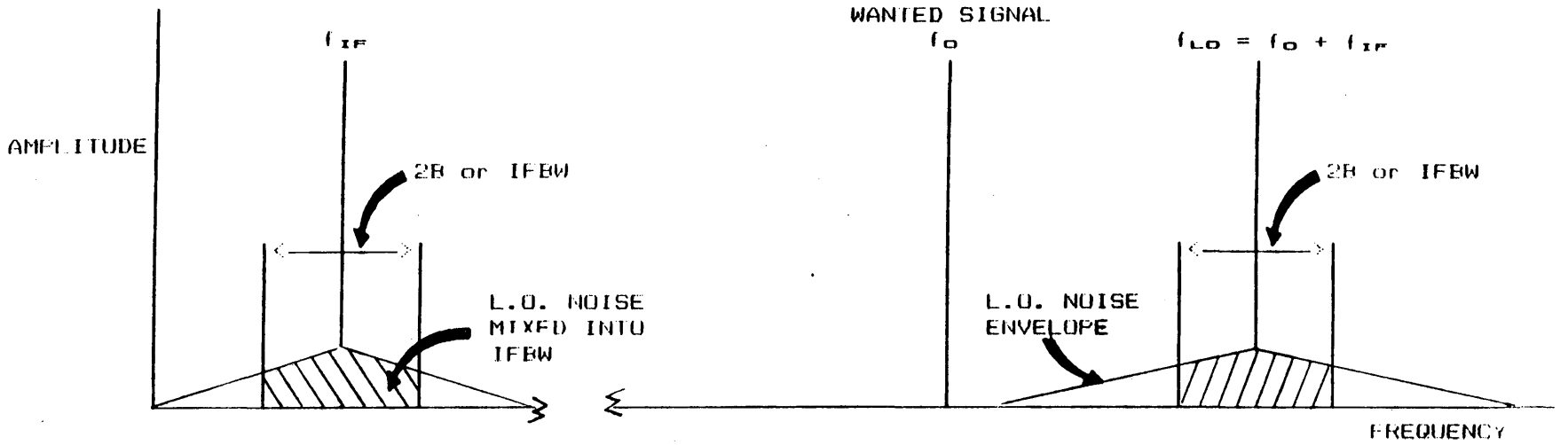


Figure 1. IF Noise Without Interfering Signal

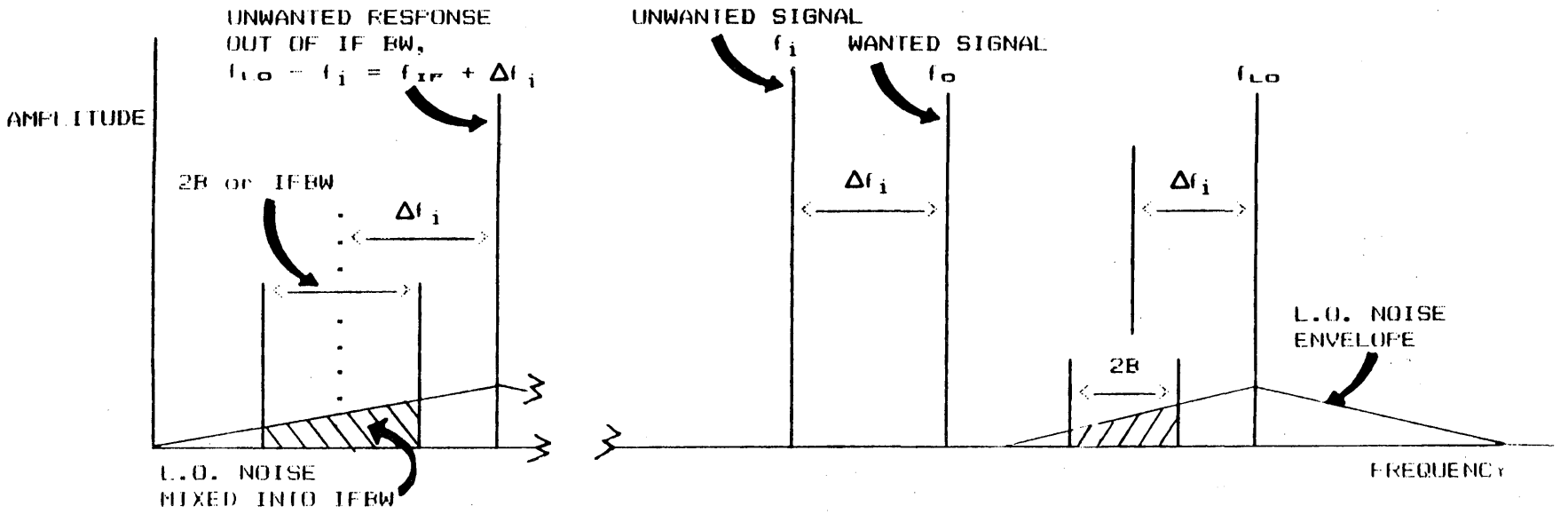


Figure 2. Additional IF Noise From Mixing L.O. Noise With Interfering Signal

where:

p_s is the level of the wanted signal at the receiver input, in mW; and

p_{co}, d, E_1, E_2, u_1, and u_2 are receiver model parameters (derived in the Annex and determined by methods shown in paragraph 3) expressed in mW, kHz, dBmW, dBmW, dB/decade, and dB/decade, respectively.

The relationship of the permissible interference level (P_n) to Δf_i may be approximated by two straight lines on a logarithmic scale. The intersection of these two lines establishes the break-point (Δf_i=d) for determining the appropriate equation, (1) or (2). At small frequency offsets ($\Delta f_i < d$), P_n is determined principally by the local oscillator noise characteristics, while at large frequency offsets ($\Delta f_i > d$), P_n is principally determined by the RF preselector characteristics. Graphic examples of the model are presented in Figures 3 and 4. (CAUTION: p_s and p_{co} are in mW in equations (1) and (2), but have been converted to P_s and P_{co} in dBm in Figures 3 and 4). In Figure 3A, P_n is related to P_s, at a constant Δf_i = 200 kHz, and in Figure 4, P_n is related to Δf_i at full receiver sensitivity. These figures are based on measurements of a particular receiver.

3. Method of determining the receiver model parameters

The receiver model parameters are measured according to the block diagram in Figure 5. Generator 1 provides the wanted signal f_o and generator 2 provides the unwanted signal f_i. If necessary, a band elimination filter is connected to the f_i generator output to attenuate the f_i generator noise in the receiver RF band.

To determine the model parameter p_{co} with f_i disconnected, the relationship of S/N as a function of P_s is producted. By varying the wanted signal level P_s at the receiver input in steps from the receiver sensitivity level to the maximum practical level, we can determine the corresponding S/N at the receiver output. The data obtained are used to plot a curve approximated by two straight lines as shown in Figure 3B. The intersection of the two straight lines corresponds to the value $P_s=P_{co}$.

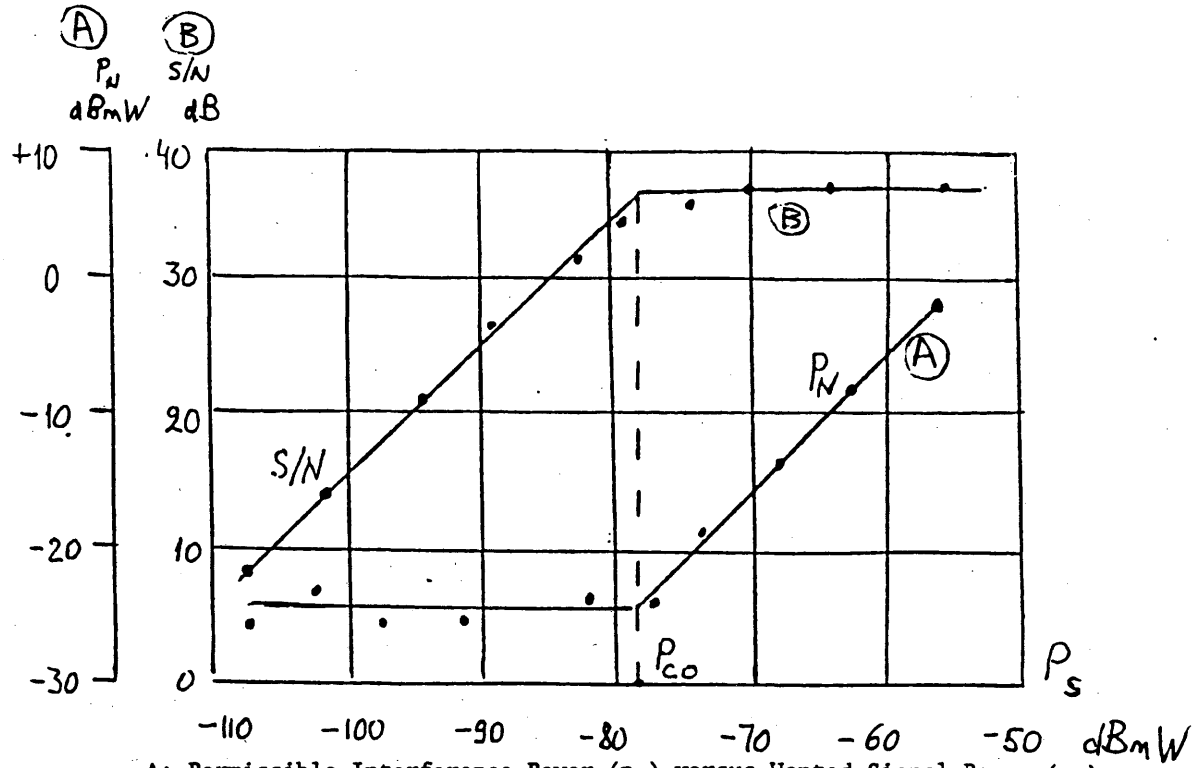
To determine the model parameters d, E_1, E_2, u_1, and u_2 the relationship of P_n as a function of Δf_i is produced, holding P_s constant. For this purpose, P_s is output from generator 1 at a level equal to the receiver sensitivity, with a modulating tone of 1 kHz at the nominal frequency deviation, while generator 2 provides an unmodulated f_i at a selected Δf_i, increased in level until the S/N at the receiver output is reduced by 6 dB. The resultant interference power value provided to the receiver input is the maximum permissible level of a single interfering signal, P_n, for this particular Δf_i. Repeating this process for a variety of Δf_i, we obtain data to plot a curve approximated by two straight lines on a logarithmic scale as shown in Figure 4. The parameter d is determined as the abscissa of the intersection of the two sloping straight lines. The parameters u_1, u_2, E_1, E_2 are calculated by the following formulae from Alter, L. Sh. (1984).

$$u_1 = (P_n'' - P_n')/(\log d - \log d') \quad (3)$$

$$E_1 = P_n'' - u_1 \log d \quad (4)$$

$$u_2 = (P_n''' - P_n'')/(\log d''' - \log d) \quad (5)$$

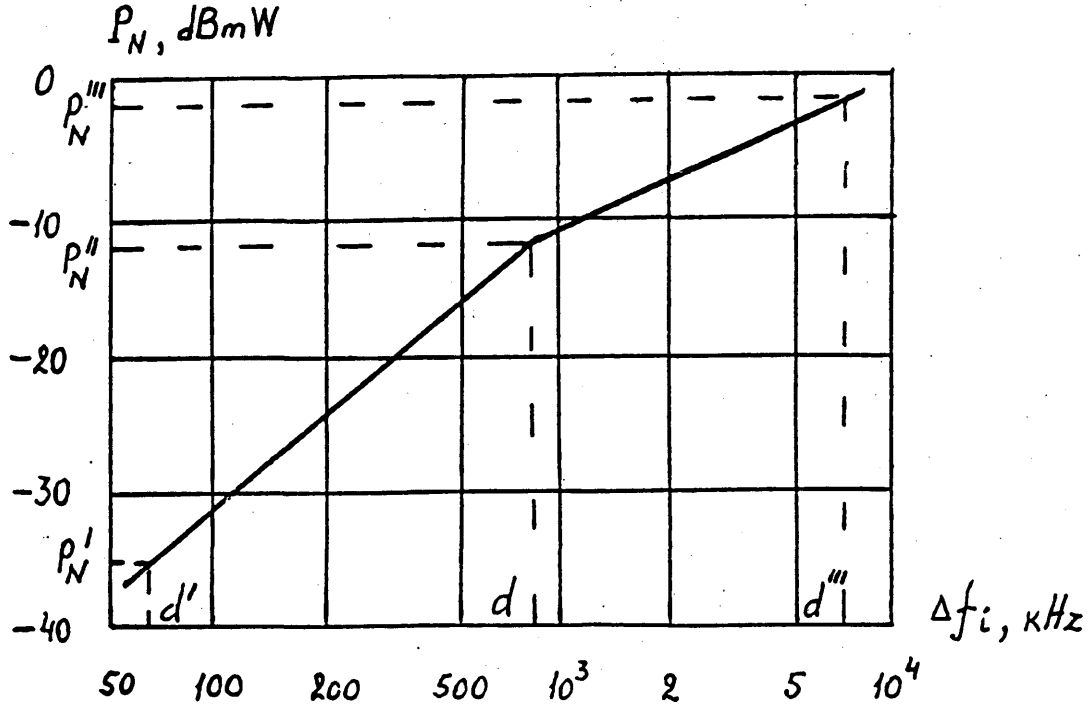
$$E_2 = P_n''' - u_2 \log d''' \quad (6)$$



A: Permissible Interference Power (p_n) versus Wanted Signal Power (p_s) at a constant Δf_i = 200 kHz

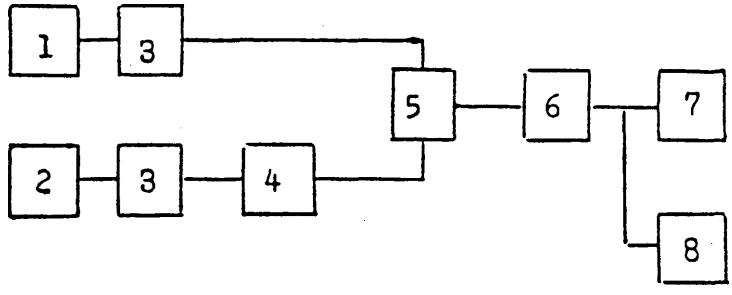
B: output signal/noise ratio (S/N) versus Wanted Signal Power (p_s) with Δf_i disconnected

FIGURE 3



Permissible Interference Power (p_n) versus Δf_i with Wanted Signal Power (p_s) at receiver sensitivity

FIGURE 4



1 - Wanted signal generator
2 - Interfering signal generator
3 - Attenuator
4 - Band elimination filter
5 - Matching device
6 - Receiver
7 - Load
8 - Non-linear distortion meter

FIGURE 5

TEST ASSEMBLY FOR MODEL PARAMETER DETERMINATION

Figure 6 illustrates a family of P_n verus P_s curves for various Δf_i. As shown by the term $(1 + p_s/p_{co})$ in equations (1) and (2), for each Δf_i, the permissible interference signal level is nearly a constant as long as the wanted signal level is below P_{co}. Above P_{co} the permissible interference level rises directly with the wanted signal level (P_s). Detatils of the method are presented in the Annex.

4. Sample calculation

For the receiver used to develop Figure 3, the parameters E_1, E_2, u_1, and u_2 may be determined using equations (3-6). From the figure:

$d' = 60$ kHz $\quad P_n' = -35$ dBm

$d = 800$ kHz $\quad P_n'' = -12$ dBm

$d''' = 7000$ kHz $\quad P_n''' = -2$ dBm

Using these values:

$$u_1 = [-12 - (-35)] / \log (800/60)$$

$$= 20.45 \text{ dB/decade}$$

$$E_1 = (-12) - 20.45 \log 800$$

$$= -71.4 \text{ dBm}$$

Similarly:

$$u_2 = 10.6 \text{ dB/decade}$$

$$E_2 = -42.76 \text{ dBm}$$

To check, let $\Delta f_i = 1$ MHz ($\Delta f_i > d$) and $P_s = -60$ dBm:

$$P_n = 10 \log [1+(10^{-6}/10^{-7.8})] + (-42.76) + 10.6 \log 1000$$

$$= 7.1 \text{ dBm}$$

See sample point on Figure 6.

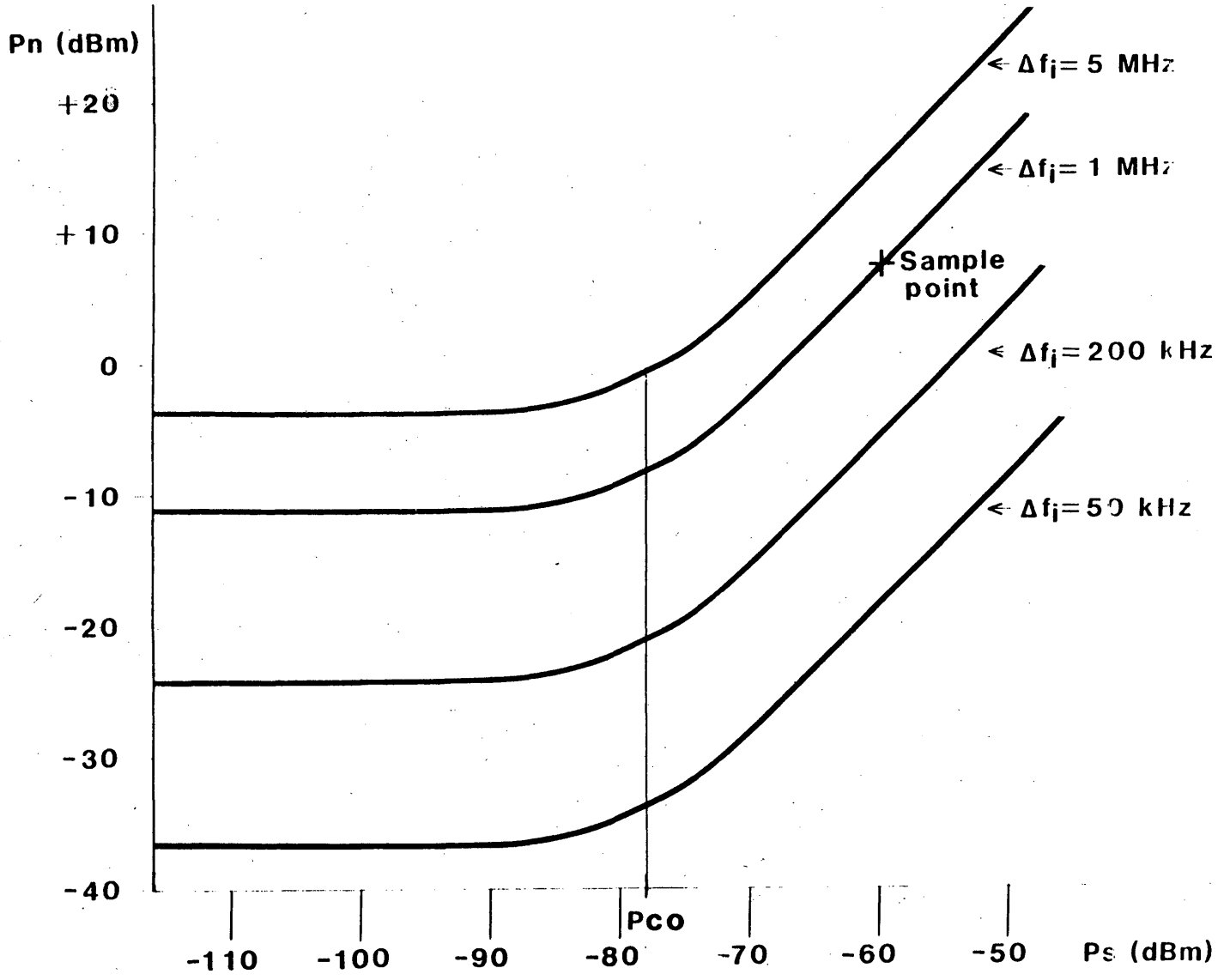


FIGURE 6

Permissible Interference Power (Pn) vs Wanted Signal Power (Ps) for a sample FM receiver (from equations one and two)

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ANNEX

The local oscillator noise power from the band ($f_{LO} \pm B$), mixed into the IF channel by the wanted signal, and the local oscillator noise power from the band ($f_{LO} + \Delta f_i \pm B$) mixed into the IF channel by the interfering signal, are equal, respectively, to:

$$P_{n1} = 2BS_{ho}P_s \prod_{m=1}^{z} k_m^2 \qquad (1)$$

$$P_{n2} = 2BS_h(\Delta f_i)p_i \prod_{m=1}^{z} [k_m y_m(\Delta f_i)]^2 \qquad (2)$$

where

S_{ho} = normalized mean spectral noise power density in the band ($f_{LO} \pm B$)

$S_h(\Delta f_i)$ = normalized mean spectral noise power density in the band ($f_{LO} + \Delta f_i \pm B$)

P_s = receiver wanted input signal, mW
p_i = receiver unwanted input signal, mW
k_m = gain of the m^{th} stage of RF
$y_m(\Delta f_i)$ = discrimination factor of m^{th} stage of RF
z = number of stages of RF amplification including the mixer

The signal-to-noise (S/N) ratio at the mixer output including the effect of interfering power (p_i) present at the receiver input is:

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$$(p'_s / \sum_{j=0}^{2} p_{nj}) = \frac{p_s \prod_{m=1}^{z} k_m^2}{p_{no} + p_{n1} + p_{n2}} \quad (3)$$

where:

p_{no} = the inherent mixer output noise power in the case, when $p_s = p_i = 0$.

Let us take as the interference criterion a reduction (C) of the receiver output S/N, which corresponds to the same reduction of the S/N ratio at the limiter input.

In this case, we obtain p_n, a permissible level of interference power, p_i at the receiver input of:

$$p_i = \frac{p_{no}(1 + p_s/p_{co})\ (C-1)}{2BS_h(\Delta f_i) \prod_{m=1}^{z} [k_m y_m(\Delta f_i)]^2} \quad (4)$$

where:

$$p_{co} = p_{no}/2BS_{ho} \prod_{m=1}^{z} k_m^2 \quad (5)$$

To determine the frequency dependence of p_n, we must consider the amplitude-frequency curves $S_h(\Delta f_i)$ and $y_m(\Delta f_i)$ • $S_h(\Delta f_i)$ can be approximated as follows:

$$S_h(\Delta f_i) = S_{hq}(\Delta f_i / F_h)^{u_{1.1}} \text{ where } \Delta f_i < F_h \quad (6)$$

$$= S_{hq} \text{ where } \Delta f_i \geq F_h \quad (7)$$

where:

$F_h = f_{LO}/2Q_h$, and

S_{hq}, $u_{1.1}$, and F_h are parameters of approximation of the local oscillator noise curve, and

Q_h is the figure of merit of the local oscillator circuits.

The amplitude-frequency curve of the preselector may be approximated by the following:

$$y(\Delta f_i) = (1 + \lambda^{-2}\ \sigma^{2\lambda})^{-z/2} \quad (8)$$

$$\sigma = 2\ Q_p\ \Delta f_i / f_o$$

where:

λ = 1 or 2 for 1 or 2 circuit filters

Q_p = figure of merit of the preselector filters

In this case:

$$y(\Delta f_i) = 1 \text{ where } \Delta f_i \leq F_p \quad (9)$$

$$= (\lambda^{-1}\ \sigma^{\lambda})^{-z} \text{ where } \Delta f_i > F_p \quad (10)$$

and $F_p = f_o/2Q_p$

Introducting the symbols of the receiver model:

$$E_{1,1} = \frac{P_{no}(C-1)}{2B \prod_{m=1}^{z} k_m^2} \left[\frac{F_h}{S_{hq}}\right]^{u_{2.1}} \quad \text{and} \quad u_{2.1} = \lambda z \tag{11}$$

$$E_{2,1} = \frac{P_{no}(C-1)}{2B \prod_{m=1}^{z} k_m^2 \ \lambda^z} \left[\frac{2Q_p}{f_o}\right]^{\lambda z} \tag{12}$$

Using equations (3) and (4), and considering that generally $F_h \tilde{=} F_p$, we may express the permissible interference level p_n as follows:

$$p_n = (1 + p_s/p_{co})\, E_1(\Delta f_i)^{u_{1.1}} \quad \text{where} \quad \Delta f_i \leq d \tag{13}$$

$$= (1 + p_s/p_{co})\, E_2(\Delta f_i)^{u_{2.1}} \quad \text{where} \quad \Delta f_i > d \tag{14}$$

REPORT 324-3

METHODS OF MEASURING THE BANDWIDTHS OF EMISSIONS

(Study Programmes 60A/1 and 60C/1)

(1966-1970-1978-1990)

1. Measurement of x dB bandwidth

To determine in practical conditions, the degree of "optimality" of emissions in keeping with Recommendation 328, § 2, in the proximity of operating transmitters and at monitoring stations, x dB bandwidth measurements may be carried out by means of spectrum analyzers (Recommendation 327, § 1.1). The operating characteristics of spectrum analyzers in various conditions and the requirements concerning the choice of spectrum analyzer parameters in various operating regimes are given in § 2.3 and in Annex I of Recommendation 327.

In making such measurements, various x dB level values may be applied, for example, -26 dB and -6 dB relative to the peak emission value determined by the spectrum analyzer, if the required relationships between the x dB bandwidths at these levels and the necessary bandwidth are known. To determine this peak value, the sweep is made inoperative and the filter bandwidth of the analyzer made sufficiently large to accept all significant components of the emission.

Reports 275 and 420 give, for a number of classes of emission, x dB level values at which the x dB bandwidth readings are equal or nearly equal to the necessary bandwidth, provided that the "optimality" condition of the emission is fulfilled.

According to USSR Standards [1976], a standard level of -30 dB relative to the unmodulated carrier level (for telegraphy classes of emission) or the maximum level of spectral power density in the frequency sideband (for broadcasting and telephony classes of emission) is used for checking of emission "optimality" in the USSR. The "reference bandwidth" determined at this level must be the necessary bandwidth for the class of emission concerned in the relationship shown in Table I or must be narrower. The x dB bandwidth values at lower levels, namely -40, -50 and -60 dB, are taken as criteria of the permissible out-of-band spectrum (Recommendation 328, § 1.10).

In many cases, when essentially determined by the filter characteristics, the slope in dB per logarithmic unit of frequency of those parts of a spectrum which are distant from the central part becomes asymptotic to a line of constant slope and may therefore be expressed by the power functions:

$$S_1(f) = S(f_m)\left(\frac{f_m}{f}\right)^{\gamma} \qquad (1)$$

$$\gamma = 0.33\ N$$

where $S(f_m)$ is the power on a given frequency f_m, and N is the number of dB by which the spectrum envelope is reduced within a single octave of band widening. However, within parts near the limits of the necessary bandwidth, it is frequently observed that spectra show a slope almost constant in dB per unit frequency and may therefore be expressed by the exponential functions:

$$S_2(f) = S(f_m)\ \exp\left[-\frac{0.23\ N_1}{f_m}\ (f-f_m)\right] \qquad (2)$$

where N represents the number of dB corresponding to the first octave of band widening.

On the basis of such an approximation of spectrum envelopes, [CCIR, 1974-78] contains an evaluation of the corresponding relative errors of δ_i in x dB level bandwidth measurement related to the error in Δb (in dB) in the reproduction of the spectrum envelope levels by the analyzer. The results of the calculation are given in Fig. 1. The figures alongside the curves refer to the values of N. Figure 1 shows that, in order to obtain sufficient measurement accuracy, an analyzer accuracy within ± 1 dB is required.

TABLE I

Class of emission	Relationship between B_m and B_n or formulae for calculation of B_m	B_n value	Method of determining "0 dB"
A1A, A1B, A2A, A2B	$B_m = B_n$	Rec. 328	Using unmodulated carrier level (Report 275, § 9)
A3EGN Sound broadcasting	$B_m = 1.2\ B_n$	Idem	Apply maximum spectral power density level in side band (using data in Report 325, § 9.1)
A3EJN Telephony	$B_m = 1.9\ B_n$	Idem	Idem
R3E	$B_m = 1.15\ B_n$	Idem	Idem
B3E	$B_m = 1.05\ B_n$	Idem	Idem
H3E, J3E	$B_m = 1.15\ B_n$	Idem	Idem
F1B	$B_m = 4.3\ \sqrt{mB}$ for $1.5 \leq m < 5.5$ $B_m = (m + 7)\ B$ for $5.5 \leq m \leq 20$	Idem	Using unmodulated carrier level (Report 275, §§ 7 and 9)
F3C	$B_m = 1.45\ B_n$ for $0.14 \leq m < 0.77$ $B_m = 1.35\ B_n$ for $0.77 \leq m \leq 1.7$ $B_m = 1.25\ B_n$ for $1.7 < m \leq 2.75$	Report 418-1 (Kyoto, 1978)	Idem
F7BDX	$B_m = B_n$	Rec. 328	Idem

Notes. – B - the modulation rate in bauds
m - modulation index
B_m – reference bandwidth

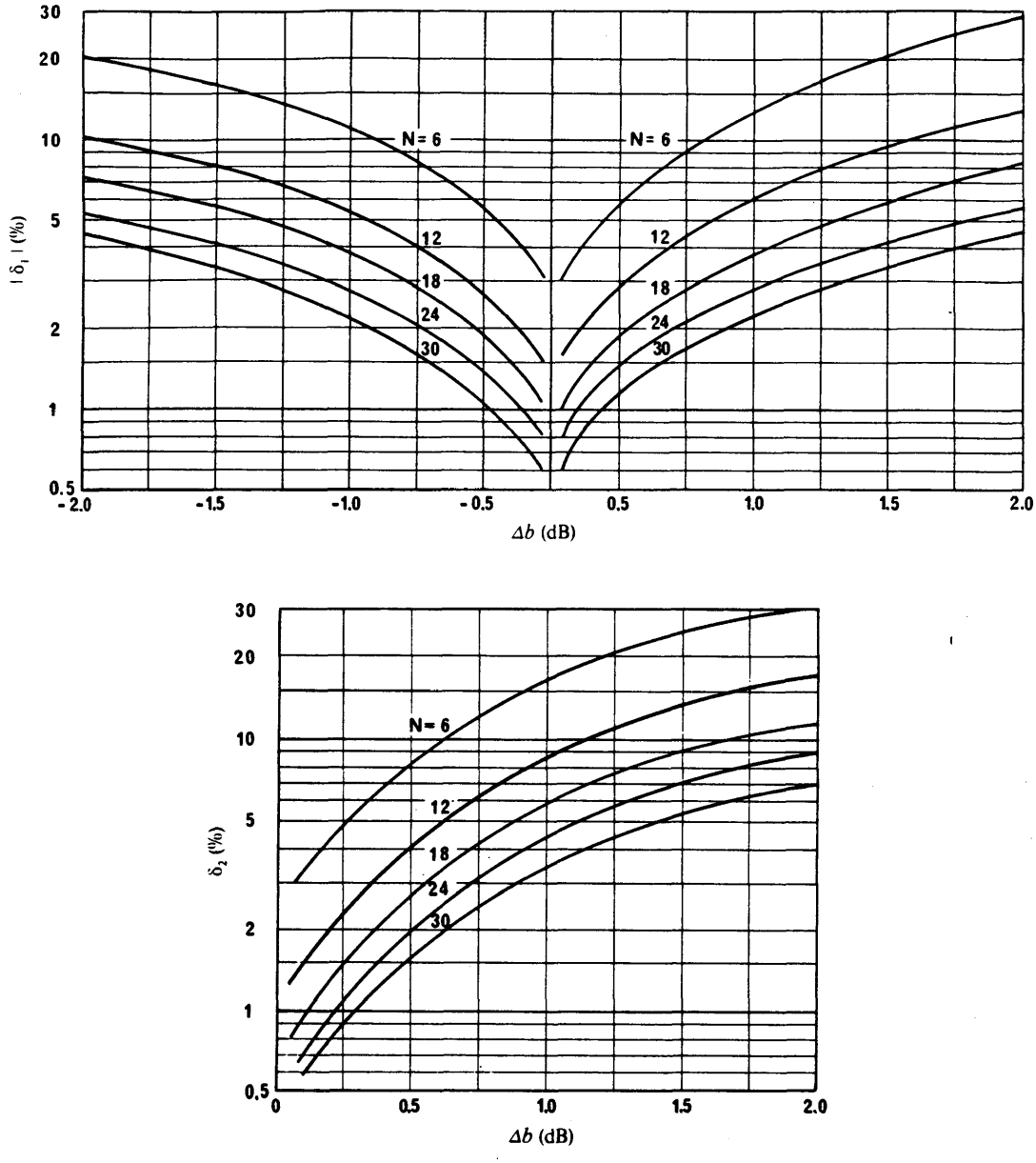


FIGURE 1 — *Relationship between the percentage error ($\delta_1 \delta_2$) in measuring the x dB bandwidth and the error (Δb) in reading the corresponding level, for different values of N*

2. Methods of measuring occupied bandwidth

2.1 *Direct measurement methods*

Measurements of occupied bandwidth by direct methods are conducted as specified in Recommendation 327. The relation of the errors in the measurement of occupied bandwidth δ' to the errors in the power comparison v is obtained from the spectrum envelope approximation described in § 1 and shown in Fig. 2. The solid lines correspond to the approximation according to Formula 1 and the dashed lines to the approximation according to Formula 2. Figure 2 shows that, for the most common values of N = 12 to 20 dB/octave, it is sufficient to carry out the power comparison at a very low accuracy of about ± 15 to 20% to ensure an occupied bandwidth measurement accuracy of ± 3 to 7%.

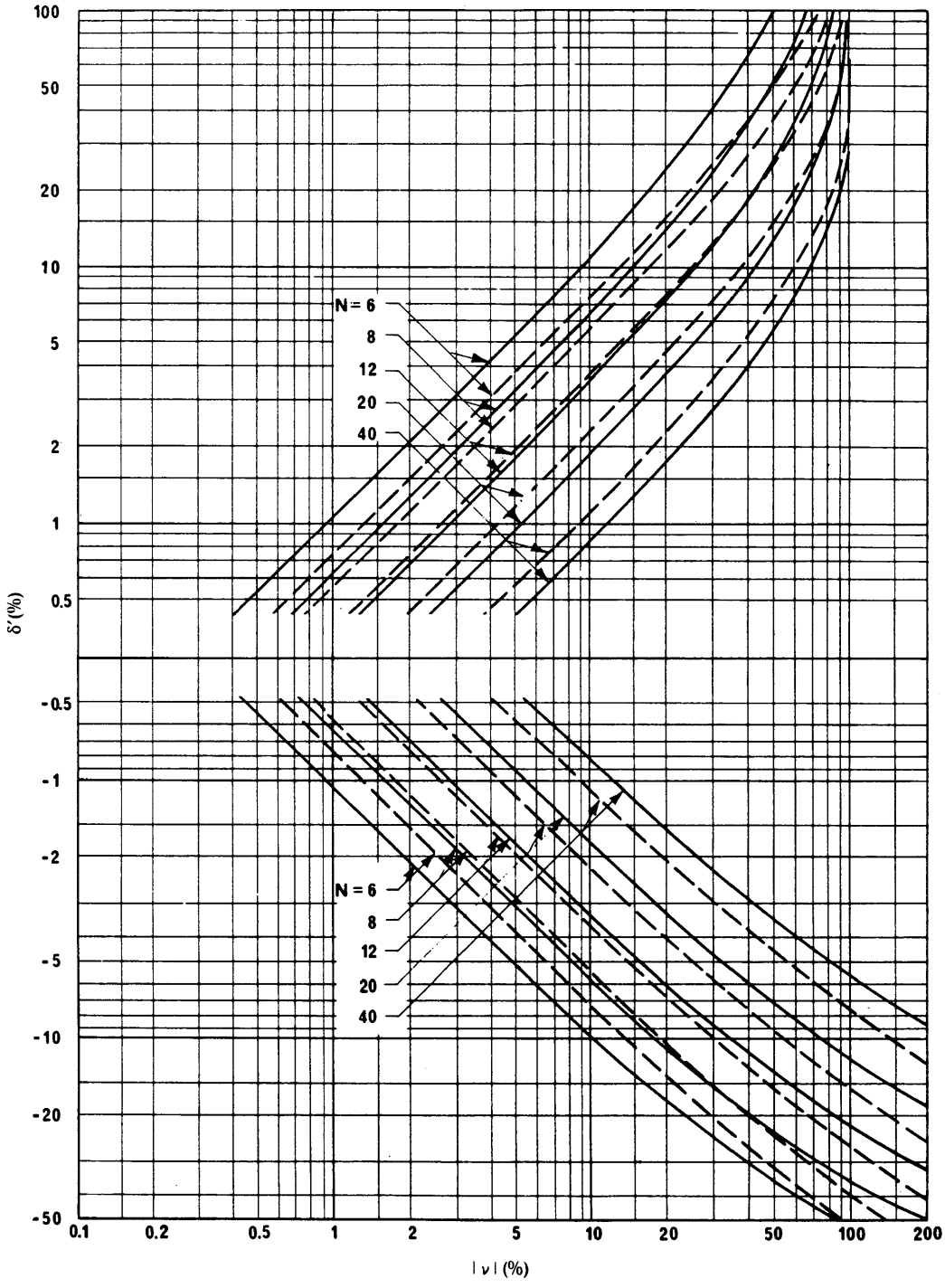


FIGURE 2 — *Relationship between the percentage error (δ') in measuring the occupied bandwidth and the percentage error (ν) in the power comparison, for different values of N*

2.2 *Indirect methods of measuring occupied bandwidth*

2.2.1 *Method of determination of occupied bandwidth by out-of-band spectrum slope*

On the basis of the spectrum envelope approximation described in § 1 the occupied bandwidth can be determined from the spectrum slope by integrating the spectral power density from the frequency around the presumed limit of the occupied bandwidth up to infinity. A detailed description of this method is contained in [CCIR, 1963-66]. As shown in § 11 of Report 275, the effect of interference on the measurement accuracy achieved with this method may be substantially reduced by determining the spectrum slope from the in-band part, i.e., at sufficiently high levels.

2.2.2 *Method of determining the bandwidth of a frequency-modulated single-channel radiotelephone emission by measuring the r.m.s. frequency deviation*

The bandwidth of an emission produced by a frequency-modulation transmitter provided with an amplitude limiter in the modulation channel, followed by a low-pass filter, can be approximated by measuring the r.m.s. value of the frequency deviation.

This method [CCIR, 1966-69], justification of which is given in [Gourevitch, 1963], considerably simplifies the approximate measurement of bandwidth as it does not require any special instrumentation, such as a spectrum analyzer.

2.3 *Effect of noise on the accuracy achieved in the measurement of emission bandwidth by various methods*

The effect of noise on the accuracy achieved in the measurement of emission bandwidth by various methods is discussed in Report 275, § 11.

3. Comparative bandwidth measurements by various methods

Document [CCIR, 1974-1978b] describes a measurement method and the results of occupied bandwidths and x-dB bandwidths measurement for transmitters operating in classes of emission AIB, FIB, F3C and F7BDX. The x-dB bandwidths were measured by means of a spectrum analyser, the parameters of which were selected in accordance with Annex IV to Recommendation 328-6, and the x-dB levels in accordance with Table I of Report 275-4. The results of the x-dB bandwidth measurements were compared with the results of direct and indirect measurments of occupied bandwidth carried out in accordance with sections 4.2 and 4.3 of Recommendation 328-6, and also with the results of the theoretical calculations of these bandwidths. The measured and calculated occupied bandwidth values were based on the quantity $\beta/2 = 0.5\%$ (section 1.13 of Recommendation 328-6). The measurements were carried out in the far field but close enough to the transmitter to prevent fading.

The results of the measurements conducted by various methods and of the calculations showed a good agreement, which confirms the practical validity of these methods, at least in stable field conditions.

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CCIR Documents

[1963-66]: 1/42 (Japan).

[1966-69]: 1/9 (USSR).

[1974-78]: a. 1/154 (USSR); b. 1/56 (USSR).

REPORT 841-2

SPECTRUM MANAGEMENT AND COMPUTER-AIDED TECHNIQUES

(Decision 27-2)

(1982-1986-1990)

1. General

The CCIR has the task of studying the necessary technical bases to facilitate the introduction of computer-aided frequency management procedures, the development of recommended interference analysis techniques and the development of an agreed format for the exchange of computer generated coordination information. A handbook on "Spectrum Management and Computer-Aided Techniques" has been prepared by the CCIR at the request of the WARC-79 (Recommendation No. 31). This Report summarizes the findings in the handbook and describes spectrum management techniques, spectrum engineering models, data bases and computer-aided techniques.

The data files and analysis techniques used by administrations vary from simple to very complex systems. If the basic file size is small (less than approximately ten thousand data records), there is no long-term expected growth, and complex computations are not required, a manually-operated system may be sufficient. However, once the data file contains tens of thousands of records and continuing growth is expected, serious consideration should be given to the automation of the data file and relevant analysis techniques.

Automated analysis techniques have numerous advantages over manual techniques and become indispensable as the data processed becomes voluminous. However:

- the changeover from a manual to an automated process will initially create new types of problems;
- the initial period of automated system development and implementation may be costly. The user should realize that it requires time before he can receive all the advantages and financial benefits of an automated system;
- there is an infrastructure that should be analyzed, planned and carried out before starting an automated system.

2. Spectrum management techniques

2.1 *Background*

The ever advancing development in communications led to the creation of new fields of activity and a steady expansion of radiocommunications. Because of the high utilization of the radio frequency spectrum, it is no longer possible to comply with all requests for the allocation of interference-free frequencies to individual radio services with conventional means and methods.

The requirements of individual radio stations or services can be met only by sharing time, space, or frequency. Moreover, bilateral or multilateral agreements will increasingly be required. This means that general assignment procedures in the form of simple network plans for the use of frequencies cannot always be employed in the near and especially in the far future. It will become necessary to make electromagnetic compatibility (EMC) calculations before frequencies are assigned. The bulk of technical and administrative work involved will require detailed knowledge not only of the equipment used but also of the physical characteristics of propagation over the whole frequency spectrum. The solution of the aforementioned problems necessitates the introduction of new spectrum management methods.

Spectrum management is the combination of administrative and technical procedures necessary to ensure the efficient operation of radiocommunication services without causing harmful interference.

The following describes national spectrum management procedures and then introduces "spectrum engineering techniques" that can be used to improve the efficiency of spectrum management.

2.2 *National spectrum management*

2.2.1 *Need for management of the use of frequencies at the national level*

When an administration ratifies the International Telecommunication Convention or accedes to it, it may enact domestic legislation to make the provisions of that Convention and of the Radio Regulations annexed to it applicable to its administration. Generally speaking, the frequency spectrum can be efficient only if it is properly managed:

- by imposing restrictions on certain technical characteristics that must not be *less strict* than those specified in the international agreements;
- by providing for the future well in advance through national planning of frequency usage.

The first condition necessitates the issuing of national regulations which should not be solely confined to the international provisions and may comprise everything which the legislative considers necessary to:

- enable each user to carry on a service under specified conditions;
- make sure that international obligations are fulfilled.

The second condition implies the existence of an administrative body to ensure coordination between the different users. This body must have the technical resources and administrative means to check whether the domestic legislation is being applied; in other words, to manage frequency spectrum usage in the national context.

2.2.2 *Role of a national spectrum management authority*

The national spectrum management authority will thus have several roles, which may be summarized as follows:

- standardization, which consists of effecting coordination between various users with a view to defining technical standards to be imposed on users and, if necessary, on equipment manufacturers, in the implementation of domestic legislation;
- planning, which consists of effecting coordination between various users with a view to defining the use to be made of the frequency bands in compliance with the Table of Frequency Allocations, and in planning future uses so that they may be included in this Table by an administrative conference competent to review it;
- management proper, consisting of granting authority for frequency usage in accordance with international regulations and domestic legislation, and ensuring that those frequencies are actually used in conformity with the terms of the authorization. This role also includes compliance with the obligations embodied in the Convention and the Radio Regulations with respect to other administrations (for instance, the handling of requests for coordination received from other administrations, etc.).

2.3 *Organization and operation of a national spectrum management authority*

Any national spectrum management body, whatever its structure, must have the necessary authority to assign frequencies to the various users and regulate their use on a rational basis at the national level. It should also be responsible for international coordination within the framework of the ITU Convention and the Radio Regulations. The authority and supervisory functions of this body may, therefore, include responsibility for ensuring that frequency assignments are notified to the IFRB in accordance with the provisions of the Radio Regulations in force.

Operational policies and procedures are the basis for administering the regulations discussed above. They detail the activities associated with licensing and enforcement and are used by the personnel entrusted with spectrum management in their day-to-day activity.

Policies and procedures should be constantly reviewed to ensure that social and technological changes are taken into account. Computerized data bases, containing technical and administrative data on existing radio systems, can provide support to this function. For example, statistical analyses can be used to create trend reports for spectrum planning purposes.

2.4 *Spectrum engineering techniques*

Efficient spectrum management can be accomplished only through a joint technical and administrative effort requiring the application of rapid record keeping, quantitative analysis and judgement based upon experience. In many cases the results of an analysis process serve as improved advisory information to the human decision-maker and are used to expand the understanding of the problem. It is only from the application of rapid spectrum management techniques that timely and complete solutions to the problem can be obtained.

The approach to spectrum engineering analysis should be broad with the maximum degree of flexibility in data and analytical procedures. The nature of some tasks dictates the need for data files and much of the approach is oriented with this in mind. Efficient use of large volumes of data logically requires high-speed digital computer capability and this application for both data and analysis is considered. However, the need for manual approaches to many problems is also recognized, and many of the data sources and techniques will also support such activity. In practice, a hybrid approach combining automated and manual techniques may be used for the solution to many problems.

3. Spectrum engineering models

3.1 *Introduction*

Spectrum management problems are analyzed with spectrum engineering models and appropriate data bases. An engineering model of a physical process has a number of advantages over operationally trying out various changes to equipment. Using a model is quicker and cheaper, since parameter changes in a model can be easily made without affecting the equipment. The accuracy to which models simulate a process is determined by the time, effort and data available to their design and use. The following model techniques are described in detail in the handbook.

3.2 *Model techniques*

3.2.1 *Analytical models*

Analytical model techniques involve the application of formulae that continuously describe some process. They invariably use common mathematical functions and often have other functions e.g., integration, convolution, etc., which have to be computed separately if they are not available in the computer's scientific software library. Numerical methods may also be used; for example, a process could be described by graphs for the range of parameters to be encountered and the graphs described by curve fitting procedures. Then, an interpolation between two graph points would give the same result (to an intended accuracy) as a set of equations and functions by using simple arithmetic rather than sophisticated mathematics.

3.2.2 *Empirical models*

These are similar to the numerical method above but the graphs are obtained from plots of observations or estimations. For example, the exact detail of the side lobes of a dish antenna is not of practical interest. An empirical Cartesian model would use a cosine curve to represent the main beam, and a series of straight lines or a simple curve could be used to describe the envelope of the side lobes. Such a model requires only the maximum gain of the main lobe to completely describe a pattern. In the extreme, an empirical model is one that produces the right result but is not itself founded on or derived from the wanted process (canonical).

3.2.3 *Statistical models*

A fundamental limitation of interference prediction and many frequency assignment models is that they are not statistical in character. With the exception of propagation calculations in a few models, statistical representation of parameters and relationships is not usually encountered (Report 656). However, many basic parameters of spectrum-using systems cannot be realistically described except in statistical terms. For example, the gain of a scanning antenna is not a fixed value, but varies with time and angle. A statistical description of the gain in terms of its amplitude probability density function may be essential to an adequate understanding of a problem involving this antenna. This type of data has not always been available, so that analysts have had nothing available other than estimates of the mean value of antenna gain to use in an analysis (Report 524).

3.3 *Element models*

An interference prediction process is intended to identify the level of unwanted energy that is coupled into a victim receiver from one or more interfering sources. For most spectrum utilization studies, antenna coupled interfering signals and not those penetrating the cabinet are of primary concern. Consideration of this coupling includes elements of wanted and unwanted transmitters, antennas, transmission medium, receivers, and information processing.

4. Spectrum management data

4.1 *Introduction*

In collecting data, an administration must consider the trade-off between the cost of the data acquisition and its requirements for data to support its spectrum management function. Administrations that require spectrum engineering models must be careful to ensure that the data necessary to use those models is available.

Spectrum management requirements vary widely from administration to administration and it is important that each administration develops the data bases necessary to meet its particular requirements. In determining those requirements, an administration should consider international agreements, multilateral agreements and national needs. In the interest of a speedy and economical application of electronic data processing (EDP) to spectrum management, it is advisable that an administration include only those data elements, data files, and data bases that are necessary for it to meet its spectrum management requirements. It is considered necessary to include a list of the data elements required in radio-frequency management. The development of different procedures for the collection and storage of the required data elements must be avoided. To facilitate the introduction of EDP in radio-frequency management, the data elements to be agreed upon should be included in future agreements and arrangements. Therefore, the definition, format and possible coding of the most important data elements should be internationally agreed upon as early as possible. An early agreement on the data elements will be in conformity with the work of the CCIR which tries, at international level, to agree upon the solution of problems at the earliest possible date. This also means that a list of the required data elements cannot be final, but has to be adaptable to new findings and requirements.

4.2 Hardware compatibility

To facilitate the transfer of data between administrations which must coordinate with each other, a standard method of transferring data is required. Currently the most acceptable method for transferring data is by use of 9-track magnetic tape. The compatibility to transfer data on a 9-track magnetic tape recorded at either 800 or 1600 bits per inch using odd parity and using either the ASCll or EBCDIC character set can satisfy requirements for data exchange. New, high-capacity four and eight millimeter digital tape cartridges and CD-ROM provide more efficient and cost effective alternatives to meet the several applications involving large volumes of data. Floppy disks (MS-DOS formatted, 5 1/4", 360 kB or 1.2 MB; 3 1/2", 760 kB or 1.44 MB) provide an effective medium for the exchange of programs and data files of small to moderate size. The technology for data storage is rapidly changing and new data storage techniques may be available in the future.

4.3 *Data groups*

In the interest of establishing a common approach to data acquisition, maintenance and retrieval, it is recommended that spectrum management data be grouped according to the following categories:

- frequency allocation data,
- frequency assignment and notification data,
- licence holder data,
- equipment characteristic data,
- planning parameter data,
- geographical data,
- frequency monitoring data.

The total number of data items proposed in the handbook to support these data groups is rather large. Administrations planning to set up data bases are invited to follow the guidelines referred to in the handbook. This does not mean, however, that all listed data elements should be included, if they are not needed. In most cases, the required data may be reduced to a limited number of basic data elements, depending on the particular requirements of an administration. The need for many data elements is largely influenced by the aim and purpose of the user. For example, the amount of data required to achieve meaningful and valid EMC computation grows with the state of congestion of the spectrum, which is related to the density of radiocommunication equipment in use in a country and thus to the infrastructure of this country.

5. Computer techniques

5.1 *Introduction*

Preceding sections have summarized the application of analysis techniques and data files used in management of the frequency spectrum. Automation can improve the implementation of analysis techniques and data bases. Computer systems offer the means to store data in readily retrievable form, to manipulate the data, to produce reports relating to all or part of the data and to perform analytical studies. Cost effective computer systems that process large amounts of data or perform complex analytical studies have been available for some time. Technological developments have reduced the cost of computer systems, and the application of computer techniques to spectrum management has become a feasible approach for many administrations. Where the amounts of data are large and the requirements for analytical studies are complex and diverse, computer techniques have become necessary.

As the costs of maintaining manual spectrum management systems increase because of growth in data volume, number of transactions, and number and complexity of analytical operations, the use of automated spectrum management systems becomes economically attractive. Computer hardware that includes the processing and data storage capability to offer substantial performance at reasonable cost is now available. The requirements of such hardware concerning installation, operating environment and operating knowledge are not critical. In addition to standard programming languages, it is now feasible to support, in terms of processing and data storage requirements, more advanced software packages that provide facilities suitable for management of the data files previously discussed.

To evaluate the application of computer systems to a specific spectrum management situation, the various types of computer hardware and software available must be analyzed. Techniques for the development and implementation of automated data processing systems have been extensively documented. The following paragraphs briefly describe some of the characteristics of hardware and software available.

5.2 *Hardware*

5.2.1 *Central processors*

Computer system hardware consists of several elements: central processing unit, main memory unit, auxiliary data storage devices, input devices, and output devices. The available equipment represents a wide range of capabilities, capacities, and costs. Three categories of computer systems are considered; they are microcomputers, minicomputers, and large-scale computers.

A comparison of *typical* central processors representative of the three categories is given in Table I.

TABLE I — *Comparison of representative central processors*

Categories of systems	Micro	Mini	Large-scale
Memory capacity (8 bit bytes)	4 000-64 000	32 000-512 000	32 000-16 000 000
Central processing unit add time (microseconds)	1-20	0.1-3.0	0.02-3.0
Data transfer rate (bytes/second)	120-1 000 000	20 000-16 000 000	20 000-100 000 000
Word length (bits)	8-16	8-16	32-64

5.3 *Software*

The sequence of instructions followed by the computer in performing a specific task is called a program. An integrated collection of programs that perform a specific activity is called a system. Programs and systems are classed as an operating system (or executive system) if their primary purpose is to control the processing of work done by the computer. Programs and systems are classed as an application system if their primary purpose is to process user data and produce specific outputs required by the user. Computer instructions, such as operating systems, utility routines and library routines, and application programming systems are called software. An application program that provides access to and management of an organized collection of data is called a data base management system (DBMS). In a more general sense, software also includes operating manuals and other documentation, user training and equipment servicing details.

5.3.1 *Data base management system*

To be useful, spectrum management data must be updated as changes occur, must be accessible by spectrum engineering models, and must be used to produce various reports. In automated systems the functions of updating, accessing and reporting data are performed by applications software. The applications software that provides these functions can be either an application related file management program or a data base management system (DBMS). A DBMS permits efficient management of large amounts of varied information and convenient manipulation of the information by users not familiar with highly technical file management procedures. The DBMS is a package of routines that includes modules used to define records and files of data, to build and maintain such files, to retrieve sets of data from the files, and to generate defined reports of the data. The files of data are stored so they can serve multiple applications; they can be linked together through key data items so unnecessary redundancy of stored data is eliminated. The organization and language syntax of such systems enables a user without comprehensive knowledge and without programming experience to have convenient access to the data. The ultimate objective of a DBMS is to make application development easier, cheaper, faster and more flexible.

Annex V to the Handbook contains generic data base management system characteristics which can be used as a guide in selecting a data base management system to be used to support automated spectrum management. Table AV-I contains generic data base management system characteristics for a microcomputer system. Table AV-II contains generic data base management system characteristics for a minicomputer system.

5.4 *Selecting a computer system*

Annex VI to the Handbook contains a description of micro and minicomputer characteristics for spectrum management.

The required power of the processor in a frequency management system is determined by several factors; the sizes and rates of transaction associated with data files, the complexity and rate of application of engineering models, and the desired response time for carrying out specified procedures. The volume factors (sizes and rates) are usually determined by an administration's geographical size and the state of development of telecommunications usage; the frequency manager must decide upon necessary response times in order to provide a satisfactory level of service to users and the administrative authority. The level of service must be provided within appropriate budgetary constraints. Although a powerful computer can process an amount of data or perform lengthy, complex calculations in a short period of time, a slower and less expensive computer may be able to process the required amount of data or perform the required calculations within time constraints acceptable to the manager. Increased processing time will also reduce the hardware and software costs associated with more complex data storage/access methods.

The frequency manager may be required to utilize an existing computer system serving other users, or he may be required to acquire a computer system for his specific task. In the first instance he will usually have access to a large general-purpose system. Those systems are capable of performing large-scale data processing and solving complex engineering problems; however, the frequency management application will be required to co-exist with the other applications on the computer system. This could cause constraints on available storage and calculation time. Where the computer system is required to satisfy only the spectrum management function, a minicomputer system offers several advantages. It can be dedicated to the single set of associated problems and placed at the location of its particular uses; the users can interact directly with the computer. Hardware can be obtained in smaller, specific amounts and types in order to satisfy the spectrum management function. The compact, rugged equipment does not require special operating environments so that installation can occur in common office facilities and the equipment can be easily maintained. All of these factors lead to a low overall cost and a task-efficient approach.

Computer usage usually expands and grows after an automated system has been implemented. An automated system is usually designed to alleviate a particular problem, and it is often found that the original problem had obscured other different problems. The incremental cost of solving these new problems is often small in comparison to the benefits which are obtained. A computer system should be designed to permit expansion due to additional applications being automated as well as to accommodate the normal growth expected from existing applications. A system should be designed with excess storage capacity, main memory and auxiliary storage, as well as with the ability to replace input/output devices with higher speed models or to add additional input/output devices without a major system overhaul. Where possible, central processors should be selected so as to allow increasing the processor capability through upgrading while maintaining software compatibility.

Annex VII to the Handbook contains a description of the application of a microcomputer to spectrum management.

6. Catalogue of spectrum management models

The handbook describes spectrum engineering models that enhance the efficient spectrum management of radiocommunication systems.

A list of computer programs for spectrum management has been published separately.

REPORT 836-2*

NECESSARY BANDWIDTH CALCULATIONS

(Study Programme 60B/1)

(1982-1986-1990)

1. Multi-channel FDM-FM emissions

Appendix 6 to the Radio Regulations (1985) "Additional Characteristics for the Classification of Emissions; Determination of Necessary Bandwidths Including Examples for their Calculation and Associated Examples for the Designation of Emissions", includes under its Part B, Table III-B, the necessary factors for use in computing peak frequency deviation of multi-channel frequency division multiplex/frequency modulation (FDM-FM) emissions. Peak frequency deviation is a critical factor in Carson's rule, $B_n = 2M + 2DK$, used to calculate necessary bandwidth for frequency spectrum allocation purposes. Table III-B is reproduced as Table I.

The factors 2.6, −1, and −15 in the Table are the average power (dBm0 (see Note)) values that were found in a standard, commercial telephone, switched public network circuit. The values were, in fact, based upon measurements of "talker volume" conducted in 1960 [McAdoo, 1963] that were previously agreed to in the CCIR and eventually at the World Administrative Radio Conference, Geneva, 1979, as applicable, for the purposes of necessary bandwidth calculation.

Note. – "dBm0" refers to the power, in decibels, relative to 1 mW referred to a point of zero relative transmission level.

In 1975 and 1976, further measurements of speech signal power were made in the same circuits and networks, using similar methodology, so as to allow direct comparison of results to the earlier work. The later measurements [Ahern *et al.*, 1978] have been under study, both in industry and government, since that time, finally leading to modifications [FCC, 1983] of the United States domestic telecommunications regulations.

To summarize the 1975-1976 study, which included the reasons why differences from the earlier study occurred, it was found that substantial changes have aggregated over time to produce an average decrease in actual average talker power level of approximately 4.6 dB. The changes have tended to increase the uniformity of service in the switched public networks from the viewpoint of speech volumes. These include a decrease in the proportion of toll grade battery systems, loss plan improvements, upgraded telephone sets, and an increase in direct trunking. Direct distance dialling has become normal practice with new loop and trunk design techniques. Furthermore, advanced digital data acquisition technology facilitated the measurement in 1975-1976 of speech signal power with greater precision than possible in 1960 when volume unit (VU (see Note)) meters were used for that survey. Standard deviation of all measurements between the 1960 and 1975-1976 surveys, was reduced by an average of about one third, from 7 to 4.6 volume units. A multi-stage statistical sampling scheme (near-end and far-end) of talker power measurements on more than 10 000 calls, originating from approximately 2500 loops, was used. Average conversational signal power (averaged over the entire observation interval), and a new measure of speech level known as equivalent peak level (EPL) [Brady, 1968] were the measures used to characterize talker signals. Loop d.c. current, class of service, switch type, and call destination were recorded as part of the 1975-1976 study.

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* This Report should be brought to the attention of Study Groups 4 and 9 and the CCITT with reference to its Recommendation G.223.

TABLE I – Multi-channel FDM-FM emissions

MULTIPLYING FACTORS FOR USE IN COMPUTING *D*, PEAK FREQUENCY DEVIATION, IN FM FREQUENCY DIVISION MULTIPLEX (FM/FDM) MULTI-CHANNEL EMISSIONS

For FM/FDM systems the necessary bandwidth is:

$$B_n = 2M + 2DK$$

The value of *D*, or peak frequency deviation, in these formulae for B_n is calculated by multiplying the rms value of per-channel deviation by the appropriate "Multiplying factor" shown below.

In the case where a continuity pilot of frequency f_p exists above the maximum modulation frequency *M*, the general formula becomes:

$$B_n = 2f_p + 2DK$$

In the case where the modulation index of the main carrier produced by the pilot is less than 0.25, and the rms frequency deviation of the main carrier produced by the pilot is less than or equal to 70% of the rms value of per-channel deviation, the general formula becomes either

$$B_n = 2f_p \quad \text{or} \quad B_n = 2M + 2DK$$

whichever is greater.

Number of telephone channels N_c	Multiplying factor [1]	
	(peak factor) × antilog	$\left[\frac{\text{value in dB above modulation reference level}}{20}\right]$
$3 < N_c < 12$	4.47 × antilog	$\left[\frac{\text{a value in dB specified by the equipment manufacturer or station licensee, subject to administration approval}}{20}\right]$
$12 \leq N_c < 60$	3.76 × antilog	$\left[\frac{2.6 + 2 \log N_c}{20}\right]$
$60 \leq N_c < 240$	3.76 × antilog	$\left[\frac{-1 + 4 \log N_c}{20}\right]$
$N_c \geq 240$	3.76 × antilog	$\left[\frac{-15 + 10 \log N_c}{20}\right]$

[1] In the above chart, the multipliers 3.76 and 4.47 correspond to peak factors of 11.5 dB and 13.0 dB, respectively.

Note. – Volume unit is one way of measuring speech level with a power-level indicator calibrated in terms of dB for a steady sine-wave voltage, with 1 mW in 600 Ω taken as a reference. The indicator is not frequency weighted in response. VU readings are averages achieved by a particular set of meter ballistic (mechanical) characteristics.

The later study indicated that with (1975 and 1976) switched public network systems, there was little dependence of speech signal power on call destination or originating class of service (residence, business, local, toll, and combined). Small differences are explained for the most part by loop characteristics. There is little, if any, variation in speech signal power attributed by study conclusions to psychological factors such as call distance or perception of received volume. The average values measured indicate that the switched telecommunications network in 1975-1976 was essentially transparent to customers in the sense that talker signal power was not found sensitive to call distance, local or toll call classification, or other factors that are outside local loop circumstances. In summary, it is believed that the later measurements of telephone "talker level" indicate the same, normal speech levels of the population when not using the telephone. It is concluded that little or no further change in representative talker level is expected in the future on the switched public telephone network where the measurements were taken. Consequently, the CCIR may wish to take this fact into account in an examination of the necessary bandwidth formulae for FDM-FM systems when telephone speech comprises a signficant part of the total FDM-FM circuit.

In the B_n formula as shown in Table I, the three factors 2.6, -1, and -15 are components of the multiplying factor used to determine D, or peak frequency deviation. Practical consequences of a reduction in "talker power" on the telephone circuit feeding an FDM-FM radio circuit is less peak deviation. There are three independent parameters that determine the peak deviation of the FM signal, all of which are constrained by the design of the system so as to limit the maximum value of each so that D does not exceed a predetermined value (in FDM-FM systems). Those variables are:

- r.m.s. value of per channel deviation,
- average power in a message channel,
- number of total channels in the multiplexed message load.

If the average power of speech signals can be reduced, as indicated in the 1975-1976 study, a trade-off would be possible with the two other parameters. This could be achieved by:

- increasing the number of channels in the same frequency bandwidth,
- increasing spectrum efficiency by reducing the bandwidth for the same number of channels,
- a combination of the two methods.

For example, in systems using 4 kHz voice grade message channels, it would be possible to vary the ratio of data to voice traffic. A user could choose to remain within a specified RF bandwidth and then select an average message channel power level which could be used to make a trade-off between an increase in the number of message channels against their individual frequency deviation. For a given ratio of data traffic to voice traffic, a user might consider that a too large increase in the number of 4 kHz message channels would reduce the per channel deviation to the point where signal quality would become degraded. However, an average message channel power level could be chosen which keeps within the specified RF bandwidth and which can permit some increase in the number of message channels while still allowing a proportional increase in the individual channel frequency deviations. By this method, instead of using all the available spectrum to maximize the number of message channels, it would be possible to obtain an increase in the value of the per channel deviation which would provide an improvement of the signal-to-noise ratio and a probable reduction of bit error ratio of the data traffic.

A microwave system can therefore be designed for which the parameter for the average message power level is chosen between the level used in the current equations and one 4.6 dB lower. There are valid reasons for changing (i.e. increasing) the other available parameters in the system when the average message power level (talker power) is decreased. However, a decision to carry out the change should be made by the carriers (or governments) concerned, taking account of the advantages and disadvantages of the possible trade-offs.

The important point for consideration within the CCIR is recognition of the changed average "talker power" level. To take account of the decrease in average "talker power" by 4.6 dB in speech telephone circuits, the United States of America has domestically replaced the 2.6, −1, and −15 values by the variable X which can then range between limits comprised of the present values, and corresponding values that are 4.6 dB lower, depending upon the total number of circuits in the FDM-FM system, and the composition of the system itself.

Thus, the necessary bandwidth, B_n is:

$$B_n = 2M + 2\left[d \times 3.76 \times \text{antilog}\left(\frac{X + Y \log N_c}{20}\right)\right] K$$

where:

M: maximum modulating frequency (Hz),

d: per channel deviation,

N_c: number of circuits in the multiplexed message load,

K: unity,

X: -2 to $+2.6$ for $12 \leqslant N_c \leqslant 60$ and for $Y = 2$,

X: -5.6 to -1.0 for $60 < N_c < 240$ and for $Y = 4$,

X: -19.6 to -15.0 for $N_c \geqslant 240$ and for $Y = 10$.

The term enclosed in brackets is the peak deviation, D. The numerator $(X + Y \log N_c)$ of the fraction represents the average power of the composite signal delivered to the modulator input of the transmitter. The value of 3.76, as noted in Table I, corresponds to a peak factor of 11.5 dB [Holbrook and Dixon, 1939; Smith, 1964].

To correctly select a value for X in the formula for B_n, it is helpful to summarize applicable conditions under which it is used in an FDM-FM system. Final choice within the 4.6 dB range may be empirical. It is clear from the study details compiled that average "talker power levels" of −2, −5.6, and −19.6 dBm0 should replace the corresponding numbers for $N_c \geqslant 12$ in the B_n formulation with an FDM-FM system that is used to provide trunk connections for modern commercial public telephone circuits where most of the FDM-FM channels are used for speech.

In smaller, private, or older FDM-FM systems, particularly those with $N_c < 12$ or those containing data (non voice) in most of the channels, the original values, as shown in Table I nominally apply. Typical multi-channel data circuits operate at power levels from −13 to −15 dBm0. Therefore, the composite loading limit will be determined by using the value $X = -13$ to -15 for systems with a large percentage of data circuits for $N \geqslant 240$. Individual channel signalling, as opposed to common channel signalling, is an indication that the −15 dBm0 level is applicable (systems with $N_c \geqslant 240$).

The choice over the range (4.6 dB) of signal power level is, as stated above, largely one of experience. The RF frequency band itself, is unrelated to choice of "talker power" parameters. It has been noted [FCC, 1983] that, as an example of practical application of the newer parameters, as many as 1800 telephone channels could be operated in the same terrestrial fixed point-to-point microwave radio-frequency bandwidth now designated for only 1500 circuits, a significant development in improved spectrum efficiency.

2. Unmodulated pulses emissions

The necessary bandwidth for the common cases of unmodulated pulses with either a trapezoidal or rectangular pulse shape are specified. A value of 20 dB below the peak of the theoretical envelope of the pulse spectrum is the criteria used to establish the necessary bandwidth for unmodulated pulses. The values for the envelope of the pulse spectrum of the unmodulated pulses were determined using simple calculation techniques [Mason and Zimmerman, 1960].

Table II lists the calculated necessary bandwidths for unmodulated pulses. In Case 1 (trapezoidal pulse) the pulse width, t, is the time between 50% amplitude points and t_r (rise time) is the time between 10% and 90% amplitude points. The fall time, t_f, is equal to the rise time t_r. In Case 2 the rise time t_r does not necessarily equal the fall time t_f. The fall time is the time between the 90% and 10% amplitude points. In Case 3 (rectangular pulse), t is the pulse width.

TABLE II – Unmodulated pulse emissions

Description of emission	Necessary bandwidth		Designation of emission
	Formula	Sample calculation	
Unmodulated pulse emission	Case 1: $B_n = \frac{1.79}{\sqrt{t \cdot t_r}}$ Hz at points 20 dB below the peak envelope value of the spectrum of a trapezoidal pulse	$t = 3 \times 10^{-6}$ s $t_r = 0.06675 \times 10^{-6}$ s Bandwidth: 4×10^6 Hz	4M00P0N
	Case 2: $B_n = 1.27 \sqrt{\frac{\frac{1}{t_r} + \frac{1}{t_f}}{t}}$ at points 20 dB below the peak envelope value of the spectrum of an unsymmetrical trapezoidal pulse	$t = 3 \times 10^{-6}$ s $t_r = 0.06675 \times 10^{-6}$ s $t_f = 0.167 \times 10^{-6}$ s Bandwidth: 3.36×10^{-6} Hz	3M36P0N
	Case 3: $B_n = \frac{6.36}{t}$ Hz at points 20 dB below the peak envelope value of the spectrum of a rectangular (ideal) pulse	$t = 1.41 \times 10^{-6}$ s $B_n = 4.5 \times 10^6$ Hz	4M50P0N

3. K Factors for Digital Modulation

The formulas for the necessary bandwidths of digital data modulations (Appendix 6 of the Radio Regulations) include a factor K which accounts for trade-offs which are made in designing a system. Trade-off decisions usually involve system power, bandwidth, and performance (bit error rate). For example, by utilizing higher levels of modulation, digital line of sight systems trade power for bandwidth. The higher level digital modulations, such as 16 QAM, 64 QAM, and 256 QAM, for a given amount of spectra can transfer more bits/seconds than the lower levels of modulation but require more power (i.e., higher carrier to noise). Conversely, in satellite systems where on-board power is limited, spectrum is often traded for power and the lower level modulations are utilized. Digital signals often require significant filtering to achieve adjacent channel protection requirements.

The filtering of the digital signals may take place as premodulation filtering of the baseband transmitter signal or at one or more other points within the transmitter or receiver. The compensation for the addition of this filtering is to increase the system C/N. The numerical values of the K factor depend upon the amount and shape of the filter selectivity characteristic. Filter characteristics are designed on a case-by-case basis. Thus, it is not practical or advantageous to include in the necessary bandwidth formulas single numerical values for K.

TABLE III shows some example values of K for several digital modulations. The K values are either for computed or measured spectra. TABLE III includes the necessary bandwidth formulas used, the K value and other parameters such as power, E_b/N_o, and bit error rate. The formula for the necessary bandwidth (B_n) of digital frequency modulation is from Appendix 6 of the Radio Regulations. R is the bit rate, S is the number of signaling states, and D is the frequency deviation. The formula for the necessary bandwidth of phase shift keying modulation [Cohen, 1984] is based on the principle that this bandwidth should be some multiple of the symbol rate.

TABLE III - Example K values for digital modulation under specified conditions

Modulation & Conditions	Necessary bandwidth formula	Example K value	P% fractional containment bandwidth
BPSK (unfiltered) S=2 (computed)	$B_n = \frac{2RK}{\log_2 S}$	10.28 2.0	99% 95%
BPSK (filtered, BER=10^{-3}) S=2 (computed)	$B_n = \frac{2RK}{\log_2 S}$	1.0 1) 0.75 2)	100% 100%
MSK (unfiltered) S=2 (computed) D=R/4	$B_n = \frac{R}{\log_2 S} + 2DK$	0.36 3.52	99% 99.9%
GMSK 3 dB premodulation Guassian filter bandwidth = 0.25R S=2 (computed) D=R/4	$B_n = \frac{R}{\log_2 S} + 2DK$	-0.28 0.18	99% 99.9%
Digital FM (Continuous Phase FSK) rectangular pulses S=2 (computed) D=0.35R	$B_n = \frac{R}{\log_2 S} + 2DK$	0.89	99%
16 QAM Microwave Digital S=16 (experimental, 90 Mbit/s)	$B_n = \frac{2RK}{\log_2 S}$	0.51 0.43	99% 95%

Notes: 1) For this case E_b/N_o = 7.5dB

2) For this case E_b/N_o = 9.3dB

The MSK, GMSK [Murota and Hirade, 1986 and Report 903], and digital FM modulations [Cohen, 1984] are examples of a class of modulations which are of constant amplitude and continuous phase. These continuous phase signals can be viewed as simultaneously phase and frequency modulation (Sundberg, 1986). Either of the necessary bandwidth formulas (frequency or phase) might be used for this class of modulation. Since the signals are typified by a modulation index (2D/R), the FSK formula for necessary bandwidth was used for these signals in Table III. Also, for QAM (Docherty and Pallemaerts, 1980) modulation, the PSK necessary bandwidth formula is used. The rationale is that the 16 QAM signal is made up of a statistical sum of 16 signal locations of PSK signals with three different signal amplitudes.

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REPORT 977-1

SPECTRA AND BANDWIDTH OF EMISSIONS

(Study Programme 60A/1)

(1986-1990)

1. Introduction

This Report deals with the results of studies performed by various administrations on spectra and bandwidth of emissions produced by transmitters in the immediate vicinity of the transmitting station.

The Report includes the results of measurements and calculations made to establish the curves showing the power or power density spectra for various classes of emission and provides, in certain cases, also the formulae for deriving these curves and for calculating bandwidth and out-of-band power.

For certain classes of emission, the Report also includes details concerning the test conditions to be used for measuring bandwidth and out-of-band radiation in accordance with Recommendation 328.

For the methods which may be used to measure these characteristics reference is made to Recommendation 328 and Report 324.

For methods of measurement appropriate to monitoring stations, see Recommendation 443 and Report 275, and for the measurement of x dB bandwidth, see Reports 324 and 420.

The Report comprises the following parts:

Part A – Amplitude-modulated emissions for telegraphy and data transmission

Part B – Angle-modulated emissions for telegraphy and data transmission

Part C – Frequency-modulated multi-channel emissions employing frequency division multiplex

Part D – Double-sideband amplitude-modulated emissions for telephony and sound broadcasting

Part E – Single-sideband and independent-sideband amplitude-modulated emissions for telephony and multi-channel voice-frequency telegraphy

Part F* – Frequency-modulated emissions for telephony and sound broadcasting

Part G* – Vestigial-sideband emissions for television broadcasting

* It should be noted that this part will be included in this Report when the necessary information is available.

PART A

AMPLITUDE-MODULATED EMISSIONS FOR TELEGRAPHY AND DATA TRANSMISSION

1. Classes of emission A1A and A1B

1.1 *Shape of the spectrum envelope*

The amplitude of the spectrum envelope relative to the amplitude of the continuous emission is shown in Fig. 1 (of Part B) as a function of the order of the sideband components, assuming that the envelope of the RF signal is a square wave [CCIR, 1958]. In this figure, the order n, of the sideband component is given by:

$$n = \frac{2f}{B}$$

where:

f: frequency separation from the centre of the spectrum (Hz),

B: modulation rate (Bd).

1.2 *Occupied bandwidth*

According to [CCIR, 1970-74] the occupied bandwidth, L, (Hz) for an out-of-band power ratio $\beta = 0.01$ may be calculated from the following empirical formula:

$$L = \left(\frac{1}{0.05 + \alpha} - 1\right) B \qquad (1)$$

where:

B: modulation rate (Bd),

α: relative build-up time of the shortest pulse of a telegraph signal as defined in § 1.21 of Recommendation 328,

β: out-of-band power ratio as given in § 1.12 of Recommendation 328.

The maximum divergence between the results obtained by using this formula and the results of accurate calculations is $2B$ when $\alpha < 0.02$; and B when $\alpha \geqslant 0.02$. This has also been confirmed by measurements. Equation (1) may therefore be used for the indirect measurement of occupied bandwidth of A1A and A1B emissions.

1.3 *Shaping of the telegraph signal by means of filters*

Increasing the build-up time of the telegraph signal to the maximum value compatible with the proper operation of the receiving equipment is a suitable means of reducing occupied bandwidth.

CCIR [1958] describes the spectra produced by signals with different build-up times; references 3, 4 and 8 of the document give details regarding the filters that may be used for shaping the signal.

The minimum value of the ratio, T, of the 6 dB passband of such filters to half the modulation rate in bauds, is largely dependent on the synchronization requirements of the receiver terminal equipment, the frequency stability of both the transmitter and receiver and, in the case of actual traffic, also on the propagation conditions. The minimum value may vary from 2, when synchronization and stability are extremely good, to 15 when the frequency drift is appreciable and teletype equipment is used.

Minimum overshoot filters should be preferably used in order to fully utilize the transmitter power.

The table below shows, as a function of T, the percentage of time during which the signal element is flat within 1% for a minimum overshoot filter:

$\frac{\text{Length of flat portion}}{\text{Length of signal element}}$	0% (sinusoidal signal)	50%	90%	100% (rectangular signal)
$T = \frac{\text{6 dB passband}}{\text{Half the modulation rate}}$	1.6	3.2	16	∞

Since the ratio T is predetermined, it may be necessary to use a filter consisting of several sections to sufficiently reduce the components in the outer parts of the spectrum.

1.4 *Adjacent-channel interference*

Interference to adjacent channels depends on a large number of parameters and its rigorous calculation is difficult. Since it is not necessary to calculate the values of interference with great precision, semi-empirical equations and graphs can be used, examples of which are given in [CCIR, 1958].

1.5 Results of measurements carried out by various methods

Document [CCIR, 1974-78] contains information on the results of comparative measurements of occupied bandwidths and x-dB bandwidths for class of emission A1B using various methods. The measurement methods are briefly described in Report 324-2, section 3.

The measurements conducted by various methods and their comparison with the calculation results showed a fairly good agreement.

REFERENCES

CCIR Documents

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[1970-74]: 1/42 (USSR).

[1974-78]: 1/56 (USSR).

PART B

ANGLE-MODULATED EMISSIONS FOR TELEGRAPHY AND DATA TRANSMISSION

1. Class of emission F1B

1.1 *Shape of the spectrum envelope*

The shape of the RF spectrum for class of emission F1B is described in § 1.1.1 to 1.1.3 below for various shapes of the telegraph signal.

1.1.1 *Telegraph signal consisting of reversals with zero build-up time*

The amplitude of the spectrum envelope relative to the amplitude of the continuous emission ($A(n)$) is shown in Fig. 1 (solid lines) as a function of the order of the sideband component for a telegraph signal consisting of reversals with zero build-up time and equal mark and space durations [CCIR, 1958a].

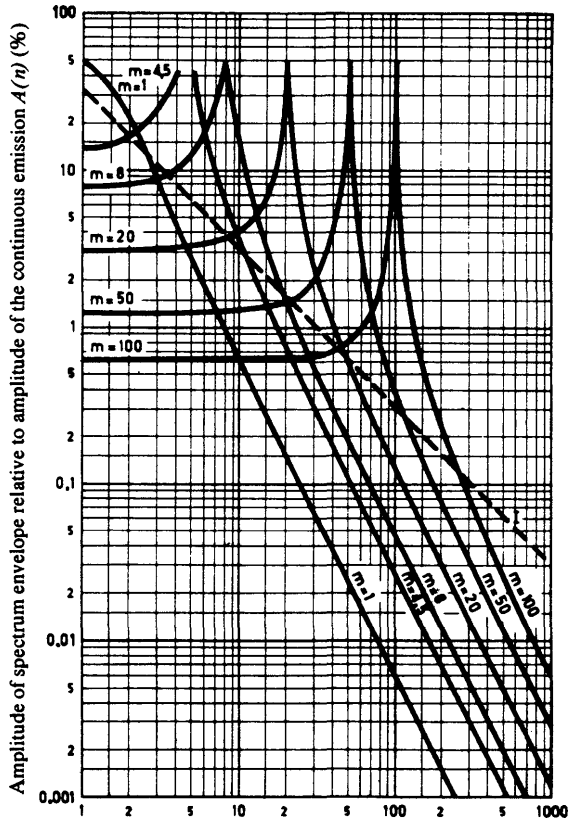


FIGURE 1 – *Envelopes of RF spectra for a telegraph signal consisting of reversals*

m: modulation index

—— class of emission F1B

– – – classes of emission A1A and A1B

The linear or asymptotic parts of the solid curves shown in Fig. 1 may be approximated with the aid of the formula:

$$A(n) = \frac{2m}{\pi n^2} \tag{2}$$

where:

n: order of the sideband component $= 2f/B$,

f: frequency separation from the centre of the spectrum (Hz),

B: modulation rate (Bd),

m: modulation index $= 2D/B$,

D: peak frequency deviation or half the frequency shift (Hz).

1.1.2 *Periodic telegraph signals with finite build-up time*

According to [CCIR, 1958b] the amplitude, $A(x)$ of the envelope of the spectrum produced by a telegraph signal consisting of reversals with a finite build-up time and equal mark and space durations is given by the following empirical formula:

$$A(x) = E\frac{2}{\pi}\frac{1}{m}x^{-u}(x^2 - 1)^{-1} \qquad \text{(for } x > 1\text{)} \tag{3}$$

where:

x: f/D,

E: amplitude of the continuous emission,

$u = \sqrt{5D\tau}$

τ: build-up time of signal (s) of the telegraph signal, as defined in Recommendation 328, § 1.20,

f, D and m: as defined in § 1.1.1 above.

Equation (2) shows that the shape of the spectrum envelope depends only on the product $D\tau$ and that for a given value of this product the amplitude, $A(x)$, of the envelope is inversely proportional to the modulation index m. See Fig. 2 where the product $mA(x)$ is shown as a function of x for various values of $D\tau$.

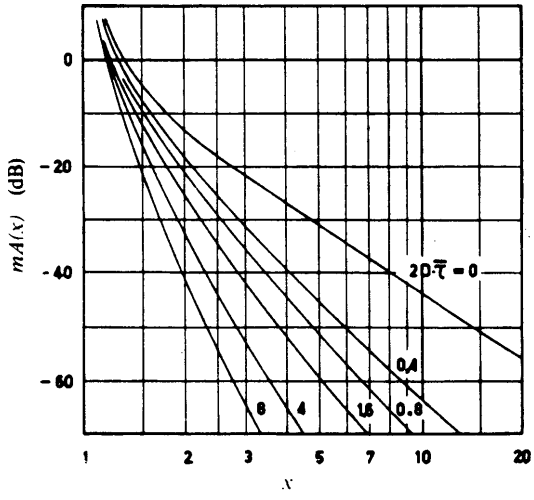


FIGURE 2 – *Spectrum distribution of F1B emission calculated from the empirical formula (3)*

It has been shown that the effect of the build-up time on the shape of the spectrum envelope is small for values of $D\tau$ which are less than 0.15 or are between 1 and 5. When the mark and space durations are unequal, the shape of the spectrum envelope depends largely on the product of $D\tau$ and the duration of the shortest signal element, but is always similar to that produced by a signal consisting of reversals with the same build-up time.

In Fig. 3 the results of measurements made on various spectra are compared with those obtained by calculating the corresponding values from equation (3). The agreement is satisfactory for values of x greater than 1.2, but decreases for decreasing values of the product $D\tau$.

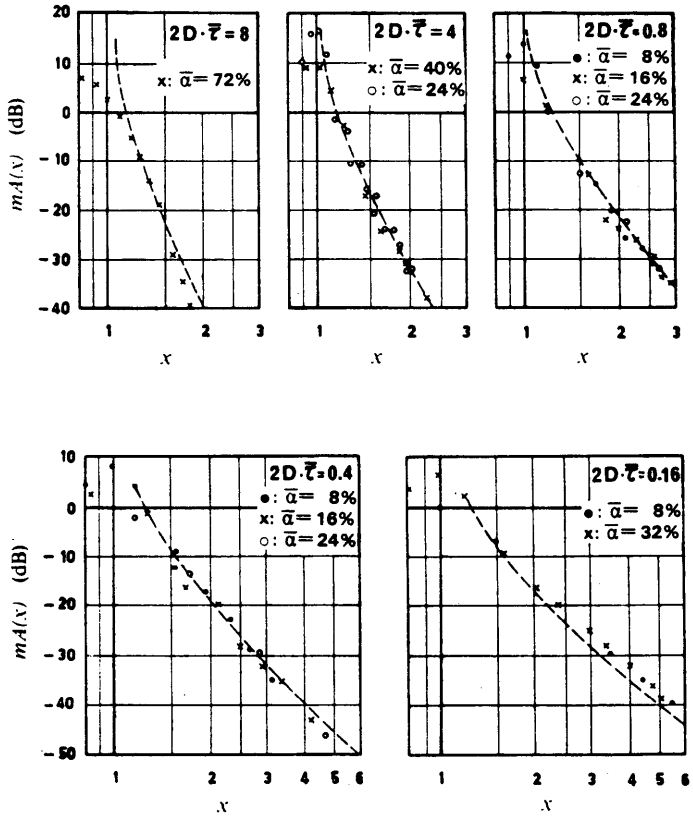


FIGURE 3 – *Spectra of F1B emissions*

– – – – calculated from empirical equation (3)
• o × measured values
ᾱ: relative build-up time (%)

1.1.3 *Non-periodic telegraph signal with finite build-up time*

When the signal is non-periodic, as may be the case under actual traffic conditions, the spectrum distribution should be represented in the form of a power density spectrum.

According to [CCIR, 1958b] the average power density per unit of bandwidth, $p(x)$, is given by the empirical formula:

$$p(x) = \frac{P_0}{B} \frac{4}{\pi^2} \frac{1}{m^2} x^{-2u} (x^2 - 1)^{-2} \quad (4)$$

where:

P_0 : total power of the emission,

B, m, x and u : as defined in § 1.1.1 and 1.1.2 above.

Also in this case, the shape of the spectrum envelope depends only on the product of frequency shift and build-up time.

1.2 *Out-of-band power and occupied bandwidth*

The out-of-band power, P^1, as defined in § 1.11 of Recommendation 328 may be determined by integrating the power density given by equation (4) between two frequency limits.

Figure 4 shows the values of bandwidth, L, calculated in terms of m and $2D\tau$, for $\beta = 0.01$ and $\beta = 0.001$, where β is the out-of-band power ratio $(P^1)/(P_0)$.

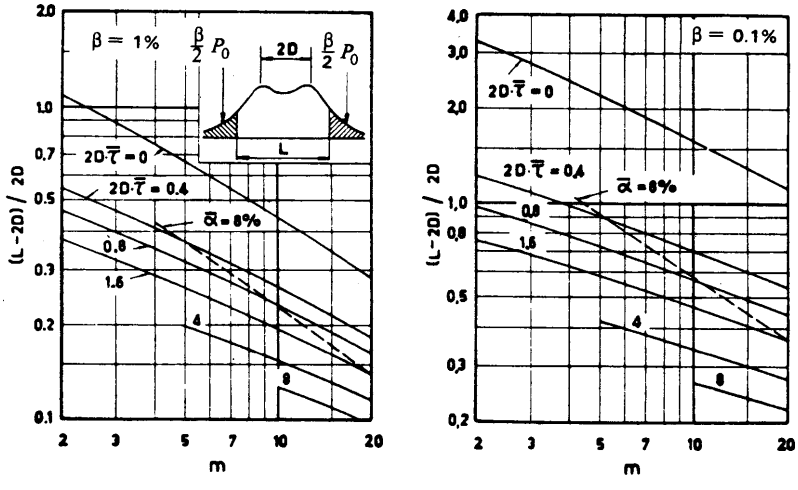


FIGURE 4 – *Bandwidth calculated from empirical equation (4)*

According to [CCIR, 1958b] the occupied bandwidth L (Hz) for $\beta = 0.01$ may also be calculated from the empirical equation:

$$L = 2D + D(3 - 4\sqrt{\alpha})\, m^{-0.6} \quad (5)$$

where α is the relative build-up time of the shortest pulse of the telegraph signal, as defined in § 1.21 of Recommendation 328.

The occupied bandwidth so calculated is hardly affected by the shape of the telegraph signal, whereas the out-of-band spectrum depends largely on this shape.

The maximum divergence between the results obtained by using equation (5) and those obtained by exact calculations, as summarized in [CCIR, 1951] is as follows:

$$\begin{aligned} 3\% \text{ for } \alpha &= 0 \quad ; 2 \leqslant m \leqslant 20 \\ 9\% \text{ for } \alpha &= 0.08; 1.4 \leqslant m \leqslant 20 \\ 10\% \text{ for } \alpha &= 0.24; 2 \leqslant m \leqslant 20 \end{aligned}$$

The list above shows the limits within which equation (5) can be used with reasonable accuracy. The percentages indicated apply to the lower limit of *m*. They are less for the higher limit.

Finally, Fig. 5 shows the results of calculations and measurements of occupied bandwidth employing different methods.

1.3 *Shaping of the telegraph signal by means of filters*

See Part A, § 1.3. However, the use of minimum overshoot filters is not essential, when the transmitter is required to operate at more than two frequencies, for example in the case of four-frequency diplex.

1.4 *Adjacent-channel interference*

See Part A, § 1.4.

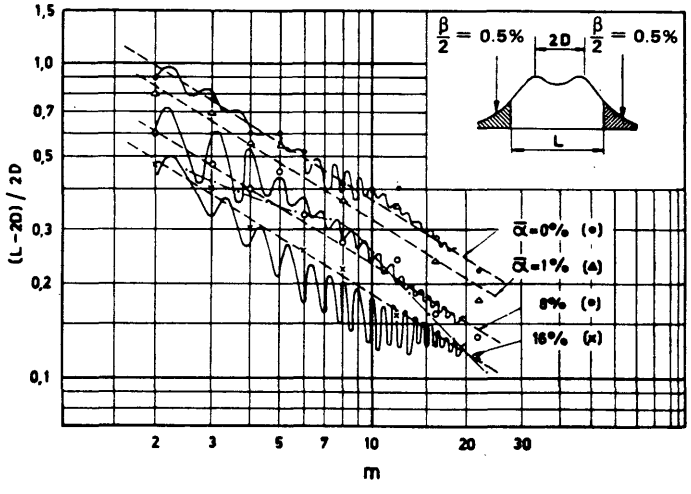


FIGURE 5 – *Comparison of the results of calculations and measurements of occupied bandwidth*

−·−·−·− calculated from the equation given in § 3.6 of Recommendation 328
− − − − calculated from equation (5)
——— calculated from spectra obtained by means of a spectrum analyzer
• Δ o × Values measured with the aid of the equipment described in § 4.2 of Recommendation 328
$\bar{\alpha}$: relative build-up time

1.5 Results of measurements carried out by various methods

Document [CCIR, 1974-78] contains information on the results of comparative measurements of occupied bandwidths and x-dB bandwidths for classes of emission F1B and F7BDX using various methods. The measurement methods are briefly described in Report 324-2, section 3.

The measurements conducted by various methods and their comparison with the calculation results showed a fairly good agreement.

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[1958] Geneva: **a.** I/31 (United States of America); **b.** I/23 (Japan).

[1974-78]: 1/56 (USSR).

PART C

FREQUENCY-MODULATED MULTI-CHANNEL EMISSIONS EMPLOYING FREQUENCY DIVISION MULTIPLEX

1. Introduction

The output signal of a frequency-modulated multi-channel transmitter using frequency division multiplex can be simulated by a signal which is frequency-modulated with white noise. This applies also to the output signal of a transmitter with a limited number of channels if band-splitting privacy devices are used in each of the channels.

It is difficult, however, to make a theoretical analysis of the spectrum of a signal which is frequency-modulated with white noise, unless the frequency deviation is either very large or very small, compared with the maximum frequency of the band-limited white noise.

However, emissions with modulation indices between the limits mentioned above are important in actual communication systems. CCIR [1966-69] reports on studies carried out to establish curves for the spectral distribution and the bandwidth of such signals.

2. Shape of the spectrum envelope

For larger values of the frequency deviation the envelope of the spectrum may be derived from the equation given by Medhurst [1956] (this equation is also mentioned in [CCIR, 1966-69]). Measurements have been carried out in order to find an empirical equation which could be applied for median values of the frequency deviation [Kurimura, 1959]. The equation mentioned in this last reference may be considered as an extension of the equation given by Medhurst [1956]. Curves derived from the empirical equation are shown in Fig. 6.

For small values of the frequency deviation, the distribution of the power density may be calculated from an equation given by Stewart [1954]. However, this equation is valid only for that part of the spectrum which lies within the frequency limits defined by plus and minus the maximum frequency of the noise signal.

Measurements have demonstrated that the spectra beyond these limits decay almost linearly. Therefore the slopes of the spectra were determined and used to complete the curves representing the spectral distribution (see Fig. 7).

It should be noted, however, that these slopes do not continue without limit. Because of the noise generated internally within the transmitter, the spectrum has a lower bound, or floor, the level of which depends upon the type of radio-frequency output stage.

3. Out-of-band power and bandwidth

Curves giving the out-of-band power of emissions with median values of frequency deviation are shown in Fig. 8. These curves have been derived from the empirical equation mentioned in the first paragraph of § 2.

Curves relating to emissions with a small frequency deviation are given in Fig. 9. This figure has been obtained from Fig. 7 by graphical integration.

Figure 10 has been obtained from Figs. 8 and 9 and shows the normalized bandwidth for different values of the out-of-band radiation. The irregularities in the vicinity of the points given by $D_p/M = 0.5$ may be attributed to the fact that Figs. 6 and 7 were deduced using different approaches, starting from the two extreme cases of modulation index.

Experimental data obtained by using the method of measuring the bandwidth as mentioned in § 4.2 of Recommendation 328 have been plotted in Fig. 10, and clearly demonstrate the validity of Figs. 6 to 10.

Symbols used in Figs. 6 to 10:

M: the maximum frequency of the band-limited white noise;

D_0: r.m.s. frequency deviation, i.e. the r.m.s. value of the difference between the instantaneous frequency and its arithmetic mean;

$D_p = D_0\sqrt{2}$, i.e. the peak frequency deviation when the white noise modulating signal is replaced by a sinusoidal signal having the same power;

f: frequency separation from the centre of the spectrum;

P_0: total power of the emission;

P^1: power outside the frequencies $-f$ and $+f$ in the spectrum, i.e. the out-of-band power;

β: out-of-band power ratio $(P^1)/(P_0)$, as mentioned in § 1.12 of Recommendation 328;

$p(f)$: power density of the spectrum at frequency f.

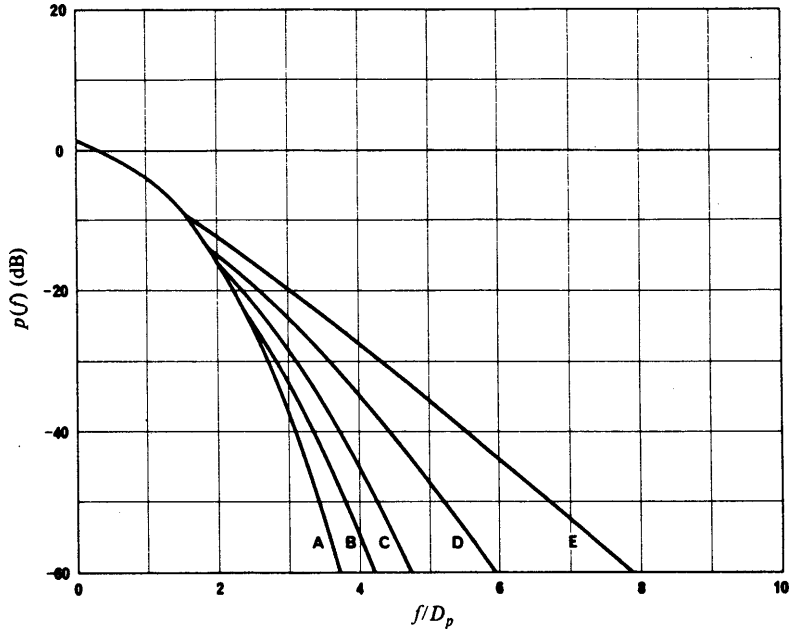


FIGURE 6 — *Power spectral distribution for* $\frac{D_p}{M} > 0.5$

0 dB = $P_0/2\ D_p$

Curve A: $D_p/M > 2.5$ Curve C: 1.0 Curve E: $D_p/M = 0.5$
Curve B: 1.8 Curve D: 0.7

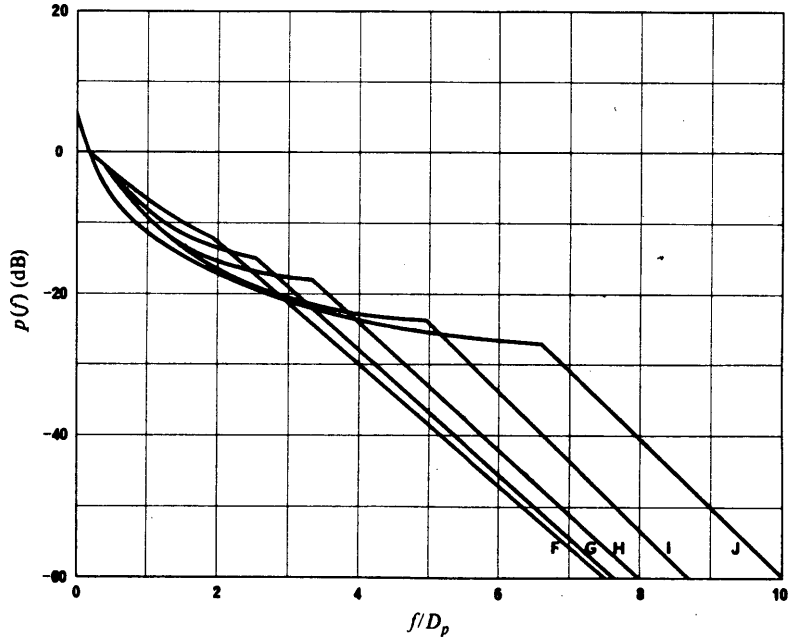


FIGURE 7 — *Power spectral distribution for* $\frac{D_p}{M} < 0.5$

0 dB = $P_0/2\ D_p$

Curve F: $D_p/M = 0.5$ Curve H: 0.3 Curve J: 0.15
Curve G: 0.4 Curve I: 0.2

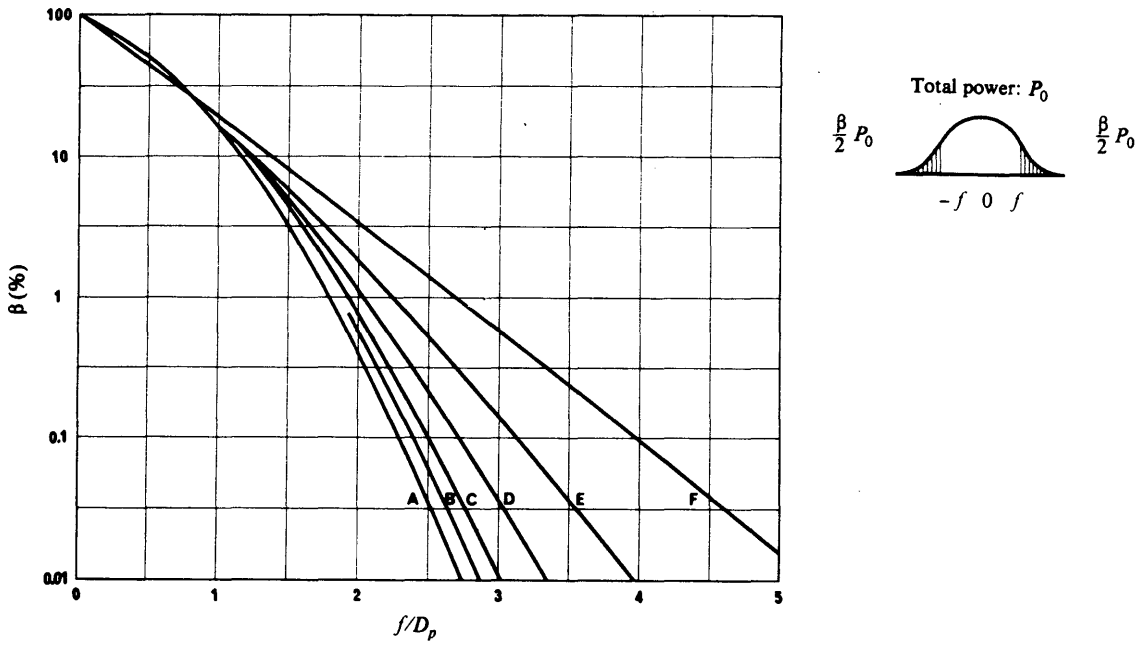


FIGURE 8— *Out-of-band power of the spectra for* $\frac{D_p}{M} > 0.5$

Curve A: $D_p/M > 2.5$ Curve C: 1.4 Curve E: 0.7
Curve B: 1.8 Curve D: 1.0 Curve F: 0.5

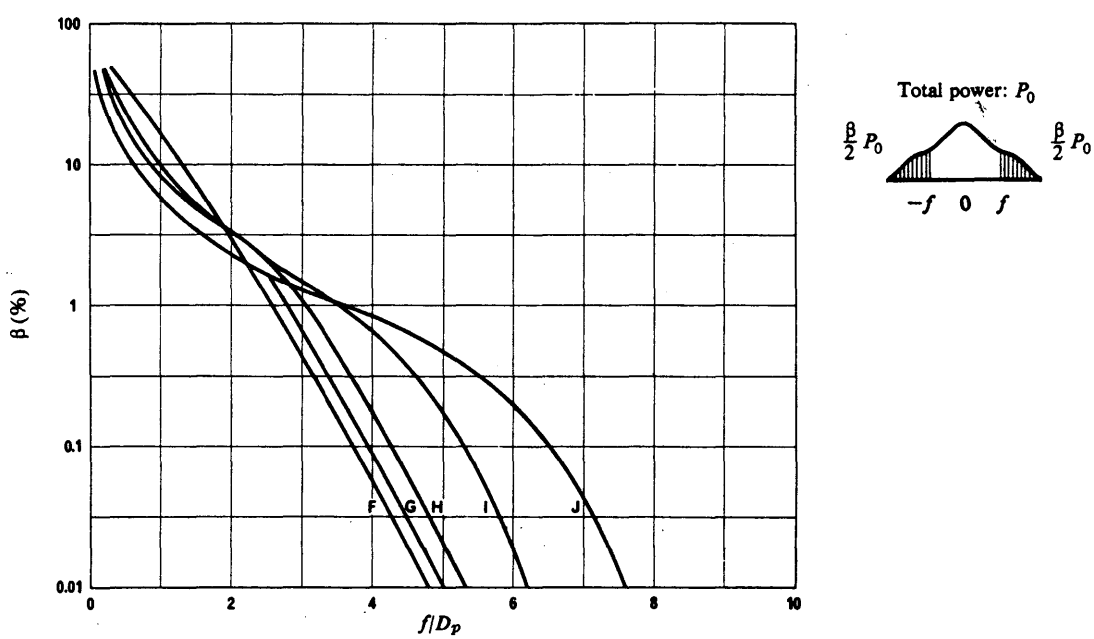


FIGURE 9 —*Out-of-band power of the spectra for* $\frac{D_p}{M} < 0.5$

Curve F: $D_p/M = 0.5$ Curve H: 0.3 Curve J: 0.15
Curve G: 0.4 Curve I: 0.2

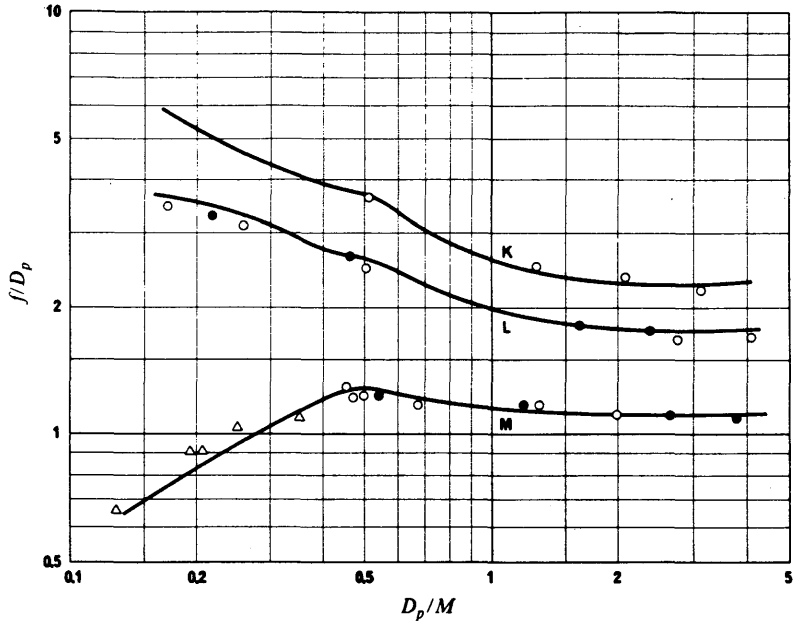


FIGURE 10 — *Bandwidth, in terms of D_p, for specific percentages of the out-of-band power*

Curve K: $\beta = 0.1\%$
Curve L: $\beta = 1\%$
Curve M: $\beta = 10\%$

●
○ } indicate the results of measurements
△

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PART D

DOUBLE-SIDEBAND AMPLITUDE-MODULATED EMISSIONS FOR TELEPHONY AND SOUND BROADCASTING

1. Introduction

The occupied bandwidth and out-of-band radiation of amplitude-modulated emissions carrying analogue signals depend, to a varying degree, on several factors such as:

- type of modulating signal;
- the input signal level determines the modulation loading of the transmitter;
- the passband which results from the filters used in the audio-frequency stages and in the intermediate and final modulating stages of the transmitter;
- the magnitude of the harmonic distortion and inter-modulation components at the frequencies of the out-of-band spectrum.

The results of measurements are also dependent upon the passband of the selective measuring device employed and on its dynamic characteristics, such as the integration time of the meter, or any other devices used in conjunction with the selective measuring device.

2. Class of emission A3E telephony

2.1 *Type of modulation signal and adjustment of the input signal level*

CCIR [1966-69a] deals with the adjustment of the level of the modulating signal at the transmitter input when the occupied bandwidth and the out-of-band radiation of a transmitter is to be determined with the aid of a modulating signal consisting of noise weighted according to the curve reproduced in Annex II to Recommendation 328.

As the statistical distribution of the noise amplitude is almost independent of bandwidth and is not significantly altered when a linear weighting network is used, the following procedure is suitable for simulating the loading of a transmitter under actual traffic conditions.

The transmitter is first modulated with a sinusoidal signal to a modulation factor of 100%. Next, the sinusoidal signal is replaced by a noise signal, the level of which is adjusted until the r.m.s. voltage after linear demodulation of the radio-frequency signal is equal to 35% of the r.m.s. voltage which was produced by the sinusoidal signal.

With this adjustment, which applies as well to a modulating signal consisting of white noise or of weighted noise, the envelope of the noise-modulated signal will not exceed the level corresponding to a modulation factor of 100% for more than about 0.01% of the time, according to the curve shown in Fig. 11.

The levels should preferably be measured at the output of the transmitter, as explained above, in order to avoid errors due to different values of the noise level is determined at the input or at the output of the band limiting filters used in the transmitter.

2.2 *Relationships between the 0 dB reference level for determining the out-of-band spectrum in accordance with Recommendation 328 and the levels of other spectral components of the emission*

2.2.1 *Relationship between the 0 dB reference level and the level corresponding to maximum spectral power density*

According to [CCIR, 1970-74a] the 0 dB reference level defined in § 3.4.1.3 of Recommendation 328 is about 5 dB below the level corresponding to the maximum power density in either sideband when the transmitter is modulated with white noise weighted in accordance with the curve mentioned in § 3.4.1.2 and shown in Annex II to the same Recommendation.

The value of 5 dB is valid for a modulation frequency bandwidth with an upper frequency limit of 3 kHz or 3.4 kHz.

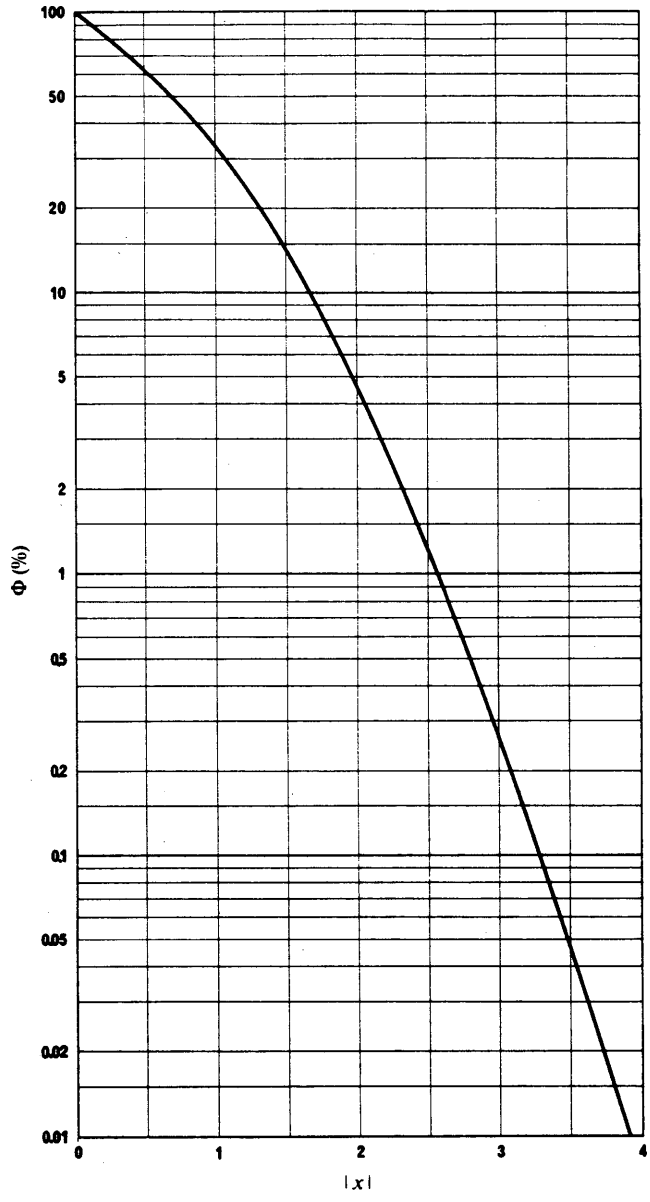


FIGURE 11 – *Time Φ (in %) during which the instantaneous value of the white noise exceeds the threshold voltage ± u, as a function of the ratio x*

x is given by $|x| = |u| / U_{rms}$

where:

U_{rms}: r.m.s. noise voltage

u: threshold level

2.2.2 *Relationship between the 0 dB reference level and the carrier level*

According to [CCIR, 1970-74b] the ratio α_B (dB) of the 0 dB reference level to the carrier level is given by the equation:

$$\alpha_B = 10 \log \left(\frac{m_{rms}^2}{2} \times \frac{B_{eff}}{F} \right) \quad (6)$$

where:

m_{rms}: r.m.s. modulation factor of the transmitter,

B_{eff}: effective noise bandwidth of the analyzer,

F: necessary bandwidth for the emission.

Hence the reference level depends on:

– the power of the sideband P_s, given by the formula:

$$P_s = \frac{m_{rms}^2}{2} \times P_c \quad (7)$$

where P_c is the carrier power,

– the necessary bandwidth F,

– the effective noise bandwidth B_{eff} of the analyzing instrument used.

Figure 12 shows the ratio α_B calculated from equation (6) as a function of the necessary bandwidth for different values of the r.m.s. modulation factor.

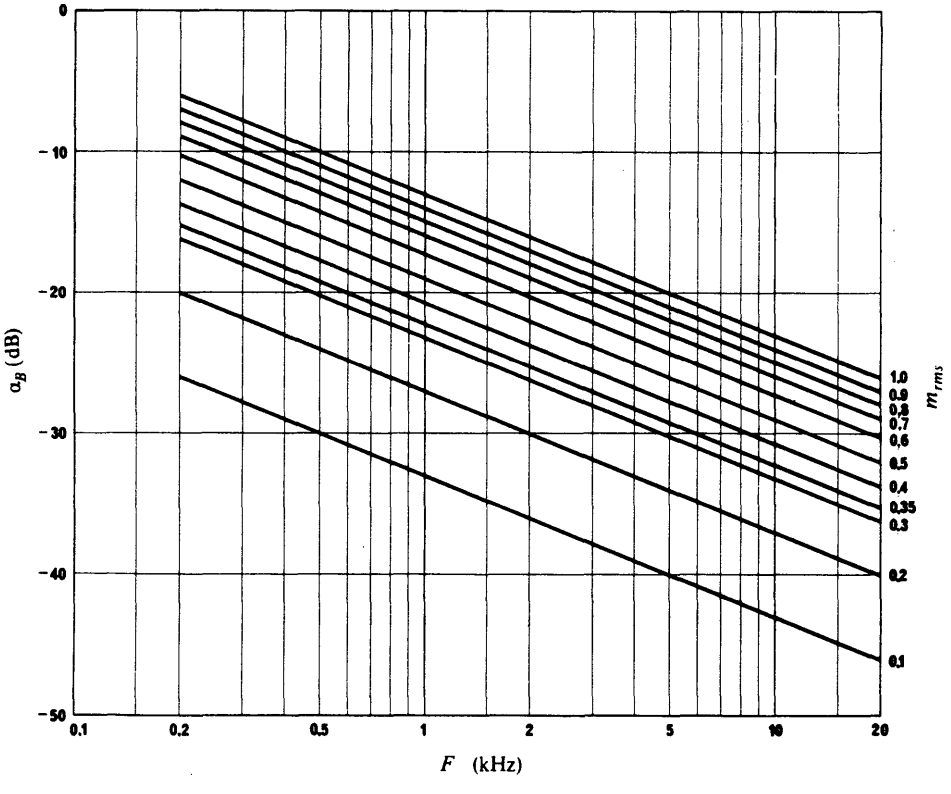


FIGURE 12 – *The ratio α_B (dB), between the 0 dB reference level for the limiting curve of the out-of-band spectrum for class of emission A3E and the level of the carrier, as a function of the necessary bandwidth F, for an effective noise bandwidth of the analyzing instrument of 100 Hz with the r.m.s. modulation factor (m_{rms}) of the transmitter as a parameter*

For certain practical applications, for example in monitoring stations, an r.m.s. modulation factor of the transmitter of 35% may be assumed in cases where the actual modulation factor cannot be determined precisely. Equation (6) may then be simplified as follows:

$$\alpha_B = 10 \log \left(\frac{B_{eff}}{F} \right) - 12.1 \qquad (8)$$

Figure 13 shows the ratio α_B calculated from the simplified formula (8) as a function of the necessary bandwidth for different values of the effective noise bandwidth.

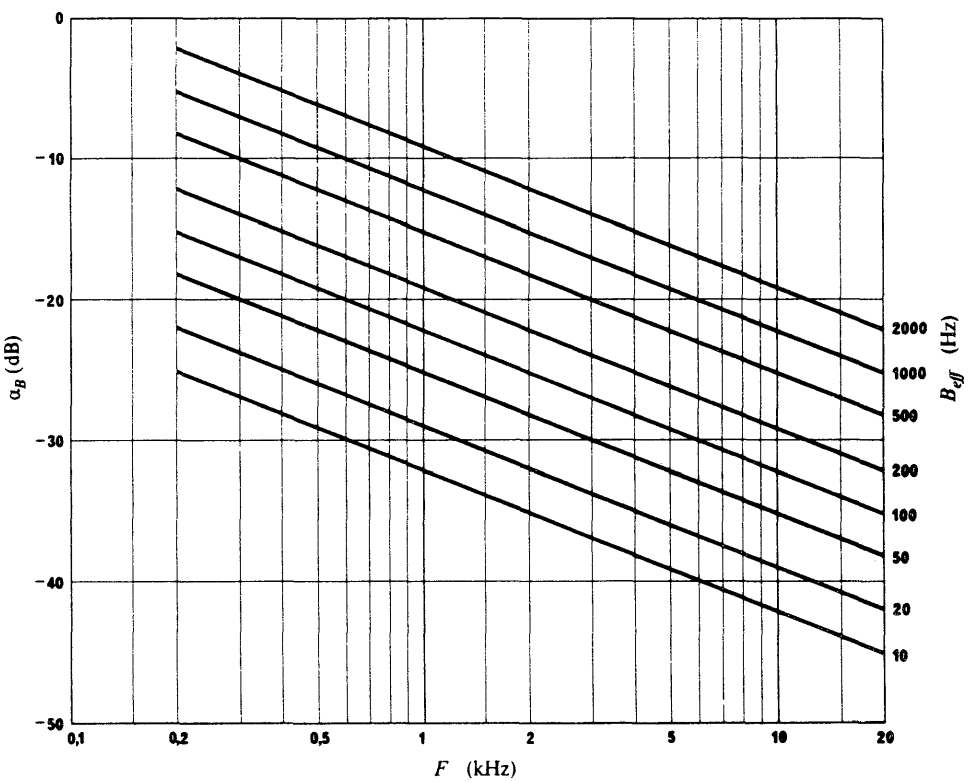


FIGURE 13 – *The ratio α_B (dB), between 0 dB reference level for the limiting curve of the out-of-band spectrum for class of emission A3E and the level of the carrier as a function of the necessary bandwidth F (kHz) for an r.m.s. modulation factor of 35% with the effective noise bandwidth (B_{eff}) of the analyzing instrument as a parameter*

3. Class of emission A3EGN sound broadcasting

3.1 *Type of modulating signal and adjustment of the input signal level*

According to [CCIR, 1966-69b] the adjustment procedure described in § 2.1 above may also be applied to transmitters for sound broadcasting, except that in this case, the noise is weighted in accordance with the curve mentioned in § 3.5.1.2 of Recommendation 328, i.e. Fig. 2 of Recommendation 559.

3.2 *Relationship between the 0 dB reference level for determining the out-of-band spectrum in accordance with Recommendation 328 and the levels of other spectral components of the emission*

3.2.1 *Relationship between the 0 dB reference level and the level corresponding to maximum spectral power density*

According to [CCIR, 1970-74a] the 0 dB reference level defined in § 3.5.1.3 of Recommendation 328 is 8-10 dB below the level corresponding to the maximum power density in either side band when the transmitter is modulated with white noise weighted in accordance with the curve mentioned in § 3.5.1.2 of the same Recommendation (i.e. Fig. 2 of Recommendation 559).

The value of 8 dB is valid for a modulation frequency bandwidth with an upper frequency limit of 4.5 kHz or 6 kHz. The value of 10 dB is applicable when the upper frequency limit is 10 kHz.

3.2.2 *Relationship between the 0 dB reference level and the carrier level*

See § 2.2.2 above, which is also applicable in the case of sound broadcasting.

4. General conclusions

The following conclusions may be drawn with respect to the test conditions to be used for measuring bandwidth and out-of-band radiation in accordance with Recommendation 328:

- a white noise signal weighted in accordance with the appropriate curve representing the statistical distribution of power in the band of modulation frequencies concerned is a satisfactory substitute for the modulating signal of class A3E emission for telephony and sound broadcasting;
- to prevent the transmitter from being overloaded, the noise at the input should be adjusted to a level such that the level corresponding to a modulation factor of 100% is not exceeded for more than a specific small percentage of the time:
- *transmitters for radiotelephony:* the level of the noise signal should be adjusted to obtain a specific modulation factor, determined by means of a true r.m.s. responding instrument. Results of measurements have shown that a value of 0.1%, for the percentage of time during which the level corresponding to 100% modulation was exceeded, could represent an acceptable value; however, further measurements should be made to confirm this assumption;
- *transmitters for sound broadcasting:* the level should be adjusted to obtain a modulation factor of 35%, determined by means of a true r.m.s. responding instrument; under these conditions the percentage of time is approximately 0.01%.

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PART E

SINGLE-SIDEBAND AND INDEPENDENT-SIDEBAND AMPLITUDE-MODULATED EMISSIONS FOR TELEPHONY AND MULTI-CHANNEL VOICE-FREQUENCY TELEGRAPHY

1. Introduction

See Part D, § 1.

2. Shape of the spectrum envelope for class J3E and class J7B emissions modulated with white noise

This section deals with the results of measurements made by several administrations on different designs of transmitters for classes of emission J3E and J7B.

The major characteristics of the transmitters and the test conditions relating to the measurements are summarized in Table I.

TABLE I – *Transmitter characteristics and measurement test conditions for J3E and J7B emissions*

Item No.	(1)	(2)	(3)	(4)	(5)
CCIR references	[1963-66a]	[1963-66b]	[1966-69a]	[1966-69b]	[1970-74a]
Class of emission	J3E	J3E	J3E	J3E; J7B	J3E
Transmitter characteristics:			Various transmitters		Various transmitters
– peak envelope power P_p (two tones) (1) (kW)	Different values	0.150	2.5-30		Several kilowatts to some tens of kilowatts
– third order intermodulation distortion α_3 (1) (dB)	Different values	About −40	Different values		
Type of modulating signal:	White noise	White noise	White noise	White noise Weighted noise	White noise
– bandwidth	Slightly smaller than B_p (2)	Limited only by B_p (2)	Limited only by B_p (2)		
Input signal level (1) adjusted to a value such that:					
– at the *input*, P_m (noise) =	P_m (two tones)		P_m (two tones)		
– at the *output*, P_m (noise) =				0.25 P_p (two noise)	
– at the *output*, P_p (noise) =		P_p (two tones)			
Type of mesuring device:		Spectrum analyzer	Spectrum analyzer		Spectrum analyzer
– passband (Hz)		300			⩽ 0.05 F (2)
Shape of spectrum	See Fig.15				See § 2.1.5

(1) In all tests, the transmitter is first modulated with two sinusoidal signals of equal amplitude. Next, the peak envelope power, P_p (two tones), and the third order intermodulation distortion level, α_3, are determined in accordance with the methods given in Recommendation 326. Finally, the two sinusoidal signals are replaced by a noise signal, the level of which is adjusted to obtain one of the conditions mentioned under "input signal level", where P_m denotes mean power and P_p denotes peak envelope power.

(2) B_p is the passband resulting from the filters in the transmitter, and F is the necessary bandwidth.

2.1 The results of the measurements may be summarized as follows:

2.1.1 *The tests described in item (1) of Table I:*

Assuming that the transmitter is operated under the conditions mentioned in item (1) of Table I and also assuming that the out-of-band radiation is mainly caused by intermodulation in the radio-frequency stages following the final modulator, the following may be concluded:

– the centre part of the radio-frequency spectrum exhibits substantially a rectangular form and is superimposed on a curve showing the out-of-band radiation which extends symmetrically with respect to the centre frequency (see Fig. 15);

- the difference α_N between the level of the flat portion of the top of the spectrum and the level at which the out-of-band radiation starts is generally equal to the level of the third order intermodulation component α_3 (see Fig. 16);
- the slope, expressed in dB per hertz, of the curve representing the out-of-band radiation, is inversely proportional to the bandwidth B of the noise signal at the input;
- the slope is constant, at least in the neighbourhood of the limits of the bandwidth, and has a value between 10 and 20 dB per bandwidth B, dependent on the character of the distortion (see Fig. 17);
- the bandwidth occupied by the emission is equal to the width of the main spectrum, provided that α_3 is at least 20 dB.

The above conclusions are expected to be also valid in those cases where the modulating signal is similar to white noise, such as radiotelephone emissions using a band-splitting privacy device and multi-channel voice-frequency radiotelegraph emissions.

2.1.2 *The tests described in item (2) of Table I*

The results, particularly with respect to the level at which the out-of-band radiation starts, correspond very closely to those obtained from the measurements described in item (1) of Table I and in item (1) of Table II.

2.1.3 *The tests described in item (3) of Table I*

CCIR [1966-69a] summarizes the results of measurements made on transmitters which, although of different design and power rating, used triodes in the final stage which were capable of being driven into grid current.

In one series of tests, the transmitters were fairly heavily loaded in order to determine the possible influence of grid current. Under this condition the third order intermodulation distortion level α_3 was rather poor and there appeared to be a fairly large difference between the value of α_3 and the level α_N in the power spectrum at which the out-of-band radiation starts.

In a second series of tests, α_N and α_3 were determined as a function of the modulation input level. For the lower values of this level the relation $\alpha_3 = \alpha_N$ was approximately satisfied.

Furthermore, it has been observed that under the modulating conditions mentioned in item (3) of Table I, the mean power of the noise-modulated radio-frequency signal was about 1 dB greater than the mean power of the radio-frequency signal modulated with two sinusoidal signals. This causes the peak envelope power to be exceeded for a considerable percentage of the time. This condition does not correspond to the practices generally adopted in actual traffic and further experiments seem to indicate that it might be necessary to adjust the level of the noise signal to a value which is 2-3 dB lower than that used in the tests just mentioned.

2.1.4 *The tests described in item (4) of Table I*

The adjustment of the input signal level mentioned in item (4) of Table I applies to both transmitters for class of emission J3E and transmitters for class of emission J7B. In this case, the following relationship is satisfied with respect to the power of the radio-frequency signal:

$$P_m \text{ (noise)} = 0.5\ P_m \text{ (two tones)} = 0.25\ P_p \text{ (two tones)} \qquad (9)$$

Under this condition the envelope of the noise-modulated signal will not exceed the level corresponding to the rated peak envelope power for more than about 2% of the time.

If, with a transmitter for class of emission J3E, the noise signal is weighted, the same adjustment can be used.

2.1.5 *The tests described in item (5) of Table I*

If frequency is plotted as the abcissa in logarithmic units, the reference frequency being assumed to coincide with the centre of the necessary bandwidth F, and if the power densities are plotted as ordinates, in decibels, the curves representing the out-of-band spectra produced by a number of transmitters of different power rating for class of emission J3E lie below two straight lines, starting at point ($+0.5F$, 0 dB), or at point ($-0.5F$, 0 dB), and finishing at point ($+0.6F$, -30 dB) or ($-0.6F$, -30 dB), respectively. Beyond the latter points and down to the level -60 dB, the curves lie below two straight lines, starting from the latter point and having a slope of 12 dB per octave.

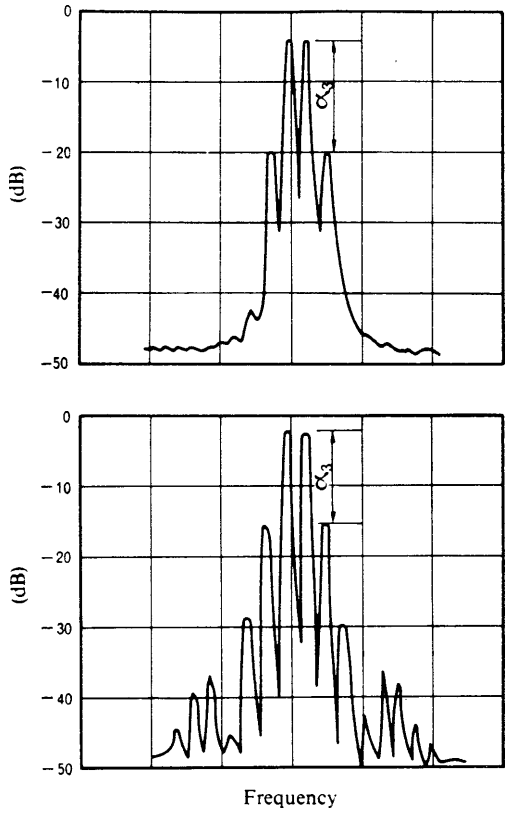


FIGURE 14 – *Spectrum envelope of class J3E emission modulated with two sinusoidal signals*

α_3: third order intermodulation level

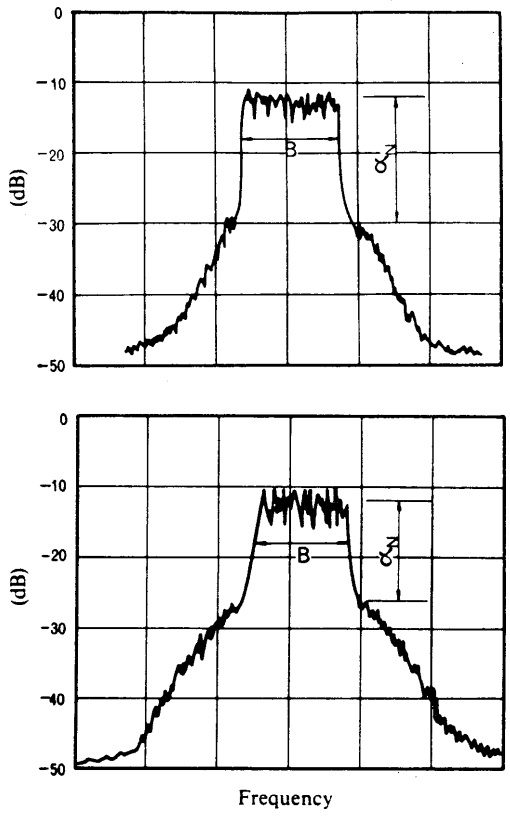


FIGURE 15 – *Spectrum envelope of class J3E emission modulated with white noise*

α_N: see text

B: bandwidth of the noise signal

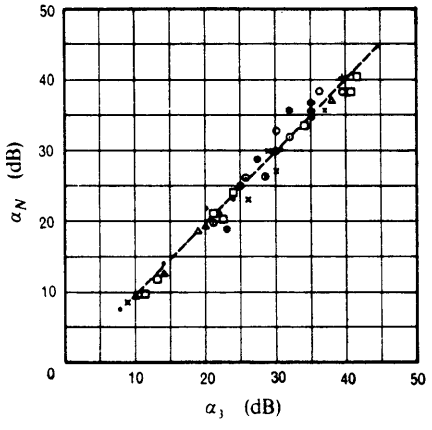


FIGURE 16 – *The value of α_N shown in Fig. 15 for different values of α_3*

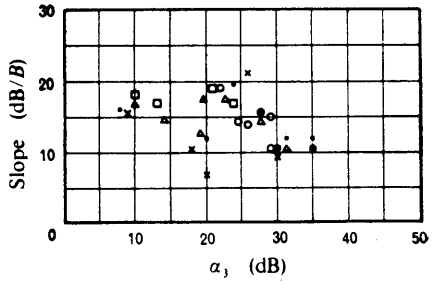


FIGURE 17 – *The value of the slope in the vicinity of the bandwidth for different values of α_3*

3. Shape of the spectrum envelope for class B8E and class R7J emissions modulated with white noise

This secton deals with the results of measurements made by several administrations on transmitters of different design for classes of emission B8E and R7J.

The major characteristics of the transmitters and the test condition relating to the measurements are summarized in Table II.

TABLE II – *Transmitter characteristics and measurement test conditions for B8E and R7J emissions*

Item N°	(1)	(2)	(3)
CCIR references	[1963-66c]	[1970-74a]	[1970-74b]
Class of emission	B8E	B8E	B8E; R7J
Transmitter characteristics:		Various transmitters	Various transmitters
– peak envelope power P_p (two tones) (1) (kW)	20	Several kilowatts up to some tens of kilowatts	Different values
– third order intermodulation distortion α_3 (1) (dB)	$\leqslant -35$		
– number of channels active during the measurement	2, in lower sideband	2 and 4	
– bandwidth of speech channel (Hz)	3000		
– carrier suppression (dB) relative to peak envelope power	−50		
Type of modulating signal:	White noise	White noise	White noise
– bandwidth	30 Hz-20 kHz ± 1dB		100 Hz-6 kHz per sideband
Input signal level (1) adjusted to a value such that:			
– at the *output*, P_m (noise) =	0.25 P_p (two tones)		0.25 P_p (two tones)
Type of measuring device:	true r.m.s. selective measurement device	Spectrum analyzer	Spectrum analyzer
– passband (Hz)	Curve C: 3800 Curve D: 100	$\leqslant 0.05\ F$ (2)	
Shape of spectrum	See Fig. 18	See § 3.1.2	

(1) In all tests, the transmitter is first modulated with two sinusoidal signals of equal amplitude. Next, the peak envelope power, P_p (two tones), and the third order intermodulation distortion level, α_3, are determined in accordance with the methods given in Recommendation 326. Finally, the two sinusoidal signals are replaced by a noise signal, the level of which is adjusted to obtain one of the conditions mentioned under "input signal level", where P_m denotes mean power and P_p denotes peak envelope power.

(2) B_p is the passband resulting from the filters in the transmitter, and F is the necessary bandwidth.

3.1 The results of the measurements may be summarized as follows:

3.1.1 *The tests described in item (1) of Table II*

Only the lower sideband was used, the upper sideband being suppressed to at least −60 dB by means of the filter incorporated in the transmitter. The carrier was suppressed to approximately −50 dB (class J3E) and the audio-frequency bandwidth was approximately 6000 Hz.

The bandwidth of the noise signal, was limited only by the filter characteristic of the transmitter; see curve A of Fig. 18. In this connection it should be noted that, if the radio-frequency spectrum produced by only one speech channel were to be determined, the bandwidth of the test signal should be limited before it is applied to the transmitter, since its overall bandwidth is considerably larger than the width of one speech channel.

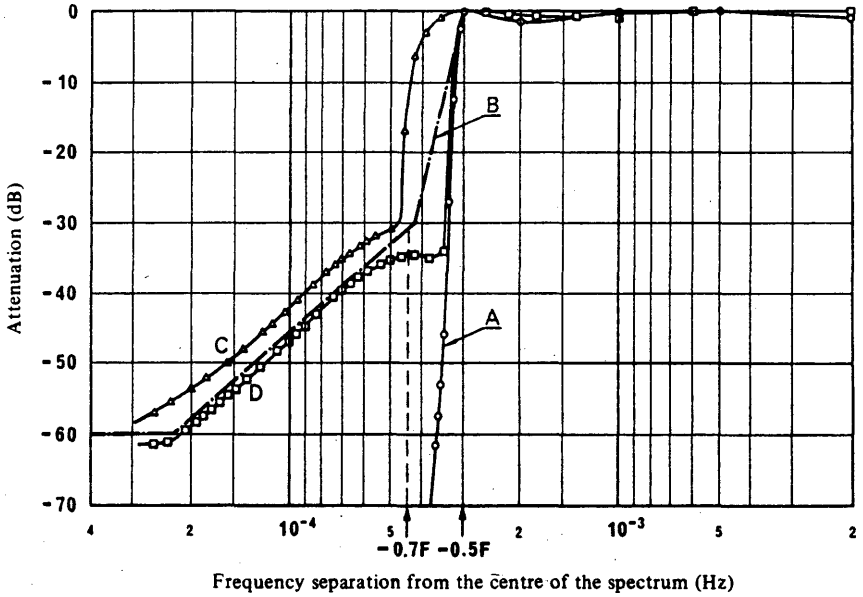


FIGURE 18 – *Envelope of power density spectrum for class of emission B8E*

F: necessary bandwidth (6000 Hz)

Curves A: filter characteristic of the transmitter
B: limiting curve specified in § 3.4.2.3 of Recommendation 328
C: measured with an analyzer having a passband of 3800 Hz
D: measured with an analyzer having a passband of about 100 Hz

One series of measurements was carried out using an analyzer with a bandwidth of about 100 Hz. An analyzer with a bandwidth of 3.8 kHz and a very steep attenuation slope was employed for the other series.

The results are shown in Fig. 18 curves D and C respectively. These curves represent the envelopes of the spectra of the lower sideband, measured in the lower radio-frequency range. Curves similar to those given in Fig. 18 were obtained for the higher frequency range.

If the spectrum measured with the aid of a narrowband equipment is, as in the present case, just within the limiting curve B, the spectrum analyzed by means of wideband receivers will exceed this limit. As wideband measuring equipment does not take account of the fine structure of the spectrum, particularly in the region where its slope is steep, the use of narrow-band devices for such measurements is recommended.

It can be further concluded from Fig. 18 that the out-of-band radiation starts at a level nearly equal to the level of third order intermodulation components, viz. at −35 dB. The out-of-band radiation remains almost constant in the immediate vicinity of the limits of the bandwidth; for frequencies remote from these limits the curve gradually decays, at first proportional to frequency, then reaching an ultimate slope of about 12 dB per octave. In Fig. 19 a linear frequency scale has been used at the abscissa, to illustrate more clearly the envelope of the spectrum mentioned above.

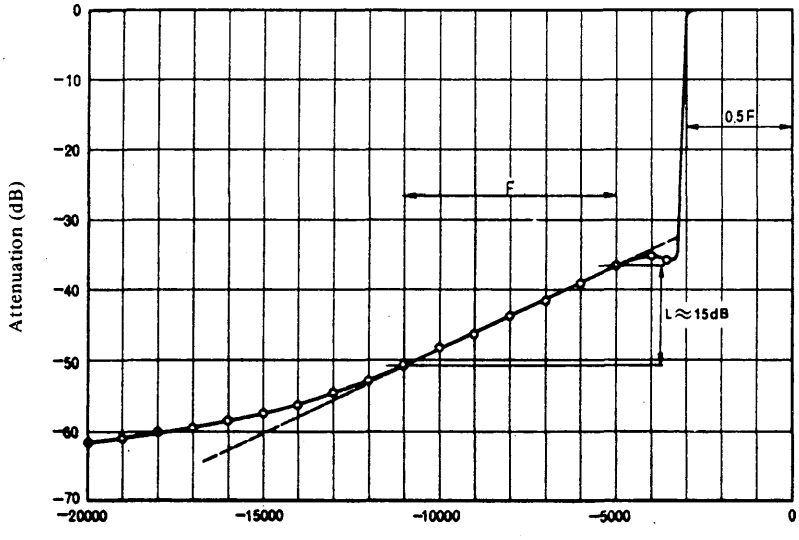


FIGURE 19 – *Curve D of Fig. 18 shown on a linear frequency scale*

L: linear region equal to about the necessary bandwidth *F*

3.1.2 *The tests described in item (2) of Table II*

If frequency is plotted as the abscissa in logarithmic units, the reference frequency being assumed to coincide with the centre of the necessary bandwidth *F*, and if the power densities are plotted as ordinates, in decibels, the curves representing the out-of-band spectra produced by a number of transmitters of different power rating for class of emission B8E (eight two channels or four channels simultaneously active) lie below two straight lines starting at point ($+0.5F$, 0 dB), or at point ($-0.5F$, 0 dB), and finishing at point ($+0.55F$, −30 dB) or ($-0.55F$, −30 dB), respectively. Beyond the latter points and down to the level −60 dB, the curves lie below two straight lines starting from the latter points and having a slope of 12 dB per octave.

Further measurements of out-of-band spectra in B8E and J3E emissions should be carried out with narrow-band analyzers by other administrations to obtain a more accurate shape of the corresponding limiting curves of out-of-band spectra specified in Recommendation 328 for the mentioned classes of emissions.

3.1.3 *The tests described in item (3) of Table II*

The test equipment was arranged to facilitate intermodulation distortion measurements to be made either by the two-tone method or the white-noise method, so that comparisons could be made between the two methods. When using the white-noise method, the white noise generator output was passed through filters to limit the noise bandwidth to the maximum bandwidth normally expected on traffic i.e. 100-6000 Hz per sideband. A band stop filter provided a slot in which "in-band" distortion products could be measured using a 30 Hz filter in the spectrum analyzer. A band-stop filter with a minimum bandwidth of 500 Hz at 3 dB and a 60 dB shape factor of 3.5 to 1 was found necessary to permit adequate resolution by the 30 Hz spectrum analyzer filter when measuring distortion ratios approaching 50 dB.

The majority of the white-noise loading tests were made with a mean output power level of −6 dB relative to peak envelope power rating which confirms the relationship mentioned in § 2.1.4, equation (9), of this part of the Report.

The tests confirm and extend the earlier conclusions and establish the use of a white-noise signal as a valid substitute for the modulating signal of two types of multiplex emissions, B8E and R7E, in common use. Further, the tests disclose a useful and stable experimental relationship between in-band intermodulation distortion and out-of-band radiation. However, there was no clear agreement between two-tone intermodulation distortion ratios and equivalent white-noise loading distortion.

Finally, [CCIR, 1970-74b] discusses factors affecting transmitter loading when using Lincompex and recommends loading levels for 2, 3 and 4 channel operation in which the limiting criteria is the out-of-band radiation (see Recommendation 455).

4. General conclusions

The following conclusions may be drawn with respect to the test conditions to be used for measuring bandwidth and out-of-band radiation in accordance with Recommendation 328:

- a white-noise signal, limited to a bandwidth somewhat larger than the bandwidth resulting from the filters used in the transmitter, is a satisfactory substitute for the modulating signal of single channel and multi-channel telephony emissions using a band-splitting privacy device, multi-channel voice-frequency telegraphy emissions and multiplex emissions carrying a very large number of speech channels;
- a white-noise signal weighted in accordance with the curve representing the statistical distribution of power in the band of speech frequencies concerned is a satisfactory substitute for the modulating signal of single channel emissions and multi-channel emissions with a limited number of speech channels, not employing any privacy device;
- the level of the noise signal at the input should be so adjusted that the mean power of the noise-modulated radio-frequency signal is 6 dB below the peak envelope power of the transmitter. Under this condition, the peak envelope power is only exceeded for about 1% to 2% of the time, which approximately corresponds with the values frequently encountered in actual traffic;
- a wideband measuring equipment does not take account of the fine structure of the spectrum, particularly in the region where its slope in steep, narrowband measuring devices with a passband of the order of 0.05 times the necessary bandwidth should be used for these measurements.

With regard to the shape of the power density spectrum of single-sideband emissions with suppressed carrier and of independent-sideband emissions, both modulated with white noise, it may be concluded that:

- the steep part of the spectrum just outside the limits of the necessary bandwidth follows very closely the attenuation curve of the filters used in the transmitter;
- the difference between the flat portion of the centre part of the spectrum and the level at which the out-of-band radiation starts is approximately equal to the level of the third order intermodulation component;
- over a small frequency interval in the immediate vicinity of the limits of the necessary bandwidth, the level of the out-of-band radiation may be constant;
- at frequencies remote from the limits of the necessary bandwidth the spectrum decays, at first directly proportional to frequency, then reaching an ultimate slope of about 12 dB per octave.

REFERENCES

CCIR Documents

[1963-66]: **a.** I/49 (Japan); **b.** I/6 (United States of America); **c.** I/43 (Federal Republic of Germany).

[1966-69]: **a.** I/12 (the Netherlands); **b.** I/15 (Federal Republic of Germany).

[1970-74]: **a.** 1/41 (USSR); **b.** 1/4 (United Kingdom).

REPORT 972

PEAK POWER RESPONSES TO INTERMITTENT INTERFERENCE SIGNALS

(Question 44/1)

(1986)

1. Introduction

Report 654 describes basic measures that quantify the interactive effects between an interferer and a receiver for various frequency distance separations. One of the measures is frequency dependent rejection (*FDR*), which is used in equation (1) of Report 654 to calculate *I*, the average interference power in the receiver bandwidth. The average power is of interest when the interference produces a response that is continuous in time. When the response is intermittent, however, the peak interference power $\hat{I}$, rather that the average power, and the duty cycle of the response waveform produced by the interfering signal, must be determined to quantify the interactive effects.

This Report presents a modification of equation (1) in Report 654 for estimating the peak interference power and the duty cycle of the response waveform.

2. Definitions

Equation (1) in Report 654 is modified as follows to express the peak interference power (dBm) in the receiver bandwidth:

$$\hat{I} = \hat{P}_t + G_t + G_r - L_b(d) - FDR\,(\Delta f) + \eta_i - \eta_0\,(\Delta f) \qquad (1)$$

where $\hat{P}_t$ is the peak power (dBm) of the interferer transmitter, and G_t, G_r, $L_b(d)$, $FDR\ (\Delta f)$, and Δf are defined in Report 654. The terms η_i, the duty cycle of interfering waveform at the receiver input (dB), and $\eta_0\ (\Delta_f)$, the duty cycle of the response waveform produced by the interfering signal in the receiver bandwidth (dB), are defined as follows:

$$\eta_i \triangleq \overline{P}_i - \hat{P}_i \tag{2}$$

$$\eta_0\ (\Delta f) \triangleq \overline{P}_0\ (\Delta_f) - \hat{P}_0\ (\Delta_f) \tag{3}$$

where:

$\overline{P}_i$: average power of the interfering waveform at the receiver input (dBm);

$\hat{P}_i$: peak power of the interfering waveform at the receiver input (dBm);

$\overline{P}_0\ (\Delta f)$: average power of the receiver response waveform produced by the interferer (dBm);

$\hat{P}_0\ (\Delta f)$: peak power of the receiver response waveform produced by the interferer (dBm).

The duty-cycle terms, which depend on the type of modulation used by the interferer and the bandwidth conditions, can be obtained with the expressions listed in Tables I to IV. The various types of input and output waveforms associated with the conditions listed in the tables are shown in Figs. 1 to 4.

TABLE I – *Duty-cycle functions for interfering pulse train*
(see Fig. 1)

$\eta_i = 10 \log (\tau \cdot PRF)$		
Bandwidth conditions	$\eta_0\ (\Delta f)$ (dB)	Figure No.
$B_r\tau \leqslant 1,\ \lvert\Delta f\rvert \geqslant 0$	$10 \log (PRF/B_r)$	1c)
$B_r\tau > 1,\ \lvert\Delta f\rvert \leqslant B_r/2$	$10 \log (\tau \cdot PRF)$	1b)
$B_r\tau > 1,\ \lvert\Delta f\rvert > B_r/2$	$10 \log (\tau \cdot PRF) - \lambda_r\ (\Delta f)\ 10 \log (B_r\tau/2)$	1d)

TABLE II – *Duty-cycle functions for interfering signals from chirp radars*
(see Fig. 2)

$\eta_i = 10 \log (\tau \cdot PRF)$		
Bandwidth conditions	$\eta_0\ (\Delta f)$ (dB)	Figure No.
$B_r/B_c > 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert \leqslant B_r/2$	$10 \log (\tau \cdot PRF)$	2b)
$B_r/B_c > 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert > B_r/2$	$10 \log (\tau \cdot PRF) - \lambda_r\ (\Delta f)\ 10 \log (B_r\tau/2)$	2e)
$B_r/B_c \leqslant 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert < B_c/2$ and $B_r^2\tau/B_c > 1$	$10 \log (\tau \cdot PRF \cdot B_r/B_c)$	2c)
$B_r/B_c \leqslant 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert > B_c/2$ and $B_r^2\tau/B_c > 1$	$10 \log (\tau \cdot PRF \cdot B_r/B_c) - \lambda_c\ (\Delta f)\ 10 \log (B_r^2\tau/2B_c)$	2e)
$B_r/B_c \leqslant 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert < B_c/2$ and $B_r^2\tau/B_c < 1$	$10 \log (PRF/B_r)$	2d)
$B_r/B_c \leqslant 1,\ B_r\tau > 1,\ \lvert\Delta f\rvert > B_c/2$ and $B_r^2\tau/B_c < 1$	$10 \log (PRF/B_r) + 3\lambda_c\ (\Delta f)$	2e)
$B_r/B_c \leqslant 1,\ B_r\tau < 1,\ \lvert\Delta f\rvert \geqslant 0$ and $B_r^2\tau/B_c < 1$	$10 \log (PRF/B_r)$	2d)

TABLE III – *Duty-cycle functions for interfering digital signals (2-PSK, 4-PSK, MPSK, O4-PSK, MSK, 2-FSK)*

Continuous input waveform (see Fig. 3) η_i = 0 dB		
Bandwidth conditions	$\eta_0(\Delta f)$ (dB)	Figure No.
$B_r/R_s \leqslant 1$, $\lvert\Delta f\rvert \geqslant 0$	0	3d)
$B_r/R_s > 1$, $\lvert\Delta f\rvert \leqslant B_r/2$	0	3b)
$B_r/R_s > 1$, $\lvert\Delta f\rvert > B_r/2$	$-\lambda_d(\Delta f)$ 10 log $(2B_r/R_s)$	3c)

Gated input waveform (see Fig. 4) η_i = 10 log (τ_g/T_g) dB		
Bandwidth conditions	$\eta_0(\Delta f)$ (dB)	Figure No.
$B_r/R_s \leqslant 1$, $1/B_r < \tau_g$, $\lvert\Delta f\rvert \geqslant 0$	10 log (τ_g/T_g)	4d)
$B_r/R_s \leqslant 1$, $\tau_g < 1/B_r < T_g$, $\lvert\Delta f\rvert \geqslant 0$	10 log $1/(B_r/T_g)$	4e)
$B_r/R_s \leqslant 1$, $1/B_r > \tau_g$, $\lvert\Delta f\rvert \geqslant 0$	0	4f)
$B_r/R_s > 1$, $\lvert\Delta f\rvert \leqslant B_r/2$	10 log (τ_g/T_g)	4b)
$B_r/R_s > 1$, $\lvert\Delta f\rvert > B_r/2$	10 log (τ_g/T_g) $-\lambda_d(\Delta f)$ 10 log $(2B_r/R_s)$	4c)

TABLE IV – *Duty-cycle functions for interfering analogue signals*

Continuous input waveform η_i = 0 dB	
Bandwidth conditions	$\eta_0(\Delta f)$ (dB)
B_r/B_i : any value, $\lvert\Delta f\rvert \geqslant 0$	0

Gated input waveform η_i = 10 log (τ_g/T_g) dB	
Bandwidth conditions	$\eta_0(\Delta f)$ (dB)
$B_r/B_i > 1$, $\lvert\Delta f\rvert \geqslant 0$	10 log (τ_g/T_g)
$B_r/B_i \leqslant 1$, $1/B_r < \tau_g$, $\lvert\Delta f\rvert \geqslant 0$	10 log (τ_g/T_g)
$B_r/B_i \leqslant 1$, $\tau_g < 1/B_r < T_g$, $\lvert\Delta f\rvert \geqslant 0$	10 log $(1/B_rT_g)$
$B_r/B_i \leqslant 1$, $1/B_r > T_g$, $\lvert\Delta f\rvert \geqslant 0$	0

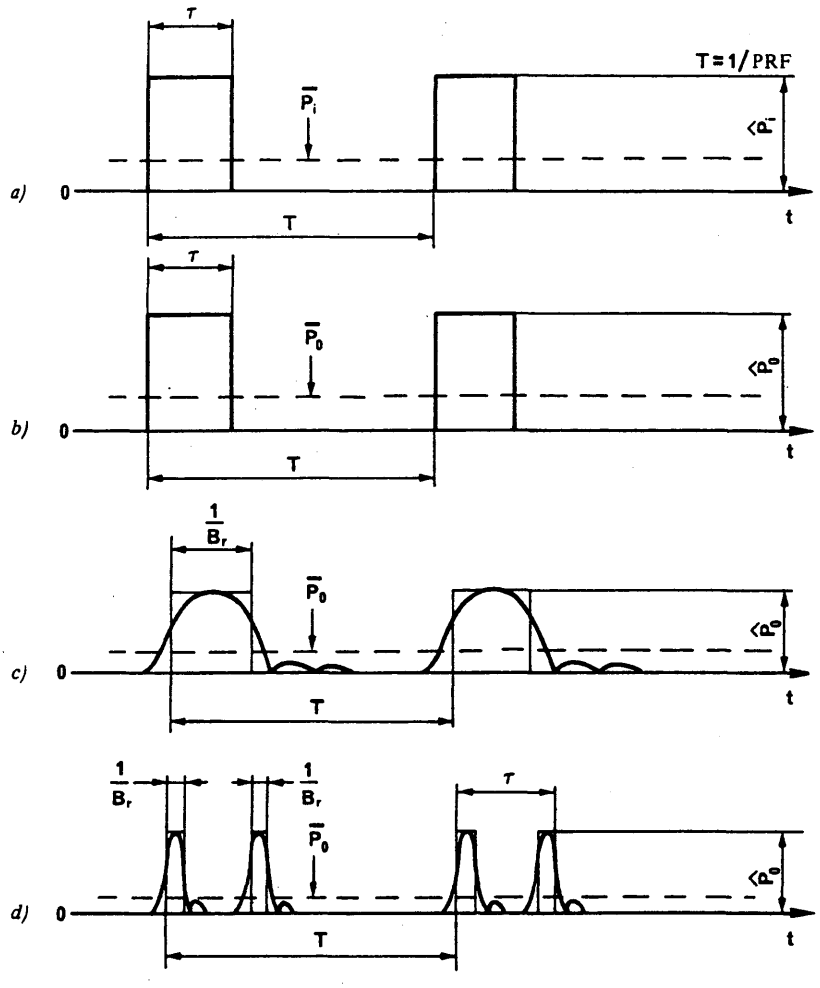


FIGURE 1 – *Pulse train and responses for various bandwidth conditions*

a) Pulse train input

b) Undistorted response

$B_r\tau > 1, |\Delta f| \leq B_r/2$

c) Impulsive response

$B_r\tau \leq 1, |\Delta f| \geq 0$

d) Double-impulsive response

$B_r\tau > 1, |\Delta f| > B_r/2$

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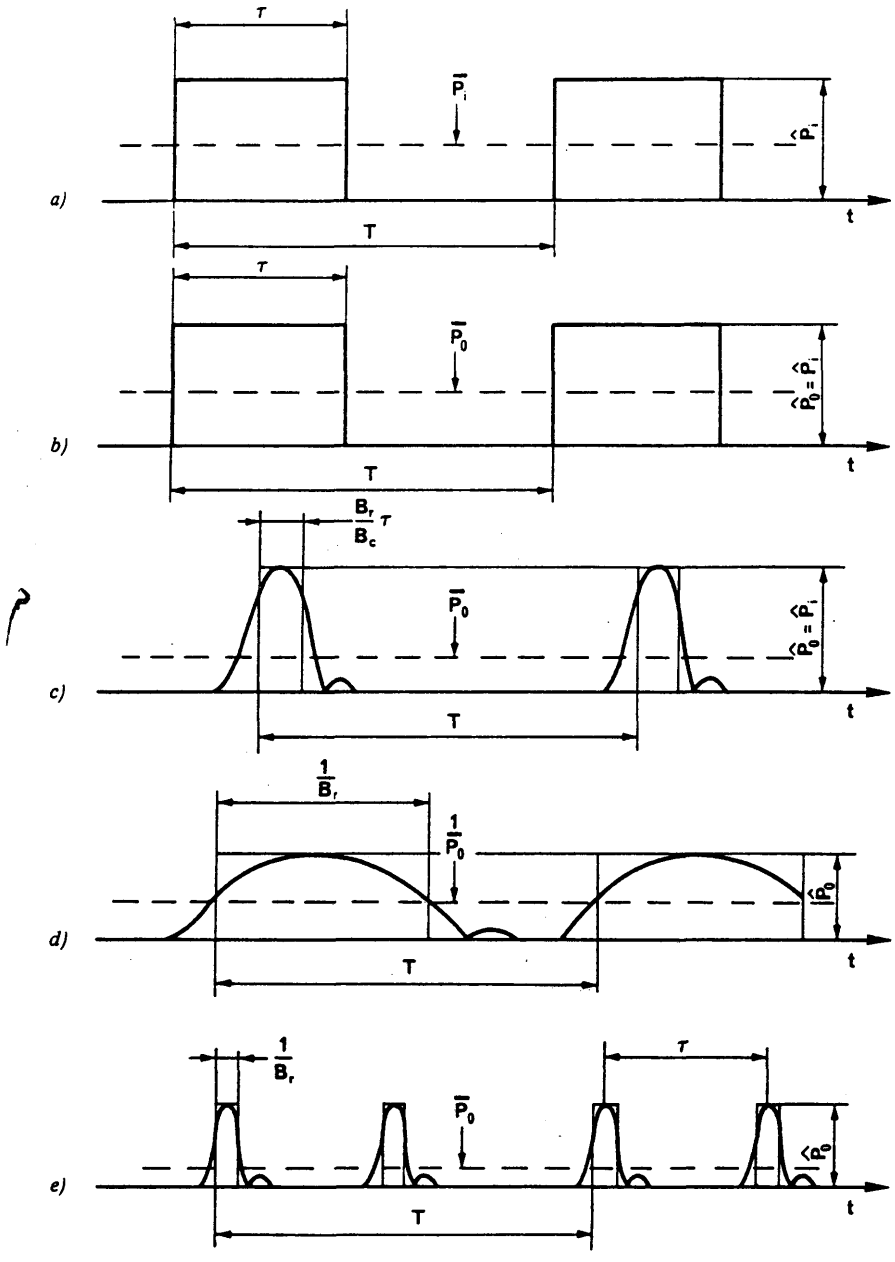


FIGURE 2 – *Chirp radar pulse and responses for various bandwidth conditions*

a) Chirp pulse train input

b) Undistorted response

$B_r/B_c > 1, B_r\tau > 1, |\Delta f| \leq B_r/2$

c) Partially undistorted response

$B_r/B_c \leq 1, B_r\tau > 1, |\Delta f| < B_c/2$

d) Impulsive response

. $B_r/B_c \leq 1, B_r\tau > 1, |\Delta f| < B_c/2, B_r^2\tau/B_c < 1$, or

$B_r/B_c \leq 1, B_r\tau < 1, |\Delta f| \geq 0, B_r^2\tau/B_c < 1$

e) Double-impulsive response

$B_r/B_c > 1, B_r\tau > 1, |\Delta f| > B_r/2$, or

$B_r/B_c \leq 1, B_r\tau > 1, |\Delta f| > B_c/2$

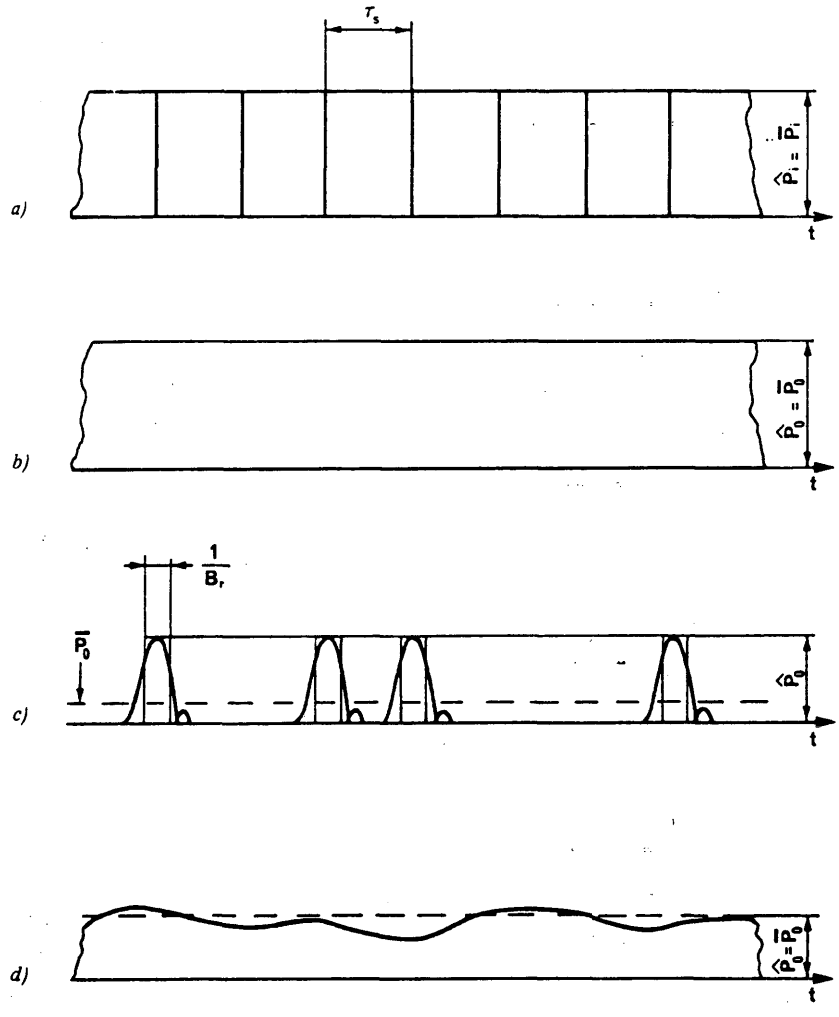


FIGURE 3 – *Digital signal and responses for various bandwidth conditions*

a) Digital signal input

b) Undistorted response

$B_r/R_S > 1, |\Delta f| \leq B_r/2$

c) Impulsive response

$B_r/R_S > 1, |\Delta f| > B_r/2$

d) Noise-like response

$B_r/R_S < 1, |\Delta f| \geq 0$

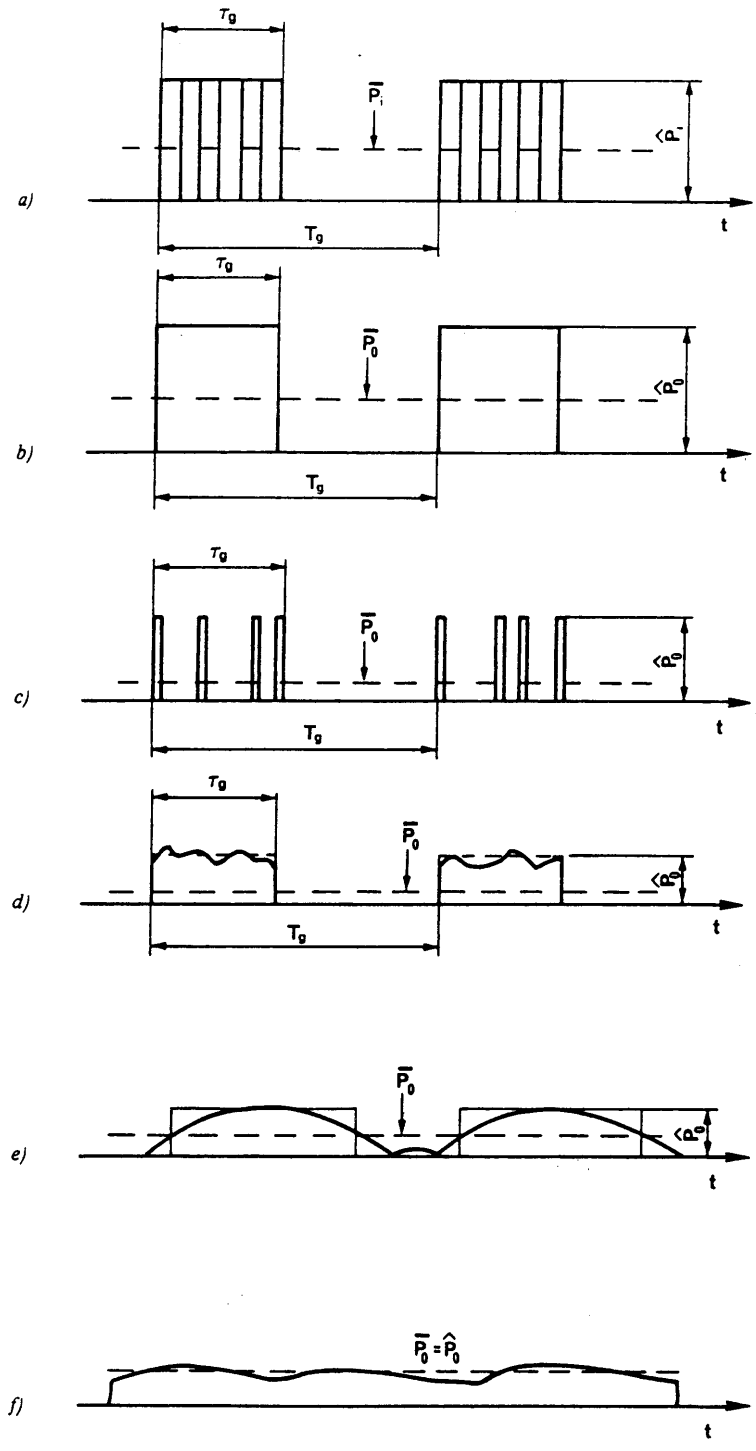


FIGURE 4 – *Gated digital signal and responses for various bandwidth conditions*

a) Gated digital signal input

b) Undistorted response

$B_r/R_s > 1, |\Delta f| \leq B_r/2$

c) Impulsive response

$B_r/R_s > 1, |\Delta f| > B_r/2$

d) Noise-like-burst response

$B_r/R_s \leq 1, 1/B_r < \tau_g, |\Delta f| \geq 0$

e) Impulsive response

$B_r/R_s \leq 1, \tau_g < 1/B_r < \tau_g, |\Delta f| \geq 0$

f) Noise-like output

$B_r/R_s \leq 1, 1/B_r > \tau_g, |\Delta f| \geq 0$

The various terms and functions listed in the tables and shown in the figures are defined as follows:

MPSK : M-ary phase shift keying,

2-PSK : binary phase shift keying,

4-PSK : quadriphase shift keying,

O4-PSK : offset 4-PSK,

MSK : minimum shift keying,

2-FSK : binary frequency shift keying,

B_r : 3 dB bandwidth (Hz) of the receiving channel,

R_s : symbol rate (Bd) of a digital interferer,

τ_g : gate length (s) of time gated interferer signal,

T_g : average spacing (s) of gates,

τ : pulse length (s),

PRF : pulse repetition frequency (pulses/s),

B_c : total frequency shift (Hz) in a linear frequency-modulated pulse, as used in chirp radar,

λ_r, λ_c, λ_d : smoothing functions.

When the interfering signal is a pulse train (see Fig. 1a) and $B_r\tau > 1$, the shape of the response waveform is approximated by a rectangle with a width equal to τ (see Fig. 1b) when $|\Delta f| \leqslant B_r/2$, and by a pair of rectangles each of width $1/B_r$ (see Fig. 1d) when $|\Delta f| > B_r/2$. Actually, as $|\Delta f|$ is increased from 0 to $\gg B_r/2$, the response waveform changes as depicted in Fig. 5. Similar waveform transitions occur in the responses to other types of digital interfering signals. For purposes of simplification in calculating η_0 (Δf), arbitrary smoothing functions are used to account for the transitional wave shape (Fig. 5c) that occurs when $|\Delta f| \approx B_r/2$.

The smoothing function $\lambda_r(\Delta f)$ used with pulse train (radar) signals is:

$$\lambda_r(\Delta f) = 1 - \exp\left[-\alpha_r\left(\frac{\Delta f}{B_r} - \frac{1}{2}\right)\right] \qquad \text{for } |\Delta f| \geqslant B_r/2 \tag{4}$$

where:

$\alpha_r = 3.91/[(\Delta f_r/B_r) - (1/2)]$

Δf_r : frequency at the intersection of 10 log $H(f)$ and 10 log $P(f)$ + 20 log $(B_r\,\tau/2)$ − 10.*

$H(f)$ and $P(f)$ are defined in Report 654.

The smoothing function used for interfering chirp radar signals is:

$$\lambda_c(\Delta f) = 1 - \exp\left[-2.61\left(\frac{\Delta f}{B_c} - \frac{1}{2}\right)\right] \qquad \text{for } |\Delta f| \geqslant B_c/2 \tag{5}$$

The smoothing function $\lambda_d(\Delta f)$ for digital signals is:

$$\lambda_d(\Delta f) = 1 - \exp\left[-\alpha_d\left(\frac{\Delta f}{B_r} - \frac{1}{2}\right)\right] \qquad \text{for } |\Delta f| \geqslant B_r/2 \tag{6}$$

where:

$\alpha_d = 3.91/[(\Delta f_d/B_r) - (1/2)]$

Δf_d : frequency at the intersection of 10 log $H(f)$ and 10 log $P(f)$ + 20 log (B_r/R_s) − 10.**

* If 10 log $H(f)$ exceeds the other expression for all values of f so that there is no intersection, $\lambda_r(\Delta f) = 0$.

** If 10 log $H(f)$ exceeds the other expression for all values of f so that there is no intersection, $\lambda_d(\Delta f) = 0$.

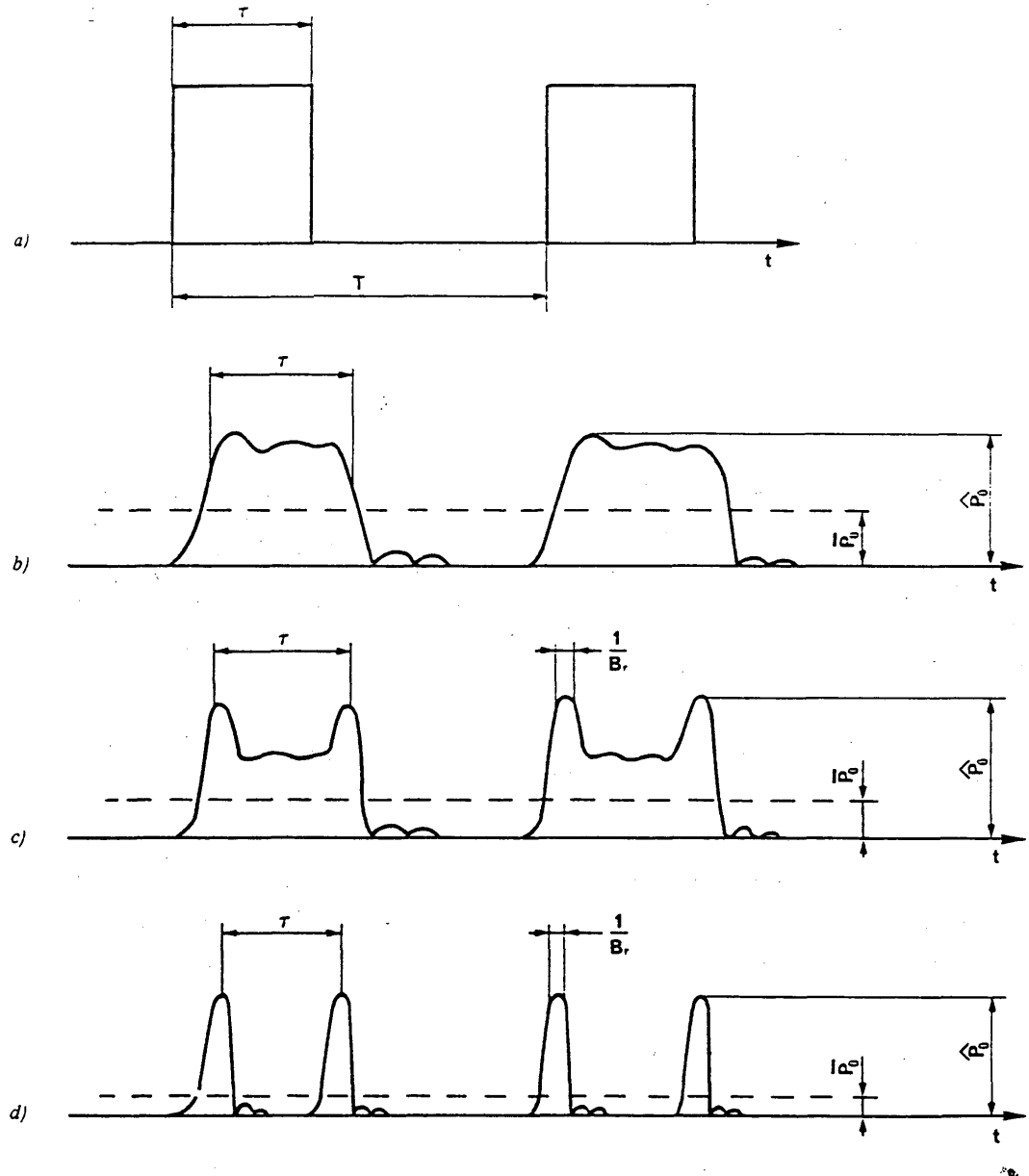


FIGURE 5 – *An example of the response to a pulse train when* $B_r\tau > 1$

a) Input pulse
b) Response waveform when $|\Delta f| = 0$
c) Response waveform when $|\Delta f| \simeq B_r/2$
d) Response waveform when $|\Delta f| \geqslant B_r/2$

3. Computation of $\hat{I}/N$

It is often required to calculate the ratio of the peak interference power, $\hat{I}$, and the noise power, N_0, where $\hat{I}$ is obtained using equation (1) of this Report and N_0 is obtained with equation (3) of Report 654.

$$\hat{I}/N = \hat{I} - N_0 \quad \text{dB} \tag{7}$$

4. Conclusion

Expressions are given for calculating the peak interference power in the receiver bandwidth when the receiver is subjected to one of several types of unwanted emissions. The results obtained with the expressions can be used to determine the frequency-distance separation required between the interferer and the receiver to prevent unacceptable degradation.

REPORT 829

CALCULATION OF THE PROBABILITY OF INTERFERENCE

(Question 45/1)

(1982)

1. Introduction

The Radio Regulations, Nos. 160, 161 and 162, give definitions of "interference", "permissible interference" and "acceptable interference" which were formulated on the basis of Recommendation 506 (Kyoto, 1978). Because of the variabilities of equipment characteristics, transmission loss, and (for some services) location and traffic intensity, only probabilistic estimates of interference are realistic. Report 656 discusses this in more detail. In this Report, a formula is derived for the probability of interference that corresponds to the definition of interference in No. 160 of the Radio Regulations. The relationship of this formula to others that have been used is shown.

2. Derivation of probability of interference

The Radio Regulations, No. 160, defines *interference* as "the effect of unwanted energy ... upon reception in a radio communication system, manifested by any performance degradation, misrepresentation, or loss of information which could be extracted in the absence of such unwanted energy".

2.1 *The probability of instantaneous interference*

Let: A denote "The desired transmitter is transmitting".

B denote "The wanted signal is satisfactorily received in the absence of unwanted energy".

C denote "Another equipment is producing unwanted energy".

D denote "The wanted signal is satisfactorily received in the presence of the unwanted energy".

All of these statements refer to the same small time period. Then according to the definitions, interference means "A and B and C and D^*," where D^* is the negation or opposite of D. Let $P(x)$ denote the "probability of x", and $P(x \mid y)$ denote the "probability of x, given y". Then, the probability of interference during the small time period is

$$P(I) = P(A \text{ and } B \text{ and } C \text{ and } D^*) \tag{1}$$

In Annex I, it is shown that this can be expressed in terms of known or computable quantities:

$$P(I) = [P(B \mid A) - P(D \mid A \text{ and } C)]\, P(A \text{ and } C) \quad (2)$$

It may be preferable to consider the probability of interference only during the time that the wanted transmitter is transmitting. This probability is:

$$P'(I) = P(B \text{ and } C \text{ and } D^* \mid A) \quad (3)$$

which can be reduced to

$$P'(I) = [P(B \mid A) - P(D \mid A \text{ and } C)]\, P(C \mid A) \quad (4)$$

2.2 *Discussion of equations (2) and (4)*

First consider the difference between equations (2) and (4). The probability of interference can be interpreted as the fraction of time that interference exists. In equation (2), this fraction is the number of seconds of interference during a time period divided by the number of seconds in the time period. In equation (4), the fraction is the number of seconds of interference divided by the number of seconds the wanted transmitter is transmitting during the time period. This second fraction is larger than the first unless the wanted transmitter is on all the time.

$P(B \mid A)$ is just the probability that a wanted signal will be correctly received when there is no interference, often expressed as the probability that $S/N \geqslant R$, where S is the signal power, N is the noise power, and R is the signal-to-noise ratio required for satisfactory service. In some services this probability is called the reliability, and is often computed when the system is designed. It can be computed if system parameters (for example, transmitter and receiver location, power, required S/N) are known using statistical data on transmission loss (for example, Recommendation 370) and statistical data on radio noise (for example, Reports 322 and 670).

Many systems, such as satellite or microwave relay point-to-point systems, are designed so that $P(B \mid A) \approx 1$. In other services, such as long-distance ionospheric point-to-point services, or mobile services near the edge of the coverage area, $P(B \mid A)$ may be quite small. In this latter case, the probability of interference will be small regardless of the other probabilities.

$P(D \mid A \text{ and } C)$ is the probability that the wanted signal will be correctly received even when the unwanted energy is present. It can be computed if there is sufficient information about the location, frequency, power etc., of the source of unwanted energy. For examples, see the references in Report 656.

Notice that it has been assumed that $P(D \mid A \text{ and } C) \leqslant P(B \mid A)$; that is, if the signal can be received satisfactorily in the presence of unwanted energy, then it can surely be received satisfactorily in the absence of the unwanted energy. Thus, $P(I)$ cannot be negative.

$P(A \text{ and } C)$ is the probability that the wanted transmitter and the source of unwanted energy are on simultaneously. In some situations, the wanted transmitter and source of unwanted energy may be operated independently. For example, they may be on adjacent channels, or beyond a coordination distance. In this case, $P(A \text{ and } C) = P(A)P(C)$, where $P(A)$ is the fraction of time that the wanted transmitter is emitting, and $P(C)$ is the fraction of time that the unwanted source is on.

In other situations, the operation may be highly dependent. For example, the transmitters may be co-channel stations in a disciplined mobile service. In this case, $P(A \text{ and } C)$ is very small, but perhaps not zero, because a station can be located so that it causes interference even when it cannot hear the other transmitter.

The two transmitters might both operate continuously. For example, one might be part of a microwave point-to-point service, and the other a satellite sharing the same frequency band. In this case, $P(A \text{ and } C) = 1$, and the probability of interference depends entirely on the factor in square brackets in equation (2).

Similarly, $P(C \mid A) = P(C)$ if the transmitters operate independently; $P(C \mid A)$ is very small if the two transmitters are co-channel stations in a disciplined land mobile service; and $P(C \mid A) = 1$ if the unwanted transmitter is on all the time.

In general, all the terms in equations (2) and (4) affect the probability of interference, although their relative importance is different in different services.

2.3 *Special cases*

In this section, equations (2) and (4) are reduced to formulae that have been sometimes used to compute the probability of interference. The assumptions necessary to make the reductions illuminate the features of the equations.

First consider a system designed for high reliability, so that $P(B \mid A) = 1$. Now,

$$1 - P(D \mid A \text{ and } C) = P(D^* \mid A \text{ and } C) \tag{5}$$

which is the probability that the wanted signal is not received in the presence of unwanted energy. Then

$$P(I) = P(D^* \mid A \text{ and } C)P(A \text{ and } C) \tag{6}$$

an expression that is sometimes used for the probability of interference (for example, see Juroshek [1975]).

Under the assumption, equation (4) becomes

$$P'(I) = P(D^* \mid A \text{ and } C)P(C \mid A) \tag{7}$$

If in addition, both transmitters operate continuously, or at least on the same schedule, so that $P(A \text{ and } C) = P(C \mid A) = 1$, then:

$$P(I) = P(D^* \mid A \text{ and } C) = P'(I) \tag{8}$$

a formula that has been used [Freeman, 1972; Jasinski, 1975] to calculate the probability of interference even in services for which $P(A \text{ and } C) \neq 1$.

2.4 *Probability of interference during a transmission*

Equations (2) and (4) give the probability that interference will occur at an instant of time. A more conservative view is that interference occurs if any part of a transmission is lost; that is, if the unwanted energy causes loss of information anytime during the wanted transmission. This is particularly applicable to digital transmission systems. In this case, we replace the factor $P(C \mid A)$ in equation (4) with the probability that the wanted and unwanted transmissions overlap.

If both the wanted transmission and the unwanted energy are present all the time, this probability is one. The interesting case is when one or both transmit intermittently.

Let:

T_w: the length of a transmission by the wanted transmitter;

N: the average number of unwanted emissions per unit time; and

T_{ua}: the average length of an unwanted emission.

Make the queueing theory assumption that the beginning time of the unwanted emission is Poisson distributed, and the length of the quiet times is exponentially distributed [Kleinrock, 1975]. It is shown in Annex I that the probability that the wanted and unwanted transmissions overlap is:

$$P(overlap) = 1 - (1 - NT_{ua}) \exp [- T_w N/(1 - NT_{ua})] \tag{9}$$

Substituting into (2), the probability of interference at some time during a transmission of length T_w is:

$$P(interference/T_w) = [P(B/A) - P(D/A \text{ and } C)] \times \left\{1 - (1 - NT_{ua}) \exp [- T_w N/(1 - NT_{ua})]\right\} \tag{10}$$

Notice that NT_{ua} is the fraction of time that the unwanted energy is present. If the unwanted energy is present all the time so that $NT_{ua} = 1$, then $P(overlap) = 1$. If NT_w and NT_{ua} are both much smaller than 1 (both operations are very intermittent) then $P(overlap) \sim NT_w + NT_{ua}$.

3. An illustrative example

To clarify the role of the different terms in equations (2), (4), and (10), consider the following hypothetical example: let the wanted signal be transmitted on a mobile-to-mobile link in a land mobile radio service. Suppose that $P(B \mid A) = 0.85$; that is, because of terrain shielding, transmission loss variability, and variations in radio noise, the wanted signal is correctly received only 85% of the time when there is no interference. Suppose that the

unwanted energy comes from a powerful adjacent channel base station on one edge of the coverage area, and causes interference when the interfered mobile station is in that part of the coverage area. Given data about the systems' characteristics, it is possible to compute the probability of communicating in such a situation using computer methods such as those described in Report 655 and by Berry [1977]. To be specific, suppose that the calculated value of $P(D \mid A \text{ and } C) = 0.65$. Finally, suppose that the wanted mobile-to-mobile link operates 0.25 of the time, and the unwanted base station operates 0.5 of the time. Because they are not on the same channel, they operate independently and $P(A \text{ and } C) = P(A)\,P(C) = 0.125$. Then the probability of interference at a specific instant of time according to equation (2) is

$$P(I) = [0.85 - 0.65]\,(0.125) = 0.025 \tag{11}$$

Also, independence implies that $P(C \mid A) = P(C)$, so $P'(I) = 0.1$.

If the formula (8), which is the probability of interference given that both stations operate continuously, were used the probability of interference would be 0.35, (more than 10 times larger than the probability given by formula (2) and 3.5 times that given by formula (4)). The reduction in probability of interference results from the recognition that not all signals would be received correctly even when the unwanted transmitter is off.

To find the probability that interference will occur some time during a message using equation (10), assume that the length of a mobile-to-mobile message (T_w) is 0.01 h. Assume that the average length of the interfering base station transmissions (T_{ua}) is also 0.01 h. Then N must be 50 since $NT_{ua} = P(C) = 0.5$. Then the probability of overlap is 0.816, and the probability of interference during a transmission of length 0.01 h is 0.163.

4. Conclusion

Equations (2) and (4) give the probability of instantaneous interference based on the definition of interference in the Radio Regulations. Equation (2) gives the fraction of time that interference occurs; equation (4) gives the fraction of time that interference occurs when the wanted transmitter is on.

Equation (10) shows the probability of interference during a transmission of specified length when the statistics of the intermittent interfering emission are known.

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ANNEX I

Using the notation and assumptions of § 2 of the Report, it will be shown that:

$$\begin{aligned} P(I) &= P(A \text{ and } B \text{ and } C \text{ and } D^*) \\ &= [P(B \mid A) - P(D \mid A \text{ and } C)]\,P(A \text{ and } C) \end{aligned} \tag{12}$$

First change the order of the statements and use the definition [Zehna, 1970]:

$$P(x \text{ and } y) = P(x \mid y)\,P(y) \tag{13}$$

$$\begin{aligned} P(I) &= P((B \text{ and } D^*) \text{ and } (A \text{ and } C)) \\ &= P(B \text{ and } D^* \mid A \text{ and } C)\,P(A \text{ and } C) \end{aligned} \tag{14}$$

Assume that if a signal can be correctly received in the presence of unwanted energy then it can be correctly received when the unwanted energy is absent. This means:

$$P(B \mid D) = 1 \tag{15}$$

Apply the probability identity [Zehna, 1970]:

$$P(x \text{ and } y^*) = P(x) - P(x \text{ and } y) \tag{16}$$

to the right side of formula (14):

$$P(I) = [P(B \mid A \text{ and } C) - P(B \text{ and } D \mid A \text{ and } C)]\, P(A \text{ and } C) \tag{17}$$

Because C is irrelevant to B,

$$P(B \mid A \text{ and } C) = P(B \mid A) \tag{18}$$

Furthermore, holding the condition "A and C" constant and not writing it,

$$P(B \text{ and } D) = P(B \mid D)\, P(D) = P(D) \tag{19}$$

where the last equality from formula (19) results from formula (15). Substituting formulae (18) and (19) into formula (17) yields

$$P(I) = [P(B \mid A) - P(D \mid A \text{ and } C)]\, P(A \text{ and } C) \tag{20}$$

Similarly,

$$P(B \text{ and } C \text{ and } D^* \mid A) = [P(B \mid A) - P(D \mid A \text{ and } C)]\, P(C \mid A) \tag{21}$$

Formula (20) is the probability of interference at any instant of time. A more conservative view is that interference occurs if any part of a transmission is lost; that is, if the unwanted energy causes loss of information anytime during the wanted transmission. A modification to formula (20) for this view is derived below.

Let W be the wanted transmitter and let U be the unwanted transmitter. Assume W begins a transmission at time t_0, and find the probability that U comes on during T_w, the length of a transmission by W.

$$\begin{aligned} P(\textit{overlap}) &= P[(U \text{ on at } t_0) \text{ or } (U \text{ off at } t_0 \text{ and turns on during } T_w)] \\ &= P(U \text{ on at } t_0) + P(U \text{ off at } t_0 \text{ and turns on during } T_w) \\ &= P(U \text{ on at } t_0) + P(U \text{ turns on during } T_w \mid U \text{ off at } t_0)\, P(U \text{ off at } t_0) \end{aligned} \tag{22}$$

Let:

N: the average number of transmission per second by U;

T_u: the length of a transmission by U with average value T_{ua};

S_u: the length of the quiet periods between transmissions by U with average value S_{ua}.

Then:

$$N = 1/(T_{ua} + S_{ua}) \text{ or } S_{ua} = (1 - NT_{ua})/N \tag{23}$$

$$P(U \text{ on at } t_0) = NT_{ua} \text{ and} \tag{24}$$

$$P(U \text{ off at } t_0) = 1 - NT_{ua} \tag{25}$$

Make the assumption that S_u is exponentially distributed:

$$\begin{aligned} P(S_u \leqslant t) &= 1 - \exp(-t/S_{ua}) \\ &= 1 - \exp[-tN/(1 - NT_{ua})] \end{aligned} \tag{26}$$

Now:

$$\begin{aligned} P(U \text{ turn on during } T_w \mid U \text{ off at } t_0) &= P(S_u \leqslant T_w) \\ &= 1 - \exp[-T_w N/(1 - NT_{ua})] \end{aligned} \tag{27}$$

Substituting formulae (24), (25) and (27) into formula (22), yields:

$$P(overlap) = NT_{ua} + \left\{1 - \exp\left[-T_w N/(1 - NT_{ua})\right]\right\}(1 - NT_{ua}) \quad (28)$$
$$= 1 - (1 - NT_{ua}) \exp\left[-T_w N/(1 - NT_{ua})\right]$$

The probability of interference at some time during a transmission of length T_w is the product of the probability of overlap and the first factor on the right side of formula (21):

$$P_i(1) = \left\{P(B \mid A) - P(D \mid A \text{ and } C)\right\}\left\{1 - (1 - NT_{ua}) \exp\left[-T_w N/(1 - NT_{ua})\right]\right\} \quad (29)$$

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REPORT 837

METHODS FOR CALCULATING PULSED RADAR EMISSION SPECTRUM BANDWIDTH

(Study Programme 60A/1)

(1982)

1. Introduction

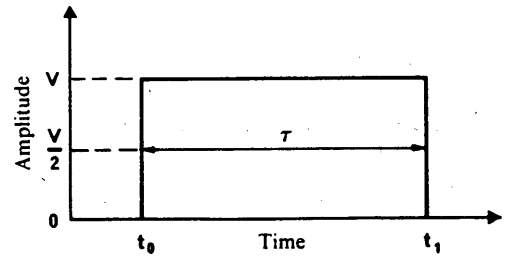
Recommendation 337 describes the primary factors which determine frequency separation between channels. These factors take into consideration the receiver selectivity and the transmitter emission spectrum. Report 654 describes a method of calculating adjacent band interference by taking into consideration the selectivity and the transmitter emission spectrum. In order to address technical issues relating to radar spectrum utilization, it would be useful to have available a means of calculating the radar emission spectrum based on radar technical characteristics. A method of easily relating spectral bounds to radar technical parameters is required. Various techniques have been developed to perform this type of analysis. The methods presented in this Report develop, through straightforward calculations, a straight line first-order approximation of the spectral boundaries of a trapezoidal radar pulse and a linear frequency modulated (FM), "chirped", radar pulse. The techniques outlined below are based on five major pulse characteristics which affect the radio spectrum emission characteristics of pulsed radars. These are the pulse amplitude, pulse duration, pulse rise time and fall time, and frequency modulation of the RF carrier during the pulse (chirping).

2. Calculation of radar emission spectrum

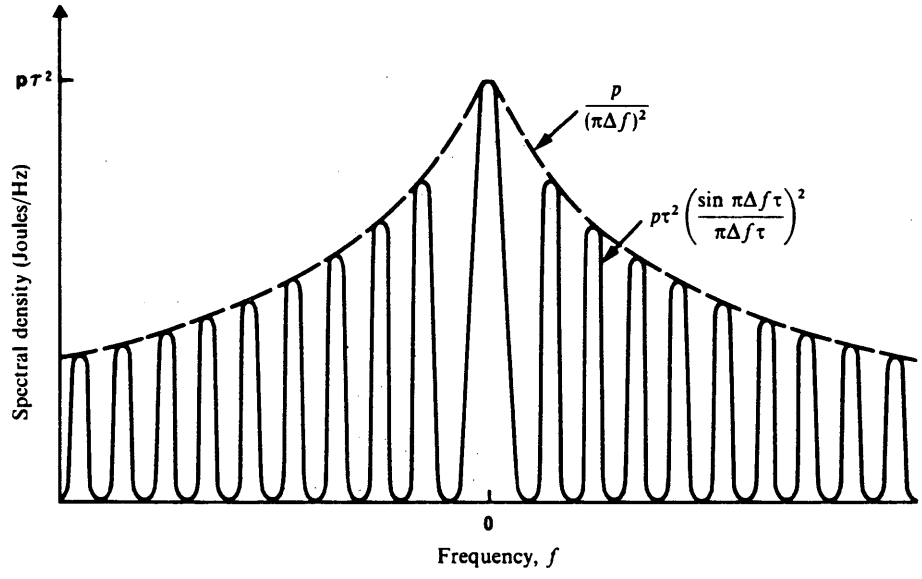
To precisely calculate the emission spectrum of a radar, the time domain waveform is transformed into the frequency domain through a Fourier transform. The theoretical radar pulse is an ideal rectangular pulse whose time and frequency domain representations are shown in Figs. 1a and 1b.

As a practical matter, a rectangular pulse cannot be obtained. The leading and trailing edges of the pulse will have a finite slope because of practical circuit design considerations. This results in a trapezoidal pulse shape (Fig. 2a). The Fourier transform of the square of the amplitude of the waveform yields the energy-density spectrum of the pulse. The spectrum takes the form shown by the solid curve in Fig. 2b.

In EMC analysis, it is appropriate as well as convenient to represent the spectrum by a bound, such as the dotted curve in Fig. 2b. It is common practice to normalize the graph of the spectral density by dividing by the peak value $p\tau^2$ (where p: peak pulse power level (W), and τ: pulse duration between half amplitude points (s)) and express the ratio in decibels (dB). This normalizing process yields the relative energy density in dB.

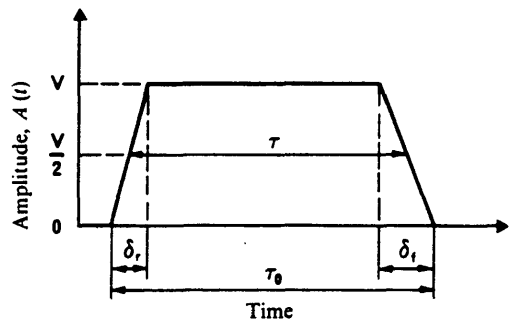


(a) Time domain

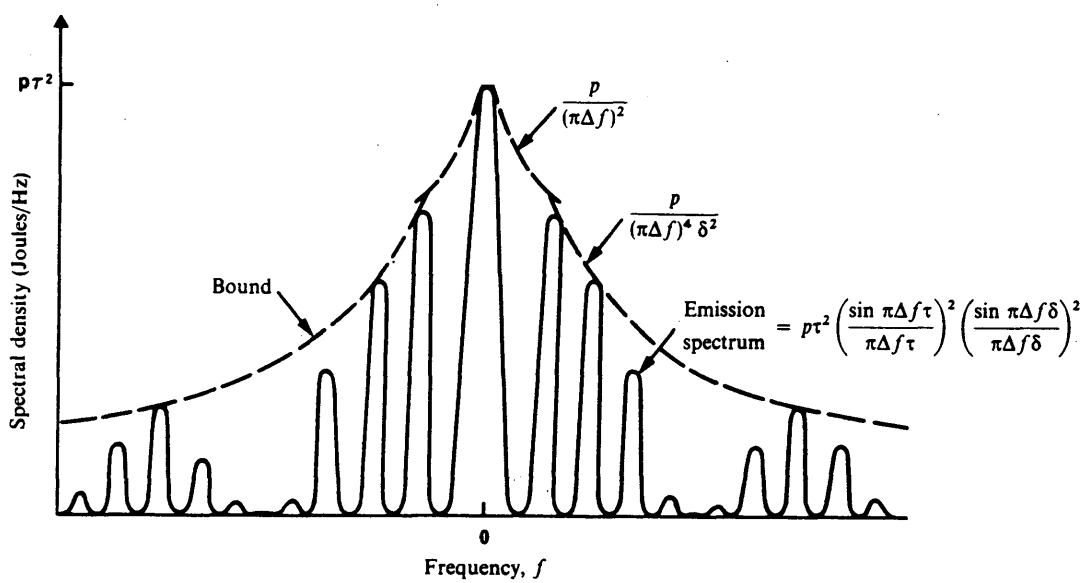


(b) Frequency domain

FIGURE 1 — *Rectangular pulse*



(a) Time domain



(b) Frequency domain

FIGURE 2 — *Trapezoidal pulse (without FM)*

The bounds on the spectrum of a trapezoidal pulse can be obtained by a simple calculation technique [Mason and Zimmerman, 1960]. This procedure involves four steps to draw an approximation of the bounds of the trapezoidal pulse spectrum (see Fig. 3). When a logarithmic scale is used for the abscissa (frequency) and a decibel, (dB) scale is used for the ordinate (relative to the peak spectral power density), the curves representing the bounds of the spectrum can be drawn as straight lines, as shown in Fig. 3. The frequency is referenced to f_c, i.e. $\Delta f = f - f_c$. To plot the spectrum bounds, the following steps are required:

Step 1: Calculate the critical frequencies, Δf_2 and Δf_3 using the equations below:

$$\Delta f_2 = \pm \frac{1}{\pi \tau} \tag{1a}$$

$$\Delta f_3 = \pm \frac{1}{\pi \sqrt{\tau \delta}} \quad \text{where} \quad \frac{1}{\delta} = \frac{1}{2}\left[\frac{1}{\delta_r} + \frac{1}{\delta_f}\right] \tag{1b}$$

where:

τ : pulse duration between half amplitude points, (s)

δ_r : pulse risetime, (s) } defined as the time interval from

δ_f : pulse falltime, (s) } 0 to 100% of the voltage-amplitude.

Note. – The spectrum is symmetrical about the carrier frequency (f_c) whether or not the pulse is symmetrical, (i.e. δ_r equal or not equal to δ_f).

Step 2: On semi-logarithmic paper draw line 1 horizontally through 0 dB.

Step 3: Starting on line 1 at Δf_2, draw line 2 with a slope of -20 dB/decade.

Step 4: Starting on line 1 at Δf_3, draw line 3 with a slope of -40 dB/decade.

The spectrum is bounded by lines 1, 2 and 3 of Fig. 3.

The peak energy density level (p_d), corresponding to the 0 dB level in Fig. 3, is calculated as follows:

$$p_d = (p\tau^2) \qquad \text{Joules/Hz} \tag{2}$$

where:

p : peak pulse power level (W).

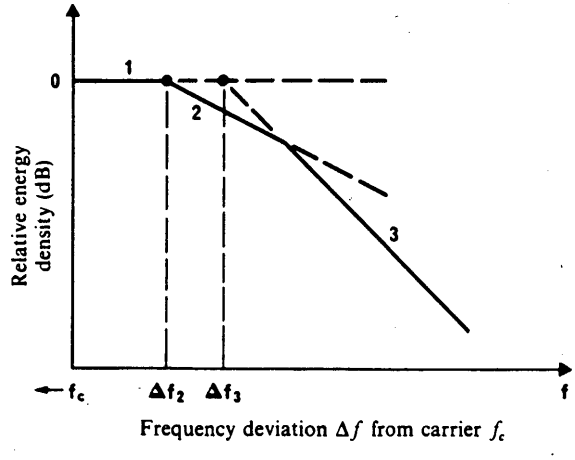


FIGURE 3 — *Spectrum bounds (Mason-Zimmerman approximation) of a trapezoidal pulse spectrum envelope*

When linear frequency modulation is applied to the pulsed carrier frequency as shown in Fig. 4, the peak of the spectrum is reduced in amplitude from $p\tau^2$ to $p\tau/\beta$ and spectrally broadened, but spectrum bounds further removed from the carrier frequency remain unaffected by the frequency modulation (Fig. 5). If the pulse is symmetrical about some time, t_0, (i.e. $\delta_r = \delta_f$) the spectrum will be symmetrical with reference to the nominal carrier f_c. If the pulse is asymmetrical, (i.e. $\delta_r \neq \delta_f$), the spectrum will be asymmetrical as shown in Fig. 6. The direction in which the spectrum is shifted will depend on whether the deviation is negative or positive with respect to time. The spectrum boundaries further removed from the carrier frequency will be symmetrical about a frequency, f_0, which is displaced from f_c, as described in equation (3) [Newhouse, 1982].

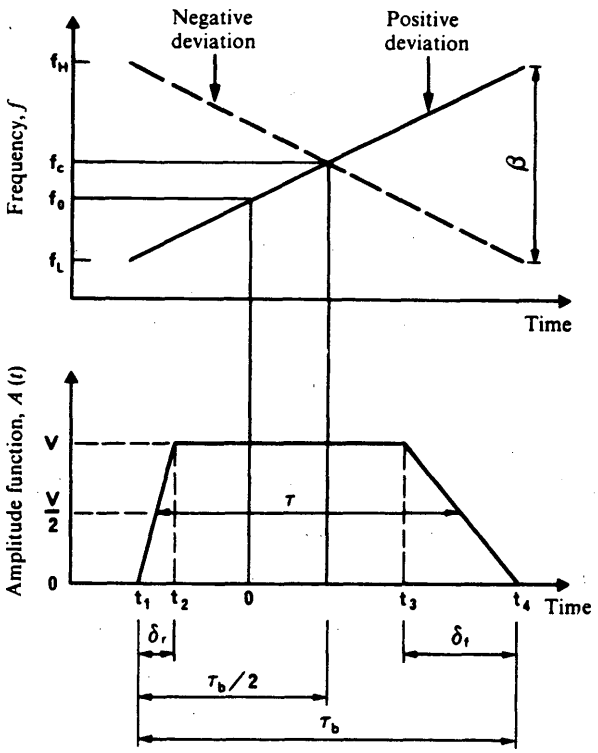


FIGURE 4 — *Parameters describing a chirp pulse*

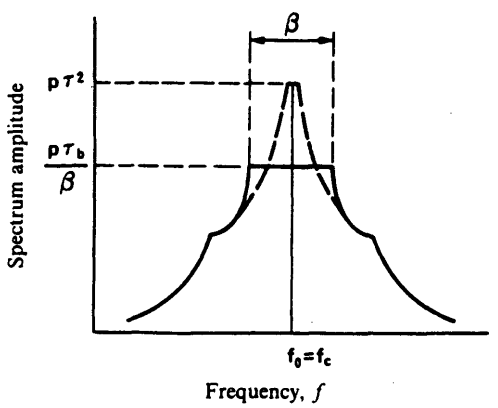


FIGURE 5 — *Shape of spectrum of chirp pulse when* $\delta_r = \delta_f$

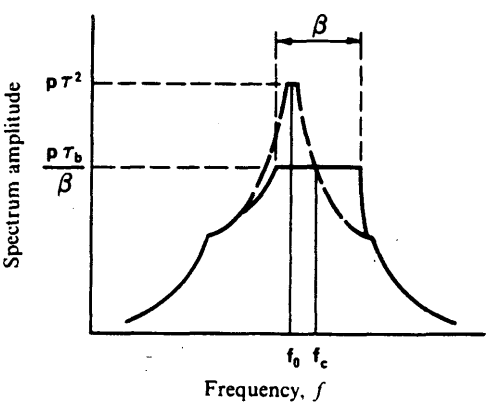


FIGURE 6 — *Shape of spectrum of chirp pulse when* $\delta_r < \delta_f$ *and frequency deviation is positive*

The energy-density spectrum is generally normalized by dividing by $p\tau/\beta$ and expressing the ratio in decibels to obtain the relative energy-density spectrum in dB,

$$f_0 = f_c + Q\left[\frac{\beta(\delta_r - \delta_f)}{2(\delta_r + \delta_f)}\right] \tag{3}$$

where Q is defined in Table I and β is defined in Fig. 4.

The mathematical description of the frequency spectrum of a FM pulse entails the use of Fresnel integrals. Evaluation of the integral is very tedious and must be performed with a high degree of precision to obtain meaningful results. Even when a computer is used the required precision may not be attained unless special care is taken in preparing the algorithm [Newhouse, 1973]. To overcome this difficulty, an approximate bound on the emission spectrum of a frequency-modulated pulse is developed here which is similar in technique to the procedure presented above. For the FM pulse, there are two cases which must be considered. First, when the product of $\beta\tau \leqslant 2/\pi$ the spectrum approximation reduces to the procedure presented above for the trapezoidal pulse. Second, when the product $\beta\tau > 2/\pi$, the procedure expands to seven steps and involves more complex calculations. The procedure for calculating the spectrum bounds for $\beta\tau > 2/\pi$ is presented below [Newhouse, 1982] and shown in Fig. 7. Step 1(A) is used when the pulse risetime is equal to the falltime ($\delta_r = \delta_f$). Step 1(B) is used when the pulse risetime is not equal to the falltime ($\delta_r \neq \delta_f$).

Step 1(A): If the pulse is symmetrical, ($\delta_r = \delta_f$) and $\beta\tau > 2/\pi$, calculate the critical frequencies, Δf_2, Δf_3, Δf_a, Δf_b as shown by equations (4a), (4b), (4c) and (4d):

$$\Delta f_2 = \pm\frac{1}{\pi}\left[\frac{\beta}{\tau_b}\right]^{1/2} \tag{4a}$$

$$\Delta f_3 = \pm\frac{1}{\pi}\left[\frac{\beta}{\tau_b}\right]^{1/4}\left[\frac{1}{\delta}\right]^{1/2} \quad \text{where} \quad \frac{1}{\delta} = \frac{1}{2}\left[\frac{1}{\delta_r} + \frac{1}{\delta_f}\right] \tag{4b}$$

$$\Delta f_{a\pm} = \pm\frac{\beta}{2}\left[1 - \frac{\delta}{\tau_b}\right] \tag{4c}$$

$$\Delta f_{b\pm} = 2\Delta f_{a\pm} \tag{4d}$$

where:

$\beta = f_H - f_L$ from Fig. 4; and

$f_0 = f_c = (f_H + f_L)/2$

The spectrum is now symmetrical about f_c, so the same plot can be used for positive and negative halves, i.e. $f_0 = f_c$.

Step 1(B): When the pulse is asymmetrical, ($\delta_r \neq \delta_f$) and $\beta\tau > 2/\pi$, critical frequencies may be calculated using the following equations:

$$\Delta f_2 = \pm\frac{1}{\pi}\left[\frac{\beta}{\tau_b}\right]^{1/2} \tag{5a}$$

$$\Delta f_3 = \pm\frac{1}{\pi}\left[\frac{\beta}{\tau_b}\right]^{1/4}\left[\frac{1}{\delta}\right]^{1/2} \quad \text{where} \quad \frac{1}{\delta} = \frac{1}{2}\left[\frac{1}{\delta_r} + \frac{1}{\delta_f}\right] \tag{5b}$$

When the spectrum is asymmetrical, positive and negative halves must be plotted separately, because $f_0 \neq f_c$ as shown in equation (3).

For positive Δf:

$$\Delta f_{a+} = \frac{M\beta}{\delta_r + \delta_f}\left[1 - \frac{\delta_r + \delta_f}{2\tau_b}\right] \tag{6a}$$

$$\Delta f_{b+} = \begin{cases} 2\Delta f_{a\pm} & (\text{when } \beta\delta \leqslant 1/\pi) \\ \dfrac{M\beta}{\delta_r + \delta_f}\left[\dfrac{1}{1 - \sqrt{\dfrac{N}{2(\delta_r + \delta_f)}}}\right] & (\text{when } \beta\delta > 1/\pi) \end{cases} \tag{6b}$$

For negative Δf:

$$\Delta f_{a-} = \frac{-N\beta}{\delta_r + \delta_f}\left[1 - \frac{\delta_r + \delta_f}{2\tau_b}\right] \tag{7a}$$

$$\Delta f_{b-} = \begin{cases} 2\Delta f_{a-} & (\text{when } \beta\delta \leqslant 1/\pi) \\ \dfrac{-N\beta}{\delta_r + \delta_f}\left[\dfrac{1}{1 - \sqrt{\dfrac{M}{2(\delta_r+\delta_f)}}}\right] & (\text{when } \beta\delta > 1/\pi) \end{cases} \tag{7b}$$

For M, N and Q values used here, see Table I.

Step 2: Draw line 1 horizontally through 0 dB (Figs. 7 and 8).

Step 3: If Δf_b is less than $1/(\pi\delta)$, use Fig. 7 and draw line 2 with a slope of -20 dB/decade starting on line 1 at Δf_2.

If Δf_b is equal to or greater than $1/(\pi\delta)$, skip this step and use Fig. 8 for the next step.

Step 4: Draw line 3 with a slope of -40 dB/decade starting on line 1 at Δf_3 in Fig. 7 or Fig. 8.

Step 5: Locate point "a" or "a'" at 6 dB down from 0 dB at Δf_a or $\Delta f_a'$ in Fig. 7 or Fig. 8 respectively.

Step 6: Locate point "b" or "b'" at Δf_b or $\Delta f_b'$ on line 2 in Fig. 7 or on line 3 in Fig. 8 respectively.

Step 7: Draw line 4 or line 4' through points "a" and "b" in Fig. 7 or through points "a'" and "b'" in Fig. 8.

TABLE I — *M, N and Q values for equations (3), (6) and (7)*

Constants	Positive frequency shift	Negative frequency shift
M	δ_f	δ_r
N	δ_r	δ_f
Q	$+1$	-1

Spectrum bounds: For the situation where $\Delta f_b < 1/(\pi\delta)$, the spectrum is bounded by a curve as described by lines 1, 4, 2 and 3 in Fig. 7. If $\Delta f_b \geqslant 1/(\pi\delta)$, the spectrum would be bounded by a curve as described by lines 1, 4 and 3 in Fig. 8.

The peak energy density level (p_d) corresponding to the 0 dB level in Figs. 7 or 8 is calculated as follows:

$$p_d = (p\tau_b/\beta) \qquad \text{Joules/Hz} \tag{8}$$

where:

β : frequency deviation during pulse (Hz),

τ_b : pulse width at the base of the pulse (s).

The techniques presented above are methods to calculate a first order approximation of a pulsed radar emission spectrum. The bounds obtained with these techniques are within one dB of the peaks of the main lobe and major side lobes of the pulsed emission spectrum envelope. These techniques are a reasonable approximation of the emission spectrum boundary up to a value of Δf corresponding to a spectral power density between 60 and 70 dB below the peak energy level. In this region, noise and distortion generated in the high power output stages create a noise floor which dominates the spectrum bounds for larger values of Δf.

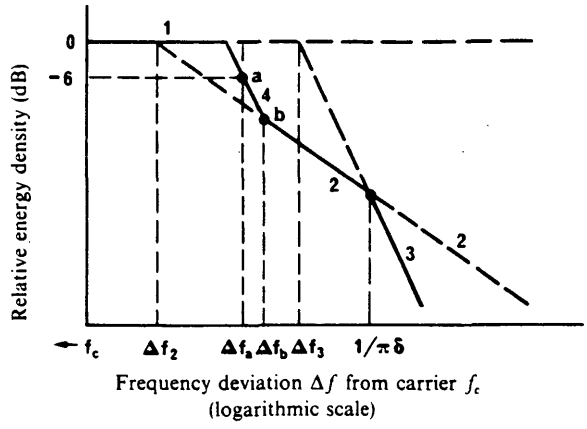


FIGURE 7 — *Spectrum bounds for a linear FM pulse when* $\Delta f_b < \frac{1}{\pi\delta}$

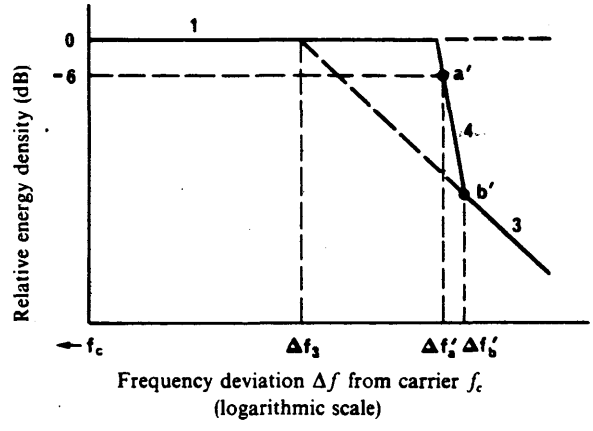


FIGURE 8 — *Spectrum bounds for a linear FM pulse when* $\Delta f_b' \geq \frac{1}{\pi\delta}$

3. Sample calculations of radar emission spectrum envelope

Example 1. — Trapezoidal pulse

$\tau = 6 \times 10^{-6}$ s
$\delta_r = 0.2 \times 10^{-6}$ s
$\delta_f = 0.35 \times 10^{-6}$ s

Step 1: Calculate critical frequencies Δf_2 and Δf_3 using equations (1a) and (1b).

$$\Delta f_2 = \pm \frac{1}{\pi\tau} = \pm \frac{1}{\pi (6 \times 10^{-6})} = 53.1 \text{ kHz}$$

$$\frac{1}{\delta} = \frac{1}{2}\left[\frac{1}{\delta_r} + \frac{1}{\delta_f}\right] = \frac{10^6}{2}\left[\frac{1}{0.2} + \frac{1}{0.35}\right] = \frac{1}{0.25 \times 10^{-6}}$$

$$\Delta f_3 = \pm \frac{1}{\pi (\tau\delta)^{1/2}} = \frac{1}{\pi [(6 \times 10^{-6})(0.25 \times 10^{-6})]^{1/2}} = 259.9 \text{ kHz}$$

Step 2: In Fig. 9, draw line 1 horizontally through 0 dB on the ordinate representing the peak energy level.

Step 3: Starting on line 1 at Δf_2 draw line 2 with a slope of -20 dB/decade.

Step 4: Starting on line 1 at Δf_3 draw line 3 with a slope of -40 dB/decade.

The spectrum bounds are described by lines 1, 2 and 3. To determine the accuracy of this approximation technique, the emission spectrum envelope was calculated. The calculated spectrum is plotted on Fig. 9 using the symbol "·" to designate representative points on the spectral power density function.

The peak energy density level corresponding to the 0 dB point is calculated as follows:

Assume: peak power (p) = 1×10^6 W, and

from equation (2):

$$p_d = (p\tau^2)$$
$$= (1 \times 10^6)(6 \times 10^{-6})^2$$
$$= 3.6 \times 10^{-5} \quad \text{Joules/Hz.}$$

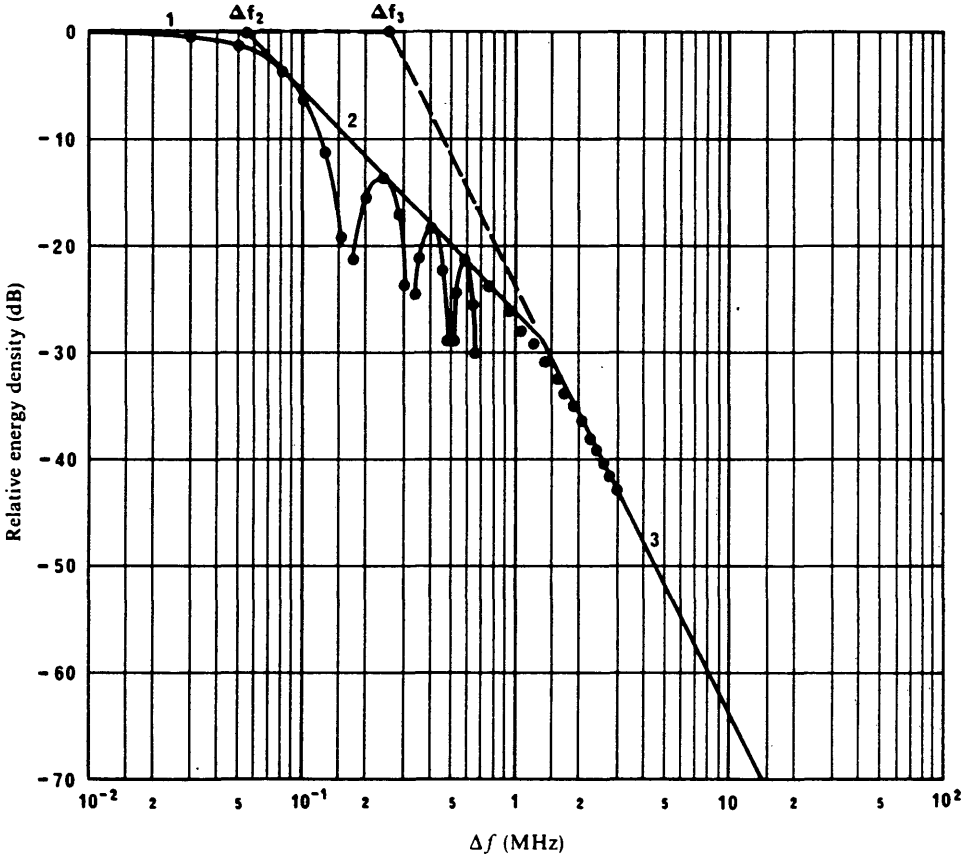


FIGURE 9 — *Relative energy density of a trapezoidal pulse in Example 1*

Example 2 – Linear chirp pulse

To illustrate the procedure for determining the approximate bounds on the spectrum of a typical chirp pulse, the following example is considered where the pulse parameters are:

$\beta = 1 \times 10^6$ Hz, positive deviation
$\tau_b = 102 \times 10^{-6}$ s
$\delta_r = \delta_f = 1 \times 10^{-6}$ s
$\tau = 101 \times 10^{-6}$ s

Calculations:

Step 1: Determine if $\beta > 2/(\pi\tau)$ and if the pulse is symmetrical.

$2/(\pi\tau) = 0.0063 \times 10^6$ which is less than β; and
$\delta_r = \delta_f$ and therefore the pulse is symmetrical.

Since the pulse meets these conditions calculate Δf_2, Δf_3, $\Delta f_{a\pm}$, and $\Delta f_{b\pm}$ according to Step 1(A).

Step 1(A):

$$\Delta f_2 = \pm \frac{1}{\pi}[1 \times 10^6 / 102 \times 10^{-6}]^{1/2} = 0.032 \times 10^6 \text{ Hz}$$

$$1/\delta = \frac{1}{2}\left[\frac{1}{10^{-6}} + \frac{1}{10^{-6}}\right] = 10^6 \text{ s}^{-1}$$

$$\Delta f_3 = \pm \frac{1}{\pi}[1 \times 10^6 / 102 \times 10^{-6}]^{1/4}[1 \times 10^6]^{1/2} = 0.10 \times 10^6 \text{ Hz}$$

$$\Delta f_{a\pm} = \frac{10^6}{2}\left[1 - \frac{10^{-6}}{102 \times 10^{-6}}\right] = 0.50 \times 10^6 \text{ Hz}$$

$$\Delta f_{b\pm} = 2\Delta f_{a\pm} = 1 \times 10^6 \text{ Hz}$$

Now plot the spectrum bounds as follows and as shown in Fig. 10.

Step 2: Draw line 1 horizontally through 0 dB.

Step 3: Is $\Delta f_b > 1/(\pi\delta)$?

$1/(\pi\delta) = 10^6/\pi = 0.32 \times 10^6$ Hz which is less than Δf_b. Therefore line 2 is not used in the approximation of the spectrum bounds. Proceed to step 4.

Step 4: Starting on line 1 in Fig. 10, at $\Delta f = \Delta f_3$, draw line 3 with a slope of -40 dB/decade.

Step 5: Locate point *a* 6 dB down at $\Delta f = \Delta f_{a\pm}$.

Step 6: Since $\Delta f_{b\pm} > 1/(\pi\delta)$, for this example, locate point *b* at $\Delta f = \Delta f_{b\pm}$ on line 3.

Step 7: Draw line 4 through points *a* and *b*.

The approximate bounds of the spectrum are formed by the solid lines 1, 4 and 3 on Fig. 10. Line 2 is not used in this particular example because $\Delta f_b > 1/(\pi\delta)$. The pulse is symmetrical therefore, the spectrum is also symmetrical and the plot can be used for either positive or negative values of Δf.

The peak energy density level corresponding to the 0 dB point is calculated as follows:

Assume: peak power (p) $= 1 \times 10^6$ W.

From equation (8):

$$p_d = [p\tau_b/\beta]$$
$$= [(1 \times 10^6)(102 \times 10^{-6})/1 \times 10^6]$$
$$= 10.2 \times 10^{-5} \quad \text{Joules/Hz.}$$

To indicate the accuracy attained in the above example, the spectral power density function, calculated using fast Fourier transform, is also plotted in Fig. 9 using the symbol "·" to designate representative points on the spectral power density function.

The accuracy of bounds obtained with the method described varies over the spectrum, being best for values of Δf of greater than 3β. At this value, the difference between the approximated bound and the theoretical bound is less than 1 dB. As Δf is increased further, the theoretical bound tends to approach line 3 asymptotically and the error rapidly approaches zero. In the region represented by line 1, the error usually is in the order of 1 dB.

The largest error occurs in the region represented by line 4 (Fig. 10). For long, symmetrical pulses, the error here is usually less than 6 dB, but for asymmetrical pulses, especially when τ_b is not much larger than either δ_r or δ_f, the error may be as much as 12 dB. Because line 4 is used to represent only a small part of the overall spectrum, recognition that an error can exist in this region is sufficient for most spectrum analysis.

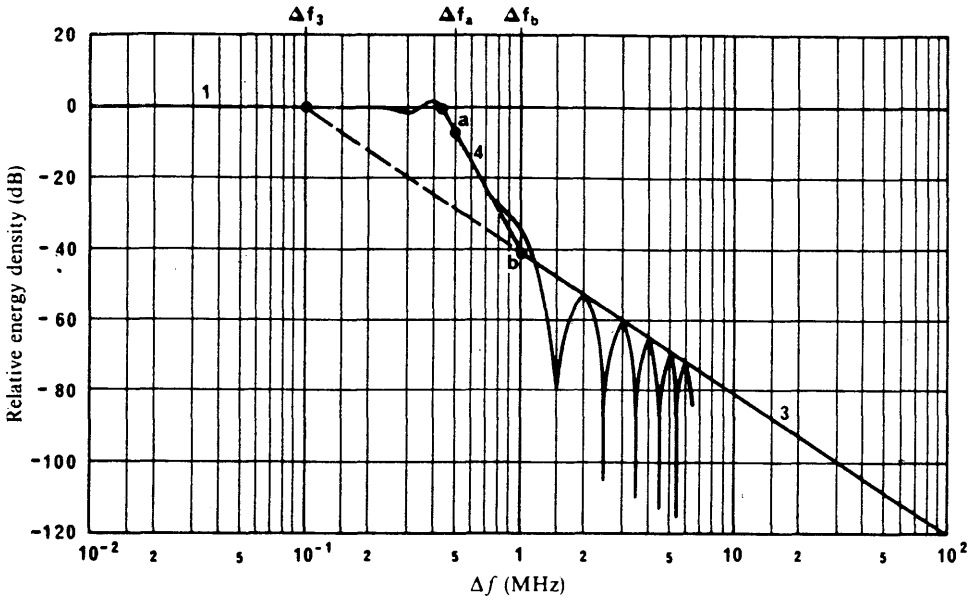


FIGURE 10 — *Relative energy density of a linear FM modulated pulse in Example 2*

4. Conclusions

It is important to recognize that the techniques presented above are first-order approximations of pulse radar emission spectrum shapes. However, the techniques summarized in this Report are of value in most spectrum analysis problems because of the straightforward calculations involved for a reasonably accurate spectrum approximation. Some examples of spectrum analyses to which these techniques would be applicable are: preliminary assessment of radar interference potential as described in Report 654; and radar frequency assignment procedures. These procedures can also be of value in assessing the spectrum requirements for existing radar systems.

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REPORT 1110

FREQUENCY MANAGEMENT DATA BASE SYSTEMS USING SMALL COMPUTERS

(Question 18/1)

(1990)

I. Introduction

The frequency management data base serves as the foundation of most frequency management activities. While a data base may consist of many kinds of data in various forms, CCIR IWP 1/2 has primarily been concerned with selected files stored in computer usable form.

Small computers can perform many of the functions required by frequency managers that have traditionally been performed by larger machines. However, the total size of the data files will, in some cases, be more restricted, requiring more careful planning of files and their contents. File processing will also be slower in some applications.

Small computers offer the frequency manager many attractive advantages, including ease of use and of maintenance, low cost, small size, minimum requirements for air conditioning or other special treatment, ease of use, reliability, and convenient support in most countries.

The growing capability of small computers (micro-computers), their low cost, and their wide availability make them especially attractive for many frequency management applications. Continuing advances make their use even more effective. This report considers frequency management data base implementation using these machines.

Data bases may take a variety of forms. While this paper emphasizes the use of automated (computerized) data bases, readers are reminded that in some cases, it will prove cost-effective or necessary to retain some data on the original paper forms. The CCIR IWP 1/2 has developed recommendations for frequency management data bases and the approaches discussed here rely extensively on these guidelines. Practical data base management systems must adapt these guidelines to meet specific requirements and constraints of individual administrations.

II. Data Base Definitions

Data base terminology is discussed in the Handbook and in other IWP 1/2 reports. Terms important to this report are briefly reviewed here.

Data Base: A collection of data referring to a common subject or a common purpose. Note that a data base is not necessarily automated. Data bases may be manual (paper documents, microfilm, etc.), automated (computer records), or a combination of both.

Data File: A collection of documents (or records), each dealing with the same subject and (usually) kept in the same storage location. In automated files and in some manual files, each document usually conforms to the same format.

Data Record: One document in a data file, usually containing information pertinent to one specific unit of the data file. In a frequency file, a data record typically will pertain to a single frequency assigned to one user at a specified location.

Data Element (field): A single item of data stored in a file. Several data elements make up a data record. Typical frequency management data stored in data elements are frequency, power, latitude, longitude, licensee name, etc.

Data Base Manager (Data Base Management System): A computer program that allows the creation, modification, and use of files in an automated data base.

III. Frequency Management Data Base Planning

Planning a frequency management data base system becomes an increasingly important and complex procedure as the size of the data base increases. Important considerations include which data items to include, the design of appropriate frequency application and licensing forms, and the procedures that will be used to assign frequencies, grant licenses, and authorize radio ownership and use, etc.

The data base to be used by a frequency management organization must recognize the special requirements of individual user organizations. [1]

The Master Frequency File is the most important part of a frequency management data base. This report emphasizes this type of file, although a number of other files will be included in some data bases. Note that while the master frequency file may be divided into several file segments, it is considered in this paper as a single file. (File segments will each contain similar information, but for different frequency bands, geographic locations, etc.)

A. Data Base Contents

While some data bases will contain a number of files, this paper concentrates on the master frequency file.

Primary frequency assignment and notification data file contents are described in table A IV-II of the Handbook. Most practical applications will require only some of the data items from the table. Choosing data items must be done with an understanding of the effect of overall file size, cost of creating and maintaining the resulting data file, and the speed and reliability with which data can be extracted from the file.

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[1] Perhaps the most important planning consideration is the existing data base and the procedures used to maintain and use it.

B. Record Formats and Record Sizing

It is important to plan record formats to insure that essential information is included, but that record size is kept to a practical minimum, especially in those cases where a large number of records will be kept in the mature data base.

The record format should be planned to contain essential information with needed precision. As record size grows, however, so too does the size of the total data base. Careful planning which recognizes the practical needs of the user can provide for efficient file record formats. Even greater efficiency is possible if the data base management program makes provision for inference rules, in which case a single data space (field) in the record can be used to store different kinds of information in different records. The data will be properly interpreted depending on the assigned frequency or other characteristics in the record.

Provision should be made in the size of a data element (field) to allow for needed precision. For example, if all frequencies are to be stored using MHz as units, the data element must be sufficiently large to provide for 4 decimal points if differences of 100 Hz are to be recognized.

C. File Size and File Structure

After the required data elements have been chosen, the way in which these data are arranged in the file can be decided. Two factors are especially important. First, are all records to be exactly the same size? "Same size" records are referred to as fixed length records. Variable length records allow each record to vary in length at increased costs in program complexity and running time. A fixed length record format is recommended for most frequency management data base systems.

A second factor to be considered is the supplementary information that will be required (and created) by the data base management system. The most common situation which creates this additional information is referred to as "file indexing", in which separate, sorted lists are created for each data element the frequency manager specifies as being of special interest. The special interest data elements are referred to as "keys" or "key fields". Frequency, for example, will be a special interest key field for every frequency manager.

The use of indexed files and key fields allows much faster access to required data. Indexing does increase the overall size of the file. Therefore, key fields should be selected with care to control the size of the file. Some data elements recommended as key fields for frequency managers include:

Frequency
Call Sign (or identification code or number)
License Number (a unique identifier)
Latitude
Longitude

These considerations are discussed further below. (See Data Access Considerations.)

The size of the data file is dependent on a number of factors that will vary depending on the method used to organize the file to meet different requirements for efficiency. These factors include:

The number and size of standard records.

The number and size of non-standard records (when variable length records are permitted.)

The number and size of key fields. This is the most important factor in determining file size after the basic record size has been determined. In figure 1, key field size is represented as a fraction of the record length. Thus the curve for a value of Key = 0.5 represents the case where the total size of all the key fields is half the record length. If record length is 512 bytes, then the total length of all keys is 256 bytes.)

The number and size of duplicate key fields. (Duplicate key fields are those where two or more records are allowed to have the same value for the key. Frequency fields are an example, since a frequency management data base should allow more than one record to contain the same frequency.)

The number and size of unique key fields. (Unique key fields are those in which no two records are allowed to have the same value. A serial number or unique identifier might be stored in such a field.)

Computer memory allocation to the indexing function. (The amount of memory allocated to organizing and maintaining file indexes and other organizational information will affect the speed of processing and the size of the files.)

The file size approximations in Figure 1 are based on a representative set of values for each of these factors. The number and size of key fields is represented as a fraction of the total record length. Actual file sizes will also depend on the combination of data in key fields. In practical cases, as much as a 50 percent reduction in the "key data" space will be achieved. The "key data" space is that space required for the file in excess of that needed to store only the original data. In Figure 1, the "key data" space is the difference between the upper curves and the bottom curve.

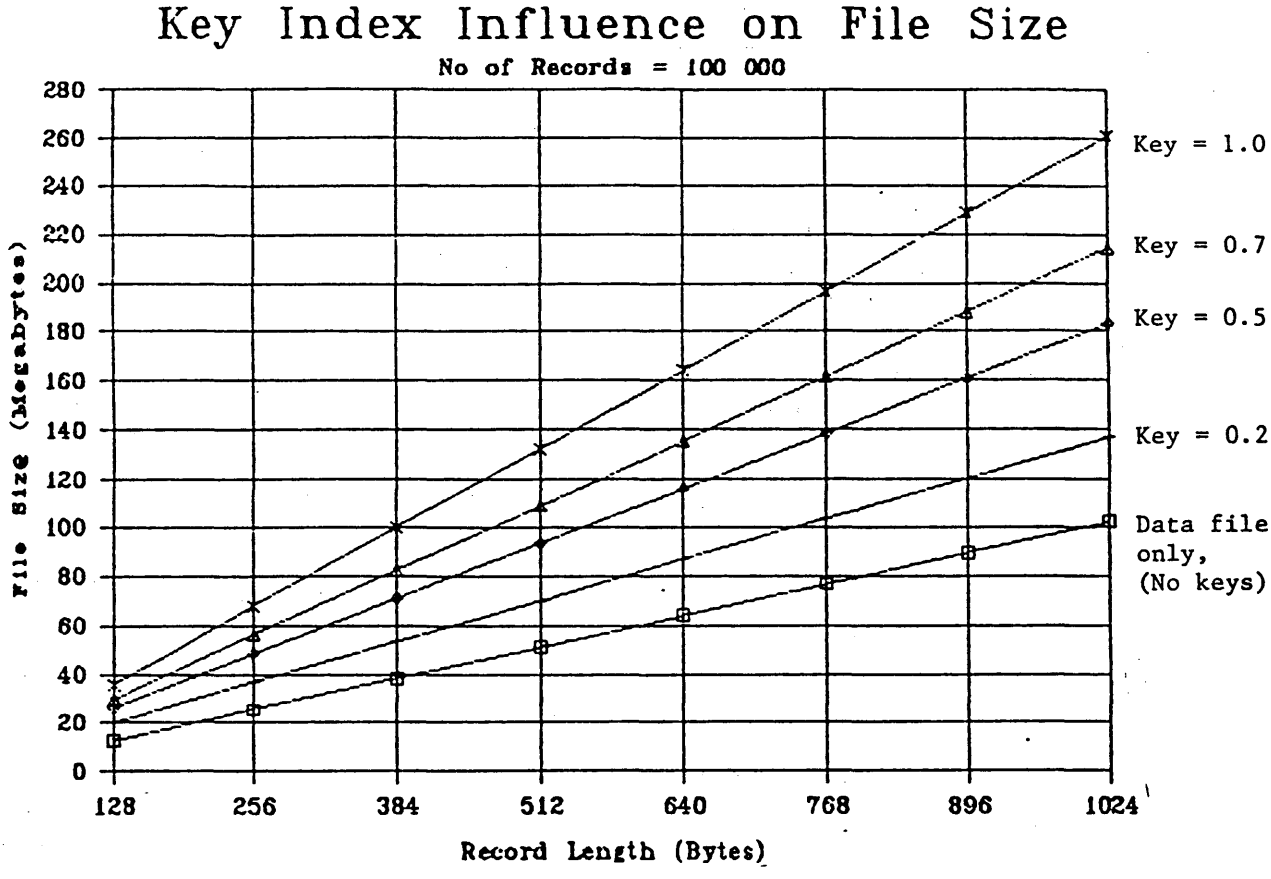

Key Index Influence on File Size
No of Records = 100 000
File Size (Megabytes)
280
260
240
220
200
180
160
140
120
100
80
60
40
20
0
128
256
384
512
640
768
896
1024
Record Length (Bytes)
Key = 1.0
Key = 0.7
Key = 0.5
Key = 0.2
Data file only, (No keys)

FIGURE 1

IV. Hardware Considerations in File Sizing

File characteristics will affect the choice of a computer system required to effectively create and use the data base. Conversely, the characteristics of an existing computer will impose constraints on the data files and the data base management system that can be used. Three equipment characteristics are of special concern: data storage capacity, addressable memory size, and processing speed.

Addressable memory size and processing speed will affect the speed with which large data files can be used to provide answers to specific requests. While high processing speed is a convenience, in many cases slower speeds can still provide cost-effective results, especially with moderate size data bases.

Data storage capacity must be large enough to store all the files in the frequency management data base (see Figure 2). However, not all files need to be "on-line" simultaneously. ("on-line" means that the computer has direct electrical access to the file contents at any time.) Effective frequency management procedures can be developed to use selected parts of the data files on-line, with the remainder stored on tape or as removable disks.

Small computer based data base systems can support very cost-effective frequency management data base requirements.

	Storage (Million Characters)----------	
Device	On-Line Storage	Total Storage
Floppy Disk	.36 - 1.2	Unlimited. *
Hard Disks, fixed	10 - 160	Same as on-line
Hard Disks, Removable	20	Unlimited. **
Tape	20 - 60	Unlimited. **

* Total off-line storage for these devices in unlimited, but storing and cataloging floppy disks will be very difficult.

** Off-line storage is practical for very large data bases.

FIGURE 2

Storage Capacities

Disks provide working storage for files that allow frequency managers quick access to any records of interest. Tape does not allow such rapid access, but is useful for data exchange and for long term, low cost storage of important data. (Such files, used for long term storage, are often referred toles as "archive" files.)

V. Software Considerations, Data Base Management Systems (DBMS)

Computer programs are needed in any automated data base to perform these essential functions:

- Data Entry
- Data Editing
- Data Storage
- Data Access and report generation
- Data Conversion (from existing files)

Programs to meet these needs may be developed using general purpose commercial data base management programs. A number of such systems are commercially available for small computers as well as for large machines. Special purpose data base management systems are also available designed specifically for the frequency manager. Of course, the frequency manager may choose to have a Data Base Management System developed by his own staff, although the development of such systems is both complex and costly.

Factors to be considered when choosing from among these options include:

- Total Cost of the system. Costs include the cost of purchase or development of the program, training costs, cost of maintenance, and costs associated with daily use of the program, such as data entry time, error rate, and error correction. In many cases, those costs incurred after software purchase can be substantially larger than the initial purchase cost. These costs should be seriously considered when planning for a frequency data base management system.

- Utility for Frequency Management.

Some data base management systems are intended for a wide range of users. The advantages of these general purpose programs lie principally in their low cost and flexibility. However, these general purpose programs impose significant disadvantages for frequency managers. Some of these programs impose undue limits on the number and size of records that can be supported. Running time is often excessive for practical frequency management applications. Perhaps most significantly, these programs require varying degrees of programming skill to use or to adapt to specific frequency management applications. As a result, higher skill levels are required of the users in some applications.

General purpose data base management systems also impose technical or functional limitations on frequency managers. For example, it is often very difficult to select records based on their distance from a designated location. This function, which is often used by frequency managers, can be very difficult or impossible to obtain in a general purpose DBMS.

As an alternative; data base management programs specifically designed for frequency management are becoming available. These frequency management DBMS overcome many of the limitations in general purpose systems. While they are more costly than the general purpose programs because of their much more limited market, they offer the frequency manager greatly improved ease of use, reduced training and start-up requirements, faster processing, and results formatted to meet the important frequency management needs. These programs make intensive use of the features of small computers and provide capabilities that can exceed those on bigger machines. Frequency management features like range-selection are included. Cost savings in reduced personnel expenses, reduced training,

and ease of use for both the professional frequency manager and supporting clerical staff will often make the use of such programs very cost-effective.

- Programmability

In some cases, the ability to change the formats of output reports or to otherwise revise the DBMS will be important. However, such revisions will still require a knowledgeable, trained person to make reliable changes. An effective alternative for those frequency management organizations with limited programming staff is to define a series of standard report formats to be provided by the DBMS supplier. Routine use of the DBMS can then be accomplished by making appropriate use of one or a combination of reports. In such cases, programming expertise will not be needed by the frequency management organization. Suppliers of such DBMS programs should provide for revised programs as future requirements dictate.

- File Compatibility

The DBMS program should allow for the use of data files prepared according to other standards. Similarly, output from the DBMS should be available in standard file formats for use by other programs.

- Processing and Access Speed

Small computers will usually provide slower processing speeds than larger machines, although speed differences are diminishing with new advances in small computers. A more important consideration is the ability to rapidly find one or a group of records through effective file organization. Good DBMS can provide rapid access to large numbers of records using efficient file organization procedures. Data base software for small computers should provide for efficient file organization with no special requirements imposed on the user other than the initial identification of key "Special Interest" data items.

VI. Data Access Considerations

A specific data item can be found in a standard file by checking the first data record, then the second, and so on until a matching data element is found. Such a procedure is very slow, requiring on the average a search of half the data file for each request. A more efficient means of finding required information is needed for any type of computer. A data base management system should provide for efficient procedures for finding data required by frequency managers.

Efficient file searching procedures rely on special kinds of file organization. This organization should be provided by the DBMS and not require special knowledge or procedures of the user.

Efficient file organization will require additional storage to provide supplementary file indexing information or special file organization. The amount of additional storage required will depend on the number and size of "key fields" chosen by the frequency manager when the data base is first designed.

VII. File Access Considerations, Multiple Users

Preparing input for a large data base file will usually require more than one operator working simultaneously. The DBMS should allow for mutliple users, at least for the preparation of input. Small computers can support this requirement in several ways. First, some small computers will allow simultaneous access to a common file from several other

computers, generally using Local Area Networks (LANs). Secondly, input files can be prepared on multiple, independent small computers and later merged into a single master file. This latter option represents the most cost effective approach in many cases. DBMS software should support whichever of these options is chosen.

VIII. Reporting

The most important function of the DBMS system is to provide the frequency manager with information extracted from the available files. Much of the complexity in a DBMS comes from the development of output reports.

Three report types meet the frequency manager basic requirements. The standard report presents a standard list of data elements for each record meeting the user's specifications. A number of standard report formats might be available to a frequency manager. Special reports are tailored to meet different requirements on a case-by-case basis. Universal reports provide output for the entire record, typically as a list of values for all data elements in a record, 1 record per page.

Contents of reports will be controlled by the selection criteria used (on a case by case basis) and by the format in which data from selected records are presented. Selection criteria should allow any field or combination of fields to be used to select desired records. However, recognition should be given to the efficiencies that will be obtained by using key fields.

A. Standard Reports

Standard reports present the output from selected records in a standard format. Data in selected elements of each selected record is displayed, typically on one of several lines of printed output. The use of standard reports allow lower cost software, easier use and training, and lower cost operation. There are no special requirements for programming expertise.

B. Special Reports

Special reports allow the user to change the format of the output display. This capability requires a more complex DBMS program and more programming skill on the part of the users.

C. Universal Reports

Universal reports display all of the contents of each selected record.

A combination of Standard Reports and Universal Reports will provide a cost effective way of meeting many frequency management data requirements with lower software costs and without the need for experienced programming staff.

IX. Data Collection and Data Input Considerations

Frequency management data is usually collected on paper forms which are stored as part of the paper (hard copy) archives portion of the data base. Proper design of these forms can provide for improved efficiency in data input processing. In most cases, the data

collection process is already taking place. Consideration must then be given to making the most effective use of existing forms. In some cases, it may be possible (and important) to revise the existing forms.

Data input procedures, including the data base software, should be designed to provide rapid, error free processing. Design principles to be observed include:

A. The input data sequence should be the same as the order in which data appears on the original form.

B. Data items on the paper form that are not to be entered into the computer record should be kept on a separate part of the form.

C. Data input units should be the same as those on the input form, even when provision is made for computer conversion of the input data.

D. Critical data items should be entered first.

E. The input program should provide automatic error detection and correction wherever possible. Well designed input programs will recognize permissible input values. Permissible ranges can be adjusted based on the values of critical input items. For example, once frequency is entered, allowable ranges for power or other data items can be automatically selected based on national regulations, commercial catalog values, etc.

F. Standard data should automatically be entered.
Standard data may be derived from national regulations, commercial catalog specifications, or the previous record. (It is important to design files so that standard data that can be readily obtained elsewhere is <u>not</u> included in the file. Even so, a number of records will contain duplicate data, as when several licenses are issued to the same organization. Each record may be identical except for frequency, location, and call sign.)

G. Data Automation Criteria

Efficient DBMS require careful consideration of data to be included in the data files. Recommended criteria for identifying data items to be included are:

- Functional need: There should be a clearly recognized need for the data item that can not be efficiently satisfied in some other way. One useful test is to identify those data items that are required for standard reports.

- Data item size. Large data items, such as general comments, should be avoided where possible, since they make files much larger and slow down processing.

- Redundant data. Data elements containing the same information for a large number of users should be saved once in a separate file and a suitable cross reference field included in the main file.

Select data that will play an active role in the day to day frequency management functions.

- Size of the data item.
- Format of the data item. (How standard is it?)

X. Frequency Management Data Files

The Data Base Management System should be designed with a consideration of each of the several types of files required by
the frequency manager. Major file categories include:

- Master Frequency File
- Licensing
  - equipment
  - users
- Equipment Technical Characteristics
- Spectrum Monitoring
  - spectrum usage
  - equipment operating characteristics
    - transmitted spectrum
    - antenna patterns
  - Interference (measured or observed)
  - Unauthorized usage
- Interference
  (Analysis, reported, special considerations)

XI. Data Base Management Systems for Frequency Management (FM-DBMS)

Special purpose Frequency Management - Data Base Management Systems (FM-DBMS) are now becoming available for the frequency management community. These programs, designed especially to run on small computers, provide the ability to establish and maintain data bases that meet the unique frequency management needs. Emphasis is on simplicity of use in recognition of the often limited support staff available.

REFERENCES

CCIR, Handbook on Spectrum Management and Computer-Aided Techniques, (Annex IV) Geneva, 1983, revised 1986.

CCIR IWP 1/2 has addressed the data base question in a number of documents; this paper refers particularly to the Handbook and documents IWP 1/2 - 170 and 1-186.

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SECTION 1B: SPECTRUM SHARING AND PLANNING PRINCIPLES AND TECHNIQUES

REPORT 1098*

SHARING BETWEEN THE LAND MOBILE SERVICE AND THE BROADCASTING SERVICE IN THE VHF AND UHF BANDS WHERE THE SERVICE AREAS ARE GEOGRAPHICALLY SEPARATED

(Question 45/1, Study Programme 45A/1)

(1990)

1. Introduction

The WARC-79 and the WARC MOB-87 provided new opportunities for sharing between the broadcasting service and the land mobile service in the frequency bands below 1 GHz.

Sharing between two or more services may be accomplished by:

- separation of the services by frequency within the shared band;
- separation of the services by time;
- separation of the services by geographical area.

This report addresses the third of these alternatives and provides means of determining the required geographical separation in order that adequate protection is afforded to both services. It does not address the implication of this sharing on the flexibility in TV allotment planning.

2. Frequency bands considered

The following frequency bands have been identified as having a sharing environment between land mobile and broadcasting services, world-wide, regionally, sub-regionally, nationally or in one country where in a neighbouring country stations in the land mobile service operate:

44 - 74 MHz, 76 - 108 MHz, 137 - 144 MHz, 167 - 238 MHz, 246 - 254 MHz and 470 - 960 MHz.

In the range 66 - 108 MHz, it is understood that sound broadcasting services operate whilst television services operate in the ranges 44 - 100 MHz, 137 - 154 MHz and 470 - 960 MHz.

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* This report should be brought to the attention of Study Groups 5, 6, 8, 10 and 11.

3. Method proposed

3.1 Sound broadcasting

The method and criteria proposed for this sharing situation are essentially these defined in the Final Acts of RARC-84, for land mobile services having permitted status, and take account of the frequency separations between the carriers of the broadcasting and land mobile emissions.

3.2 Television broadcasting

Here a number of narrow-band systems will be operating within a relatively wide broadcasting channel using the characteristics of the television signal to avoid undue interference. In the case of a high power broadcasting transmitter it will be necessary to avoid the vision and sound carriers for some considerable distance. In the reverse direction, the land mobile stations will produce a cumulative nuisance field in the broadcasting station's service area and a method is proposed in Annex II for calculating this parameter.

4. Sharing criteria

Annexes I and II contain the advised arrangements for sharing between the land mobile service and the broadcasting service in the VHF and UHF bands. Methods of assessing mutual interference and criteria to be applied are given, and particular areas requiring further study are identified.

5. Future draft Recommendation

A number of administrations present at the Interim Meetings were of the opinion that a Recommendation is required in the short term [CCIR, 1986-90], concerning sharing between the land mobile service and the broadcasting service especially for the requirements of Region 3 countries. Other administrations, however, were of the opinion that any thought of a Recommendation was premature at this time, since several issues are still under study, and particularly in consideration of the issue of relevancy of Recommendation 655 to sharing between land mobile and television, noting that the protection ratios given apply to television versus television interference only.

REFERENCES

CCIR Documents

[1986-90]: 1/33 (United Kingdom).

ANNEX I

PARAMETERS RELATING TO THE PROTECTION OF THE BROADCASTING SERVICE IN SHARED FREQUENCY BANDS IN RESPECT OF INTERFERENCE FROM THE LAND MOBILE SERVICE USING ANGLE MODULATION

1. The median field strength for which protection against interference is provided in planning should, in the case of television broadcasting, be as given in Recommendation 417:

- 48 dB(μV/m) in the range 44 - 100 MHz;
- 55 dB(μV/m) in the range 137 - 254 MHz;
- 65 dB(μV/m) in the range 470 - 582 MHz;
- 70 dB(μV/m) in the range 582 - 960 MHz,

at an antenna height of 10 m above ground level.

Different values of field strength and antenna heights may be considered, e.g., for providing a service to areas where lower field strength values may be appropriate (see Report 409) or for relay transmitters where higher field strength values may be appropriate or for certain city areas where typical building heights may be significantly higher than 10 m above ground level. This will require further study.

It is noted that currently world-wide a range of median protected field strength values are utilized. These are given below and include the values given in Recommendation 417 and Report 409, the values used for the 625-line (OIRT) system, those used in North America and those contained in the report to the Second Session of the Conference (Regional Administrative Conference for the Planning of VHF/UHF Television Broadcasting in the African Broadcasting Area and Neighbouring Countries, First Session, Nairobi, 1986 (RARC AFBC(1))):

- low VHF 46 - 48 dB(μV/m);
- high VHF 49 - 56 dB (μV/m);
- low UHF 58 - 65 dB(μV/m);
- high UHF 64 - 70 dB(μV/m);

Additionally at RARC AFBC(1) provisional figures for protection from land mobile services were also agreed of 46, 49, 53 and 58 dB(μV/m) respectively. However, Report 1023 states that the use of lower values than those given in Recommendation 417 militates against sharing.

2. The required protection ratios for television broadcasting should be determined from Recommendation 655 using the overlapping-channel curves for the case of interference from a CW or angle-modulated signal with non-controlled frequency. Where appropriate, allowance should be made for continuous interference conditions (see Annex I of Recommendation 655). Further investigation is required to confirm the appropriate values of protection ratios.

3. In the case of sound broadcasting the minimum protected median field strength for the VHF FM broadcasting service, as given in Recommendation 412, is:

- 54 dB(μV/m) for stereophonic service in the range 66 - 108 MHz;

- 48 dB(μV/m) for monophonic services in the range 66 - 108 MHz,

at an antenna height of 10 m above ground level.

Different values of protected field strength and antenna heights may be considered especially for providing service to urban or city areas. This will require further study.

4. For sound broadcasting the protection ratio should be as detailed in Report 659 (see also Recommendation 412).

5. The value of the protected field strength is calculated as follows:

a) The interference potential of a base station in the land mobile service should be determined by the concept of a "nuisance field". The nuisance field F dBμV/m, is calculated from:

$$F = E(50, T) + A + B;$$

where:

A is the protection ratio (in dB) determined in accordance with § 2 or 4 above;

B is the appropriate antenna discrimination (in dB) for the service and polarization in question. This is -15 dB in the case where the television broadcasting emission is horizontally polarized, this value having been agreed at RARC AFBC(1). However, in some countries in Region 2, a value of -9 dB is used. For the case where the broadcasting emission is vertically polarized, the curves given in Recommendation 419 should apply. For sound broadcasting the antenna discrimination should be calculated from the information given in Recommendation 599. Where mixed polarized signals are used, no polarization discrimination should be taken into account;

E(50, T) is the field strength (dBμV/m) of the interfering transmitter exceeded at 50% of the locations for T% (between 1 and 10%[1]) of the time with a receiving antenna height of 10 m, determined from Recommendation 370 and Report 239.

For transmitter effective antenna heights of less than 37.5 m, corrections should be applied to the appropriate curves in Recommendation 370.

In the VHF bands, for effective antenna heights of 20 m and 10 m, additional curves can be derived by applying -5 dB and -11 dB respectively for distances up to 25 km, and 0 dB in both cases for distances in excess of 250 km, with linear interpolation for intermediate distances. For effective antenna heights between 0 and 10 metres the values derived for 10 m are used (see RARC-84). For the UHF bands the case of effective antenna heights of less than

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[1] The precise value for T within the range 1 to 10% should be specified.

37.5 m requires further study. Different propagation calculation methods may be considered as a result of further study.

b) The effect of multiple interference arising from base stations operating at the same site is calculated by means of the power sum method:

$$e_o = \sqrt{\sum_{i=1}^{n} (F_i)^2}$$

where F_i is the nuisance field arising from the ith interfering transmitter expressed in $\mu V/m$.

c) The nuisance field arising from base stations operating at different sites together with the multiple interference from base stations operating at the same site should be calculated by means of the simplified multiplication method (see Report 945). Calculations should include all base stations within an appropriate coordination distance to be agreed between administrations.

d) In the VHF bands the nuisance field from land mobile stations are assumed to be around 20 dB less than the nuisance field arising from the corresponding base stations. For the UHF bands further study is required.

e) An appropriate factor requires developing to take account of the statistical probability that within the bandwidth of a television emission not all land mobile assignments are likely to be carrying traffic at the same time.

ANNEX II

PARAMETERS RELATING TO THE PROTECTION OF THE LAND MOBILE SERVICE USING ANGLE MODULATION IN SHARED FREQUENCY BANDS IN RESPECT OF INTERFERENCE FROM BROADCASTING SERVICES

1. The minimum protected median field strength for the land mobile service, using 25 or 30 kHz channel spacings, is given in Table I below:

TABLE I

Frequency range (MHz)	Field strength dB(μV/m) signal quality grade (see Report 358, and see Note 1 and Note 2, below)	
	3	4
44 to 68	16	19
68 to 87.5	15	20
87.5 to 108	14	20
137 to 174	14	21
174 to 254	14	21
470 to 582	20	24
582 to 960	30	38

For 12.5 or 15 kHz channel spacing the values should be 3 dB higher. For channel spacings greater than 30 kHz (wideband equipment), appropriate

values of minimum protected field strengths have still to be determined. The protected median field strengths given in Table I are preliminary and further work is required to confirm the appropriate values for various sharing situations.

2. Calculations for the interfering field strength should be carried out using Recommendation 370. Appropriate figures for 10% time and 50% of locations should be used. See also Report 567.

3. For signal quality grade 3, for sharing with the sound broadcasting service, the protection ratio should be as detailed in Table II below:

TABLE II

Frequency separation between carriers of the two services (kHz)	Protection ratio (dB) for the land mobile service using 12.5 kHz channel spacing
0	8
25	6
50	-5.5
75	-17.5
100	-27.5

In the case of sharing between television broadcasting and the land mobile service the protection ratio should be 10 dB.

Further study is required to determine the values of protection ratio for other signal quality grades and for other channel spacings used in the land mobile service.

4. In the case of sharing with television broadcasting the power measured in a reference bandwidth of 7 kHz with respect to the power notified for the vision carrier should be as detailed in Table III. The frequency separation between the vision and sound carriers should be as detailed in Report 624.

TABLE III

Assumed mask for television channel	Power relative to vision carrier (dB)
Sound carrier $<\pm 0.2$ MHz	-10
Sound carrier $>\pm 0.2$ MHz $<\pm 0.7$ MHz	-20
Sound carrier $>\pm 0.7$ MHz	-30
Vision carrier $>\pm 0.5$ MHz	-30
Vision carrier $<\pm 0.5$ MHz	0

The mask outlined in Table III should not constitute a restriction on the spectral content of the signal of the television service operating with its authorized characteristics, which should be agreed between the concerned administrations.

5. The values of antenna polarization discrimination for horizontally polarized broadcasting emissions are:

- 18 dB for base stations;

- 8 dB for land mobile stations.

Where vertical or mixed polarized broadcasting emissions are used no antenna polarization discrimination should be taken into account.

Note 1 - The following formula may be used to determine the power (dBm) induced into a 50-ohm load by an electric field (dBμV/m):

$$Pr = E(dB\mu V/m) - 20\log F(MHz) - L_c(dB) + G_r(dBi) - 77.2 \ (dBm)$$

where:

Pr = received power (dBm);

E = electric field strength (dBμV/m);

F = frequency (MHz);

L_c = cable losses between the antenna and receiver (dB);

G_r = receiver antenna gain (dBi).

Note 2 - Quality grade 3 produces "annoying" interference; quality grade 4 produces "noticeable" interference (see Report 358).

REPORT 828-1 *

THEORETICAL AND EXPERIMENTAL RESULTS FOR SPECTRUM SHARING BETWEEN FDM-FM AND RADAR SYSTEMS USING PULSE BLANKING WITH FREQUENCY DIVERSITY RECEPTION

(Question 45/1)

(1982-1986)

1. Introduction

During recent years, user demands on the electromagnetic frequency spectrum have increased significantly. All indications are that this growing requirement will continue at a more accelerated pace in the years to come. One approach that is being pursued by many investigators to help satisfy this increasing demand is frequency spectrum sharing.

This Report addresses some of the theoretical aspects, and examines some of the empirical data available, for the specific case of spectrum sharing between pulse radar systems and frequency division multiplex/frequency modulation (FDM-FM) communications systems by using receiver blanking techniques. In the general case, it is shown that it is difficult to develop simple analytic expressions which quantify the degree of communications interference created by radar transmissions. For certain conditions, however, reasonably accurate analytical approximations can be developed from theoretical analysis, and validated through laboratory measurement. For the case where the radar peak power is much greater than the FDM-FM carrier power it is shown that the interfering radar signal effectively and naturally blanks the output of an FM demodulator when the radar and FDM-FM carrier frequencies are equal. When the carrier frequencies are unequal, the demodulator output contains interfering pulses with amplitudes related to the frequency difference between the two carriers. For this case, intentional blanking of the demodulator output can be analytically shown to reduce the level of interference. Lastly, it is noted that pulse blanking is not necessarily appropriate for all of the operational modes of FDM-FM communications systems. Specifically, pulse blanking is not appropriate for FDM-FM systems that incorporate data and/or high quality picture signal transmissions in the vacant parts of the baseband.

2. Generalized interference model

The generalized interference model derived below investigates the processing by a FDM-FM receiver of a wanted plus an unwanted received signal. The carrier and interfering signals at the FDM-FM receiver input can be mathematically represented as follows:

$$C(t) = \sqrt{2C}\cos\psi(t) \tag{1}$$

$$I(t) = \sqrt{2I}\,M(t)\cos\varphi(t) \tag{2}$$

where $\psi(t)$ and $\varphi(t)$ are the instantaneous, time-dependent phases of the desired and interfering carriers respectively; C and I are the average FM carrier power and interfering radar peak power referred to the receiver input respectively; and, $M(t)$ is the envelope of the interfering carrier signal.

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* This Report should be brought to the attention of Study Groups 4, 8 and 9.

For an FDM-FM carrier,

$$\psi(t) = 2\pi \int_0^t [f_c + B(x)]\, dx \tag{3}$$

where f_c is the FDM carrier frequency and $B(x)$ is the baseband multiplex.

In general, the resultant signal at the input to the receiver can be represented by:

$$D(t) = \sqrt{2C}\, A \cos\left[\psi - \arctan \frac{R \sin(\psi - \varphi)}{1 + R \cos(\psi - \varphi)}\right] \tag{4}$$

where $D(t)$ is the combination of the wanted ($C(t)$) and unwanted ($\hat{I}(t)$) signals, and A and R are functions of time, related to the input signal parameters as follows:

$$A = [1 + R^2 + 2R \cos(\psi - \varphi)]^{1/2} \tag{5a}$$

$$R = \sqrt{\hat{I}/C}\, M(t) \tag{5b}$$

In an ideal frequency demodulator, the output voltage, E_o, is proportional to the time derivative of the phase difference of the input signals; thus,

$$E_o = \frac{k}{2\pi}\left\{\psi - \frac{d}{dt}\left[\arctan \frac{\rho \sin(\psi - \varphi)}{1 + \rho \sin(\psi - \varphi)}\right] - \omega c\right\} \tag{6}$$

$$= kB(t) + k\,[\rho' \sin\Delta + \rho\,\Delta' \cos\Delta + \rho^2\,\Delta'] / 2\pi\tilde{A}^2 \tag{7}$$

where $\Delta = \varphi - \psi$, $\rho = \sqrt{\hat{I}/C}\ \tilde{M}(t)$, and the functions $\tilde{M}(t)$ and $\tilde{A}$ represent the pulses after passing through the receiver IF stage. Equation (7) represents the output of a typical discriminator, provided the functions ρ, ρ' and φ realistically represent the characteristics of the radar, and the effects of dispersion are considered.

3. Empirically observed interference results

From equation (7) it is apparent, at least theoretically, that the output signal from the frequency demodulation section of a communications receiver is a function of the desired communications signal and the unwanted interfering signal. Several organizations have experimentally investigated this phenomenon by measuring the interference noise power (*INP*) levels and/or the signal and noise power densities observed within a particular frequency slot of a communications system. Table I summarizes the empirical data collected by Wachs and Arroyo [1978] on the interference effects of pulsed radars on 24, 252 and 432 channel FDM-FM communications systems (this data was observed when the radar and the communications carrier signal frequencies were equal (i.e. $\Delta f = 0$)). The first column in Table I indicates the number of FDM-FM channels in the observed communications systems. The second to fourth columns note the modulation characteristics of the pulsed interference introduced into the communications system. The fifth and sixth columns give the noise power ratio (*NPR*)* observed in both the top and bottom channels of each communications system when the peak radar signal power ($\hat{I}$) was much greater than the average communications signal power (C). Figure 1, which is a graphical display of both measured and computed data from the work of Wachs and Arroyo [1978], illustrates the impact of varying the communications signal power to peak radar power ratio on the observed *NPR*.

* The noise component of the noise power ratio reported by Wachs and Arroyo [1978] includes white noise, IF thermal noise and pulsed interference. For this reason, this *NPR* may not be exactly equivalent to the *NPR* as defined by the CCIR.

TABLE I - *Measured values of NPR from Wachs and Arroyo [1978]*

FDM-FM channel (1)	Radar duty			Top channel	Bottom channel
	PW (μs) (2)	*PRF* (kHz) (3)	Duty cycle (dB) (4)	NPR_{max} (dB) (5)	NPR_{max} (dB) (6)
24	1.0	1	-30	36	28
24	10.0	1	-20	25.5	17
24	1.0	0.3	-35.2	(1)	35
24	1.0	3	-25.2	(1)	22.5
252	0.1	1	-40	44	37
252	1.0	1	-30	34.5	28
252	10.0	1	-20	27	22.5
252	1.0	0.3	-35.2	40	35.5
252	1.0	3	-25.2	29	23.5
432	0.1	1	-40	42.5	36
432	1.0	1	-30	33.5	29.5
432	10.0	1	-20	28.5	23
432	1.0	0.3	-35.2	40	35
432	1.0	3	-25.2	29	24.5

(1) No data available.

PW: Pulse width.
PRF: Pulse repetition frequency.

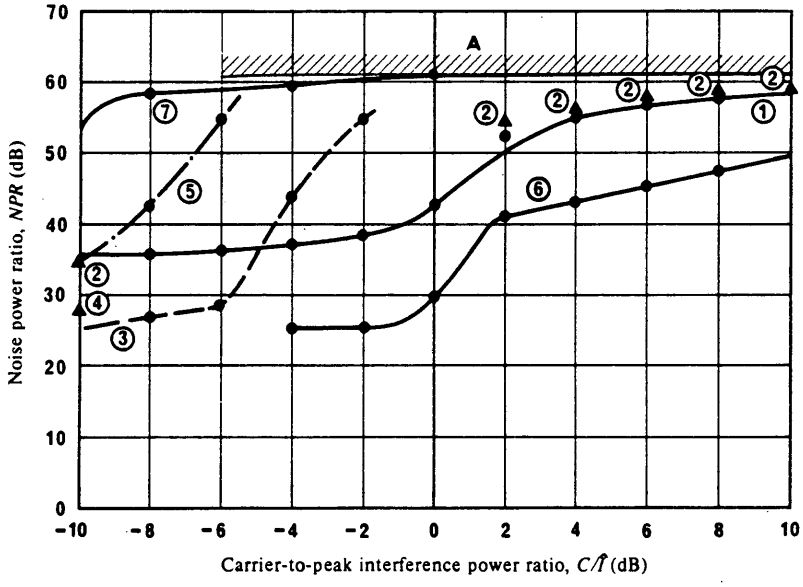


FIGURE 1 - *Noise power ratio vs. C/Î from Wachs and Arroyo [1978]*

24 channel FDM-FM, carrier-top channel; radar *PRF* = 1 kHz; channel bandwidth = 4 kHz

A measurement noise floor

① Measurement, radar co-frequency, τ = 1 μs

② Computation, radar co-frequency, τ = 1 μs

③ Measurement, radar frequency offset = 1.5 MHz, τ = 1 μs

④ Computation, radar frequency offset = 1.5 MHz, τ = 1 μs

⑤ Measurement, radar frequency offset = 2.0 MHz, τ = 1 μs

⑥ Measurement, radar co-frequency, τ = 10 μs

⑦ Measurement, radar co-frequency, τ = 0.1 μs

Table III presents a partial listing of the data collected by Hernandez [1978] on the interaction between radar modulation characteristics and observed interference in a 600 channel FDM-FM communications system. The data in Table III reflects the maximum interference noise power (*INP*) * observed in a lower channel of the communications system while the communications and interference carrier signal frequencies were equal ($\Delta f = 0$) and the interference signal power was varied (increased for all other conditions remaining constant). Plotting the observed maximum interference noise power as a function of the various radar modulation characteristics yielded the following form of a general estimating equation:

$$INP_{max}\,(\text{pW0p}) = C_1\ PRF^{C_2\ PW^{C_3}} \tag{8}$$

where C_1, C_2 and C_3 are constants for a given pulse width, and the units of *PRF* and *PW* are Hz and seconds respectively. Using Hernandez's data it is possible to calculate regression coefficients approximating C_1, C_2 and C_3 for each of the observed pulse widths. Substituting the regression coefficients into the general estimating equation yields the following:

TABLE II

Pulse width (μs)	Equation
1	$INP_{max}\,(\text{pW0p}) = 1.770\ PRF^{1.3\ PW^{0.0468}}$
5	$INP_{max}\,(\text{pW0p}) = 1.020\ PRF^{1.3\ PW^{0.0468}}$
100	$INP_{max}\,(\text{pW0p}) = 3.165\ PRF^{1.3\ PW^{0.0468}}$
400	$INP_{max}\,(\text{pW0p}) = 0.280\ PRF^{1.85\ PW^{0.0408}}$
1000	$INP_{max}\,(\text{pW0p}) = 0.986\ PRF^{1.85\ PW^{0.0408}}$

The graphical data presented by Hernandez [1978] suggests that the interference noise level in a communications system increases as the signal-to-interference power ratio decreases. When the carrier frequencies of the communications and interfering signals are equal ($\Delta f = 0$), the interference noise level achieves a maximum value at approximately $C/\hat{I} = -2$ dB, and remains constant (at the maximum value) for $C/\hat{I} < -2$ dB. When the communications and interfering pulse carrier frequencies are not equal ($\Delta f > 0$), Hernandez [1978] observed that the point at which the maximum interference noise power is observed shifts away from $C/\hat{I} = -2$ dB ($C/I = C/\hat{I} -$ (radar duty cycle), the three terms of the equation being expressed in decibels). It is clear from Fig. 2 and the other data presented by Hernandez [1978], that offsetting the radar and communications carrier frequencies increases the interference observed; but, the precise mathematical relationship between the magnitude of the frequency offset and the interference observed is not readily determinable.

Hernandez [1978] also investigated a limited number of situations (six) in which direct comparisons could be made between the *INP* that can be expected for nonchirped and chirped (frequency modulated) radar pulses of the same effective levels of interfering power. Inspection of this data indicates no significant additional interference in the communications channels is attributable to radar chirping.

* *INP* is the interference power in a voice channel measured in units of pW0p; INP_{max} is the maximum interference power observed in voice channel measured in units of pW0p.

TABLE III - *Maximum interference noise power observed in a lower channel of a 600 channel FDM-FM communications system*

Duty cycle (dB)	Pulse width (μs)	Pulse repetition frequency (Hz)	INP_{max} (pW0p)	NPR_{cal} (dB)
-43	1	50	130	50.2
-36	5	50	240	47.5
-33	1	500	2 500	37.3
-30	100	10	130	50.2
-28.2	5	300	3 000	36.6
-24	100	40	1 300	40.2
-23	5	1 000	15 000	29.6
-23	1	5 000	39 000	25.4
-21	100	80	3 000	36.6
-20	1 000	10	280	46.9
-18.2	5	3 000	80 000	22.3
-18.2	1	15 000	200 000	18.3
-18	400	40	1 700	40.0
-15	400	80	13 000	30.2
-14	1 000	40	3 000	36.6
-14	100	400	50 000	24.3
-11	1 000	80	45 000	24.8
- 8	1 000	160	240 000	17.5
- 8	400	400	390 000	15.4

INP_{max} is the maximum interference power observed in a voice channel measured in units of pW0p.

NPR_{cal} was calculated using the following formulae from [Oliver, 1964]:

$$NPR_{dB} = S/N_{dB} - 16.2$$

where:

$$S/N_{dB} = -2.5 + 90 - 10 \log INP$$

INP is the observed interference noise power in units of pW0p. Interference power rather than noise power was used to calculate S/N_{dB}; therefore, in the above equations S/N_{dB} is actually a signal-to-interference ratio.

4. Blanking as a method of pulse interference suppression

As can be seen from examination of equation (7), interference exists only during periods of time when ρ and $\rho' \neq 0$, that is, when radar pulses are present at the input of the communications receiver. The interfering intervals will, in general, depend upon the shape of the radar pulses after passing through the receiver pass band; for example, pulses at the input to the receiver will be distorted when the receiver pass band is narrow with respect to the pulse spectrum, or when the pulse carrier frequency is offset with respect to the receiver band-centre. Obviously, the effective interference will be influenced in both instances; but, the degree of susceptibility of the baseband to such parametric changes is difficult to ascertain from equation (7). The parametric variation between the individual interfering components (ρ, ρ' sin Δ, etc.) as well as the fact that the spectrum consists of the sum of several convoluted integrals, makes it difficult to develop a simple generalized analytic expression for predicting the effects of pulse interference on an FDM-FM carrier. It is possible though to outline the conditions under which intentional blanking, that is, the deliberate suppression of the input signal, can logically be considered as a means of communication signal enhancement.

When the desired carrier power is of comparable, or larger magnitude than the peak radar power, Wachs and Arroyo [1978] have shown that for any given channel total signal degradation decreases, dB for dB, as the ratio of the peak radar power-to-communications power decreases. Under these conditions, that is $C/\hat{I} \geqslant 0$ dB the quality of the signal improves in response to frequency offsets between the radar and communications carrier signals. It is apparent that intentional pulse blanking of the input signal when $C/\hat{I} > 0$ dB destroys potentially recoverable communications information, and should therefore not be considered an appropriate signal enhancing technique under the stated conditions.

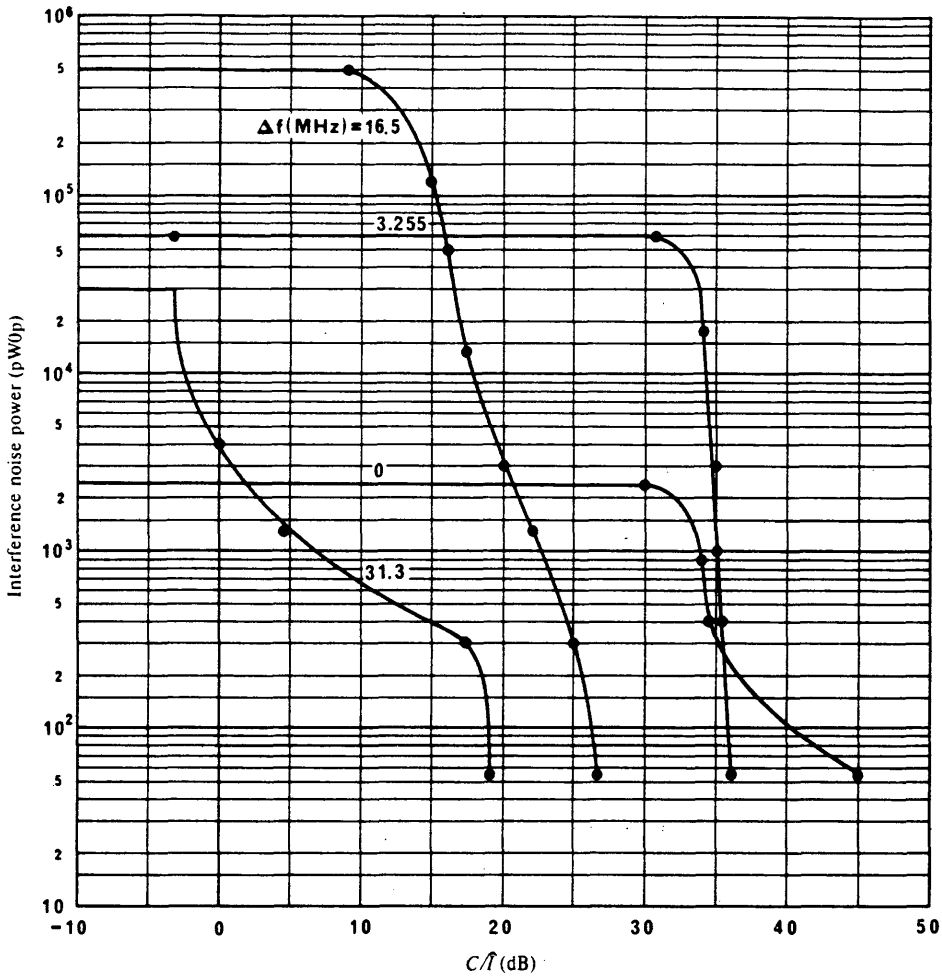


FIGURE 2 - *Measured interference power in a low baseband channel (340-344 kHz) for co-channel ($\Delta f = 0$) and off-tuned ($\Delta f > 0$), non-chirped, pulse interference; PW = 1 μs, PRF = 500 Hz*

On the other hand, when the peak radar power is much greater than the average communications signal power and $\varphi(t) = 2\pi(f_c + f_d)$ it can be shown from equation (7) that:

$$E_o \simeq \begin{cases} k\,B\,(t) & \text{for } \rho = 0 \\ k f_d & \text{for } \rho \gg 1 \end{cases}$$

where f_d is the frequency difference between the interfering and FDM-FM carrier frequencies. When $f_d = 0$, and $\rho \gg 1$; $E_o \simeq 0$. Thus, during periods of time when $\rho \gg 1$, it is apparent that:

- for $f_d = 0$, the baseband signal is naturally and automatically blanked by the interfering signal, or
- for $f_d \neq 0$, the communications signal contains interfering pulses with amplitudes proportional to the frequency difference between the FDM-FM and radar carrier signals.

It therefore appears possible to reduce the degrading effects of pulse interference when $\rho \gg 1$, and $f_d \neq 0$ by intentionally blanking the output of the FM demodulator during periods of time when interfering pulses are significant. Furthermore, when $\rho \gg 1$, intentional blanking of the demodulator output will never worsen the net interference; but, will as a minimum reduce it to the level which will exist when the frequency difference between the two carriers is zero. This presumes, of course, that a receiver can be designed, or modified to track radar pulses in the presence of an FDM-FM carrier. For certain situations, a sampling antenna/receiver system at the FDM-FM communications station, similar to current side-lobe cancellers, should permit intentional blanking.

5. Estimates of voice channel interference during blanking operations

It was noted in the previous section that when the communications and radar carrier frequencies are not equal, intentional blanking of the received signal during periods of pulse transmission can be employed to reduce the degrading effects of the interference compared with the case with no blanking. By blanking all received signals during periods of pulse interference reception, the maximum interference noise in the communications system can be limited to the maximum level observed when the communications and radar carrier frequencies are identical ($\Delta f = 0$). This is a theoretical deduction; to date, no experimental results have been published which relate the affect of intentional pulse blanking on observed interference noise levels.

With a view towards analytically quantifying the expected interference in an FDM-FM communications system, under conditions of intentional blanking, Byron and Berry [1978] derived two measures of noise for the case when the emitted radar signal is periodic and the output of the demodulator is assumed to be intentionally blanked. One of the measures is the mean squared signal-to-interference ratio (herein referred to as *NPR'*), and the other is the mean squared interference power (herein referred to as *INP'*). (*NPR'* is a signal-to-interference ratio, and should not be confused with *NPR*, which is a noise-to-noise ratio. In the blanking case *NPR'* depends only on radar parameters and the multiplex bandwidth, while *NPR* is a function of communication system parameters, e.g. noise load ratio, system non-linearities, system temperature, etc.) For any communications channel j, it can be shown that:

$$\begin{aligned} NPR'_j &= C_j^{-1} [(1 - (PW)(PRF))/(PW)(PRF)]^2 \\ &= C_j^{-1} \left[(1 - \delta)/\delta\right]^2 \end{aligned} \tag{9}$$

where:

PW is defined to be equal to the blanking time interval,

PRF is the pulse repetition frequency of the interfering radar signal,

$\delta = (PW)(PRF)$, and

$$C_j = \left(1/3100\, P_j^2\right) \sum_{n=1}^{\infty} \sum_{k=0}^{N-1} \left(\frac{\sin \pi n\delta}{n \pi \delta}\right)^2 P_k^2 \int_{f_{cj}}^{f_{cj}+4000} \left[G\left(f - |f_{ck} + n(PRF)|\right) + G\left(f - |f_{ck} - n(PRF)|\right)\right] df \tag{10}$$

In equation (10):

G: is the power spectral density function,

f_{ck}: the lower cut-off frequency of any channel, k, (for the results contained in Table IV, it was assumed that $f_{ck} = 12 + 4r$, kHz, $r = 0 \ldots N - 1$), and

P_k: the pre-emphasis factor of any channel, k, and can be determined from the expression listed below.

$$P_k^2 = \sqrt{10} \Big/ \left\{1 + 6.9 / \left[1 + 5.25 / (f_0/f_{ck} - f_{ck}/f_0)^2\right]\right\} \tag{11}$$

where, f_0, is the resonant frequency of the pre-emphasis network. For this Report $f_0 \equiv 1.25\,(f_c + 4N)$, in kHz. Equation (11) was derived from Recommendation 275.

Values of interference noise were calculated from the following expression:

$$INP'\,(\text{pW0p}) = \text{antilog}\left[\frac{(90 - 2.5 + BBP - NPR')}{10}\right] \tag{12}$$

where *BBP* is the baseband power per channel, and:

$$BBP\,(\text{dB}) = \begin{cases} -15\ \text{dB} & \text{for } C > 240 \\ -1 - 6 \log C & \text{for } C < 240 \end{cases} \tag{13}$$

where C is the number of communication channels.

Equations (9) and (12) were evaluated for several specific cases; the results are summarized in Table IV.

TABLE IV - *Voice channel interference as a function of radar parameters and channel load*

Channel number	Number of channels	Duty factor (dB)	*PRF* $\times 10^3$ (Hz)	Pulse width (μs)	*INP'* $\times 10^3$ (pW0p)	*NPR'* (dB)
1	24	-17.4	0.3	60	711.2	19.7
24	24				619.4	20.3
1	60				400.9	19.8
60	60				374.1	20.1
1	120				264.2	19.8
120	120				252.3	20.0
1	24	-24.4	0.3	12	214.8	24.9
24	24				744.7	29.5
1	60				98.4	25.9
60	60				54.1	28.5
1	120				61.9	26.1
120	120				39.1	28.1
1	24	-29.2	0.3	4	112.7	27.7
24	24				14.5	36.6
1	60				54.1	28.5
60	60				12.4	34.9
1	120				27.0	29.7
120	120				10.5	33.8
1	24	-32.2	0.3	2	35.6	32.7
24	24				5.5	40.8
1	60				36.6	30.2
60	60				4.3	39.5
1	120				18.8	31.3
120	120				4.0	38.0
1	24	-13	1	50	1959.0	15.3
24	24				1629.0	16.1
1	60				1078.9	15.5
60	60				1006.9	15.8
1	120				711.0	15.5
120	120				679.0	15.7
1	24	-20	1	10	695.0	19.8
24	24				186.0	25.5
1	432				126.0	21.5
432	432				81.3	23.4
1	24	-27	1	2	115.6	27.6
24	24				18.9	35.6
1	60				123.9	24.9
60	60				14.3	34.3
1	120				61.9	26.1
120	120				13.5	32.8
1	24	-30	1	1	31.0	33.3
24	24				5.3	41.0
1	60				42.0	29.6
60	60				5.8	38.2
1	120				40.9	27.9
120	120				4.7	37.3
1	252				21.9	29.1
252	252				4.8	35.7
1	24	-24	2	2	235.5	24.5
24	24				35.7	32.7
1	60				247.2	21.9
60	60				28.4	31.3
1	120				126.5	23.0
120	120				26.4	29.8
1	24	-27	2	1	61.9	30.3
24	24				10.5	38.0
1	60				83.8	26.6
60	60				11.6	35.2
1	120				81.7	24.9
120	120				9.4	34.3
1	24	-20	10	1	303.4	23.4
24	24				50.4	31.2
1	60				419.8	19.6
60	60				55.3	28.4
1	120				418.8	17.8
120	120				45.9	27.4
1	24	-23	10	0.5	76.2	29.4
24	24				12.9	37.1
1	60				113.0	25.3
60	60				17.1	33.5
1	120				138.7	22.6
120	120				18.3	31.4

As mentioned previously, the empirical data presented in § 3 does not reflect the results of an operational pulse blanking system; but it does provide general information on the interaction of pulse interference on communications circuits as a function of pulse modulation characteristics, signal-to-interference ratios, interference noise power and noise power ratios. This information is useful in structuring the conditions under which intentional pulse blanking is appropriate and recognizing the potential for improvement in the quality of communications. Further study is required, however, to determine in a precise manner if the FDM-FM system is adversely affected in other ways by the implementation of pulse blanking.

Figure 1 illustrates the non-linear relationship between the observed (and calculated) noise power ratio and the signal-to-peak interference ratio for several sets of radar modulation characteristics. This non-linearity suggests a natural boundary for the application of pulse blanking techniques; for instance, it is clear that when the carrier-to-peak interference ratio is large, observed noise is low; therefore suggesting that communications signal recovery is feasible. Intentional blanking in this area would degrade the quality of the communications. It is equally clear that when the carrier-to-peak interference ratio is low, the observed noise is high; therefore, the potential for communication signal recovery is severely diminished. Although it has not been specifically demonstrated in the referenced empirical investigations, it is reasonably asserted that when the $C/\hat{I}$ is less than approximately -2 dB, and the communications and radar carrier frequencies are equal ($\Delta f = 0$), the radar signal effectively and automatically blanks the communications signal by capturing the receiver demodulator section. When the communications and radar carrier frequencies are not equal, Fig. 2 indicates that the interference noise level is greater than that observed when $\Delta f = 0$ and therefore, under these specific conditions consideration of intentional pulse blanking seems to be appropriate.

Insufficient data exists regarding the transition area, where the communications signal power and the peak radar power are approximately equivalent, to suggest whether or not intentional pulse blanking can enhance the quality of the communications signal.

In the area of interest to pulse blanking (that is $C/\hat{I} < -2$ dB) it is shown in Tables I and III that the maximum observed interference noise in the communications system is a function of the modulation characteristics of the interfering signal. Equation (8) was derived from the empirical data presented by Hernandez [1978] and provides insight into the determinates of interference noise in the FDM-FM communications system under consideration.

Direct comparison of the analytical and empirical results, or for that matter, even the direct comparison of empirical results can only be accomplished on a limited basis. The predominate reasons for this situation are:

- there is no empirical data available on the impact of intentional pulse blanking on an FDM-FM communications system;
- the text conditions for the various referenced empirical investigations differ significantly in terms of the number of channels in the communications system observed, channel investigated, the modulation characteristics of the interfering signal, etc.;
- the measurement standards utilized (*NPR*, *NPR'*, *INP*, *INP'*) are dissimilar enough to void direct comparison.

Even without direct comparisons of observed and derived values, it is possible to recognize that for the purposes of pulse blanking the empirical and analytical data are in basic agreement.

6. Effectiveness of pulse interference cancellers in frequency diversity reception systems

In FDM-FM systems using frequency diversity reception, other methods more effective than simple pulse blanking can be used for purposes of interference suppression. One such method is described in Report 830. The method is based on the introduction of a time delay in one group of frequency diversity channels on transmission and in the other group of channels on reception, as a result of which different bits of information in each group of channels are affected by pulse interference at the same moment of time. By blanking the received signal of the group of channels which is affected by interference at a given time, the information to be transmitted can be restored by coherent addition of the signals only in the channel group in which pulse interference does not occur at the same given time.

The effectiveness of this method in connection with radio-relay systems has been studied by Bernoskuni *et al.* [1984]. In keeping with Recommendation 397, the quality of sound channels in a trans-horizon system has been defined by the following two parameters: mean noise power per minute $P_{T,p\%}$ for 20% of the time and by the percentage of time T_m for which the unweighted noise power in the sound channel P_Σ exceeds the value of $P_{m0} = 10^6$ pW.

The estimation was carried out with the following assumptions. In the first place, no account was taken of the interference caused by electronic switching of the channel group during the pulse interference. In the second place, it was assumed that the pulse interference was always detected by the detection circuit. In the case where the interference level significantly exceeds the wanted signal level (i.e. the most important case), the identification of the interference by the detection circuit does not cause any difficulties, and the above assumption can be regarded as valid. Although interference is naturally difficult to detect at low levels, its effect on reception noise immunity is insignificant in this case.

As is pointed out in Report 830, the method is based on the fact that during the period of pulse interference, there is a coherent addition of only half ($N/2$) of the total number (N) of the frequency-diversity-received signal components. Hence, for the mean thermal noise power per minute at the communication system output exceeded for not more than $p\%$ of the time, we can write:

$$P_{T,p\%} = (2\,\tau_d/T)P_{T,p\%}(N/2) + (1 - 2\,\tau_d/T)\,P_{T,p\%}(N) \tag{14}$$

where:

$P_{T,p\%}(N)$ and $P_{T,p\%}(N/2)$: mean minute thermal noise power at the frequency detector output in a coherent signal addition system with N and $N/2$ diversity branches, exceeded for $p\%$ of the time;

τ_d: channel group blanking time interval during the effect of pulse interference of duration $\tau(\tau_d \geqslant \tau)$;

T: pulse repetition period.

The unweighted noise power P_Σ at the demodulator output in a coherent signal addition system for given values of n_1 and n_2 – signal-to-noise ratios at the outputs of two space diversity reception channels – may be written as follows:

$$P_\Sigma = (\tau_d/T)\left[P_{T1}(n_1) + P_{T2}(n_2)\right] + (1 - 2\,\tau_d/T)P_{T3}(n_3) \tag{15}$$

where $n_3 = (n_1 + n_2)$. The values of n_1 are random. Their fluctuations are due both to "rapid" interference signal fades and to "slow" fades. The values of $P_{Ti}(n_i)$ represent the noise power at the follow-up demodulator output, with a signal-to-noise ratio at its input equal to n_i. By using equation (15) one can estimate the value of $T_m(\tau_d/T)$ – percentage of time, T_m, for which the unweighted noise power in the sound channel P_Σ exceeds the value $P_{m0} = 10^6$ pW.

Once $P_{T,p\%}$ and T_m have been defined, the effectiveness of the method can be assessed by means of the coefficient:

$$\Delta P_T(\tau_d/T) = 10 \log\left[P_{T,p\%}/P_{T,p\%}(N)\right] \tag{16}$$

and:

$$\tilde{T}_m = T_m(\tau_d/T)/T_m\left[(\tau_d/T) = 0\right] \tag{17}$$

The first coefficient indicates the number of decibels by which the mean thermal noise power per minute increases at the system output during the occurrence of pulse interference. The second coefficient refers to the increase at the system output of the percentage of time for which the unweighted noise power exceeds 10^6 pW during the occurrence at its input of pulse interference with a given pulse duty factor τ/T.

Equations (14) to (17) were used to calculate, as an example, the relations $\Delta P_T(\tau_d/T)$ and $\tilde{T}_m(\tau_d/T)$ for $N = 4$ and 8 for a trans-horizon radio-relay link of 120 channels and length 2400 km, consisting of seven hops. The transmitter power was 5 kW, with a receiving and transmitting antenna gain of 47.1 dB. The values of $P_{T,p\%}(j)$ $P_{Ti}(n_i)$ appearing in equations (14) and (16) were calculated by the method described in [Borodich, 1981].

The calculation results are given in Fig. 3. It follows that for values $(\tau_d/T) \leqslant 10^{-2}$ (a case frequently encountered in practice), this suppression method completely eliminates the effect of pulse interference on the operation of trans-horizon radio-relay links. Even at high values of $\tau_d/T = 10^{-1}$, the impairment of the channel performance parameters due to interference is insignificant ($\Delta P_T = 0.2$ dB for $N = 8$; $\tilde{T}_m = 2$). Since only one or two stations forming part of the link are normally affected by pulse interference simultaneously, the effect of the interference on the performance parameters of the whole link consisting of a number of stations will be imperceptible. Figure 3 also shows a considerable increase in reception noise immunity for an increase of the reception factor N.

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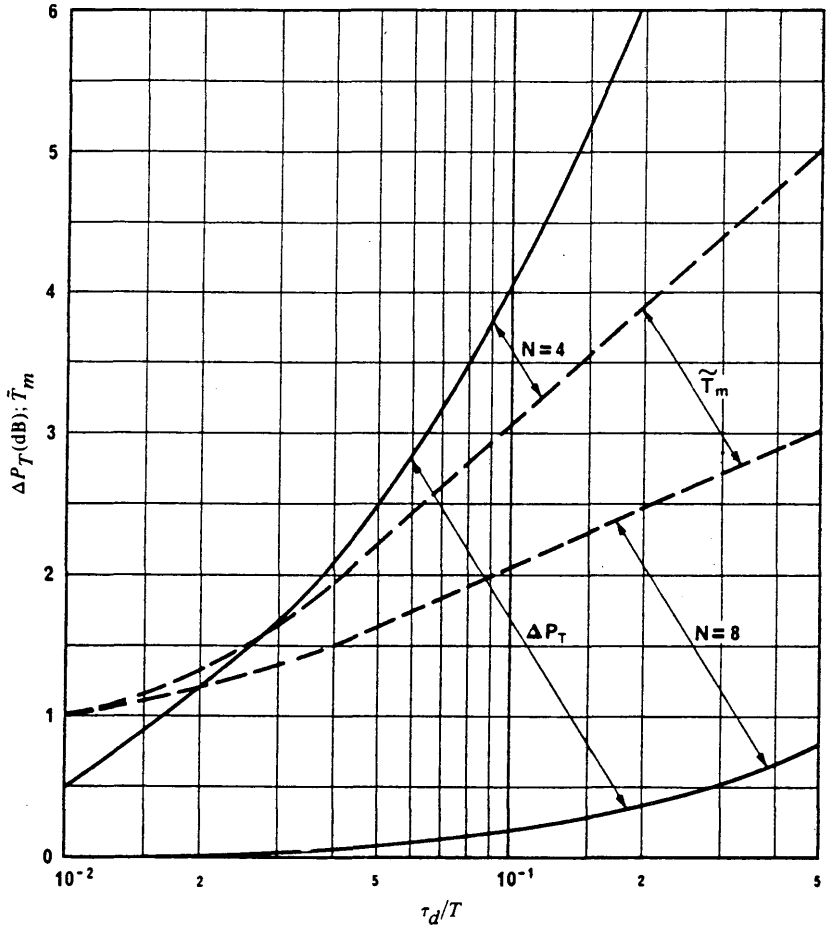


FIGURE 3 – *Mean thermal noise power/minute and* $\tilde{T}_m$ *as a function of blanking/pulse interference time ratio*

The equipment used to put this method of suppressing pulse interference into effect was tested out on a 120 telephone channel trans-horizon radio-relay link. A pulse generator with τ = 1 ms connected to a directional antenna was used as the interference source. The value of τ_d was taken as 1.5 ms. The pulse interference level at the receiver output exceeded the median wanted signal level by 15 dB. The diversity reception factor was 6.

The curves in Fig. 4, plotted from the measurement data, refer to the distribution of probabilities of values $\rho = 10 \log 10^9/P_T$ (where ρ is the signal-to-noise ratio in one upper telephone channel and P_T is the channel noise power).

Curve A relates to the absence of impulse interference; curve B to the case of interference suppression; curve C to the case of no interference suppression. A comparison of curves A and C shows that, without the use of interference suppression equipment, there is a substantial increase in telephone channel noise power (approximately by 25 dB for 20% of the time). Comparison of curves B and C shows that this method of noise suppression considerably reduces the noise power (by 15 dB for 20% of the time). It can be seen from a comparison of curves A and B that reception noise immunity with pulse interference suppression is close to that obtained in the absence of interference. The slight increase (by about 5 dB) in telephone channel noise power over the calculated results may, in practice, be attributable to the presence of switching interference.

Although this method is applicable only to systems using frequency diversity reception, it compares favourably with other known methods ("blanking", extrapolation, etc.), since it permits the effective suppression of pulse interference irrespective of its duration.

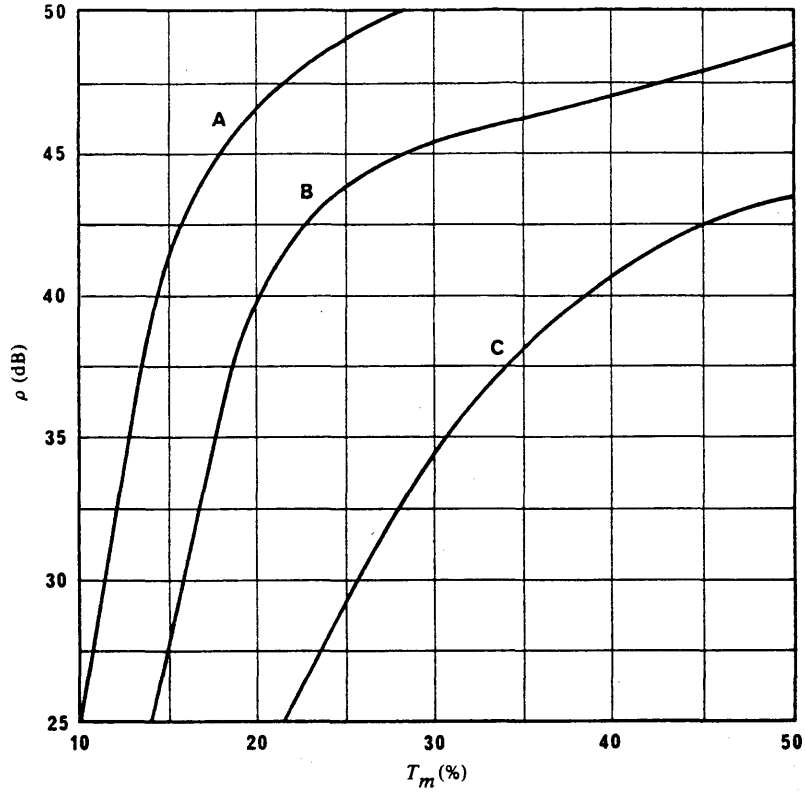


FIGURE 4 – *Distribution of probability value ρ (dB) as a function of T_m(%)*

7. Conclusions

Based on the analytical and empirical results discussed in the previous sections, the following conclusions are drawn for sharing between radar systems and certain kinds of FDM-FM systems. The previous discussion addresses only some of the operational modes of FDM-FM systems that have the noise performance objectives as specified in Recommendation 393. The results, therefore, of this Report do not apply to systems that incorporate data and/or high quality picture signal transmissions in the vacant lower and/or upper parts of the FDM-FM baseband.

For the case in which the communications signal power is greater than the peak radar signal power, methods other than pulse blanking or suppression should be used to improve the communications circuit quality.

Investigations conducted to date do not support or refute the appropriateness of pulse blanking as a means of improving communications circuit quality for the case in which the communications signal power is approximately equal to the peak radar signal power.

For the case in which the communications signal power is less than the peak radar signal power, intentional blanking may be an appropriate technique to use to improve communications circuit quality for situations in which the communications and the interfering pulse carrier frequencies are not equal. For such cases, the feasibility of sharing will depend upon the communications system receiver characteristics, the pulse source characteristics and the level of acceptable interference. Further studies, both analytical and empirical, into the interrelation of these characteristics and the resultant communications system transmission quality are required. These studies should address not only sharing by the use of pulse blanking or suppression of the communications receiver but also other methods for lowering the interference level from radar systems that is experienced by the FDM-FM systems.

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REPORT 826-1 *

EXAMPLES OF BAND SHARING BY EMPLOYING SPREAD-SPECTRUM TECHNIQUES

(Study Programme 18B/1)

(1982-1986)

1. Introduction

This Report is written in response to Study Programme 18B/1 and deals with the questions of spread-spectrum (SS) techniques and spectrum sharing. A characteristic of an SS system is that the emitted signal bandwidth is typically much greater than the bandwidth of the message being transmitted. The large bandwidths and concomitant low power spectral densities used by these systems potentially make them less likely to interfere with conventional systems operating in the same environment.

The definition of an SS system is given in Report 651. Descriptions of SS signalling techniques along with transmitter and receiver implementations are contained in the literature [Dixon, 1976] (Report 651). Three examples are given to show the potential of sharing between SS and other modulation techniques.

2. Factors to consider in band sharing

The ability of two or more systems to share a band with an acceptable level of electromagnetic compatibility involves a number of factors specific to the systems being considered. In general, however, successful band sharing may necessitate a trade-off between three conditions at the receivers of the systems.

Condition 1 – The wanted signal power delivered to the receivers must, with high probability, exceed an acceptable threshold value to assure high detection probability of the shortest time duration signal element the receiver is capable of recognizing.

The factors influencing satisfaction of this condition are the conventional considerations that apply in any link calculation. Transmitter power should be the minimum needed as determined both by receiver sensitivity and expected variation in propagation path loss. Antenna characteristics should be consistent with coverage requirements. Receiver characteristics are the result of compromise at the designing stage to achieve sensitivity and dynamic-range balance and to account for transmitted signal tolerances and relative station motions.

Condition 2 – In the presence of interference, the signal-to-interference (S/I) ratio must exceed an acceptable threshold with high probability. Typical factors influencing satisfaction of this condition are:

- interference power minimization by such techniques as transmitter power limitation, antenna nulling, low-duty-factor, and low spectral density;
- orthogonal signal structure designs that produce exploitably different characteristics where the S/I ratio can be enhanced by signal processing;
- receiver discrimination factors which take account of what existing receivers do rather than what is ideally possible.

Condition 3 – If Conditions 1 and 2 cannot be satisfied simultaneously, the application of other techniques may permit sharing. The signal design may include sufficient redundancy to permit recovery of received data when there is a detection probability failure for some fraction of unit signal elements (i.e., Condition 1 and/or Condition 2 are not always satisfied). Typical factors influencing satisfaction of this condition are:

- redundant or diversified signalling structure;
- redundant information stream with error detection or correction capability;
- designs that employ memory either to retain the most current information or to extrapolate the most current information until the next update.

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* This Report is the result of a merging of Reports 652 and 826: Report 652 is hereby cancelled.

3. Example 1 – Interference from SS systems to conventional voice systems

3.1 *General*

Example 1 investigates the interference from SS systems to conventional voice systems. Two general types of SS signalling techniques of interest in this example are the direct-sequence (DS) and frequency-hopping (FH) techniques. The hybrid form employing both of these methods (FH/DS) is also of interest.

For two specified levels of performance, this Report provides signal-to-interference (S/I) power ratios required for the protection of AM voice, FM voice and FDM/FM voice signals of systems operating in the presence of either SS or like-modulation interfering signals (i.e. AM, FM, and FDM/FM voice). The protection ratios were obtained for the case of a single voice receiver operating in the presence of one SS or like-modulation system. For several typical SS signals and a like-modulation signal, these protection ratios are used to calculate values of minimum propagation loss required to maintain the specified performance levels of the voice receivers so that comparisons can be made between the SS situations and the like-modulation case. It is shown that lower propagation loss values are typically required for unwanted SS signals than for unwanted co-channel like-modulation signals, and therefore a greater potential for sharing exists.

3.2 *Protection ratios*

3.2.1 *Systems, signals and performance levels*

Representative AM and FM voice systems were selected for analysis from Report 525, which contains a summary of S/I protection ratios for combinations of desired and undesired modulation. A 600-channel FDM/FM system, for which measured performance degradation data had been obtained, was also used in the analysis. The significant characteristics of systems used in the subsequent analysis are given in Table I.

TABLE I - *Characteristics of A3E, F3E and F8E systems analyzed*

Modulation type	Receiver IF 3 dB bandwidth, BW_{IF} (kHz)	Receiver antenna gain (dBi)	Receiver noise figure (dB)	Modulation index	Peak deviation (kHz)	E.i.r.p.(2) (dBm)
A3E	8	3	6	1.0(1)	-	50
F3E	16	3	6	1.67	± 5	50
F8E (FDM/FM)	22 000	30	9	1.23	± 3280(3)	60

(1) This number corresponds to a root-mean-square (rms) modulation index of 0.3 as given in Recommendation 331.

(2) E.i.r.p.: Equivalent isotropically radiated power.

(3) This number represents the total peak deviation.

Protection ratios were obtained for two performance levels of these systems when operating in the presence of DS and FH SS signals. The protection ratios were obtained in terms of the mean wanted signal-to-mean interference power ratio within the intermediate frequency (IF) filter referenced to the receiver input. The ratios determined for the FH case are also applicable to hybrid, FH/DS, SS signals. The two performance levels were taken to be articulation index (AI) values of 0.7 and 0.9, where AI is a measure of voice intelligibility, and the 0.7 threshold value separates the region within which there is no degradation to intelligibility from the region within which there is degradation to intelligibility, and the 0.9 value, which is a tractable level for analysis, is between marginally commercial and good commercial quality (see Reports 525 and 526).

3.2.2 *Direct-sequence signals*

For conventional communication systems, it was assumed that approximately the same performance would be obtained for a given amount of unwanted power within the IF filter resulting from a DS signal or white Gaussian noise. Available measured data and computer simulation results indicate that this assumption is valid. While measured data supporting this assumption are not available for A3E or F3E systems, measured data permitting comparison of DS binary phase-shift keying (PSK) interference with Gaussian noise are available for a binary FSK system and for a simulated 600-channel FDM/FM system different from the one described in Table I. For these measurements, 40 Mbit/s DS signal and white Gaussian noise were individually processed through a filter having a 40 MHz 1 dB bandwidth, and adjustments were made such that the filter output power was the same in each of the two cases. The signal at the output of the filter was then used as the interfering signal at the input to the receiver being tested.

Measured values of probability of error versus receiver input S/I ratios for the FSK receiver for DS/PSK and unwanted noise signals are shown in Fig. 1. This figure shows that the probability of error is the same for the DS signal and Gaussian noise. The measured results indicate that an increase of approximately 2 dB in the S/I ratio is required to obtain the same performance as that of an ideal noncoherent binary FSK system [Stein and Jones, 1967]. The difference between the measured and ideal results is typical of the implementation loss encountered in the physical realization of such a system.

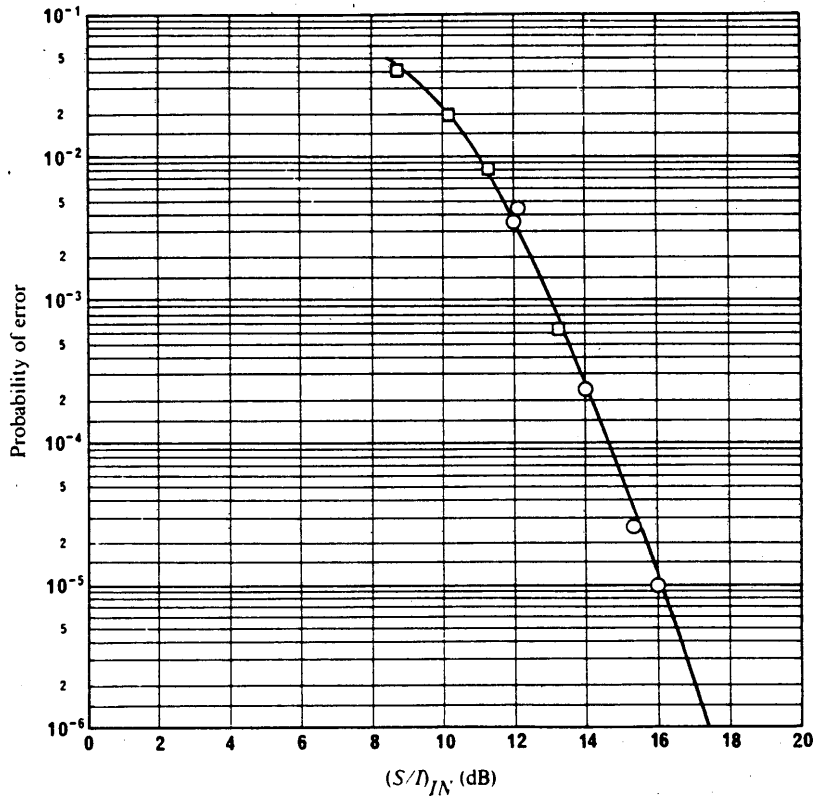


FIGURE 1 — *Measured probability of error versus S/I for noncoherent FSK with DS/PSK and noise*

□ : DS-PSK ○ : Noise BW_{IF} = 19 MHz

Illustrated in Fig. 2 are the results of measurements of "slot" interference power in an upper channel (2438 MHz) at the output of an FDM/FM receiver for a DS/PSK signal and noise. The difference between the two curves is 1.5 dB and is the result of the power spectral density of the filtered DS signal being slightly greater than that of the Gaussian noise in the channel being measured. This in turn is a result of the (sin $X/X)^2$ shape of the power spectrum of the DS signal as compared to the flat power spectrum of the Gaussian noise, and the fact that the power at the output of the 40 MHz filter is the same in each case. This indicates that the Gaussian noise assumption is valid.

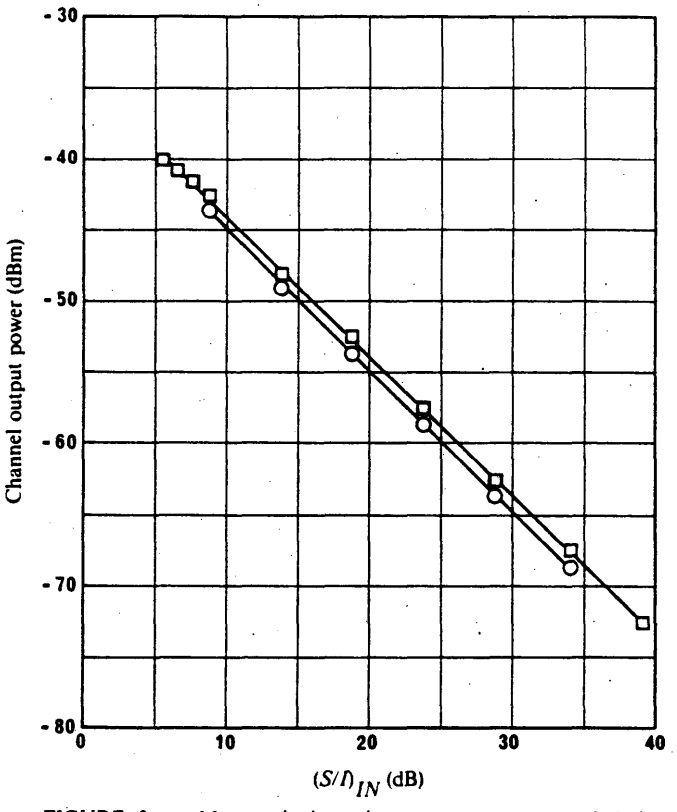


FIGURE 2 — *Measured channel output power versus S/I for FDM/FM with DS/PSK and noise*

□ : DS-PSK ○ : Noise

In addition to the measured data, computer simulation results [Farber and Cormack, 1977] permitting comparison of DS minimum-shift keying (MSK) (Report 177, Geneva, 1974) with white Gaussian noise were obtained for AM and FM voice receivers. The computer simulation model used was a modified version of the non-coherent receiver performance model described in Report 520. The modifications included adding a DS/MSK unwanted signal and changing the filtering operations so that they are performed in the time domain rather than the frequency domain. The AM and FM systems simulated were those described in Table I, and performance results were obtained individually for DS/MSK and white Gaussian noise.

For the AM system, simulated results of AI versus IF output S/I ratio for the DS/MSK signal and S/N ratio for the noise are shown in Fig. 3. The corresponding results for the FM system are depicted in Fig. 4. These figures show that approximately the same performance is obtained for the DS signal and noise of the same bandwidth. Therefore, for the DS case, the voice system performance degradation levels defined subjectively for noise interference (Recommendation 339, Report 525) may be used as well as those defined in terms of AI.

3.2.3 *Frequency-hopping signals*

For conventional communication systems, it was assumed that approximately the same performance would be obtained for a given amount of unwanted power within the IF filter resulting from an FH signal or a pulsed signal. The assumption is based on a limited number of subjective listening tests during which an FH signal in an AM receiver was reported as sounding like pulsed interference. An FH signal randomly occurs on channel, while a pulsed signal occurs regularly. However, at typical operating performance levels, a randomly occurring pulsed signal with a mean rate of occurrence equal to the pulse repetition frequency of a periodic pulsed signal of the same pulse width, will have very nearly the same effect on the intelligibility of a voice signal as will the periodic pulsed signal [Miller and Licklider, 1950].

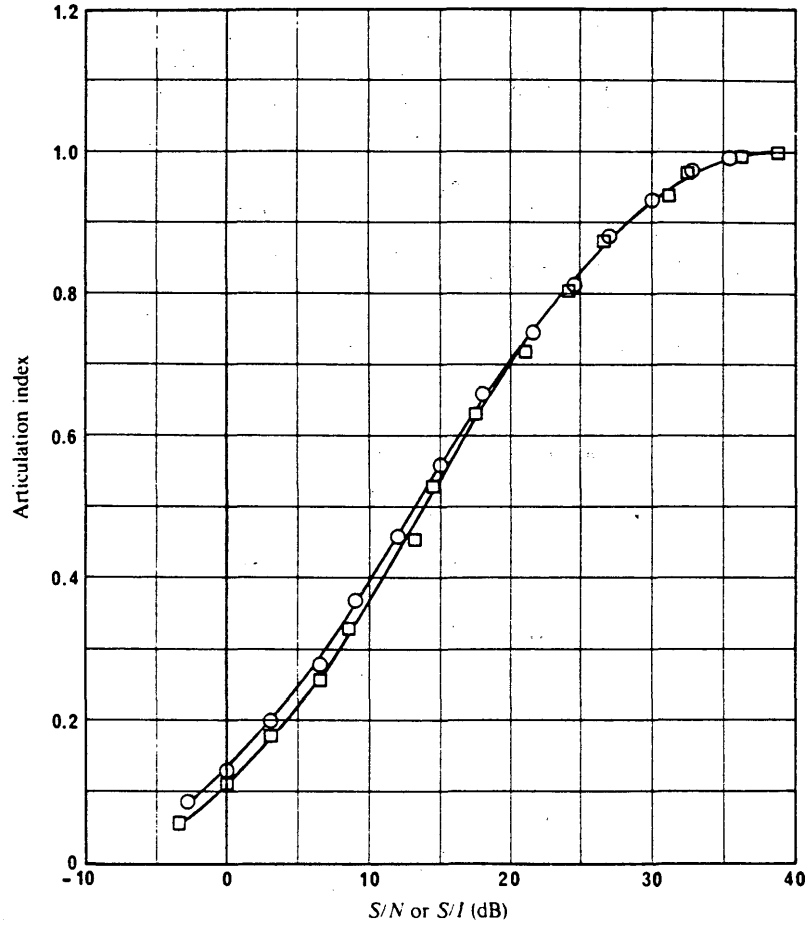


FIGURE 3 — *Simulation results for AM with DS/MSK and noise*

BW_{IF} = 8 kHz □ : DS-MSK ○ : Noise

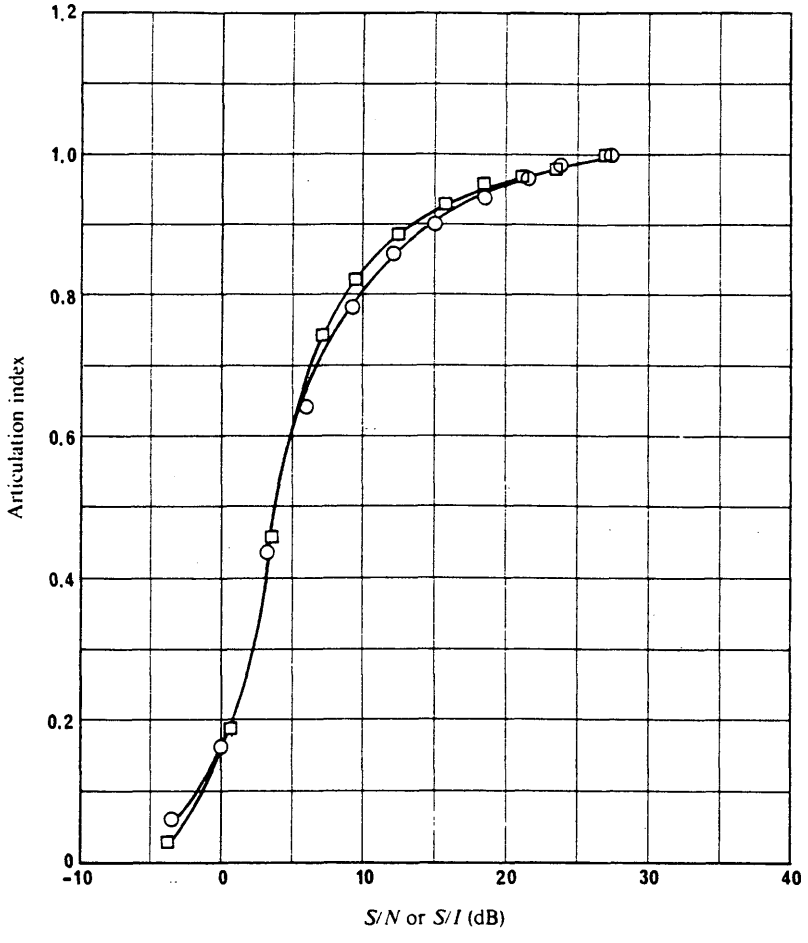


FIGURE 4 — *Simulation results for FM with DS/MSK and noise*

BW_{IF} = 16 kHz □ : DS-MSK ○ : Noise

Treating the FH signal as a pulsed signal, a significant amount of performance degradation data for pulsed interference to AM, FM and FDM/FM voice receivers was reviewed to obtain the S/I protection ratios. A summary of the trends indicated by measured data and computer simulation for the AM case is given in [Hatch *et al.*, 1971]. To obtain a given value of AI, different S/I ratios are required for different pulsed signal parameters. The S/I ratio selected as the protection ratio for each case was the greatest value required to insure an AI of at least 0.7 or 0.9 for all values of pulse width, pulse repetition frequency and off-tuning for which measurements had been performed. Therefore, these S/I protection ratios are conservative values. The pulse widths and pulse repetition frequencies for which the data varied are given in Table II.

TABLE II - *Parameter ranges for pulsed interference measurements*

Modulation type	Pulse width (μs)		Pulse repetition frequency (pps)	
	from	to	from	to
A3E	5	1000	10	400
F3E	1	1000	40	1000
F8E (FDM/FM)	0.1	1000	50	15 000

3.2.4 *Signal-to-interference ratios*

The S/I protection ratios are given in Table III, and they apply for the power in the bandwidth of the receiver. Partly due to the conservative manner in which the SS protection ratios were obtained, the protection ratio for the SS case is, in some cases, greater than that for the like-modulation situation. However, it is shown later that the large bandwidths used by the SS systems reduce the emitted power spectral density to such an extent that typically, for the specified performance levels, lower propagation losses are required for the SS systems versus the like-modulation systems. The S/I ratios for the A3E-to-A3E and DS (i.e., noise) -to-A3E cases were obtained from Report 525 for the 0.7 AI and by using the receiver performance model described in Report 520 for the 0.9 AI. For the F3E-to-F3E and DS-to-F3E cases, the values were obtained from ECAC [1973]. The protection ratios given in this Report for the F3E cases differ from those given in Report 525, because the F3E system considered in Report 525 utilizes pre-emphasis and de-emphasis, whereas the F3E system considered in this Report does not. For the F8E-to-F8E and DS-to-F8E cases, the protection ratios were obtained from measured performance degradation data. The protection ratios for the F8E receiver are for an upper (worst case) channel.

TABLE III - *Signal-to-interference protection ratios (dB)* (1)

S/N (dB)	Wanted signal modulation (2)	AI	Interference modulation				
			A3E (2)	F3E (2)	F8E (2)	DS (3)	FH (4)
45	A3E	0.7	7			21	19
		0.9	13			28	31
22	F3E	0.7		6		8	19
		0.9		20		15	30
55	F8E (FDM/FM)	0.7			17	15	15
		0.9			22	21	21

(1) All of the S/I ratios are mean power-to-mean power values.
(2) Described in Table I.
(3) Direct sequence.
(4) Frequency hopping.

3.3 *Minimum required propagation loss*

The protection ratios in Table III were used to compute the minimum required propagation loss for each system performance level for two types of SS signals and for a like-modulation unwanted signal. The minimum required propagation loss is defined as the minimum value which the unwanted signal must experience to insure an AI of at least the value specified. The two types of SS signals chosen for analysis were DS and FH/DS. For each type of SS signal, both binary antipodal PSK and MSK were evaluated for the carrier modulation. For each of these SS signals, the e.i.r.p. was taken to be equal to that of the like-modulation interference signal in the particular case being analyzed. These values of e.i.r.p. are given in Table I. Two typical chip rates (10 Mbit/s and 40 Mbit/s) were considered for the DS signal, where the term chip rate refers to the signalling speed of the spectrum spreading code sequence. The FH/DS signal was assumed to hop among 40 equally spaced carrier frequencies with a dwell time of 100 μs on each frequency. The FH/DS emission is considered to be such that each carrier frequency occurs at a mean rate of 250 times per second. Each of two values for the chip rate of the DS modulation was considered along with each of two corresponding values for the separation between adjacent frequencies. Chip rates of 2.5 and 5 Mbit/s were used along with frequency separations of 2.25 and 4.5 MHz, respectively, for the case of PSK modulation. For the case of MSK modulation, chip rates of 2.5 and 5 Mbit/s were used along with frequency separations of 1.5 and 3 MHz, respectively.

The minimum required propagation loss, L_p, in dB for an AI of 0.7 or 0.9 is given by:

$$L_p = L_{IF} - (S/N) - N + (S/I)_T \qquad (1)$$

where:

L_p: minimum required propagation loss,

I_{IF}: the transmitted interference power within the IF filter (dBm),

S/N: the mean wanted signal-to-mean Gaussian noise power ratio in the IF filter (dB),

$(S/I)_T$: signal-to-interference protection ratio, the appropriate protection ratio as given in Table III required for the 0.7 or 0.9 AI score.

For the detailed derivation of the propagation loss calculations, see Annex I.

The results of the propagation loss calculations are presented in Table IV. For the FH/DS signals in the table, the entries in the ordered sets of five numbers represent the values of parameters (n_f, τ_f, r_f, R_c, f_s)

where:

n_f: the number of frequencies being used,

τ_f: the dwell time in μs on each carrier frequency,

r_f: the mean rate of occurrence in occurrences/s of each carrier frequency,

R_c: the chip rate (Mbit/s),

f_s: the separation (MHz) between adjacent carrier frequencies.

The range of emission bandwidths covered by the selected typical SS signals is very wide and should be generally applicable. Small changes in the spread bandwidth would not lead to large changes in the minimum required propagation loss. The separation distances corresponding to the minimum required propagation losses given in Table IV may be obtained using appropriate propagation loss calculation procedures. The results in Table IV should not be used to compare A3E, F3E and F8E with each other due to the different wanted signal levels used in the three cases.

3.4 *Measurement data*

An experiment was conducted by the Canadian Administration in Ottawa, Canada, utilizing FH commercial equipment operating in the 30-88 MHz band. Measurements were made to identify co-channel interference protection criteria applicable to the land mobile service operating in the low VHF band. Specifications for the FH equipment used are given in Table V.

TABLE IV - *Minimum required propagation losses (dB)*

Emission bandwidth (1) (kHz)	Interference		Wanted signal					
			A3E		F3E		F8E	
		AI	0.7	0.9	0.7	0.9	0.7	0.9
1.4(2)	A3E		144	150				
1.5(2)	F3E				163	177		
400 (2)	F8E (FDM/FM)						143.6	148.6
9 000	DS/PSK 10 Mbit/s		127	134	137	144	141.6	147.6
6 000	DS/MSK 10 Mbit/s		129.1	136.1	139.1	146.1	141.6	147.6
36 000	DS/PSK 40 Mbit/s		121	128	131	138	139	145
24 000	DS/MSK 40 Mbit/s		123.1	130.1	133.1	140.1	141.2	147.2
180 000	FH/DS/PSK (40, 100, 250, 5, 4.5)		111.7	123.7	134.7	145.7	131.7	137.7
120 000	FH/DS/MSK (40, 100, 250, 5, 3)		113.7	125.7	136.7	147.7	133.7	139.7
90 000	FH/DS/PSK (40,100, 250, 2.5 2.25)		114.7	126.7	137.7	148.7	134.7	140.7
60 000	FH/DS/MSK (40, 100, 250, 2.5,-1.5)		116.7	128.7	139.7	150.7	136.7	142.7

(1) 3 dB emission bandwidth (the emission bandwidth to be used in determining the bandwidth over which a transmitter and receiver are co-channel).

(2) This value is with respect to the peak sideband power spectral density.

TABLE V – *Specifications of the FH equipment used*

Operating frequency (MHz)	30-88
Hop rate (per s)	100
No. of frequencies	256
FR bandwidth (MHz)	6.4
Adjacent frequency spacing (kHz)	25

3.4.1 *Signal-to-interference (S/I) protection ratios*

Three commercial FM receivers, having a channel spacing of 20, 50 and 25 kHz respectively, were used in the measurements. *S/I* protection ratios for co-channel operation were measured using the just perceptible (JP) and non-intelligible (NI) degradation levels which are about equivalent to the minimum interference threshold (MINIT) and 0.3 articulation index (AI) defined in Reports 526 and 525. The results are given in Table VI.

TABLE VI – *Co-channel signal-to-interference protection ratios (dB)*

Degradation level	SINAD (dB)	No. of interference sources	*S/I* (dB)		
			Receiver 1	Receiver 2	Receiver 3
JP	30	*n* (1)	– 12	+10	+ 10
JP	12	*n*	– 15	– 8	– 14
NI	30	1	–106	–92	– 96
NI	30	4 (2)	– 80	–78	– 92
NI	12	1	– 90	–94	–100

(1) For JP, the number of interference sources, *n*, was irrelevant since the level at which interference just becomes perceptible is independent of its rate of occurrence. Tests were done with $n = 1, 2$ and 4.

(2) For receiver 3, only two sources of interference were used.

Table III presents theoretical predictions of signal-to-interference protection ratios for an interfering FH/DS signal and a wanted FM (F3E) signal. The FH system used in the Canadian experiment had a lower hop rate than those used in Table III. Measured S/I ratios were also lower than those of Table III.

3.4.2 *Minimum propagation loss*

The minimum required propagation loss would be a better tool for comparing experimental results with predicted data. In the former case, the minimum required propagation loss is taken as the difference between the maximum interference level tolerated at the receiver input and the equivalent isotropically radiated power. Data is presented in Table VII for receiver 3, assuming an e.i.r.p. of 20 dBW.

TABLE VII – *Minimum required propagation loss (dB)*

Degradation level	SINAD (dB)	W(1) (dBW)	S/I (dB)	Propagation loss (dB)
JP	30	−129	+ 10	159
JP	12	−146	− 14	152
NI	30	−129	− 96	53
NI	30	−129	− 92	57 (2)
NI	12	−146	−100	66

(1) Wanted minimum signal level, W, in dBW for the given SINAD ratio.

(2) Two simultaneous interference sources present.

The comparison between the above results, using the JP and NI degradation levels, and that of Table IV indicates that propagation losses are slightly higher for JP than for an AI of 0.9 and lower for NI than for an AI of 0.7.

3.4.3 *Summary*

The experimental evidence indicates that large propagation losses are required to protect an FM receiver against the just perceptible level of co-channel interference, but on the other hand, surprisingly small propagation losses are required to protect against the non-intelligible harmful interference. Moreover, considering that actual land mobile users of the 30-50 MHz band use analogue transmission techniques, an interference of 10 ms duration such as the one used in the Canadian experiment, will not be annoying if it does not occur too often. For a given land mobile channel, interference from one source was present on average 0.4% of the time so that a land mobile system (analogue voice) can tolerate a greater number of FH sources, the maximum number being determined by the desired grade of service.

This experiment did not attempt to evaluate the impact of FH interference on digital systems nor did it study the relationship between the number of FH interference sources and the degradation of a land mobile signal.

3.5 *Conclusion*

The results of this example show that lower propagation loss values are typically required for co-channel SS systems versus like-modulation systems for both a quality which represents the threshold of intelligibility degradation and a quality which is between marginally commercial and good commercial quality. As indicated by the results in Table IV, the reduction in minimum required propagation loss ranged from 13.9 to 23.3 dB for the A3E system, 23.3 to 39 dB for the F3E system and 1 to 11.9 dB for the F8E system. This indicates that a potential for sharing exists. Further examination is required to determine detailed sharing criteria for the cases examined where a single additional radio channel is overlaid upon a conventional radio channel. Additional cases require examination to determine the maximum multi-system capacity that could be achieved in a given volume. Test results using a 100 hops/s FH system indicate that S/I ratios and propagation losses required to protect conventional voice systems may be lower than predicted by theoretical models.

4. Example 2 – Interference from SS systems to the television broadcast service

4.1 *General*

The example investigates the interference from SS systems to television reception. The general type of SS technique of interest in this example is the FH technique.

For a specified level of performance, this Report provides S/I power ratios required for the protection of television video and audio signals in the presence of FH interfering signals. The results are compared to situations in which degradation of the video and audio signals due to a fixed frequency interference source is experienced. It is shown that for co-channel and adjacent channel interference, lower protection ratios are required against unwanted FH signals than against fixed frequency interference sources. A good theoretical discussion of the required protection ratio for television reception is given in Report 306 (Geneva, 1982).

4.2 *Frequency hopping signals*

Interference tests were carried out in Ottawa, Canada, on five standard North American television receivers using NTSC modulation. Specifications for the FH equipment used are given in Table V. Tests were conducted in the 50-88 MHz band.

4.3 *Signal-to-interference (S/I) protection ratios*

S/I protection ratios for co-channel and adjacent channel interference were measured using the JP and NI degradation levels, which are about equivalent to impairment 4.5 and less than 1 respectively on the CCIR impairment grade scale of Recommendation 500. The wanted television signal level was adjusted in all tests to 49 dB(μV) across a 300 Ω termination.

4.3.1 *Co-channel interference*

Co-channel interference test results are given in Table VIII. Average (for five receivers) and worst case required S/I ratios are given for television channel 6 (82-88 MHz). Other television channels required similar S/I ratios. In every case, interference was much more severe on the video signal than on the audio signal.

TABLE VIII – *Television co-channel signal-to-interference ratios (dB)*

Degradation level	No. of interfering sources	Video		Audio	
		Average	Worst	Average	Worst
JP	1	44	44	18	20
JP	4	46	48	22	22
NI	1	11	14	6	8

These results, when compared to Report 306 (Geneva, 1982) Annex II data, show that television reception is less susceptible to co-channel interference from FH systems than from narrowband systems operating on a frequency close to, or at, the picture carrier or the colour sub-carrier.

4.3.2 *Adjacent channel interference*

Table IX summarizes the results of adjacent channel interference tests. Average and worst case S/I ratios are given for three modes of interference:

Mode 1 – lower adjacent channel;

Mode 2 – lower second adjacent chanel;

Mode 3 – upper adjacent channel.

Degradation of the audio signal was, in every case, much less severe than degradation of the video signal and is omitted in the Table.

TABLE IX – *Television adjacent channel signal-to-interference ratios (dB)*

Degradation level	No. of interfering sources	Mode 1		Mode 2		Mode 3	
		Average	Worst	Average	Worst	Average	Worst
JP	1	−3	+8	−18	−5	−30	−27
JP	4	+1	+11	−13	0	−23	−18

It is reported in Report 306 (Geneva, 1982) and Recommendation 418 (Geneva, 1982) that protection ratios considered to be acceptable when one interference source is present less than 10% of the time are −6 dB for lower adjacent channel interference and −12 dB for upper adjacent channel interference. Protection ratios for just perceptible degradation are said to be 10-20 dB higher. Assuming that an FH system does not create interference to a television receiver more than 10% of the time, then Table IX test results are an improvement over figures given in Report 306 (Geneva, 1982) and Recommendation 418 (Geneva, 1982).

4.4 *Conclusion*

The results of this example show that, for a given grade of service at the television receiver input, lower S/I ratios were required for co-channel and adjacent channel FH emissions than from fixed frequency emissions. This example indicates that a potential for sharing exists, especially in adjacent channel operations. Further examination is required to determine the relationship between the required S/I ratio and the number of FH transmissions applicable to various situations, as well as to explore the capabilities of FH systems to avoid transmissions at frequencies near or at the picture and colour television carriers.

5. Example 3 – Band sharing of a conventional pulse navigation system (AN/DME) and a spread-spectrum TDMA system

5.1 *General*

An example of band sharing that exploits a number of factors in addition to low spectral density is band sharing between the air navigation/distance measuring equipment (AN/DME) system and a typical bandwidth expansion system specifically designed to capitalize on the features of AN/DME so as to achieve compatibility with it (and other systems) in the band 960-1215 MHz. The system is a time division multiple access system where time is subdivided into intervals designated as time slots and these time slots are allocated to users for transmission of data in accordance with their required data rate. In each active time slot a sequence of frequency-hopped bandwidth expansion pulses is transmitted which spreads the transmitted energy over the entire AN/DME band of 960 to 1215 MHz compared to the AN/DME interrogator receiver bandwidth of 300 kHz. In addition, AN/DME system features are taken into account by the selection of:

- short 6.4 μs pulses as the basic signalling element (single pulses longer than 8.5 μs might be falsely decoded as pulse pairs in some AN/DME equipments);
- continuous phase shift modulation and side-lobe filtering to limit the instantaneous spectral density at frequencies removed from the pulse carrier;
- low-duty-factor signalling to limit the probability that a single pulse coincides with an AN/DME pulse;
- minimum frequency separation between consecutive pulses in order to prevent decoding by AN/DME receivers.

For a description of the AN/DME system, see Annex II.

5.2 *Compatibility test programme*

The electromagnetic compatibility of a bandwidth expansion system with AN/DME has been carefully evaluated [Mayher, 1978]. Extensive bench tests which collected static parametric data on the effect of the bandwidth expansion system on both airborne interrogators and ground beacons were conducted. Extensive flight tests were also conducted to assess performance in the operational environment that could not be accurately simulated and to validate performance prediction models based on the bench tests. AN/DME system elements tested included representative samples of different AN/DME equipment designs and the fully planned microwave landing system/distance measuring equipment (MLS/DME) system.

An important effect of the evaluation process was to optimize compatibility by minor design modifications to both systems. The bandwidth expansion system filtered the pulse spectrum to assure low spectral density at frequencies far from the pulse carrier frequency and modified the frequency selection algorithm to assure sufficient separation in frequency between adjacent pulses to preclude false decoding by an AN/DME receiver. A modification was made to the ground beacons to prevent false range indications when exposed to certain multipath conditions. The original modification desensitized a beacon receiver for 250 μs after receipt of any strong pulse. This design was modified to work in conjunction with pulse pair decoding to reduce sensitivity to single pulses from bandwidth expansion systems or any other pulsed source.

The measure of ground beacon performance used in the test programme was reply efficiency. Reply efficiency is the percentage of replies a ground beacon makes to some interrogation source and is the ratio of the number of replies to the number of interrogations expressed in percent.

Figure 5 is a plot of a typical AN/DME beacon performance curve that shows the reply efficiency as a function of the desired signal strength with no interference present. The −90 dBm desired signal level at a reply efficiency of 70% is defined as the beacon sensitivity. When the interference was inserted into the ground beacon at a signal level of −50 dBm, the effect on beacon performance for reply efficiencies below 70% was a desensitization in the beacon performance that varied, depending upon the particular beacon characteristics from 0 dB to 1.6 dB. Figure 5 depicts a maximum desensitization of 1.6 dB in presence of interference. Reply efficiencies below 70% are considered since this is the region in which the airborne interrogator received and tracked a given beacon.

A measure of airborne interrogator performance is the desired signal level required at a given reply efficiency for the interrogator to receive and track a given ground beacon. Figure 6 is a plot of an interrogator acquisition threshold as a function of the received signal strength and reply efficiency. The corresponding beacon performance curve is also shown in Fig. 2. The acquisition sensitivity of the interrogator is approximately −92.5 dBm. When the interference was inserted into the airborne interrogator at a level of −36 dBm with a duty factor of 40%, the effect on the airborne interrogator was a shift or decrease in the acquisition sensitivity of the interrogator. This decrease varied depending upon the particular interrogator characteristics from 0 dB to 1.7 dB. A maximum decrease in interrogator acquisition sensitivity of 1.7 dB is depicted in Fig. 6.

5.3 *Conclusion*

Spectral sharing between a bandwidth expansion and a conventional system has been shown to be feasible in this example based on existing conventional characteristics. Minor modifications of the existing system and minor design modifications of the SS system permitted band sharing and led to a higher spectrum utilization efficiency. Other factors, e.g. time-slot duty factor and the acceptable degree of decrease of interrogator acquisition sensitivity, in addition to spectral density, can be extremely important in achieving band sharing and merit careful consideration in the design process. For this example, the interfering signal power is considerably above the sensitivity of the conventional receiver; however, acceptable performance is obtained by capitalizing on careful signal design as well as low spectral density.

6. Summary

Communication systems that employ a variety of spread-spectrum (SS) modulation techniques are being increasingly developed. This Report uses several examples to explain the interference from SS systems to conventional voice communication systems, television broadcast systems and a pulse navigation system respectively.

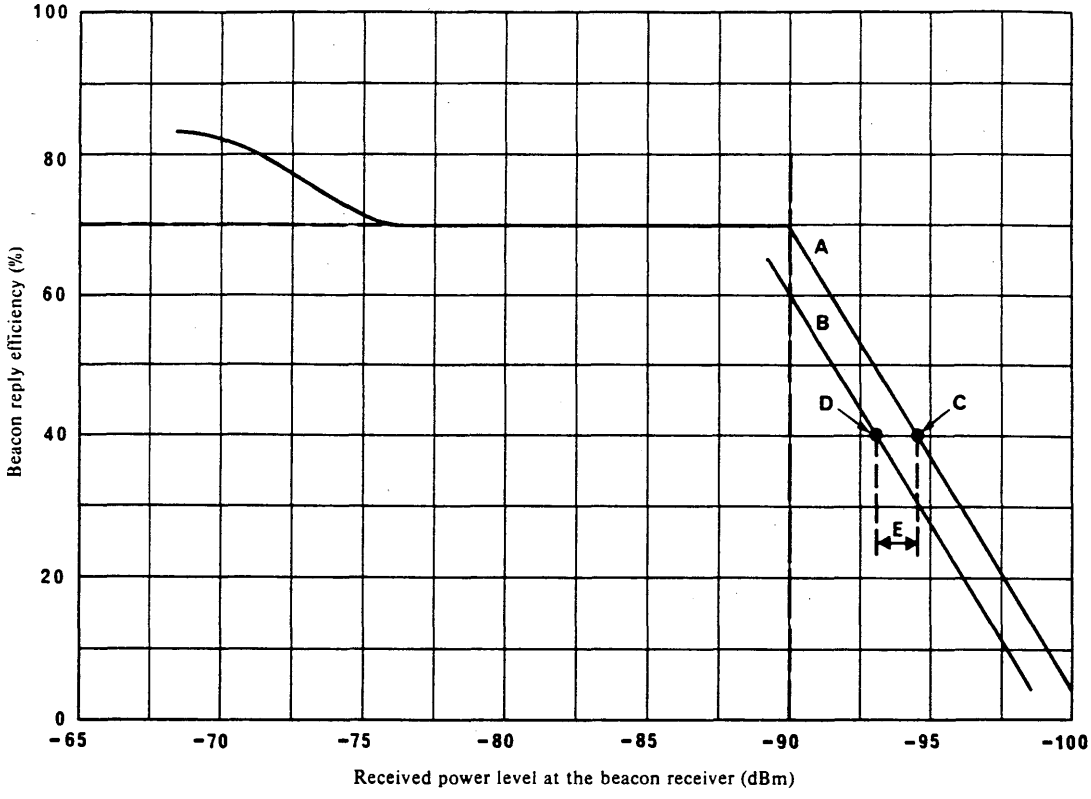


FIGURE 5 - *Typical interrogator beacon-coupled acquisition sensitivity for AN/DME beacon service*

A: beacon performance curve without interference
B: beacon performance curve with interference
C: acquisition sensitivity without interference
D: acquisition sensitivity with interference at - 50 dBm
E: maximum 1.6 dB desensitization

In the first example, for system performance levels of 0.7 and 0.9 articulation index, signal-to-interference protection ratios were given for A3, F3 and F8 (FDM-FM) voice communication systems operating in the presence of like-modulation and SS signals. Direct-sequence, FH and hybrid FS/DS SS signals were considered. For a like-modulation signal and several SS signals of the same radiated power, the minimum required propagation losses for acceptable performance of the A3, F3 and F8 systems were tabulated. The results for these cases show that lower propagation loss values are typically required for co-channel SS systems versus like-modulation systems for both a quality which represents the threshold of intelligibility degradation and a quality which is between marginally commercial and good commercial quality. Test results for an FH system substantiated the above results. Just perceptible and non-intelligible degradation levels were used in the tests.

The second example deals with the interference from FH systems to television reception in the low VHF band. Test results indicate that lower protection ratios are required for co-channel and adjacent channel FH systems versus fixed frequency systems for a just perceptible degradation level.

The third example deals with the interference to a conventional AN/DME system with low-duty factor from a spread-spectrum TDMA system. In this example, acceptable performance is obtained under the given conditions although the interfering signal power is considerably above the sensitivity of the conventional receiver.

The above examples show that with careful consideration of design factors, a potential for band sharing exists between SS and other modulation techniques.

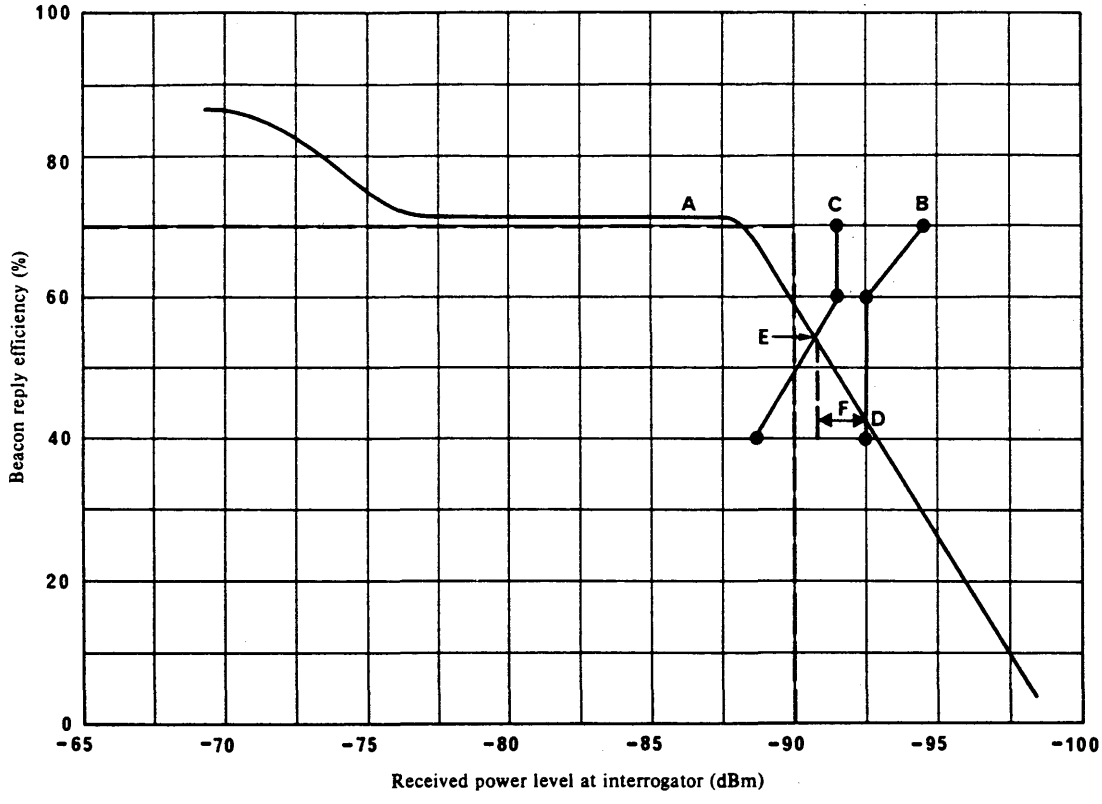


FIGURE 6 - *Typical interrogator acquisition sensitivity for AN/DME beacon service*

A: beacon performance curve
B: interrogator performance curve without interference
C: interrogator performance curve with interference at −36 dBm and time slot duty factor of 40%
D: acquisition sensitivity without interference
E: acquisition sensitivity with interference
F: maximum 1.7 dB desensitization

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ANNEX I

CALCULATION OF MINIMUM REQUIRED PROPAGATION LOSS

The minimum required propagation loss, L_p, in dB for an AI of 0.7 or 0.9 is given by:

$$\begin{aligned} L_p &= I_{IF} - (S/N) - N + (S/I)_T \\ &= I_{IF} - (S/N) + 174 - 10 \log BW_{IF} - NF + (S/I)_T \end{aligned} \quad (1)$$

where:

N: the mean noise power within the IF filter (dBm),

BW_{IF}: the IF filter 3 dB bandwidth (Hz).

(For the further glossary, see § 3.3.)

For the like-modulation interference, all of the transmitted power is contained within the IF filter so that I_{IF} may be expressed as:

$$I_{IF} = I_E + G_R \quad (2)$$

where:

I_E: the e.i.r.p. of the interfering system (dBm),

G_R: the antenna gain of the receiving system (dBi).

For the DS interference, I_{IF} is given by:

$$I_{IF} = I_E + G_R - FDR(0) \quad (3)$$

where:

$FDR(0)$ is the frequency dependent rejection in dB for $\Delta f = 0$ and $FDR(\Delta f)$ is defined by Report 654 as:

$$FDR(\Delta f) = 10 \log \left[\frac{\int_0^\infty p(f)\, df}{\int_0^\infty p(f)\, H(f+\Delta f)\, df} \right] \quad (4)$$

In equation (4):

$p(f)$: the emission spectral density (within a multiplicative constant) of the interfering signal,

$H(f)$: the receiver selectivity function due to all filters through and including the IF filter,

Δf: the difference between the carrier frequency of the interfering signal and the receiver tuned frequency.

For the FH/DS interference, the expression for I_{IF} becomes:

$$I_{IF} = I_E + G_R + 10 \log (\tau_f r_f) + 10 \log \left[\sum_{i=1}^{n_f} \left(\frac{1}{FDR(\Delta f_i)} \right) \right] \quad (5)$$

where:

τ_f: the dwell time in seconds on each carrier frequency,

r_f: the mean rate of occurrence in occurrences/second of each carrier frequency,

n_f: the number of frequencies being used and $FDR(\Delta f_i)$ may be obtained from

$$FDR(\Delta f_i) = 10^{FDR(\Delta f_i)/10} \quad (6)$$

where:

Δf_i: the difference between the i^{th} carrier frequency of the FH/DS interfering signal and the receiver tuned frequency.

The normalized power spectrum (i.e., unity maximum power spectral density) for binary PSK [Harris, 1973] is given by:

$$p(f) = \frac{\sin^2 (\pi f / R_c)}{(\pi f / R_c)^2} \tag{7}$$

and that for MSK [Mathwich *et al.*, 1974] is given by:

$$p(f) = \frac{1 + \cos (4\pi f / R_c)}{2 [1 - 16 (f / R_c)^2]^2} \tag{8}$$

where R_c is the chip rate.

The results of the propagation loss calculations are presented in Table IV.

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ANNEX II

THE AN/DME SYSTEM

AN/DME is a low-duty-factor, pulse navigation system that operates from 960 to 1215 MHz. Airborne interrogators transmit a random pattern of fixed-spaced Gaussian-shaped pulse pairs. Each pulse in the pair has a bandwidth of approximately 300 kHz and is limited to a peak power of 2 kW for aircraft equipped for high-altitude operation. Average interrogation rates range from 12 to 30 pulse pairs per second. A ground beacon receiving the interrogation transponds with a similar signal structure on a different up-link frequency and range is calculated at the interrogator from the round-trip propagation delay. Bearing data is provided by antenna-induced amplitude modulation of all pulse pairs radiated by the ground beacon, and azimuth reference bursts.

The AN/DME system has many of the attributes identified as desirable for band sharing. Strict standards exist defining service volume and the minimum signal strength provided by ground beacons within the service volume. Similarly the standards specify the signal strength required at the beacon receiver to obtain service from the beacon. Each of the elements of the system has been designed to meet these standards with adequate margin for their coverage areas of interest. Additionally, the system is designed to provide service to many aircraft simultaneously. All aircraft operate asynchronously with each other and therefore are sources of interference each to the other. Receivers are designed to tolerate this interference without disruption of service. Furthermore, airborne receivers may be within line-of-sight of more than one ground beacon on the same frequency (co-channel interference) or on an adjacent frequency (adjacent-channel interference). Receivers also tolerate this interference (within specified limits) without disruption of service.

AN/DME receivers are able to achieve their interference tolerance by utilizing the factors discussed earlier: low-duty-factor coded transmissions combined with appropriate receiver processing, redundant signalling and employment of memory to retain data between updates. Each interrogator receiver discriminates against interfering signals by integrating a number of time delayed replies to its own uniquely jittered pair interrogation pattern. Signals not having the proper pulse pair (pulse width, pulse shape, pulse spacing) are rejected. This includes almost all interference except that generated by other AN/DME equipments. Signals having the proper pulse pair characteristics but not having the interrogation jitter pattern are rejected. This includes responses to

interrogations by other aircraft. Signal not having the proper relative amplitude are rejected. This includes co-channel and adjacent-channel interference. The system is highly redundant and some interrogators can receive and track a beacon with the probability of receiving a reply to a signal interrogation as low as 30%. These interrogators operate with considerable margin under normal conditions where the reply probability is usually between 70 and 85%. Memory is often employed to provide continuous data to the operator between data updates for periods up to 10 s. After this time the interrogator reinitializes and automatically re-enters the signal acquisition mode.

As described above, AN/DME has many features which cause it to be a potential choice for band sharing.

REPORT 654-3

METHODS FOR CALCULATING INTERFERENCE POWER IN ADJACENT BANDS AND CHANNELS

(Question 44/1)

(1974-1978-1982-1986-1990)

1. Introduction

Question 44/1 asks how currently available models of system parameters of both wanted and interfering systems, e.g., emission spectra, receiver selectivity, antenna patterns, propagation attenuation, etc., may be logically combined in response to specific utilization problems. One way these parameters can be combined is to solve the problem of determining appropriate frequency and/or distance separation criteria for sharing the space-frequency domain. For example, interference problems can be created by the power from an emission in one band or channel overlapping into an adjacent band or channel and interfering with a receiver tuned close to the edge. In order to eliminate or restrict this interference problem, sharing criteria are required which specify minimum frequency and distance separation limits.

This Report describes basic measures which quantify the interactive effects among wanted signals, interference, and receiver characteristics for various frequencies and frequency distance separations. The measures are:

- frequency dependent rejection (FDR) which is a measure of the rejection produced by the receiver selectivity curve on an unwanted transmitter emission spectra;
- frequency distance (FD) which is a measure of the minimum distance separation that is required between a victim receiver and an interferer as a function of the difference between their tuned frequencies;
- relative radio-frequency protection ratio A (see Recommendation 560) which is the difference, expressed in decibels, between the protection ratio when the carriers of the wanted and unwanted transmitters have a frequency difference of Δf and the protection ratio when the carriers of these transmitters have the same frequency.

FD and FDR are measures of the interference coupling mechanism between interferer and receiver and are the basic solutions required for many interference evaluations. They aid in the solution of co-channel frequency sharing and adjacent band or channel interference problems by providing estimates of the minimum frequency and distance separation criteria between interferer and receiver which are required for acceptable receiver performance.

In amplitude-modulated sound broadcasting the effect of adjacent-channel interference is characterized by the relative radio-frequency protection ratio A (see Recommendations 559 and 560) which depends in an involved way on the technical parameters of the transmission system composed of transmitter, receiver and propagation path.

2. Definitions of frequency dependent rejection (FDR) and frequency distance (FD)

The interference level at the receiver is a function of the gains and losses the interference signal will incur between the source and the receiver and is expressed by:

$$I = P_t + G_t + G_r - L_b(d) - FDR(\Delta f) \qquad \text{dB} \tag{1}$$

where:

P_t: interferer transmitter power (dB),

G_t: gain of interferer antenna in direction of receiver (dBi),

G_r: gain of receiver antenna in direction of interferer (dBi),

$L_b(d)$: basic transmission loss for a separation distance d between interferer and receiver (dB) (see Recommendation 341),

and

$$FDR\ (\Delta f) = 10 \log \frac{\int_0^\infty P(f)\ df}{\int_0^\infty P(f)\ H(f + \Delta f)\ df} \tag{2}$$

where:

$P(f)$: emission spectra density generally normalized to unity maximum power spectral density (W/Hz),

$H(f)$: receiver selectivity,

$$\Delta f = f_t - f_r$$

where:

f_t: interferer tuned frequency,

f_r: receiver tuned frequency.

Equation (1) in Report 698 is similar to equation (2) except that equation (2) includes the normalization term $\int_0^\infty P(f)df$. The *FDR* expression, equation (2), contains only power terms. Modulation characteristics are included only through their effect on the emission spectral density $P(f)$. Interference degradation characteristics are assumed not to change as the receiver is detuned. The *FDR*, because it is considered a loss in equation (1), is defined in the manner of equation (2) to insure that expressed in dB it is ≥ 0.

In addition to the calculation performed in (1), it is often required to calculate the interference-to-noise (I/N) at the demodulation input or IF output. This measure is also useful in basic electromagnetic compatibility calculations because the I/N number is relatable to degradation, is easy to interpret, and is independent of the gain of the RF and IF amplifier stages. The noise at the receiver input referred to the IF bandwidth is given by:

$$N_0 = 10 \log kT_0 + 10 \log BW_{IF}\,(\text{Hz}) + \text{noise figure} \tag{3}$$

where:

N_0: receiver noise power in dBm,

k: Boltzmann's constant,

T_0: absolute temperature in K.

The I/N ratio is then given by:

$$I/N = I - 10 \log kT_0 - 10 \log BW_{IF}(\text{Hz}) - \text{noise figure} \qquad \text{(dB)} \tag{4}$$

The FDR can be divided into two terms, the on-tune rejection (OTR) and the off-frequency rejection (OFR), the additional rejection which results from off-tuning interferer and receiver.

$$FDR(\Delta f) = OTR + OFR(\Delta f) \qquad \text{(dB)} \tag{5}$$

where:

$$OTR = 10 \log \frac{\int_0^\infty P(f)\ \mathrm{d}f}{\int_0^\infty P(f)\ H(f)\ \mathrm{d}f} \tag{6}$$

$$OFR\ (\Delta f) = 10 \log \frac{\int_0^\infty P(f)\ H(f)\ \mathrm{d}f}{\int_0^\infty P(f)\ H(f + \Delta f)\ \mathrm{d}f} \tag{7}$$

The OTR, also called the correction factor, under conditions given by [White, 1972], can often be approximated by:

$$OTR \approx K \log\left(\frac{B_T}{B_R}\right), \qquad B_R \leqslant B_T \tag{8}$$

where:

B_R : receiver 3 dB bandwidth (Hz),

B_T : transmitter 3 dB bandwidth (Hz),

K = 10 for non-coherent signals,

= 20 for pulse signals.

The calculation of OTR can aid in co-channel sharing problems since the OTR is the rejection obtained when the receiver and unwanted transmitter are tuned to the same frequency.

When a maximum value, I_m, of acceptable receiver interference power from an interferer is specified, equation (1) is rearranged to give the required loss

$$L_b(d) + FDR(\Delta f) = P_t + G_t + G_R - I_m \tag{9}$$

Receiver performance is acceptable only if

$$L_b(d) + FDR(\Delta f) \geqslant P_t + G_t + G_R - I_m \qquad (10)$$

The required loss combines a distance dependent term $L_b(d)$ and a frequency dependent term $FDR(\Delta f)$. A curve of the various combinations of the two separations in frequency, Δf, and distance, d, which achieves a specified required loss, ξ, is an *FD* plot. Figure 1 is an illustrative example. The ordinate in Fig. 1 is the distance separation and the abscissa is the separation in tuned frequency of the interferer and receiver. The *FD* curve separates the acceptable receiver performance region from the non-acceptable receiver performance region. An *FD* plot shows the trade-offs in either frequency or distance separation which allow interferer and receiver to coexist in a space frequency environment.

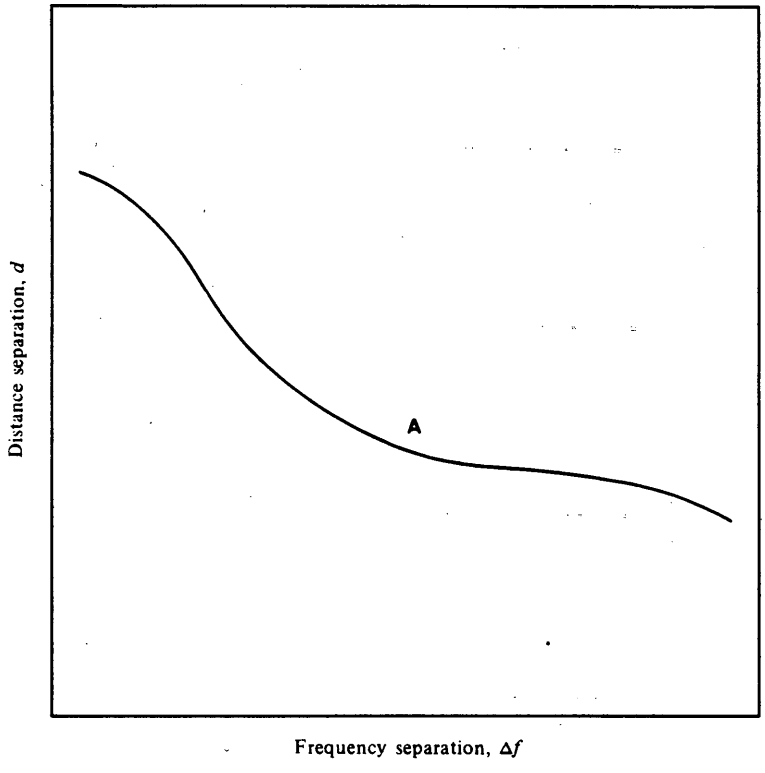


FIGURE 1 — *Separations in frequency and distance. The equation describing curve A is* $L_b(d) + FDR(\Delta f) = \xi$. *Above the curve is the region of acceptable receiver performance. Below the curve is the region of non-acceptable receiver performance.*

3. Computation of frequency dependent rejection (FDR) and frequency distance (FD)

The computation of *FDR* and *FD* is simple and has been implemented on a small computer [Cohen, 1977]. A block diagram of the computation is shown in Fig. 2. The emission spectrum and receiver selectivity are specified at a discrete number of frequencies. Extrapolation slopes define these functions beyond the discrete number of frequencies and interpolation is used to define functional values between discrete data points.

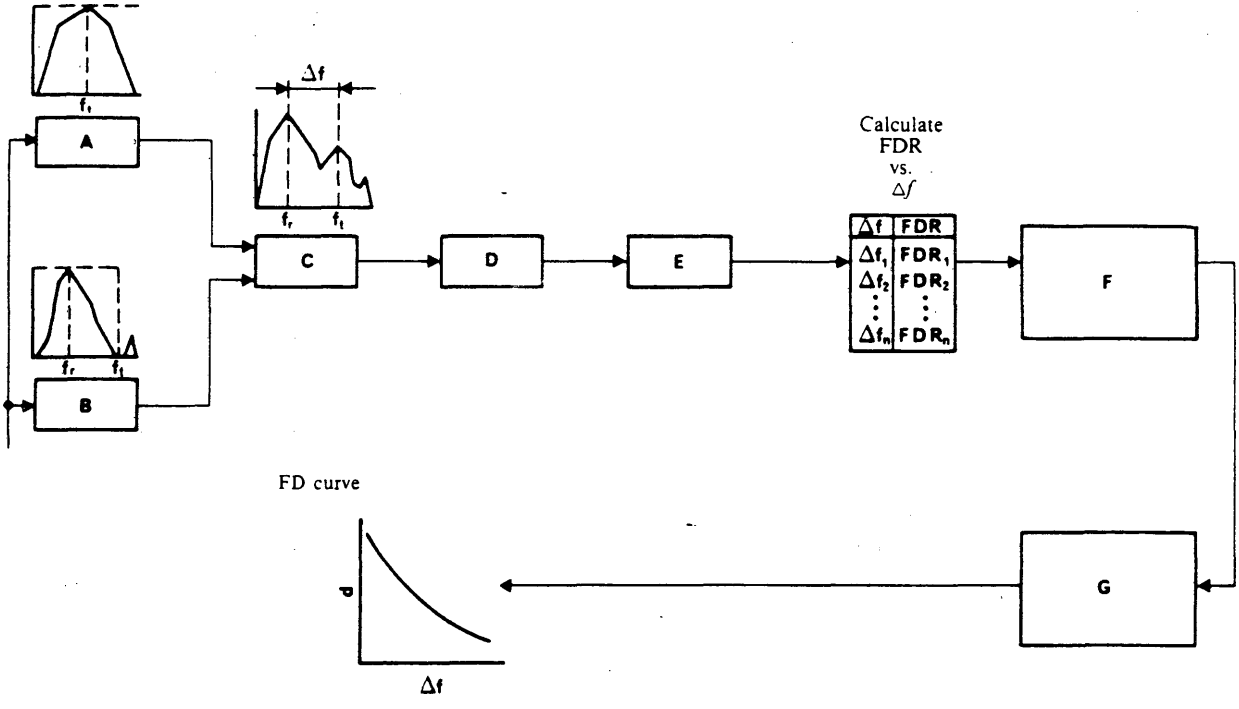


FIGURE 2 — *Block Diagram of Calculation process of FDR and FD*

A: Emission spectrum
B: Receiver selectivity
C: Multiply
D: Integrate
E: Normalize
F: Subtract required loss
G: Inverse propagation

Except for one case, in the computer model interpolation is log-linear which assumes a straight line on semi-log paper. Log-linear interpolation is not possible near the origin. In this case, linear interpolation is done. Two steps can be taken to minimize the region of linear interpolation. First, the data point representing the tuned frequency on the emission spectrum should be between two non-zero data points which are as close as possible to the zero point. Second, the selectivity curve should be very well defined since each portion of the curve may eventually be subject to linear interpolation.
The integration is evaluated numerically by the trapezoidal method using a variable integration step size.

In practice, the receiver pass bandwidth at specific selectivity levels and the width of the interference spectrum at specific attenuations are generally known. Hence the values of the number pairs f_i' and $H_R(f_i')$ $i = \overline{1,2N}$ and f''_ℓ $P_T(f''_\ell)$ may be considered as known, several numbers being selected from the series $10^{-0,1d}$, where d is an integer, as values of $H_R(f_i')$ $i = \overline{1,2N}$ and f''_ℓ $P_T(f''_\ell)$. Thus, for example, the receiver pass bandwidth and the width of interference spectrum are generally given at the levels d = 3; 30 and 60 dB.

Figure 3 shows a typical form of the relation between $H_R(f)$ or $P_T(f)$ respectively and the value of the frequency detuning relative to the receiver's central tuning frequency or the central frequency of the emitted signal (the variable f represents the frequency detuning relative to the central frequency). Figure 3 also gives the values of the receiver passband (F_{Ri}) or the transmitter emission band (F_{Ti}) at a given level equal to $10^{-0,1di}$.

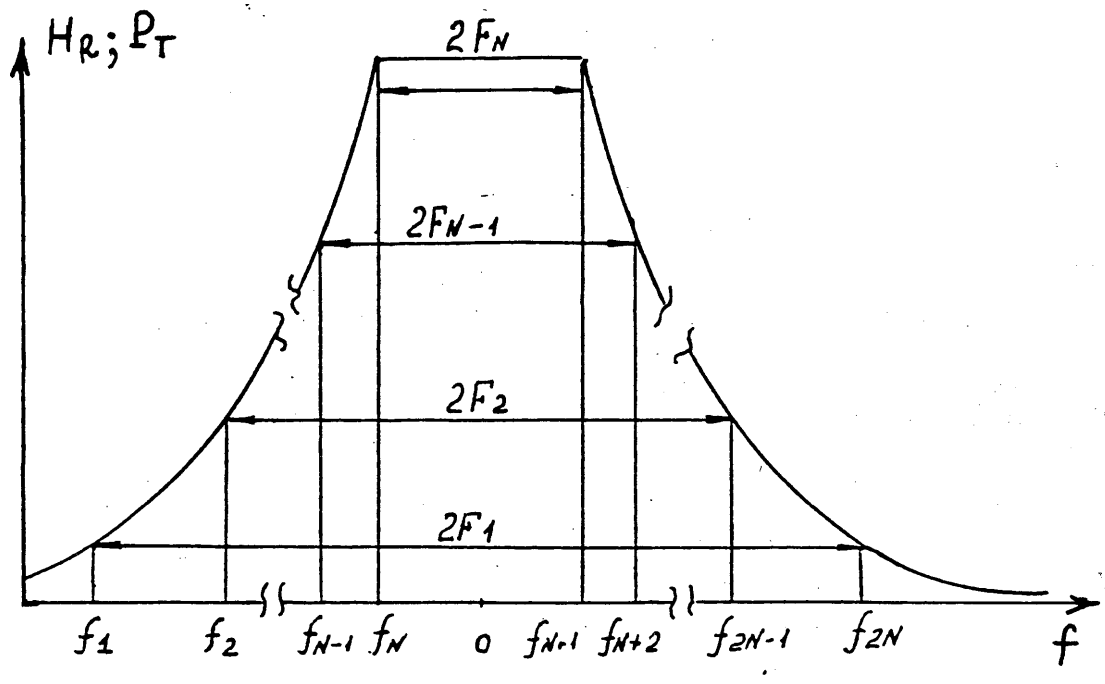


FIGURE 3

Dependence $H_R(f)$ or $P_T(f)$

The approximating equations for the function H_R (f) and P_T (f) should be selected in such a way that they match given values of these functions at given f_i' and f_ℓ'' and with sufficient accuracy approximate these functions at intermediate values of f.

Furthermore, the functional form of the approximating equations should be selected in such a way that the determination of FDR (Δf) does not require a great deal of calculation. The selection of the following equations for this purpose:

$$H_R(f) = H_{Ri}(f) = \begin{cases} 1 & \text{when } f_N' \le f \le f_{N+1}' \\ \left(\dfrac{F_{Ri}}{|f|}\right)^{n_{Ri}} & \text{when } f < f_N' \text{ or } f > f_{N+1}' \\ & f_i' \le f \le f_{i+1}' \end{cases} \tag{11}$$

$$P_T(f) = P_{T\ell}(f) = \begin{cases} \alpha & \text{when } f_K'' \le f \le f_{K+1}'' \\ \alpha\left(\dfrac{F_{T\ell}}{|f|}\right)^{n_{T\ell}} & \text{when } f < f_K'' \text{ or } f > f_{K+1}'' \\ & f_\ell'' \le f \le f_{\ell+1}'' \end{cases}$$

makes it possible to obtain recurrent formulae convenient for the calculation of FDR (Δf). The parameters of the approximations in (11) should be determined by the formulae given by [Bykhovsky, M.A. and Yashin, V.D., 1989]:

$$\alpha = (f_{K+1}'' - f_K'')^{-1}$$

$$n_{Ri} = \left[\frac{\lg\{H_R(f_i')/H_R(f_{i+1}')\}}{\lg|f_{i+1}'/f_i'|}\right]; \qquad F_{Ri} = |f_i'|\, H_R^{1/n_{Ri}}(f_i') \tag{12}$$

$$n_{T\ell} = \left[\frac{\lg\{P_T(f_\ell'')/P_T(f_{\ell+1}'')\}}{\lg|f_{\ell+1}''/f_\ell''|}\right]; \qquad F_{T\ell} = \frac{1}{\alpha}|f_\ell''|\, P_T^{1/n_{T\ell}}(f_\ell'')$$

In (12) [X] is an integer part of the number X.

To determine the FDR (Δf) coefficient, the following integral must be calculated:

$$N(\Delta f_p) = \int_{-\infty}^{\infty} H_R(f) P_T(f - \Delta f_p) df \qquad (13)$$

The calculation algorithm includes a number of stages.

In (11), the bounds of the domains in which the approximations of the functions H_R (f) and P_T (f) are correct are given by two ordered blocks of numbers $D'_{2N} = \{f_i'\}$ and $D''_{2K} = \{f''_\ell\}$. To calculate N (Δfp) the set D''_{2K} (Δfp) = $\{f''_\ell - \Delta f_p\}$ is formed and the ordered set D_2 (N + K) (fK) = $D'_{2N} \cup D''_{2K}$ (Δfp) is produced by combining the sets D'_{2N} and D''_{2K} (Δfp). The numbers f_K define the bounds of the intervals within which the approximations (11) are correct at specific values of i and ℓ, corresponding to a given value of k. In this process, N (Δfp) may be written as the relation:

$$N(\Delta f_p) = \sum_{k=0}^{M} J_k(i,\ell) = \sum_{k=0}^{M} \int_{f_k}^{f_{k+1}} H_{Ri}(f) P_{T\ell}(f - \Delta f_p) df \qquad (14)$$

where $f_0 = -\infty$; $f_{M+1} = \infty$; $M \leq 2\ (N + K)$.

Figure 4 represents a diagram of the calculation algorithm for N (Δfp), in which the determination block i and ℓ for a given k is indicated by the broken line. The symbols used in the diagram have the following meanings:

$D'_{2N} \cup D''_{2K}$ (Δfp) - combination of two sets;

$f_{i+1} \in D'_{2N}$ - verification of the condition that fi+1 belongs to the set D'_{2N}. Depending on the results of the verification of this condition, the value of i or is increased by 1.

The integrals J_k (i, ℓ) for different values of i and ℓ are calculated by means of the simple recurrent formulae given in [Bykhovsky, M.A. and Yashin, V.D., 1989], which also contains sample calculations of the coefficient N (Δfp).

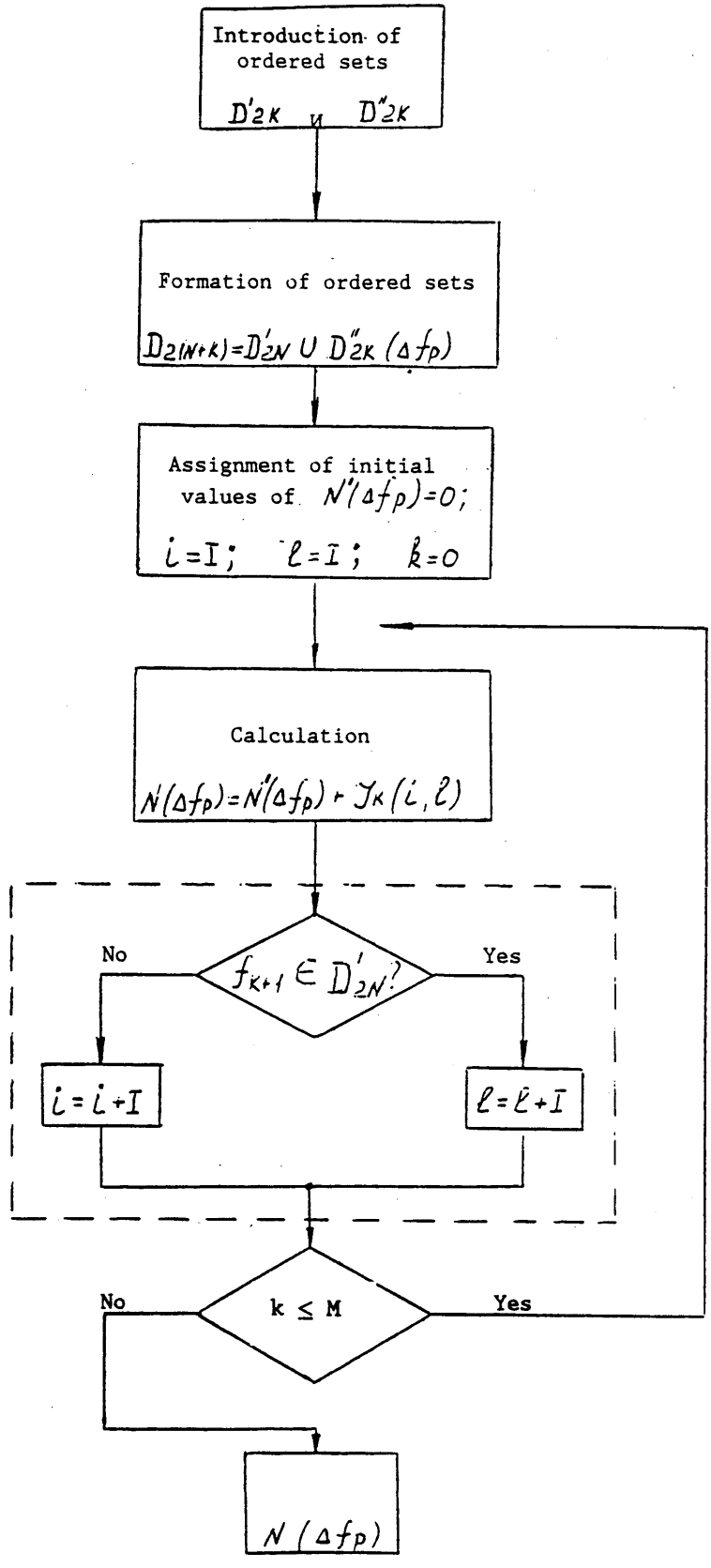


FIGURE 4

Calculation algorithm for $N(\Delta f_p)$

The propagation model might consider tropospheric, ionospheric or space paths. The propagation computation is inverse meaning that the basic transmission loss is input to the propagation model and distance is calculated. The propagation computation must be single-valued in distance for each loss value and the effect of lobes on the propagation loss cannot be included. The propagation curves included in Recommendation 370 are single-valued and have been included along with a smooth earth tropospheric propagation model [Bremmer, 1949] in the small computer computation [Cohen, 1977].

The *FD* can also be computed in a statistical manner. Jennings *et al.* [1976] have developed a computer technique which produces *FD* curves in which the parameters are the percentage of time during which interference is permissible.

For computation of *FD* in ground-wave propagation, the curves of Recommendation 368 can be used. In addition a procedure has been proposed [Råsajski and Petrović, 1981] for evaluation of the induced e.m.f. caused by a strong unwanted field, produced by a vertical dipole in a horizontal dipole receiving antenna, enabling a calculation of the *FD* expected for coast station receivers in the HF maritime mobile service with respect to unwanted hectometric (MF) broadcast transmitters.

4. A numerical method to calculate relative radio-frequency protection ratios

To investigate the influence of the parameters of an amplitude sound-broadcasting system on the relative radio-frequency protection ratio, graphical [Parreaux, 1972] and calculation methods [Gröschel, 1971; Petke, 1973] were developed which permit the determination of the relative radio-frequency protection ratio as a function of the parameters mentioned above which are in good conformity with the values measured according to the objective method of Recommendation 559. An additional description of this numerical method is contained in Recommendation 559.

5. Principle of the relative radio-frequency protection ratio calculation method

To determine the relative radio-frequency protection ratio, the physical process underlying the objective method, viz. the determination of the weighted power caused by the carrier and the modulation of the transmitter in the adjacent channel, is simulated by means of a mathematical model.

Two channels with the carrier frequencies f_T and f_R are assumed, whose frequency difference is equal to the channel spacing Δf under consideration. The transmitter power density spectrum related to the incremental bandwidth B_{eff} is simulated by the function F_T depending on the relative frequency $|f|$. This function is formed by multiplicative sub-functions (e.g. attenuation, spectral energy distribution, pre-emphasis of the higher audio-frequencies) and additive subfunctions (e.g. out-of-band radiation) [Gröschel, 1978] or it is built up by piece-wise approximation [Petke, 1973]. In a similar way, the overall response of the receiver, including weighting of the noise power with the aid of the psophometer filter, is represented by means of the function F_R depending on the relative frequency $|(\Delta f - f)|$.

In the case of double-sideband modulation, F_T and F_R are symmetrical to the respective carrier frequencies. To make allowance for the objective interference impression, involving the influence of the noise power in the adjacent-channel on the human ear, function F_R must include a weighting factor obtained by means of the psophometer curve.

The transmitter spectrum with the carrier frequency f_T produces an interference power ΔP_S in the adjacent-channel with the carrier frequency f_R, which, at given channel spacing Δf, can be calculated by integrating the product $F_T \times F_R$. As a result, one obtains:

$$\Delta P_S = \int_{f_1}^{f_1} F_T(|f|) \cdot F_R(|\Delta f - f|)\, df \qquad (15)$$

However, when performing the integration according to equation (11), for $\Delta f = 0$ (receiver exactly tuned to the transmitter frequency), one obtains the wanted receiver power ΔP_T. The relative radio-frequency protection ratio A is the ratio of interference power to wanted power in the adjacent channel

$$A = 10 \log \frac{\Delta P_S}{\Delta P_T} \qquad \text{dB} \qquad (16)$$

It is expedient to carry out the integration according to equation (11) with the aid of computers and numerical integration methods. In practice, most problems can be satisfactorily solved by numerical integration with an incremental bandwidth of $B_{eff} = 100$ Hz.

6. Accuracy and limits of the relative radio-frequency protection calculation method

In Figs. 5 and 6 a comparison is made between the calculated values and those measured according to the objective method. Figure 5 shows the results obtained for a double-sideband system and Fig. 6 those for a single-sideband system. In either case, the measured values are represented by dots. As long as the values of the relative radio-frequency protection ratio do not go beyond -40 dB, the deviations between calculation and measurement amount to about 1 dB. Beyond -40 dB, the measurements are increasingly affected by unavoidable noise inherent in the equipment used and by the intermodulation distortion in the receiver which can no longer be neglected, and is not taken into account in the calculation method.

Since the method described is based on the determination of power it does not provide the additional noise suppression by 3 dB [Tüxen, 1941] found during measurements of the co-channel interference performed on systems with linear demodulation and caused by the detector.

With the aid of this calculation method it is also possible to determine the interference components in the adjacent channel caused by the carrier (beat note) and by the sidebands, separately.

7. Summary

Computational methods for calculating interference power in adjacent bands and channels have been defined. Computer methods to perform these calculations have also been described.

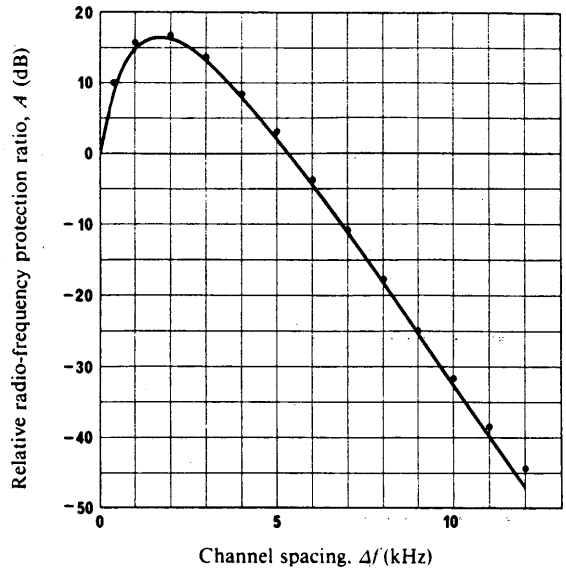


FIGURE 5 – *Comparison between the calculated relative radio-frequency protection ratio and the measured values (double-sideband system)*

——— : Calculated values
• : Measured values

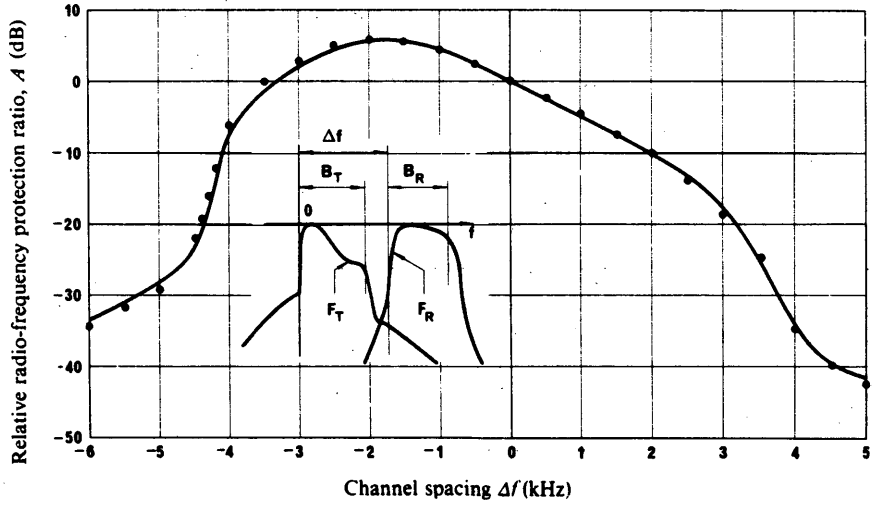


FIGURE 6 – *Comparison between the calculated relative radio-frequency protection ratio and the measured values (single-sideband system)*

——— : Calculated values
• : Measured values

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REPORT 1102*

CALCULATION OF THE RECEIVED VOLTAGE DUE TO THE RADIATION FROM MULTIPLE CO-FREQUENCY SOURCES*

(Questions 4/1 and 10/1)

(1990)

1. Introduction

There are many frequency sharing situations where interference may occur. In some situations the number and location of possible interfering sources is not known (e.g., when the interfering radiation is from domestic equipment).

There are several methods of assessment of multiple interference, some are described in Report 945-1.

This Report describes another method [Goedbloed et al, 1987]. It shows how to calculate the rms voltage at the terminals of an antenna due to the combined interfering effect from multiple co-frequency sources of a man-made nature. Different platforms for the antenna are considered and the factors particular to each are discussed. The analysis includes the probability functions for several parameters which might affect the level of interference at the receiving antenna.

2. The method

The field at the antenna due to radiation from N sources can be expressed as:

$$E = \sum_{i=1}^{N} K_i \, e^{j\phi_i} \qquad (1)$$

where N is the total number of sources in the area under consideration.

K_i and ϕ_i are the amplitude and phase of the field at the antenna due to a source i and are defined as follows:

$$K_i = E_L \, \frac{d_L}{d_i} \, B_i \, M_i \, K_s \qquad (2)$$

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* This Report should be brought to the attention of Study Groups 3, 4, 8, 9, 10 and 11.

where E_L is the specified field strength limit (μV/m) at a distance d_L (metres) from a single source and relates, for example to a CISPR limit.

B_i is a factor corresponding to the effect of the building surrounding a source i.

K_s is the source radiation pattern factor, accounting for the reduction of the mean level compared to the maximum level measured in the test

M_i is the manufacturer's safety margin resulting from the 80%/80% confidence test.

d_i is the source to antenna distance.

$$\phi_i = \phi(o) + \frac{2\pi d_i}{\lambda} + \beta_i \qquad (3)$$

where λ is the wavelength.
$\phi(o)$ is the initial phase of the signal as it leaves source i.
β_i is the phase angle due to building effect on source i.

The pattern of the receiving antenna should be taken into account so it is better to consider the voltage V at its terminals (across a 50 ohm load) rather than the filed strength E. Also, in the case of a satellite, the effect of Doppler shift on the received signal must be considered. Hence the voltage at the antenna terminals can be written as:

$$V = \sum_{i=1}^{N_D} C_A \; A(d_i) \; K_i \; e^{j\phi_i} \qquad (4)$$

where N_D is the reduced number of sources due to Doppler shift, if applicable. Otherwise N_D equals N.

C_A is the relationship between field strength at an antenna and the voltage across its terminals when matched to a 50 ohm load:

$$C_A = \frac{\lambda}{2\pi} \sqrt{\frac{50}{73}} \qquad (5)$$

$A(d_i)$ is the complex antenna factor which is a function of the source to antenna distance d_i:

$$A(d_i) = a(d_i) \; e^{j\alpha(d_i)} \qquad (6)$$

Equation (4) can therefore be written as:

$$V = \sum_{i=1}^{N_D} v_i \; e^{j\psi_i} \qquad (7)$$

where

$$v_i = C_A \; E_L \; d_L \; \frac{a(d_i)}{d_i} \; B_i \; M_i \; K_s$$

$$\psi_i = \phi(o) + \frac{2\pi d_i}{\lambda} + \beta_i + \alpha(d_i)$$

ψ_i is evenly distributed between 0 and 2π.

It is shown in annex section 1 that $|V|$ follows a Rayleigh probability law such that:

$$p(|V|) = \frac{|V|}{\xi^2} \; e^{-\frac{|V|^2}{2\xi^2}} \qquad (8)$$

where $\xi^2 = (N_D/2).\ E\ (v^2{}_i)$ and $E(v^2{}_i)$ is the expectation value of v_i^2.

The rms voltage is $v_{rms} = \sqrt{2}\,\xi$.

From the variables already stated v_{rms} can be calculated. Since $a(d_i)/d_i, B_i, M_i$ are independent probability distributions one can write:

$$v_{rms} = \sqrt{N_D}\; C_A\; E_L\; K_s\; d_L \sqrt{E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right] E\ (B_i{}^2)\ \ E\ (M_i{}^2)} \qquad (9)$$

The individual statistical terms will now be discussed in more detail.

2.1 Doppler effect

This only applies in the case of reception by satellite because of the high relative velocities involved. Briefly, the main effect is to reduce the number of sources which can be received by the satellite system. The effective number of sources is therefore given by:

$$N_D = K_D \;.\; N \qquad (10)$$

where K_D is a factor for the ratio of effective area to actual area seen by the satellite. It is unfortunate that K_D can only be evaluated numerically, nevertheless it is quite suitable for inclusion into a computer program. Table I shows typical values of K_D for a satellite at 850 km over areas of radius 15, 30 and 3120 km.

TABLE I

Radius (km)	K_D
15	0.58
30	0.30
3120	0.0067

2.2 Building effect

Building effect is due to numerous reflections, diffractions and absorptions caused by the building(s) surrounding a source. If each building affects the source signal by a factor B_i it can be shown that the distribution of B_i is log-normal. (i.e it follows a normal (Gaussian) law in terms of dB).

If m_B is the mean and σ_B is the standard deviation of -20 log B_i it is shown in annex § 2 that:

$$E\left[(B_i^2)\right] = 10^{-\frac{m_B}{10}} \; e^{2\sigma_B^2 (\ln 10/20)^2} \qquad (11)$$

A visual presentation of equation 11 is given in Fig. 1. It shows a cumulative probability for building attenuation to exceed a value on X-axis.

2.3 Manufacturer's safety margin

The many sources that are producing the interfering radiation will be built to comply with a radiation limit E_L dB(uV/m) at distance d_L meters and will have a spread due to manufacturing tolerances. This distribution of manufacturing tolerance is also log-normal.

If m_M is the mean and σ_M is the standard deviation it is shown in annex § 2 that: that:

$$E\left[M_i^2\right] = 10^{-\frac{m_M}{10}} \; e^{2\sigma_M^2 (\ln 10/20)^2} \qquad (12)$$

2.4 Antenna/distance factor

This is due to the combined effect of the receiving antenna pattern and the spread of the sources over a given area. In effect these two parameters are not independent.

Initially the simpler case of an aircraft flying over an urban area will be considered since the effects of earth curvature and Doppler shift can be neglected.

It is assumed that the aircraft antenna is a half-wave dipole with a cosine pattern and that N sources are evenly distributed over the area.

Consider the elemental area or annulus in Fig. 2 having width δ_r. The distance to the annulus is r and the radius of the whole area is R.

The probability density function p(r) is given by:

$$p(r) = 2\pi r/\pi R^2 = 2r/R^2$$

Since earth curvature is neglected the source to antenna distance is given by

$$d_i = \sqrt{r^2 + h^2}$$

Also, since we assume a cosine law for the antenna,

$$a(d_i) = h/d_i$$

Thus the antenna factor

$$a(d_i)/d_i = h/d_i^2 = h/(r^2 + h^2)$$

of which the expectation value is given by :

$$E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right] = \int_0^R \left(\frac{h}{r^2 + h^2}\right)^2 \frac{2rdr}{R^2} = \frac{1}{R^2 + h^2} \qquad (13)$$

This term is suitable for the calculation of radiation from relatively small areas such as towns where earth curvature effects can be neglected and the area can be effectively regarded as circular with radius R. It is not suitable when it is necessary to consider the total area which can be seen by an aircraft or satellite, i.e. when the horizon is the limit.

In this case, the curvature of the earth can no longer be neglected and it can be shown that:

$$E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right] = \frac{1}{2h(R_t+h)} + \left(\frac{2R_t + h}{4R_t(R_t+h)^2} \ln \frac{2R_t+h}{h}\right) \qquad (14)$$

where R_t is the radius of the Earth modified by a factor of 1.33 to allow for atmospheric refraction.

In the particular case of reception by satellite, Doppler shift on the received signal effectively reduces the area seen by the satellite to a narrow strip approximately 7 km wide. This also has a secondary effect since the shape of the source area is no longer as implied by Figure 2. The antenna-source factor is now given by an equation which, like K_D can only be evaluated numerically. However, the effect is relatively small and for a satellite at 850 km over the maximum available area the received voltage is increased by about 1.5 dB.

Returning to equation (9) the root-mean square voltage at the terminals of the distress antenna is therefore given by:

$$v_{rms} = C_A E_L K_s d_L \sqrt{K_D N} \sqrt{E(B_i^2)\, E(M_i^2)\, E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right]}$$

Hence

$$v_{rms} = C_A E_L K_S d_L \sqrt{K_D N}\ 10^{-\frac{m_B+m_M}{20}}\ e^{(\sigma_B^2+\sigma_M^2)\left(\frac{\ln 10}{20}\right)^2} \sqrt{E\left(\frac{a(d_i)}{d_i}\right)^2} \qquad (15)$$

where, for flat Earth calculations (aircraft over a town), $E[(a(d_i)/d_i)^2]$ is given by equation (13) and for curved earth calculations (aircraft or satellite over maximum available area) by equation (14).

For reception by satellite, Doppler effects must also be considered for which $E[(a(d_i)/d_i)^2]$ requires numerical evaluation.

3. Conclusion

Equation 15 gives a general formula for computing the interfering voltage at the receiver input from the summation of many interfering sources. Within this equation there are several factors of a probabalistic nature. Also some of the factors are only applicable to a particular type of antenna platform.

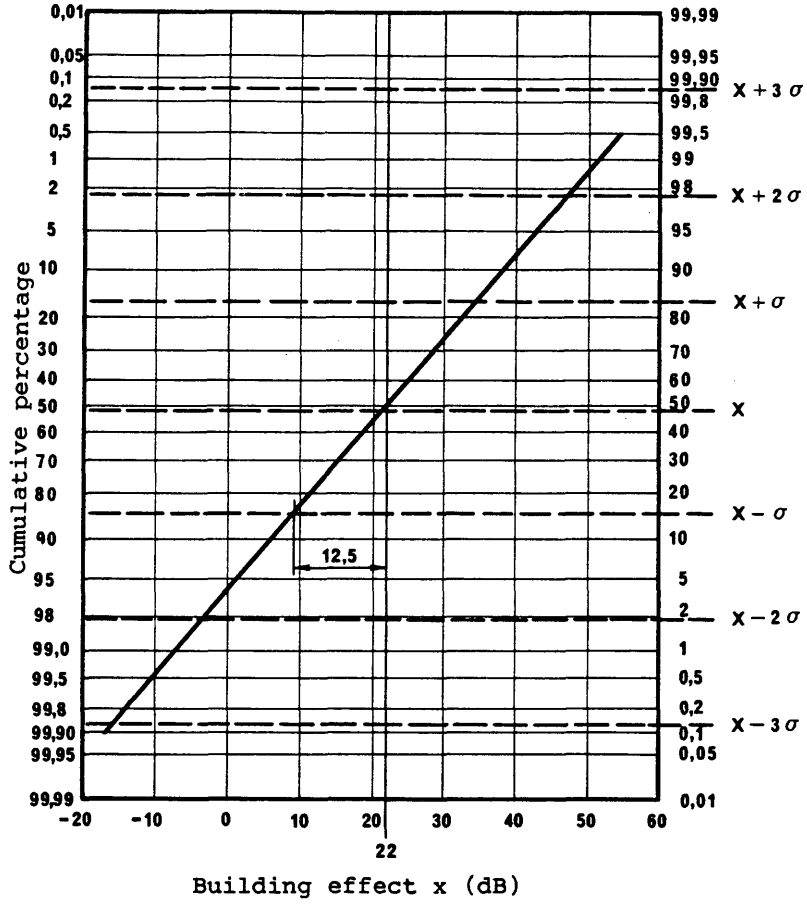


FIGURE 1 - Cumulative probability distribution of building attenuation

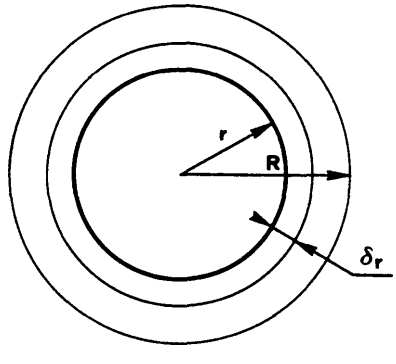


FIGURE 2 - The elemental area or annulus

REFERENCE

GOEDBLOED, J.J., PRINS, H.J. and RIEMENS K. [1987] Some RFI aspects of large-scale application of the CCIR 13.5 MHz sampling frequency standard, EMC Symposium, Zürich.

ANNEX I

1. The probability function of the voltage at the antenna terminals

From equation 7, the voltage at the antenna terminals is given by:

$$V = \sum_{i=1}^{N} v_i \, e^{j\psi_i}$$

This can be re-written as:

$$V = V_1 + j\, V_2$$

where $V_1 = \sum_{i=1}^{N} v_i \cos \psi_i$

$$V_2 = \sum_{i=1}^{N} v_i \sin \psi_i$$

V_1 and V_2 can be considered as the sums of a large number of independent random variables. The central limit theorem implies that they each converge to a Gaussian distribution. Since v_i and $\cos \psi_i$ are statistically independent, the first and the second moment of this distribution are:

$$E\left[V_1\right] = \sum_{i=1}^{N} E\left[v_i\right] \; E\left[\cos \psi_i\right]$$

and

$$E\left[V_1^2\right] = \sum_{i=1}^{N} E\left[v_i^2\right] \; E\left[\cos^2 \psi_i\right]$$

ψ_i is uniformly distributed between 0 and 2π,

hence $$E\left[\cos \psi_i\right] = \int_0^{2\pi} \cos \psi \, \frac{d\psi}{2\pi} = 0$$

Therefore $$E\left[V_1\right] = 0, \; E\left[V_2\right] = 0$$

Similarly

$$E\left[\cos^2 \psi_i\right] = \int_0^{2\pi} \cos^2 \psi \, \frac{d\psi}{2\pi} = \frac{1}{2}$$

hence $E\left[V_1^2\right] = \frac{N}{2} \sigma^2$ where $\sigma^2 = E\left[v_i^2\right]$

Similarly $E\left[V_2^2\right] = \frac{N}{2} \sigma^2$

V_1 and V_2 are independent because v_i and v_j are independent if $i \neq j$. Therefore V_1 and V_2 are two independent Gaussian variables with zero mean and the same standard deviation:

$$\xi = \sigma\sqrt{\frac{N}{2}}$$

The two-dimensional variable (V_1, V_2) follows a probability law, the density of which is given by:

$$\frac{1}{2\pi\xi^2} \; e^{-\frac{(V_1^2 + V_2^2)}{2\xi}}$$

Therefore $|V| = \sqrt{V_1^2 + V_2^2}$ has a probability density function given by:

$$p(|V|) = \frac{2\pi |V|}{2\pi\xi^2} \; e^{-\frac{|V|^2}{2\xi^2}}$$

or

$$p(|V|) = \frac{|V|}{\xi^2} \; e^{-\frac{|V|^2}{2\xi^2}}$$

This is a Rayleigh distribution

where $\xi^2 = \frac{N}{2}\sigma^2$

ξ is the most probable value;

rms value is given by $\xi\sqrt{2}$.

2. The probability function of a log-normal distribution

Let K be a log-normal random variable. If $k = -20 \log K$ then k will have a normal distribution given by:

$$p(K) = \frac{1}{\sigma\sqrt{2\pi}} \; e^{-\frac{(K-m)^2}{2\sigma^2}}$$

where m is the mean of the normal distribution and
σ is its standard deviation.

The probability function of K, the log-normal variable is related to p(k) by:

$$p(K) = p(k) \cdot dk/dK$$

Since $k = -20 \log K$

then $dk/dK = (-20/\ln 10) \cdot (1/K)$ and $K = 10^{-k/20}$

Hence $$p(K) = -\frac{20}{K \ln 10\, \sigma\sqrt{2\pi}}\, e^{-\frac{(K-m)^2}{2\sigma^2}}$$

by definition $$E[(K^2)] = \int_{-\infty}^{+\infty} K^2\ p(K)\ dK$$

Hence $$E[(K^2)] = \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^{+\infty} 10^{-\frac{K}{10}}\ e^{-\frac{(K-m)^2}{2\sigma^2}}\ dK$$

This evaluated as follows:

$$E[(K^2)] = \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^{+\infty} \exp T(K)\ dK$$

where $$-T(K) = K(\ln 10/10) + (K - m)^2/2\sigma^2$$

$$T(K) = (-1/2\sigma^2)\ K + \sigma^2\ (\ln 10/10 - m/\sigma^2)^2 + (\sigma^2/2)\ (\ln 10/10)^2 - m\ (\ln 10/10)$$

$$1/(\sigma\sqrt{2\pi}) \int_{-\infty}^{+\infty} \exp\ (-1/2\ \sigma^2)\ [K + \sigma^2\ (\ln 10/10 - m/\sigma^2)]^2\ dK = 1$$

hence the expectation value of K^2:

$$E[(K^2)] = \exp\ [(\sigma^2/2)\ (\ln 10/10)^2 - m\ (\ln 10/20)]$$

$$E[(K^2)] = 10^{-m/10} \exp\ [2\sigma^2\ (\ln 10/20)^2]$$

REPORT 842-2

SPECTRUM-CONSERVING TERRESTRIAL FREQUENCY ASSIGNMENTS FOR GIVEN FREQUENCY-DISTANCE SEPARATIONS

(Study Programme 18C/1)

(1982-1986-1990)

1. Introduction

Frequency-distance separation rules are a part of the frequency management process in many services. In a channelized service, these take the form: co-channel transmitters must be separated by at least d_0 km, adjacent-channel transmitters must be separated by at least d_1 km, transmitters separated by two channels must be at least d_2 km apart, etc. Despite the common use of such rules, it has not been possible to assign individual frequencies to arbitrarily located transmitters so as to get the maximum number of users in a specified bandwidth, or to determine how much additional spectrum is required to satisfy each additional frequency-distance separation rule for a fixed number of transmitters. The results of investigations by Hale [1980] show that it is possible to provide this information and also develop spectrum conserving frequency assignment algorithms.

The problem can be stated as follows: given a set of frequency-distance (FD) rules for a band, and a set of desired transmitter locations (arbitrary), assign frequencies to the transmitters so that the FD rules are satisfied while minimizing the total bandwidth required. A more general problem can also be stated: given a set of frequency constraints between transmitters, assign frequencies to the transmitters so that the frequency constraints are observed and the total bandwidth required is minimized. Frequency constraints are sets of forbidden channel separations restricting the frequencies that can be assigned to transmitters in order to prevent interference. These constraints can be frequency-distance rules or can be derived from signal-to-interference calculations, assigned priorities, or any other considerations. Various constraints, such as FD rules, are usually specified in terms of separation between transmitters and receivers. For certain radio services and conditions, these can be transformed into separation between transmitters. A variation of the problem is to minimize the number of channels assigned rather than the total bandwidth.

A computer algorithm that helps solve these problems efficiently can be used in several ways. First, it can be used as an allocation tool. For example, given a characterization of a proposed service, it can accurately determine the amount of spectrum required (the allocation) together with a designation of specific frequencies for use at specific locations (a table of allotments). Second, it can serve as a very fast automated tool for routine frequency assignments [Sakaki *et al.*, 1976; Pennotti and Boorstyn, 1976]. Third, it can be used as a research tool for the purposes of spectrum management. Examples of this last use are given later in this Report.

2. Progress towards an algorithm

In this Report, a computer algorithm that facilitates the assignment, to all transmitters, of frequencies which satisfy the frequency constraints is called a *sub-optimal* solution. If in addition, the algorithm minimizes the total bandwidth required, it is called a *solution* of the problem. An algorithm is *efficient*, or fast, if it always ends within a number of steps that is smaller than some power of the number of transmitter locations [Garey and Johnson, 1979]. An algorithm that is not efficient has limited value because for relatively large problems it may have intolerably long execution times even on the fastest computers existing. An ideal goal is the development of an exact, efficient solution for the frequency assignment problem.

This ideal goal is probably not realistic. Hale [1980] has shown that many frequency assignment problems are equivalent to generalized graph colouring problems, and therefore are "*NP*-hard" [Garey and Johnson, 1979], which means that the computer time required for an exact solution increases as exp (NP) where N is the number of transmitters and P is a constant. Therefore, it is very unlikely that anyone will ever devise efficient, exact solutions for these problems. A practical goal is to devise efficient, reasonably accurate, and widely applicable sub-optimal solutions for the frequency assignment problems.

Zoellner and Beall [1977] credit Metzger with recognizing that the classical graph colouring problem is analogous to frequency assignment problems where only co-channel constraints are involved and all transmitters have the same bandwidth. Other authors have commented on and applied this equivalence [Pennotti and Boorstyn, 1976; Roberts, 1979]. Struzak [1982] has also used graph colouring techniques to solve more complex frequency assignment problems.

Hale [1980] has generalized this approach and has shown that the generalized frequency assignment problem with any number of FD rules, or with more general constraints, is equivalent to the general graph colouring problem. He has developed a consistent mathematical theory for developing algorithms for frequency assignments for such problems. Hale has extended or adapted 13 graph colouring methods for the frequency assignment problem, and has implemented these on a minicomputer. The algorithms have been used on thousands of test problems which had large variations in the number of transmitters, the density of transmitters, and the frequency distance rules [Hale, 1981].

Berry and Cronin [1982] have generalized the frequency assignment algorithm to apply to mixed services. Mixed services are characterized by different frequency-distance separation rules, and there may also be frequency-distance separation rules between services. They have shown that if the cross-service separation rules are followed in both cases, mixing services within one large band never takes more total bandwidth than segregating the services in adjacent frequency bands, and usually requires less total bandwidth.

The method described by Český and others [1986, 1987] based on the principles of Metropolis and others [1953] is an alternative approach to the solution of the problem of frequency assignment. This approach uses a Monte Carlo method to minimize the total interference in a network rather than the total bandwidth used. The problem of frequency assignment for the typical task of frequency planning of networks requires only 10^3 to 10^4 iterations and can therefore be solved with a microcomputer.

3. Example applications

In addition to providing very fast solutions for allocation and assignment problems, the algorithms serve as valuable research tools. For example, they may be used in the evaluation of various proposed receiver designs or, more generally, to quantify the relationships between spectrum consumption, equipment characteristics, frequency-distance rules and operational procedures.

As a specific example of such an application, consider the problem of establishing an adjacent-channel frequency distance rule in a new channelized service. The required separation will depend on the kind and quality of service desired, the propagation characteristics in the frequency band, and on emission and equipment characteristics. In particular, it depends on the emission spectrum and the adjacent-channel rejection capability of the receivers. Within rather broad limits, these equipment characteristics are controllable by design parameters. If greater adjacent-channel rejections are required, then adjacent-channel separation distances can be smaller, but the receivers will cost more. A spectrum manager would like to trade-off the equipment costs against the spectrum costs when establishing the separation rules. The computer algorithms described in § 2 enable the minimum spectrum required for each set of separation distances to be determined, and thus provide part of the information for this trade-off. It appears that a general rule for this problem can be developed using the computer program.

The frequency assignment computer program can solve frequency assignment problems involving up to 400 transmitters. The computation time for a particular problem depends on the number of transmitters, the complexity of the frequency constraints, and the algorithm used. The computation time can range from several seconds to several hours.

The computer program has been used for a number of experiments. For most of these experiments, the transmitter locations were generated by a random number generator using a uniform probability distribution and frequency-distance rules were applied to derive the frequency constraints. Data from all experiments are used to determine which algorithms most often produce frequency assignments that use the least bandwidth required and least bandwidth assigned. The "bandwidth required" of a frequency assignment is the total amount of spectrum needed, that is, the difference between the highest frequency assigned to any transmitter and the lowest frequency assigned. The "bandwidth assigned" is the amount of spectrum that is actually assigned to transmitters. Due to adjacent channel constraints, some of the spectrum within the band may not be assigned to any transmitter. Many experiments are concerned with how much bandwidth is required for a frequency assignment as a function of the frequency-distance rules.

Figure 1 displays the overall results of two experiments concerning how the bandwidth required increases as a function of the first adjacent channel distance. Many sets of transmitter locations were generated and several frequency-distance rules were applied to each transmitter set to obtain this data. One frequency-distance rule consists of only a co-channel distance. All other frequency-distance rules include a co-channel distance and a first adjacent channel distance. The solid line represents the average increase in the bandwidth required as a percentage of the average bandwidth required when the first adjacent channel distance is 0. The scale on the vertical axis shows these values. The horizontal axis shows the first adjacent channel distance as a fraction of the co-channel distance. The data show that as long as the first adjacent channel distance is less than approximately 40% of the co-channel distance, the number of channels needed for an assignment is not much more than that needed if the co-channel distance alone were used. However, the bandwidth required increases rapidly as the first adjacent channel distance increases from 40% of the co-channel distance.

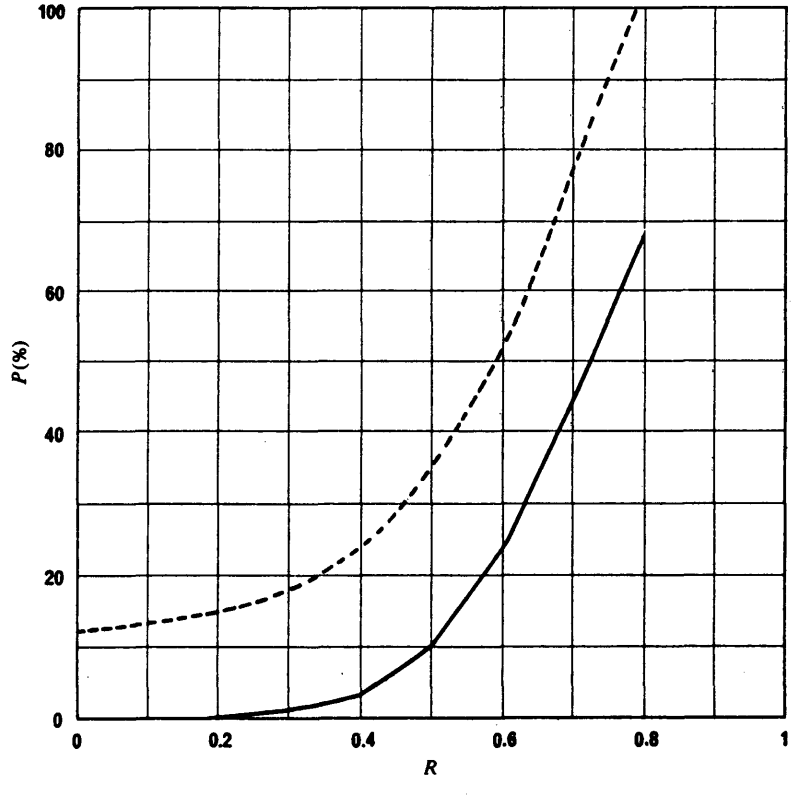


FIGURE 1 – *Percentage increase in required number of channels (P) as a function of the ratio of adjacent channel separation distance to co-channel separation distance (R)*

----- increased bandwidth required if transmitters are assigned in random order

——— best assignment

Also displayed in Fig. 1 are the results of a comparison of near optimum algorithms and a random order algorithm. The random order algorithm assigns the smallest acceptable frequency to transmitters in the order in which they are generated by the random number generator. This algorithm is similar to the current typical method of ordering transmitters to be assigned frequencies. The dotted line in Fig. 1 shows the increase in spectrum used as a percentage of the amount of spectrum needed when the adjacent channel distance is 0. The percentage of additional spectrum used by the random order algorithm over the amount of spectrum used by the best of the other algorithms increases from 12.3% to 22.1% as the first adjacent channel distance increases from 0 to 40% of the co-channel distance. Over this range, the average minimum bandwidth required by the other algorithms increases by only 3.4%. Even a few adjacent channel constraints between transmitters can cause a significant increase in spectrum usage if the transmitters are assigned frequencies in random order. As the first adjacent channel distance increases, the amount of additional spectrum used by the random order algorithm remains at approximately 21% of the amount of spectrum used by the best of the others.

One of the reasons the frequency assignment algorithms developed from graph colouring algorithms produce spectrum efficient frequency assignments is that they choose the order in which the transmitters are assigned frequencies. For many situations, transmitters are assigned frequencies in a random order, that is, the order in which they apply to an administration for a frequency. The algorithms can only choose the frequency to be assigned in these situations. If transmitters are assigned frequencies in groups, however, the algorithms can order the transmitters in each group, leaving unchanged the frequencies assigned to transmitters of previous groups. The percentage of additional spectrum used over that used when the algorithms assign frequencies to all transmitters at once has been found for several sizes of groups of 100 transmitters. Three frequency-distance rules were used to generate this data. Each consists of a co-channel distance and a first adjacent channel distance. The first adjacent channel distances are 40%, 50%, and 60% of the co-channel distance. A group size of 100 transmitters allows the algorithms to order all transmitters. A group size of one transmitter does not allow the algorithms to order the transmitters, i.e. random order. Table I shows the additional bandwidth required as a function of the group size.

TABLE I – *Additional bandwidth required to assign frequencies to 100 transmitters in small groups, compared to the bandwidth required when all 100 are assigned to one group*

Group size	100	50	20	10	1
Increase in bandwidth (%)	0	6.2	12.1	14.3	21.2

4. Example of the output for a small frequency assignment problem

The computer program begins by getting or constructing a constraint matrix. Element (i, j) of the constraint matrix contains the assignment constraints between transmitter i and transmitter j. Table II shows the lower triangle of the (symmetric) constraint matrix for a sample frequency assignment problem of 10 transmitters. All transmitters have the same bandwidth, 10 kHz. The constraint matrix and the bandwidths of the transmitters are the only data needed to define and solve frequency assignment problems. It is not necessary to know how the constraints were obtained. The transmitters are numbered arbitrarily from 1 to 10. Each element of the constraint matrix is an integer, n. For this example, all elements are either 0, 1 or 3. If $n = 0$, there are no constraints between the transmitters; if $n = 1$, the transmitters cannot be assigned the same channel; and if $n = 3$, the transmitters cannot be assigned the same channel or first adjacent channels.

TABLE II – *Lower triangle of constraint matrix for the example in the text*

If element $(i, j) = 0$, there are no constraints between transmitter i and transmitter j; if $(i, j) = 1$, there is a co-channel constraint between transmitters i and j; if $(i, j) = 2$, there is an adjacent channel constraint only; if $(i, j) = 3$, there are both co-channel and adjacent channel constraints.

		Transmitter									
		1	2	3	4	5	6	7	8	9	10
Transmitter	1										
	2	3									
	3	3	1								
	4	1	3	0							
	5	0	0	1	0						
	6	1	3	1	1	0					
	7	3	0	3	0	1	0				
	8	0	0	0	1	0	0	0			
	9	1	0	1	0	1	0	3	1		
	10	0	0	0	0	0	0	0	1	1	

Table III shows the frequency assignment results produced by one of the algorithms implemented in the computer program. The centre frequencies assigned and the degrees of each transmitter are listed for each transmitter in the order in which the transmitters were assigned frequencies. The algorithms determine the order in which the transmitters are assigned frequencies from theoretical graph considerations. Each algorithm assigns frequencies to transmitters in descending order of difficulty of finding an acceptable frequency. The algorithms differ in the method of measuring "difficulty". The degree of each transmitter is also a measure of the difficulty of finding an acceptable frequency to be assigned. The degree is a rough measure of the amount of spectrum forbidden to be assigned to a transmitter. By examining the order in which the transmitters are assigned frequencies and the degrees of the transmitters, one can determine which transmitters have the most frequency constraints imposed upon them and which, in turn, impose the most frequency constraints on other transmitters.

The lower edge of the frequency band for this example is 0 Hz. The algorithm chooses frequencies in such a way that all centre frequencies are multiples of 5 kHz.

In the lower part of Table III, some overall statistics of the frequency assignment are displayed. A clique is a complete sub-group; in this application, a group of transmitters that all constrain each other. The initial clique size is the number of transmitters in the first group to be assigned. The sum of the bandwidths of the transmitters in this clique is also shown. This is a lower bound on the bandwidth required by the frequency assignment. The sum of the bandwidths of the transmitters in the initial clique has been found to be a good estimate of the optimum bandwidth required for frequency assignment problems involving only co-channel constraints or which involve co-channel constraints and very few adjacent channel constraints.

TABLE III – *Frequency assignment results for transmitters in Table II*

The degree is the number of different interactions a transmitter has with other transmitters. The initial clique is the first group of transmitters assigned; all transmitters in the group have mutual constraints.

Transmitter	Centre frequency (kHz)	Degree
1	5	9
3	25	8
7	45	7
9	15	7
2	45	7
6	15	5
4	25	5
5	5	3
8	5	3
10	25	2

Initial clique size	Sum of bandwidths (kHz)	Required bandwidth (kHz)	Assigned bandwidth (kHz)
4	40	50	40

The bandwidth required by the frequency assignment is 50 kHz and the bandwidth assigned is 40 kHz. The frequency interval (30 kHz, 40 kHz) is not assigned to any transmitter due to first adjacent channel constraints between transmitters. The bandwidth assigned is optimum since it is equal to the sum of the bandwidths of the transmitters in the initial clique. The constraint matrix shows that the first three transmitters which are assigned frequencies constrain each other with co-channel and first adjacent channel constraints. Any frequency assignment for these three transmitters alone requires at least 50 kHz. The bandwidth required by the frequency assignment is also optimum.

5. An intermodulation FD rule

While the FD rules for co-channel and adjacent-channel transmitters are relatively straightforward, those for transmitters which might affect other receivers through the formation of intermodulation products require more complicated analysis. In the case of two-signal third-order receiver intermodulation, since two transmitters are involved in the formation of an intermodulation product, their minimum acceptable distances from a victim receiver are interrelated.

Based on the assumptions that the receiver antenna gain is equal to the receiver total loss, that the average value of minimum wanted signal level to produce 12 dB (SINAD) in the presence of noise is -145 dBW, that free-space path loss is used and that all transmitters have the same e.i.r.p. equal to 20 dBW, the FD rule for the 410-470 MHz band can be established using the empirical model described by McMahon [1974] to predict interfering power levels at the victim receiver. In this model,

$$P = 2\ P_N + P_F - 0.57 - 60 \log \delta f$$

where:

P: interfering power level at the victim receiver (dBW),

P_N: received power level from the near transmitter (dBW),

P_F: received power level from the far transmitter (dBW),

δf: frequency separation between the far and near transmitter frequencies (MHz).

By using a carrier frequency value equal to 460 MHz, the relationship between frequency and distance separations was derived by Chan [1985] who stated that two-signal third-order receiver intermodulation would occur if:

$$d \cdot \delta f \leqslant 0.17$$

where:

d: distance of an existing station from the proposed station.

A protection margin of 6 dB between the interfering power level and the wanted power level has been assumed.

Since the proposed station may be involved as a victim receiver, a far transmitter or a near transmitter in an intermodulation product, the curve of $2d$ versus $2\delta f$ has to be used with the d versus δf curve in establishing the FD rule which is depicted in Fig. 2. The area above the curve corresponds to permissible interference situations, while that below corresponds to potential interference situations. In order to assign a frequency to a proposed new station, co-channel and adjacent channel interference is first evaluated using the appropriate FD rules. After these rules are satisfied, existing stations which may be involved in intermodulation interference with the proposed station are then identified in a selection process which is subsequently followed by a detailed analysis to determine the interference power levels. The selection is performed by choosing values of d and δf to efficiently approximate the curve.

If the locations and frequencies of radio stations are stored in a data base, a computer program can be designed to do a simple search based on the above rule. However, if computer assistance is not available, any manual evaluation of a proposed frequency assignment can still be facilitated using this rule in conjunction with a rudimentary pin map.

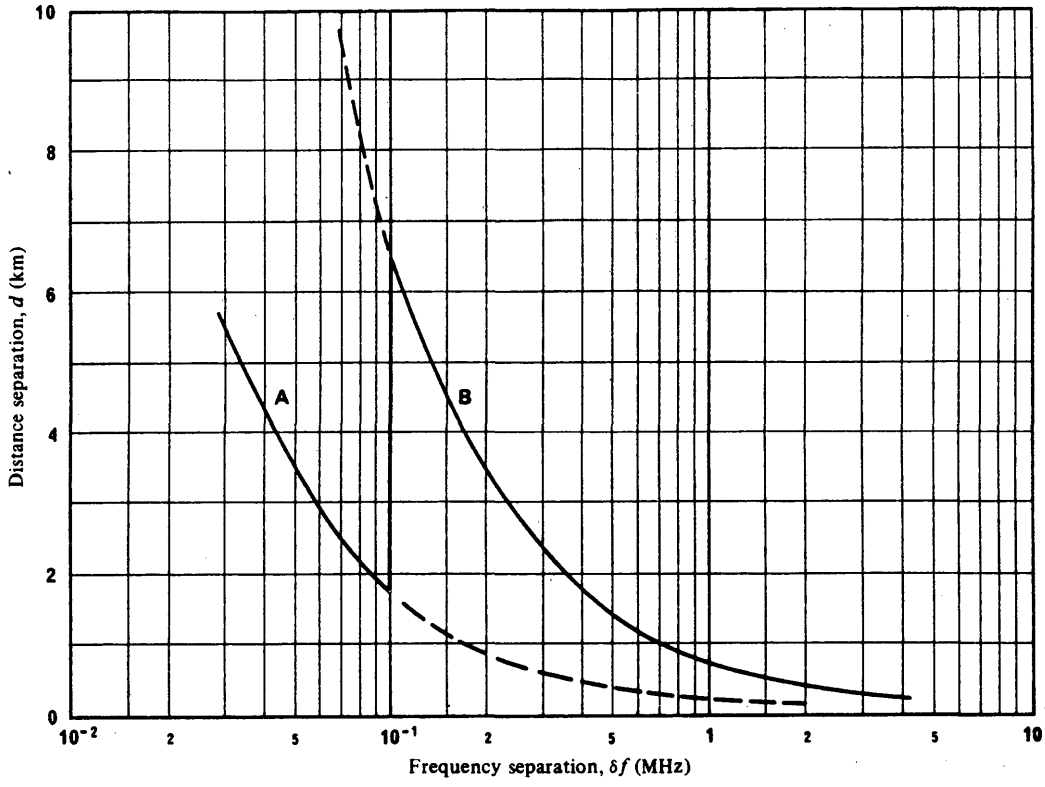


FIGURE 2 – *An FD rule for two-signal third-order receiver intermodulation interference analysis*

——— : intermodulations selection curve
Curves A: d as a function of δf
B: $2d$ as a function of $2\delta f$

6. Choice of a frequency planning algorithm

In view of the inadequate efficiency of existing non-formal spectrum management methods and the growing shortage of free frequencies, increasing attention is being paid to efforts to make use of formal methods based on the interpretation of the problem of rational frequency planning in terms of mathematical programming problems. The selection of a specific frequency assignment algorithm [Lyandres et al., 1987], if soundly based, is of importance since it can result in a more rational use of the frequency resource.

The approximate algorithms are based primarily on consideration of the characteristic features of the initial data of the problem. Given a set of possible frequency planning algorithms therefore it is possible to introduce a measure of "adequacy" for the initial data; when the planning system has to be practically implemented, this involves introducing into it a procedure for analyzing the initial data and defining one or more algorithms which are "adequate" for the data.

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The characteristics of the frequency assignment problem that are easiest to evaluate include N- the number of requirements, M- the number of available frequencies, and D_i - the density or fraction of pairs of stations that must be separated by i-Channels.

The interpretation of frequency planning problems as graph-colouring problems [Hale, 1980] makes it possible: 1) to reduce the technical problem to mathematical terms and to apply known methods of solution, 2) thanks to the clarity of representation provided, to construct new algorithms and 3) to compare the various frequency planning algorithms on an adequate basis and, thus, to obtain well-founded Recommendations on the most rational use of the frequency scheduling algorithms which will provide maximum efficiency for the planning system.

Furthermore, by using a graph-colouring interpretation, it is possible to state with confidence that a frequency planning problem is one of exponential complexity, i.e. as the number of requirements grows, the time needed to find a rigorously optimum solution grows faster than the polynomial of any order, at least asymptotically. This in turn makes the primary problem one of developing more effective approximate algorithms which do not guarantee that a precise solution to an optimization problem will be obtained over a limited period but which have a high probability that such a solution will be obtained. Algorithms of this sort possess polynomial complexity [Hale, 1980] and are heuristic.

Any algorithm for vertex colouring performs an iteration procedure, the elementary step of which consists in the selection of a specific vertex from a set of vertices which are not coloured for the iteration, the identification of a subset of colours available for the vertex and in the assignment of one of these colours to the vertex. Thus, any algorithm is defined by means of two rules: the choice of vertex and the choice of colour. Obviously, the subset of available colours (frequencies) for the vertex (requirement) selected must be made up in such a way as to take into account the colour of all the vertices coloured before it, which is equivalent to taking into account the assignments made before the requirement was dealt with.

It is the rules for the selection of the requirement and the frequency in developing approximate planning algorithms which are heuristic in character; the algorithms themselves may be classified in one of two classes: sequential and recursive.

Sequential algorithms are made up on the basis of an analysis of the initial data in accordance with a rule governing lists of requirements in accordance with which they are dealt with one after the other as the frequency schedule is compiled.

Recursive algorithms begin the assignment process with one or another way of selecting the requirement, and the selection of the next one depends on what frequency was assigned to the first, etc. The handling of each following requirement naturally reduces the set of still unsatisfied requirements and also the number of frequencies available for assignment to requirements.

Some of the known algorithms can be naturally generalized to cover the colouring of weighted graphs, in other cases such a generalization is hardly possible. The latter include in particular one of the best known colouring algorithms proposed by [Ershov, A.P., 1977]. The first category includes two sequential algorithms (described below) - the so-called "largest first" or MAXFIRST [Welsh and Powell, 1967] and the "smallest last" of MINLAST [Zoellner, J.A., Beall, C.L., 1977], and a recursive algorithm proposed by [Brelaz, 1979].

There are not many data in the literature on the comparative evaluation of algorithms and they are often contradictory. For example, [Zoellner and Beall, 1977] state that, of the two sequential algorithms MAXFIRST and MINLAST, the best in almost all vertex colouring problems is the second, but [Kubale and Dabrowski, 1978] consider the first to be definitely preferable. In this connection, the conclusion of [Zoellner and Beall, 1977] on the superiority of MINLAST even for problems involving weighted graphs is open to doubt. The recursive algorithm of [Brelaz, 1979] has potentially better characteristics than the sequential algorithms but this is achieved at the expense of operating speed and there are no data on its use with the colouring of weighted graphs.

An analytical study of the problem of heuristic algorithms for frequency assignment can practically be ruled out. Only a simulation experiment can answer the questions involved in the selection of a particular algorithm for the solution of a specific problem. This has in fact been done with a number of problems which differed in the number of requirements in a group and the densities of the edges of different weights.

The density D_0 is the number of pairs of requirements that can be co-channel (assigned the same frequency) divided by N x N; D_1 is the number of pairs of requirements that can be assigned adjacent channels (but cannot be co-channel) divided by N x N; etc.

The number of cases of each type was selected having regard for the need for providing sufficient statistical material for averaging purposes and also in the light of the limited machine time available. The numerical values for the case characteristics are as follows: 10 requirements - 100 cases, 20 requirements - 50 cases, 40 requirements - 20 cases, 80 requirements - 10 cases, 160 requirements - 5 cases; the density of the edges in the unweighted graphs $D_1 = 0.25, 0.50, 0.75$; in the graphs with a maximum edge 2, combinations of densities $D_1 = 0.5$, $D_2 = 0.16$; $D_1 = 0.33$, $D_2 = 0.33$; $D_1 = 0.16$, $D_2 = 0.5$; in graphs with a maximum edge 3, combinations of densities $D_1 = 0.5$, $D_2 = 0.125$, $D_3 = 0.125$; $D_1 = 0.25$, $D_2 = 0.25$, $D_3 = 0.25$; $D_1 = 0.125$, $D_2 = 0.375$, $D_3 = 0.25$.

The following method was adopted in carrying out the experiment. For each set of parameters (N, D) a group of cases was generated. The problem of the frequency assignment was solved using each of the algorithms under study with two densities: $D_1 = 0$ and $D_1 = 0.25$. The number of available frequencies varied from M = 1 to Mmax, equal to the number of frequencies required by the algorithm to satisfy all the requirements. At the same time the fraction of requirements using M frequencies and also the algorithm operation processor time were determined. The following algorithms were studied: (MAXFIRST), (MINLAST), (BRELAZ), (DSATURB), (RANDOM), (TRIVIAL). The last algorithm (TRIVIAL) does not contain any requirement ordering procedure and was modelled to obtain an evaluation of the effectiveness of introducing this procedure into other algorithms. For all the sequential algorithms (MAXFIRST, MINLAST and TRIVIAL) three modifications were studied. In the first and the second of these the selection of the requirement preceded the selection of the frequency. The term "first requirement - then frequency" may be used in this case. In the third modification all the possible requirements were assigned with the frequency selected in an ordered sequence ("first frequency - then requirement"). The second modification differed from the first in that scanning of the frequencies for the subject for which they were to be used was carried out in the frequency preliminary numbering order and not using the procedure (described above) for ordering the frequencies on the bases of the principle of "maximum accessibility" (discussed above).

Thus the total number of algorithms to be studied taking account of the modification of the sequential algorithms was 12.

The results of the experiment can be summarized as follows:

The differentiation between the algorithms, i.e. the difference in their efficiency, increases as the dimensionality of the problem increases.

The differentiation between the algorithms increases as the density increases.

In the areas of maximum differentiation between the algorithms the number of frequencies required vary within a range of 30% of the result obtained with the TRIVIAL algorithm.

When the shortage of frequencies is great the ordering principle "the difficult requirements are assigned first" provides no advantage and the TRIVIAL algorithm is as good as the others. As the shortage of frequencies decreases the TRIVIAL algorithm begins to lose in comparison with the other algorithms, the loss being at a maximum when there is no shortage.

When $D_1 > 0$ there is a gain for the modification of sequential algorithms with the frequencies chosen on the bases of the "minimum availability" principle. This gain is as high as 15% with the MAXFIRST algorithm. When $D_1 = 0$ the modification "first frequency - then requirement" out-performs both modifications "first requirement - then frequency".

The execution time of finite algorithms depends on the dimensionality of the problem and on a function of the vertex ordering procedure used, varying between quadratic and cubic. The fastest algorithm is the sequential (MAXFIRST).

There is a stable time gain in the (MAXFIRST) algorithm as compared to the (MINLAST) algorithm and the efficiency is practically the same. This gain increases as the dimensionality of the problem increases.

Another effective algorithm for the solution of frequency planning problems, for groups of radio-electronic equipment with a frequency separation matrix which is completed by the elements of a set {0, 1, 2} (i.e., where the triangle rule is applied for elements), is presented in [CCIR, 1989] and [Bessonov, A.P et al., 1988].

Results of calculations show that the systematic improvement, from 4% to 25%, in the total necessary band for a group of radio-electronic equipment was produced by this algorithm in relation to the heuristic algorithms mentioned above. However, it should be mentioned that computer time used by this algorithm is somewhat greater.

7. Conclusions

Approximate algorithms that enable frequencies to be assigned to a collection of transmitters, so that an arbitrary given set of frequency-distance separation rules or more general frequency constraints are satisfied and that minimize the total bandwidth needed, have been implemented. Such algorithms can be used as service allocation tools, for routine frequency assignments, and for studying the general principles of spectrum management. Example application produced curves show the additional spectrum required to satisfy an adjacent-channel separation rule as a function of the separation distance. A receiver intermodulation FD rule allows a simple selection strategy to be used. Similar general principles can be developed for more complex sets of frequency-distance rules as guidance for spectrum managers.

The choice of an efficient frequency assignment algorithm depends on the initial data of the problem, i.e. the matrix of frequency restrictions and the available frequency resource. Account should be taken of the analysis indicated in automated frequency assignment systems.

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REPORT 655-2*

METHOD OF CALCULATING INTERMODULATION INTERFERENCE IN THE LAND MOBILE SERVICE

(Question 44/1)

(1978-1986-1990)

1. Introduction

Efficient spectrum utilization in land mobile service depends on the analysis of environmental and system parameters that can best be described statistically.

This Report describes a statistical method for investigating land mobile interference that includes techniques which minimize mutual interference and optimize overall spectrum utilization. Details of land mobile receiver models and the statistical interference evaluation model are described. Finally, details of an investigation to which an automated version of the models was applied are described.

2. Outline of basic models

2.1 *Receiver model*

The wanted signal at the output of the receiver with both wanted and interfering input signals is chiefly degraded by three causes: receiver intermodulation, co-channel interference and adjacent-channel interference. The model of these modes of interference was derived from measured data on eight receivers. The measured variation between receivers was very slight. A 12 dB output $(S + N + D)/(N + D)$ criterion was used (see Recommendation 331).

2.1.1 *Receiver RF attenuation characteristics*

The receiver radio frequency (RF) attenuation characteristics have a direct effect on the level of both the receiver intermodulation and the adjacent-channel interference as described in Report 522.

The equation for the attenuation of a signal at an off-tune frequency was obtained from measured values and was estimated by applying $N = 6$ and $Q = 75$ to equation (4) of Report 524. Consequently the modified attenuation equation becomes

$$\beta(\Delta f) = 60 \log [1 + (150\Delta f/ f_0)^2] \quad (1)$$

where:

$\beta(\Delta f)$: receiver RF attenuation at Δf MHz (dB),

Δf: frequency separation between tuned and off-tune frequencies (MHz),

f_0 : tuned frequency (MHz).

* This Report should be brought to the attention of Study Group 8.

2.1.2 *Intermediate-frequency attenuation characteristics*

Interference caused by the desired and interfering signal being present simultaneously in the same channel is related to the bandwidth of the intermediate frequency (IF) amplifier. In this Report, the bandwidth of the IF amplifier is 16 kHz for the 20 kHz channel spacing considered.

2.1.3 *Co-channel rejection characteristics*

The co-channel rejection characteristics were derived from measurements of the signal level relationships at the input of the receiver using a desired signal modulated by 1 kHz tone and an interfering signal modulated by 0.4 kHz tone. Both signals are located in the centre of passband of the receiver; using a 12 dB $(S + N + D)/(N + D)$ output criterion the following conditions were obtained from the measured data:

$$P_u > P_d - 2.0 \quad \text{for} \quad -103 \text{ dBm} \leqslant P_d \leqslant -73 \text{ dBm} \tag{2}$$

where:

P_d: mean received power of a desired radio signal (dBm),

P_u: mean received power of an interfering radio signal (dBm).

Figure 1a shows the co-channel interference characteristics.

2.1.4 *Adjacent-channel interference characteristics*

The adjacent-channel interference characteristics were found from the following measurement conditions. A desired signal was modulated by a 1 kHz tone, and an interfering signal in the adjacent channel separated by 20 kHz was unmodulated. Using a 12 dB $(S + N + D)/(N + D)$ criterion with a desired and an interfering signal present simultaneously, the following equation was obtained from the measured data:

$$P_u > 0.75\ P_d + 57.5 + \beta(\Delta f) \quad \text{for } -103 \text{ dBm} \leqslant P_d \leqslant -73 \text{ dBm} \text{ and } 0.02 \text{ MHz} \leqslant \Delta f \leqslant 28 \text{ MHz} \tag{3}$$

where:

P_d: mean received power of desired radio signal (dBm),

P_u: mean received power of interfering radio signal (dBm),

$\beta(\Delta f)$: radio-frequency attenuation at Δf MHz (dB).

Figure 1b shows the adjacent-channel interference characteristics.

2.1.5 *Receiver intermodulation interference characteristics*

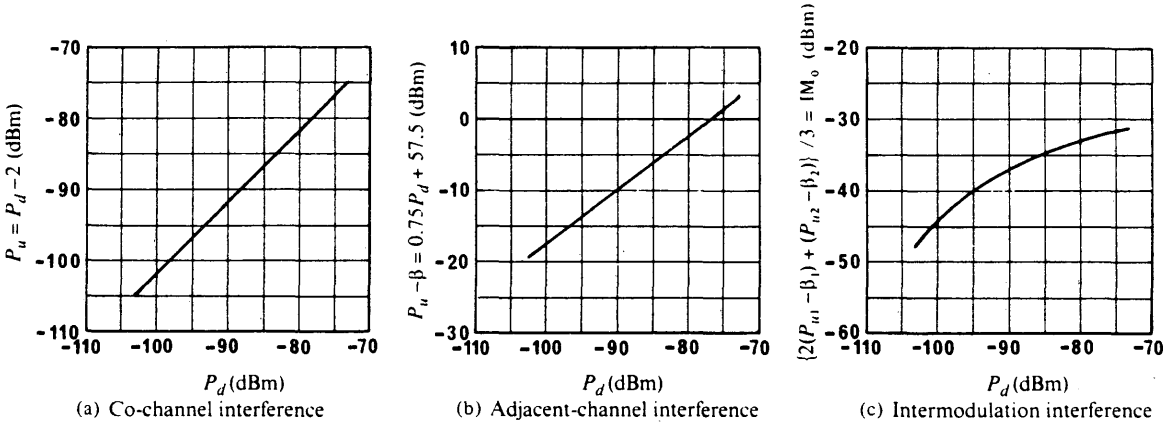
In principle, receiver intermodulation products can occur for a large number of frequencies in the non-linear circuits of the RF amplifier and mixer where two or more interfering signals are present.

Only two-signal third-order intermodulation is treated in this model because three-signal third-order intermodulation interference is a rare case in actual operation. Furthermore, the level of interfering signals required to produce two-signal fifth-order intermodulation must be about 15 dB higher.

The measurement of two-signal third-order receiver intermodulation was carried out keeping a 12 dB output $(S + N + D)/(N + D)$ ratio, and two interfering signals with the same level were modulated by the weighted noise described in CCITT Recommendation G.227 (Yellow Book, Vol. III.2).

It was found that the effect of interference could be represented according to the application of a least square approximation curve to the measured data.

Figure 1c shows the two-signal third-order receiver intermodulation.



(a) Co-channel interference (b) Adjacent-channel interference (c) Intermodulation interference

FIGURE 1 — *Level diagrams in receiver model; the solid line shows the measured results in 12 dB (S + N + D)/(N + D) ratio*

2.1.6 *Limitation of receiver model*

2.1.6.1 The model is based on measured data obtained from the averaged eight receivers which passed the type approval test required for receivers used in the land mobile telephony service that operate in the 142 MHz to 170 MHz band in Japan.

2.1.6.2 A 12 dB $(S + N + D)/(N + D)$ ratio is used as a receiver output criterion in the presence of the desired and interfering signals.

2.1.6.3 The model only considers the important case of two-signal third-order intermodulation.

2.1.6.4 Spurious responses are not included, since the frequency relationships producing spurious response exist in most cases outside the frequency band being considered.

2.1.6.5 Reference sensitivity is assumed to be -113 dBm for a 12 dB $(S + N + D)/(N + D)$ ratio.

2.2 *Land mobile operational model*

2.2.1 *Traffic density classification*

The average rates of both duration and the interval between calls are different in different businesses. These rates are classified into three typical business groups on the basis of measurements of actual operations.

2.2.2 *Connection between stations*

Channel utilization of any mobile station in a particular network is assumed to be approximately the same.

2.2.3 *Location of radio station*

The service area is divided into a number of squares with a side 0.3 km long, and a logical address is assigned to each of these squares. This value is representative of average busy durations for a vehicle speed of 40 km per hour.

The fixed ratio stations, such as base stations, have the address specified by input data. The location of each land mobile station is mathematically treated according to the address accompanied with a random walk process.

3. Particular application

In addition to the receiver and the operational models, it is necessary to develop models for the propagation and the external noise.

3.1 *Input data*

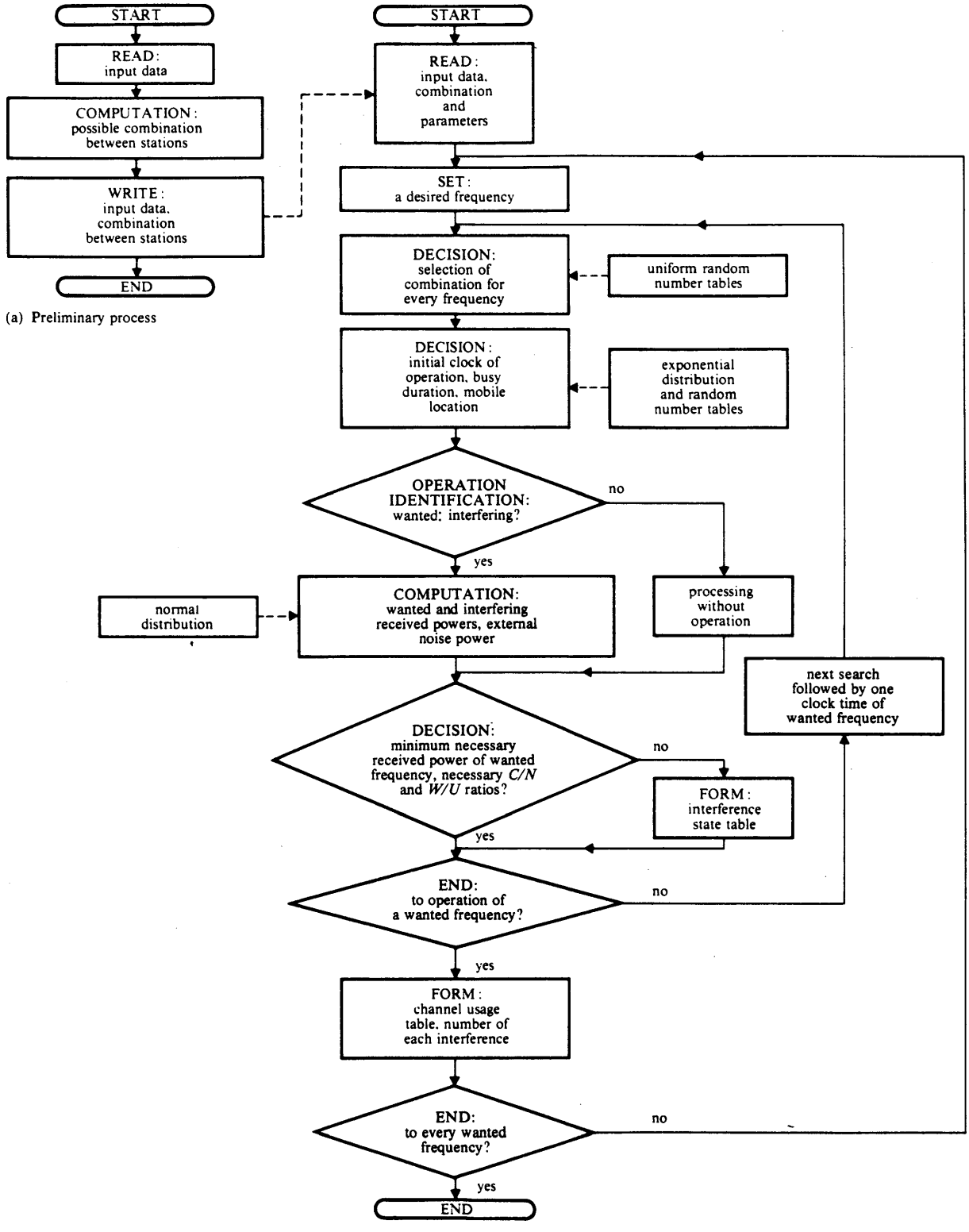
The input data involve the following main items extracted from the frequency list and the specification of radio equipment:

- name of licensee,
- assigned frequency,
- class of radio station,
- effective radiation power,
- height and directivity of antenna,
- fixed location of base station and movable range of mobile station.

3.2 *Outline of statistical prediction method*

The general land mobile statistical prediction method is summarized in the flow-chart in Fig. 2. A preliminary processor examines each element of radio equipments specifications and the computed results of potential interference combinations are stored.

The main process is carried out by initially picking a desired frequency to be examined. The selection of radio stations which can be connected, starting time and duration in operation, and locations of mobile stations are representively decided by using the operational model for different business groups and for all assigned frequencies. If a specified wanted signal and an interfering signal emit simultaneously, the received power at these frequencies is computed using the propagation model. In this case, if the transmitting and receiving stations are located in the same area, the location of the transmitting station is chosen at random as one of the surrounding logical addresses. In addition, the power ratio of the carrier-to-noise and wanted-to-interfering signals are calculated from the external noise and the propagation models respectively. When these power ratios are less than the pre-determined threshold values, the wanted frequency is treated as not communicating.



(a) Preliminary process

(b) Main program

FIGURE 2 — *Outline of computer simulation*

Secondly, the above process is repeated in conformity with the time advanced by the fraction of call intervals composed of a busy duration and empty call at a specified wanted frequency.

Finally, the statistical determination of interference is made by the main program using the random numbers generated by the distributions (e.g. the received power has a log-normal distribution).

3.3 *Application by computer simulation*

Thirty-three radio frequencies (284 stations) ranging from 146.06 MHz to 163.57 MHz in the 150 MHz band are used as an example. The channel separation is 40 kHz at the minimum and 3.14 MHz at the maximum. The 36 km^2 service is divided into 20 square cells and each cell except for the radio-relay stations, has its own identification address. Median propagation loss is calculated using the result of [Egli, 1957]. The external noise model was derived from Nakamura's data [1959].

Table I shows the simulated results which are equivalent to about 4 hours of actual operation. The output signal quality of the receiver is extremely sensitive to the external noise interference compared with the receiver interference. This is considered to be normal in reference to the interference source from many operators in this city. As a whole, the individual channel usage reflects the various specifications of each station such as effective radiated power, antenna height, business group, etc.

TABLE I - *Results of interferences of a computer simulation*

Input			Traffic density (%)	Total number of events	Event of desired signal below -113 dBm	Total number of interferences (2)				Channel usage	
Licensee name	Assigned frequency (MHz)	Business group				External noise	Co-channel	Adjacent channel	IM	Total number	Rate (%)
2512	146.06	2	28	182	26	10				146	80
1103	146.20	1	38	242	70	22				150	62
1103	146.24	1	35	233	72	26				135	58
1103	146.34	1	41	251	55	25				171	68
1106	148.01	1	41	290	28	17		1		244	84
1411 (1)	148.09	2									
1106	148.29	1	43	283	38	28				217	77
1103 (1)	148.47	1									
1919 (1)	148.85	2									
1408 (1)	149.03	2									
0108	149.17	2	33	227	0	3				224	99
2511	149.65	2	29	171	21	8				142	83
2508	149.81	2	27	189	8	3				178	94
0404	150.25	2	27	183	12	18				153	84
1119	150.29	3	42	206	33	19				154	75.
1919 (1)	150.89	2									
1911 (1)	151.05	3									
0611	151.33	3	42	216	6	19				191	88
2014	151.37	3	42	238	4	3	30			202	85
2019	151.37	3	40	215	6	14	46	1		154	72
1411 (1)	151.61	2									
1907	151.81	3	44	202	32	18				152	75
0404	152.25	2	26	167	7	12				148	89
0711	152.33	2	29	197	9	5				183	93
1103 (1)	152.55	1									
2004	153.97	2	22	153	12	15				126	82
2520	154.17	3	50	245	7	19				219	89
2005	154.59	3	44	213	12	12				189	89
1419	157.73	1	40	309	7	10				292	94
1420	159.05	3	41	202	11	14				177	88
1103	162.05	1	38	249	91	30				128	51
2025	162.93	2	30	159	13	6				140	88
1408 (1)	163.57	2									

(1) Frequencies of radio-relay system independent of desired signal only.
(2) In the case that there are more than two interferences in one clock time, each interference is counted separately.

4. Statistical method of calculating intermodulation interference and determining the zones affected by it

Models similar to those described above may be used to calculate the probability of interruption of communications as a result of intermodulation products affecting receivers and transmitters of radio stations in the land mobile service, and also for the statistical determination of zones affected by such interference.

One such exercise, based on an analysis of third-order intermodulation products, is described by Zhidkov and Klyzendo [1984]. The model of receiver intermodulation characteristics presented in § 9 of Report 522 (Alter [1986]) was used.

4.1 *Third-order intermodulation product interference*

Interference caused by third-order intermodulation products in a receiver occurs when the following into conditions are fulfilled:

$$2f_1 - f_2 \leqslant f_0 \pm B/2 \tag{4}$$

$$2P_1 + P_2 > 3W \tag{5}$$

where:

f_1, f_2 : frequencies of interfering transmitters;

f_0 : receive frequency;

B : IF receiver pass-band;

P_1, P_2 : interfering transmitter signal power at receiver input due to interfering transmitters (dBW);

W : receiver susceptibility to intermodulation interference (dBW).

The correlation for determining of susceptibility, based on measurements, has the following form [Alter, 1980; Zhidkov and Klyzhenko, 1984]:

$$W = D_3 + dP_s + \mu\log(\Delta f/\Delta f_{bm}) + P_{s0}(1 - d) \tag{6}$$

where:

D_3 : three-signal receiver discrimination (dB);

P_s : wanted signal power at receiver input (dBW);

P_{s0} : receiver sensitivity for a signal-to-noise ratio of 12 dB (dBW);

$\Delta f = |f_1 - f_2|$

d : coefficient characterizing susceptibility as a function of wanted signal power P_s;

Δf_{bm} : half pass-band of receiver radio-frequency amplifier (kHz);

μ : gradient of receiver radio-frequency amplifier frequency response curve (dB/decade).

The parameters d, Δf_{bm} and μ are determined experimentally by the three-signal measurement method.

In view of equation (6), condition (5) assumes the form:

$$2P_1 + P_2 - 3dP_s > 3R_0 \tag{7}$$

where:

$$R_0 = D_3 + \mu\log(\Delta f/\Delta f_{bm}) + P_{s0}(1 - d)$$

4.2 *Combining signals*

Recommendation 370 and Report 567 point out that, due to fading, the wanted and interfering signal levels are random values with a log-normal distribution. Hence the left side of condition (7), expressed in dBW, represents the sum of independent normal random quantities and constitutes a normal random quantity. The mean value and dispersion of the random quantity $R = 2P_1 + P_2 - dP_s$ are equal respectively to:

$$\overline{R} = 2P_{1m} + P_{2m} - 3dP_{sm}$$

$$\sigma_R^2 = 4\sigma_1^2 + \sigma_2^2 + 9d^2\sigma_s^2$$

where:

P_{sm}, P_{1m}, P_{2m}, σ_s^2, σ_1^2, and σ_2^2 are median values and dispersions of wanted and interfering signal power levels at the receiver input (determined on the basis of the data contained in Recommendation 370 and Report 567).

4.3 *Power of intermodulation products*

The power P_i of the intermodulation product occurring in the transmitter and subsequently reaching the receiver input may, according to Report 839, be written as:

$$P_i = P_2' - \beta_{12} - \beta_{10} - k_{2,1} - L_{10} \qquad \text{dBW} \tag{8}$$

where:

P_2' : interfering transmitter power (with frequency f_2) at the output terminals of the affected transmitter (with frequency f_1), in which the intermodulation products occur (dBW);

β_{12}, β_{10} : attenuation due to the output and antenna circuits of the affected transmitter of frequency f_1 to interfering transmitter signals (f_2) and to intermodulation product (with frequency f_0) respectively (dB);

k_2,1 : intermodulation conversion losses in the transmitter (dB);

L_{10} : attenuation of intermodulation product on the path between the transmitter with frequency f_1 and the receiver (dB).

The intermodulation products occurring in the transmitter affect the operation of the receiver on fulfilment of condition (4) and of the condition:

$$P_s - P_i < A \tag{9}$$

where A is the co-channel protection ratio (see Report 358).

Taking account of equation (8), condition (9) assumes the form:

$$P_2' - P_s - L_{10} > T_0 \tag{10}$$

where:

$$T_0 = \beta_{12} + \beta_{10} + k_{2,1} - A$$

The mean value and dispersion of the random quantity $T = P_2' - P_s - L_{10}$ are equal respectively to:

$$\overline{T} = P_{2m}' - P_{sm} - L_{10m}$$

$$\sigma_T^2 = \sigma_2^2 + \sigma_s^2 + \sigma_1^2$$

where:

P'_{2m}, P_{sm}, L_{10m}, σ_2^2, σ_s^2 and σ_1^2 are the mean values and dispersion of the random quantities P'_2, P_s and L_{10}.

4.4 *Probability of intermodulation products*

The probability that intermodulation products generated both in the receiver itself and as a result of intermodulation in the transmitter (conditions (7) and (10) respectively) will occur during reception is equal to:

$$\alpha = 1 - \Phi(x) \tag{11}$$

where:

$\Phi(x)$: integral of probabilities;

$x = (3R_0 - \overline{R})/\sigma_R$: on determination of the probability of intermodulation products occurring in receivers (condition (7));

$x = (T_0 - \overline{T})/\sigma_T$: on determination of the probability of interference due to intermodulation products occurring in transmitters (condition (10)).

In determining the zones affected by intermodulation interference on the basis of a given value of probability of interference α, the value of x is first determined from equation (11) and then, for a known value of P_{sm}, the permissible values of P_{1m} and P_{2m} (or P'_{2m} and L_{10m}) and the corresponding necessary geographical spacings of interfering transmitters and receiver, on which the zone affected by the interference will depend.

5. Summary

Receiver and operational models have been developed for the land mobile service based on measured data. Channel usage for a typical city has been calculated using a statistical evaluation of four types of interferences. It is expected that the development of interference prediction models that include basic models and statistical analysis will offer a useful means of promoting efficient frequency utilization for the land mobile service.

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REPORT 525-3

PERFORMANCE FUNCTIONS AND PROTECTION RATIOS OF ANALOGUE AND DIGITAL MODULATION SYSTEMS REQUIRED FOR SPECTRUM UTILIZATION

(Question 45/1)

(1974-1978-1982-1990)

1. Introduction

This Report summarizes protection ratios and prediction of performance for various combinations of desired and undesired modulation required to investigate spectrum utilization problems. The protection ratios are given for three levels of performance and are applicable to different classes of emission.

To be able to investigate a range of spectrum utilization problems, a compendium of protection ratios is desired. This Report utilizes the results and concepts of CCIR and other publications [ECAC, 1973] and summarizes protection ratios for desired and undesired modulation cases appropriate to a wide range of modulation types and threshold criteria.

This report also presents curves that may be used in the estimation of possible degradation due to interference and noise to systems that use various digital modulation techniques. The curves in this report may be used to predict probabilities of errors for communication channels experiencing interference and noise using devices such as digital demodulators, teleprinters, and digital-to-analogue converters.

2. Definition of the term "protection ratio"

The World Administrative Radio Conference (Geneva, 1979) agreed on the following definition (No. 164 of the Radio Regulations):

Protection ratio (RF): The minimum value of the wanted-to-unwanted signal ratio, usually expressed in decibels, at the receiver input, determined under specified conditions such that a specified reception quality of the wanted signal is achieved at the receiver output.

For the practical application of the WARC-79 definition given above the following points should be noted:

- the wanted-to-unwanted signal ratio is expressed in terms of a power ratio;
- this power ratio is usually expressed in terms of a mean power ratio, but may also be expressed in terms of a peak envelope-to-peak envelope, peak envelope-to-mean, mean-to-peak envelope or any other power ratio, as appropriate for the case under consideration;
- the specified reception quality is expressed in terms of a given level of performance, such as the output signal-to-interference power ratio, the articulation index *AI*, the bit error ratio, the character error rate, or the Television Allocation Study Organization (TASO) scoring grade (see Table II).

- the specified reception quality may be expressed by the double-stimulus impairment scale method in Recommendation 500 (see) Table III).

3. Protection ratios for various combinations of wanted and interfering emmissions

The protection ratios are given for three performance levels and a co-channel (the frequency separation between the desired and undesired signal is approximately zero) and off-channel (the frequency separation between the desired and undesired signal is less than half the necessary bandwidth of the wanted signal) interference. The threshold levels range from a minimum interference threshold (MINIT) to a maximum interference threshold to categorize, on the basis of parameters, the limiting definitions of performance degradation criterion.

The MINIT for a voice signal [ECAC, 1973] denotes the boundary between a region in which there is no interference and a region in which there is detectable interference, but no degradation to intelligibility at this point. MINIT is not independent of the signal-to-noise ratios and for an output signal-to-noise greater than 10 dB, the MINIT is a linear function of the signal-to-noise ratio. The MINIT is discussed in more detail in Report 526. The MINIT is obtained from a subjective evaluation process for audio information systems. This also corresponds to a 1.0 articulation index (*AI*) score. For a voice signal a 0.7 *AI* score is used for the second threshold to separate the region within which there is no degradation to intelligibility and that where there is degradation to intelligibility [Kryter and Licklider, 1963]. For a voice signal a 0.3 *AI* is used to separate the region between a degraded but usable signal and an unusable signal [Kryter and Licklider, 1963].

Performance degradation levels in voice systems have also been defined according to the subjective terms denoted in Table I. The three subjective criteria are basically defined in terms of the receiver output desired signal-to-noise ratios of 33 dB, 15 dB and 6 dB, as measured with a VU-meter (Recommendation 339). These ratios correspond to power ratios of 32, 14 and 5 dB respectively [BTL, 1970]. Although power ratios are generally useful in denoting approximate voice quality criteria, they are not unique voice intelligibility degradation measures. The above audio signal-to-noise power ratios are also used for the interfering signals listed in Table IV. For the purpose of illustration, the subjective measures associated with the above signal-to-noise power ratios are applied to the corresponding protection ratios. It has been found that for all possible cross combinations of AM, FM and SSB wanted and interfering signals, the audio signal-to-interference ratio required for a constant *AI* of 0.7 and 0.3 has a standard deviation of approximately 2.3 dB and 3.4 dB, respectively [Sorkin *et al.*, 1968]. This indicates that for the cases measured, an audio output signal-to-interference power ratio is a reasonably good measure of voice intelligibility. The articulation index score [French and Steinberg, 1947] is a measure of voice intelligibility and consequently can be used to denote various degradation regions. Since both the articulation index scores and the subjective definition in terms of power ratios have been widely used, results for both criteria are given for comparative purposes.

The minimum bit error ratio for a digital system has been chosen as a bit error probability of 10^{-6}. The maximum and middle bit error ratio for digital systems has been chosen as a bit error probability of 10^{-2} and 10^{-4}, respectively. The thresholds have been given in terms of bit error probabilities instead of character error rates, so that the results will be applicable to all systems independent of message structure. The results that have been given can be approximately converted to character error rates with the methods discussed in Report 195. It should be noted that this conversion is approximate since the distribution of the receiver output noise and interference is only approximately gaussian under special circumstances.

The minimum interference level for television systems is given in terms of a TASO score of 1. The middle and maximum criteria are given in terms of TASO scores of 2 and 2.5 respectively. Definitions of TASO grading criteria are shown in Table II.

TABLE I

Voice degradation criteria

Subjective degradation evaluation	*A I* Criteria	Signal-power to noise-power ratio at the receiver output (dB)
Minimum interference threshold (MINIT) (1)	(3)	≥ 32
Good commercial quality	(3)	32
Marginally commercial	0·84	14
Threshold of intelligibility degradation	0·7	9
Just usable quality (operator to operator)	0·57	5 (2)
Maximum interference threshold	0·3	−3

(1) This is equal to or greater than the "Good Commercial Quality" criterion.

(2) This signal-to-noise criterion corresponds to a sentence intelligibility of greater than 90% and a word intelligibility of approximately 80% (for a 1000, phonetically balanced, word list).

(3) The signal-to-noise power ratio required for MINIT and Good Commercial Quality are much greater than that required for an AI of 1.

TABLE II

TASO scoring grades *

Grade	Designation	Description
1	Excellent	The picture is of extremely high quality, as good as you could desire.
2	Fine	The picture is of high quality providing enjoyable viewing. Interference is perceptible.
3	Passable	The picture is of acceptable quality. Interference is not objectionable.
4	Marginal	The picture is poor in quality and you wish you could improve it. Interference is somewhat objectionable.
5	Inferior	The picture is very poor, but you could watch it. Definitely objectionable interference is present.
6	Unusable	The picture is so bad that you could not watch it.

* Doc. [CCIR, 1970-74] discusses both a 6-level and a 5-level impairment scale, and concludes that a 5-level scale should be used in the future to determine protection ratios. The data used in this report was taken from tests that used the 6-level TASO scale.

Protection ratios for television signals are evaluated on the basis of Recommendation 600-1 using the method described in Recommendation 500 and taking account of the information contained in Report 405. CCIR Study Group 11 has decided that television protection ratios should be evaluated by subjective assessments using the CCIR's double-stimulus impairment scale method (see Table III).

TABLE III

CCIR double-stimulus impairment scale

Grade	Criterion of interference perceptibility
5	Imperceptible
4	Perceptible, but not annoying
3	Slightly annoying
2	Annoying
1	Very annoying

In the terrestrial television service, under the effect of short-term tropospheric interference, the maximum permissible interference level should correspond to grade 3 on the CCIR scale and to grade 4 in the case of interference for greater than 50% of the time. For the satellite television service, the permissible interference level should correspond to grades 4 and 5.*

The protection ratio data for television broadcasting systems using vestigial sideband amplitude modulation are contained in Recommendation 655 and Report 1214.

For mutual interference between vestigial sideband AM television systems and systems with different types of signals, used in the fixed and mobile services, the protection ratios are given in Report 449.

Test data on protection ratios for cases of mutual interference between terrestrial and satellite TV systems are contained in Report 634-3 .

* Note that the CCIR scale values are somewhat arbitrary and will vary as a function of stimulus material, viewing ratio, etc. for television reception. Other values may be used for valid reasons, when appropriate.

The protection ratio values for TV satellite systems obtained by various administrations in cases where the wanted and interfering signals are modulated by the TV signals or, for example, multi-channel sound signals, are given in Report 634-3 .

Report 449 deals with mutual interference between satellite FM TV and fixed and mobile service systems.

Table IV summarizes the wanted and interfering modulation cases for which the CCIR desires to obtain signal-to-interference protection ratios.

Table V summarizes the protection ratios that have been completed for three performance-threshold, two frequency-separation cases and lists the reference appropriate to each signal-to-interference entry. The protection ratios are applicable to stable conditions and are given in terms of the mean desired signal-to-mean interference power ratio at the receiver input for continuous (non-pulsed) transmissions. The ratio is given in terms of peak power for pulsed signals (i.e., mean desired signal-to-peak interference power for pulsed interference; peak desired signal-to-mean interference power for a pulsed desired signal). Table IV also gives protection ratios for cases where the interference is comparable to white gaussian noise. The input signal-to-noise ratios $(S/N)_i$ specified in Table V are of sufficient magnitude to ensure that the systems are not noise limited. The $(S/N)_i$ is therefore greater than that required to produce an AI of 1.0, a TASO score of 1, or a bit error ratio of 10^{-6}.

The peak envelope power for the J3E or R3E wanted or interference case is 6 dB greater than the mean power used in Table V . The carrier power for the A3E case (30% modulation) is approximately equal to the mean signal power used in Table V . The two frequency separations are given for a case in which carrier frequency separation is approximately zero or co-channel (symbolized by CO in the Table) and an approximately worst off-tuned case (symbolized by OC in the Table) whose value is given in the parameter column. The frequency separation for the single sideband J3E or R3E cases is the difference between the assigned frequencies.

4. Digital receiver model

A simplified model of a communications receiver is shown in Figure 1. The input to the channel interface is the superposition of the desired and undesired signals appearing at the receiver antenna output. The channel interface is composed of a number of circuit elements and is characterized by a receiver selectivity and by desired and undesired signal characteristics. Several reports (e.g., Reports 654 and 524) provide means to determine the nature of desired and undesired signals at the input to the demodulator given channel interface characteristics. The most important channel interface characteristics to consider are the bandwidth relationship between the undesired signal and the channel interface, off-tuning between the receiver and the undesired signal, and nonlinear effects.

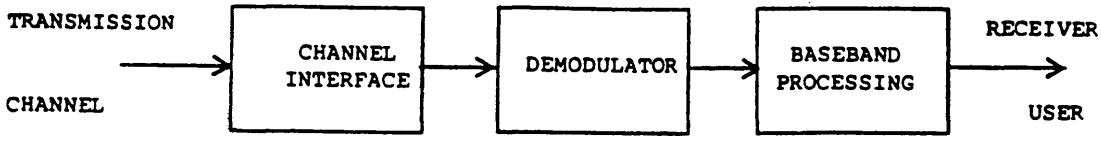


Figure 1 - General communication system receiver model.

TABLE IV - *Modulation cases to be studied*

Class of emissions (1)		Interfering signal (2)																												
Wanted signal		A1A aural	A1B printer	A2A aural	A2B 24 baud	A2B two tone	H2B	A3EJN	J3EJN or R3EJN	H3EJN	A3C	R3C	J3C	C3F	R7B	B7B	J7B	B8EJF	N0N	F1B	F3EJN	F3C	F3F	F7BDX	F8EJF	P0N	V9W	A3EGN	F3EGN	White Gaussian noise
A1A	telegraphy aural reception																													
A1B	telegraphy printer		×					×	×											×										
A2A	telegraphy aural reception																													
A2B	telegraphy 24 baud																													
A2B	two tone		×					×													×									×
H2B	telegraphy																													
A3EJN	voice (*AI* criterion)		×					×	×											×	×					×				×
	voice (*S/I* criterion)		×					×	×											×	×					×				×
J3EJN or R3EJN	SGB-SC voice (*AI* criterion)		×					×	×											×	×				×	×				×
	SGB-SC voice (*S/I* criterion)		×					×	×											×	×				×	×				×
H3EJN	SSB-FC voice (*AI* criterion)																													
	SBB-FC voice (*S/I* criterion)																													
A3C	facsimile																													
R3C	facsimile SSB reduced carrier																													
J3C	facsimile SSB suppr. carrier																													
C3F	vestigial SB picture only					×		×						×							×									
R7B	multich. vft SSB redc. car.																													
B7B	multich. vft ISB																													
J7B	multich. vft SGB supp. car.																													
B8EJF	voice mplex ISB (*AI* criterion)																													
	voice mplex IBS (*S/I* criterion)																													
N0N	(Decca)																													
F1B	telegraphy printer		×					×	×											×	×					×				×
F3EJN	voice (*AI* criterion)		×																	×	×				×					×
	voice (*S/I* criterion)		×																	×	×				×					×
F3C	facsimile																													
F3F	television picture only																													
F7BDX	4 freq duplex telegraphy																													
F8EJF	voice multiplex (*AI* criterion)		×					×												×	×				×	×				×
	voice multiplex (*S/I* criterion)		×					×												×	×				×	×				×
P0N																														
A3EGN	broadcasting																													
F3EGN	broadcasting																													

(1) If the class of emission is not defined in this Table, see Recommendation 326-1 (New Delhi 1970). In cases of automatic telegraphy and SSB telephony, cases using automatic repeat requested (ARQ) and Lincompex (Report 354) respectively should be included.

(2) × denotes cases for which values are given in Table IV.
× denotes cases for which partial results are given in Table IV.

TABLE V * - *Protection ratios (dB)***

Wanted signal \ Interference		Class of emission	500HA1B			6K00A2B			6K00A3E			3K00J3E			5M00C3F			7M00C3F-8M00C3F			1K10F1B			16K0F3E			726KF8E			1M32P0N			Noise		
		Parameters	100 baud PW = 10 ms			m_f = 1			m_f = 0.3						525 lines			625 lines			50 baud PW = 10 ms						24 channels			PW = 5 µs PRF = 300 pps			White Gaussian noise		
Class of emission	Parameters	Performance level (1)	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE	CO	OC	NOTE
500HA1B	BW_{IF} = 500 Hz 50 Baud $(S/N)_I$ = 18 dB	$P_E = 10^{-2}$	11		3				6		4	14		4							12		3	8		4									
		$P_E = 10^{-4}$	12		3				7		4										13		3	11		4									
		$P_E = 10^{-6}$	13		3				8		4										14		3	13		4									
6K00A2B	BW_{IF} = 8 kHz m_s = 1.0 $(S/N)_I$ = 18 dB	$P_E = 10^{-2}$	4		1				5		1													4		1							6	-	1
		$P_E = 10^{-4}$	4		1				5		1													4		1							9	-	1
		$P_E = 10^{-6}$	4		1				5		1													4		1								-	1
6K00A3E (2)	BW_{IF} = 8 kHz Δf = 0.5 kHz m_s = 0.3 $(S/N)_I$ = 45 dB	MINIT	44	61	1				43	48	1	50	50	1							47	55	1	48		1				20	10	1			
		0.7 *Al*	4	8	1				7	8	1	17	14	1							3	8	1	19		1				-17	-22	1	21	-	1
		0.3 *Al*	-7	-2	1				2	3	1	6	3	1							-2	4	1	8		1				-30	-37	1	10	-	1
		GCQ	39	35	2				32	42	2	44	43	2							37	41	2	40		2				-3	-2	2	41	-	2
		MCQ	21	20	2				14	24	2	26	25	2							19	23	2	22		2				-15	-20	2	23	-	2
		JUQ	12	11	2				5	15	2	17	16	2							10	15	2	13		2				-24	-28	2	14	-	2
3K00J3E or 3K00R3E	BW_{IF} = 2.7 kHz Δf = 0.5 kHz $(S/N)_I$ = 35 dB	MINIT	25	42	1				20	20	1	42	41	1							30	40	1	35		1	38		1	1		1		-	-
		0.7 *Al*	-14	-8	1				-14	-5	1	3	4	1							-25	-12	1	3		1	0		1	-38		1	9	-	1
		0.3 *Al*	-28	-24	1				-28	-19	1	-12	-16	1							-43	-37	1	-10		1	-12		1	-52		1	-3	-	1
		GCQ	10	27	2				13	30	2	31	32	2							21	30	2	27		2	26		2	-15		2	32	-	2
		MCQ	-8	9	2				-5	12	2	13	14	2							3	12	2	9		2	8		2	-33		2	14	-	2
		JUQ	-17	0	2				-14	3	2	4	5	2							-6	3	2	0		2	-1		2	-42		2	5	-	2
5M00C3F	BW_{IF} = 6 MHz 525 lines $(S/N)_I$ = 46 dB	TASO 1																																	
		TASO 2																																	
		TASO 2.5				50	15	5	50	15	5				47	25	5							50	15	5									
7M00C3F-	BWIF = 6 MHz 625 lines	CCIR 4							58	-	6							52	-	6															
8M00C3F	(S/N)I = 46 dB	CCIR 3							51	-	6							45	-	6															
1K10F1B	BW_{IF} = 1050 Hz D_{pk} = ± 425 Hz 50 Baud $(S/N)_I$ = 18 dB	$P_E = 10^{-2}$	0		1&3				2		1	10		4							6		3	0.5		1				-50		1	9		1
		$P_E = 10^{-4}$	0		1&3				3		1	13		4							7		3	1		1				-49		1	13		1
		$P_E = 10^{-6}$	1		1&3				3		1	15		4							8		3	2		1				-48		1	15		1
16K0F3E (3)	BW_{IF} = 16 kHz D_{pk} = 5 kHz Δf = 0.5 kHz De-emphasis $(S/N)_I$ = 22 dB	MINIT	38	38	1																33	33	1	31	31	1	32	32	1	-11		1	-	-	-
		0.7 *Al*	0	0	1																2	2	1	2	2	1	4	4	1	-24		1	1	-	1
		0.3 *Al*	0	0	1																0	0	1	-5	-5	1	0	0	1				0	-	1
		GCQ	13	13	2																15	15	2	14	14	2	16	16	2				11	-	2
		MCQ	2	2	2																3	3	2	1	1	2	4	4	2				5	-	2
		JUQ	-1	-1	2																1	1	2	0	0	2	1	1	2				2	-	2
4K00F8E (3)	24 channels Upper channel Δf = 44.5 kHz $(S/N)_I$ = 45 dB	MINIT	47	60	1				55	64	1										55	60	1	55	60	1	46	57	1	25	20	1	-	-	-
		0.7 *Al*	3	12	1				4	14	1										6	14	1	12	18	1	2	5	1		-34	1	9	-	1
		0.3 *Al*	0	-15	1				0	4	1										2	6	1	2	6	1	1	-3	1		-39	1	1	-	1
		GCQ	24		2				25		2										29		2				29		2				31	-	2
		MCQ	6		2				7		2										11		2				9		2				13	-	2
		JUQ	2		2				2		2										5		2				4		2				4	-	2

*For aeronautical mobile (R) service above 30 MHz, see Recommendation 441.

**The values given in this Table for the noise interference case for A3E and J3E are 2 dB higher than the corresponding values given in Recommendation 339.
This is due to the difference in the modulation specifications.

(1) GCQ = Good commercial quality MCQ = Marginally commercial quality
JUQ = Just useable quality CO = Frequency separation approximately zero
OC = Off-channel separation given by Δf Δf = Frequency separation between the centre of the wanted and unwanted signal spectra except for single sideband J3E or R3E where the frequency separation is the difference between the assigned carrier frequencies.

(2) - For UHF and VHF land and maritime mobile services, see Report 358.
- For A3E broadcasting, band 5 (LF), band 6 (MF), and band 7 (HF), see Recommendations 448-1 and 449-2, and also Report 298-3, (Geneva, 1974).
- For F3E (monophonic) broadcasting, band 8 (VHF), and band 9 (UHF), see Recommendation 412.
- For F3E (stereophonic) broadcasting, band 8 (VHF), see Report 462 (New Delhi, 1970).
- For A3E broadcasting, protection ratios to avoid interference in the band shared with broadcasting in the Tropical Zone, see Recommendation 216.

(3) The values given to protection ratio objectives identified in this category are limited to single-link considerations and are not intended to apply to multi-link terrestrial microwave radio-relay systems such as covered in Recommendations 390, 393 and 395.

NOTES

1. OT/ECAC [August 1975] Communications/Electronics Receiver Performance Degradation Handbook. The Frequency Management Support Division, Office of Telecommunications, US Department of Commerce and the Electromagnetic Compatibility Analysis Center (ECAC), ESD-TR-75-013, NTIS. (Available from US Department of Commerce, National Technical Information Service, Springfield, Va. USA, Order No. AD-A016400).
2. Obtained from transfer curves used in the Handbook referred to in Note 1.
3. Extrapolated from CCIR Recommendation 240.
4. MAYHER, R. [1972] Interference performance degradation to digital systems. Record of the 1972 IEEE International Electromagnetic Compatibility Symposium.
5. Extrapolated from CCIR Recommendation 418.
6. Evaluated in accordance with Recommendations 500, 600-1 and Report 405.

Undesired signals are characterized as follows:

1. Undistorted: the ideal waveform transmitted by the interfering transmitter. The signal may be specified in the frequency domain in terms of power spectral density.

2. Noiselike: the signal varies in amplitude according to a normal (Gaussian) distribution. The signal may have a flat spectrum and is referred to as additive white Gaussian noise (AWGN).

3. Continuous Wave (CW): a constant frequency sinusoid whose phase with respect to the receiver is assumed to be a uniformly distributed random variable.

4. Impulsive: a sequence of periodic or randomly spaced pulses, each of which is of short duration compared to the time between pulses.

Undesired signals may be either continuous or intermittent. An intermittent undesired signal may be defined as a signal whose statistics such as amplitude distribution function, mean, and variance are time-varying when observed at a victim receiver. Interference due to a co-located frequency hopper is an example of an intermittent undesired signal in the sense that the victim receiver will typically exhibit time-varying performance degradation. The recommended analysis procedure for the case of intermittent undesired signals involves partitioning the observation interval into contiguous time segments, or epochs, during each of which the undesired signal statistics are (approximately) constant. A separate degradation analysis is performed for each epoch, and the results are time-averaged. It is important that the time-averaging not be performed on signals until they have been demodulated.

For electromagnetic compatibility (EMC) analyses using the performance curves in this report, the undesired signal at the receiver input can usually be assumed to be either undistorted (i.e., the output of an interfering transmitter with known waveform characteristics) or noiselike. The channel interface characteristics are then used to determine the undesired signal at the demodulator input. The curves show the demodulator output bit-error rate as a function of the ratio of the desired symbol energy-to-noise power spectral density (E/N_o) or the ratio of the desired symbol energy-to-interference energy (E/I_e) at the demodulator input. The noise is assumed to be Gaussian, and the interference is assumed to be continuous-wave. The analyst must determine whether the undesired signal at the input to the demodulator more closely resembles noise or CW interference or some combination of the two. This determination may include a prediction regarding the nature of the interfering signal spectrum at the demodulator input based on the passband of the channel interface and the interfering signal RF characteristics. Material [Kennison et al., 1978] is available to assist the analyst.

The remainder of the report addresses the individual sections of the receiver model shown in Figure 1 following the channel interface. The output of any particular section may be found by concatenating the effects of that section and any preceding sections.

5. Performance of Digital Demodulators

Typical M-ary digital demodulator performance is given in terms of P_s versus E/N_o and E/I. These terms are defined as follows:

M: the number of possible distinct symbols. For binary signaling, M = 2.

P_s: the symbol-error probability. The bit-error probability P_b, which is also often used, cannot exceed P_s. When M = 2, $P_b = P_s$.

E/N_o: the ratio of average symbol energy in joules-to-noise power spectral density in watts/Hz, as specified at the demodulator input,

E/I_e: the ratio of average signal energy per symbol (or per bit)-to-interference energy per symbol (or per bit), as specified at the demodulator input.

Power ratios, in particular the signal-to-noise ratio (S/N), may be used instead of energy ratios by noting that

$$E/N_o = (S/N)\ (BT) \qquad (1)$$

where

B = the receiver bandwidth, in Hz
T = the symbol duration, in seconds
S/N = measured at the demodulator input.

TABLE VI summarizes the types of modulation presented, which curve to use for each, and the undesired signal (CW or noise) for which probabilities of bit or symbol errors may be obtained. The curves identified in TABLE VI are provided in Figures 2 through 24. These curves relate receiver performance in terms of symbol- or bit-error rate in the presence-of noise and/or interference. The noise is assumed to be Gaussian, and the interference is assumed to be CW. The curves have been developed assuming optimum receiver design, i.e., bandwidths associated with the demodulator were designed for the associated system bit durations and data rates. The table also includes a listing of primary sources that were consulted to provide either curves or the methods to develop the curves.

When using the curves, be alert to the parameters used. Bit energy-to-noise ratio (E_b/N_o) rather than symbol energy-to-noise (E/N_o) was used in most of the figures comparing M-ary schemes for different values of M in order to simplify the graphs. The value E represents the average symbol energy. The relationship between symbol energy and the equivalent bit energy is

$$E = E_b \log_2 M \qquad (2)$$

Some of the figures presented have the probability of a bit error, P_b, rather than the probability of a symbol error, P_S. The relationship between the two, for orthogonal signaling (CFSK, NCFSK, MSK), is,

$$P_b = \frac{2^{k-1}}{2^k - 1} P_S \qquad (3)$$

where $k = \log_2 M$ (number is equivalent bits).

For M-ary CPSK and DPSK (Gray encoding is assumed) and for CASK and NCASK the relationship is:

$$P_b = P_s/k \qquad (4)$$

TABLE VI - SUMMARY OF DIGITAL MODULATION TYPES CONSIDERED

Modulation Type	Plot of	Interfering[1] Signal	Figure Number	Source
CPSK, M-ary (Coherent Phase Shift Keying)	P_S VS E_b/N_o	N	2	[Proakis, 1983]
CPSK, M = 2	P_S VS E/N_o, E/I_e	I_e	3	[Juroshek, 1976]
CPSK, M = 4	P_S VS E/N_o, E/I_e	I_e	4	[Juroshek, 1976]
CPSK, M = 8	P_S VS E/N_o, E/I_e	I_e	5	[Juroshek, 1976]
CPSK, M = 16	P_S VS E/N_o, E/I_e	I_e	6	[Juroshek, 1976]
QPSK (Quaternary PSK)	P_S VS E/N_o, E/I_e	I_e	4	[Juroshek, 1976]
O-QPSK (Offset-QPSK)	P_S VS E/N_o, E/I_e	I_e	4	[Juroshek, 1976]
DPSK M = 2 (Differential Phase Shift Keying)	P_S VS E/N_o, E/I_e	N, I_e	7	[Rosenbaum, 1969]
DPSK M = 4	P_S VS E/N_o, E/I_e	N, I_e	8	[Rosenbaum, 1969]
DPSK M = 8, 16	P_S VS E/N_o, E/I_e	I_e	9	[Rosenbaum, 1969]
CFSK M-ary (Coherent Frequency Shift Keying)	P_b VS E_b/N_o	N	10	[Proakis, 1983]
CFSK M = 2	P_S VS E/N_o, E/I_e	I_e	11	[Proakis, 1983]
MSK (Minimum Shift Keying)	P_S VS E/N_o, E/I_e	I_e	4	[Juroshek, 1976]
NCFSK M-ary (Noncoherent Frequency Shift Keying)	P_S VS E_b/N_o	N	12	[Proakis, 1983]

TABLE VI (continued)

Modulation Type	Plot of	Interfering[1] Signal	Figure Number	Source
NCFSK M = 2	P_S VS E/N_o, E/I_e (Interference in one channel)	I_e	13	[Wang, 1974]
NCFSK M = 2	P_S VS E/N_o, E/I_e (Equal interfering tones in each channel)	I_e	14	[Wang, 1974]
NCFSK M = 4	P_S VS E/N_o, E/I_e	I_e	15	[Massaro, 1975]
NCFSK M = 8	P_S VS E/N_o, E/I_e	I_e	16	[Massaro, 1975]
CASK M-ary, bipolar (Coherent Amplitude Shift Keying)	P_S VS E_b/N_o	N	17	[Proakis, 1983]
CASK M = 16, biplolar	P_S VS E/N_o, E/I_e	I_e	18	[Juroshek, 1976]
CASK M-ary, unipolar	P_S VS E_b/N_o	N	19	[Arthurs and Dym, 1962]
CASK M = 2, unipolar (Also called OOK or on off keying)	P_S VS E/N_o, E/I_e	I_e	20	[Juroshek, 1976]
NCASK, M-ary (Noncoherent Amplitude Shift keying)	P_S VS E_b/N_o	N	21	[Arthurs and Dym, 1962]
QAM, M-ary (Quadrature Amplitude Modulation)	P_S VS E_b/N_o	N	22	[Proakis, 1983]
QAM, M = 4	P_S VS E/N_o, E/I_e	I_e	4	[Juroshek, 1976]
QAM, M = 16	P_S VS E/N_o, E/I_e	I_e	23	[Juroshek, 1976]
QAM, M = 64	P_S VS E/N_o, E/I_e	I_e	24	[Juroshek, 1976]

[1] N = AWGN, I_e = CW interference.

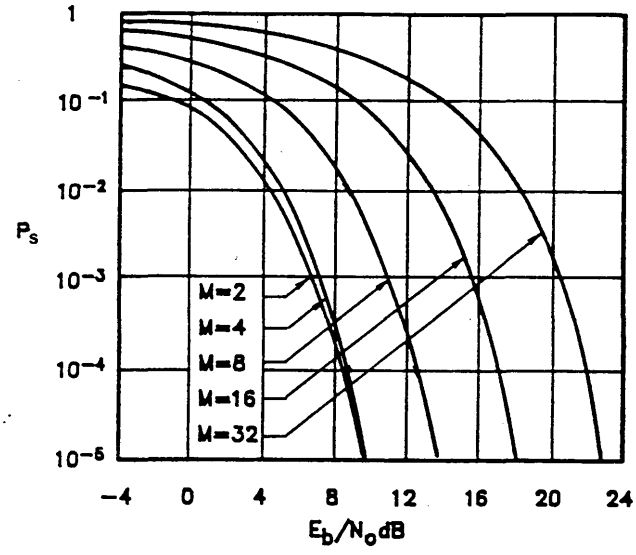


Figure 2 - P_S vs. E_b/N_o for M-ary CPSK.

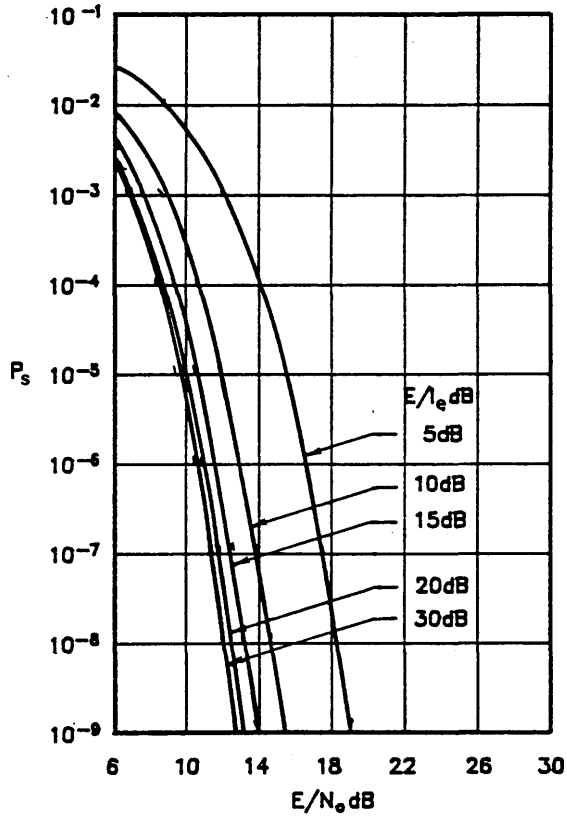


Figure 3 - P_S vs. E/N_o and E/I_e for binary CPSK.

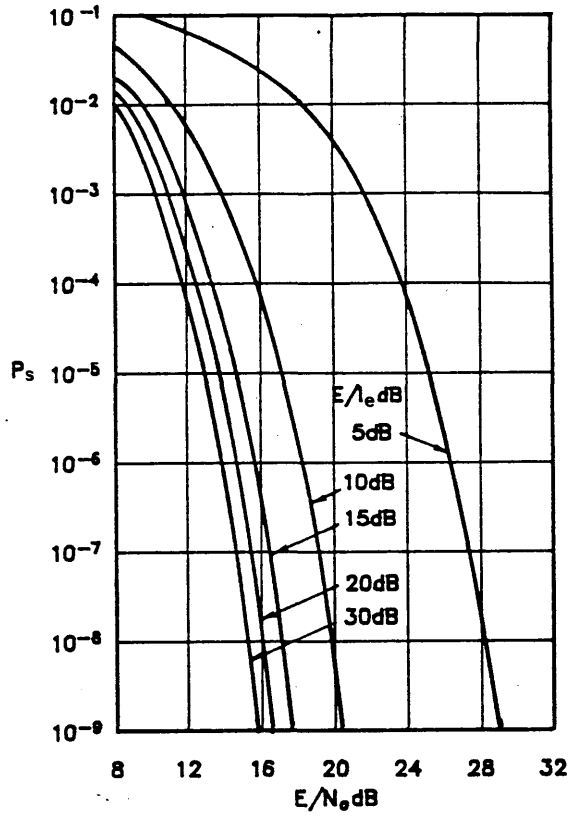


Figure 4 – P_S vs. E/N_o and E/I_e for 4-ary CPSK, QPSK, O-QPSK, MSK and 4-ary QAM.

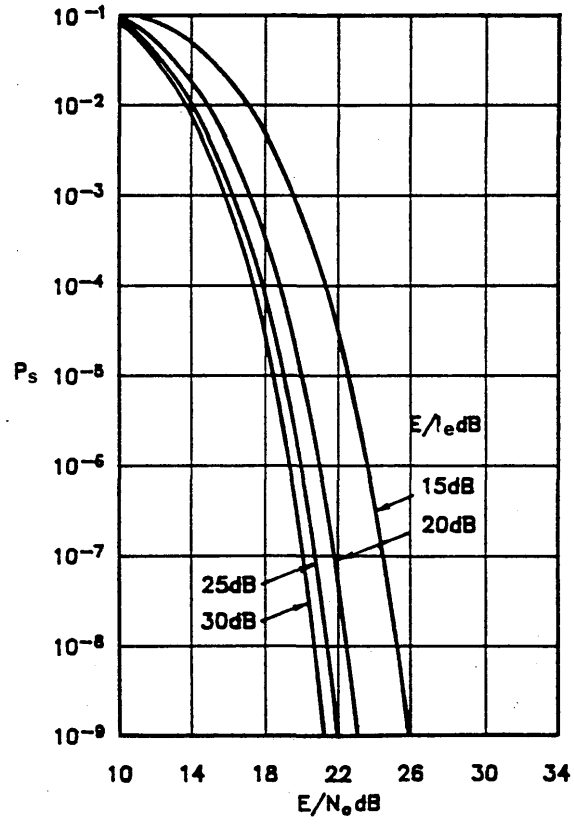


Figure 5 – P_S vs. E/N_o and E/I_e for 8-ary CPSK.

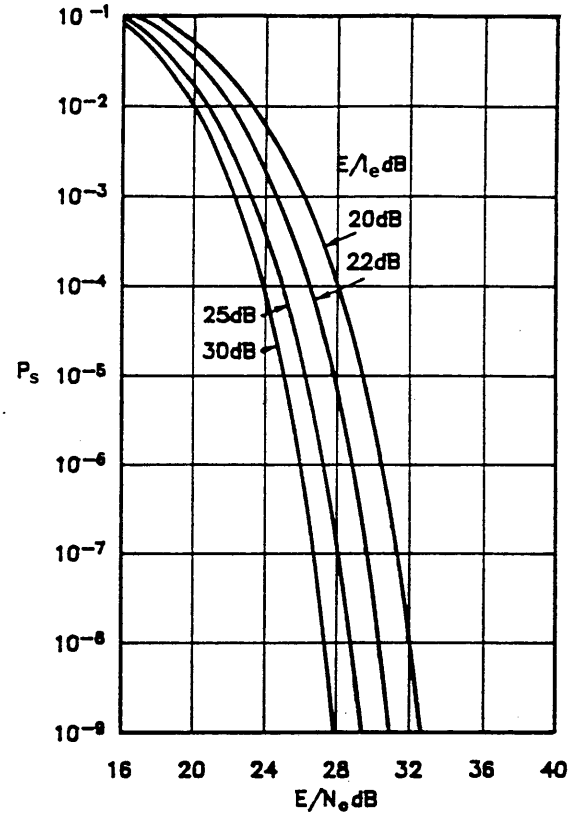


Figure 6 - P_S vs. E/N_o and E/I_e for 16-ary CPSK

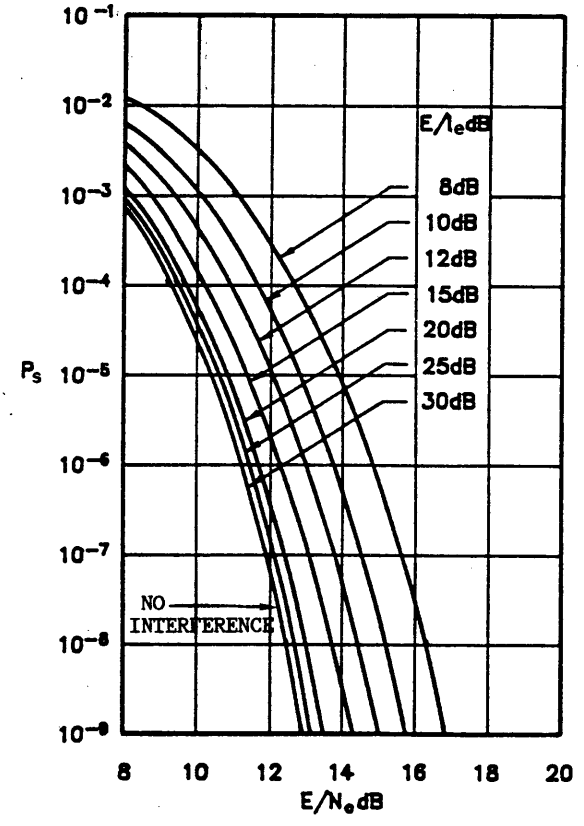


Figure 7 - P_S vs. E/N_o and E/I_e for binary DPSK

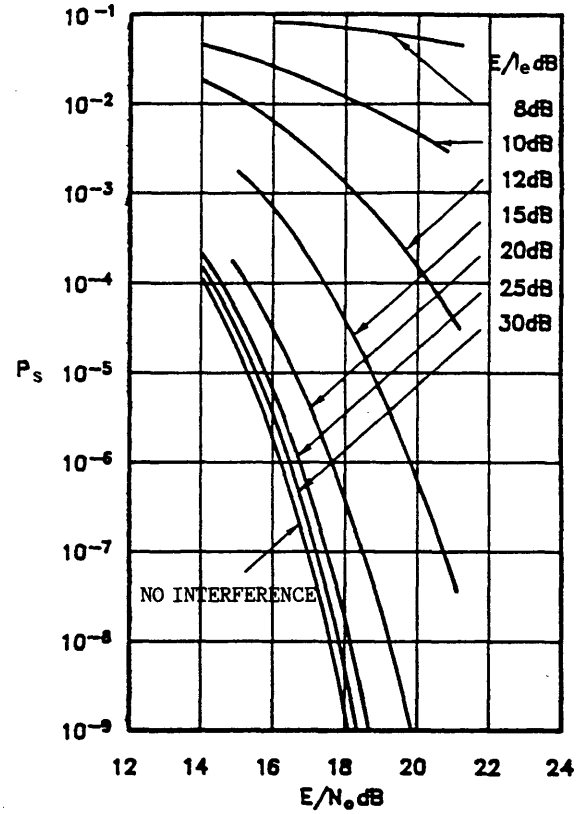


Figure 8.- P_S vs. E/N_o and E/I_e for 4-ary DPSK

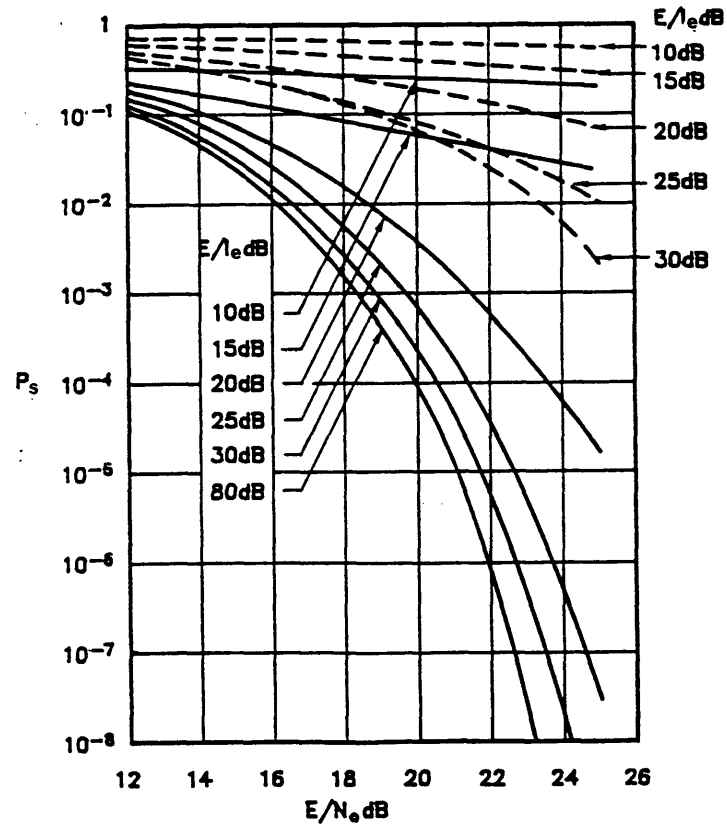


Figure 9 - P_S vs. E/N_o and E/I_e for 8-ary and 16-ary DPSK

– – – M=8

——— M=16

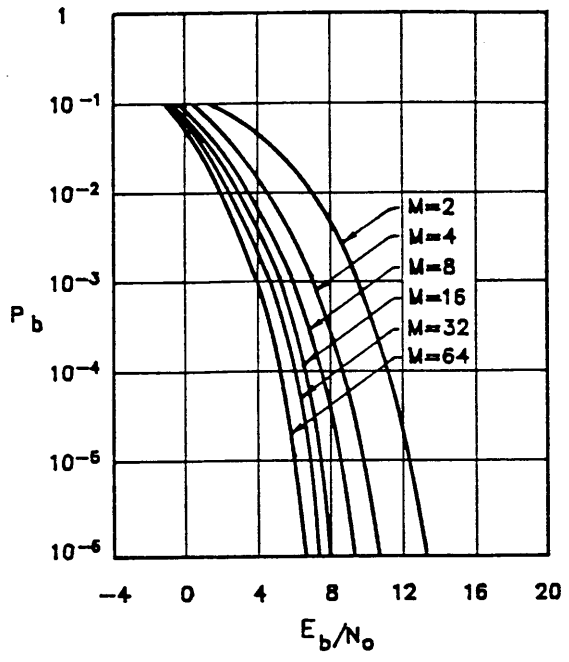


Figure 10 - P_b vs. E_b/N_o for M-ary CFSK.

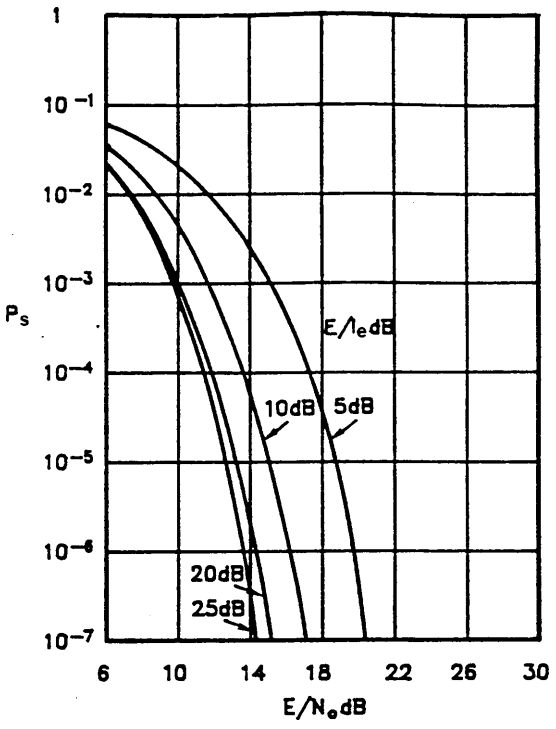


Figure 11 - P_S vs. E/N_o and E/I_e for binary CFSK.

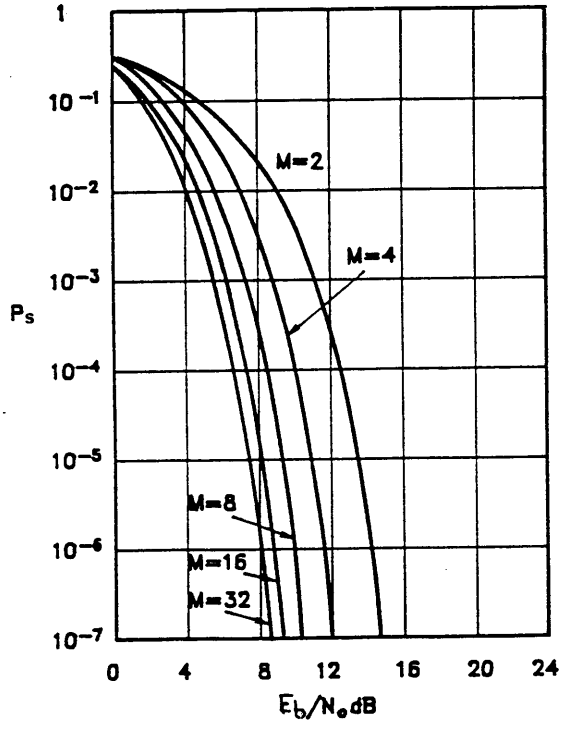


Figure 12 - P_S vs. E_b/N_o for M-ary NCFSK.

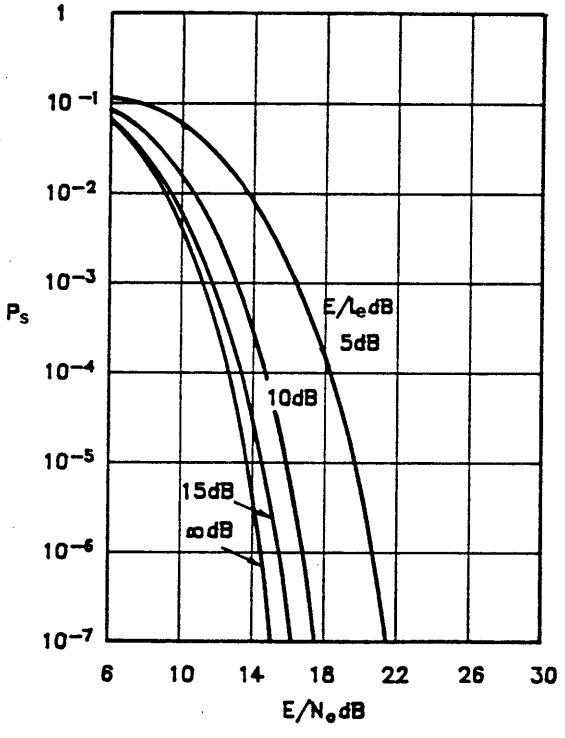


Figure 13 - P_S vs. E/N_o and E/I_e for binary NCFSK with interin one channel.

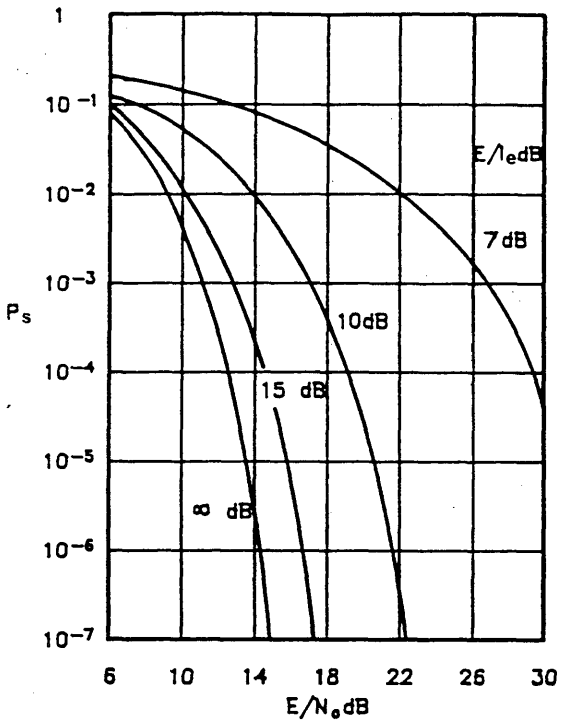


Figure 14 - P_S vs. E/N_o and E/I_e for binary NCFSK with equal interfering tones in each channel.

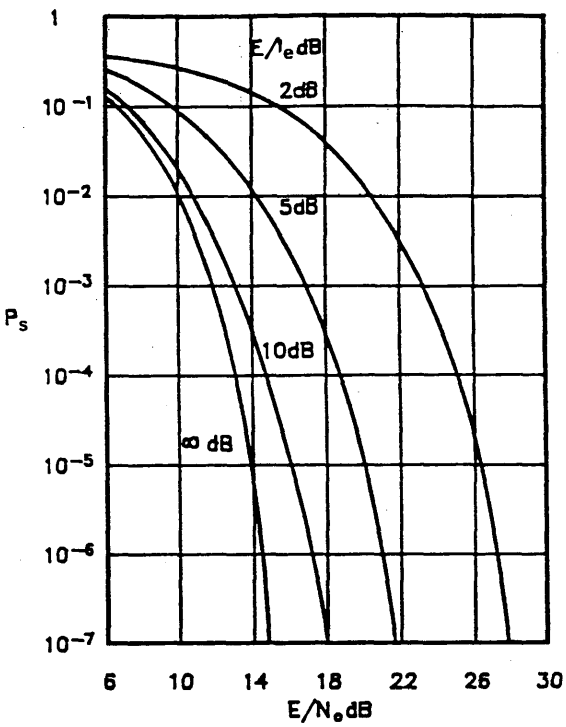


Figure 15 - P_S vs. E/N_o and E/I_e for 4-ary NCFSK.

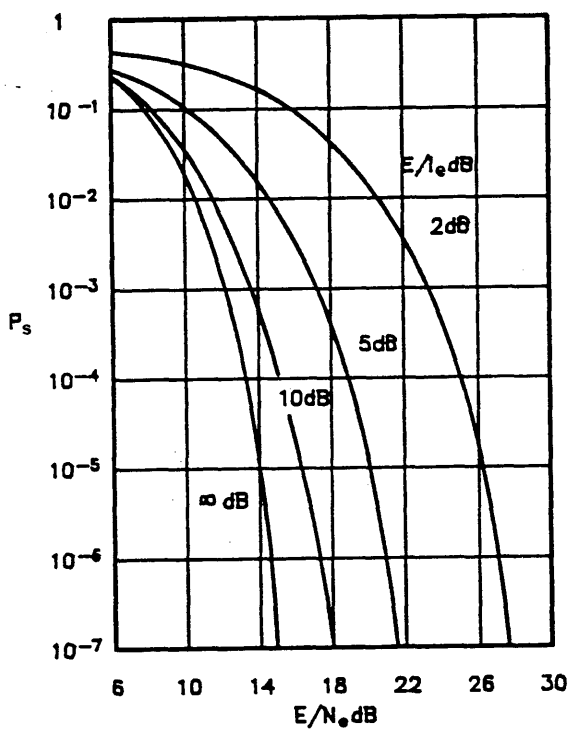


Figure 16 - P_S vs. E/N_o and E/I_e for 8-ary NCFSK.

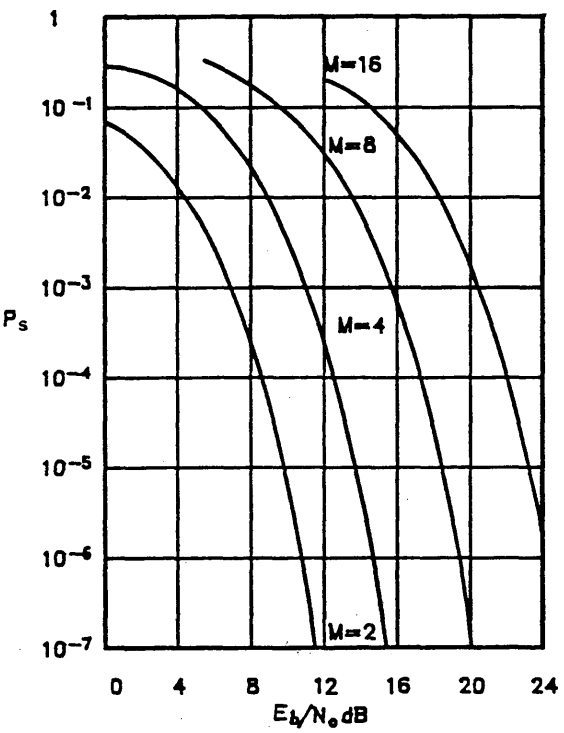


Figure 17 - P_S vs. E_b/N_o for M-ary bipolar CASK.

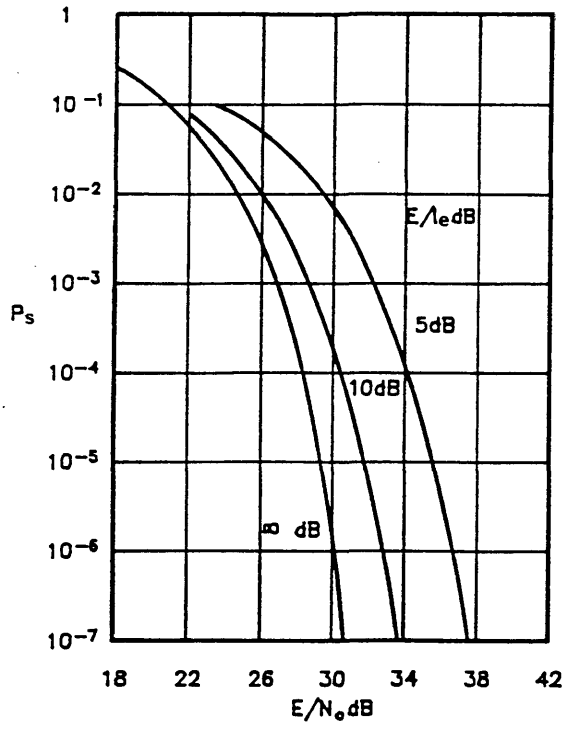


Figure 18 – P_S vs. E/N_o and E/I_e for 16-ary bipolar CASK.

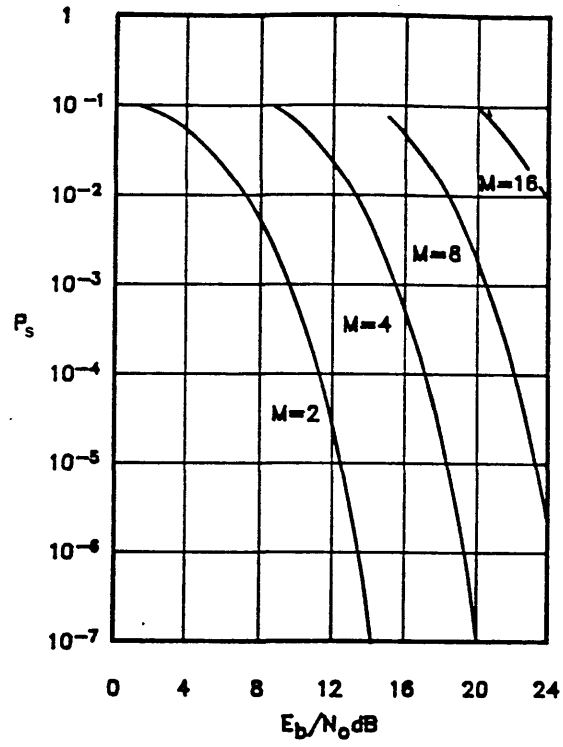


Figure 19 – P_S vs. E_b/N_o for M-ary unipolar CASK.

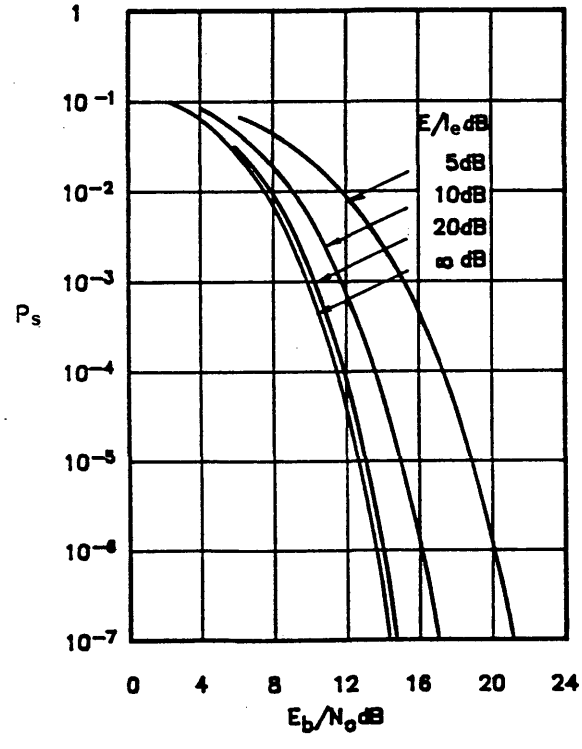


Figure 20 – P_S vs. E_b/N_o and E/I_e for binary unipolar CASK (OOK).

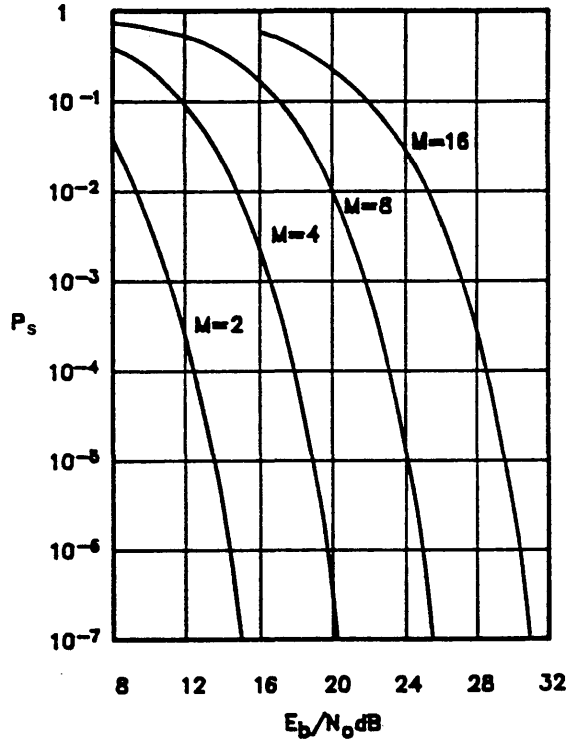


Figure 21 – P_S vs. E_b/N_o for M-ary NCASK.

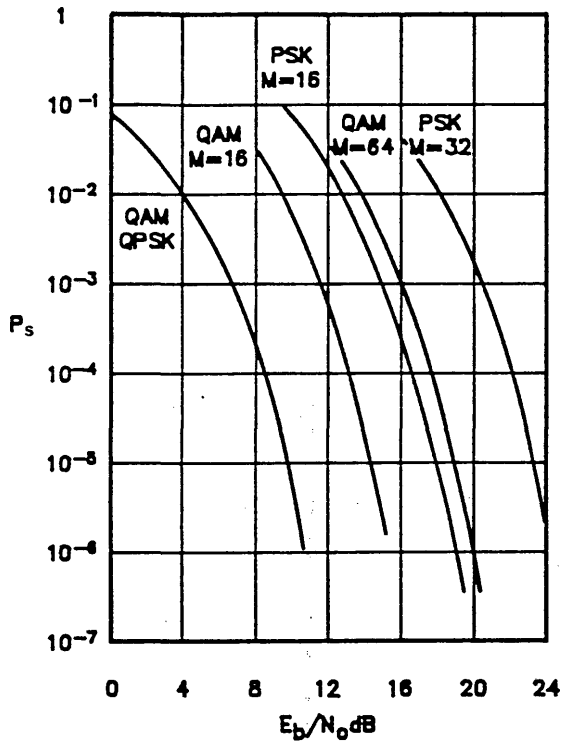


Figure 22 - P_S vs. E_b/N_o for M-ary QAM and comparison to selected M-ary PSK.

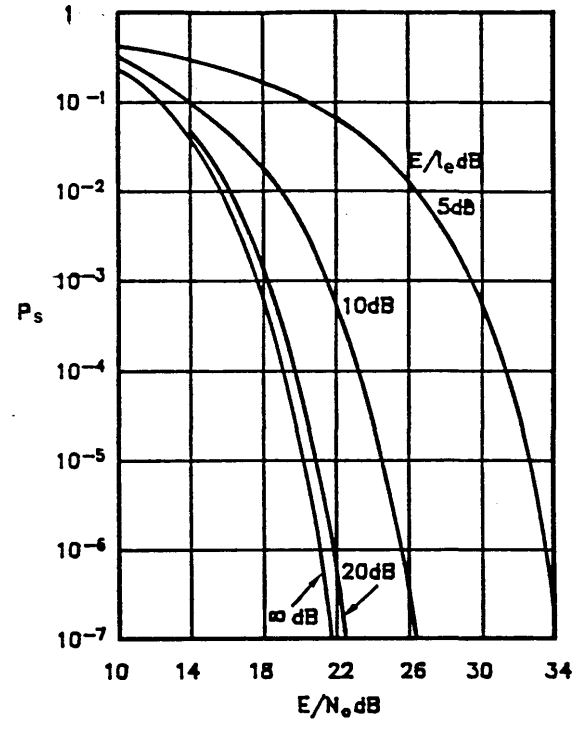


Figure 23 - P_S vs. E/N_o and E/I_e for 16-ary QAM.

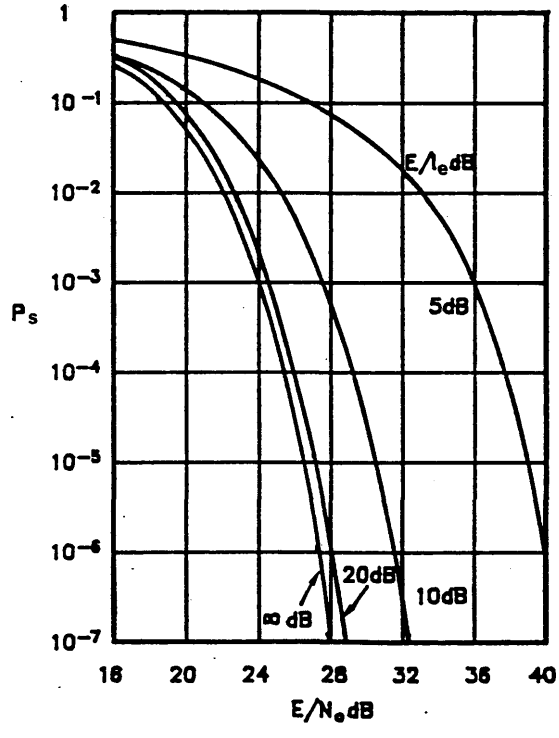


Figure 24 - P_S vs. E/N_o and E/I_e for 64-ary QAM.

6. Performance Degradation of Teleprinters

A teleprinter is a communications terminal device that converts an input bit sequence into an output sequence of printed characters. The bit-to-character transformation is essentially a decoding operation; several consecutive bits are taken to represent a single printable character or control operation. The coding scheme is provided by a so-called International Telegraph Alphabet (ITA). Five ITAs are currently defined.

In order to correctly decode r-bit characters from a serial bit stream the teleprinter must have some means of identifying the beginning and end of each character. This process is known as framing a character. There are two general techniques for achieving proper framing, referred to as asynchronous and synchronous. In an asynchronous system, the transmitter appends a start bit and a stop bit to the data bits that comprise the coded character. The communications system maintains an idle state until a character is transmitted. The detection of a start bit by the teleprinter signifies the suspension of the idle state and results in the next r bits being decoded and printed. A stop bit follows, signifying the end of the decoding process for that particular character and permitting the printing mechanism to recover in anticipation of another character. Following the stop bit will be either a return to idle or the start bit for another character.

A synchronous system does not rely on individual start/stop bits for character framing. Special synchronization characters are appended to the beginning of a message. The detection of these characters by the teleprinter results in character sync being obtained; the teleprinter can then correctly frame and print the characters that comprise the transmitted message.

Bit errors in the teleprinter input sequence result in character errors in the output sequence. The relationship between the input bit-error rate (BER) and the output character-error rate (CER) is the teleprinter degradation transfer function. The input BER is dependent on parameters of previous stages of the receiver such as modulation type and channel coding. It is convenient to consider the synchronous and the asynchronous cases separately.

6.1 Synchronous Systems

Assume that the teleprinter has obtained and is maintaining character sync. Also assume that the input BER is not time-varying and that the occurrences of bit errors are statistically independent events. This latter condition is sometimes denoted the "random error" case to distinguish it from the "burst error" case in which some clustering of bit errors is to be expected. For the correct decoding of an r-bit character, it is necessary that every one of the r bits be correct. Thus, the CER is determined by

$$CER = 1 - (1 - BER)^r \quad (5)$$

For small values of BER, Equation 5 can be approximated by the bound

$$CER \leq r\ BER \quad (6)$$

These assumptions can now be relaxed somewhat. Suppose bit errors are not independent; in particular, suppose that the errors exhibit some tendency to cluster (the "burst error" case). Then Equation 5 is not entirely accurate, however, it can be taken as an upper bound on error performance. That is

$$CER \leq 1 - (1 - BER)^r \quad (7)$$

Consequently, Equation 6 also holds. To justify this result, consider the burst-error case where several errors occur in adjacent bit positions in an n-bit sequence. Since one bit error is sufficient to cause a character error any further bit errors in that character are irrelevant. As a result, maximum degradation to the CER occurs when the bit errors are "spread out", i.e., distributed randomly.

If the input BER is time-varying, the CER will also be time-varying. In this case, it may be appropriate to time-average the CER as follows. Partition the observation time into N contiguous epochs during which the BER is approximately constant. For each epoch, determine the CER:

$$CER_i = 1 - (1 - BER_i)^r \quad (8)$$

The average CER may be computed:

$$\overline{CER} = \frac{1}{N} \sum_{i=1}^{N} d_i\ CER_i \quad (9)$$

where d_i is the fraction of the time that CER_i prevails.

Note the restriction that BER_i be approximately constant during the i-th epoch. The validity of Equation 9 improves as the length of the epochs increase. Epochs of at least several characters in length are required for Equation 9 to produce meaningful results.

The case of modelling burst errors has been studied by Vogler [1986], who developed a model that accounts for the first order effects of error clustering.

The assumption that the system is operating in character sync should also be considered. If the sync condition is not met, the system will not work at all; transmission will simply be suspended until sync is established.

6.2 Asynchronous Systems

A distinction must be made between real-time and buffered systems. In a real-time system, characters are transmitted as they are generated. Due to typing speed limitations, the system invariably returns to the idle state between transmitted characters. Consequently, character errors are independent, and Equations 5 through 9 hold with r replaced by r+2. The additional two bits are the start and stop bits.

For buffered systems, the character errors are not independent. An error in a start or stop bit in a particular character will result in the decoder searching for the appropriate control bit. It has been determined experimentally [Watt, 1958] that an average of five characters will be lost as the teleprinter attempts to regain character sync. It is necessary, therefore, that not only the bits in the character presently being decoded be correct, but that the start and stop bits in the previous five characters also be correct.

The Equations 5 through 9 hold (subject to the restrictions mentioned) with r replaced by r+12. The exponent 12 is due to the five start and five stop bits in the five characters prior to the one presently being decoded, as well as the start and stop bits in the present character. It is seen that a synchronous system offers somewhat better error tolerance than a buffered asynchronous system.

7. Performance Degradation of Digital-to-Analogue Converters

A digital-to-analogue converter (DAC) is a device that converts an input bit sequence into an analogue output. A DAC, along with its associated analogue-to-digital converter (ADC), is used in a hybrid communication system, i.e., a system in which digital processing and transmission techniques are used to communicate analogue information. There are various techniques available for performing the analogue-to-digital and digital-to-analogue conversions; the particular technique employed has an effect on the fidelity of the reconstructed analogue signal as well as the sensitivity of the system to channel-induced bit errors.

All hybrid communication systems introduce distortion due to the quantization process. A fundamental assumption in a performance degradation analysis of hybrid communications system is that the noise due to quantization and the noise due to channel-induced bit errors are basically additive. It is appropriate, therefore, to specify a DAC degradation transfer function that

gives output S/N as a function of input BER. Another measure of the output fidelity of a DAC for voice systems is articulation score (AS). Performance curves are provided for two types of DAC that are used in many practical systems: pulse-code modulation (PCM) and continuously-variable-slope-delta (CVSD) modulation.

7.1 Pulse-Code Modulation

Pulse-code modulation (PCM) is a straightforward conversion algorithm. The analogue waveform is sampled at regularly spaced intervals established by the sampling frequency. Each sample is then quantized and coded into an r-bit digital word according to its amplitude. The quantization is necessary because only 2^r digital words are available to represent all the possible amplitudes. The analogue waveform is thus represented by a sequence of r-bit digital words. The resulting bit stream is transmitted; channel impairments typically result in a number of errors in the received demodulated bit stream. The DAC decodes the digital words and produces a sample sequence. Signal conditioning operations such as interpolation and low-pass filtering are then performed on the samples, resulting in a reconstructed analogue waveform that should be nearly identical to the original information signal.

The output S/N and AS curves for PCM are shown in Figures 25 and 26, respectively. The S/N versus BER relationship is shown for r-bit PCM with r = 4, 5, 6, 7, 8. The sampling rate in each case is 8 kHz. The AS versus BER relationship is shown for PCM with r = 6,7,8. It is seen that AS does not vary with r for r = 6, 7,8 for the BER values in the figure. This is probably due to the fact that AS measures intelligibility not quality.

7.2 Continuously-Variable-Slope-Delta Modulation

CVSD is a variation of adaptive delta modulation that is currently popular due to its simplicity and resistance to channel errors. CVSD systems are capable of adequate performance at bit-error rates as high as 10^{-1}. For this reason CVSD systems are appropriate for use in applications where systems must function properly under adverse channel conditions. The CVSD algorithm is described more thoroughly in [Jayant and Noll, 1984]. S/N and AS curves for 16 k-bit CVSD are provided in Figures 27 and 28, respectively.

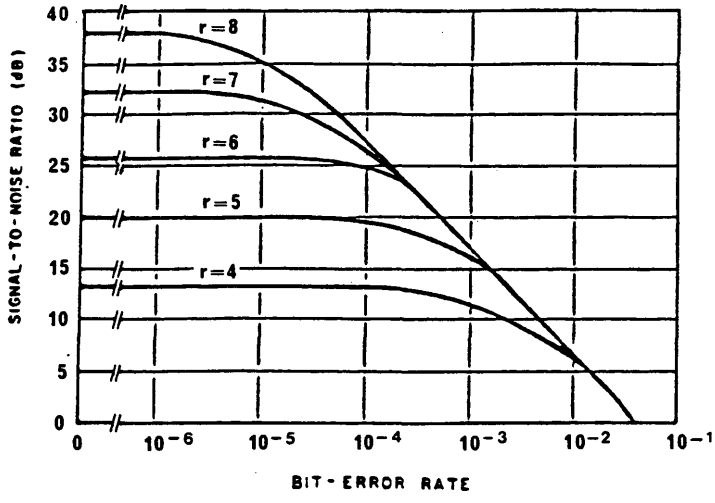


Figure 25 - Performance degradation of PCM - S/N vs. BER.

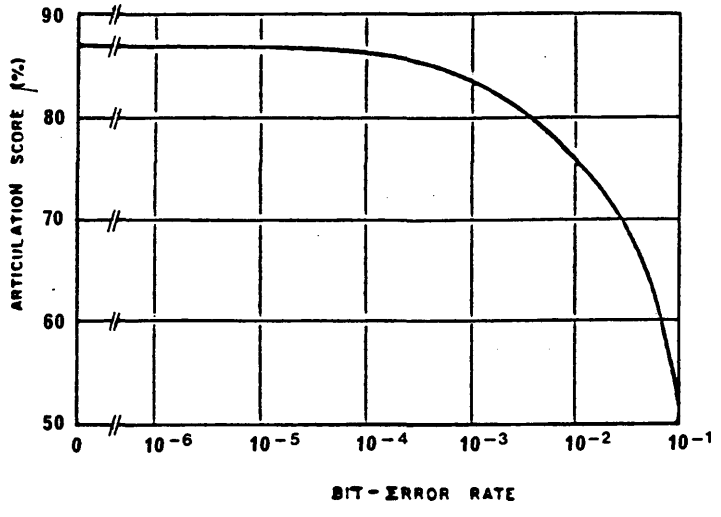


Figure 26 - Performance degradation of PCM - AS vs. BER (for 6, 7, and 8 bit PCM)

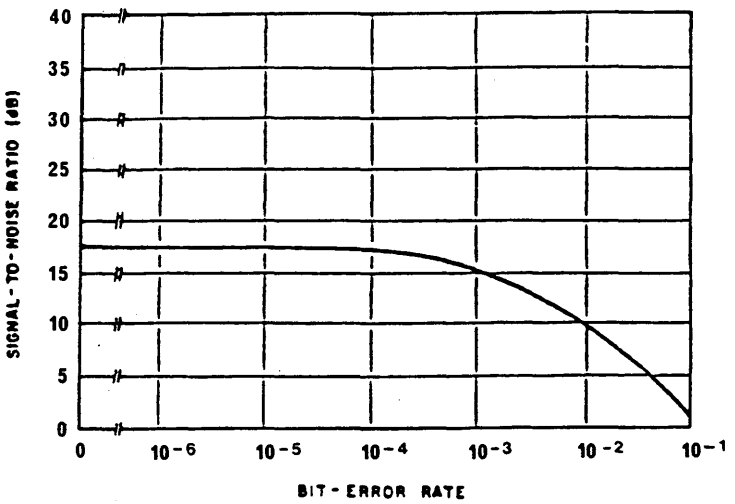


Figure 27 - Performance degradation of16 k-bit CVSD - S/N vs. BER.

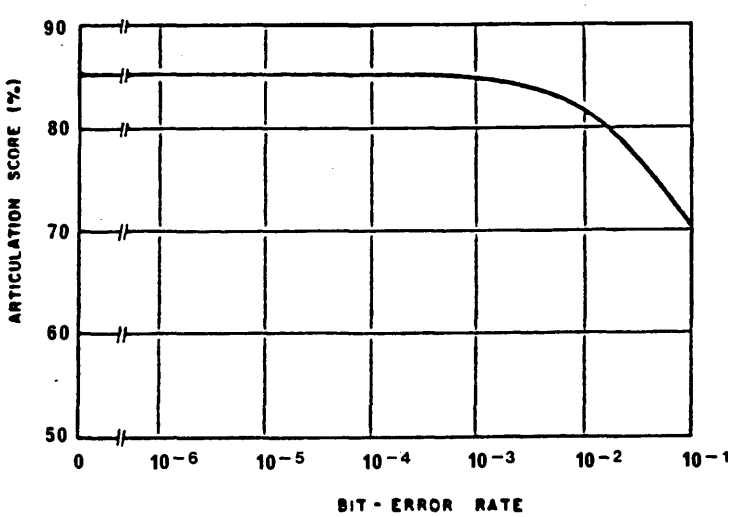


Figure 28. - Performance degradation of 16 kbit CVSD - AS vs. BER.

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[1970-74]: 10/383 (Joint Working Group 10/11B).

GLOSSARY

AI: articulation index
Baud: one bit per second in a train of binary signals
BW_{IF}: 3 dB intermediate frequency bandwidth (Hz)
D_{pk}: maximum frequency deviation
m_I: interference signal modulation index
m_s: desired signal modulation index
MINIT: minimum interference threshold
N: white gaussian noise
P_E: probability of bit error
PW: pulse width
$(S/N)_I$: the mean signal-to-noise power ratio in dB at the receiver input.

REPORT 659*

RF-PROTECTION RATIOS BETWEEN THE FM BROADCASTING SERVICE AND THE LAND MOBILE SERVICE

(Question 45/1)

(1978)

1. Introduction

Different radio services in adjacent frequency bands may affect each other by their overlapping frequency spectra. RF-protection ratios for the FM sound broadcasting service in band 8 (VHF) are given in Recommendation 412. These protection ratios are valid only within the broadcasting service itself. At the edges of the bands allocated to the broadcasting service mutual interference between the broadcasting service and other services may also occur. These services may operate in the double-sideband AM or FM mode. This Report gives results of measurements carried out under the conditions stated [Netzband and Petke, 1977].

The measurements were restricted to the case when the wanted transmitter was a broadcasting station and the unwanted transmitter was a base station of the land mobile service. The case when the unwanted transmitter is a broadcasting station is not treated in this Report.

This Report includes figures which indicate how the interference situation changes when using FM rather than AM at the unwanted transmitter.

2. Method of measurement

The method of measurement is described in detail in Recommendation 559 and Report 796. Some changes regarding the modulation mode and the modulation signal were made. Coloured noise should be used according to Recommendation 559 in simulating the broadcasting programmes. This was replaced by a sinusoidal tone of 1150 Hz, since in this case the unwanted transmitter is not modulated with a broadcasting programme. This tone modulating the wanted transmitter defines the reference level of the wanted signal at the receiver audio frequency output, and it serves as modulation at the unwanted transmitter. The reception mode was in all cases broadband FM (peak frequency deviation $\Delta f \pm 75$ kHz).

The RF protection ratio values were measured with a receiver ouput signal to interference plus noise ratio of 54 dB. The output noise was weighted with a psophometric network according to Recommendation 468.

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* This Report is brought to the attention of Study Groups 8 and 10.

The transmission mode of the unwanted transmitter was either AM or FM. In the case of AM, the modulation factor was m = 95% and with FM a modulation index of m = 1 was used.

The RF-protection ratios were determined by calculating the average value resulting from measurements of 16 domestic receivers. Receivers of different price classes and from different manufacturers (Europe, USA, Japan), were investigated.

3. RF wanted-to-interfering signal ratio in the case of an unwanted transmitter operating in the AM mode

The RF wanted-to-interfering signal ratios were measured at various receiver input levels. On account of the large spread of the results obtained, the geometrical average value was calculated

$$(X = \sqrt[n]{x_1 \cdot x_2 \cdot \cdots \cdot x_n}) \quad (1)$$

This corresponds to an "arithmetic averaging" in the logarithmic scale.

Curve M_2 in Fig. 1 depicts the RF wanted-to-interfering signal ratios for monophonic reception and a receiver input level corresponding to 1 mV at 75 ohms. The RF protection ratios of the worst and best receiver measured differ more than 30 dB at frequency separations of 300 kHz. The dashed line for frequency separations between 0 and 50 kHz indicates a linear estimation.

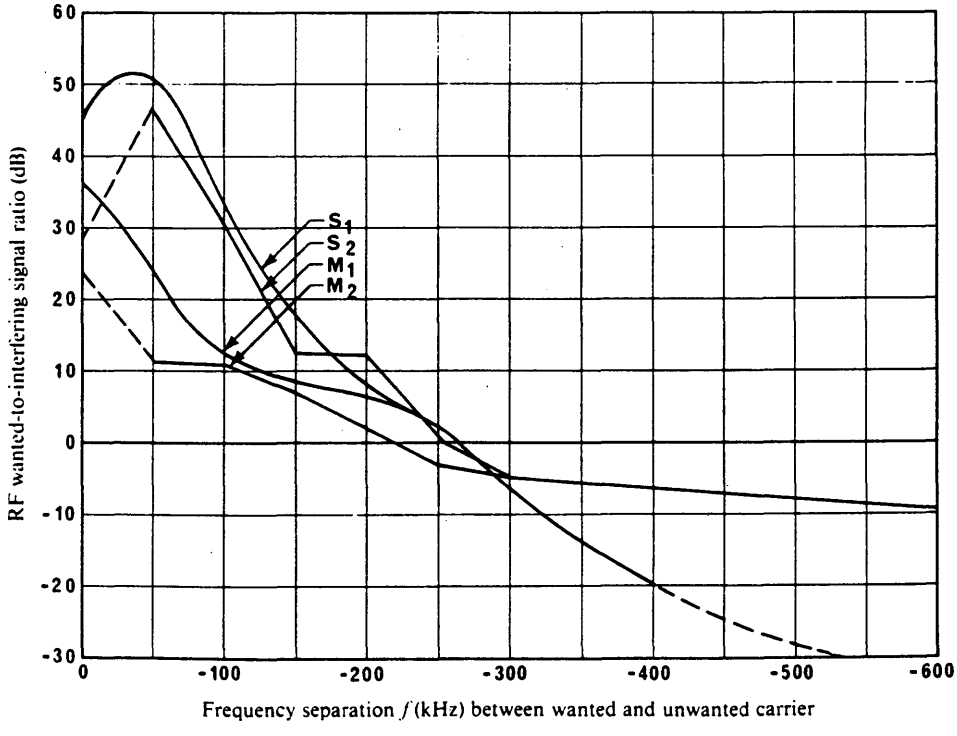


FIGURE 1 — *Average curves of the RF wanted-to-interfering signal ratios measured on 16 domestic receivers in comparison to Recommendation 412*

S_1: Stereophonic RF protection ratio curve of Recommendation 412.

S_2: Stereophonic RF wanted-to-interfering signal ratio curve measured on the 16 domestic receivers. Receiver input voltage 1 mV.

M_1: Monophonic RF protection ratio curve of Recommendation 412.

M_2: Monophonic RF wanted-to-interfering signal ratio curve measured on the 16 domestic receivers. Receiver input voltage 1 mV.

M: Mono

S: Stereo

For frequency separations between 250 kHz and 600 kHz the RF wanted-to-interfering signal ratios are nearly constant, the reasons being non-linear distortion effects at the front-end of the receivers (e.g. overloading effects, oscillator-pulling, etc.).

As curve S_2 in Fig. 1 shows, the RF wanted-to-interfering signal ratio curves for stereophonic operation differ from the curves obtained for monophonic operation only at frequency separations up to 300 kHz.

For comparison purposes the corresponding RF-protection-ratio curves for monophonic and stereophonic reception, as given in Recommendation 412 are also shown. It can be seen that up to 300 kHz frequency separation the average curves resulting from the 16 domestic receivers roughly follow the protection-ratio curve of Recommendation 412. The difference at larger frequency separations results from non-linear effects and insufficient selectivity. Above 400 kHz the recommended curve was extrapolated (dashed), since in the Recommendation, values are only given up to 400 kHz.

The measurements further pointed out that for different receiver RF input levels (e.g. 3.0, 1.0, 0.3 mV) the slopes of the curves are almost identical. The RF wanted-to-interfering signal ratios primarily depend on the level of the unwanted transmitter signal.

4. RF wanted-to-interfering signal ratios in the case of an unwanted transmitter operating in the FM mode

Figure 2 depicts the resulting RF wanted-to-interfering signal ratios for the 16 receivers investigated. The lower curve applies in the case when the unwanted transmitter is operating in the FM mode. A modulation index of 1 was used.

For comparison purposes the AM-curve is reproduced in the same Figure. Except for the co-channel condition the improvement obtainable when FM rather than AM is used at the usable transmitter is remarkable. At 600 kHz frequency separation this improvement is of the order of 40 dB.

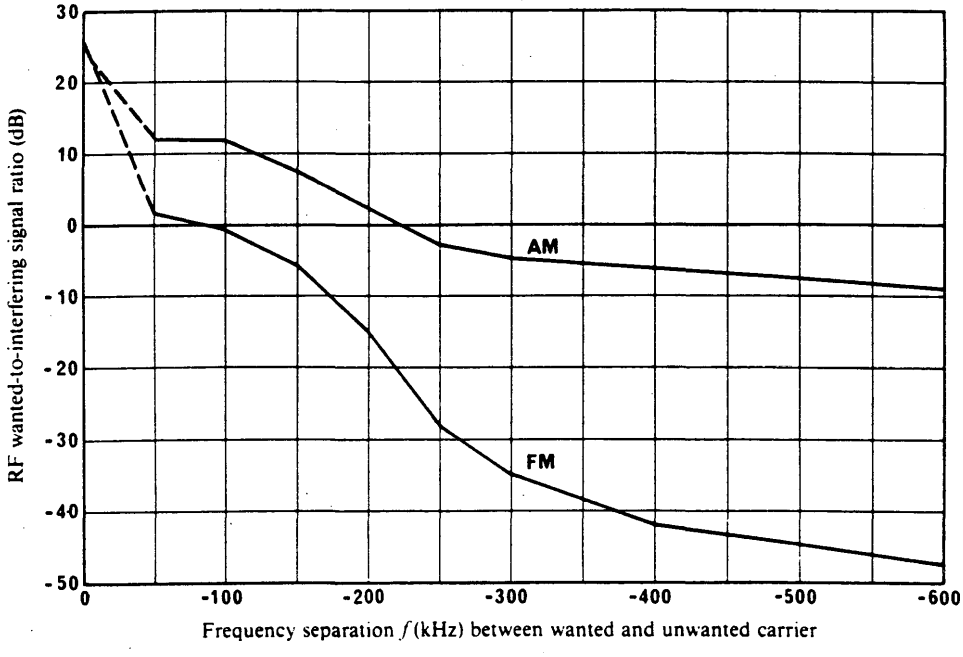


FIGURE 2 — *Comparison of the RF wanted-to-interfering signal ratio in the case of narrowband FM and AM*

AM: Unwanted signal amplitude modulated (modulation depth m = 95%). Receiver input voltage 1 mV.

FM: Unwanted signal frequency modulated (modulation index m = 1).

5. Conclusions

An FM broadcast receiver shows different susceptibility to double-sideband AM interference than FM interference. This is generally independent of the quality of the receiver. The RF signal-to-interference ratios for monophonic and stereophonic reception differ only at frequency separations of less than approximately 300 kHz. Interference caused by narrowband FM emissions to monophonic and stereophonic receiver is on average – depending on the modulation index and the modulation depth used – up to 40 dB less susceptible than AM interference for frequency separations greater than 400 kHz. The actual choice of the type of modulation may however depend on other factors (e.g. topographical condition, cost, etc. [Wey, 1967]).

Improvement of the technical characteristics of future broadcasting receivers, such as immunity from direct signal break-through, improved selectivity and large signal behaviour, would reduce the interference by the land mobile service to the reception in the FM broadcasting band. Such improvements are, however, of importance only at larger frequency separations between the two services. An improvement of the interference situation at frequency separations below 300 kHz would require a high cost in broadcasting receivers lest the reception quality be degraded.

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REPORT 528-1

CO-CHANNEL INTERFERENCE EFFECTS ON MULTIPLE PHASE SHIFT KEYING (MPSK) SYSTEM PERFORMANCE

(Questions 18/1, 44/1 and 45/1)

(1974-1978)

1. Introduction

Several CCIR Questions, such as 18/1, 44/1 and 45/1, seek methods and results of communication theory that would increase the efficiency of spectrum use. A case of considerable present – and even greater future – interest to high speed data technology deals with the performance of MPSK systems (short for coherent M-ary or Multiple Phase Shift Keying, $M = 2, 3, 4, \ldots$) in the presence of noise and co-channel interference.

2. Definitions

Assume that each M-ary symbol, same as a binary or non-binary signal element, has duration T, and that the received signal waveform in absence of other inputs is:

$$s(t) = \sqrt{2S} \cos(\omega_0 t + \theta(t)) \quad (1)$$

where the instantaneous coherent phase $\theta(t)$ is some $2\pi m/M$, with m an integer $0 \leqslant m < M$. The signal power is S and the signal energy per symbol is ST. The received noise $n(t)$ is white Gaussian with one-sided spectral density N_0. To signal and noise we add interference $i(t)$. This interference is co-channel when its centre frequency is also ω_0.

Typical MPSK performance is given in P_e versus S/N plots, with one or more parameters identifying the receiver type, filters in the signal path, signal distortions, interference conditions, and so forth.

The two primary terms are:

P_e: The symbol error probability. The bit error probability, which is also often used, cannot exceed P_e, nor can it be less than $P_e/\log_2 M$. The two probabilities are equal for $M = 2$.

S/N = $10 \log_{10} (ST/N_0)$ is signal symbol energy per noise spectral density in decibels. It exceeds the bit energy over N_0, in decibels, by $10 \log_{10} (\log_2 M)$ decibels, and can be interpreted as signal-to-noise power ratio $S/(N_0 T^{-1})$ in decibels.

3. Results

3.1 *Theoretical results*

The performance of various MPSK systems has been studied by many workers [Aein and Turner, 1973; Akima and Nesenbergs, 1973; Fang and Shimbo, 1973; Goldman, 1971; Jeruchim and Lilley, 1972; Palmer and Lebowitz, 1973; Prabhu, 1969; Rosenbaum, 1969, 1970 and 1973; Shimbo and Fang, 1973].

Figure 1 shows the performance of a sampler or matched filter receiver in the presence of noise and no interference [Lucky *et al.*, 1968; Stiffler, 1971]. The curves are parametric in $M = 2$, 4, 8 and 16.

The inclusion of interference $i(t)$ starts with specification of the number of distinct constant envelope, random angle, interfering signals contained in $i(t)$. Let us denote this multiplicity by K. Thus:

$$i(t) = i_1(t) + i_2(t) + \ldots + i_K(t)$$

and the total interference power

$$I = I_1 + I_2 + \ldots + I_K$$

correspondingly. The signal-to-interference ratio (S/I) is defined as the ratio of desired-to-undesired (interfering) signal power in decibels. It is the same as the ratio of signal energy per symbol (or per bit) to interference energy per symbol (or per bit) in decibels.

The simplest case occurs when the interference is a single unmodulated carrier, $K = 1$, at centre frequency ω_0 and with a random uniformly distributed phase. The performance degradation caused by this interference to ideal MPSK is shown in Figs. 2a to d.

The effect of $K > 1$ is to further deteriorate the performance. The effect is shown in Figs. 3a to 3b, each chosen for a fixed S/I value but both with $M = 2$. It seems that $K = \infty$ should have the worst effect of all choices of K.

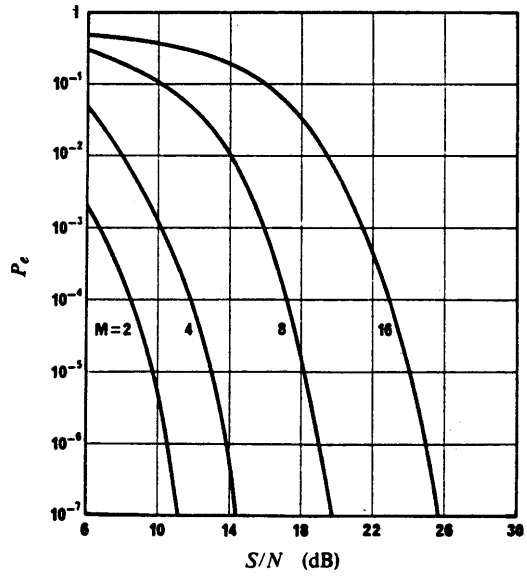


FIGURE 1 — *Performance of ideal MPSK with no co-channel interference*

M = 2, 4, 8 and 16

When the co-channel interference contains modulated constant envelope signals, the effects become far more complex and are not well documented. While theoretical estimates for a single, $K = 1$, angle modulated interference suggest performance degradation shown in Fig. 4, more results can be deduced through simulation. Fig. 5 shows results from a simulated QPSK ($M = 4$) receiver being interfered with by a modulated undesired QPSK signal [Palmer and Lebowitz, 1973]. The model and details of the simulation are as described in Report 519. The data show that for high signal-to-interference ratios the theoretical derivation agrees reasonably well with the simulated results. For low signal-to-interference ratios there is considerable difference between the two procedures which is caused by the approximations inherent in the analytic approach. The results in general indicate that the theoretical bounds are valid and for low signal-to-interference ratios additional analytic complexity needs to be considered. In particular, the simulation described in Report 519 should be extended to the arbitrary M phase case and to a multiplicity K of interferers.

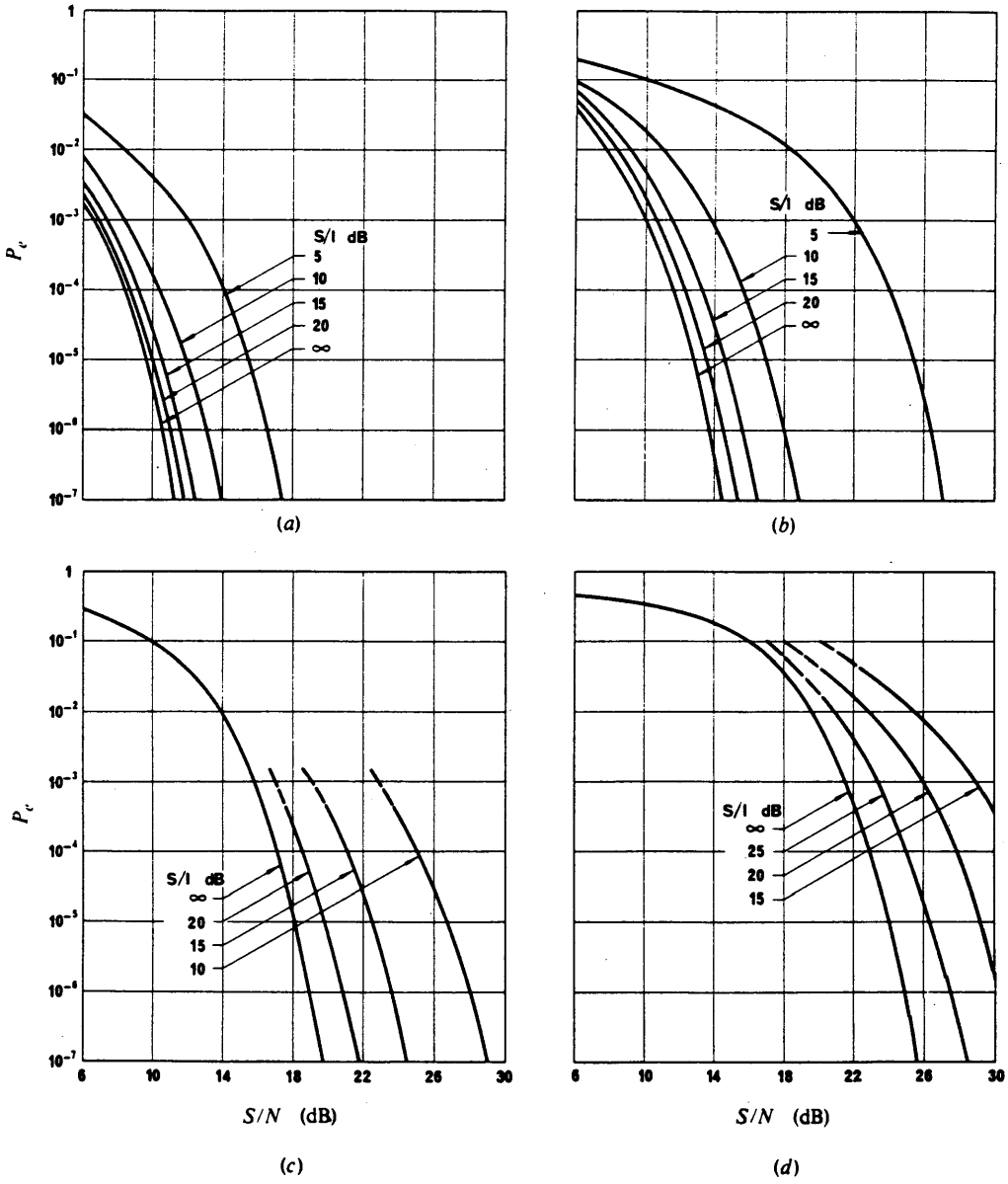


FIGURE 2 — *Performance of ideal MPSK with a single, K = 1, unmodulated interference*

(*a*) $M = 2$ (*c*) $M = 8$
(*b*) $M = 4$ (*d*) $M = 16$

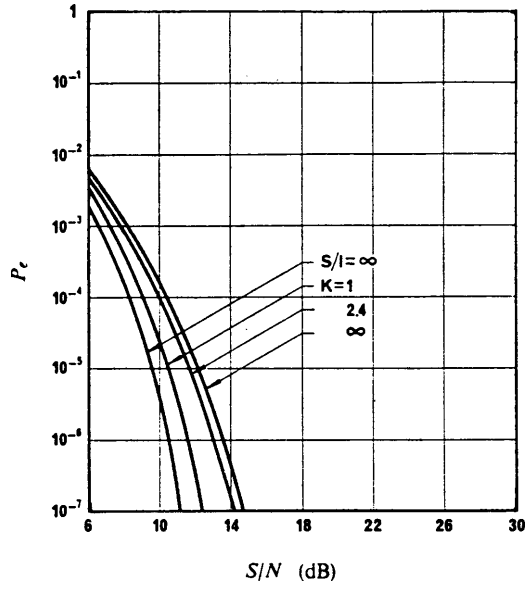


FIGURE 3*a*

The effect on binary (M = 2) MPSK performance by K = 1, 2, 4 . . . unmodulated interfering signals S/I = 15 dB

K: number of interfering signals

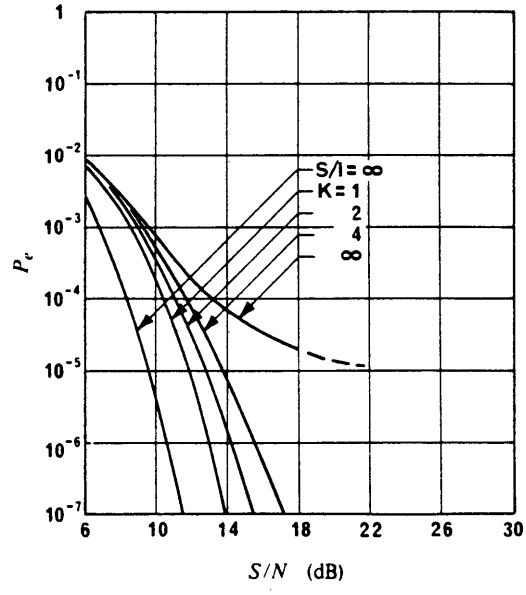


FIGURE 3*b* — *The effect on binary (M = 2) MPSK performance by K = 1, 2, 4, . . . unmodulated interfering signals, S/I = 10 dB*

K: Number of interfering signals

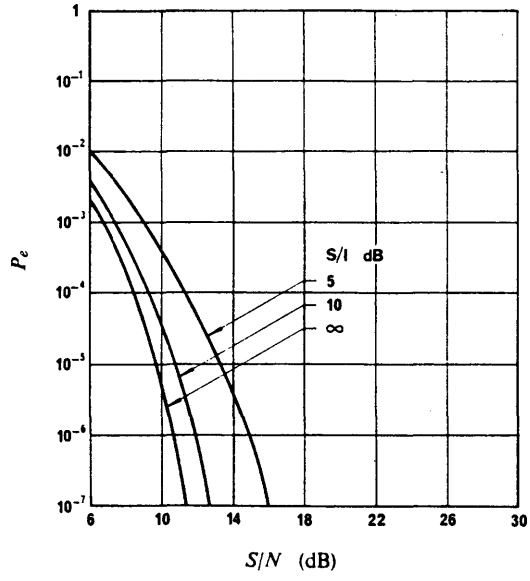


FIGURE 4

Binary (M = 2) MPSK performance estimate for a single (K = 1) interference with arbitrary angle modulation

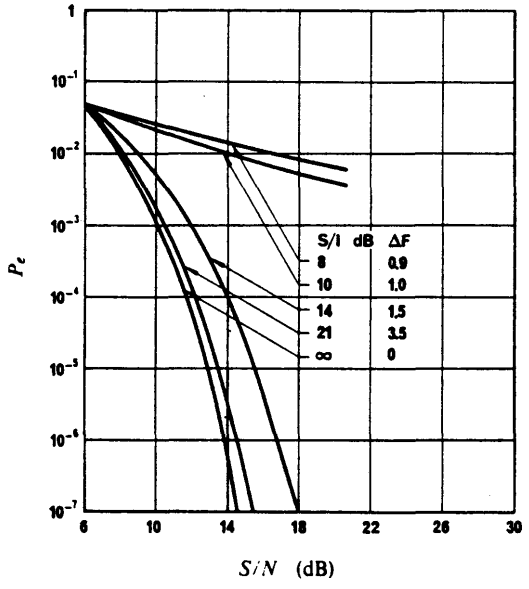


FIGURE 5 — *QPSK (M = 4) performance simulation results for a single (K = 1) QPSK interference that has a normalized frequency offset (ΔF). CW*

$$\Delta F = \frac{f_1 - f_0}{\text{Symbol rate}}$$

3.2 *Measured results*

The probability of errors for a 4-phase PSK receiver subjected to interference were measured using the test set up shown in Fig. 6. The PSK demodulator included a carrier recovery circuit with a single tuned filter that had a bandwidth of 400 kHz and a detection/decision circuit.

The desired signal was a 4-phase PSK signal modulated at 30 Mbauds. The interfering signals (I_1, I_2) are unmodulated sinusoidal waves. In Fig. 7, for $K = \infty$, band limited white noise from noise generator II was used on the interference source. The interfering signal in Fig. 9 was a 4-phase PSK signal modulated at 30 Mbauds.

Figure 7 gives measured results on the relation between wanted signal to noise power ratio (S/N) and bit error ratio (P_e) with the number of interfering signals (K) as a parameter. The results have the same tendency as the calculated results (Fig. 3b).

Figure 8 shows the measured results on the relation between S/N and P_e with Δf and S/I as parameters. It can be seen from this figure that bit error ratio increases when the value of Δf approaches zero. This seems to result from the effects of interfering signals on the carrier recovery circuit.

Figure 9 shows equivalent change in S/N versus S/I when the desired signal is interfered by 4-phase PSK signal modulated at 30 Mbauds. The equivalent change in S/N is the difference between S/N required to obtain a given bit error ratio (10^{-4} or 10^{-6}) in the absence of interference and the S/N required for the same error ratio in the presence of interference. The above results indicate that the effects of the interfering signal on the carrier recovery circuit cannot be neglected in the region of small value of S/I.

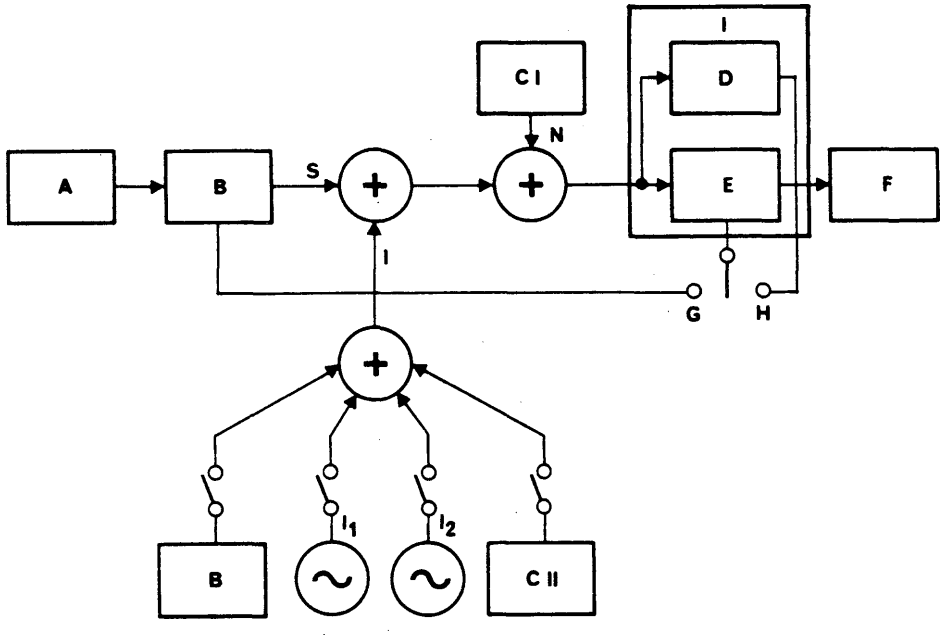


FIGURE 6 — *Simplified block diagram for the measurement*

A: *PN* generator
B: 4 PSK modulator
C: Noise generator
D: Carrier recovery
E: Detection decision
F: P_e measurement
G: Original carrier
H: Recovered carrier
I: Demodulator

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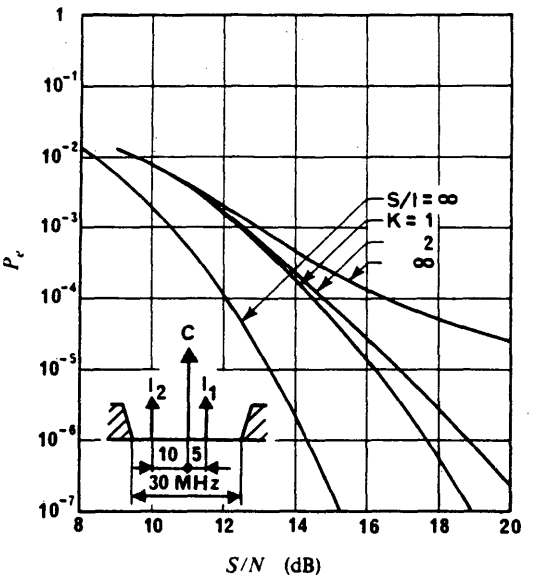


FIGURE 7 — *Measurement results on performance of 4-phase PSK with multiple unmodulated interferences*

K: number of interfering signals

S/I = 14 dB for curves with K = 1, 2 and ∞

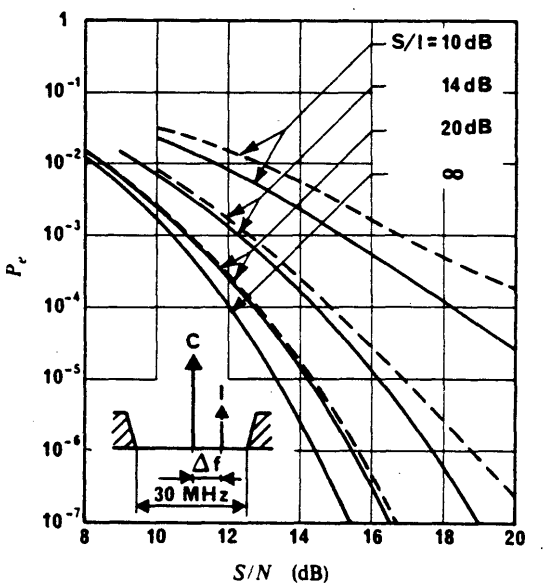


FIGURE 8 — *Measurement results on performance of 4-phase PSK with a single unmodulated interference*

——— : Δf = 5 MHz
– – – – – – : Δf = 0.5 kHz

M = 4
K = 1 CW

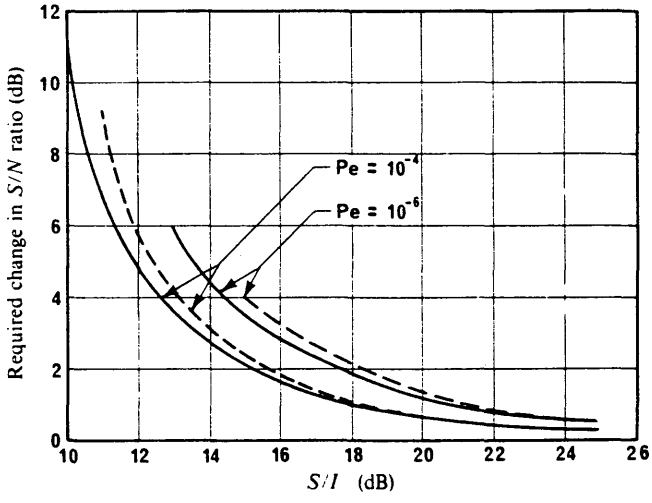


FIGURE 9 – *Change in S/N ratio versus S/I ratio for error rates P_e of 10^{-4} and 10^{-6}*

———— : with original carrier
– – – – – – : with recovered carrier (see Fig. 1)

M = 4
K = 1 Modulated

4. Conclusions

Co-channel interference degradation of the error probability performance of coherent MPSK modems under the quoted conditions can be estimated from the given theoretical curves in Figs. 2 to 4, the simulated results in Fig. 5 or the experimental measured results given in Figs. 7 through 9.

The above results reflect a relatively small portion of the co-channel MPSK interference effects. Many refinements and extensions need be pursued, notably for different number K of interfering signals with different powers, modulations, and different frequency offsets.

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REPORT 1111

PLANNING CONSIDERATIONS FOR MANAGEMENT OF THE RADIO SPECTRUM

(Question 55/1)

(1990)

1. Introduction

The purpose of any planning effort is to organize and focus thoughts and actions for the efficient, effective achievement of directed or agreed goals and objectives derived from national requirements. A plan promotes action, rather than reaction. It may be applicable for a specific period or event with finite time bounds, or it may be a standing statement, updated periodically to reflect changing policy or events dictated by circumstances. It may be written or verbal, formal or informal, general or specific; each form has its advantages and disadvantages.

Management and use of the radio frequency spectrum demands the direction of effort and cohesion provided by the act of planning and producing a recognized plan if this limited resource is to adequately support the national goals and objectives of the various administrations. Most importantly, spectrum management planning must support, and follow, the major events or drivers of the spectrum users.

Spectrum management's role is to find ways to fulfil a spectrum requirement, after an administration's requirements and operational planning determines what is to happen. Due to the demands, the complexity, and the political, economic and technological considerations attendant to spectrum use, planning over the long range is prudent. This contemplates, perhaps, a five-year national planning period extending up to 15 years for international considerations.

The purpose of this text is to present certain considerations that warrant addressing during the formulation of any administration's long range plan for management and use of the radio frequency spectrum. A long range plan outline is provided as a checklist of typical items to be considered.

2. Content of a plan

The planning activity falls into at least two categories, namely, planning related to technical utilization of the spectrum - sometimes referred to as spectrum planning, and planning related to the spectrum management function within an administration. The latter tends to vary from administration to administration depending upon legislative structures and government organization. As a result this can only be dealt with in very general terms.

The planning associated with management of the spectrum can be dealt with in a generic fashion, as it deals with the technical utilization of the same resource, irrespective of administration.

The precise form that an administration's long range plan may assume will be highly dependent on the emphasis placed on each of the many factors bearing on spectrum use in that administration. These factors also greatly influence the availability of resources with which to execute the various planned functions of spectrum use and management. There is considerable merit in planning alternative methods and techniques for maintaining continued, viable spectrum management.

The following discussion provides an outline which considers the more critical elements of a long range plan for spectrum management, and describes the major elements which can be used in developing the plan. Emphasis is not included, as this can only come from the administration conducting the planning. It may also not provide for any and all elements and factors which might be considered critical due to the peculiar demands on use of the spectrum by a specific administration. It is a generalized long range planning guide.

The material is organized into spectrum planning, which can be regarded as technical in nature, and administrative matters.

2.1 Spectrum planning

2.1.1 Spectrum use objectives

The radio frequency spectrum is a natural resource which should be effectively utilized for the benefit of a country. The specific uses of the spectrum can be to support security and defense, safety and law enforcement, commerce, transportation, education, etc., as required by that administration. Again, it is emphasized that spectrum use objectives must support and provide direction to meet the requirements of operational users.

2.1.2 The spectrum resource

A description of the various allocated uses of the spectrum within the administration's sphere of interest and influence, and its current level of use by frequency band or radio service is necessary. This would consist of a national table of allocations, which would be in conformity with Article 8 of the Radio Regulations, and the list of frequencies in use. It should identify spare capacity and areas where there may be shortages.

2.1.3 Future spectrum demand

The plan should identify, to the degree possible, future use of the spectrum derived by forecasting pertinent technological and economic influences. To the extent practicable the future implications of current and projected use of the spectrum resource should be identified. Plans for proposing and negotiating necessary changes to national allocation tables, to be pursued at cognizant national and international fora, should be developed.

2.1.4 International activities

The principle focus of international spectrum management activities is through the International Telecommunication Union (ITU) and its associated WARC and CCIR activities.

There are other international or regional organizations concerning functions such as aviation (ICAO), meteorology (WMO), etc. which generate requirements for radio spectrum, and suitable liaison on a national basis is necessary.

The planning activity associated with international matters can arise when changes are required to the international or regional allocation table to accommodate a new technology or service, or the growth of an existing service. Alternatively, it might arise if an administration wished to have a particular sharing model, for example, endorsed internationally by the CCIR in the form of a report or recommendation.

2.1.5 Spectrum engineering

This involves the basic technical aspects of spectrum management, including the development of equipment standards, channel plans, compatibility models, protection ratios and sharing models. Spectrum engineering ensures the practical feasibility of the various possible uses of the spectrum and enables the options to be evaluated in deriving a future plan for spectrum use. Finally, the engineering methods and procedures to be followed in the spectrum management day-to-day activities should be included.

2.1.6 Long range plans

This should provide a clear statement of the specific long term actions planned for each element discussed in the preceding portions of the plan. This activity may be oriented by frequency band, radio service, or spectrum management functional area, but should result in a concise statement of the requirement. The planning action should include exactly what is planned, the desired results (objectives), and the strategy for pursuing the planned action. This part of the effort must also thoroughly consider the interrelationship which exists between the various frequency bands and radio services of the spectrum. This interrelationship may be described as a three dimensional array, or matrix, as depicted in Figure 1.

2.1.7 Schedule

A graphical or tabular presentation of the schedule for initiating, pursuing and achieving each planned action must be developed. The interdependence of related actions should be shown, and significant milestone events should be indicated.

2.2 Administrative matters

2.2.1 Authorities

The plan should include the authorities provided to spectrum managers by the enabling legislation and the procedures to follow in carrying out those authorized activities.

2.2.2 Resource requirements

Since it is fundamental that no action may be accomplished without the expenditure of some resources, it would not be realistic to develop a plan that did not consider the resources required to support the achievement of the plan's objectives. This portion of the plan should address the time, staff, hardware, software, facilities, supporting material and actions, and funding required to

maintain and advance each of the planning elements. The financial resources required can be obtained from licence fees, direct charges for services, direct government funding, or a combination of these. Style or policies may dictate that this portion of the plan not be bound in the final document, but it must be developed in close coordination with the planning effort if adequate support of the plan is to be realized.

2.2.3 Spectrum management functions

This portion of the plan should provide a detailed discussion of the organization and activities of each of the ongoing or proposed functions of spectrum management conducted by the administration. Such elements as policy development and implementation, decision-making processes, management processes (e.g. service to the public, licensing and fee collection, compliance), disaster recovery, management support (e.g. electronic data processing) and training should be included.

2.2.4 Implementation

To be effective and realizable, a long range plan for management and use of the spectrum must be thoroughly coordinated during its development, and must be implemented through promulgation by a competent authority. Since the radio spectrum supports the users and operators of telecommunications systems and equipment, the plan for its supporting use must be developed in consonance with the activities of these entities, and must reflect their current and projected requirements. And, to acquire the status necessary for viability, the plan must issue from a recognized source with the authority to ensure compliance.

Upon promulgation, the plan should be disseminated to all entities having a statutory or practical requirement for the information. Implementation planning should also provide for scheduled reviews of milestone and objective achievement, and - because of the dynamic nature of telecommunications - for periodic revision.

3. Long range plan outline

A generic outline for a typical, or representative, long range plan is postulated in the attachment. The outline includes the factors covered in this discussion, but need not be limited to them. Administrations having unique requirements or arrangements for spectrum management must certainly address those points. Conversely, not every administration will need, nor desire, to include all of the listed factors.

The primary purpose of the outline is, therefore, to serve as a checklist. As such it will ensure that, as a minimum, the most critical factors are addressed. Further, the act of working through the outline while formulating an administration's long range plan will be conducive to stimulating the planner's thoughts, thus leading to consideration of additional factors which may be peculiar to that particular administration.

While developing the outline into a long range plan, the planner also has the opportunity to place the required additional emphasis on those factors which are key to spectrum management within that administration.

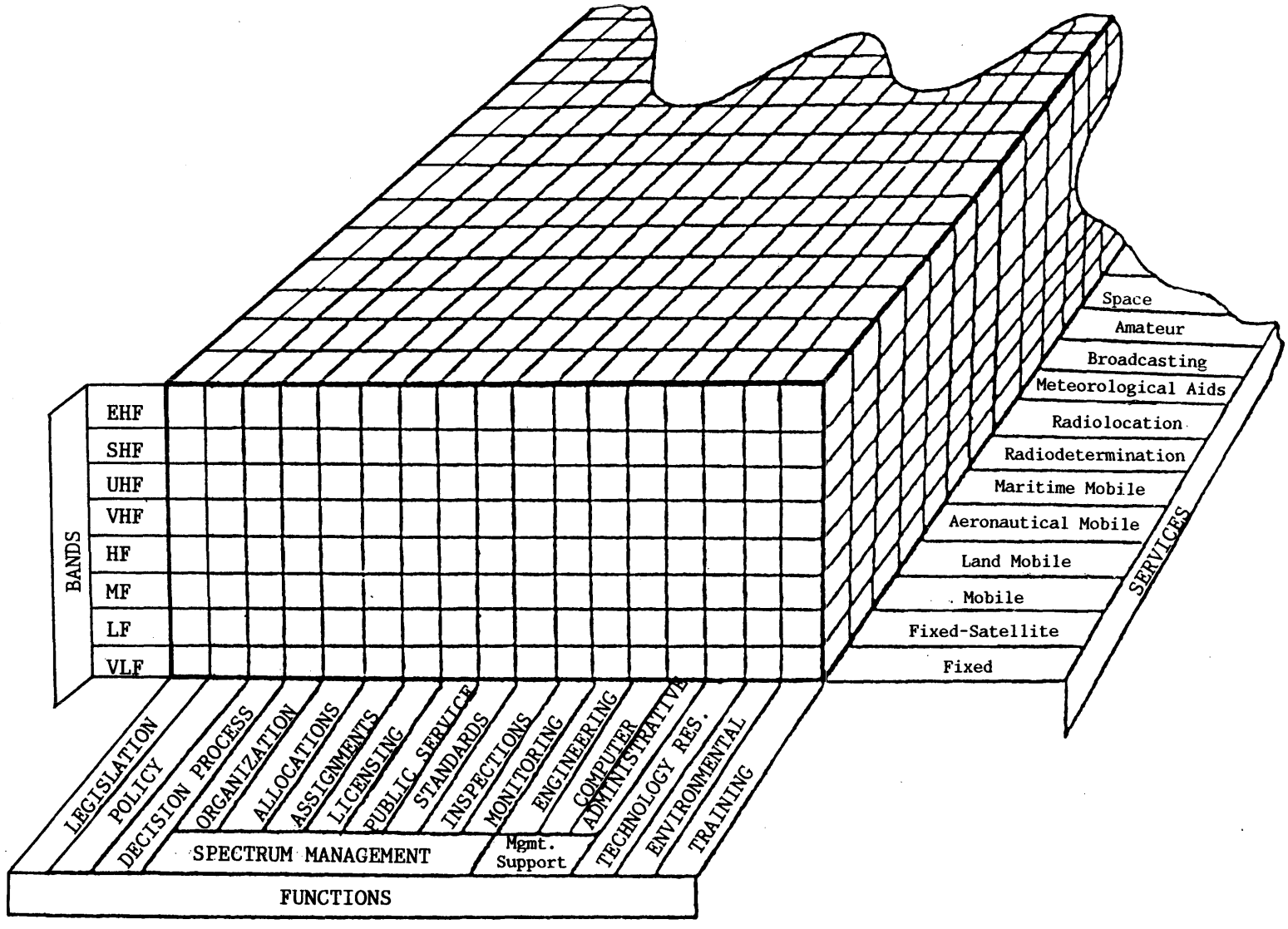


FIGURE 1

Interrelationship of planning factors

4. Summary

The limited spectrum resource must be effectively managed if it is to adequately support the national goals and objectives of an administration. Planning for use of the spectrum over the long range, and gaining recognition of this planning effort, facilitates the spectrum management effort.

By careful selection and consideration of the key planning factors, and emphasizing those most critical to the administration's spectrum management programme, those which are most important will receive the degree or visibility and attention necessary to compete for the resources needed to accomplish the desired actions.

Elements of the long range plan must be thoroughly coordinated with cognizant spectrum management and user organizations during the planning process to assure that adequate support will be rendered to, and by, the plan in its implementation phase.

Upon completion of planning and coordination, the plan must be issued by a recognized authority in order to ensure adherence to the provisions and guidance contained in the document.

OUTLINE

LONG RANGE PLAN FOR MANAGEMENT OF THE RADIO SPECTRUM

Spectrum planning

1. Spectrum use objectives

Objectives meeting requirements of users in accordance with national policies, for example:

- safety and law enforcement;
- commerce and transportation;
- education;
- national security;
- broadcast.

2. Spectrum resource

- National table of frequency allocations in conformity with Article 8 of the Radio Regulations.
- List of frequencies in use.
- Spare capacity and shortages.

3. Future spectrum demand

- Demand and economic and technological forecasts.
- Impact of international regulations.

4. International activities

- Strategies for participation in the ITU and other international and regional organizations.
- International agreements.

5. Spectrum engineering

- Equipment standards, channel plans, compatibility models, protection ratios, engineering methods and procedures.

6. Long range plans

- Specific long-term actions for each element above:
  - to spell out the interrelationships between services, frequency bands and other planning factors.
- Schedule of activities and targets.

Administrative matters

1. Authorities

   - Enabling legislation.
   - Delegated authority.
   - Procedures.

2. Resource requirements

   - Source of funding.
   - Personnel expertise.
   - Computer resources.
   - Equipment.

3. Spectrum management functions

   - Policy development.
   - Enforcement and regulation.
   - Licensing and fee collection.

4. Implementation

REPORT 1112

MEASURES OF AGGREGATE SPECTRUM USE IN A FREQUENCY BAND

(Question 47/1)

(1990)

1. Introduction

Question 47/1 asks for ". . . appropriate methods of describing 'spectrum utilization' . . .," and Report 662-2 defines the measure of spectrum use as the product of the bandwidth, geometric space, and time denied to other users.

This report describes two methods for computing the spectrum utilization for all systems authorized in a frequency band. One method, described in section 2, computes an absolute measure of the spectrum denied to potential users [Mayher et al., 1988]. It is illustrated with an application to a band containing point-to-point relay systems and a few aeronautical mobile systems. The second method (spectrum quality indicator (SQI)) described in section 3, computes the utilization achieved to assess how effectively the spectrum is being managed within the assignment strategy used relative to the demand for spectrum in the region evaluated [Hafez and Towaij, 1989].

2. Absolute measures of spectrum utilization

As pointed out by Report 662, the geometric space of interest depends on the type of service being analysed. For terrestrial services, area is the relevant space. The mapping of spectrum utilization involves the two-dimensional representation of the spectrum resources denied to potential new systems by existing systems. The basic technique is to calculate the level of spectrum utilization at points representing incremental areas within a region (for example, a country) and to create a map shaded according to those values. At a point, the measure of spectrum utilization is the product of bandwidth and time denied to other potential users. The total spectrum utilization for the area being studied is then the sum of the products of spectrum utilization and incremental area for each point.

As stated in Report 662, the dimension of time can be ignored for many applications. In this case, the spectrum utilization of a frequency band at a point is simply the total bandwidth denied to other potential users. This assumption will be used in the rest of this report.

The computer model described in this report [Mayher, et al, 1988] calculates the spectrum utilization of authorized systems in a specified band. Report 662 points out that tranmitters deny areas to receivers only, and receivers (by virtue of the protection afforded them) deny locations to transmitters, so measures of both kinds of use must be computed. For most applications, it is simpler to produce a single map using the union of the receiver and transmitter denial as the total system utilization.

It is clear that denial depends on the characteristics of the equipment being denied. For example, receiver denial by a transmitter depends (among other things) on the height of the receiver's antenna -- a higher receiver antenna will deny a greater area than a lower antenna. Denial of a transmitter by an existing receiver depends (among other things) on the power of the denied transmitter. Thus every measure of spectrum utilization depends on the (perhaps hypothetical) characteristics of the denied equipment.

2.1 Example application to fixed services in the USA

The general concept described in Section 2 has been applied to the fixed services and implemented in a computer program [Mayher, et al, 1988]. A computerized data base of all authorized systems is used as input.

The calculation procedure is developed by Mayher et al., [1988] and Haines [1989] and is summarized here. Except as noted, these calculations are for denial of potential new receivers by existing transmitters. The procedure for computing the denial of transmitters by receivers is similar.

The general procedure is to compute the total bandwidth denied at each of a regular grid of geographic points. At each point, the power available from every authorized transmitter in the region is computed and compared with an interference threshold. If the interference threshold is exceeded, the bandwidth of the authorized transmitter is marked as denied at that point. Both co-channel and adjacent-channel interference are considered. When all authorized systems have been examined, the total bandwidth denied at the point is determined and recorded.

The area and bandwidth denied to a reference fixed service receiver by the existing transmitters depends on the orientation of the receiver antenna main beam. Transmitters at locations in the reference receiver main beam will deny far more area than those at locations in the nulls of the receiver antenna pattern. This fact results in two measures of spectrum utilization. The first, called the spectrum use bandwidth (SUB), treats all existing transmitters as though they were in the receiver antenna mainbeam when calculating interference power. This results in a measure that is the upper bound of spectrum utilization in units of Megahertz.

The second measure of spectrum utilization, called the spectrum use factor (SUF), multiplies the fraction of the total band represented by the denied bandwidth by the fraction of receiver antenna pointing angles that will result in interference. This results in dimensionless measure of spectrum utilization with smaller relative values than the SUB.

The antenna radiation patterns used are determined from the antenna gain and formulas in Reports 391-5 and 614-2.

The data base for the program includes the following relevant information for each transmitter: site latitude and longitude, site elevation, antenna height above site, main beam azimuth, antenna gain, transmitter power, center frequency, and transmitter bandwidth. The user specifies the following characteristics of the receiver that is to be denied: antenna height above site, antenna gain, bandwidth, received signal strength, and required protection ratio.

For the calculations in this report, the characteristics of the denied system were the median values for all authorized systems in the band.

2.2 Example calculations

The spectrum utilization in the United States in the band 1710-1850 MHz was calculated using the program. There are over 5000 assignments assignments in this band in the US. Most of the assignments are in the fixed service, but there are some mobile and other assignments. For the example calculations, the denied service was the fixed service. The utilization of the transmitters and the receivers of the authorized systems were combined. Figure 1 is a map of the United States showing the spectrum use bandwidth (SUB) for this band (Mayher et al., [1988] show a map for the entire United States). The band is heavily used. The large circular areas are due to aeronautical mobile assignments.

Figure 2 shows the total area and percent of area of the US having each value of SUB. The entire 140 MHz band is used in about 30 % of the United States. The total spectrum-area use product in this band in the US is 8.9×10^8 MHz-km^2.

The spectrum use factor for this band is shown in Figure 3. This factor includes consideration of the spatial discrimination due to the narrow beam antennas used by the fixed services, so the measure of spectrum-space utilized is smaller.

2.3 Applications

There are several ways that the measures of spectrum utilization can be used by administrations [Haines, 1989a]. These include:

- Maps of spectrum utilization can show areas of spectrum congestion where restrictive standards and intensive coordination are necessary to assure efficient use of the spectrum.

- Spectrum planning for allocation to specific services can be aided by quantitative comparisons of the intensity of utilization of different bands in each geographic region.

- Periodic calculations of the spectrum use in each band can reveal trends that can be used for strategic planning.

2.4 Summary

A method for calculating spectrum-space and mapping spectrum utilization has been developed and implemented in a computer program. The measure of spectrum use depends on the characteristics of the system being denied. The results can be used for frequency management and spectrum planning.

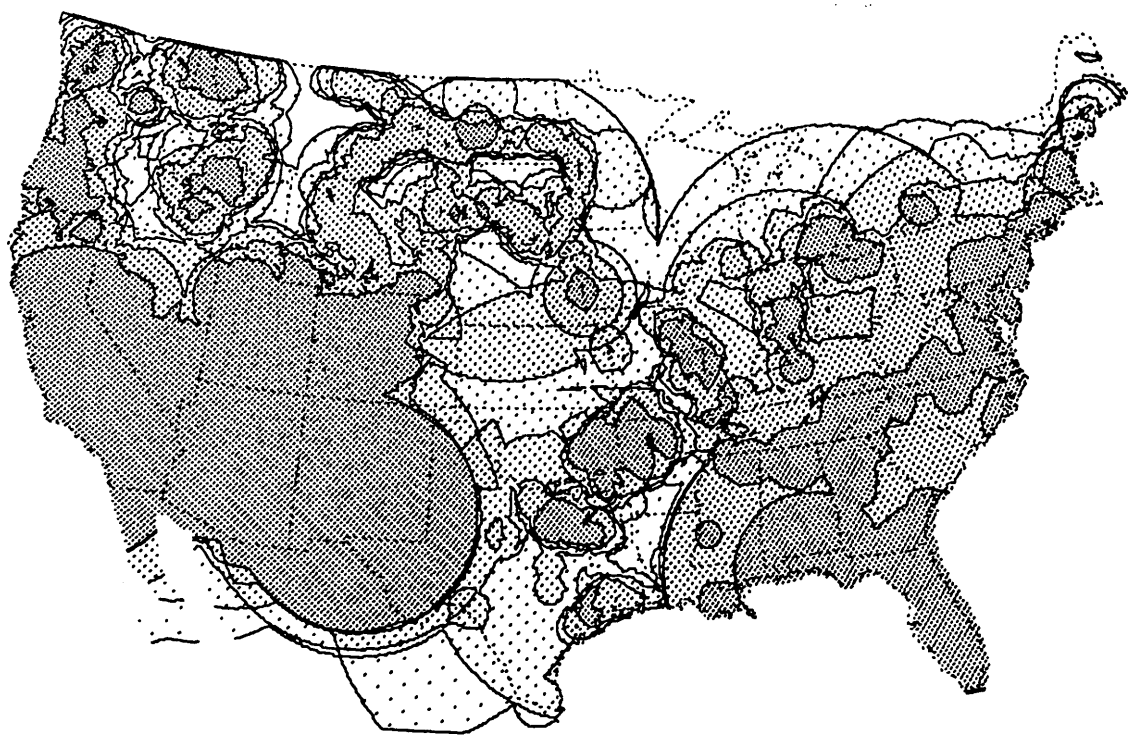
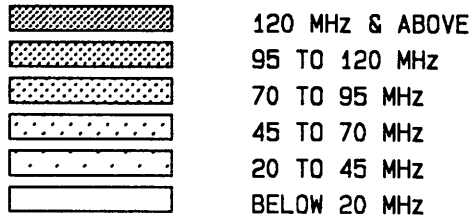


Figure 1. Spectrum use bandwidths in the 1710-1850 MHz band in the 1710-1850 MHz band in the United States.



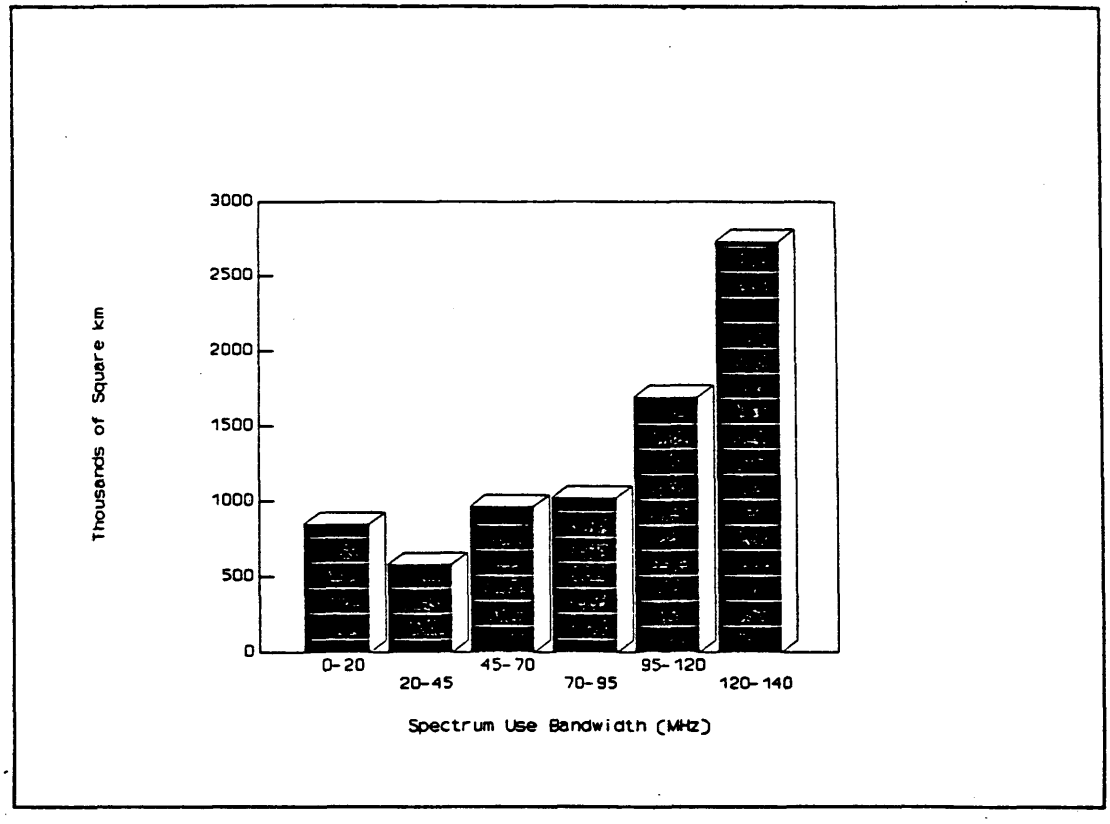


Figure 2. Area used for various values of the spectrum use bandwidth in the 1710-1850 MHz band in the United States.

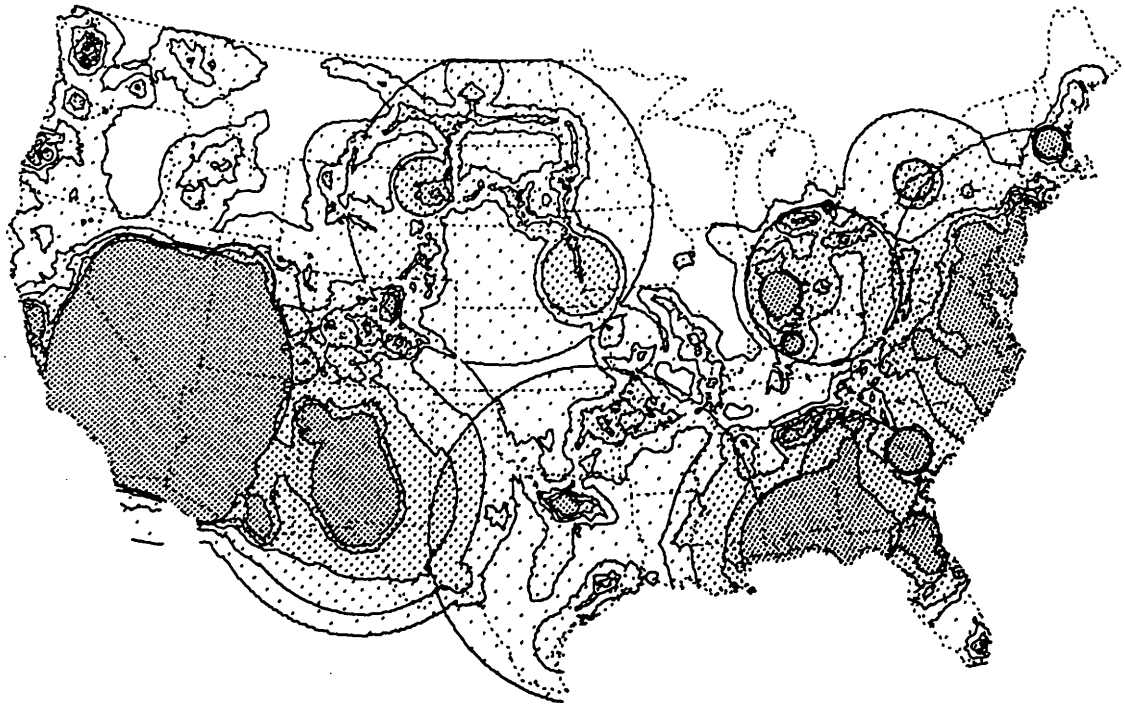
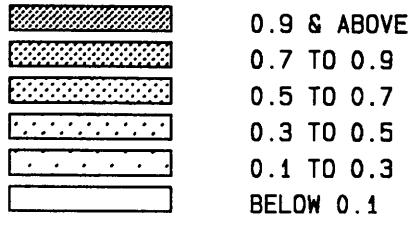


Figure 3. Spectrum use factor in the 1710-1850 band in the United States.



3. Spectrum utilization relative to demand - The spectrum quality index (SQI)

3.1 Definitions

3.1.1 Spectrum dimensions

According to Report 662, spectrum utilization is measured by the number of spectrum units it occupies, where

spectrum unit = Hz • sec • metre^3

3.1.2 Occupied Spectrum

The "Occupied Spectrum" is the number of spectrum units required by the radio system to maintain the radio communications between the various stations operating within the coverage area of the system. It is the product of the coverage volume, the total occupied time during a peak hour and the occupied radio channel bandwidth.

3.1.3 Denied Spectrum

The "Denied Spectrum" is the number of spectrum units denied to other users as the result of the operation of a radio system. To calculate the "Denied Spectrum", all the interference effects such as, intermodulation, adjacent, co-channel, desensitization etc. must be taken into consideration coupled with region of influence and the time of activity.

3.1.4 Spectrum Value

Spectrum value is a measure of the degree of demand on the radio spectrum for a given service within a specified geographic area.

3.1.5 Spectrum quality

The spectrum quality is a measure of spectrum occupied relative to the spectrum denied in a frequency band and geographic region

3.2 Measure of spectrum quality

3.2.1 Spectrum value factor Γ

Within a specific service demand on the radio spectrum is not equally distributed throughout a geographic area. For example, in the land mobile service, demands are concentrated in the urban areas and the peak of the demand is found in the center cores of the metropolitan areas. Accordingly spectrum is more valuable in areas of high demand than those with low demands. Spectrum has no value in an area where it is not required. Denied spectrum is more critical in the areas where the spectrum is in high demand than in deserted areas.

It may not be possible to know the actual demand for some kinds of spectrum in urban areas. A desire for the service provided by the spectrum in question may not be expressed by a license application because the potential user knows that the band is saturated, or because of regulatory restrictions. However, the amount of spectrum units occupied may be taken as a first approximation to the relative demand in an area. This approximation is made throughout this section of this report.

To understand the Spectrum Value factor Γ, consider a geographic region covered with a volume V. Let V be divided into equal-size cubes. The demand in each cube is different and is represented by the number of spectrum units occupied in each cube (product of time, frequency and volume). The greater the demand in a cube, the higher the spectrum value factor. This value may be represented mathematically by Γ and has numerical value ranging from 0.0 to 1.0, with the greater number indicating higher spectrum value. Mathematically, the Spectrum Value factor in cube $\Gamma(n)$ is directly proportional to the number of spectrum units $\beta_t(n)$ demanded in the cube;
Accordingly $\Gamma(n) = \beta_t(n) / \beta_{total}$

where β_{total} is the total demands in volume V

3.2.2 Spectrum quality indicator (SQI)

It is a measure of how efficiently the radio spectrum is being managed for a given service within a geographic area. The SQI is not applicable as a relative measure to compare the spectrum utilization efficiency of different services.

Spectrum Quality Indicator is accordingly given by the relation

SQI = TOTAL WEIGHTED OCCUPIED SPECTRUM/
TOTAL WEIGHTED(OCCUPIED + DENIED) SPECTRUM

$$SQI = \Sigma\, \Gamma(n)\, \beta_{t(n)} / \Sigma\, \Gamma(n)\{\beta_t(n) + D_t(n)\} \qquad (1)$$

where $\Gamma(n)$ and $\beta(n)$ have been defined, and $D_t(n)$ is the number of spectrum units denied in cube n in addition to those used for communications. $D_t(n)$ is called the "marginal denied spectrum".

3.2.3 Effect of Spectrum Value on Spectrum Quality Indicator Calculations

To evaluate the effect of Spectrum Value weighting factor Γ on SQI, let us assume that the traffic demands β_{total} is concentrated into one cube (n1). The Spectrum Value factor $\Gamma(n1)=1.0$ in cube (n1) and $\Gamma(n)=0.0$ in all the other cubes. Accordingly in volume(V),

$$SQI = \beta_{total} / (\beta_t(n1) + D_t(n1))$$

for the unweighted case, $\Gamma(n)=1.0$ in all the cubes, accordingly

$$SQI(\text{unweighted}) = \beta_{total} / \Sigma_{n=1}^{N} (\beta_t(n) + D_t(n))$$

Since $\Sigma_{n=1}^{N}(\beta_t(n) + D_t(n)) > (\beta_t(n1) + D_t(n1))$

therefore SQI > SQI (unweighted)

Accordingly, the inclusion of the "Spectrum Value" factor Γ into the calculation of SQI effectively reflects relative spectrum demand distribution in the area of interest. This model can therefore provide a very good indication of how effectively the spectrum is being managed to meet the demand and will be used in assessing spectrum utilization.

3.2.4 Application of SQI

The SQI can be used for absolute as well as comparative measures within a specific service. Absolute measures are used when all system parameters are known. The purpose of the absolute measure is to periodically assess how effectively the spectrum is being managed in a specific area and within a specific service. In other words, to compare the actual with the near optimum assignment strategy.

In the comparative measure, different schemes or techniques, such as spread spectrum vs FDMA/FM, digital vs analogue modulation, etc., can be compared for a specific service. It is not possible to use SQI to compare a technique used in different services as the models for the two services may be different.

3.2.5 Factors Affecting Spectrum Quality

There are a number of factors that can affect spectrum quality and the major classification is as follows:

- Propagation Characteristics
- Demand Distributions
- Available Technology
- Performance Requirements (Grade of Service)

3.3 Application of the spectrum quality indicator to the land mobile service

3.3.1 Introduction

In this section a spectrum quality indicator (SQI) is applied to the land mobile service and is used as an absolute measure of how efficiently the spectrum is being managed.

Absolute measures are used when all system parameters are known. In dispatch systems, the channel spacing, the modulation technique and the interference criteria are all known before hand. The purpose of the absolute measure is to periodically assess how efficiently the spectrum is being managed within a specific geographic area. In other words, to compare the actual with the near optimum assignment strategy [Delfour and Towaij, 1989].

3.3.2 The SQI land mobile model

The Optimum assignment is maximum number of frequency assignments at predetermined locations that can be made within a given geographic area without violating the defined interference criteria. The optimum conditions will be very difficult to obtain and will not provide any assignment location flexibility. A practical assignment strategy should not only take into consideration traffic demand distribution but also provide a reasonable assignment location flexibility. In the Canadian model, the NOAS strategy provided the maximum number of frequency assignments within the location flexibility of an assignment with an area of one km square [Delfour and Towaij 1989].

In Canada, there are two major types of users in the land mobile service, dispatch and public mobile telephone. The typical dispatch users are taxi, police, delivery, etc. Their systems consist of base stations and the associated mobiles operating within a specified coverage area. Depending on the number of mobiles, a dispatch service may have to share its frequency with other users.

The SQI model is applied in this section to the land mobile service with the following assumptions:

a) Future demands are likely to follow present traffic demographic distribution.

b) For the purpose of analysis, the geographic area of interest is divided into a grid of equal squares the size of which is determined by the interference criteria used in the band under consideration.

c) The time unit used is directly related to the average peak traffic hours loading.

d) The frequency unit used is the value of the bandwidth of a single radio channel used in the band under evaluation.

e) Spectrum unit demand within a square $\beta_t(n)$ is directly related to the total occupancy in Erlang $O_t(n)$ within that square.

$$\beta_t(n) = C\ O_t(n)$$

Spectrum units demanded in the (i)th channel in the (n)th square is approximated by

$$\beta(n,i) = C\ O(n,i)$$

where C is a constant determined by the size of the grid and the channel bandwidth utilized in the band under consideration.

The SQI was evaluated using equation (1) for both the weighted and unweighted cases. The marginal denied spectrum, $D_t(n)$, included unused communications capacity in Erlangs when the channel was not loaded up to the nominal capacity. Different loading factors were used for public safety services than for other services, and allowance was made when several systems shared the capacity. In addition, $D_t(n)$ included considerations of adjacent channel separation distances, intermodulation products, etc.

3.3.3 SQI results for major cities in Canada

Spectrum usage in the VHF land mobile bands were evaluated in Toronto, Montreal and Vancouver which are the three largest cities in Canada. Table 1 provide a general statistics on these three cities.

Table 1. VHF land mobile usage in Canada

CITY ITEM	TORONTO	MONTREAL	VANCOUVER
DIMENSION OF EVALUATION AREA (km x km)	75 x 75	75 x 75	75 x 75
TOTAL TRAFFIC DURING A PEAK HOUR (ERLANG)	175.98	226.79	171.81
ACTUAL NUMBER OF ASSIGNED CHANNELS	720	608	580
NUMBER OF NEAR OPTIMUM ASSIGNED CHANNELS	1066	1066	927
AVERAGE PEAK HOUR MESSAGE OCCUPANCY (ERLANG)	0.244	0.373	0.296
TOTAL NUMBER OF SHARED CHANNELS	333	337	348
TOTAL NUMBER OF UNSHARED CHANNELS	387	271	232
SQI*	0.518	0.652	0.441
SQI(MAXIMUM LOADING OF 0.6)	0.828	0.823	0.772

* A composed channel loading per channel of 0.60 Erlang was used in the calculation of SQI

3.4 Discussion

A model for calculating SQI was developed and applied to the land mobile radio service. This model employs the concept that a specific geographic area has a spectrum value relative to the total traffic demand in the area. The concept also points out that due to the three dimensional nature of the spectrum, volume, time and bandwidth, occupied spectrum units by a radio system will result in denied spectrum units to other users. The amount of denied spectrum is determined by level of interference other

systems will receive from operation in the vicinity of the radio system. The land mobile SQI mode utilizes actual distribution of demands as average peak hour channel loadings. In calculating the equivalent channel loading factor BL, percentage distribution of users (i.e. public safety, non public safety and special users) is to be determined and taken into consideration.

3.5 Conclusions

Based on the model described above the following observations can be made:

a) Spectrum quality is dominated by urban core communication traffic loading. Effectiveness of spectrum management can be measured by the maximum number of interference free frequencies available in the heaviest traffic cores of major cities.

b) When assigning frequencies outside heavy communication traffic urban cores, care must be taken to avoid preempting core assignments.

c) For new and replanned bands, the near optimum assignment strategy, that can provide the maximum number of interference free frequency assignments to meet the demands.

d) The near optimum assignment strategy can result in higher spectrum quality and at the same time simplify the assignment process due to the preselected interference free frequencies throughout the area of interest.

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REPORT 976

A SYSTEMATIC PLANNING METHOD FOR THE USE OF UNASSIGNED FREQUENCY BANDS

(Study Programme 18C/1)

(1986)

The method described in this Report has been used successfully for many years on a country-wide basis in the Federal Republic of Germany (approximately 400 000 frequency assignments have been handled) and recently, by six other central European countries in the 80, 160 and/or 460 MHz bands. Though it is now used only for the non-public land mobile service, it can be applied to many terrestrial services operating in the VHF/UHF frequency bands. This system was originally developed for the broadcasting service.

The method consists, in short, in the creation of a grid of rhombs (or hexagons) based on equilateral triangles and assigning to each rhomb (or hexagon) a number of frequencies (channels). The frequencies for each rhomb are selected in such a way that intermodulation products, and adjacent channel interference, are at a minimum. To fulfil these conditions, the transmitters should ideally be placed at the centre of the rhombs; in this ideal case, there would be a repeating distance of three rhombs between transmitters. Increasing off-centre positions will, however, reduce these conditions.

A rather common pattern used in the non-public land mobile service in the Federal Republic of Germany is a basic grid consisting of nine rhombs to which 37 sequential channels are allocated. To each rhomb, four channels are allotted (Figs. 1 and 2, the latter is an assignment model in the 900 MHz band). One channel, No. 34 in this case, cannot be assigned to any of the rhombs without increased risk of intermodulation interference. In practice, however, a way by which it can be assigned to one rhomb or another can very often be found.

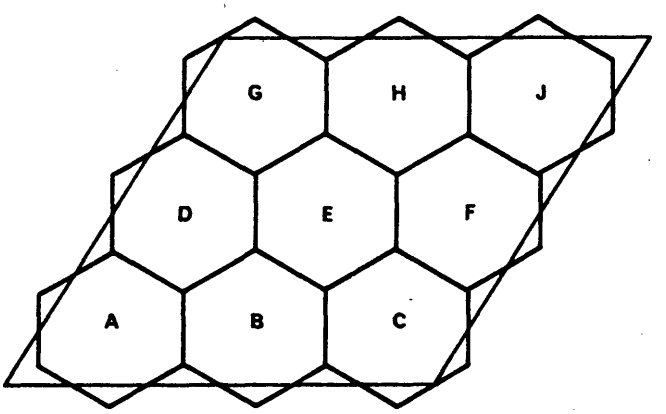


FIGURE 1 – *A basic grid for frequency allotment*

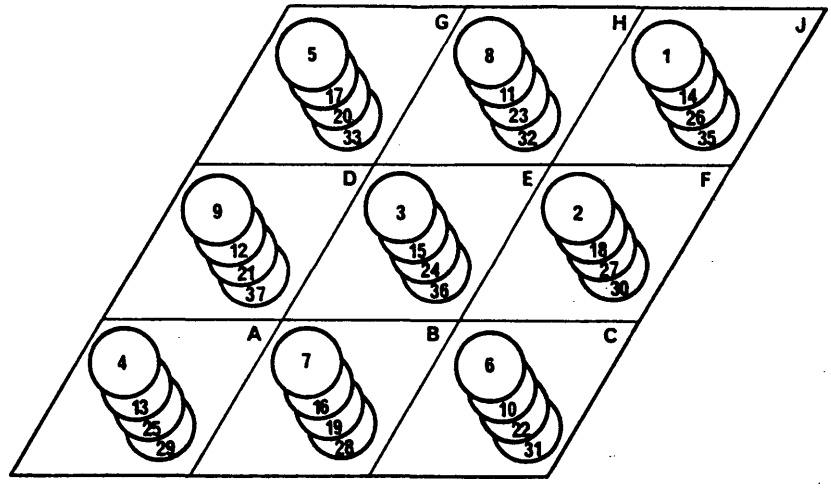


FIGURE 2 – *Frequency allotment plan for 37 channels*

The adherence of equipment to standardized technical parameters is a prerequisite for the application of this method. The basic pattern consisting of nine rhombs is repeated periodically so that there is, dependent upon the position of each transmitter in its rhomb, a minimum repeating distance of two rhombs between two transmitters using the same frequency, which is a minimum requirement for satisfactory service.

The considerations in this Report are of value to those services where only one frequency is assigned to a station. It is only applicable with adaptation and with certain limitations to the public land mobile service where mobile units traverse cells with automatic transfer of an ongoing call from one base station to another.

A major advantage is that the system allows one-time frequency planning in an economical way for large regions or for a country. Later frequency assignments could then easily be handled on a decentralized basis by local offices, by reference to the pre-planned grid.

The major limitation of this system is that it cannot be generally applied where the terrain or population distribution is uneven.

The use of a planning system similar to the one described can assist in international frequency coordination, if all the basic technical parameters are the same on both sides of the border (see Fig. 3).

Such a system may be applicable where portions of the VHF/UHF frequency bands are available for such assignment procedures. It can be developed and implemented without the assistance of a computer-based management system.

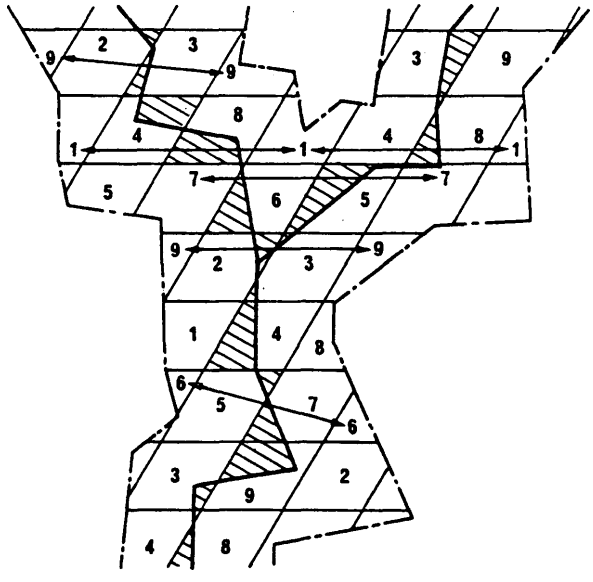


FIGURE 3 – *Result of frequency coordination in border areas for three-country border*

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REPORT 1113

A METHOD OF FREQUENCY-REASSIGNMENT TO IMPROVE CHANNEL UTILIZATION

(Study Programme 18C/1)

(1990)

1. Introduction

This report presents methods that are useful in the systematic engineering and realignment of land mobile frequency assignments in an area of highly concentrated communications where saturation of a particular frequency band is imminent or has occurred. The methods will apply whether saturation results from incremental growth of users or from uncontrolled growth where a complete frequency assignment method was not available or utilized. Application with propagation modifications to the 40-1000 MHz range and utilizing omni-directional antennas is possible. The method will encompass change over to an allotment plan, and usually results in satisfaction of new communication requirements by providing some un-used channels.

A seven step process is applied sequentially: (1) a spectrum utilization survey from current records, (2) an updated survey with the communicators, (3) a network loading analysis, (4) a grouping of networks, (5) re-engineering of the spectrum utilization for the area, (6) negotiation of specific frequency changes, and (7) the realignment plan implementation. Re-engineering includes re-capture of unused frequency assignments, resolution of duplicating and overlapping assignments, and consolidation of common purpose networks.

2. Goals

The primary goals are to relieve the frequency assignment congestion problem and to accommodate all known requirements by realigning a minimum of frequency assignments based on an engineering method and approach. The frequency assignment realignment is to be accomplished without wholesale equipment changes, so that the costs will be limited to crystal replacements.

3. Method

Seven major steps (and five required reports) have been established to properly consider all aspects of the realignment problem. Step one is to obtain a listing of frequency assignments from the national frequency assignment data system. Then, in step two, a detailed survey of each area is completed. The assignment listing is updated with new/deleted frequencies, new communications and use, equipment types, power, antenna heights, and receiver sensitivities. The survey of area users will prove valuable in later meetings required to advise communicators of the need and practical gains to be accomplished by the realignment project. At the end of this step, a Spectrum Utilization Report is produced, for approval by the communication personnel and for use in the next step.

The third step is to identify and separate those networks that are candidates and non-candidates for possible frequency change. Several realignment factors are applied to the architecture of each network to determine if frequency changes are possible while computer-aided electromagnetic radiation patterns are calculated and plotted for each network. The fourth step, using the data gathered or calculated for each network, separates candidate networks into similar groups, emphasizing mission importance, location, and purpose. Since exclusive use of channels is a luxury that can no longer be assumed, judicious frequency sharing must be planned for these similar network groups. The results of steps three and four are documented in a Frequency Analysis Report (FAR).

Step five includes the actual engineering of the FAR data to determine the load of candidate nets to be carried by each channel. Channel loads are selected to include the maximum number of networks and radio equipment that can effectively operate on the same channel at each location, based on the numbers of users and the frequency of use of these networks. A goal of 100 equipments is selected as the maximum target figure for effectively sharing the same channel at a given location although results of the analysis could reduce that number. Pertinent factors are message duration, occurrence of the busy hour, and presence of blocked message queuing. The realigned use of the band that results from this step is published in a Spectrum Utilization Plan.

The sixth step determines changes to the current assignments that are required to achieve the channel loading objectives, prepared as frequency change requests in the standard national format. These frequency change requests are published in a Frequency Realignment Plan (FRP). This plan has to be coordinated with communications and spectrum management specialists who address omissions, overloads, or errors. The seventh step addresses all last minute changes to the FRP, the re-engineering required, the plans, the possible communicator or contractor logistic problems in changing crystals, and the costs projected to arrive at a final Realignment Implementation Plan (RIP). After approval by proper authorities, the frequency realignment is accomplished as scheduled in the RIP.

4. Technical Discussion

4.1 Network Architecture

The network system architecture is a brief statement of the operational requirement in terms of the operating frequency(s), simplex or duplex usage, operating unit that is using the network, the purpose/use of the network, the allotment plan for the frequency(s) used, the station class and the number and types of equipments listed by station class.

4.2 Realignment Criteria

Some general realignment factors must be established prior to specific considerations of each network. Off-channel operations (not on allotment plan channels) are to be realigned on the nearest channel, except for communication networks not included in the analysis. Channel loading considerations result in the combination of low density-of-use networks to provide maximum sharing and to facilitate future modernization of equipment to tone-coded squelch systems. Channels are loaded with systems similar in occupied bandwidth, so that future allotment plans could possibly sub-allocate similar systems. Networks using fewer than 35 radios are considered candidates for realignment since consolidation of

these networks onto one frequency in the same or close geographical areas is used to provide adequate communication on an infrequent, aperiodic basis. Networks with many users in several locations are not considered for realignment, if they are on an alloted channel, since possible improvements are not adequate to offset the technical and economic costs.

Terrain shielding (masking) of transmissions is utilized to minimize interference. When a repeater station located on a mountain peak is involved, terrain masking does not generally apply, and repeater networks preclude the use of that frequency by other networks. Therefore, repeater stations are not changed except when there are less than 100 mobile radios with infrequent reply traffic density. Similarily, the area of coverage of a particular network may be greater than the stated required service volume, and the interference range of these networks could be reduced by lowering the antenna height, or by reducing the power output. These changes are made only when reduction of gross antenna height is necessary to attain channel-loading efficiency. Finally, all networks that experience Electromagnetic Interference (EMI) from border countries or domestic sources, as well as networks demonstrating potential for co-channel and adjacent-channel EMI with other users at a co-site, are candidates for change.

Realignment criteria for each network are derived from network importance and usage, as listed in Table 1, while realignment criteria for selecting candidate networks for frequency change are provided in Table 2, and criteria to prevent selection for frequency change are shown in Table 3. Each network is assigned a sequence number(s) which represents a selection or non-selection criteria. Due to budget constraints or other considerations, it is sometimes prudent to leave an assignment as is and change other small networks. In some cases, a network meets the selection criteria both as a candidate and as a non-candidate for change, and the non-candidate criteria prevails.

Table 1 Realignment Criteria

NETWORK IMPORTANCE	USAGE
I. Emergency Services II. Police or Guards III. Industrial Control IV. Research or Testing V. Administrative Services	A. 1-4 Hours Daily B. 4-8 Hours Daily C. 8-16 Hours Daily D. Continuous Use E. Aperiodic Infrequent Use

Table 2 Candidate Network Selection Criteria

SEQUENCE NUMBER	CRITERIA
C1	Networks operating on off-channel frequencies; networks with operating frequency not in compliance with the allotment plan.
C2	Networks consisting of 35 radios or less.
C3	Networks creating or receiving disruptive EMI to/from another co-channel or adjacent-channel user at the same or different locations. Includes co-site frequency separation problems as well as EMI from border country sources.
C4	Duplicate networks on multiple channels (i.e., two or more channels used by the same communicators, for the same purpose at the same location and usually associated with multiple channel radios).
C5	Common-purpose networks operating on discrete channel. A network which is not operating on the same common-purpose channel used by similar networks in the local area (i.e., a network operating on a discrete channel when other common-purpose channels are available for shared use).
C6	Near-term requirement. A communication need that is justified but has not been fulfilled due to the lack of an available channel for assignment.
C7	Deletion. Move the assignment to an appropriate band.

Table 3 Non-Candidate Network Selection Criteria

SEQUENCE NUMBER	CRITERIA
NC1	Networks operating on channels not included in the analysis. All net-on others' channels are not considered for realignment regardless of number of radios involved. Exception. Networks operating on off-channel frequencies in the other portions of the band are selected as candidates for realignment.
NC2	Networks located 145 to 160km from the area of concern. All networks operated from these area are not considered for realignment since they do not present a co-channel EMI threat to/from each other. Exception. Networks experiencing disruptive EMI from border country sources operating on off-channel frequencies are selected as candidates for realignment.
NC3	Simplex networks consisting of 36 or more radios. A single network consisting of 36 or more radios is not considered for realignment.
NC4	Paging system networks. Paging systems consisting of 36 or more radios are not considered for realignment.
NC5	Repeater system networks. Repeater networks consisting of 36 or more radios are not considered for realignment.
NC6	Common-purpose/use networks using same channel. Multiple networks using the same channel for the same common-purpose/use in the same area are not considered for realignment if justified by the importance or if the total number of radios per channel exceeds desired channel-loading objectives of the FRP.
NC7	Remote control/sensing systems. Networks used for command data links, such as fire/intrusion alarm sensors, targets, well pumps, energy monitoring/control, or other remote control/sensing devices are not considered for realignment. Such devices use coded bit streams of short duration only on an aperiodic basis, and present minimal EMI threat to/from co-channel users at the same location.
NC8	Special-purpose research or test systems, used for sending data from range timing systems or radiolocation systems are not considered for realignment based on the unique nature of the system and the prohibitive cost of frequency changes.

4.3 Electromagnetic Radiation Patterns

Three electromagnetic radiation patterns are compiled by computer for each network. The first pattern computed for each network is the effective mobile operating range or service volume talk-back range which defines the practical operating range of the mobile stations. The equation, used to obtain the 90% coverage area that is required to sustain two-way communications between the base station and the mobile stations, is based on the transmit power of the mobile station and the receiver sensitivity of the base station. The second pattern is the maximum range of the base station or talk-out range wherein marginal two-way communications or one-way paging is sustained at 90% coverage between the base station and the mobile stations. This pattern depends on the base station transmit power and the mobile station receiver sensitivity. To predict radii of the worst case interference range, the third pattern uses the potential interference range or base station talk-out range at 10% coverage depending on the transmit power of the base station or repeater station and the victim receiver sensitivity (highest anticipated level). The interference margin was defined as the distance between patterns formed by the radii bounding the range of potential interference of two stations operating in a co-channel assignment. Figure 1 illustrates an interference margin, and also depicts the three calculated ranges.

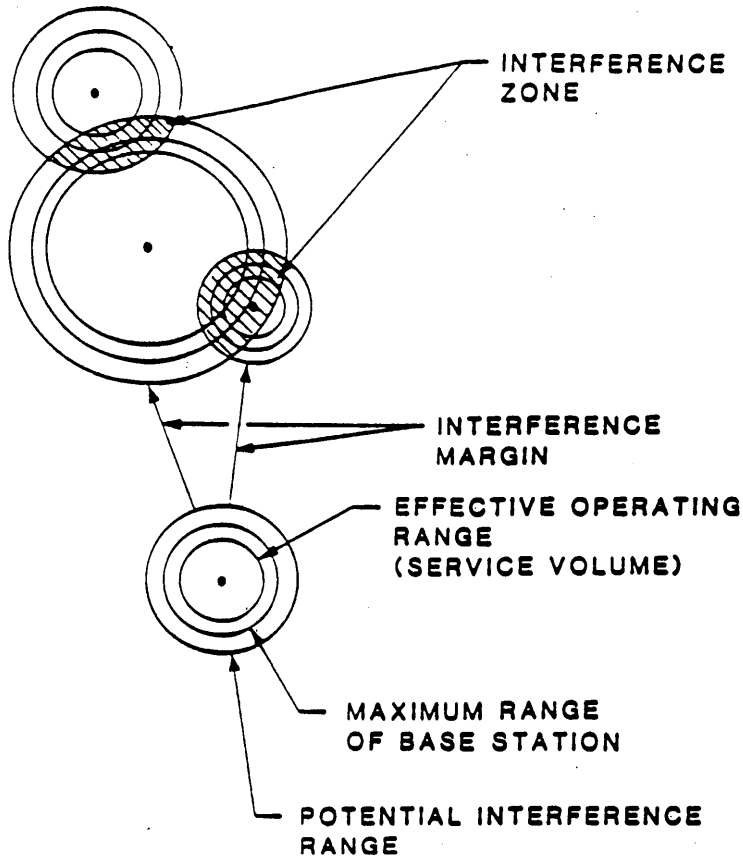


Figure 1. Sample Co-channel Radiation Patterns

These three electromagnetic radiation patterns provide the worst-case situation to be expected over irregular terrain, although hilly to mountainous terrain would attenuate potential unwanted signals even further. The interference margins and the three ranges for each operating network are used as the basis for determining the channel-packing density. For Simplex networks, all three distances are calculated. For Duplex networks, the repeater needs no talk back range, while the mobile unit requires no talk out range.

A mathematical model by Egli (1957) has been developed to quantify irregular terrain loss, and all three computer-generated electromagnetic radiation ranges are calculated by use of the Egli formula for the median received power over irregular terrain, between half-wave dipoles:

$$p_r = 0.345\left[\frac{h_t\, h_r}{d^2}\right]^2 \left[\frac{40}{f}\right]^2 p_t \times 10^{-14} \qquad (1)$$

where:

p_r is median received power, watts
h_t is transmitting antenna height, feet
h_r is receiving antenna height, feet
d is distance from transmitter, miles
f is frequency, $40 \leq f \leq 1000$ MHz
p_t is transmitted power, watts

Rearranging to obtain the median propagation loss with half-wave dipoles, l_{ph}, gives:

$$l_{Ph} = P_t/P_r = \left[0.345\left[\frac{h_t\ h_r}{d^2}\right]^2\left[\frac{40}{f}\right]^2 \times 10^{-14}\right]^{-1} \quad (2)$$

Converting to decibel format,

$$L_p = 112.6 - 20 \log h_t h_r + 40 \log d + 20 \log f \quad (3)$$

Since the Egli model used two half-wave dipoles, for isotropic antennas the actual loss is greater by twice 2.2 dBi = 4.4 dBi, giving:

$$L_p = 117 - 20 \log h_t h_r + 40 \log d + 20 \log f \quad (4)$$

From general field theory, however,

$$P_r = P_t + G_t + G_r - L_p \quad (5)$$

Where P_t is transmitter power in dBm
G_t is gain of the transmitter antenna in dBi
G_r is gain of the receiver antenna in dBi
L_p is propagation loss in dB

Substituting (4) into (5),

$$P_r = P_t + G_t + G_r - 117 + 20 \log h_t h_r - 40 \log d - 20 \log f \quad (6)$$

Received power, P_r, can also be specified in terms of receiver sensitivity and a protection margin specified for a particular receiver. Thus,

$$P_r = R_s + S/I \quad (7)$$

Where R_s is the receiver sensitivity in dBm
S/I is the protection margin in dB

Substituting equation (7) into equation (6) makes it possible to solve for d, the median distance from the transmitter (which is either the effective median mobile range, the median base station range, or the potential median interference range, depending on the system parameter selected).

$$d = \log^{-1}\left[\frac{P_t + G_t + G_r - 117 - S/I + 20\log (h_t h_r/f) - R_s}{40}\right] \quad (8)$$

Also from Egli, for the cases of 90% coverage area and 10% coverage area utilized for the three radiation patterns, a variable terrain correction factor, $K_\%$ in dB, can be introduced into (8), where:

$$K_{10\%} = + 8 + 6 \log (f/40) \quad (9)$$

$$K_{90\%} = - K_{10\%} \quad (10)$$

Thus, the final equation to be programmed for use in the many repetitive distance calculations is:

$$d_{\%} = \log^{-1}\left[\frac{P_t + G_t + G_r - 117 + K_{\%} - S/I + 20\log(h_t h_r/f) - R_s}{40}\right] \quad (11)$$

4.4 Network Grouping

The fourth step uses realignment criteria (network importance and usage), candidate/non-candidate selections, radiation patterns, and radiation distances to separate the networks into the prioritized groupings of Table 4.

Table 4 - Network Groupings

Group	Description
(a)	Critical, highly visible, high-priority networks.
(b)	Location. All networks used at the same location.
(c)	Purpose/Use. All networks at the same location used for the same purpose by any user.
(d)	All networks at the same location used by the same organization.
(e)	Specialized Purpose/Use. All networks used at the same location for a specialized purpose.
(f)	High EMI Potential. All networks with a current configuration that creates a high potential for EMI to/from other users of the same channel at other locations.
(g)	Low EMI Potential. All networks with a current configuration that presents little or no EMI threat to/from other users of the same channel at other locations.

4.5 Realignment Planning

The fifth step realigns the networks to the available channel resources so that the maximum number of networks and radio equipments effectively operate on the same channel at each location, based on a goal of 100 subscribers. A series of matrices depicting frequency channel versus operating organization are developed to facilitate the change planning.

The working matrices use various color-coding schemes to indicate architectural constraints, groupings, realignment candidacy, number of users, and network importance. Channel loading becomes more apparent and trial changes are made and evaluated using the matrices, achieving the maximum number of cleared channels through judicious loading while minimizing costs.

5.0 Example of Realignment Analysis

An example has been partially described by Dobberteen [1988], and analysis results are provided in Table 5. The resultant cleared channels at each location support the concept that a logical and well engineered approach to frequency realignment can provide relief to congested areas. Although action through channels is still in progress to coordinate new frequencies with border governments, each area has some clear channels available for future growth. Even more channels with partial loading are available at each location.

Table 5 Analysis Results

Area	Total Networks	Cleared Channels	Changes
A	400	20	2,742 of 19,387 systems
B	110	38	981 of 6,193 systems
C	250	30	900 of 6,000 systems
D	132	22	12 frequencies

By keeping the number of frequency assignment changes to a minimum, the costs for each operating organization are also minimized. No users are forced to purchase new radios, but as age takes its toll, new systems with tone-coded squelch addressing will be justified by the occasional irritating traffic from other users of the same channel.

Because the frequency changes recommended by the realignment plan are carefully engineered using known parameters, proposed changes in system configuration must be submitted to the frequency coordinator for coordination, review, and approval.

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REPORT 662-3*

DEFINITION OF SPECTRUM USE AND EFFICIENCY

(Question 47/1)

(1978-1982-1986-1990)

1. Introduction

Question 47/1 calls for methods of describing "spectrum utilization", and for a definition and methods of calculation of "efficiency of spectrum use". This Report provides preliminary general definitions of these two terms, and illustrates their application with examples in the Annexes.

2. General definitions

First, the relationship of a measure of spectrum utilization to a measure of spectrum utilization efficiency is stated. A measure of spectrum utilization merely shows how much of the spectrum-space has been used (just as, for example, "occupied bandwidth" is a measure of how much radio bandwidth has been used). By itself, it tells nothing about the efficiency of use. To determine this, the amount of information carried must be known. Spectrum efficiency is the ratio of the information transferred to the amount of spectrum-space used. It can be computed for a specific system or for the aggregate of all systems in a frequency band and region.

2.1 *Measure of spectrum utilization*

Efficient use of the spectrum is achieved by (among other things) the isolation obtained from antenna directivity, geographical spacing, frequency sharing, or orthogonal frequency use and time-sharing or time division (Report 660) and these considerations should be reflected in any definition of spectrum utilization. Therefore, the measure of spectrum utilization is defined to be the product of the frequency bandwidth, the geometric (geographic) space, and the time denied to other potential users: the measure is $U = B \cdot S \cdot T$, where B is the radio bandwidth, S is the geometric space (usually area) and T is time.

The geometric space of interest may also be a volume, a line (e.g. the geostationary orbit), or an angular sector around a point, as it is in the application in § 3. The amount of space denied depends on the spectral power density, as shown in the Annexes. For many applications, the dimension of time can be ignored, because the service operates continuously. But in some services, for example, broadcast and single channel mobile, the time factor is important to sharing and all three factors should be considered simultaneously, and optimized.

Spectrum use measures which are the product of bandwidth, geometric space, and time have been proposed in the literature by [Gifford, 1966; Joint Technical Advisory Committee, 1968; Ewing and Berry, 1973; Colavito, 1974; Luk'yanova, 1974] and [Vinogradov, 1974 (without time)]. Spectrum use measures which include the effect of spurious emissions have been proposed by Vinogradov and Vinogradov [1976 and 1979].

The measure of spectrum use may be computed by multiplication of a bandwidth bounding the emission (e.g., occupied bandwidth) and its interference area, or may take into account the actual shape of the power density spectrum of the emission and the antenna radiation characteristics as is shown by Berry [1977].

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* Administrations, as well as other Study Groups of the CCIR, are requested to transmit contributions and comments on this Report to Study Group 1.

Traditionally, radio transmitters have been considered the users of the spectrum resource. They use the spectrum-space by filling some portion of it with radio power – so much power that receivers of other systems cannot operate in certain locations, times and frequencies because of unacceptable interference. Notice that the transmitter denies the space to *receivers* only. The mere fact that the space contains power in no way prevents another transmitter from emitting power into the same location; that is, the transmitter does not deny operation of another transmitter.

Receivers use spectrum-space because they deny it to transmitters. The mere physical operation of the receiver interferes with no one (except as it inadvertently acts as a transmitter or power source). Even then the space used physically is relatively small. However, the authorities deny licenses to transmitters in an attempt to guarantee interference-free reception. The protection may be in space (separation distance, coordination distance), in frequency (guard bands) or even in time (in the United States of America, some MF broadcasting stations are limited to daylight operation). This denial constitutes "use" of the space by the receiver. The radio astronomy bands are a familiar example of the recognition of receiver use of the spectrum space.

One way to incorporate these facts into a unit of measure of spectrum space is to partition the resource into two spaces – the transmitter space and receiver space – and define dual units to measure the usage of each space. Where simplicity is most important, the two units can be recombined into a single measure of system use. Ewing and Berry [1973] and Berry [1977] provide such definitions. Luk'yanova [1974], Vinogradov [1974, 1980, 1982], Vinogradov and Vinogradov [1976, 1979], Dodo *et al.* [1980] and Moreno [1982] give detailed definitions and example calculations for specific situations.

If the general concepts of these definitions are accepted, much work is still needed to determine precisely how the boundaries of the volume shall be specified, and how its values are to be computed when two areas overlap (occupy the same space).

2.2 *Measure of spectrum utilization efficiency*

Efficiency is defined in many diverse fields as the ratio of output to input. The output may not be the total output of the system, but merely the output of immediate interest. Similarly, the input used in the definition may be only one of the inputs – the one of special interest. These ideas are used to define spectrum efficiency. The desired output of a telecommunications system is "information transferred over a distance", and the input of interest to spectrum efficiency is spectrum-space used, that is, the space denied to others. So the general form of a measure of spectrum efficiency is:

$$\frac{\text{information transferred over a distance}}{\text{spectrum space used}} \quad \text{or} \quad \frac{\text{communications achieved}}{\text{(bandwidth)} \cdot \text{(space)} \cdot \text{(time)}}$$

Considerable work is needed to transform this general concept into calculated values for each particular service as illustrated by the examples in Annexes I, II and III and in the References.

Another attempt to create a unified approach, based upon this general concept, was undertaken by [Ramadanović, 1986].

Where the configuration of the radio (broadcasting) network or the radio installation siting plan is known and the interference situation can thus be analyzed, another criterion for evaluating spectrum utilization efficiency can be used to advantage [Bykhovsky, 1979; Bykhovsky and Pavliouk, 1986; Bykhovsky and Pavliouk, 1987]. This criterion is based on a comparison of the frequency bandwidth (F_c) required to transmit a given volume of information (i.e. to provide a given number of communication or broadcasting channels) in a

real network with real radio installations with a frequency bandwidth (F_{opt}) of an idealized radio system with the same communication capacity. Such a system is understood to be a regular broadcasting network (Report 944) or an equivalent network for other radio services in which use is made of optimum (in the Shannon sense) communication systems which in addition possess ideal characteristics (from the point of view of unwanted transmitter radiation, non-fundamental receiver reception channels, antenna parameters, etc.). The criterion for spectrum utilization efficiency (μ) is defined by the expression

$$\mu = F_{opt}/F_c .$$

An optimum communication system is characterized by the highest gain in the signal/noise ratios which can be obtained at the output (ρ_o) and the input (ρ_s) of the receiver by increasing the frequency bandwidth of the modulated signal (F_m) in comparison with that of the wanted message (F_o). The magnitudes ρ_s and ρ_o are linked by the following relation (see Annex IV):

$$\rho_s = (1 + \rho_o)^{Fo/Fm} - 1.$$

An example of the evaluation of spectrum utilization efficiency in radiocommunication and broadcasting networks is given in Annex IV. As will be seen from this annex, the criterion given is used to optimize the networks and also the parameters used in these radio networks, with reference to the limit values provided by an idealized radio system.

ANNEX I

CRITERIA FOR THE INTEGRATED EVALUATION OF SPECTRUM OCCUPANCY AND UTILIZATION EFFICIENCY

According to § 2.2 of this Report, spectrum utilization efficiency (E) of a radio communication system can be expressed by:

$$E = \frac{M}{\Delta f \cdot V} \qquad (1)$$

where

M: useful result obtained from the radio equipment considered,

Δf: frequency spectrum occupied,

V: volume of space occupied.

Depending on the purpose of the radio system and on the standpoint from which the efficiency of spectrum use is evaluated, M may be expressed in physical terms (channel-kilometres, dimensions of service areas, etc.) or by equivalent indicators (transmission capacity in binary units versus distance, financial profits, etc.).

Radio systems are evaluated by applying the E criterion after they have been arranged in groups, each of which has the same M value.

If there is a demand for the simultaneous use of the same bandwidth by a number of users in conditions in which it is difficult to ensure electromagnetic compatibility, the order of priority of the systems should be considered as suggested by Vinogradov [1980].

The volume of space occupied (V) can be considered by taking as an example the interference caused on a link of length L_c in free space between a transmitter station (A) and a receiver station (B), which form an independent system, by another equally independent system with stations A′ and B′ operating on a line of the same length L_c (Fig. 1).

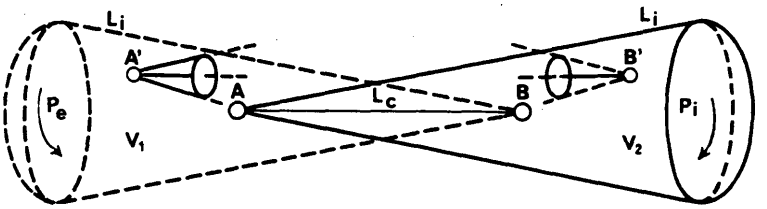


FIGURE 1 - *Representation of space occupied volumes*

This figure shows the areas of space covered by the radiation patterns of the transmitter and receiver antennas. They are approximated by spherical sectors. The solid line show the boundaries of the interference space of volume V_1 determined by the level of the maximum permissible noise power P_i which may be caused on B′ by A. The broken lines show the boundaries of the volume V_2' determined by the equivalent power of emissions from station A′ ($P_e = P'G_1$, where P' = power radiated in the necessary bandwidth, G_1 = transmitting antenna gain), which may cause interference P_i at station B. The two sectors have the same radius L_i.

A station B′, of the same type as B, and belonging to a different link, cannot be sited within the solid line contour whenever A may be located within the limits of the radiation pattern of B′, since A will cause to B′ interference in excess of the permissible level. Likewise, a station A′ of the same type as A and belonging to a different link cannot be sited within the broken line contour whenever B may be within the zone of the radiation pattern of antenna A′.

Outside these contours, there are no restrictions on the siting of transmitter and receiver stations for reasons of mutual interference.

Expressing V by the parameters of transmitting and receiving equipment [Vinogradov, 1974; Vinogradov and Vinogradov, 1979] the spectrum efficiency indicator E can be expressed by the formula:

$$E = \frac{M}{F_m}\left(1 + \frac{F_{st}}{F_m} + \frac{F_{sr}}{F_m}\right)^{-1} \tag{2}$$

As the measure M, it is convenient to take the maximum capacity of the radio channel in bit/s, which can be expressed by Shannon's formula taking generally into account the extent of the radio system served (volume, area, length of channel, etc.).

$$M = L_c \Delta F_o \log_2\left(1 + \frac{P_{so}}{P_{no}}\right) \tag{3}$$

where:

ΔF_o: the operating radio channel frequency band at the radio system output,

P_{so}/P_{no}: the signal-to-noise ratio in the band ΔF_o.

In the case of free space propagation:

$$L_c = L_{cmax} \left(\frac{P_n}{P_n + xP_i} \right)^{1/2} \quad (4)$$

where:

L_{cmax}: maximum possible circuit length between transmitter and receiver at a given quality (of reception) and in the absence of external interference ($P_i = 0$);

P_i: maximum permissible power of interference from external sources at the receiver input;

P_n: internal noise power referred to the receiver input;

x: coefficient of transformation of the various forms of spectra P_i to interference with a continuous uniform spectrum similar to P_n, allowing for the difference (compared with P_n) in the degree of the harmful effect of P_i on the required ratio P_{so}/P_{no}.

F_m, F_{st} and F_{sr} are the aggregate (frequency-space) values of spectrum utilization by the main emitting and receiving channel, by spurious emissions of the transmitting station and spurious reception of the receiving station respectively.

The formulae for the determination of F_m, F_{st} and F_{sr} are determined, on the assumption that the volume occupied by the radio equipment has the same value of radio-wave attenuation as of free space. These formulae in their integrated form represent the many characteristics that determine the conditions for the electromagnetic compatibility of radio equipments [Vinogradov, 1974, 1980 and 1982; Vinogradov and Vinogradov, 1976 and 1979].

The maximum distance to the edge of the interference zone in free space propagation conditions is expressed by the formula:

$$L_i = L_{cmax} \left(q \frac{P_n}{P_i} \right)^{1/2}$$

where:

q: ratio, required for normal channel operation, of the radio signal power (P_s) to the aggregate power of internal noise (P_n) and outside noise (P_i) at the receiver input.

In planning the locations of radio systems, acceptable values of P_n/P_i can be selected on the basis of the relations of L_c/L_{cmax} and L_c/L_i to P_i/P_n.

Conclusion

(a) The formulae for the values E, M, F_m, F_{st} and F_{sr} allow assessment of the efficiency of radio systems using frequencies related to each other and show the extent of the effect produced by each of the many indicators commonly used to determine that efficiency.

(b) Additional parameters are needed to define F_{st}. Some of the factors upon which F_{st} depends are, L_{cmax}, the spurious and out-of-band emissions of the transmitter, the transmitting antenna main lobe and side-lobe sharpness at spurious emitting frequencies, the power of the spurious emissions, and parameters relating to signal and natural source power flux-densities.

For the planning of radio system networks and the assignment of frequencies, the values of L_{cmax}, L_i (the length of the interference path), P_n/P_i, χ (a measure of the sensitivity of the radio system to interference) and cost factors may serve as guidelines. Additional work on this subject of the spectrum utilization efficiency indicator is needed.

(c) The following remain to be determined: the characteristic temperatures of radio interference resulting from all the natural sources acting simultaneously on the isotropic receiving antenna; the characteristic power of the wanted and unwanted emissions caused by radio equipments throughout the band. These factors as well as others can be taken as standard in assessing the efficiency of radio systems in the use of radio frequencies.

ANNEX II

SPECTRUM USE EFFICIENCY IN THE TERRESTRIAL POINT-TO-POINT RADIO-RELAY SYSTEM

1. Introduction

For radio-relay systems that operate continuously, the dimension of time may be ignored. Referring to § 2.2 of this Report, the spectrum utilization efficiency, E, can be written as:

$$E = \frac{C}{B \cdot S} \tag{5}$$

where:

C: measure of communications capacity, for example telephone channels or bits/s;

B: required radio bandwidth;

S: geometric measure, for example, area, or the angle between branching links at a node.

2. Spectrum utilization efficiency for a long artery with branching links at the nodes

Normalized communication capacity, η, which gives the spectrum use efficiency for the terrestrial point-to-point radio-relay system, is defined as:

$$\eta = \frac{M \cdot A}{B} \tag{6}$$

where,

A: transmitting capacity (e.g., number of telephone channels) per radio channel,

B: required radio bandwidth per radio channel,

M: allowable number of branching links (that is, two-way radio routes) for one repeater station.

This formula is derived by expanding the one given by Tillotson *et al.*, [1973] to include the geometric measure, M (M depends on the allowable angle between branching links).

Spectrum use efficiency in the terrestrial point-to-point radio-relay system was calculated for telephone transmission using the above formula.

The assumptions used are:

- telephone signal is transmitted;
- probability of fading is the same as that given in Report 338;
- circuit length is 2500 km; and circuit model is as shown in Fig. 2;

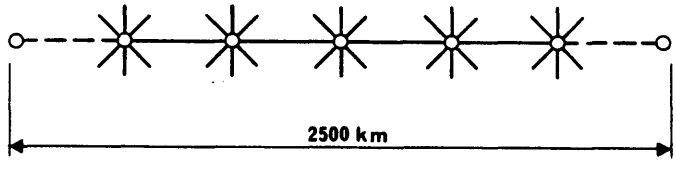


FIGURE 2 – *Circuit model*

- required carrier-to-noise ratio (C/N) is expressed as:

$$C/N = 10 \log [(2^n - 1)/3] + 11.8 \quad \text{dB}$$

where n is n-state QAM;

- one tenth of the overall radio-relay circuit noise for the 2500 km circuit is assigned as the interference noise from other routes;
- interference from other routes has the same frequency as the wanted signal;
- a reference antenna diagram for a circular antenna in Report 614 and a dual offset tri-reflector antenna used in Japan for a digital microwave radio, as shown in Fig. 3, are used;
- links with random branching angles.

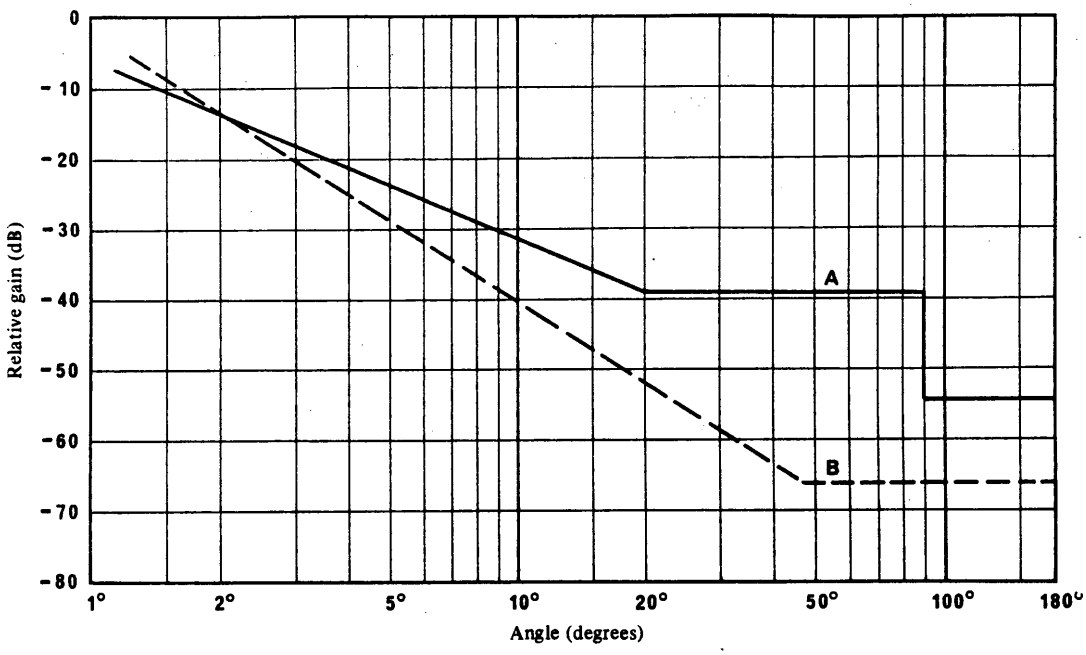


FIGURE 3 – *Antenna pattern*

A: preliminary CCIR reference pattern
B: offset tri-reflector antenna pattern

The normalized communication capacities for these two types of antenna were calculated and are shown in Fig. 4. The performance of the circular antenna in Report 614 is insufficient to estimate the spectrum use efficiency of high-level modulation systems. As the results depend on antenna performance, if a high performance antenna can be used, higher level modulation such as 256 QAM is effective [CCIR, 1982-86].

3. Spectrum utilization efficiency in randomly arranged radio-relay links

3.1 *Formulation*

Figure 5 shows a radio-relay link X-Y with another radio station (Z) operating on the same frequency. Station Z is randomly located on a circle around station Y.

Station (Y) receives a desired signal of frequency f_1 from station (X). Station (Z) transmits a signal of the same frequency f_1 in an arbitrary direction.

The normalized communication capacity, η, which gives the spectrum utilization efficiency, is defined as:

$$\eta = \frac{M \cdot A}{B} = \frac{\overline{p}}{p} \cdot \frac{A}{B} \tag{7}$$

where:

A: transmitting capacity per radio channel,

B: required RF bandwidth per radio channel,

M: number of radio links possible using the same frequency $\approx \overline{p}/p$.

The probability p that station (Y) receives interference exceeding the acceptable limit is calculated by considering the combination of the antenna pattern of stations (Y) and (Z) [Dodo *et al.*, 1980] and $\overline{p}$ is the maximum permissible probability of interference.

As the accumulation of interference from two or more stations has been neglected, some margin should be provided in any actual application.

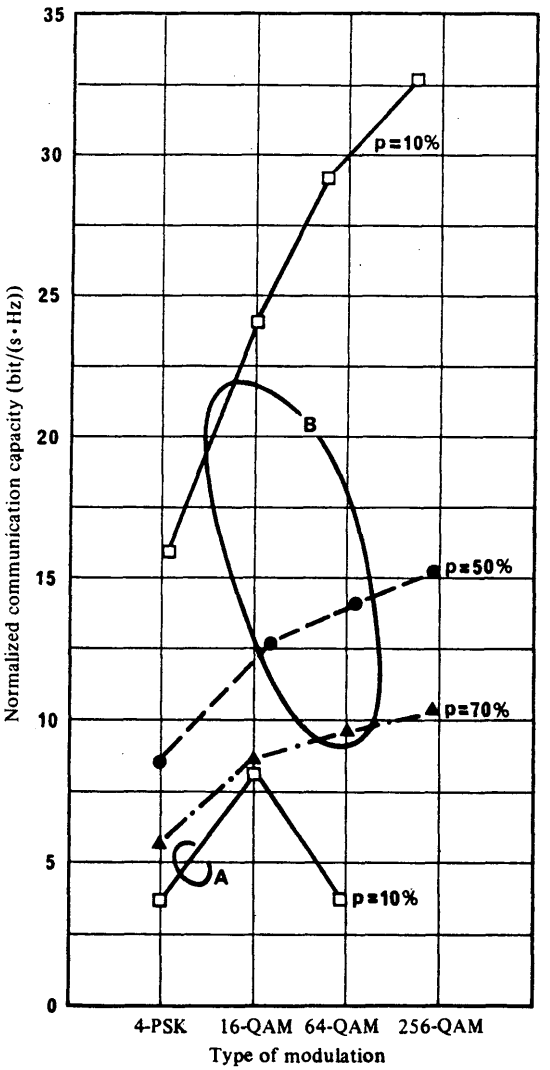


FIGURE 4 – *Normalized communication capacity*

Fading margin = 20 dB
Repeater spacing = 50 km
Channel capacity = 64 kbit/s
Allotment to branch noise = 10%

p: probability of interference
A: preliminary CCIR reference pattern
B: offset tri-reflector antenna pattern

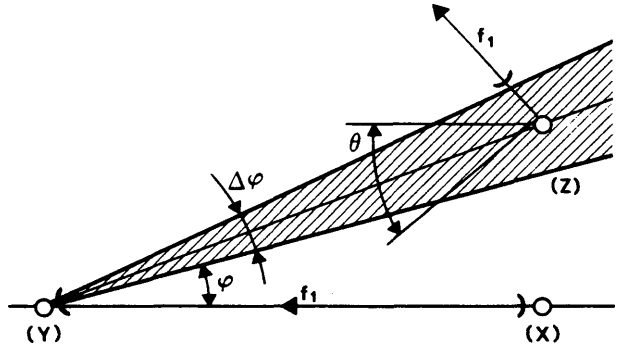


FIGURE 5 – *Random layout*

○ : station

3.2 *Application: spectrum efficiency in 2 GHz band radio-relay systems*

The spectrum utilization efficiency for a small-capacity terrestrial point-to-point radio-relay system operating in the 2 GHz band was calculated for telephone transmission using the above formula.

The relative spectrum utilization efficiency for 1.8 m diameter antennas was calculated using the permissible interference ratio and corresponding efficiency for each type of modulation in Table I. The results are shown in Fig. 6.

The digital system is superior to the analogue system for smaller fading margins. In this study, the attenuation due to fading is the same as the degradation of W/U (wanted signal level to unwanted signal level ratio) caused by interference. If space-diversity techniques are used, the necessary fading margin is lower. In general, digital systems tend to deliver superior spectrum utilization efficiency.

For digital modulation, a change from 2-phase to multi-phase or multi-state requires less bandwidth, but it may have lower spectrum utilization efficiency when interference is high. The exact value depends on the antenna characteristics, etc. but the 4-phase PSK system may be optimum from the macroscopic viewpoint in cases where other radio links operating around the repeater station are randomly located in an area.

3.3 *Spectrum utilization efficiency in a random mesh network*

A detailed model of spectrum utilization efficiency in an area is discussed in [Moreno, 1982], where it is shown that spectrum efficiency can be expressed as suggested in § 2.2.

In order to perform a fair comparison of modulation techniques, one can assume an interleaved frequency plan with a channel spacing corresponding to a given performance degradation caused by adjacent channel interferences. Table II gives tentative values of the normalized channel spacing X (defined in Report 608, Kyoto, 1978) and the corresponding spectrum efficiency in bit/(s · Hz). Even if different results could be derived, based on other assumptions, it should be noted that the calculated results of Table II [Moreno, 1982] are quite near the values which could be derived from specific channel arrangements, as suggested by CCIR Recommendations (for example 140 Mbit/s, with 16-QAM modulation and 40 MHz channel spacing between cross-polarized channels). Measured values might be different from these calculated values.

TABLE I - *Parameters of various modulation types in the 2 GHz band*

Modulation type		Permissible *S/N* or error ratio	Interference reduction factor (IRF)		Permissible wanted signal/ unwanted signal ratio *W/U*	Parameters related to *B*		Spacing to adjacent channels *B*	Number of channels *A*	*A/B*(1) (channels/kHz)
Analogue transmission	FM	58 dB	20 dB		38 dB	Frequency deviation for test tone: 100 kHz r.m.s.		520 kHz	24	0.046
	SSB	58 dB	9.5 dB		48.5 dB	Highest baseband frequency: 108 kHz Filter coefficient: × 2 Frequency tolerance: 20 kHz		236 kHz	24	0.1
Digital transmission			(*C/N*)	(Degradation)		Clock frequency	Filter coefficient			
	2-phase PSK	10^{-6}	10.7 dB	5.5 dB	16.2 dB	1544 kHz	× 1.3	2 MHz	24	0.012
	4-phase PSK	10^{-6}	13.7 dB	5.5 dB	19.2 dB	772 kHz	× 1.4	1.1 MHz	24	0.022
	8-phase PSK	10^{-6}	19.1 dB	5.5 dB	24.6 dB	515 kHz	× 1.5	0.77 MHz	24	0.031
	QPRS	10^{-6}	16.8 dB	5.5 dB	22.3 dB	772 kHz	× 1.1	0.85 MHz	24	0.028
	16-QAM	10^{-6}	21.4 dB	5.5 dB	26.9 dB	386 kHz	× 1.6	0.62 MHz	24	0.039

(1) The proper efficiency for each type of modulation

The assumptions used are:

— acceptable interference and spectrum efficiency for each modulation type are as shown in Table I. Eighty per cent of the total circuit noise is allotted to interference;

— distances between a station subject to interference (station Y) and the interfering stations are assumed to be the same; this assumption is considered to cause little error in efficiency calculation since the free-space losses of two links differ by only 6 dB even if they differ in length by a factor of two;

— fading in the wanted signal and in the interfering signals is assumed to have no correlation;

— the antenna radiation pattern is the reference diagram in Report 614;

— all stations have the same transmitting output power;

— the limit on the probability of interference, $\bar{p} = 0.1$.

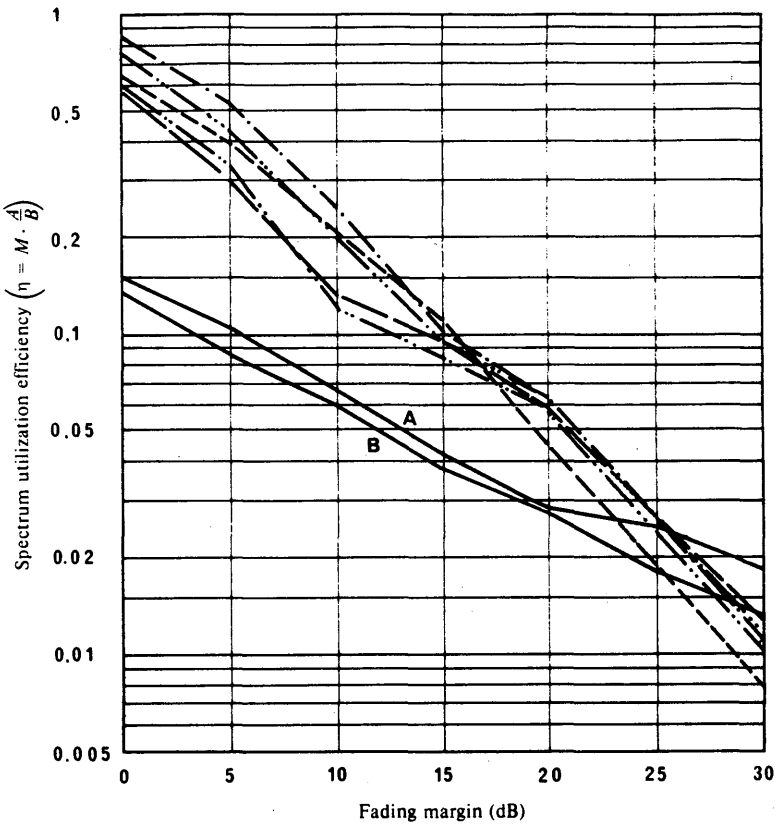


FIGURE 6 – *Spectrum utilization efficiency of random layout*

Frequency band: 2 GHz
Capacity: 24 Channels

—·— : 4-PSK (4 phase - phase-shift keying)
—··— : 8-PSK
– – – : 2-PSK
—···— : QPRS (quadrature partial response modulation)
—— : 16-QAM (16-state quadrature amplitude modulation)
—— **A** : SSB (single-sideband amplitude modulation)
—— **B** : FM (frequency modulation)

TABLE II

Modulation method	Normalized channel spacing X	Spectrum efficiency bit/(s · Hz)
4-PSK	1.88	2.13
8-PSK	2.16	2.77
16-QAM	2.23	3.59

Notes. - Degradation due to adjacent channel interference: 0.5 dB.
Channel filters: raised cosine roll - off 0.5.
Decoupling between cross - polarized channels (residual XPD): 12 dB.

The antenna radiation pattern used in the analysis is shown in Fig. 7; it is for a typical parabolic antenna. Moreno [1982] shows in detail how to compute the area using the concepts from this Report. It has been assumed that performance degradation (and a bit error ratio of 10^{-3}) due to co-channel interference from other links is not greater than 1 dB. It is assumed that the interfered-with link is at the threshold, with 40 dB fade margin, while the interfering link is receiving its nominal value.

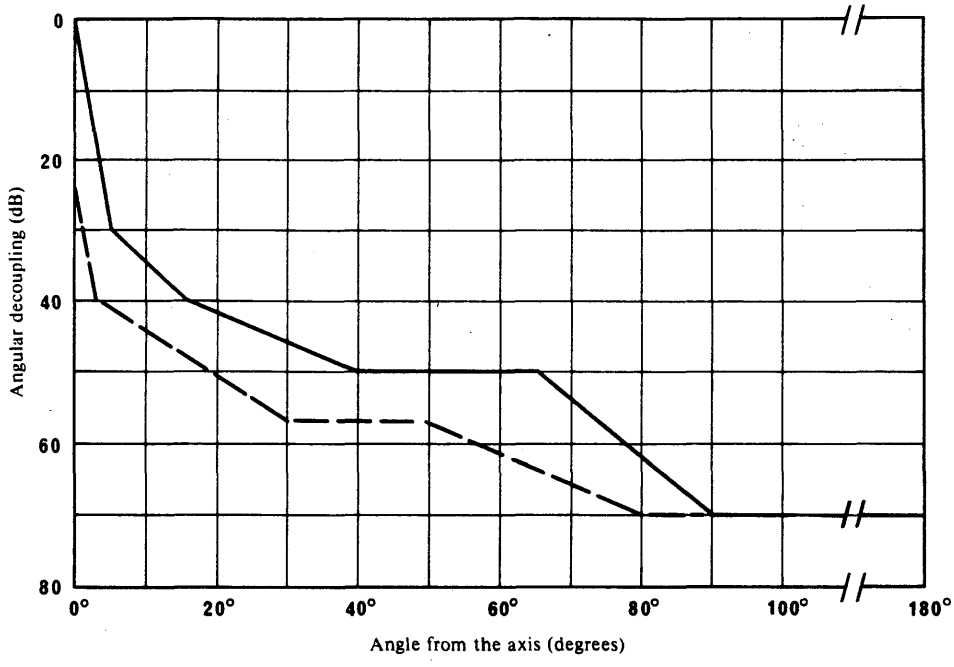


FIGURE 7 – *Antenna radiation masks*

Parabolic antenna, $D/\lambda = 75$

——— Co-polarization

— — Cross-polarization

A normalized network density (γ) has been defined as:

$$\gamma = \frac{2N\rho^2}{\text{overall area covered by the network}} \tag{8}$$

where:

N: number of radio nodes in the network,

ρ: mean square hop length.

The results of Fig. 8 show that in high density networks the highest efficiency is achieved with 4-PSK modulation. However, the modulation method moves in favour of 8-PSK or even 16-QAM when the network density is lower. This is typical of other results by Moreno [1982] which show that the spectrum utilization efficiency of modulation methods depends on the interference environment.

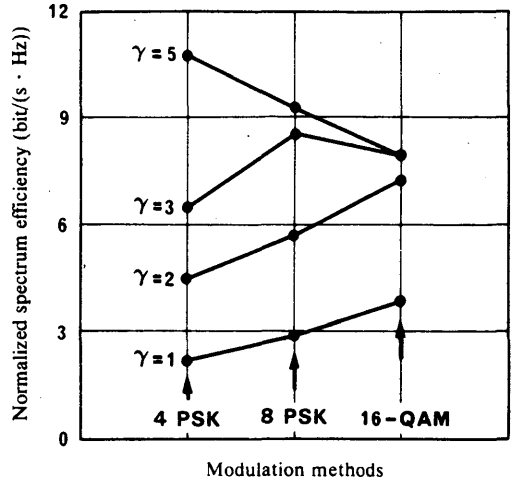


FIGURE 8 – *Spectrum efficiency in a mesh network*

Antenna radiation mask of Fig. 7
Performance degradation due to frequency re-use: 1 dB.

ANNEX III

A LINK AND NETWORK ENERGY LOAD CRITERION FOR THE FREQUENCY BANDS USED FOR RADIO-RELAY SYSTEMS

One comprehensive yet extremely specific way of quantifying the spectrum use efficiency of radio-relay systems is to apply link and network band energy load criteria.

For a single isolated radio-relay link of length (L km), the link criterion of energy load (I_L) for operation in a band Δf is expressed by the equation:

$$I_L = \frac{\Delta f \sum_{i=1}^{N} P_{ti} G_{ti}}{LK} \qquad \left(\frac{\text{W} \cdot \text{Hz}}{\text{km} \cdot \text{volume of transmitted information}} \right) \tag{9}$$

where:

N: number of transmitting installations on the link;

P_{ti}: transmitter power of the ith station;

G_{ti}: gain of the ith station transmitting antenna;

K: volume of transmitted information (number of telephone channels, transmission rate of digital information in bit/s, number of television channels).

The value of the radio-relay equipment energy parameters $P_{ti} G_{ti}$ in equation (9) is that at which channel performance objectives at the output of the link are met when taking into account the statistics of random signal level variations on all the intervals of the link at the specified working frequency, f, and possible multipath distortions in the band Δf, but disregarding impaired channel performance factors due to interference from other radio systems.

Since $P_{ti} G_{ti}$ is proportional to the volume of the antenna pattern of an antenna with gain G_{ti}, and LK is the information transmitted over distance L, I_L is inversely proportional to the spectrum efficiency defined in § 2.2. Obviously, all other factors being equal, most efficient use of the band Δf will correspond to the smallest value of I_L. The value $P_{ti} G_{ti}$ in equation (9) defines the value of the equipment energy parameters in the direction of maximum radiation of the ith station transmitting antenna. Hence, from the standpoint of the electromagnetic compatibility of the link with other radio systems sharing the same frequency bands, equation (9) defines limiting conditions corresponding to the maximum coordination distances between stations.

To evaluate the spectrum efficiency of a radio-relay link located on any territory of arbitrary configuration with an area of S km^2, the network energy load criterion I_N may be applied in accordance with the equation:

$$I_N = \frac{\Delta f l \sum_{k=1}^{l} \sum_{j=1}^{M_{fk}} \sum_{i=1}^{N_{fk}} P_{tijk} G_{tijk}}{SK} \qquad \left(\frac{\text{W} \cdot \text{Hz}}{\text{km}^2 \cdot \text{volume of transmitted information}} \right) \tag{10}$$

where:

M_{fk}: number of links in the network, operating in the kth frequency band;

N_{fk}: number of stations of the jth link, operating in the kth frequency band;

P_{tijk} and G_{tijk}: respectively, the transmitter power and transmitting antenna gain of the ith station of the jth radio-relay link operating in the kth frequency band;

l: number of frequency bands required to set up the radio-relay network with a given density over a territory S (km)^2.

In equation (10) it is assumed that the occupied bandwidth Δf is the same in each frequency band; all the links in each frequency band are located as far as is necessary from one another, to ensure permissible levels of channel performance impairment due to reciprocal interference among all stations on the links operating in the same frequency band. This impairment is determined by the permissible level of the interfering signals, the antenna patterns, the information transmission method, the polarization decoupling, etc.

Clearly, a radio-relay network displaying the lowest value of I_N will be that with the greatest efficiency of spectrum use.

The value of the equipment energy parameters $P_t G_t$ in equations (9) and (10) may be adopted, either averaged over the whole band Δf or applied to any appropriate frequency band Δf_i (e.g. Δf_i = 4 kHz). In the latter case, allowance must be made for the radiated power spectrum irregularity.

ANNEX IV

SPECTRUM UTILIZATION EFFICIENCY IN RADIOCOMMUNICATION AND BROADCASTING NETWORKS

1. Introduction

This annex briefly discusses the results obtained in a study which has been made of the variation in spectrum utilization efficiency in radiocommunication and broadcasting networks as a function of the modulation method and the parameters selected, the quality requirements for the information to be transmitted and the parameters characterizing the organization of the network, on the basis of a comparison of the frequency bands required for network organization on the basis of optimum and real radio installations. Details of this analytical study will be found in [Bykhovsky, 1979; Bykhovsky and Pavliouk, 1986 and 1987].

2. Broadcasting and land mobile service networks

2.1 Characteristics of an idealized network

For the sake of simplicity and to obtain finite analytical expressions, this annex deals with the simplest models of a network in the form of ideal lattice and propagation conditions in the UHF band. However, the general laws will be the same for more complex models in a real network.

The network is made up of squares of equal dimensions with the central station in the case of mobile communication systems or the broadcasting transmitter being located in the centre of the area (Figure 9). The dimension r of the service area is considered to be given. In areas bearing the same digit in Figure 9, the same frequency channel can be used if the distance R between these areas provides sufficient interference attenuation. The antennas of the radio installations operating in the network are not directive in the horizontal plane and only use one type of polarization.

All the transmitters have the same power and a stable carrier frequency and they do not produce any out-of-band or spurious radiation; the receivers have ideal selectivity characteristics.

As the noise immunity in the reception of signals for different modulation techniques increases with increasing width of the communication channel frequency band F_m used for the transmission, an increase in F_m can be accompanied by a decrease in R, i.e. more frequent use of the same frequency channel in different areas. For a certain value of F_m, the total bandwidth required for the network can be minimised.

2.2 Analysis of interference in the network

In the idealized network under consideration, if M_c frequency channels are used in each area, the total number of channels N required to cover the whole region by this network is:

$$N = M_c \, ([R/2r] + 2)^2,$$

where [A] is the integer part of number A. The frequency bandwidth F_c allocated to the network is then equal to:

$$F_c = NF_m. \qquad (11)$$

In the analysis it is convenient to consider the values of these bands in relation to the frequency bandwidth of the modulating signal F_o, i.e.

$$\tilde{F}_m = F_m/F_o \quad \text{and} \quad \tilde{F}_c = F_c/F_o.$$

As can be seen from Figure 9, the input of a receiver located at the edge of the service area is subject to the effect, not only of the wanted signal, but also of co-channel interference from transmitters in the four adjacent areas located at about distance R and from transmitters in four other areas located at about distance $R/\sqrt{2}$. Expressing the power of the wanted and interfering signals by the formula

$$P = \frac{P_c G_1 G_2}{(4\pi)^2} \left(\frac{\lambda}{L}\right)^2 V(L)$$

where P_o is the power of the transmitter; G_1 and G_2 are gain coefficients of the receiving and transmitting antennas, λ is the wavelength, L is the distance from the receiver to the corresponding transmitter and V(L) is the attenuation factor on the propagation path,

and assuming the following commonly used model to calculate the attenuation factor:

$$V(L) = \alpha/L^k,$$

where k, depending on the height at which the antenna is suspended and the distance, may vary between 2 and 6, it is possible to calculate $R/2r$ as a function of the signal/noise ratio at the receiver input [Bykhovsky, Pavliouk, 1987]. The following relation for the calculation of the total frequency bandwidth allocated to the network:

$$\tilde{F}_c = \tilde{F}_m m_c \left\{ \left[1/2 \quad \left(4\rho/\beta \right)^{1/k+2} \right] + 2 \right\}^2 \qquad (12)$$

where ρ is the signal/noise ratio at the receiver input and

$$\beta = \left[1 + 2^{-(k+2)/2} \right]^{-1}$$

The signal/noise ratio ρ at the receiver input must not be less than the protection ratio ρ_s, i.e. the value at which the necessary signal/noise ratio ρ_o is obtained at the output. Once the relationship ρ_s (ρ_o) is obtained for different modulation methods and parameters, it is possible to evaluate the influence of these factors on the number of channels required, i.e. on the total frequency bandwidth F_c required for the organization of the network. The values of the bandwidths F_{opt} and F_c necessary for the organization of the network on the basis of optimum and real radio installations are compared and the spectrum utilization efficiency is defined:

$$\mu = F_{opt}/F_c.$$

2.3 Protection ratios for various systems for the transmission and reception of traffic

2.3.1 Optimum (according to Shannon) systems (OPT)

According to the principles of information theory [Gallager, 1968], the communication capacity of a communication channel on which a subscriber or a listener receives a wanted communication is determined by the relation $C_o = F_o \, \mathrm{Ln}(1 + \rho_o)$, where F_o is the bandwidth of the wanted communication and ρ_o is the signal/noise ratio at the receiver output. If the signal/noise ratio at the receiver input is equal to the protection ratio ρ_s and the bandwidth of the communication channel over which the signals are transmitted is equal to F_m, then the communication capacity is $C_p = F_m \, \mathrm{Ln}(1 + \rho_s)$. It must exceed or at least be equal to the communication capacity of the channel over which the subscriber receives a wanted communication, i.e. $C_p \geq C_o$. Hence the minimum possible value of the protection ratio ρ_s at which the subscriber will receive a communication with a signal/noise ratio equal to ρ_o is defined as:

$$\rho_s = \left(1 + \rho_o \right)^{1/\tilde{F}_m} - 1,$$

2.3.2 Frequency modulation (FM)

In accordance with [Gallager, 1968], it is possible to obtain:

$$\rho_s = \frac{\kappa^2 \rho_0}{3\tilde{F}_m (\tilde{F}_m - 2)^2},$$

where κ is the traffic peak factor (normally κ = 3 - 5).

2.3.3 Amplitude modulation with single sideband (AM SSB)

It is obvious that for this type of modulation $\rho_s = \rho_0$.

2.3.4 Pulse-code modulation (PCM) used for the transmission of signals at n levels and transmission mode using a class of signals with M-ary phase shift keying (MPSK)

In accordance with [Viterbi, 1966] we obtain:

$$\rho_s = \frac{2(n/\tilde{F}_m + 1)\,\mathrm{Ln}2 + \mathrm{Ln}\,\rho_0}{\sin^2 (\pi \cdot 2^{-2n/\tilde{F}_m})}$$

2.3.5 16 quadrature amplitude modulation (16QAM)

In accordance with [Bykhovsky and Pavliouk, 1986 and 1987], for this type of modulation:

$$\rho_s = 10\,(3.22 + \mathrm{Ln}\sqrt{\rho_0})$$

2.4 Analysis of calculation results

Equation $\rho = \rho_s$ in (12) and using the relationships ρ_s (ρ_0) given in section 2.3, it is possible to determine the relationship $F_c(F_m)$ for different types and parameters of modulation. The results of calculations are given below for k = 3, n = 8 and M equals 4 and 8 for PCM-M-PSK and 16 for QAM, as well as for SSB and FM.

The relationship $\tilde{F}_c(\tilde{F}_m)$ appears as saw-toothed lines as shown in Figure 10 for a network using optimum radio installations, for two values of ρ_0, and also for FM with one value of ρ_0. The squares of integers (4, 9, 16 etc.) shown in the breaks in the lines indicate how many frequency channels N are required to set up a network with one frequency channel per area, i.e. with M_c = 1. For PCM the relationships $\tilde{F}_c(\tilde{F}_m)$ are discrete points lying on the same lines.

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Figure 10 shows that the increases in $\tilde{F}_m$ will only be advantageous if at the same time there is a sudden reduction in the R/r ratio, i.e. N and $\tilde{F}_c$. The minimum values of the saw-toothed lines $\tilde{F}_c(\tilde{F}_m)$ are shown in Figure 10 by means of broken lines. This figure clearly shows the presence of optimum values of F_{mo} at which the relationships $\tilde{F}_c(\tilde{F}_m)$ have minimum values equal to $\tilde{F}_{co}$, i.e. maximum spectrum utilization efficiency.

Figure 11 shows the relationships $\tilde{F}_{co}(\rho_o)$ for optimum radio installations and also the relationships $\mu(\rho_o)$ for all the other types of modulation considered under the same circumstances. For all types of modulation $\tilde{F}_{co} = \tilde{F}_{co(OPT)}/\mu$. Figure 11 illustrates the spectrum utilization efficiency in a network for all the types of modulation considered in relation to optimum radio installations.

As can be seen from Figure 11, FM has the lowest spectrum utilization efficiency, since when this type of modulation is used, the bandwidth required for network organization is approximately five times greater than in the case of optimum radio installations. On the other hand, the type of modulation which is closest to the optimum for all values of ρ_o is 16QAM, which for the organization of the network requires only 1.5 times the bandwidth needed in the case of optimum radio installations, even for the reception of traffic with very high quality.

Of the PCM-M-PSK digital methods of transmission, the best for the conditions being considered is the system with M = 8 (8 - PSK).

If the reception quality requirements are not very high, the closest with respect to optimum radio installations is AM-SSB. However, the efficiency of AM-SSB drops appreciably as the reception performance requirements are increased, particularly if account is taken of the effect of the frequency instability of real transmitters.

The increase in bandwidth required for network organization as a result of transmitter frequency instability can be evaluated by replacing in (11) F_m by an assigned frequency band value F_p equal to

$$F_p = F_m + 2\Delta F,$$

where ΔF is the tolerance for transmitter frequency instability.

For a land mobile service network with AM-SSB (5 MHz band) and with FM (30 MHz band), with frequency instability tolerances for AM-SSB of 50 Hz and for FM of 20 parts in 10^6, the value of $2\Delta F$ is 100 and 1 200 Hz, respectively. The calculated values of the relation $\mu(\rho_o)$ with the transmitter frequency instability taken into account are shown in Figure 11 by means of broken lines. Comparison of these curves shows that the value μ for AM-SSB decreases as a result of transmitter frequency instability in approximately the same way as with FM although the tolerance for AM-SSB was taken to be twelve times less than for FM.

This example of the effect of transmitter frequency instability shows that it is possible, when applying the spectrum utilization efficiency criteria, to take into account the characteristics of a real receiving-transmitting installation.

Consideration was given above to the operation of optimum and real radio installations in one and the same idealized radio network. If the values of $\tilde{F}_{co}$ in this network are compared with the frequency band allocated for the operation of a real (not regular) network, it is possible to determine the additional losses caused by the non-ideal nature of the network.

3. Analysis of a rural network with single section radio-relay links

In [Bykhovsky, 1979] a similar approach was adopted in analyzing the problems of optimum frequency planning for a rural network based on radio-relay links. In the model adopted for this network each regional station has a specific service area in which it is linked by single-section links to four stations located in different rural localities.

The spectrum utilization efficiency in such a network was considered in relation to optimum radio systems as a function of a large number of parameters: the frequency band used, the type of frequency and polarization-space planning of the network, the height at which the antenna was suspended and the antenna diameter, and also the type and parameters of the modulation (FM and PCM with M-PSK). A determination was also made of combinations of parameters which would ensure maximum spectrum utilization efficiency and at the same time compliance with the network's performance characteristics.

Table III gives the results of calculations of spectrum utilization efficiency in the case of network operation in the 8 GHz band with antenna suspension heights of 45 m, antenna diameters (D) of 1.5 and 3 m, two and four frequency operation (K = 2 and K = 4, respectively) and two types of polarization plan - with the use in the network of one (1P) and two (2P) types of polarization. For PCM the data are given for 2 PSK, 4 PSK and for a value (M_{max}) which ensures maximum spectrum utilization efficiency. The dashes in Table III indicate that with the given combinations of parameters the performance characteristic standards cannot be met.

TABLE III - Spectrum utilization efficiency (μ)

Modulation	Frequency Plan	Parameters of modulation	D = 1.5 m		D = 3.0 m	
			1P	2P	1P	2P
FM	K = 2		0.27	0.3	0.285	0.285
	K = 4		0.2	0.25	0.25	0.25
PCM	K=2	M=2	-	-	0.055	0.049
		M=4	-	-	-	-
		Mmax	-	-	0.055	0.055
	K=4	M=2	0.125	0.0625	0.0625	0.0625
		M=4	0.125	0.125	0.125	0.125
		Mmax	0.25	0.25	0.25	0.5

As Table III shows, the use of PCM with PSK produces a gain in the spectrum utilization efficiency only with the optimum number of modulation conditions ($M_{max} \geq 8$) and antennas with a diameter of 3 m. Detailed recommendations on the optimization of a network are given in [Bykhovsky, 1979].

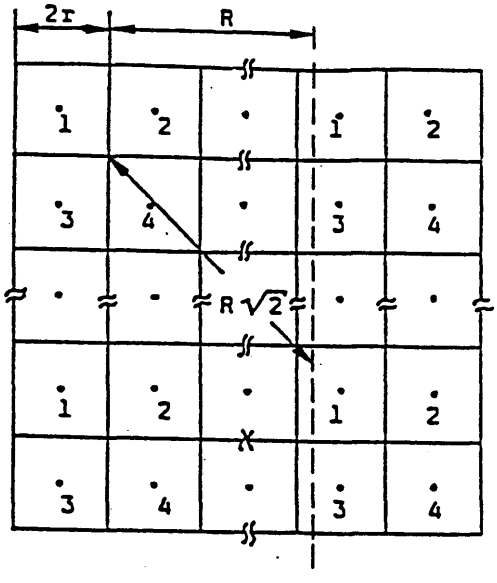


FIGURE 9

Network model

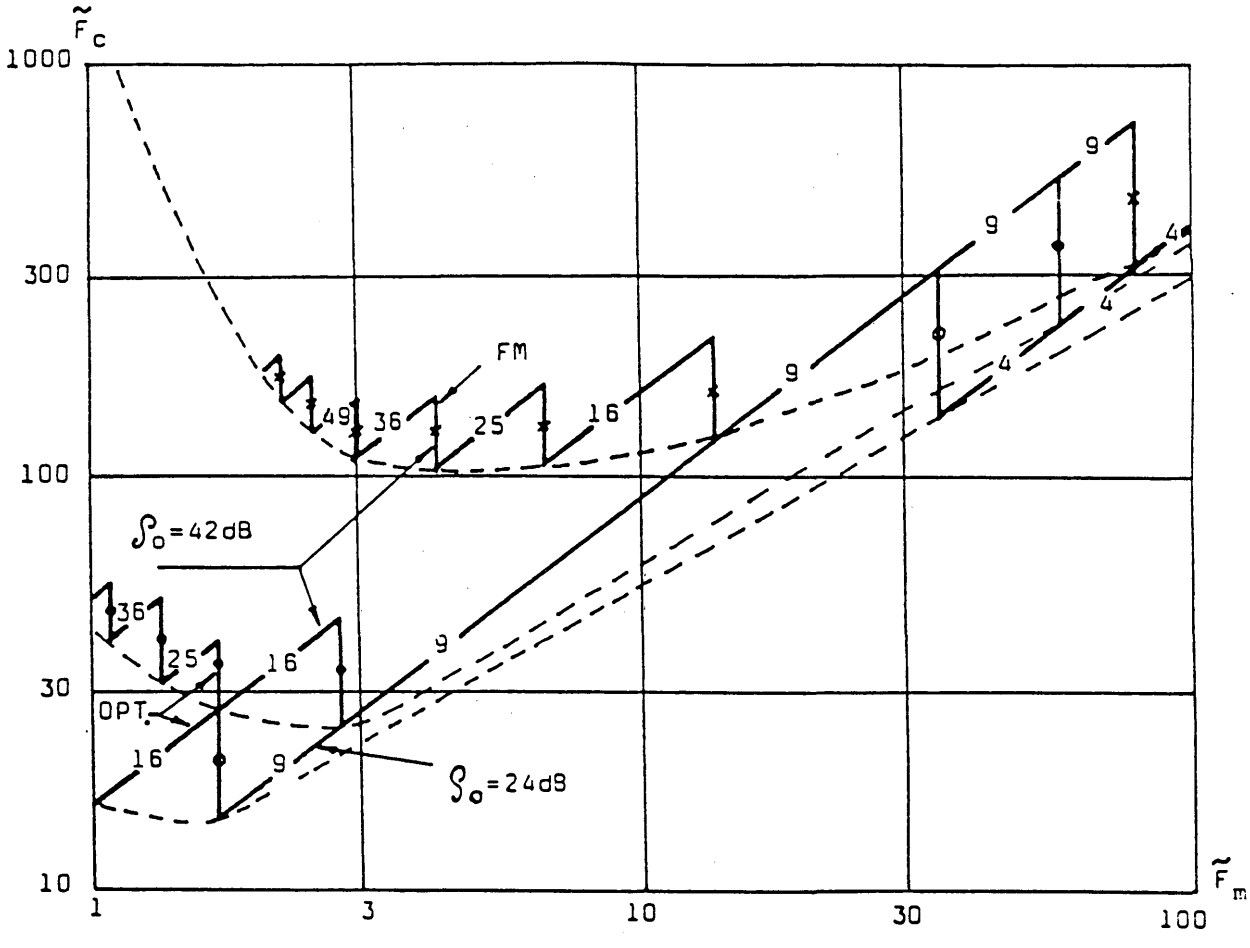


FIGURE 10

Comparison of frequency bands in the considered network

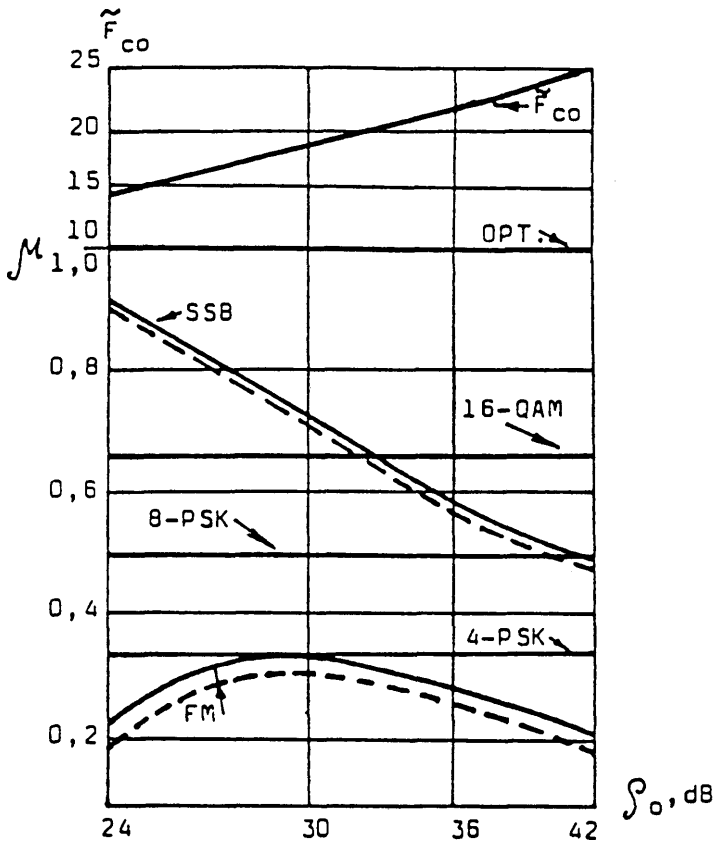


FIGURE 11

Spectrum utilization efficiency in a network with different types of modulation in the considered network

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REPORT 656

EFFICIENT SPECTRUM UTILIZATION USING PROBABILISTIC METHODS

(Question 45/1)

(1978)

1. Introduction

Communications technology has advanced rapidly during the past decade. As a result, the use of radiocommunication services by administrations has multiplied and placed new demands on the radio spectrum. These demands have traditionally been met by simply advancing the upper limit of the usable spectrum through development of new modulation techniques and equipment that operate at higher frequencies.

As more systems use the available spectrum space, performance that was once limited by background environmental noise has become increasingly affected by interference signals. More efficient spectrum utilization may depend on the acceptance of performance criteria developed through the application of probabilistic methods. These methods would help evaluate spectrum utilization against required performance standards, considering the statistical characteristics of both the desired and interference signals.

This Report reviews the spectrum utilization question from the probabilistic viewpoint, reviews selected sharing criteria contained in the Radio Regulations, and provides several examples of probabilistic analysis techniques.

2. Current sharing criteria

Efficient spectrum utilization is related to the effect of interference and its impact on system performance. Harmful interference has been defined as, "interference which endangers the functioning of a radionavigation service or of other safety services or seriously degrades, obstructs, or repeatedly interrupts a radiocommunication service operating in accordance with these Regulations." Radio Regulations No. 163. This definition relating to radionavigation and other safety services denotes that if the inteference "endangers the functioning" it is harmful and cannot be tolerated. However, the second part of this definition raises questions regarding how to specify what levels "degrade, obstruct or repeatedly interrupt" the radiocommunication services, and how to relate these conditions to an acceptable level of required system performance.

The problems of dealing effectively with spectrum utilization questions and procedures have been recognized (Recommendation No. 61 of the WARC-79). The concept of permissible interference, Radio Regulations No. 161, and the frequent reference to the probability of interference in the Radio Regulations reinforces the need to quantify the interference prediction process with probabilistic considerations.

Many frequency sharing and coordination provisions appear to be based on engineered frequency allocation plans or probabilistically derived criteria that represent the very pessimistic tails of the desired and interfering signal distributions. That is, the guidelines and sharing criteria were based on the most unfavourable interference assumptions, e.g. co-channel operations, terrestrial system mainbeam coupling, common volume mainbeam rain scattering, 0.01% of the time propagation path transmission losses, and antenna patterns that encompass approximately 90% of the pattern peaks. To achieve more efficient spectrum utilization, techniques must be developed to characterize the probability of interference and its impact on system performance. To achieve this, an approach to the interference prediction process that considers the statistical properties of the desired signal, system performance and interference signal characteristics needs to be advanced and accepted.

3. Interference prediction process

An interference prediction process quantitatively assesses potential band sharing problems between systems. The process is intended to identify the level of unwanted energy which is coupled into an electronic device. For spectrum utilization studies, antenna coupled interference signals are of primary concern. Consideration of this coupling mode includes as constituents, interference sources, radiators, a transmission medium, receptors, detection devices, and information processing elements. This configuration and a general characterization of the elements is illustrated in Fig. 1. This one-to-one interference interaction is typical of what is initially considered during many spectrum sharing studies. To assess fully the interference potential of introducing a new system into the environment, the effects of multiple interference sources and victims of interference should be considered. This is typically accomplished by using detailed information on actual equipment deployments or estimates of equipment distributions when detailed information is not available.

To assess the potential for band sharing between systems requires detailed consideration of numerous factors. The prediction of receiver input signal-to-interference $(S/I)_{IN}$ ratios in a system is a major item and can be calculated for continuously operating systems by:

$$\begin{aligned}(S/I)_{IN} &= (S/N)_{IN} - (I/N)_{IN} \\ &= (S/N)_{IN} - (P_t + G_t - L_b + G_R - R_s - BW_{CF} - OFR)\end{aligned} \tag{1}$$

where

$(S/N)_{IN}$: input desired signal-to-noise ratio (dB)

P_t: transmitter power output (dBm)

G_t: transmitter antenna gain (dB)

L_b: basic transmission loss between isotropic antennas (dB)

G_R: receiver antenna gain (dB)

R_r: receiver noise level (dBm)

BW_{CF}: bandwidth correction factor (dB)

OFR: off-tune rejection (dB)

$(I/N)_{IN}$: receiver input interference-to-noise ratio (dB)

Each term can be represented by a single deterministic value, the use of which produces predictions that are single valued and, in many cases, overly pessimistic. In reality, each of the terms represent a variable in the prediction process that has a probability distribution of values associated with it. The combined effect of these distributions can be easily recognized if independence and log normal characteristics for each variable are assumed. Under these conditions, the standard deviation of the $(S/I)_{IN}$ distribution is described by the square root of the sum of the individual squared deviations for each variable. Through a process such as this, the effects of the individual parameters are considered and the resultant $(S/I)_{IN}$ distribution accounts for parameters that have large as well as small variations about the mean. It is more realistic to account for these variations together than to consider a combination of "mean" values, 1σ, 2σ, etc. for the individual parameters in the calculation of $(S/I)_{IN}$.

It is recognized that the log normal assumption does not hold true in many situations; hence, Monte Carlo (Report 524) and other simulation techniques, as well as closed form analytic solutions, have been used.

To assess the potential for band sharing, it is necessary to evaluate a set of statistically derived receiver $(S/I)_{IN}$ data relative to the operation of the interacting systems. Factors such as the respective operating areas, operational duty cycles, probability of system interception, the telecommunication service being provided, the desired service probability, and minimum service requirements, etc. require consideration. The ability to bring together the statistical, operational and performance data in an organized fashion to produce acceptable sharing criteria related to performance requires discipline, persistence and insight.

4. Typical probabilistic models

A number of statistical interference prediction techniques have been developed. The level of detail varies and most have been in response to specific spectrum management problems or the need to enhance heretofore deterministic models. A review of several models is presented here and represents a typical cross-section of models. Models have been previously reported within the CCIR which also address the statistical aspects of spectrum sharing (e.g., Reports 520, 522, 523 (Geneva, 1982), 524, 525 and 651). All of these models encompass Monte Carlo simulation or some form of closed form analytic solution.

4.1 *Frequency distance trade-offs*

One technique used in spectrum sharing studies encompasses the trade-off between equipment separation distances and tuned frequency. The basic factors considered in deriving a frequency distance (FD) curve are the interfering transmitter power spectrum, the transmitting antenna, the propagation path to the antenna of the victim receiver, the receiving antenna, radio frequency filters, and the bandpass characteristic of the receiver. Given these characteristics, the amount of spectrum dependent (SD) power in the IF output of the victim receiver due to the interfering source can be computed.

This level can be compared to a maximum power level which the receiver can tolerate. To identify the distance between the systems required to satisfy the maximum tolerable level of interference, an inverse solution to the computation of path loss is necessary. This provides for the generation of frequency and distance trade-off curves similar to that illustrated in Fig. 2 (Report 654). Most applications of the (FD) concept have considered only the discrete, single-valued specification of all parameters of interest. A more useful concept results when the parameters are statistically characterized.

Jennings *et al.* [1976] extended the application of (FD) trade-offs to consider the statistical distribution of antenna gain and propagation path loss. The major problem in developing a statistical (FD) separation model was to properly combine the probability distributions of the several contributing variables.

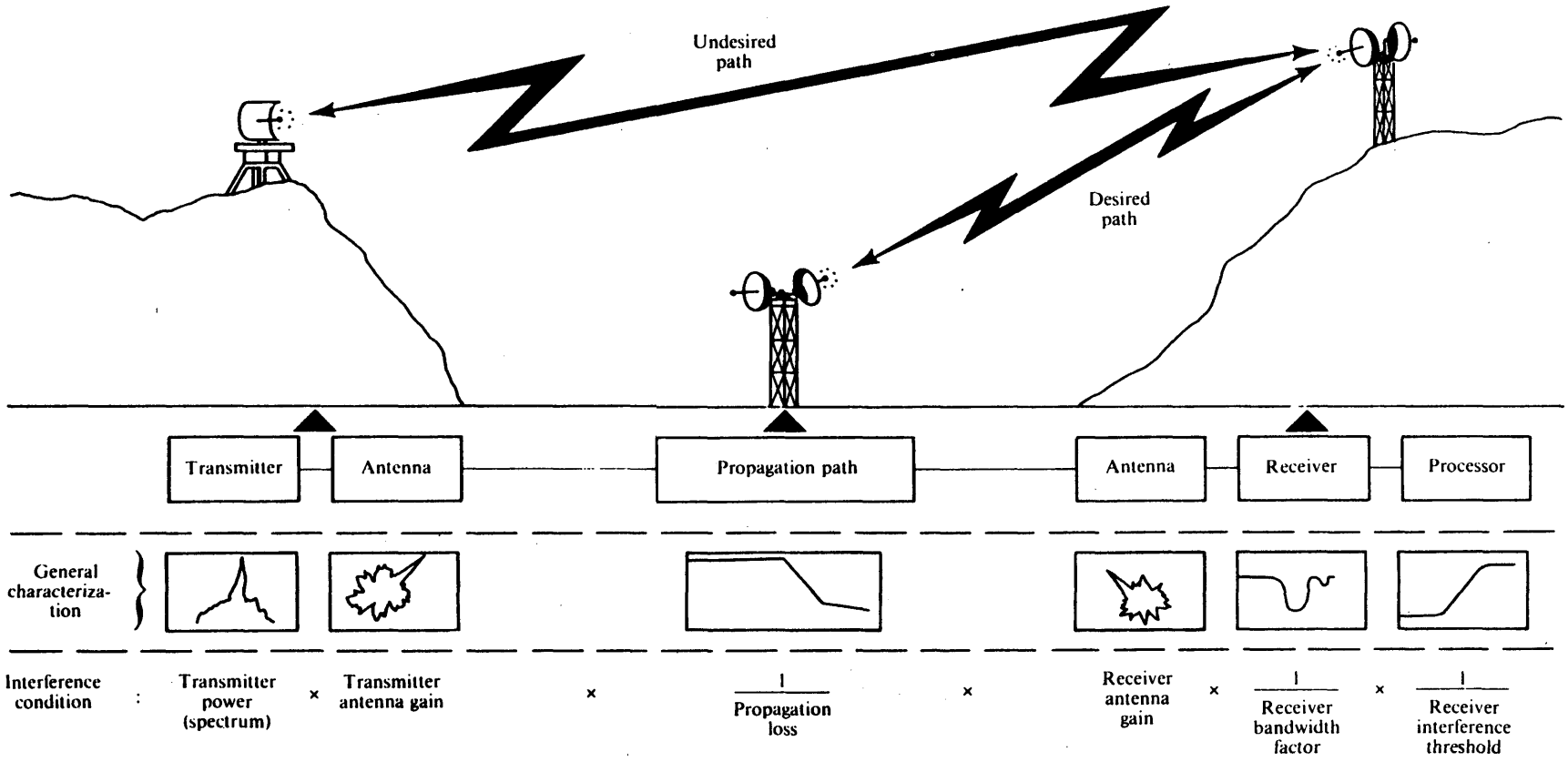


FIGURE 1 — *Generalized interference prediction process*

In summary, for each frequency separation increment, distance values are calculated for the required probability levels. The process of determining the SD, and then convolving the statistics of the antenna gain and propagation loss is repeated over the range of frequencies and distances of interest. This process produces curves similar to those in Fig. 3 and illustrates the trade-off between frequency and distance separation, and the percentage of time a specified interference-to-noise ratio will be exceeded.

4.2 *Assessment of airborne interference*

Terrestrial line-of-sight microwave communications systems are typically designed with large median carrier-to-noise ratios to prevent loss of information during periods of fading. The excess carrier power may serve to mask interfering signals when fading is not occurring. Weissberger *et al.* [1975] presented a procedure for predicting interference in such cases. The measure of interference was the increase in outage time (time during which the signal-to-interference-plus-noise ratio falls below the minimum acceptable value) produced by an airborne interfering source.

This analysis procedure considered the expected desired signal power level, the interfering signal power and background noise in the microwave receiver. Variations in the desired signal level would result from fading and equipment failure, whereas variations in the interfering signal level would be due to movement of the interfering transmitter, changes in the orientation of its antenna, fading, and periods of no transmission. Variations in background noise would be due to fading and interference in the preceding hops of a multiple-hop system. The probability of outage, $P(O)$, was obtained using computerized Monte Carlo techniques.

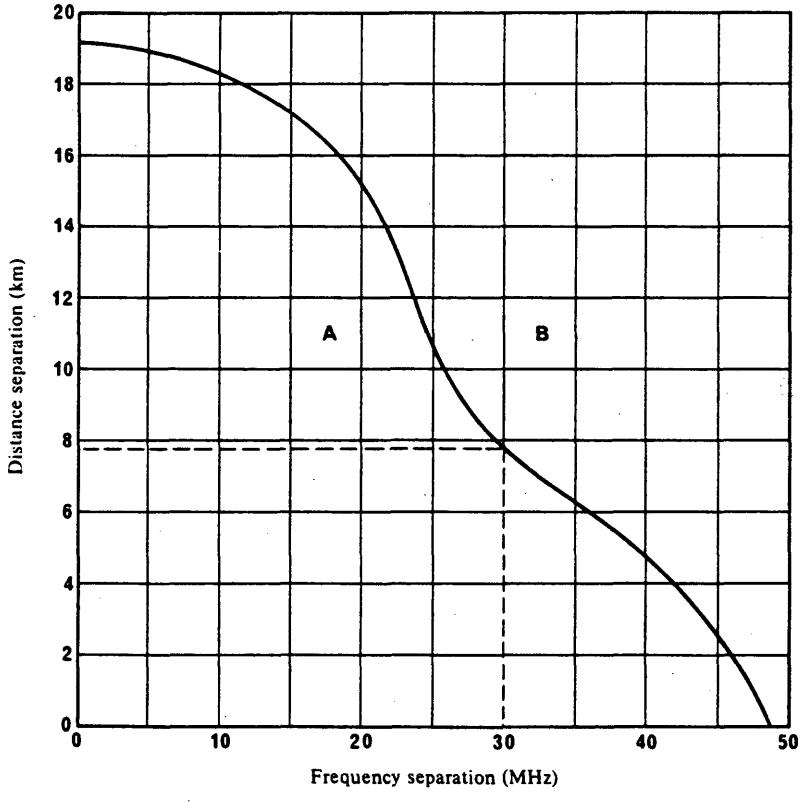


FIGURE 2 - *Typical frequency distance trade-off curve*

A: interference
B: no interference

I/N = 0 dB

Other investigators have taken the work of [Weissberger *et al.*, 1975] and developed meaningful extensions. Wheeler and Haakinson [1977] computerized a technique for calculating and plotting contours of median signal-to-interference in fixed terrestrial microwave systems from airborne interference sources. By varying the airborne sources' characteristics, this technique can be used to identify changes in the S/I contours, determine the sensitivity of the terrestrial system to these changes, and assist in determining the potential for spectrum sharing. The airborne source characteristics which can be varied include the transmit frequency, antenna pattern, effective isotropic radiated power, bandwidth, operating altitudes/area of the airborne source, and the location of the receiver to which the airborne source is communicating. Consideration is given to the location, pointing angle, antenna pattern, received signal levels and bandwidths of the terrestrial microwave system(s).

This automated technique predicts the impact on one or more microwave receivers (e.g., multiple hop links that repeat frequencies every other hop) and produces a composite plot of input S/I similar to that illustrated in Fig. 4. Statistical tabulations are also provided which identify the percentage of the operational area which falls within specified S/I levels.

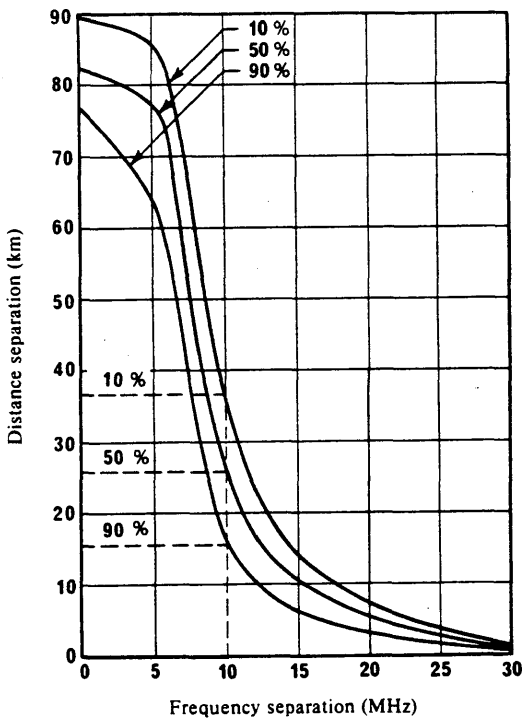


FIGURE 3 - *Sample statistical frequency-distance separation curve*

I/N = 10 dB

To generate these S/I contours, which are analogous to the coordination contours in Appendix 28 to the Radio Regulations, reliable frequency assignment data must be available. In general, this capability provides an efficient and realistic method by which to assess the impact of airborne emissions on terrestrial microwave systems. The S/I contours and the statistical data can be used in conjunction with information on the various links to statistically assess the interference potential of the airborne sources.

Mayher and Parlow [1976] presented a procedure to develop statistical sharing criteria between airborne interference sources and terrestrial microwave systems. In effect, they stated that the long-term probability of penetration of specific S/I contours by an airborne source can be calculated that will not reduce terrestrial system performance below a specific minimum level. The specification of this minimum performance level is dependent on the required service probability and criticality of the service being supported, e.g. radionavigation, other safety services, or services having routine or non-critical functions. This approach was developed for both analogue and digital systems.

Using the basic design data for the digital link, the probability of error can be calculated under various levels of fade (0 to max.) with and without interference. The condition of interference is directly related to specific S/I contours and, for example, if I is noise-like, to its probability of error. If an acceptable change in the no-interference-probability of error can be identified, then the probability (P_I) with which the airborne source can penetrate specific S/I contours can be calculated. Data of this sort allows quantitative assessment of the effects, possible restrictions, design changes and operational guidelines during development of possible sharing criteria.

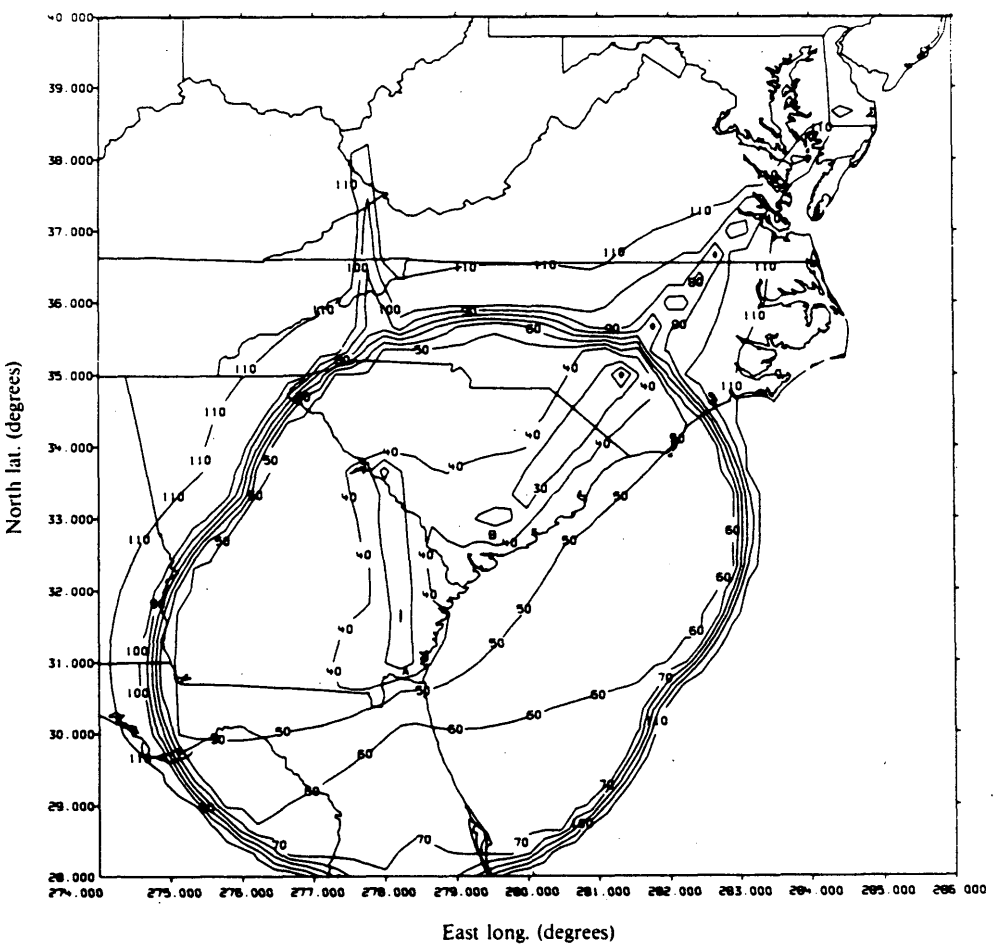


FIGURE 4 — *Median signal level of the terrestrial receiver(s) relative to co-channel aircraft interference*
Typical $(S/I)_{IN}$ contours

A: Terrestrial receiver # 1. Main beam azimuth = 356° E of N

B: Receiver # 2 main beam azimuth = 33° E of N. Aircraft altitude = 5 km.

4.3 *Probabilistic trade-offs and efficiency*

Berry, [1977] developed a technique based on the assumption that statistical distributions for each of the prediction parameters (i.e. antenna gains, power, etc. for equipment in the environment) will exist and that these distributions, rather than individual equipment characteristics and interactions, can be used directly to develop trade-off relationships. The intent of these trade-offs is to develop optimum parameters for classes of equipment or operations, rather than the fitting of one new operation into a crowded band. With this approach, changes in regulations, equipment specifications, operational tactics, or assignment strategies can be expressed as changes in the statistical distributions and used to compute the probability of successful communications, i.e. the probability that the signal-to-interference (or signal-to-(interference + noise)) ratio exceeds specified values.

4.4 *System model for band sharing*

To assess the potential for band sharing, it is necessary to evaluate a set of statistically derived $(S/I)_{IN}$ data in relationship to the operation of the interacting systems. Sailors *et al.* [1977] described a technique for computing in a probabilistic manner the electromagnetic compatibility of communication systems. The effect of interference was characterized by a parameter called circuit compatibility. This was defined as the ratio of the estimated circuit reliability considering the operation of an interfering transmitter (without considering other potential interferers) to estimated circuit reliability under interference-free conditions.

Circuit compatibility is a function of the probability of simultaneous transmission of the desired and undesired signals, the joint probability that the received signal power-to-noise density ratio and $(S/I)_{IN}$ is above that required for the desired grade of service, and (for HF sky-wave systems) of the joint probability of ionospheric layer support for both the desired and undesired paths. Formulae are given for circuit reliability under interference-free conditions, techniques for extending these methods to estimate the expected performance of skywave systems in the presence of skywave interference are provided, and generalized to other propagation modes and frequency bands.

4.5 *Analytical solution to spectrum sharing*

In a series of reports, [Middleton, 1974, 1976] has examined the use of analytical methods to derive statistical measures of telecommunication system performance in interference environments. His principal goal was to show how analytic models could be used.

Middleton has developed communication-performance models based on the assumption that emitters and receivers in operational environments are positioned according to a Poisson process in time and space. The signal structure for this type of interference environment has been shown by Middleton to be highly impulsive and not simply Gaussian.

Using models based on these principles, Middleton has developed expressions which allow the calculation of amplitude probability distribution (APD), or portion of the time the envelope of the interference exceeds some level, phase statistics of the signals, and other statistics of the interference process, e.g. the probability density function (PDF) of the signals' instantaneous amplitude. Spaulding and Middleton [1977] applied this type of model to develop predictions of digital system receiver performance.

This generalized type of canonical model (i.e. their analytic forms are invariant with regard to the particular interference source and its quantifying parameter values, levels, etc.) has promise. This concept requires further investigation and operational application to existing spectrum management problems. Middleton is working with others [Cohen, 1977] to further develop and apply his models.

5. Summary

There are numerous sharing criteria contained in the Radio Regulations, many of which are deterministic or have minimum statistical considerations. The need for more statistical procedures that consider factors such as the nature of the service, number of interference sources, percentage of time the interference is present, and pertinent operational factors has been stated in Recommendation No. 61 of the WARC-79. Numerous probabilistic models have been developed which deal with problems of spectrum utilization/sharing that encompass Monte Carlo simulation or analytic techniques. This move towards more probabilistic considerations during the development of spectrum-sharing options is an important trend since many system designers have for years been using statistical methods to determine if system designs meet required performance levels and service probabilities.

To promote the development of statistical frequency sharing criteria, a limited set of telecommunication services, that operate exclusively or share with others, which would benefit from the availability of such criteria should be identified. For these services, statistical-sharing criteria should be developed that account for the desired service probability, the criticality of the service being supported (i.e. safety of life or other critical services) and the expected interference. Development, assessment, evaluation, and acceptance of these criteria could form the basis for more extensive use of statistically derived frequency-sharing criteria and more efficient spectrum utilization.

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REPORT 1097

SPECTRUM EFFICIENCY OF NARROW-BAND MODULATION TECHNIQUES

(Question 18/1)

(1990)

1. Introduction

The demands on the radio spectrum continue to increase so there is a continuous need to make more efficient use of the spectrum. One approach is to reduce channel bandwidth and recent trials in the United States [Shelton et al., 1984], the United Kingdom [Bateman et al., 1987] and Japan [Daikoku et al., 1988] have compared selected narrow-band systems. The amplitude compandored single sideband (ACSSB) and real-zero single sideband (RZ SSB) technologies have been found suitable for both analogue and digital systems with possible applications as diverse as land, aeronautical and maritime mobile, satellite sound broadcasting and mobile satellite systems. The simultaneous use of more than one modulation type in the land mobile service is discussed in Report 1018.

2. NARROW-BAND METHODS

2.1 CONVENTIONAL SSB

Conventional SSB (eg emission J3E) is one of the most efficient techniques for reducing channel bandwidth, however, it is very sensitive to amplitude and phase fluctuations which make it unsuitable for mobile use due to the rapidly changing multipath environment. Conventional SSB is also susceptible to co-channel and adjacent channel interference. Many of these problems have been overcome by the use of advanced digital signal processing techniques which are now possible, at a reasonable cost, due to advances in VLSI technology.

2.2 PILOT TONE ACSSB

The transmission of a reduced carrier as a pilot tone carrier has long been the established technique for determining any frequency offset and regenerating the signal at the receiver. The amplitude and phase variations of the pilot carrier can also be monitored at the receiver and suitable inverse functions generated to remove the effects of multipath distortion from the signal. Feedforward Signal Regeneration (FFSR) is one technique for removing the channel imperfections caused by multipath fading. The advantage of this technique is that the channel can be made transparent for speech or data. The main disadvantage is that the pilot carrier is on the edge of the band where the amplitude and phase characteristics are least stable and, therefore, correlation between pilot and signal fading may not be high. The pilot, being on the edge of the band, is also susceptible to adjacent channel interference.

Tone above band (TAB) is used by some commercial systems as it allows some flexibility in the location of the pilot tone. However, in general, TAB systems have the same advantages and disadvantages as pilot carrier systems.

The tone in band (TIB) concept increases the correlation between pilot tone and signal by placing the pilot tone in the middle of the signal band. The pilot tone is inserted into a null created by removing part of the signal with a notch filter. A high correlation between pilot and signal fades is achieved and as the pilot is in the middle of the band, the adjacent channel protection is good. The main disadvantage of this technique is that the signal band is no longer transparent.

The transparent tone in band (TTIB) concept significantly reduces this disadvantage by splitting the signal band into two halves and moving them apart in the frequency domain to create a gap for the pilot tone. In the receiver the reverse process is employed to regenerate the signal and with an ideal implementation a transparent channel is maintained. This approach has the advantage that no information is lost and the separation between the two halves of the signal spectrum can be made variable to convey information such as the clocking reference frequency for the digital system. In practice TTIB systems can be implemented using digital signal processing to obtain a near transparent channel, however, there will be some distortion of the signal especially at mid band.

A switched tone system has been suggested, which overcomes the signal distortion by using a tone in band for signal acquisition but reverts to tone above band during signal transmission. This compromise system requires a complex frequency control circuit and does not provide the good correlation between signal and pilot fades provided by a tone in band system.

2.3 RZ SSB

i. RZ SSB is a combination of two extant technologies; SSB signaling and Phase Modulation (PM) reception. SSB signaling retains narrow-band properties and PM reception brings great immunity against fading.

ii. RZ SSB transmitters are similar to ordinary SSB transmitters except for carrier addition.

iii. The RZ SSB signal can be demodulated with an amplitude limiter, frequency detector, integrator and linearizer. The linearizer is required for removing intermodulation distortions generated from the demodulation process itself. Therefore, the RZ SSB is free from the degradations due to a mistuned carrier.

iv. The RZ SSB receiver can safely use the amplitude limiter for removing amplitude variation without any loss of information signal. The amplitude limiter is powerful enough to remove amplitude degradation caused by fading. Therefore, the RZ SSB does not require an Automatic Gain Control (AGC) circuit to provide amplitude correction under fading conditions."

v. Conventional SSB must be equipped with both AGC and Automatic Frequency Control (AFC) in order to maintain good demodulated signal quality during fading. The TIB scheme makes it possible to increase the correlation between pilot tone and signal fades. Good correlation is very important for the amplitude correction, that is, AGC operation.

vi. RZ SSB is a pilot carrier system, it does not require the TIB scheme because RZ SSB can demodulate the signal without using AGC and AFC schemes.

vii. The RZ SSB transceiver is transparent for the frequency range of public telephone, i.e., from 300 Hz to 3400 Hz.

viii. RZ SSB can be used for data transmission. The experimental results verified that voice-band MODEM signals with transmission speeds of 4.8 and 9.6 kbps are transmitted with good quality in fading and non-fading environments.

ix. RZ SSB has a high level of transparency which makes it possible to connect ordinary voice-band MODEMs which are available for very low cost. A transparent analogue transmission system like RZ SSB is very attractive, because the application of digital modulation like GMSK to mobile communication always requires specialized MODEMs which are expensive.

2.4 NARROW-BAND FM

The design of narrow-band FM (NBFM) is virtually identical to that of conventional FM with three exceptions:

- the peak deviation of the transmitter is reduced to ± 2.5 kHz;
- the IF filter and discriminator bandwidths are reduced; and
- the frequency tolerance is also reduced.

Conventional communication FM modulation equipments require channel bandwidths of 25 or 30 kHz while NBFM equipments utilize a channel bandwidth of 12.5 kHz.

3. MODULATION COMPARISONS

The two narrowband modulation techniques, NBFM and ACSSB, were found to be effective spectrum efficient modulation methods that can be implemented in the VHF and UHF bands allocated to the Fixed and Mobile Services. Although the testing was conducted in the VHF band, the results are believed to be also applicable to the UHF band. In an overall comparison of NBFM and ACSSB technologies for analog voice applications, each modulation method provided its own advantages and limitations.

NBFM offers expected advantages over ACSSB in such factors as interoperability with existing FM equipment, common acceptance testing procedures and test equipment and maintenance. The channel spacing associated with both techniques limits the data rate handling capability to values less than that of conventional FM. In addition, neither of the narrowband modulation methods can accommodate currently available digital voice techniques. The overall requirements of the user must be considered when choosing one or the other of the narrowband modulation methods. Since the equipments for the two narrowband modulation types are not uniformly available one or the other of the modulation types may offer more advantages in terms of acquisition and availability of maintenance.

The function of FFSR technique used in the ACSSB receiver can be resolved into two parts; a random Frequency Modulation (FM) noise canceller and feed-forward AGC. The RZ SSB receiver employs a random FM noise canceller whose fundamental configuration is the same as that in the FFSR scheme.

4. COMPARATIVE TESTING OF THE NARROW-BAND MODULATION TECHNIQUES

[Shelton et al., 1984] have conducted comparative testing of narrow-band land mobile systems to evaluate their usefulness for improving the spectrum efficiency of the land mobile service. All subsequent measured results are from [Shelton et al., 1984]. The tested systems were a conventional single-channel FM system with a nominal necessary bandwidth of 16 kHz, a NBFM system with a nominal necessary bandwidth of 11 kHz, and a ACSSB system with a necessary bandwidth of 3 kHz. The systems were tested in the laboratory and in the field. The comparative testing results are summarized in Table I.

The communication ranges for voice signals was similar for ACSSB, NBFM, and conventional FM, but with slightly lower voice quality for the narrowband modulations at the fringe of the communication range. Field tests conducted at equivalent powers over the same routes, and using the same base station antenna (omni whip, 3 dBi gain, 10 meters height) found that the range for good communication was 7.5 km for NBFM and 8.0 km for ACSSB and conventional FM. The communication ranges for the three different modulation types are comparable for the same equivalent power and terrain conditions.

TABLE I

COMPARATIVE FACTORS FOR NARROW-BAND LAND MOBILE TECHNOLOGIES

Comparative Factors	Evaluation Factors Conventional FM	ACSSB	NBFM
1. Channel Spacing (kHz)[a]	25	5 - 6.25	12.5
2. Spectrum Efficiency (Relative to 25 kHz FM)	1.0	2.5	1.8
3. Adjacent Channel[b] Protection Ratio Provided (dB)	-80	-47 to -58[c]	-80
4. Communications Range	Approximately the same for all sets		
5. Interoperability with Existing Equipment (conventional FM)	Yes	No	Yes, with minor performance degradation
6. Availability of Tone-Coded Squelch	Yes	Yes	Yes
7. Maintenance/Testing Procedures Compared to Existing Environment (conventional FM)	Same	Modified	Same
8. Standardized Equipment Design	Yes	No	Yes
9. Maximum Data Handling Capability of Audio Bandpass (Kilobauds) (for the same Audio Bandwidths)	1.2 - 1.8	1.2 - 1.8	1.2 - 1.8
10. Maximum Data Handling Capability Using Full Channel Spacing (Kilobauds) (direct modulation of the main carrier)	10 - 12	1.2 - 1.8	5 - 6
11. Channel Spacing Compatible with Current Digital Voice Techniques	Yes	No	No

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[a] Values in this Table are based on these channel separations.

[b] 25 kHz FM vs 25 kHz FM, ACSSB vs ACSSB and NBFM vs NBFM, respectively.

[c] -47 dB for 5 kHz channel spacing
-58 dB for 6.25 kHz channel spacing

The voice quality of both ACSSB and NBFM is similar to conventional FM in a high signal to noise environment. In a low signal to noise environment, the voice quality of both narrowband techniques is below that of conventional FM with the voice quality of the NBFM being slightly better than that of the ACSSB. In environments of low signal to noise, the NBFM voice signal is subject to distortion probably due to noise burst and capture, while the ACSSB voice signal is subject to distortion probably due to the companding, pilot tone circuitry, and reduced audio bandpass. ACSSB has a slight gravelly or raspy quality with less voice recognition. Its voice quality tended to degrade at a fairly constant rate with distance over the entire communication range. The NBFM system appeared to maintain a higher degree of voice quality for a greater distance. Signal distortion was more rapid as the range approached the maximum.

The dominant factor affecting the adjacent signal performance of conventional FM is considered to be its own receiver selectivity characteristics. The adjacent-channel protection offered by a conventional FM system with ±5 kHz peak frequency deviation is on the order of 80 dB.

The dominant factor affecting the adjacent channel signal performance of ACSSB was found by measurement and analysis to be the transmitter out-of-band emissions power (sideband spectrum). The out of band power for 5 kHz channels was measured to be 47 dB down and computer analysis estimated it to be 58 dB down for 6.25 kHz channels.

Improvements in transmitter linearity and pilot carrier processing appear feasible in this relatively young ACSSB technology. A change in channel spacings from 5 kHz to 6.25 kHz would offer 11 dB improvement in adjacent channel protection. At 5 kHz spacing, the important third-order intermodulation emission products fall within the passband of an adjacent receiver. At a 6.25 kHz separation, the third-order intermodulation products are substantially avoided. A 6.25 kHz spacing would be compatible with channeling schemes that involve 12.5 kHz NBFM and 25 kHz FM. However, a 6.25 kHz spacing would not be directly compatible with a 30 kHz FM channeling scheme.

NBFM equipments were found to have better than 80 dB adjacent signal protection at 12.5 kHz spacing. It is not always possible or practical to replace an entire communication network at one time. The introduction of NBFM into the environment offers advantages over ACSSB in that, with some performance degradation (approximately a 5 dB reduction in sensitivity compared to conventional FM), it is interoperable with existing conventional FM networks by adjusting the base and repeater stations for a peak deviation of ±2.5 kHz, with only a nominal reduction in communication range. When the conversion of mobile units is complete, the base and repeaters can be converted to complete the update. ACSSB radios are not interoperable with existing FM equipment.

The SINAD quality of RZ SSB appraised by a tone signal is similar to 12.5 kHz in fading and non-fading conditions. Laboratory tests using a 1 kHz tone modulating signal have shown that the subjective voice quality of RZ SSB is also similar to 12.5 kHz even in fading.

5. Spectrum efficiency comparison

Ideally, based on the bandwidth savings [Bateman et al., 1987] the spectrum efficiencies without frequency reuse of AC SSB and RZ SSB may be four to five times that of conventional 25 kHz systems. However, it is expected [Shelton et al., 1984] that the practical spectrum efficiencies of AC SSB are in the range 1.0 to 2.5 times as spectrum efficient as conventional FM. The range of the NBFM spectrum efficiencies is in the range 1.7 to 1.8, as efficient as conventional FM

assuming the co-channel protection ratio of NBFM and conventional FM are the same. These practical spectrum efficiency values take into account providing adjacent channel protection in environments of varying amounts of congestion which require frequency reuse. The expected spectrum efficiency practical values were calculated using the methods of Report 662.

Daikoku et al., 1988 calculates the spectrum efficiency for a 5 kHz channel spacing RZ SSB system to be two times that of 12.5 kHz FM for a frequency resuse system which includes co-channel protection ratios for RZ SSB = 13 dB and 12.5 kHz FM = 10.5 dB. The calculation did not include taking account of adjacent channel protection.

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REPORT 1114

AN APPLICATION OF THE CONCEPT OF SPECTRUM EFFICIENCY TO RADIO-RELAY SYSTEMS

(Question 47/1)

(1990)

1. Introduction

The choice of digital radio-relay systems for a given application involves many factors, including technical, economic and regulatory. Since the spectrum is a limited natural resource, the spectrum management community has a major interest in identifying spectrum conservation techniques that will provide more efficient spectrum utilization. Efficient use of the spectrum involves consideration of tradeoffs in technical design factors.

This Report assesses the spectrum conserving properties of advances in new technology for digital radio-relay systems in antennas, modulation schemes and signal processing. The concept of spectrum use and efficiency given in CCIR Report 662-2 was used. CCIR Report 662-2 defines spectrum utilization efficiency as the ratio of communications achieved to the spectrum-space used. Spectrum-space is defined as the product of time, bandwidth, and the spatial volume (area) denied to other users due to potential interference. For this study on analysis of spectrum conservation techniques for fixed microwave systems, a computer model [1] was developed.

The system parameters used in the analysis have been selected to provide an indication of a single path analysis of spectrum efficiency for the various cases considered. As such, the parameters may not be representative of realizable systems, particularly those using higher order modulation schemes. The results therefore are illustrative of an application of the concept of spectrum efficiency to radio-relay systems, and administrations should employ representative parameters in any spectrum efficient analysis.

2. Analysis Approach

To assess the spectrum conserving properties of various design factors or technology options, the relative spectrum utilization efficiency that can be achieved must be quantitatively evaluated. For the study the concepts in CCIR Report 662-2 were extended and defined as the spectrum utilization efficiency (SUE):

$$SUE = VC/(T \cdot A \cdot B) \quad (1)$$

where: VC = Number of voice channels

T = Fraction of time a system is used (defined to be equal to 1 for this analysis)

A = Denial area (km^2)

B = Occupied bandwidth (MHz)

Equation (1) was chosen because it takes into account both spectrum and spatial (area) denial in assessing the spectrum-conservation properties of a system. The denial area is the area in which another system can not operate without degradation in system performance below a specified performance criteria. The denial area is a function of the system antenna-pattern characteristics, transmitter output power and the receiver interference threshold level.

The algorithm used to calculate the denial area involves the segmentation (quantization) of the transmitter antenna gain pattern into a number of segments, angular sectors, which accurately represent the antenna pattern. The transmitter antenna gain pattern is an input to the model. The program calculates the denial area by summing the area in each segment. Geometrically, each segment is a angular sector, the area of which may be calculated using the formula:

$$\text{Area of angular sector} = \pi R^2 \theta / 360 \quad (2)$$

where: R = Radius of sector (R_1, R_2, ..., R_n).
θ = Vortex angle of sector (θ_1, θ_2, ..., θ_n).
n = Number of angular segments.

The radii R_n for each segment were calculated using the relationship:

$$L(R) = P_t + G_t(n) + G_r - I_{MAX}$$

where: $L(R)$ = Required propagation loss, in dB.
P_t = Transmitter output power, in dBm.
$G_t(n)$ = Transmitter antenna gain for sector n, dBi.
G_r = Receiver antenna gain, 0 dBi non-directional.
I_{MAX} = Maximum permissible interference level, in dBm.

Then using a smooth earth inverse propagation model, the distance R corresponding to the required loss is determined. This facilitates the evaluation of the denial area for each angular sector (Equation 2).

To apply Equation (1) to point-to-point radio-relay systems, it is necessary to establish characteristics of a reference system between two microwave sites. These characteristics include path length, path attenuation, antenna gain, insertion losses, fade margin and system gain. It is also necessary to establish certain modulation characteristics for the modulation types addressed. The digital modulations considered in this investigation are 16, 64 and 256 quadrature amplitude modulation (QAM). The characteristics assumed for the digital radio-relay systems for this analysis are based on the North American Standard and are as follows:

DIGITAL RADIO-RELAY SYSTEM PARAMETERS

1.	VOICE CHANNELS (VC)	=	1344 for 16-QAM 2016 for 64-QAM 2688 for 256-QAM
2.	BIT RATE (BR)	=	90 Mb/s for 16-QAM 135 Mb/s for 64-QAM 180 Mb/s for 256-QAM
3.	BIT ERROR RATE (BER)	=	10^{-6}
4.	RECEIVER NOISE FIGURE (F)	=	4 dB
5.	SYSTEM GAIN (G_s)	=	103 dB

The analysis utilized theoretical transmission efficiency and carrier-to-noise ratio $(C/N)_t$ levels for the different modulation types to ensure a just comparison.

The following is a discussion of the application of Equation (1) to the major design areas of antennas, modulation and signal processing.

3. Antennas

Spatial denial is a key factor in addressing spectrum conservation. One of the major radiocommunications system components contributing to the spatial denial is its antenna. In recent years, significant advances in the antenna-design areas of polarization discrimination and sidelobe reduction have provided the capability for enhanced spectrum efficiency in point-to-point microwave radiocommunications.

Frequency re-use can be achieved by implementing antenna-design spectrum-conservation techniques. Spatial denial can be minimized if sidelobe levels are minimized. The antenna radiation patterns, and therefore sidelobe distributions, vary with antenna type. Three antenna types commonly used in point-to-point microwave transmission are:

1. Standard-dish (STD)
2. Shrouded-dish (SHD)
3. Conical horn reflector (CHR)

Typical radiation patterns for these antennas, with a 43 dBi gain, are shown in Figure 1. The antenna pattern characteristics shown in Figure 1 were used in the model.

3.1 Spectrum Utilization Efficiency

A plot of the transmitter output power versus denial area for a receiver interference threshold of -102.5 dBm is shown in Figure 2 for the three types of antennas. Although, the mainbeam gain for all the antennas is the same, the results shown in Figure 2 indicate that the CHR antenna has less denial area than the other two types of antennas. Also the difference in denial area for the three types of antennas is small until the transmitter power is greater than 30 dBm. This is understandable because the contribution to the denial area caused by sidelobe/backlobe antenna characteristics is small until the transmitter power is increased beyond 30 dBm. At transmitter output powers greater than 30 dBm, the difference in denial area for the three types of antennas is significant. The denial area is also a function of the receiver interference threshold.

Since the denied area for the three types of antennas is a function of the transmitter output power and the maximum permissible interference level (I_{MAX}), the spectrum conserving properties for the three antenna types must also be related to system modulation type. Thus the spectrum efficiency enhancement properties of the STD, SHD and CHR antennas will be discussed in the modulation section.

4. Modulation

The evaluation of spectrum conservation properties for different modulation schemes is very complex in that both spectrum and spatial denial are affected by the choice of modulation type used in a system. In general, system parameters, such as occupied bandwidth, required receiver input carrier-to-noise ratio $(C/N)_I$, and maximum permissible interference level are all functions of the modulation type and have a direct bearing on spectrum utilization.

The analysis is based on theoretical transmission efficiency and receiver input carrier-to-noise ratios for the different modulation types to ensure a just comparison. To evaluate the spectrum-conservation properties of the different modulation types, the occupied bandwidth (B) and required transmitter output power level (P_t) for each were determined. These parameters are given in TABLE 1.

The occupied bandwidth (B) for the digital modulations was determined using the relationship:

$$B\ (MHz) = \text{Bit Rate (Mb/s)/Transmission Efficiency (b/s/Hz)} \quad (3)$$

where the Bit Rate (See digital system parameters) and the Transmission Efficiency (See TABLE 1) are functions of the modulation type.

To establish the required transmitter output power for each modulation type, the required receiver input, $(C/N)_I$, for specified performance criteria, was determined. A bit error rate (BER) of 10^{-6} was used as the performance criteria and the theoretical required $(C/N)_I$ was obtained from literature.

The receiver input noise levels, N_I, given in TABLE 1 were determined using a receiver bandwidth of 22.5 MHz and a receiver noise figure of 4 dB. The required minimum carrier level (C_{MIN}) at the receiver input was then determined from the relationship:

$$C_{MIN}\ (dBm) = (C/N)_I + N_I \quad (4)$$

The required transmitter power (P_t) given in TABLE 1 was determined using the expression:

$$P_t\ (dBm) = C_{MIN} + G_S \quad (5)$$

where G_S represents system gain, which was set equal to 103 dB.

The denied area is also a function of the victim receiver maximum permissible interference level (I_{MAX}). The receiver maximum permissible interference level associated with each modulation was determined assuming that the victim receiver has the same modulation type as the interfering transmitter. In this analysis, the maximum permissible interference level was determined using the criteria established in the Electronic Industries Association Standard (EIA Bulletin No. 10-D). For the digital systems, the performance criteria was an increase in BER from 10^{-6} to 10^{-5}, which corresponds to approximately a 1 dB increase in receiver noise level. This is equivalent to an interference-to-noise ratio (I/N) = -6 dB (i.e., I_{MAX} = -102.5 dBm for 16-QAM, 64-QAM and 256-QAM).

4.1 Spectrum Utilization Efficiency

TABLE 2 contains the calculated spectrum utilization efficiency (SUE) values using the bandwidth and transmitter output power given in TABLES 1 and the maximum permissible interference level of -102.5 dBm. The entries for SUE in TABLE 2 are for three different modulation types and three types of antennas (STD, SHD, and CHR). Systems with higher SUE values are more efficient from the spectrum utilization point-of-view. It should be emphasized that the calculated results clearly point out that, for the SUE varies considerably from one antenna type to another. For example, the SUE for 64QAM is 0.044 for the STD antenna as compared to 0.073 and 0.478 for the SHD and CHR antennas, respectively. Therefore, the results shown in TABLE 2 clearly indicate that SUE can be optimized only when the effects of the antenna and modulation are both considered.

Also, the analysis results show the SUE for 64-QAM to be higher than that for 256-QAM for all three antenna types. Hence the analysis results indicate that 64-QAM is a more "spectrum efficient" modulation from the spectrum utilization point-of-view. TABLE 3 helps provide an explanation of why 64-QAM is more "spectrum efficient" than 256-QAM for an ultra high performance (SHD) antenna. The input parameters to the model are provided in the table. The number of voice channels (VC) is 2016 for 64-QAM and 2688 for 256-QAM. The required system bandwidth, B, is the same for both 64 and 256 QAM (B = 22.5 MHz). However, the required transmitter power level, P_t, for 256-QAM is significantly higher than 64-QAM (39.6 dBm as compared to 33.5 dBm). Since the transmitter power has a major effect on denied area to another user (see Figure 2), the denied area for 256-QAM is significantly more than for the 64-QAM modulation thus causing the 64-QAM modulation to be more spectrum efficient than 256-QAM.

However, the spectrum conserving potential of a system is a function of several design factors all of which must be taken into consideration when evaluating the spectrum efficiency of a system. That is one can not say that a system with a particular modulation is more spectrum conserving than a system with another modulation without considering all other design factors (e.g., antennas, signal processing, RF filters, etc.).

TABLE 2 can also be used to determine the relative improvement in spectrum conservation of using a shrouded dish (SHD) antenna or conical horn reflector (CHR) antenna over a standard dish (STD) antenna. As stated earlier, the improvement in spectrum conservation for the SHD and CHR antenna is dependent on the modulation type. This is due to the fact that the denial area produced by a particular antenna type is a function of the required transmitter output power (P_t) which is modulation dependent (See Figure 2). TABLE 4 shows the percentage improvement in Spectrum Utilization Efficiency (SUE) for the various modulations addressed using the SUE data in TABLE 2. For the digital modulations, TABLE 4 clearly shows that as the digital modulation order increases the spectrum conservation improvement increases for a particular antenna type. The greatest improvement occurs for 256-QAM modulation with a 130 and 1500 percent increase for the shrouded dish (SHD) and conical horn reflector (CHR) antennas respectively.

5. Signal Processing

In a fixed radio-relay systems, signal processing is done at the transmitter and receiver terminal. Signal processing consists of electrical operations on a signal in order to produce certain desired characteristics. Signal processing can affect such parameters as amplitude, frequency, phase, signal level and reliability. The use of signal processing techniques can improve the processing gain of a system, permitting lower transmitter output power levels for specified receiver output performance criteria. Thus, through the use of signal processing techniques, the transmitter output power can be lowered reducing the spatial (area) denied to other systems. However, it should be noted that signal processing techniques are used by the

microwave link designers to improve link reliability and are not generally considered for the purpose of spectrum conservation.

5.1 Error Correction/Coding

Forward error correction (FEC) coding is a method of improving the BER performance of digital microwave systems, particularly when the system is power limited. The utilization of FEC coding techniques permits a limited number of errors to be corrected at the receiving end by means of special coding and software (or hardware) implemented at both ends of a circuit. This improvement in BER can be traded off for a reduction in required receiver input carrier-to-noise, $(C/N)_i$, ratio to meet a specified BER performance, thus reducing the denial area to other systems. The reduction in $(C/N)_i$ is referred to the coding gain. The performance of a coding technique is described by the coding gain and coding rate. However, the coding rate has an impact on the system occupied bandwidth, thus increasing the denied spectrum to other users of the spectrum.

To show the effect of coding on spectrum conservation, 64-QAM was selected as the modulation for study. Four types of FEC codes were selected. TABLE 5 shows the coding rate, R, the bandwidth expansion factor, 1/R, the occupied bandwidth after coding, the obtainable reduction in $(C/N)_i$ for a BER of 10^{-6} and the required transmitter output power, P_t, after taking into consideration the obtainable reduction in $(C/N)_i$. The values for occupied bandwidth, B, and transmitter output power, P_t, shown in TABLE 5 were input to the spectrum utilization efficiency (SUE) model to evaluate coding as a spectrum conservation technique. TABLE 6 shows the SUE for the three types of antennas: STD, SHD and CHR. The SUE for 64-QAM without coding is also shown in the TABLE 6 for a baseline comparison of with and without coding.

In summary, the SUE values given in TABLE 6 indicate that signal processing techniques such as error correction/coding which utilize RF bandwidth versus carrier-to-noise trade-offs only provide improvement in spectrum conservation, higher SUE values, when high-efficiency coding techniques (i.e., coding techniques with high coding rates and coding gain) are used. Also the relative improvement in spectrum conservation is greater when the system has an STD type antenna than an SHD or CHR type antenna. This is due to the fact that the reduction in denied area is greater for standard dish antennas because of the higher sidelobe/backlobe characteristics.

5.2 Adaptive/Transversal Equalizers

Adaptive/transversal equalizers improve the digital system performance in the presence of multipath fading, linear distortion, or both. The equalizers can only mitigate the dispersive aspects of multipath fading. These adaptive equalizers reshape the pulse so as to minimize the intersymbol interference. An approximate 4 to 6 dB improvement in the composite fade margin can be achieved with these equalizers in 64-QAM receivers. The major drawback of adaptive equalizers is their expense. The model was run for a system bandwidth, B, of 22.5 MHz and a transmitter output power, P_t, of 29.5 dBm (A 4.0 dB reduction in P_t for 64-QAM). TABLE 7 shows the SUE for the three types of antennas: STD, SHD and CHR. The SUE values without adaptive equalizers is also shown in the table for baseline comparison with adaptive equalizers.

For 64 QAM, The use of adaptive equalizers can improve the spectrum conservation properties of a system from approximately 30 to 90 percent with the greatest improvement in systems that use STD type antennas.

5.3 Error Correction/Coding and Adaptive Equalizers

Some digital systems utilize both error correction/coding and adaptive equalizers to improve system performance. For 64-QAM, the utilization of error correction/coding (18/19 code rate) and adaptive equalizers can reduce the required input carrier-to-noise, $(C/N)_i$, ratio by 7 dB for BER = 10^{-6}. The application of the Spectrum Utilization Efficiency (SUE) model for a occupied bandwidth, B, of 23.74 MHz and a transmitter output power, P_t, of 26.5 dBm (A 7.0 dB reduction in P_t for 64-QAM) is shown in TABLE 8. TABLE 8 shows that the use of error correction/ coding and adaptive equalizers can improve the spectrum conservation properties of a system from 45 to 190 percent with the greatest improvement in systems that use dish type antennas.

6.0 Results

The following are the significant results of the study.

1. The spectrum utilization efficiency (SUE) defined in CCIR Report 662-2 is a technical concept that can be used to quantitatively assess the spectrum conserving potential of new technologies and to develop new spectrum standards.

2. The spectrum conserving potential of a system is a function of several design factors all of which must be taken into consideration when evaluating the spectrum utilization efficiency of a system. That is one can not say that a system with a particular modulation is more spectrum conserving than a system with another modulation without considering all other design factors such as antennas, signal processing, RF filters, etc.

3. The higher order digital modulations (modulations with higher transmission efficiency, b/s/Hz) require higher transmitter output power levels. Therefore, when the CCIR definition of spectrum use and efficiency which takes into consideration denied area is used, modulations which have higher transmission efficiency (b/s/Hz) may not necessarily be more spectrum conserving. Thus the transmission efficiency (bits/s/Hz) of the digital modulation may not suffice as an indicator of spectrum efficiency.

7.0 References

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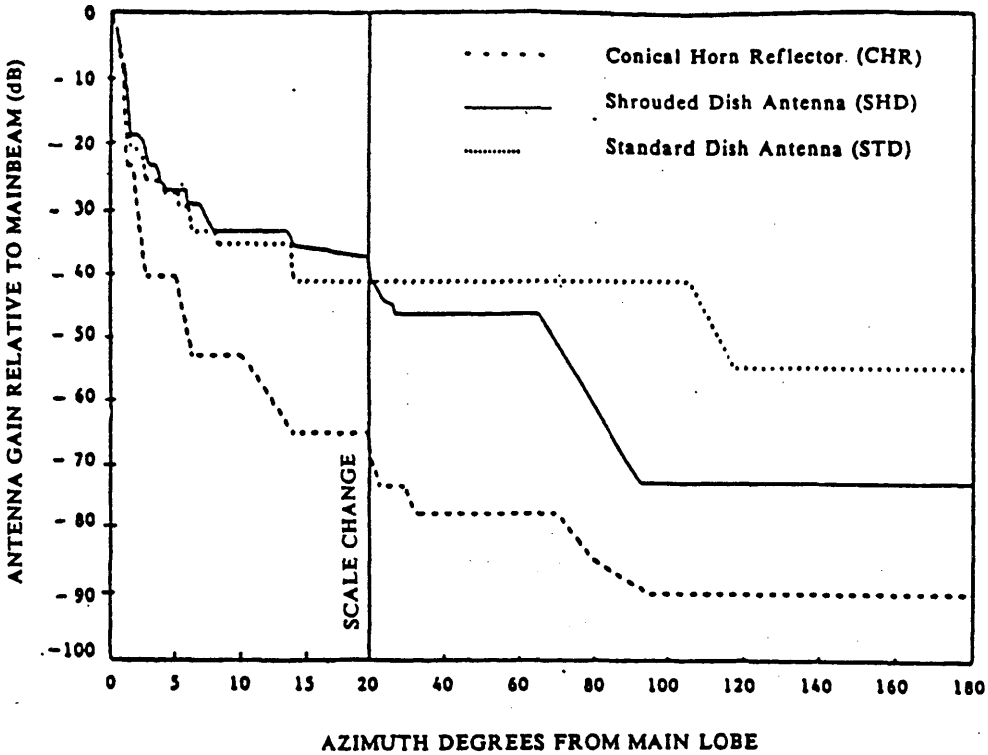


Figure 1. Typical radiation patterns for STD, SHD AND CHR antennas.

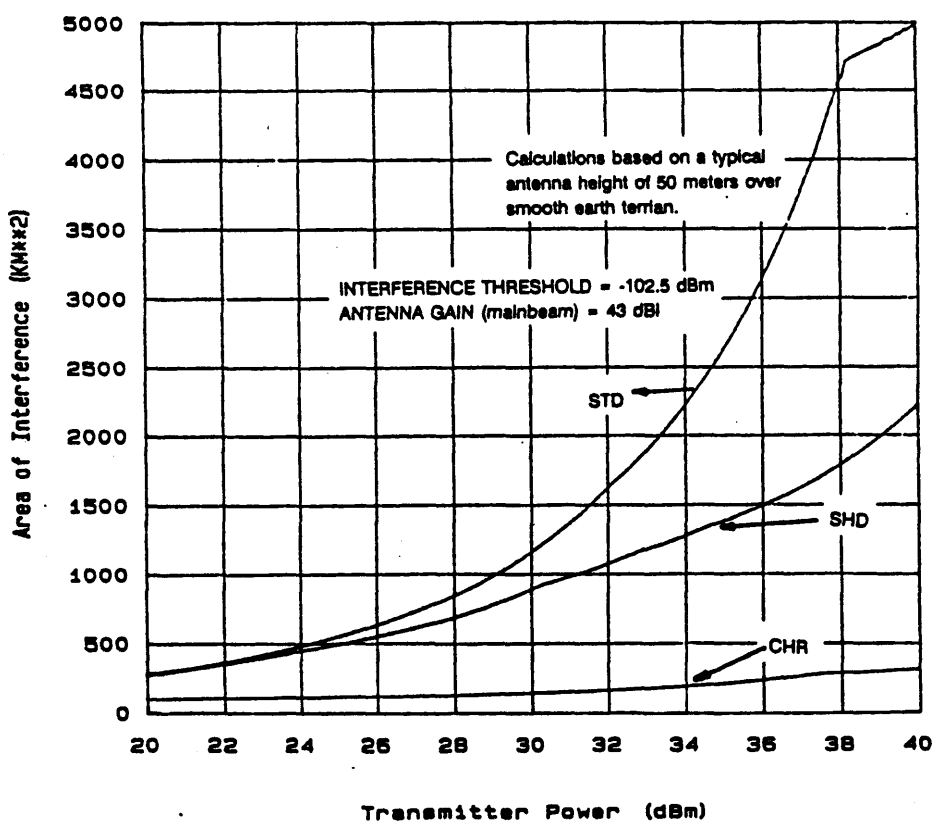


Figure 2. Denied area s a function of antenna type ans transmitter output power for an interference threshold of -102.5.

TABLE 1

DIGITAL SYSTEM PARAMETERS

Modulation Type	Transmission Efficiency (b/s/Hz)	Occupied Bandwidth (MHz)	Required Carrier-To-Noise (dB)	Noise Level (dBm)	Minimum Carrier Level (dBm)	Transmitter Power Level (dBm)
16-QAM	4	22.5	21.0	-96.5	-75.5	27.5
64-QAM	6	22.5	27.0	-96.5	-69.5	33.5
256-QAM	8	22.5	33.1	-96.5	-63.4	39.6

TABLE 2

SPECTRUM UTILIZATION EFFICIENCY

Ranking Order	Spectrum Utilization Efficiency For Different Antenna Types		
	STD (SUE)	SHD (SUE)	CHR (SUE)
1	16-QAM (0.077)	16-QAM (0.093)	64-QAM (0.478)
2	64-QAM (0.044)	64-QAM (0.073)	16-QAM (0.464)
3	256-QAM (0.024)	256-QAM (0.056)	256-QAM (0.390)

TABLE 3

SPECTRUM UTILIZATION EFFICIENCY COMPARISON OF
64- AND 256- QAM FOR SHD ANTENNA

PARAMETER	64-QAM	256-QAM
VC	2016	2688
B (MHz)	22.5	22.5 (See TABLE 1)
P_t (dBm)	-33.5	-39.6 (See TABLE 1)
I_{MAX} (dBm)	-102.5	-102.5
A (km^2)	1226	2120 (See Figure 2)
SUE	0.073	0.056 (See TABLE 2)

TABLE 4

SPECTRUM UTILIZATION EFFICIENCY IMPROVEMENT
FOR
SHROUDED DISH AND CONICAL HORN REFLECTOR
ANTENNAS AS A FUNCTION OF MODULATION TYPE

Modulation Type	Improvement In Spectrum Utilization Efficiency (SUE)	
	Shrouded Dish (SHD)	Conical Horn Reflector (CHR)
16-QAM	20%	500%
64-QAM	65%	990%
256-QAM	130%	1500%

TABLE 5

ERROR CORRECTION/CODING (64-QAM MODULATION)

Signal Processing	Coding Rate	Bandwidth Expansion Factor	Occupied Bandwidth MHz	Reduction In C/N (dB)	Transmitter Power (dBm)
Error correction coding	1/2	2	45.00	6	27.5
	3/4	1.333	30.00	3.5	30.0
	7/8	1.142	25.70	2.0	31.5
	18/19	1.055	23.74	3.0	30.5

TABLE 6

SPECTRUM UTILIZATION EFFICIENCY FOR ERROR CORRECTION/CODING (64-QAM MODULATION)

		Spectrum Utilization Efficiency For Different Antenna Types		
Signal Processing Type		STD (SUE)	SHD (SUE)	CHR (SUE)
Without signal processing		0.044	0.073	0.478
	Coding Rate			
Error correction coding	1/2	0.057	0.069	0.348
	3/4	0.058	0.076	0.460
	7/8	0.053	0.076	0.488
	18/19	0.067	0.090	0.564

TABLE 7

SPECTRUM UTILIZATION EFFICIENCY FOR
ADAPTIVE EQAUALIZER (64-QAM MODULATION)

Signal Processing Type	Spectrum Utilization Efficiency For Different Antenna Types		
	STD (SUE)	SHD (SUE)	CHR (SUE)
Without signal processing	0.044	0.073	0.478
With adaptive equalizers	0.084	0.108	0.631

TABLE 8

SPECTRUM UTILIZATION EFFICIENCY
ERROR CORRECTION/CODING AND ADAPTIVE EQUALIZERS
(64-QAM MODULATION)

Signal Processing Type	Spectrum Utilization Efficiency For Different Antenna Types		
	STD (SUE)	SHD (SUE)	CHR (SUE)
Without signal processing	0.044	0.073	0.478
Error correction/coding and equalizers	0.126	0.147	0.688

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SECTION 1C: SPECTRUM MONITORING TECHNIQUES

REPORT 1105

THE RELATIONSHIP BETWEEN SPECTRUM MANAGEMENT AND RADIO FREQUENCY MONITORING

(Decision 27-2, Question 29/1)

(1990)

1.0 INTRODUCTION

Decision 27-2 seeks information about the necessary principles and techniques for efficient spectrum management, including specification of functional requirements for data management and frequency assignment. The Handbook on Spectrum Management and Computer-aided Techniques, the Handbook for Monitoring Stations and the IFRB/CCIR "Booklet on National Frequency Management" describe the basic functions of frequency management and monitoring. These functions include the key areas of frequency assignments, record keeping and monitoring. The "Second Meeting On The Development of National Radio Frequency Management" addressed the problem of an implementation strategy for the development of a spectrum management capability. This strategy indicates that the spectrum management functions should be prioritized, the spectrum management process should be sized to identify the resources required and that a time-phased plan to implement the spectrum management functions should be developed. This report addresses the problem of implementing a spectrum management system with particular emphasis on spectrum monitoring requirements and the integration of spectrum monitoring data with the basic objectives of spectrum management.

2.0 GENERAL

2.1 Spectrum Management Functional Requirements

Annex VIII to the ——— Handbook on Spectrum Management and Computer-Aided Techniques describes the interrelationships among various elements of the functional requirements of spectrum management. This Annex indicates that the spectrum management functional requirements generally consist of licensing, field inspections, license examinations, spectrum monitoring, international relations and coordination, spectrum planning, the development of spectrum standards and frequency assignment and record keeping. In order to effectively implement a national

spectrum management system a number of subjects relative to these elements need to be addressed. These include to:

1. prioritize the functions that should be carried out

2. estimate the magnitude of the resources required to implement the process

3. develop a time phased implementation plan that identifies near term and far term goals

4. consider automation support and a timeframe for its introduction.

These implementation factors should be considered relative to the functional requirements described above. This report addresses factors relating to monitoring and key elements of the spectrum management process. These key elements are the use of monitoring data to support the spectrum management process in the areas of frequency assignment and planning. It is critical in this context that the frequency assignment process be operational, that data file requirements be defined and that near term and far term planning requirements be established. These key areas should be implemented and their relationship between monitoring and spectrum management clearly defined before a monitoring activity is implemented. If these key elements of the spectrum management process have not been adequately implemented there is only a limited requirement for spectrum monitoring data and large resources spent for monitoring activities will not be effectively used.

2.2 Radio Frequency Monitoring

The purpose of radio frequency monitoring is to support the, spectrum management process in general, and frequency assignment and planning functions. Radio frequency monitoring accomplishes this by supplying real-time measured information that can be used to determine compliance with the technical and operational standards set by the spectrum management process. The spectrum management programme also supports the monitoring process by providing accurate and complete data on current assignments and by making specific job assignments to the monitoring activity. Monitoring is used to perform the following functions:

1. general measurements of channel and band usage, including assessment of channel availability and the effectiveness of frequency assignment procedures

2. resolution of interference problems

3. compliance with spectrum management policies, including

a. verification of proper technical and operational characteristics of transmitted signals

b. detection and identification of illegal transmitters

4. generate and verify frequency records.

Measurement of radio spectrum usage is an expensive process, and the results are limited to the radio signals received by the measurement system. Therefore, monitoring is usually undertaken to detect unauthorized use and to obtain information that is not available from the normal spectrum management files or from analysis and simulation. As the following list shows, there are important connections between monitoring and spectrum management. For this reason, close coordination should be maintained between the two offices. The principle areas of interaction between monitoring and spectrum management include the following.

1. Spectrum management provides official lists of assigned frequencies to monitoring. Signals not on this list can be assumed to be operating unauthorized and are of special interest to monitoring.

2. Spectrum management provides general guidance to monitoring on bands of interest, level of support required, suspected problem areas, and future monitoring requirements.

3. Spectrum management provides specific job requests to monitoring involving:

    a. interference complaints for monitoring to resolve

    b. special EMC problems, where specific measurements are needed

    c. usage measurements on proposed assignable frequencies

    d. measurements required for the IFRB and CCIR.

4. Monitoring provides usage information to frequency management on:

    a. clear or least used frequencies for assignment

    b. general band crowding assessments.

5. Monitoring provides feedback to frequency management on the general state of technical and administrative discipline in radio bands, in addition to identifying particular problem areas.

6. Monitoring supports and enhances frequency management decisions by identifying transmitters not in compliance with rules and allowing appropriate action to be taken.

In general, spectrum management should provide guidance and direction to monitoring, especially in the area of long-term planning for the type and amount of support needed from the monitoring programme. It is vital that continuing close coordination be maintained between these organizations to ensure that the monitoring activities performed are most useful to frequency management.

Monitoring activities can provide a wide range of information to support spectrum management activities. The following sections contain a description of several types of information derived from monitoring and the measurement systems used to obtain that data.

3.0 SPECTRUM USAGE DATA

Usage data indicates how much of the time a particular frequency has had a signal present, either for a single frequency or for all of the frequencies in a frequency band. Individual usage measurements on a single frequency can be combined to show how usage varies during a 24-hour day (including hours of peak and minimum usage). Note that frequency assignment records indicate only that the use of a frequency is authorized, but may not give adequate information on whether the frequency is actually used. Therefore, a frequency band that appears crowded on the basis of frequency assignment records may or may not in fact be crowded. Usage data from many frequencies can be combined to show the average usage for all frequencies in a band or for a selected group of users.

Usage information is a valuable tool for several spectrum management functions. This information can be used to identify unused channels in a band or to prohibit adding more assignments to heavily used channels. It can also be used to prompt an investigation when usage is present on channels that are unassigned according to the frequency management records or when no usage is seen on frequencies with assignments. Finally, this information can be used to plan for the allocation of additional bands when existing bands become too crowded.

Spectrum usage measurements will be particularly helpful in the land mobile radio (LMR) bands and similar channelized communication bands. Other mobile bands would also be good

candidates for usage requirements. Standards of use would be established by the SM process for reference purposes in each of the services involved.

Usage data is obtained by measuring at a frequency or across a band of frequencies to determine the amount of time a signal is present at each frequency. A wide range of measurement techniques can be employed to measure usage, each tailored to the signals being measured and the measurement technology at hand. A computer-controlled measurement system, for example, can detect channel activity in the LMR bands at the rate of more than 100 channels per second. A system using a stable receiver tuned to a single frequency can be used with a carrier-operated audio recorder to demodulate and record any usage on that frequency. Although such a system gives usage data on only one channel at a time, it provides the additional advantage of yielding valuable information on the identification of users on the channel.

The relationship between monitoring information and the spectrum management records is not straightforward. Usage information tells only that a frequency is used. It does not indicate which transmitter is producing the signal. The existence of an assignment at a particular frequency and the presence of a signal at that frequency does not necessarily mean that the measured signal was transmitted by the assigned transmitter. Aural monitoring for call signs or similar information to identify a transmitter is needed to resolve this ambiguity. In addition, the absence of a measured signal does not necessarily indicate the absence of a frequency assignment. An assigned transmitter may not have been used during the monitoring period, or it may have been used outside of the limited reception range of the monitoring station.

4.0 VERIFICATION OF TECHNICAL AND OPERATIONAL PARAMETERS

Monitoring is used to obtain detailed information on the technical or operational characteristics of radio systems. This might typically include a detailed measurement of the emission spectrum of a transmitter or the antenna pattern. These measurements can be made to provide information needed in a particular EMC analysis, to verify conformance with the characteristics specified in a particular frequency assignment record, or as part of a type-acceptance process to ensure that a particular type of equipment will operate compatibly with other equipments in the frequency band. Finally, measurements might be made merely to ascertain that a given transmitter was operating in accordance with specified limits.

Although many types of technical parameters can be measured, probably the most important one is the emission spectrum of a transmitter. A measurement technique must be selected that will allow various types of signal modulation to be measured in a

quantitatively useful way. Thus, a measurement system should have a variety of bandwidths and video signal processing, selected individually for the signal being measured.

5.0 RESOLUTION OF INTERFERENCE AND UNAUTHORIZED TRANSMITTERS

Frequency monitoring measurements are useful in resolving the cause of interference to authorized transmitters. Such measurements may detect the presence of unauthorized transmitters causing the interference or more subtle conditions resulting from a combination of transmitters, unintended spurious emissions, etc. Although a wide variety of combinations of spectrum measurements and engineering analyses may be needed to resolve some types of interference, spectrum measurements often play a key role in this process. Aural monitoring will often be useful in learning the identity of transmitters involved in the interference.

Unauthorized transmitters may be suspected as the cause of interference. For this reason, there will often be a close association with the ability to solve interference problems and the ability to detect and identify unauthorized transmitters. A major problem with the detection of unauthorized transmitters is that it is often difficult to separate the legal signals from those that are unauthorized. This is especially difficult in crowded bands where unauthorized transmitters share the same frequencies.

Detection of unauthorized signals can sometimes be accomplished by aural monitoring on frequencies where authorized users have complained of interference or where frequency assignment records show no legally assigned users. Direction-finding, mobile-tracking vans, and intelligence from aural monitoring are helpful in identifying and locating illegal transmitters after they have been detected.

6.0 EFFECTIVE USE OF MONITORING INFORMATION

Automated data bases can be used to improve the effectiveness of frequency monitoring in two major ways. First, the value of data derived from monitoring can be enhanced by making it easily available to different frequency management processes. For example, spectrum usage data that is easily available to help make interference-free frequency assignments can also be used to get statistics on the growth of frequency use. Second, automated data bases can be used to increase the effectiveness of the monitoring process itself, by helping to efficiently sort out signals for further study (comparison of usage with assignments) and by providing a more efficient way to record and process the results of the inspections on selected signals.

This means that the use of data bases and computers should be encouraged, and that many of the monitoring records that are kept in log books should, when the record size is large, be kept on a computer in a form that is more easily manipulated and sorted by data base programs. Table IV-V of the Handbook on Spectrum Management and Computer-Aided Techniques describes frequency monitoring data used for manual monitored emissions. This Annex also describes other data used by spectrum management that should be coordinated with frequency monitoring (e.g., allocation data, frequency assignment data, equipment characterized data etc). Equally important, there should be a continuing close coordination between monitoring and spectrum management to ensure that data is recorded in compatible formats to allow easy use of the same information by both offices.

7.0 SUMMARY

Spectrum Management involves the operation of a number of functional units that includes frequency assignment, record keeping, near and far term spectrum planning, setting technical standards and radio frequency monitoring. It is critically important in implementing a spectrum management system that the first priority is to make the frequency assignment process operational, to specify data file requirements. If these key elements of the spectrum management process have not been adequately implemented there is only a limited requirements for spectrum monitoring data and large resources spent for monitoring activities will not be effectively used. The ratio frequency monitoring activity supports the spectrum management process by mainly supplying information for frequency assignment and spectrum planning activities.

REPORT 1115

TECHNIQUES OF RADIO MONITORING AS RELATED TO SPECTRUM MANAGEMENT

(Question 32/1)

(1990)

1. Introduction

Radio frequency spectrum is a "limited" natural resource which, unlike other natural resources, is "used" and not "consumed" thus, making it "non-depleting". Its non-use or improper, unplanned and inefficient use results in waste of the resource. Though the spectrum may appear to be extensive, but, in practical terms, it is "limited" due to various considerations like constraints of available technology, bewildering array of radio services, astronomically increasing demand on the spectrum for multifarious types of usages and propagational/operational constraints of different frequency bands, geographical terrain, etc. It is, therefore, essential that radio frequency spectrum is used in a planned, orderly and efficient manner so that not only that the present requirements are catered for but the future demands on the spectrum are also adequately met and all the operations can co-exist in an interference-free environment. Spectrum management is a unique, highly specialized and distinct discipline having regulatory, legal and technical inter-related dimensions with national as well as international commitments.

2. Role of radio monitoring in spectrum management

Radio monitoring is one of the most essential features of spectrum management. Since usage of spectrum is susceptible to interference it is essential that planning, standardization, coordination, assignment and other functions are based on technical studies and data as well as operations are constantly checked, observed and overseen. The radio spectrum is generally given the dimensions of space, time and frequency and, thus, monitoring of various emissions is necessary to ensure interference-free operation of various services. Monitoring also provides data for effective planning and management of spectrum, optimum parameters for satisfactory operation of circuits, site selection and other allied functions. Cases of harmful interference, detection of unauthorized transmissions, determination of occupancy, vacancy and circuit loading of frequencies also need field data by way of monitoring.

The task of monitoring is quite complex and different from operation of normal communication services. In the first place, monitoring activity is not confined to a particular service or a particular type of emission or modulation characteristic. Its scope is of diverse nature since it has to cater for various types of services with different technical parameters and characteristics. Secondly, propagation and circuit conditions are not necessarily optimally suited for reception of emissions and taking measurements from the locations

where monitoring is being carried out. Presence of interfering signals makes the task all the more complicated. In principle, monitoring activity has to start from basically an "unknown" environment to making it "known" by way of identification and observation of technical and allied parameters. It is almost as if one is sitting in a "black box" and tries to establish contents by technical observations. The task is not merely mechanical or of a set pattern but varies depending on the problem and conditions.

3. Categorization of radio monitoring and other allied activities

While there may not be a distinct dividing line between various types of monitoring and other allied activities or between national and international monitoring, various functions can, however, be broadly categorized as follows:

3.1 Surveillance of radio emissions for detection of infringements

Frequencies authorized by an administration are systematically and periodically monitored so as to ensure their adherence to various technical parameters and regulatory provisions and conditions. Activity is somewhat analogous to police patrolling and the infringements observed are reported to users for necessary remedial actions. Infringements generally include excessive frequency offset beyond permissible limits, non-use or incomplete use of callsign, bad quality of transmission (including modulation parameters like percentage modulation, frequency deviation), unauthorized class of emission, unauthorized period, unauthorized frequency, unauthorized traffic, unauthorized link, unauthorized location, excessive bandwidth, unstable frequency, excessive harmonic/spurious emissions, etc. It includes monitoring of maritime and aeronautical mobile bands especially distress frequencies to check adherence to various radio regulatory provisions.

3.2 Spectrum surveillance to determine occupancy, vacancy and circuit loading

For effective frequency management, spectrum surveillance is carried out with a view to determining vacant frequency spots, non-activation of authorized frequencies and density of traffic on a particular channel. A particular band or a particular frequency is, however, required to be monitored over a longer period for coming to a conclusion in this regard.

3.3 Investigation of cases of harmful interference

Specific interference complaints are investigated through extensive monitoring and data analysis for identifying source of interference and for finding suitable solutions to remove harmful interference. Nature of monitoring specifically includes identification, measurement of technical parameters, assessment of extent of interference, etc. Different strategies are to be worked out on a case-by-case basis depending on type of problem, type of service, frequency band, propagational characteristics, etc. The task is often complex because of presence of the wanted as well as unwanted signal simultaneously.

3.4 Rechecking of emissions for ensuring corrective actions

Monitoring is required to be taken up for rechecking various frequencies with a view to ensuring that necessary corrective actions have been carried out by the users for reported infringements, solutions to interference problems, etc.

3.5 Detection of unauthorized transmitters

Monitoring is conducted to detect, identify and locate unauthorized transmitters, so that necessary action may be taken in accordance with the national law.

3.6 Participation in International Monitoring System

Monitoring campaigns are undertaken from time-to-time under the aegis of the International Telecommunication Union for providing data on an international basis for effective spectrum management and allied activities. The liaison with other administrations may be undertaken, as necessary.

3.7 Technical assistance to users

Technical measurements and radio monitoring may be conducted so as to provide technical assistance to users, on request, for keeping their operations within permissible technical limits. These may, in general, include frequency measurement, frequency stability, quality of transmission and measurement of other related parameters.

3.8 Selection of interference-free sites for transmitting and receiving stations

Radio monitoring and technical measurements are carried out with a view to selecting interference-free sites for transmitting and receiving stations.

3.9 Standardization of technical parameters

Technical studies are taken up for providing practical field data with a view to standardizing various technical parameters and framing regulatory provisions for optimum and effective utilization of spectrum.

3.10 Studies on electromagnetic compatibility and propagation

Various EMC studies are conducted for providing data for evolving frequency allocation plans, channelling plans, sharing criteria, protection ratios, antenna and propagational characteristics, mutual compatibility, etc. Scientific studies on propagation may also be carried out.

3.11 Radio noise measurements

Radio noise measurements are closely linked to radio monitoring and measurements are undertaken on industrial, scientific and medical equipment which include various types of electrical appliances, power generation and distribution systems, vehicular traffic etc. for mutual compatibility with radiocommunication services.

3.12 Inspection of radio installations

Periodic inspection of radio installations is yet another measure to ensure compliance of technical, operational and regulatory provisions of spectrum management.

3.13 Equipment type approval

Radiocommunication equipment are required to conform to various technical standards as related to spectrum management. Development, production and usage of equipment is, therefore, to adhere to such stipulations. Laboratory testing of such equipment is, therefore, essential before type approval can be granted.

4. Monitoring observations and measurements

For carrying out various tasks of monitoring, several observations and measurements are taken depending on the nature of assignment. In general, these include:

- station identification;
- frequency measurement;
- field strength measurement;
- bandwidth measurement;
- class of emission;
- quality of transmission;
- modulation characteristics;
- traffic data;
- spectrum analysis of transmission;
- panoramic display/printing;
- spectrum occupancy recording;
- direction finding;
- special technical characteristics for a particular type of service.

5. Typical functional organizational structure

Number of monitoring stations and monitoring channels will depend on quantum of radiocommunication activity, monitoring needs and administrative/operational logistics. Monitoring and other allied activities are dependent on various factors like types and extent of problems, regulatory requirements, frequency bands to be covered, concentration of radiocommunication activity, etc. An example of functional organizational structure of radio monitoring setup is given in Figure 1.

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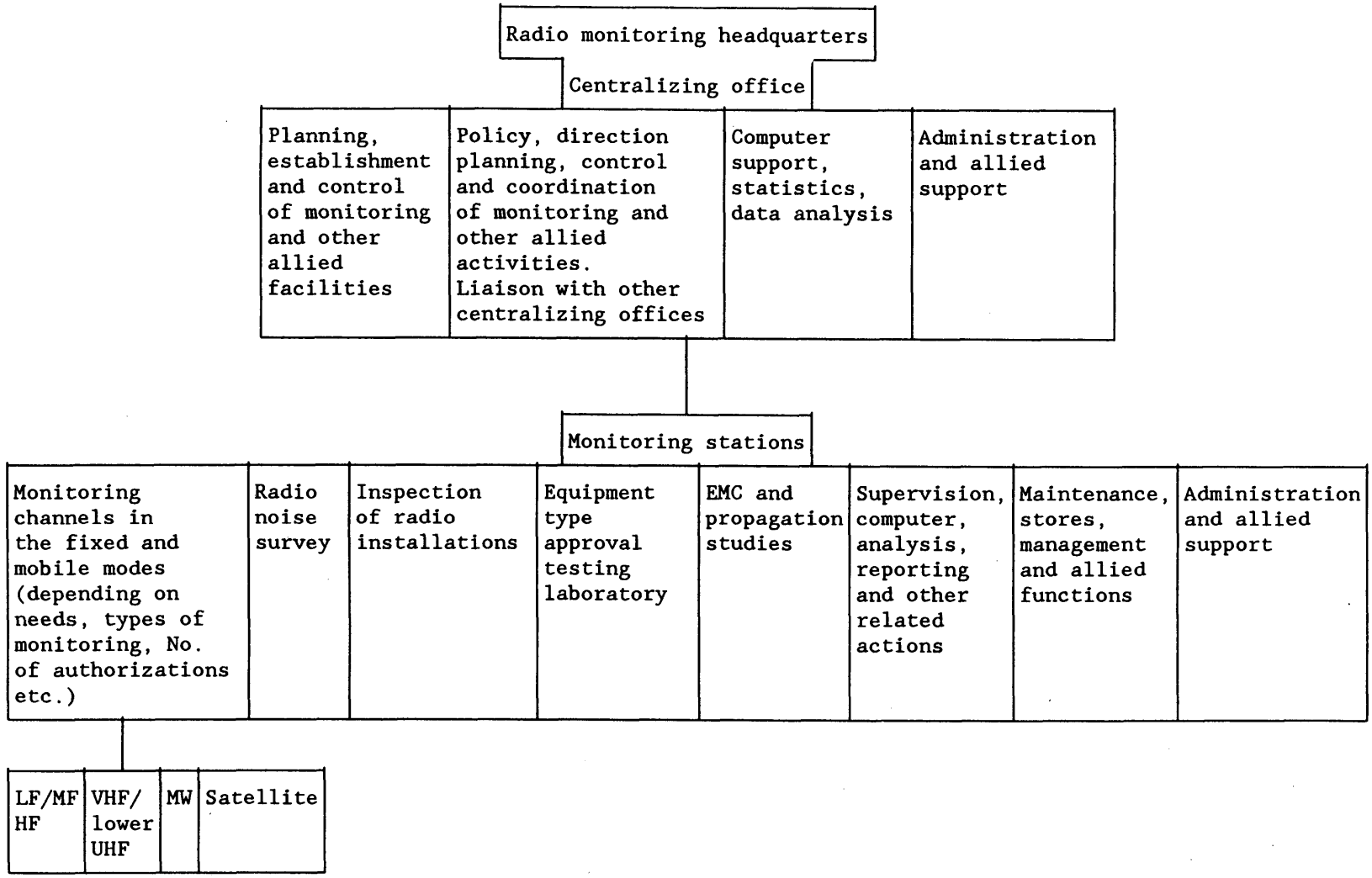


FIGURE 1

An example of functional organizational structure

REPORT 371-2

MONITORING SERVICES IN THE DEVELOPING COUNTRIES

(Question 32/1)

(1966-1974-1990)

1. Introduction

Question 32/1 seeks to discover what kind of monitoring services should be started by developing countries, the guidance that can be given in the establishment and organization of such services and the facilities to be provided by monitoring equipment, including suitable characteristics. This Report as well as several Chapters of the CCIR Handbook for Monitoring Stations give information concerning this subject.

2. Tasks which a monitoring station should be able to perform

2.1 Selection of frequencies for new communication channels

In the 10 kHz - 30 MHz band, it is necessary to:

- select internationally listed frequencies suitable for new communication channels, taking into account, of course, radio wave propagation and the provisions of the Radio Regulations on frequency-band allocation and harmful interference;
- then make a study of the use of the radio-frequency spectrum in the bands selected over a sufficiently long period to prevent interference with other already authorized services.

In bands above 30 MHz, national coordination with the users concerned, and sometimes international coordination with adjacent countries, is necessary. There is no reason why mobile monitoring stations should not then make sure that no unwanted emission disturbs the coordinated frequencies.

2.2 *Monitoring of emissions*

By this is meant the monitoring, either periodically or whenever a check seems required, of the country's own transmitters, and, if necessary, of those of their correspondents. This monitoring consists in:

- measuring the frequencies and bandwidths of the emissions;
- checking the quality of the modulation;
- checking for freedom from spurious emissions and, in some cases, measuring the strength of the radiated field.

2.3 *Measures relating to interference*

In all cases of interference affecting a circuit, measures should be taken promptly to:

- determine the cause of interference,
- identify the interfering emission,
- determine the technical characteristics of the interfering emission,
- propose steps for eliminating the interference.

2.4 *Participation in the international monitoring system*

In addition to the help which it should be able to give to another country or another monitoring station within the framework of § 2.1, 2.2 and 2.3, a monitoring station must also be able to participate in the international monitoring system under the conditions specified in Article 20 of the Radio Regulations.

Such participation consists in carrying out all the monitoring of operations requested by the International Frequency Registration Board (IFRB).

3. Organization and operation of a monitoring station

3.1 General

As regards the organization of a monitoring service, there is no specific plan in existence to meet all needs. The first thing to be done is to determine the nature of the duties to be performed. The factors to be taken into consideration are as follows:

- frequency bands monitored;
- area of the region to be covered.

These factors determine the advisability of installing a direction-finder system and providing mobile monitoring stations to supplement the fixed station or stations required. See also Chapter 17 of the Handbook for Monitoring Stations.

3.2 *The choice of site for the monitoring station*

The site of a monitoring station must be easily accessible by road, easily connected to essential public services such as power and water supplies, as well as telephone communications and should be near to a residential area where the staff can live. At the same time, the site selected must be far enough away from built-up areas to avoid industrial interference. The interference range of a residential area is approximately 400 m and that of an industrial zone approximately 5 km. For HF monitoring, the height of the terrain is of no importance, but since most monitoring stations operate also in the VHF and UHF bands, it is better to choose a site higher than the surrounding ground. As a general rule, it is also better to site the station on flat-land covering an area of approximately 100 000 m^2, with good conductivity and with no obstacles within a radius of approximately 5 km. When planning the establishment of MF and HF direction finders, a larger area should be envisaged.

The theoretical HF coverage area of a monitoring station can be predicted by using computer programmes based on up-to-date information on ionospheric propagation [CCIR, 1970-74].

Assuming that the developing country in question starts off with the very minimum of communication equipment, and probably no monitoring facilities at all, the monitoring installations might quite easily be combined with the receiving station of the communications installations, at least during the initial period of operation. Each transmitting station should be capable of measuring its own frequencies, to be able to operate correctly. Each receiving station might be similarly equipped, so as to be able, if necessary, to render assistance to its counterpart transmitting station, in measuring the interfering signals and to carry out reciprocal monitoring of the frequencies of its correspondents. Besides resulting in a saving of staff (the same engineers and technicians being available for both communications and monitoring operations), performance of these various functions in conjunction would save the expense of establishing the monitoring station on separate ground and in separate premises. Moreover, the various antennas of the receiving station would be of the greatest assistance to monitoring operations.

Obviously, in countries where the receiving stations are operated by private operating agencies other than the national Administration, the question of establishing a monitoring station in a receiving station should be examined jointly by the Administration of the country and the private agency operating the receiving station.

For additional information see also Chapter 2 of the Handbook of Monitoring Stations and Recommendation 575.

3.3 Buildings and power supply

The monitoring station building should comprise at least an equipment room, a power room and an office.

The power should be supplied to the station by underground cables to prevent any interference. It is useful to instal a generating set. An uninterrupted power supply is essential if computer systems are being used as monitoring or management aids.

For additional information, see also Chapter 3 of the Handbook for Monitoring Stations.

3.4 *Antennas*

The question of the antennas for the equipment of monitoring stations is dealt with in Chapter 4 of the Handbook for Monitoring Stations.

3.5 *Staff*

3.5.1 *Establishment*

The following staff is absolutely essential for proper working of the monitoring station:

- a specialist in charge of the station,
- four operators, capable of performing the more common operations and working shifts so that one of them is always on duty,
- a good all-around technician, responsible for new developments, maintenance and fault clearance and also capable of working as an operator.

3.5.2 Recruitment and training

This is the most difficult problem and the one that will take longest to solve. Some information is given in Chapter 18 of the Handbook for Monitoring Stations.

4. Technical equipment and procedure

4.1 *Basic requirements*

Although some degree of monitoring coverage would be possible with even the simplest and most inexpensive installation (e.g., a receiver and a suitable antenna), the following might be considered as a basic monitoring installation which will satisfy the requirements for the more important monitoring functions.

4.1.1 Receivers

The choice of receivers depends in the first place on whether the bands of frequencies to be monitored are below or above 30 MHz. Whatever the frequency band, the receivers should at least possess all the qualities of the receivers used at receiving stations:

- adequate frequency indication accuracy;
- rapid tuning;
- adequate RF selectivity;
- variable IF selectivity;
- very good sensitivity;
- demodulation of all the standard emissions: CW, A3, A3+, A3-, F3;
- adequate protection against:
  - image frequencies;
  - receiver intermodulation products;
  - receiver oscillator radiation.

It is essential to use a panoramic adaptor unit providing a visual display of spectrum occupancy with the receiver in order to locate interfering HF frequencies.

For additional details, see also Chapter 5 of the Handbook for Monitoring Stations.

4.1.2 Frequency measurement

Majority of the existing receivers or field-strength meters can indicate frequency accurately enough, but at a fixed station it is advisable to have a frequency standard which permits measurement of the frequency by the Lissajous and the off-set methods with high accuracy.

Crystal, rubidium and caesium oscillator frequency standards are available with stabilities better than $1x10^{-9}$, $1x10^{-11}$ and $1x10^{-12}$ respectively, but continuity of power supplies must be assured to maintain this stability.

For further details on frequency measurements, see Recommendation 377 and Report 272, and also Chapter 6 of the Handbook for Monitoring Stations.

4.1.3 Bandwidth measurement or spectrum analysis

A method that can be used is to sweep a very narrow filter slowly over the frequency band to be measured. This can be done by connecting a radio-frequency spectrum analyser to the IF output of a receiver or by using an independent spectrum analyser. Also there is a modern digital spectrum analyser (with feasibility of maximum hold) that can make cumulative measurements of the levels received, thus simplifying emission analysis.

Report 275 describes typical spectrum analysers, while Recommendation 443 and Report 275 describe bandwidth measurements. Chapter 9 of the Handbook for Monitoring Stations also deals with this subject.

4.1.4 Field-strength measuring equipment

A field-strength measuring set generally consists of several units which are normally assembled by a manufacturer and form a single set. These units are:

- an antenna with known characteristics;
- a receiver with a step-by-step attenuator which allows to adjust the sensitivity;
- a generator to calibrate the receiver sensitivity;
- a measuring calibrated device.

Nowadays the field-strength measurement receivers include the last three units in one piece.

Modern field-strength meters used with a calibrated antenna give the field-strength value directly. Otherwise all that need be done is to use the field-strength meter with an antenna for which the value of the correction to be made to the value read on the field-strength meter is known.

Because of technological reasons it is necessary to have two types of receivers (field-strength meters) to cover whole the 9 kHz - 1 000 MHz band; one for the band below 30 MHz, another for the band above 30 MHz.

Details of field-strength measurement techniques are given in Recommendation 378, Reports 273 and 368 and Chapter 7 of the Handbook for Monitoring Stations.

4.1.5 Identification equipment

The identification of radio signals is one of the most difficult tasks that a monitoring station has to perform. The difficulty is partly due to the infrequent transmission of identifying signals and partly to the fact that the different types of emission are many and varied.

The facilities referred to above, namely, direction-finding, frequency-measurement, spectrum analysis and field-strength measurement equipment, are valuable aids for identification. Identification of telegraph transmissions is now made easier by the use of automatic telegraph decoders, which can among other things measure the transmission rate. It is always advisable to use a magnetic tape-recorder for difficult emissions by complex transmission systems.

A telegraph signal decoding system is a valuable aid for emission identification. There are decoders which can decode all the main types of telegraph modulation.

See also Recommendation 379, Report 280 and Chapter 13 of the Handbook for Monitoring Stations.

4.2 *Additional equipment*

Although equipment mentioned in § 4.1 will perform many of the monitoring functions, certain additional instruments will permit more effective operation of a monitoring station and will expand its capabilities. These facilities include:

4.2.1 Direction finders for localizing sources of interference to a particular area and as an aid in identifying emission sources based upon knowledge of their location. Generally, only the larger monitoring systems, having a minimum of three direction-finding stations, will benefit greatly from fixed direction finders, since a minimum of three bearings are needed to obtain a reasonably accurate fix. However, even one or two direction finders will provide some useful information. Where the direction-finder networks of two or more Administrations can be coordinated so that the bearing data obtained may be shared, improved fix accuracy should result in many instances. For detailed information concerning the characteristics and relative merits of various types of direction finder, see Report 372.

4.2.2 Automatic monitoring equipment for unattended surveillance of selected portions of the radio-frequency spectrum, with the aid, for instance, of a computer system to obtain, for example:

- frequency or frequency-band occupancy;
- signal level in terms of time;
- bandwidth occupied.

Report 278 gives further details on this question.

4.2.3 Visual monitoring equipment. Although spectrum analysers will provide displays of limited areas of the radio-frequency spectrum, their design is directed primarily towards permitting a detailed analysis of an emission. Visual monitoring equipment and techniques are discussed in greater detail in Report 279.

4.2.4 Monitoring of frequencies could also be done by mobile equipment on motor vehicles equipped as laboratories, comprising:

- a field-strength meter (capable of being used as a receiver) combined with a panoramic adapter. Existing field-strength meters can measure frequencies with sufficient accuracy;
- a spectrum analyser for fine analysis of emissions;
- an array of calibrated antennas suitable for the type of monitoring to be effected.

4.2.5 Apparatus is available for the automatic comparison and control of a frequency source, relative to a VLF standard-frequency transmission. (A list of standard-frequency transmissions is contained in Report 267).

4.2.6 Monitoring of frequencies above 30 MHz generally calls for mobile equipment on motor vehicles. This could comprise:

- a suitable receiver (possibly of the panoramic analysis type),
- a frequency measuring device, capable of acting as a calibrated generator with an accuracy appropriate for the emission to be measured (see Recommendation 377),
- a field-strength measurer, covering up to 1000 MHz,
- a spectrum analyser, with which a single emission or several adjacent emissions with frequency shift may be observed, while a calibration, based on the calibrated generator, offers the possibility of measuring the nominal frequency, actual occupied bandwidth, frequency variation, etc.,
- an array of antennas, suitable for the type of monitoring to be effected.

5. Doc. [CCIR, 1963-66]

The European Broadcasting Union has published a Technical Monograph (No. 3102) entitled: "Receiving and measuring stations for broadcasting purposes" (obtainable from: Centre technique de l'UER, 32 Avenue Albert-Lancaster, Bruxelles 18) which describes the various measuring methods and apparatus used in several European monitoring stations. In this monograph, a number of passages describe the techniques and equipment likely to be of particular interest to developing countries; in particular, the following paragraphs should be mentioned:

Chapter III: Measuring methods

1. The verification and comparison of the frequencies of the standards

2.2 Frequency measurement by the double-beat method

2.3 Frequency measurement by the interpolation method

2.6 Special cases of frequency measurement

3. The determination of the occupation of the frequency spectrum

4.1 Field-strength measurement procedure

6. Measurement of the depth of modulation

Chapter IV: Apparatus required

1. Antennas
2. Receivers
3. Frequency standards

4.1 Constitution of a frequency-measuring position

5.1 Constitution of the automatic recorders of the occupation of the spectrum

6.1 Constitution of the field-strength measuring equipment

6.2 Essential characteristics of field-strength measuring sets

8.1 Device for measuring the average depth of modulation.

REFERENCES

CCIR Documents

[1963-66]: VIII/80 (EBU).

[1970-74]: 1/115 (Canada).

REPORT 277-3

MEASUREMENTS AT MOBILE MONITORING STATIONS

(Questions 22/1, 24/1, 26/1)

(1959-1963-1966-1970-1986)

1. Types of measurement

Subject to the limitations necessarily imposed by mobile operation on the size of suitable instruments and power consumption, it is considered to be practical to employ mobile stations to make all the measurements of emissions normally made at fixed monitoring stations.

Typical measurements and ranges are:

1.1	Frequency measurement	10 kHz to above 10 GHz
1.2	Field-strength measurement	10 kHz to above 10 GHz
1.3	Bandwidth measurement	10 kHz to above 10 GHz
1.4	Direction finding:	
	– with loops	10 kHz to 20 MHz
	– with dipole arrays	20 MHz to above 1 GHz
	– with horns	1 GHz to above 10 GHz
1.5	Automatic monitoring of occupancy	10 kHz to above 300 MHz
1.6	Percentage modulation, FM deviation	10 kHz to 1 GHz
1.7	Video wave-form measurement of television emissions	40 MHz to 960 MHz

Mobile monitoring stations are particularly valuable for monitoring low-power stations and stations at frequencies above 30 MHz whose range is limited.

In addition to the use of equipment installed in a vehicle, certain measurements can also be made using portable equipment. Experience shows that separate mobile monitoring units, specifically designed for selected types of measurement, lend themselves more readily to operational use rather than a single vehicle equipped to make all classes of measurement.

2. Types of equipment

It is considered undesirable to have to use special equipment for the mobile monitoring service and regular stock types of equipment with small dimensions low weight and low power consumption are preferred. Since mobile equipment is subjected to vibration and shock, robust construction, together with shock-mounting of components, is important. With the exception of some portable units, the equipment is normally designed for operation at standard mains voltage and frequency.

It is preferred that instruments, such as frequency standards, which are sensitive to temperature changes, be designed to reach and maintain operating temperature in as short a time as possible, so that continuous application of power will not be necessary or so that only limited circuits will require continuous power. As an alternative, standards which are locked in synchronism with VLF standard-frequency emissions should prove useful.

2.1 *Frequency standards and associated measuring equipment*

The frequency standard should be capable of maintaining a frequency stability of one part in 10^8 per day or better, after nominal warm-up periods and under the operating conditions normally encountered in the mobile service. Currently available frequency standards, specifically designed for mobile use, are capable of stabilities of $\pm$ 1 part in 10^8 or better under field conditions.

2.2 *Field-strength measuring equipment*

2.2.1 *General remarks*

The extent to which field-strength measuring instruments should be provided in a mobile monitoring station depends upon the purposes of the measurements. For regulation enforcement purposes, continuous coverage up to at least 1000 MHz is desirable. For obtaining propagation data, highly sensitive recorders operating on a fixed frequency have certain advantages, although the instruments used for enforcement measurements may also be used for this purpose with some compromise in sensitivity and frequency stability.

This discussion will deal only with tunable continuous coverage instruments.

Wherever possible, field-strength measurements should be made in an area where the signal field is relatively undisturbed by local reflections and re-radiation. Where there is strong dependence on location, measurements may be made in terms of median values, the recording being made continuously as a function of location or time.

The mobile station itself may contribute errors and the calibration includes antenna, receiver and recording equipment. It is desirable, therefore, that the equipment be portable so that measurements can be made away from the vehicle. To avoid directivity effects when measurements are made from the vehicle in motion, omnidirectional antennas are used. Since measurements of harmonic attenuation may involve comparatively weak signals, it is desirable that the field-strength meter be capable of the measurement of signal levels of the order of 1 μV/m. The instrument should likewise be immune to overloading from strong signals of 1 V/m or more. To attain satisfactory accuracies, considering that undisturbed fields are the exception at frequencies above 30 MHz, the errors contributed by the instrument itself should be less than ± 2 dB. It is desirable that circuitry be incorporated for determination of peak and quasi-peak levels of pulsed emissions, as well as for the average field strength of amplitude- and frequency-modulated and similar emissions.

2.2.2 *Field-strength measurements in a moving vehicle*

Report 668, § 8.2, describes a system with an increased capability for measuring field strengths in a moving vehicle.

2.3 *Bandwidth measuring equipment*

Authorized bandwidths, for stations assigned to the various VHF and UHF services, vary from less than 1 kHz for radiotelegraph services to several MHz for television broadcast services. Mobile monitoring stations should, therefore, be equipped with spectrum analyzers having very flexible characteristics, including a sweep width of 20 kHz or less for narrow-band emissions and with the capability of displaying at least 5 MHz of spectrum at a time on certain broadband emissions.

The sweep rate should be variable from about 1 to 30 sweeps per second to permit optimum utility for various types of observation. Provision should be made for either linear or logarithmic display of signal levels and calibrated scales should be provided for direct measurement of level ratios.

Mobile measurements of bandwidth may be made by a frequency occupancy recorder, for those types of modulation for which no measurement method of higher accuracy exists, or when it seems impracticable to carry out such measurements in the mobile service because of the high technical expense involved. With telegraph signals for instance, the bandwidth can be determined from the signal element with the shortest rise time (see [CCIR, 1959]). In the VHF band, estimated bandwidths can also be read off the oscillagraph screen of a panoramic set with a calibrated frequency scale. The measurement of bandwidth can also be made by a method of frequency analysis, using a very narrow (± 100 Hz) filter to measure the magnitude of the spectrum components passing through the filter in relation to the carrier.

2.4 *Direction-finding equipment*

It is desirable that the monitoring receivers in a mobile installation also serve as receivers for the mobile direction-finder. Light weight, compactness, high sensitivity, effective-shielding, stability, robust construction and low power-consumption are desirable features for mobile monitoring receivers. Furthermore, receivers for mobile monitoring should provide continuous tuning coverage of as wide a frequency range as possible. Accurate calibration of the order of 1 kHz at 30 MHz and below, and 10 kHz above 30 MHz is highly desirable. Accuracy of frequency resetting of the order of 1 part in 10^4 or better is also desirable. Although band-switching is preferable, plug-in radio-frequency units for different bands may reduce the circuit complexity.

The loop rotater should preferably be installed in the approximate centre of the roof of the vehicle, with the loop socket extending through the roof and with the azimuth scale and rotating mechanism (usually a hand wheel) inside the vehicle at a convenient location. A number of plug-in loops with different characteristics will permit optimum results to be obtained over a wide range of frequencies.

Directional antennas are needed to extend the frequency range above the limits of loop antennas. Desirable characteristics are: high directivity. broadband frequency coverage, high gain, horizontal and vertical polarization and small size. Antenna types in use are horizontal and vertical dipoles, Yagi-beam, helical and rotatable H-type Adcock, while parabolic, horn type or corner-reflector antennas can be used above 300 MHz.

An antenna finding increasing acceptance above 1 GHz is a broadband horn covering approximately a frequency range of 2 to 1. For added gain and directivity, the horn may be mounted in a parabolic reflector.

For antennas in the VHF/UHF range, it may be very useful to use telescopic rotating supports controlled from inside the vehicle employed as a mobile station. With these supports, it is possible to surmount obstacles near the place of measurement and to take measurements at various heights with the antenna steered in different directions. The antenna height and azimuth should be indicated accurately on the dials inside the vehicle.

2.4.1 *Portable direction-finding equipment*

A portable direction-finding equipment (see Report 372) can be used for the following tasks:

- direction finding of VHF/UHF signal sources by means of a manual DF antenna;
- detection of undesired parasitic radiation (RFI) caused by devices of all kinds (e.g. radio and television receivers, electric motors and combustion engines, electronic data processing (EDP) systems, process computers or high-voltage insulators);
- protection against wire tapping by detecting non-authorized transmitters.

2.5 *Automatic monitoring of occupancy measuring equipment*

Monitoring of spectrum occupancy, at specified locations or along specified paths, may be made in the mobile monitoring unit, by means of a frequency spectrum recorder or by means of a panoramic adaptor.

The latter permits visual observations to be made over a relatively small band of frequencies while the vehicle is in motion.

2.6 *Modulation measurement equipment*

It is desirable to provide facilities for meaurement of percentage modulation of both amplitude- and frequency-modulated emissions. For the former, a cathode-ray oscilloscope with a wideband vertical amplifier usable up to 10 MHz or more is required. For measuring modulation deviation of FM emissions, an instrument capable of accurate measurement of carrier deviation, over a range from ± 5 kHz or less to ± 100 kHz or more, is required and should be capable of indicating instantaneous values. An overall accuracy of 5% or better is highly desirable. A cathode-ray oscilloscope may be used as an indicator for the carrier deviation meter to permit measurements of instantaneous peaks of deviation of FM stations.

2.7 *Television measuring equipment*

Measurements of the characteristics of the video waveform of television transmissions can be made with instruments specially designed for the purpose. Cathode-ray oscilloscope display of the various portions of the video signal is desirable, so that individual picture lines or segments thereof, as well as individual synchronizing pulses, may be observed. Provision should also be made for viewing the entire picture on a screen of adequate size to permit evaluation of the picture quality.

2.8 *Microwave monitoring and measuring equipment*

Signal reception and measurements at frequencies above about 1 GHz are often difficult to accomplish due, in part, to the general reduction in the effective sensitivity of the receiver as compared with the value at lower frequencies. Although the use of very high-gain antennas would greatly alleviate this problem, antenna size and weight become critical factors in mobile installations. Another approach to improved system sensitivity involves reducing the noise contribution of the receiving equipment. Since the signal-frequency stages, up to and including the first mixer, are generally the chief sources of noise in a microwave receiver, careful selection of circuitry and components for low-noise contribution in this portion of the receiver is of prime importance. Significant improvement (of the order of 10 to 15 dB) in the sensitivity of microwave receivers has been made during the past few years.

Some of these receivers are provided with impulse-type level calibrators and with calibrated attenuators so that they may be used in making field strength measurements.

2.8.1 *Antennas for microwave receivers*

Various types of calibrated antennas are available for use with the receivers; one type which has been found to be particularly useful for general monitoring, a well as for field-strength measurement work in the 1 to 10.5 GHz range, consists of a log-periodic antenna permanently installed in a 60 cm (24 in) parabolic reflector. The antenna assembly is installed on a specially designed mount which permits orientation over 360° of azimuth and from −30° to +90° in elevation (0° = horizontal). Automatic or manual positioning of the antenna is accomplished by means of controls installed at the operating position in the monitoring vehicle. Special horn antennas, with suitable parabolic reflectors, are also available to extend the frequency range to 26 GHz. The typical gain of a 1 to 10.5 GHz antenna, of the type mentioned above, is shown in Fig. 1.

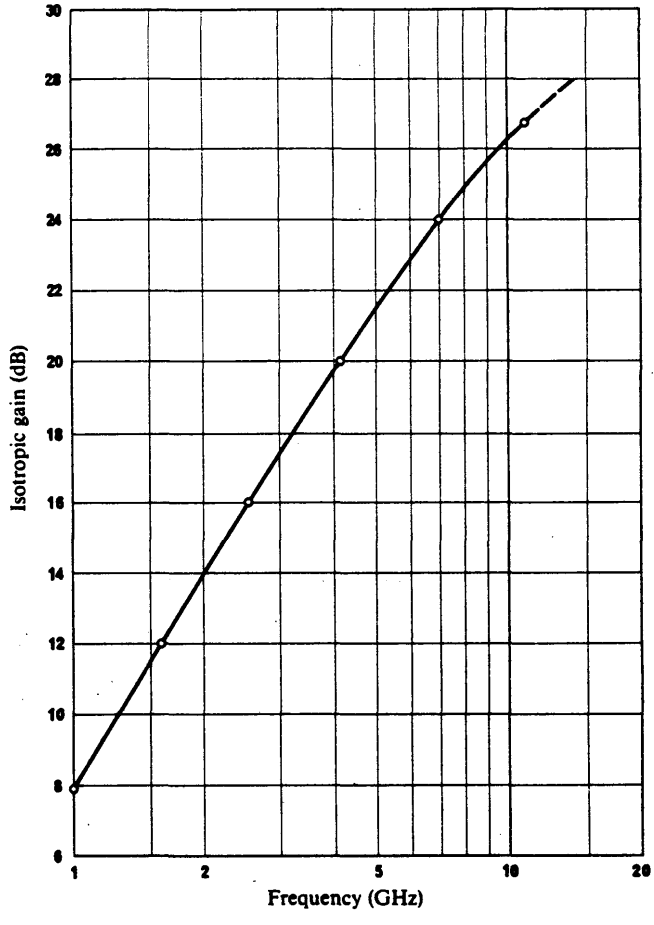


FIGURE 1 — *Typical gain of a calibrated directional log-periodic antenna with a 60 cm (24 inches) diameter reflector, as a function of the frequency*

2.8.2 *Spectrum analysers*

Although there have been no revolutionary changes in the techniques for spectrum analysis or equipment, continuing improvements have been made in recent years, including the use of phase-lock circuitry to provide a trace of greater stability. Some of the newer analysers are provided with variable-frequency input controls, so that a single instrument may be used with a variety of receivers including the microwave receivers mentioned above. One such instrument may be tuned over a range of centre frequencies from less than 10 kHz to 110 MHz, permitting it to be operated from an intermediate frequency output of the microwave receiver. For the observation of the baseband or demodulated spectrum of radio-frequency signals, output may be obtained from the video-frequency or the discriminator outputs of the receiver.

In such analysers, several widths of sweep are available up to 25 MHz or more. Generally, the receiver selectivity characteristics, rather than limitations of the spectrum analyser, will determine the sweep width used. The resolution filters, through which the sweeping-mixer ouputs are passed, may be set to various bandwidths, from less than 1 kHz to 100 kHz or more. The continuous sweep may be disabled, to permit manual adjustment of the frequency presented to the resolution filter so that any desired portion of an emission may be examined in detail.

2.8.3 *Frequency measuring equipment*

In the measurement of frequency of microwave emissions, in the range from 1 to 10.5 GHz, the use of a stable transfer-oscillator with an electronic frequency counter has been found best, consistent with simplicity of operation and required frequency tolerance. Although there may be a slight residual frequency-modulation of the oscillator, the short term stability will permit measurement of frequency to an accuracy of about 1 part in 10^6 in the range from 1 to 10.5 GHz.

2.9 *Power supplies*

To provide adequate power for large mobile monitoring units, a separate power unit, either integral with the vehicle or mounted on a trailer, may be used. An engine-alternator unit of 5 kW rating is usually adequate for this purpose. For the smaller mobile monitoring units, an alternator of about 500 W driven by the vehicle fan belt is a useful source of power.

3. Accuracies and limitations of measurement

Most of the measurements listed in § 1 of this Report can be as readily performed with mobile as with fixed facilities, generally with little or no compromise in desirable accuracy.

3.1 *Frequency measurements*

Mobile frequency standards are available which will maintain stability better than ± 1 part in 10^8 per day under field conditions.

By means of frequency synthesizers or electronic counters, it is often possible to obtain direct measurements of the frequency of emissions, without recourse to auxiliary interpolating instruments, over a range of frequencies extending above 100 MHz. Transfer oscillator techniques may be used at much higher frequencies, with interpolation between the transfer oscillator and the frequency standard being made with an electronic counter or other appropriate means. Measurement accuracies, within the desirable limit of ± 1 part in 10^7, may be accomplished under optimum conditions when the nature of the carrier is such that a precise zero beat may be obtained or where direct counting techniques may be used.

On FM emissions, frequency averaging techniques with an electronic counter will provide an accuracy of 1 part in 10^6 or better.

3.2 *Field-strength measurements*

The accuracy of field-strength measurements above 30 MHz is limited, not so much by the instruments themselves, as by the local conditions and transmission path which tend to distort the electromagnetic field, so that a considerable variation in field strength over relatively short distances may be expected. This problem may be alleviated to some extent by making several measurements at each measuring location, moving the antenna a few feet horizontally between measurements. Very good results can also be obtained by taking measurements at different heights using the telescopic supports described in § 2.4 [Gentile, 1966].

Under average conditions, even when appropriate precautions are taken to ensure optimum accuracy, the accuracy cannot be depended upon to be better than ± 2 dB. If large numbers of measurements are made as functions of time, location or distance, etc., the statistical evaluation will yield values of the quasi-maximum, quasi-minimum and median values with an accuracy of about ± 3 dB or better, depending somewhat on the distribution of field strengths. A major limitation of portable or mobile field-strength meters is the limited sensitivity, particularly at the higher frequencies. In the lower portion of the spectrum, up to several MHz, atmospheric noise is likely to be a limiting factor on the usable sensitivity, while at higher frequencies the noise introduced by the instrument itself becomes important. Under average field conditions and depending upon the frequency involved, the minimum field strength, which may be measured with portable instruments with the accuracies mentioned above, will range from about 2 to 100 μV/m in the frequency range from 10 kHz to 1000 MHz.

3.3 *Bandwidth measurements*

Generally, the same limitations apply on measurements of bandwidth by mobile installations as for measurements at fixed monitoring stations. The same measuring instruments are applicable to both types of installation. One advantage of mobile, as contrasted with fixed installations, is that, in the case of the former, measurements can often be made immediately adjacent to the transmitter, eliminating some of the measurement problems (such as those associated with interference and noise), which are mentioned in Report 275.

3.4 *Direction finding*

A major limitation of loop-type direction finders is their polarization error in the presence of signals with significant sky wave components. A further serious limitation of a direction finder, which depends upon the direction of arrival of the wave front, is the likelihood of local reflections or re-radiation from nearby objects such as buildings, transmission and telephone lines, etc. In an area where the wave is relatively free from distortion due to local conditions, where the path of arrival is essentially horizontal (vertically polarized wave), and taking into account possible errors introduced in orienting the direction-finder azimuth scale, accuracies of the order of ± 5° with vehicle-mounted loop-type direction finders are typical, with somewhat greater accuracy at frequencies below about 5 MHz. However, where the wave has been distorted by local conditions or where sky-wave components are present, bearing nulls are likely to be obscured or to deviate widely from the true direction of the emission source.

Where the ground wave is predominant (i.e. the ratio of ground wave to reflected waves being at least 6 dB), direction finding loops may be used in the frequency range 10 kHz to 30 MHz. The accuracy obtained depends upon the signal level and the equipment used.

Typical values for a direction finding null 1° wide are:

10 kHz : 50 μV/m,
1000 kHz : 15 μV/m,
20 MHz: 10 μV/m.

For the bearings of horizontally polarized signals (line-of-sight) at frequencies above 30 MHz, Yagi-beam antennas may be used in a method of maximum signal direction finding with an angular width of bearing indication of ± 15°.

For measurements made from a vehicle in motion, no proper bearing seems possible because of multiple reflections from nearby objects. However, when equipped with a visual indicator, mobile direction-finding sets will allow determination of a median value of the varying indications and hence yield an approximate direction.

3.5 *Monitoring of occupancy of the radio-frequency spectrum*

The accuracy obtainable from a frequency spectrum occupancy recorder depends upon the receiver and the overall calibration. Equipment used in the Federal Republic of Germany has the following characteristics:

- Sensitivity: better than 1 μV/m.
- Accuracy of field-strength indications: ± 3 dB at angles of incidence below 60°.
- Frequency stability of receiver: ± 5 parts in 10^5 ± 40 Hz (where ± 40 Hz is the bandwidth of the filter).
- Accuracy of reading time-markers: ± 1 min or ± 30 s.
- Adjustable maximum frequency-spread on the recording strip by choice of appropriate gear reduction: 100 Hz per mm. When monitoring above 30 MHz by means of panoramic sets the resolving power may be about 400 Hz.

3.6 *Percentage modulation*

Cathode-ray oscilloscopes, with the desirable characteristics for measurement of percentage modulation of amplitude-modulated signals, are available. Accuracy of such measurements can usually be maintained within ± 5%. For measurement of carrier deviation, instruments are available which meet the desirable characteristics listed in § 2.6. Self-contained calibrators are provided to maintain an instrument accuracy better than 5%.

3.7 *Television measurements*

Instruments are available for performance of all desirable television measurements and observations mentioned in § 2.7. Measurement accuracy of the various instruments in most cases approaches or equals that of measurements which can be performed at a fixed location. Comparative level measurements of components of the composite video waveform, for example, can usually be made with an accuracy of 5 to 10%.

Measurements and observations, which may be performed with the equipment installed in a mobile television enforcement unit in the United States of America, include:

- frequency measurements of television station carriers (including chrominance sub-carrier frequency of colour emissions);
- the precision frequency measuring equipment covers a frequency range of 20 to 1000 MHz with an accuracy under optimum conditions of 5 parts in 10^7;
- measurements of synchronizing pulse waveform and timing, video wave form and of colour video phase relationships;
- checking of video levels with respect to synchronizing pulse levels;
- measurements of bandwidth of television emissions including observations of spurious emissions at all frequencies between 54 MHz and 890 MHz;
- percentage of modulation rate of vision and sound emissions;
- measurements of distortion and noise level at audio frequencies.

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REPORT 1116

AIRBORNE MONITORING STATIONS

(Questions 28/1, 31/1, 32/1)

(1990)

1. Introduction

Present radio frequency spectrum monitoring systems are generally based on fixed and suitably equipped ground mobile stations. The integration of an airborne station in the monitoring system could allow some advantages as:

- quick investigations over broad areas;
- acquisition of larger amount of data of optically propagated signals;
- more realistic monitoring inside aeronautical protection volumes;
- complementary performances (e.g. analysis of interference to satellite transponders).

This report outlines the methodology and relevant arising operational problems in the use of an aircraft monitoring station, taking into account the current experience on aeronautical flight-test procedures and systems.

2. General considerations

Due to the necessary particular configuration of an airborne station and the environment within which it is operated (e.g. 3D movement), the monitoring activities that could generally be performed are based upon the following measurements:

- frequency measurements (see Report 272);
- field-strength measurements (see Report 273);
- bandwidth measurements (see Report 275);
- direction finding (see Report 372);
- identification and location procedures.

A form of "spectrum occupancy" in space domain could be also performed; in fact, while the observations are not continuous, the location and identification data of each station in connection with the characteristics of the emission could be recorded and monitored again during periodic flight inspections, so as to plot (tridimensional) maps of the areas under observation. That type of presentation could be used,in computer aided techniques, for channel assignment analysis and for solution of particular compatibility problems.

Particular applications could be exploited using an helicopter (e.g. detection of antennas radiation pattern), nevertheless that type of platform may not be the better solution for radio monitoring applications due to operational (weather conditions), technical (interference from rotating wings, mechanical vibrations), and economical (high cost to load ratio) considerations.

3. The platform

3.1 Aircraft operational requirements

Taking into consideration the on-field spectrum monitoring activities and measurements to be performed by a mobile station (see also Report 277-3/1), the operational requirements of an aircraft housing an airborne monitoring system should be as follows:

- flight capabilities for all weather operations;
- manoeuvrability and good stability at low speed/altitude;
- adequate load capability in order to accommodate all necessary equipments and monitoring staff.

3.2 Aircraft technical requirements

Technical characteristics required for an aircraft supporting an airborne monitoring system should perform the following criteria:

- propulsion

  reliable multiengine system; safe operation should be maintained in mono engine conditions also at full load;

- autopilot

  facilities should permit surveillance operations by prefixed procedures (e.g. grid or circular patterns);

- avionics

  navigational equipment on board should allow full instrument flight operations; for certification reasons the reference navigation system interconnected to the monitoring processing unit should be independent;

- range

  it should be possible to operate for at least two hours in the area under observation in full load condition and below 6,000 ft;

- speed

  speed range should be large enough to allow both quick transfers and homing procedures at low altitude;

- operational altitude

  desired radio horizon determines the corresponding operational altitude (e.g. 20,000 ft for about 250 km); higher altitude could permit quick transferring;

- noise and vibrations

  should be as limited as possible to avoid degradation of reliability and performance of the equipment on board;

- power supply

  stable power source, to guarantee correct operation of equipment; auxiliary power supply batteries should be provided.

4. The station

4.1 General requirements for an on-board system

All airborne system set-up must be designed and implemented very carefully in order to ensure maximum safety and reliability of the aircraft.

Some of the parameters which can directly influence flight operation are:

- weight and location of apparatus on board, so as to avoid any upset of aircraft stability;

- structural stability of the aircraft following eventual modifications;

- aircraft aerodynamic following monitoring antennas installation;

- thermic and electrical power balance.

The above listed parameters show the importance of an accurate cross-checking to be performed in order to improve the system reliability and to limit operating cost.

Also, simplified maintenance (modular design and built-in-test system) and possibility to remove and substitute fault equipments in a easy way for laboratory check and calibrations, are desired qualities in the aeronautical field; that philosophy of maintenance requires adequate availability of spare parts and equipment sub-systems.

4.2 On-board system set-up criteria

The main station sub-systems should be based on the following criteria:

- Self-relying reference navigation system

  the reference navigation system connected to the monitoring station must be able to determine its own position in order to perform signal source fixing without any external auxiliary trajectografic system.

  Moreover, by coupling the reference navigation system (e.g. inertial and/or satellite system, cross DME readings) to the DF processing unit, it will be possible to compute the position of radio signal sources for all possible heading and trim assumed by the aircraft.

- Signal source multiple inspection

  the monitoring system should be capable of scanning a predetermined list of frequencies, processing and storing the characteristics of each emission, and comparing new signals at every cycle with previously fixed radio sources.

- Efficient man-machine interface

  aircraft radiomonitoring operations generally involve limited task time and minimum personnel on board. Therefore monitoring equipment must be capable of performing programmable and automatic measuring procedures, reserving manual operations only for very specific tasks.

  In any case data presentation should be of an immediate interpretation for technical staff in flight.

4.3 On-board system configuration

In general, the configuration of the on-board monitoring system should include the following sub-system:

- antennas;
- reference navigation system;
- receivers and DF system;
- computing unit;
- audio and data record system.

A communication system could be implemented for data transfer and/or coordination activities with other fixed or mobile monitoring stations.

4.3.1 Antennas

The monitoring antenna system should be preferably installed below the main body of the aircraft, so to be free when in flight.

Usually, from the physical point of view:

- clearance from ground during landing and take-off have a direct influence over dimensioning and positioning of antennas;
- design and positioning influence the general performances of the aircraft.

Experience acquired by at least one administration (see Annex 1) points out that problems connected to antenna dimensions and positioning have, especially in case of small/business class aircraft, a direct impact also in the choice of DF concept system (i.e. Doppler, adcock/interferometry) (see also Report 372).

Further investigation and studies are needed regarding design of small dimensioned DF aircraft antenna systems.

4.3.2 Reference navigation system

A reference navigation system, connected to the computing unit, should provide the following navigation parameters:

- latitude and longitude;
- heading informations;
- ground speed;
- height/flight level;
- trim (pitch/roll).

Those parameters can be provided by:

INS/GPS integrated unit;

and/or VOR/DME radionavigation receivers;

and/or LORAN-C unit;

magnetic/gyro compass;

barometric/radio altimeter.

While the VOR/DME cross-reading needs to be processed, the LORAN-C gives the position of the aircraft directly in geographical coordinates. An inertial navigation system could instead give, as well as geographical coordinates, information about acceleration and trim; those informations could be very useful when processing the bearings accuracy (see Report 372, § 5).

Pointed out that the accuracy of fixing depends upon the reference navigation system adopted, an analysis of the requirement should be made taking also into account areas usually under observation (e.g. urban or rural areas).

Usually the information derived from aeronautical equipment are available both in the form of analogic and digital signals.

4.3.3 Monitoring and recording equipment

Depending on the load and volume capabilities of the aircraft, the measurement equipment used for ground mobile station applications could also be adopted for aircraft monitoring stations; some modifications could be required due to the power source (112 Vac/400 Hz or 12 Vdc) and the electromagnetic compatibility requirements as established from the aviation authorities.

5. Operations

5.1 Activities

The main type of activities which can be carried out making use of an airborne monitoring station, could be organized as follows:

- periodic scheduled activities
  (e.g. spectrum utilization surveillance);

- non periodic scheduled activities
  (e.g. new implementations);

- non scheduled activities (e.g. interfering signals detection).

5.2 General purpose procedures

Flight missions should be duly scheduled and well planned.
This implies that some coordinating activities should take place following the general guideline listed below:

a) Request for flight mission indicating time duration.

   Monthly or weekly schedules can be fixed only when the activities to be performed and the relative task time are known, together with aircraft and crews availability.

b) Flight activities coordination

   Aeronautical authorities responsible of air spaces where radio monitoring should take place, together with the ones in charge of aircraft and crew logistic support, must be notified in advance.

   Flight procedures required for monitoring may in fact require some special coordination, especially in heavy air traffic areas.

   Likewise, ground fixed and mobile monitoring stations, which should take part in joint coordinated activities, should be advised.

c) Pre-flight preparatories activities

In addition to operational planning and preparation of flight mission (flight plan, aircraft checkout, fueling etc.), data obtained from measurements made in previous missions, or available otherwise, should be carefully examined in order to preselect the functions and the automatic procedures of the monitoring equipment to be used during the flight.

d) Mission execution

All necessary flight operations must be duly performed, depending on the required monitoring activities.

5.3 Operational procedures

Monitoring flight mission implies different operational procedures depending upon the requested activities.

Such procedures must be carefully planned before each mission. For instance, it is likely that during monitoring mission the aircraft will have to fly on predetermined paths, such as loops, at suitable speed and altitude.

6. Conclusions

An airborne radio monitoring system offers large possibilities for efficient and quick inspections of the radio spectrum, also in the areas where fixed or ground mobile stations present some limits.

Free sight and possibility to perform direction finding measurements from far places in a limited time indicates that an aircraft station can efficiently be integrated with ground fixed or mobile station, especially in those countries where mountainous lands represent a serious limitation to the development of networks of remote-controlled stations.

Nevertheless, in-flight trials conducted from one administration (see Annex 1) indicate that further studies and investigation are needed, with particular reference to:

- effects deriving from the platform speed and trim on frequency measurements and DF bearings (Doppler system);
- design and performance of smaller DF aeronautical antenna system;
- computer aided procedures for automatic monitoring;
- interoperability with fixed and/or mobile stations using the aircraft station as remote-controlled station.

Annex I

AIRBORNE MONITORING STATION

The Italian experience

1. Introduction

The Italian Administration carried out, during 1987, an experimental campaign concerning the application in the spectrum monitoring service of an aircraft fitted out with suitable equipment.

The purposes of the experiment were:

- to acquire knowledge of technical and managerial aspects connected to the use of such particular craft;
- to study problems that could arise in the on-board use of equipments usually installed on ground vehicles;
- to outline a good configuration of a system to be installed on board;
- to verify the practical usefulness of an airborne monitoring station in spectrum surveillance activity and location of sources of interference.

A total amount of 40 hours of trial flights, as well as ground tests and initial tests of functionality, was done both in real and in simulated situations.

The following sections describe briefly the system tested, the results obtained and some general considerations.

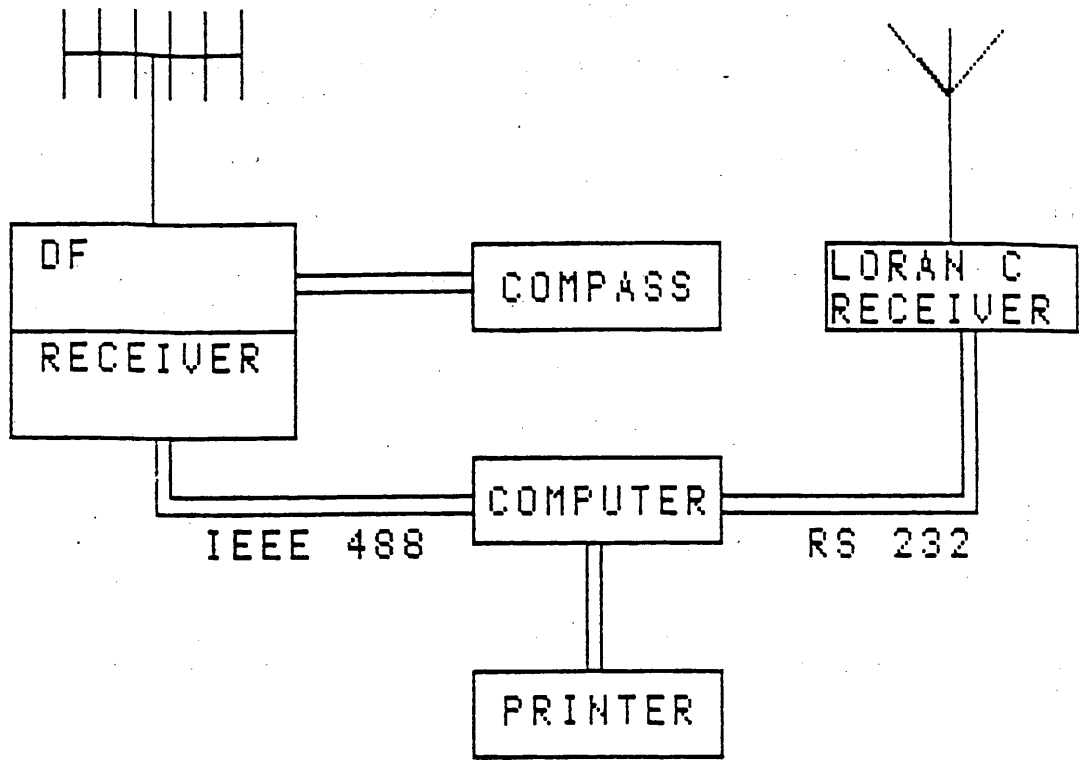
2. Equipment used

The monitoring system was composed of (Figure 1):

- VHF/UHF direction finder;
- VHF/UHF monitoring receiver;
- LORAN-C receiver;
- gyroscopic compass;
- computer.

The computer was connected with the DF by means of an IEEE-488 interface and with the LORAN-C receiver by means of an RS-232 interface. The compass was directly connected with the DF that is provided with a special input for this purpose. The power was supplied by a small inverter, operative only after ignition.

No serious problems were detected with regard to EMC compatibility or environmental conditions (e.g. vibrations).



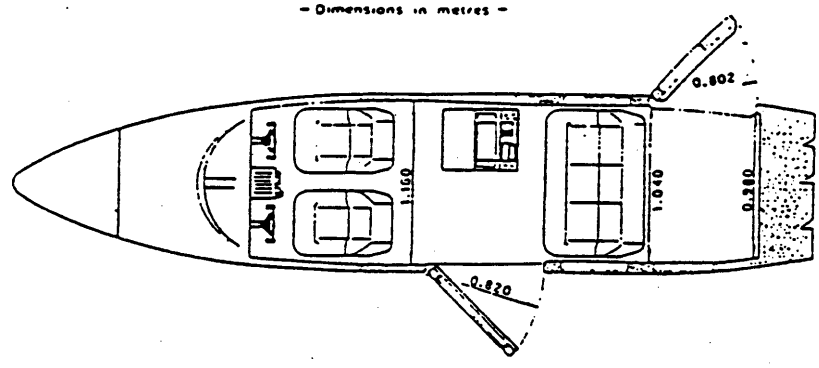


FIGURE 1
System block diagram and onboard arrangement

2.1 Direction finder and monitoring receiver

The direction finder adopted was an equipment based on Doppler principle, covering the frequency range from 20 to 1 000 MHz.

That type of equipment allows both digital indication of the bearing and the analogous pattern indication on a cathode ray tube which enables an easy evaluation of the DF quality impaired by wavefield distortion.

A receiver, especially designed for radiomonitoring, was connected to the reference antenna of the DF, without any interference from DF procedures.

During the experiment two DF antenna systems were used, covering respectively the frequency sub-bands 20 to 500 MHz and 500 to 1 000 MHz. Each antenna system is made up of 16 dipoles arranged on a circle and a central reference antenna.

The radome had to be especially designed and installed as shown in Figure 2; each antenna had to be mounted one at a time according to the frequency range work programme.

The antennas' installation caused the major problems due to their significant dimensions and the necessity to be kept away from the body of the aircraft.

2.2 LORAN-C receiver

In order to determine the aircraft position, a LORAN-C receiver provided with RS-232 interface, was used.

It caused many system deficiencies, since the position data were often affected by unacceptable errors and in certain areas of the country it did not work at all.

Its antenna consists of a simple rod 1 metre long and it was installed under the DF antenna; however, in order to obtain better results, such a proximity and any shielding effect due to the aircraft should have been avoided.

2.3 Gyroscopic compass

A gyroscopic compass, provided with an electric output, was directly connected to the DF, so to obtain the bearing indication referred to the north, instead of the longitudinal axis of the aircraft.

On the digital DF display, indeed, were obtained informations either referred to the antenna alignment or to the north.

The informations on the analogue CRT display were always referred to the antenna alignment in order to indicate to the pilot the best route to follow during the localization of a single emission source (homing).

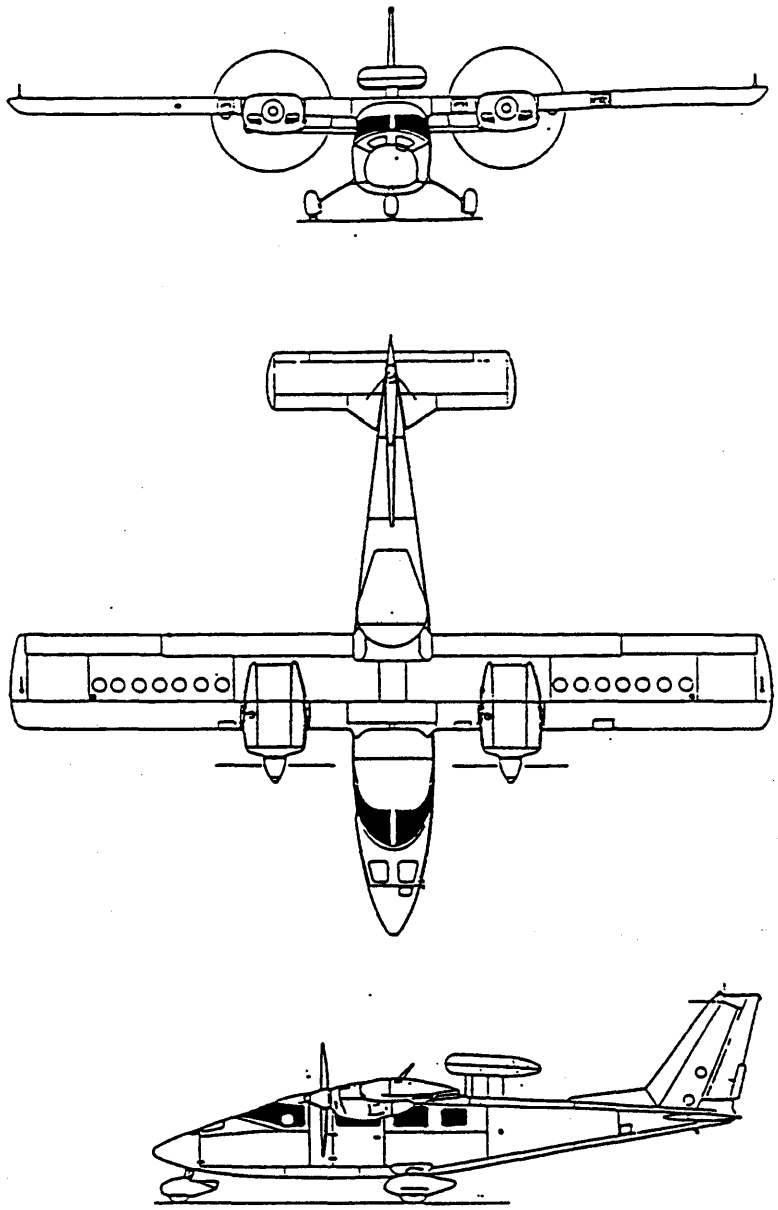


FIGURE 2

Direction finder antenna installation

2.4 Computer

A small computer with 16 bit microprocessor, two floppy-disk drivers, 256 Kbytes RAM, MS-DOS operating system, was used.

The software developed was very simple, consisting only in data acquisition and reproduction.

A basic program consisting in a cyclic routine, took data from the DF and the LORAN-C receiver, showing on the screen and simultaneously recording on a floppy-disk, the following elements:

- date and hour;
- latitude and longitude;
- frequency;
- bearing.

It must be noted that there was neither control of the instrumentation nor data elaboration by the computer, but only a situation recording each second.

The operator was free to manually set the equipment in any condition he wished.

3. Measures

The main causes which affected the measures of bearing can be listed as below:

a) instrumentation errors
   - electric errors of DF;
   - difference between aircraft axis and antenna axis;
   - compass errors;
   - LORAN-C receiver errors;

b) site errors
   - wavefield distortion due to the metallic aircraft body;
   - shielding effect of the metallic aircraft body;
   - mutual interference between DF antenna and other antennas installed.

Some of such errors are frequency dependent, and in particular those linked to the field distortion are inversely proportional to the DF antenna "aperture" related to the wavelength.

While the bearing accuracy mean was in the order of ± 4 degrees in the frequency range from 20 to 300 MHz and ± 7 degrees in the frequency range from 500 to 800 MHz, it has also to be mentioned that in certain conditions there were many measures with errors contained within ± 1 degree; it always happened when the DF antenna had fully free sight of the source (e.g. free from the wings shielding), in the upper part of the frequency sub-range of each antenna.

The operational conditions during the measurements were:

altitude	between 2,500 ft and 5,000 ft
ground speed	about 115 kts.

The bearings considered were those acquired in fully balanced aircraft conditions.

4. Considerations

From the global consideration of the problems encountered during the experimental campaign and of the results of the measures carried out, the following indications can be given with regard to the craft, the equipment and the suitability to the monitoring service.

4.1 Main equipment

Results showed a fairly good DF precision, a few degrees in certain conditions, which can be considered sufficient for the localization of sources of emissions, taking into account that on the same source as many measures as needed can be taken, in order to increase precision. Nevertheless it has to be expected that in many cases the need of a more precise localization requires ground vehicle intervention.

The monitoring receiver used is provided with a small panoramic display, very useful for monitoring activities. Moreover, a larger screen representing all the radio frequency spectrum instead of the IF spectrum only, would be preferable.

The antennas dimension, the need to take them away at least 30-40 cm. from the body of the aircraft, the problems connected to the replacement depending on the frequency spectrum under observation, focused on antenna system the main limitations. In the experiments, indeed, the radome containing the antennas had to be arranged on the top of the tailcone, imposing limitations to aircraft performance and influencing greatly the measurements.

A more satisfactory solution could be found using the same DF equipment with an appropriate antenna system designed for avionic applications, or using a DF equipment operating on a different principle requiring an antenna system more suitable for such application. Installations in the lower side of the aircraft are preferable.

4.2 Additional equipment

Trials demonstrate that additional equipment (LORAN-C receiver, compass, computer) are absolutely necessary in order to fully benefit the craft capabilities. Among other things, it was pointed out that the capabilities of an airborne monitoring station are strongly linked to the use of a computer-controller with software developed to carry out procedures-automation, data

acquisition and effective man-machine interface. Bearing in mind the particular environment where the operator has to work, it is indeed clear that both rapid actions and intensive use of predefined procedures are required.

The possibility of installing a more sophisticated LORAN-C receiver or a different radiodetermination system should be studied, since the position information has a great impact on the obtainable results.

4.3 Aircraft

The aircraft used in the experiment proved to be very easy to manage, but also revealed limitations that are not negligible. From its small dimensions, problems arise from antennas installation, for equipment sizes and, last but not least, for environmental conditions of personnel who operate on board. Furthermore, its operational safe capabilities in bad weather are limited, therefore interventions in every circumstance are not guaranteed. It has also to be underlined that, while operational speed and manoeuvrability appeared to be suitable for spectrum monitoring activity, maximum speed disposable to reach the operational area is too low. This does not affect interventions quickness, but mainly affects costs deriving from the total amount of flight time required for each intervention.

The convenience to employ a bigger machine should be examined.

4.4 Usefulness in the service

The experience made could be considered positive from this point of view. The possibility to carry out a spectrum surveillance service and source of emission localization over large areas in a very short time, is not comparable with capabilities of ground means used so far.

Geographical obstacles and difficulties in large urban areas due to reflections, usually encountered by ground means, disappear with an aircraft. The use of an aircraft can also be considered precious for rapid localization of plants interferring aeronautical communication and radionavigation services. In such cases often the interference is experienced by a plane flying at high altitude and the source may be positioned very far from the vertical, so ground mobile stations might have to explore large areas to locate it.

REPORT 834

RECENT DEVELOPMENTS IN LOW-COST MOBILE HOMING SYSTEMS

(Question 28/1)

(1982)

1. Introduction

Question 28/1 seeks information relating to the techniques and equipment available for direction finding at all frequencies for fixed, mobile, and portable applications. This Report will address developments which have taken place in mobile homing apparatus for practical use in bands from MF to UHF. The equipment uses components which are low in cost and complexity and which are widely available. The equipment is suitable for use with a variety of receivers. It is also compatible with simultaneous aural monitoring and recording, if desired.

2. Principle of operation

One general class of mobile direction finding equipment (used in car, truck or aircraft) operates on the principle of a homing system which continuously indicates the direction to the transmitter relative to the heading of the vehicle on which the direction finding apparatus is mounted. The operator of the vehicle moves in the general direction of the indicated bearing, making appropriate choices of available branching routes to eventually arrive at the location of the transmitter. Thus the homing system is not used for plotting bearings in the traditional sense of direction finding, but finding a transmitter can be done in about the time required to drive to it, assuming the transmitter is radiating. A high degree of accuracy is not required in the relative bearing indication as the distance to the transmitter is constantly diminishing thus reducing the absolute value of errors resulting from the system. The directional information is derived by comparing the instantaneous values of the outputs of two cardioidal antenna systems carried on the vehicle.

Three kinds of homing antennas (spaced whips, rear window defroster and dual defroster for estate cars) are described in this Report. The relative gain of each is shown in Fig. 1. Note that values of gain are identical for both defroster-type antennas. It is important to note that although gain is relatively low, the antennas are designed principally for concealed direction finding. Other characteristics are secondary. Low gain performance of the antennas can be somewhat compensated by low noise amplifiers, assuming signal to noise ratio is adequate in the first place. Furthermore, as the homing antennas described are intended to be used inconspicuously, they naturally are usable closer to transmitters where less gain is needed. As system voltage output is proportional to transmitter power, the system works well close to higher power transmitters operating at frequencies where antenna gain is low. For example, Fig. 1 shows the defroster type unit has only −70 dBi gain in the middle of the MF broadcast band, but it does still work well within several kilometres of a 10 to 20 kW broadcast transmitter. Often a direction finding assignment is passed from a long range fixed monitoring system to one or a group of mobile homing vehicles. In this case the mobile units may be directed to within a small area where the distance to the signal source is small. This enables a relatively low gain antenna of this type to operate satisfactorily.

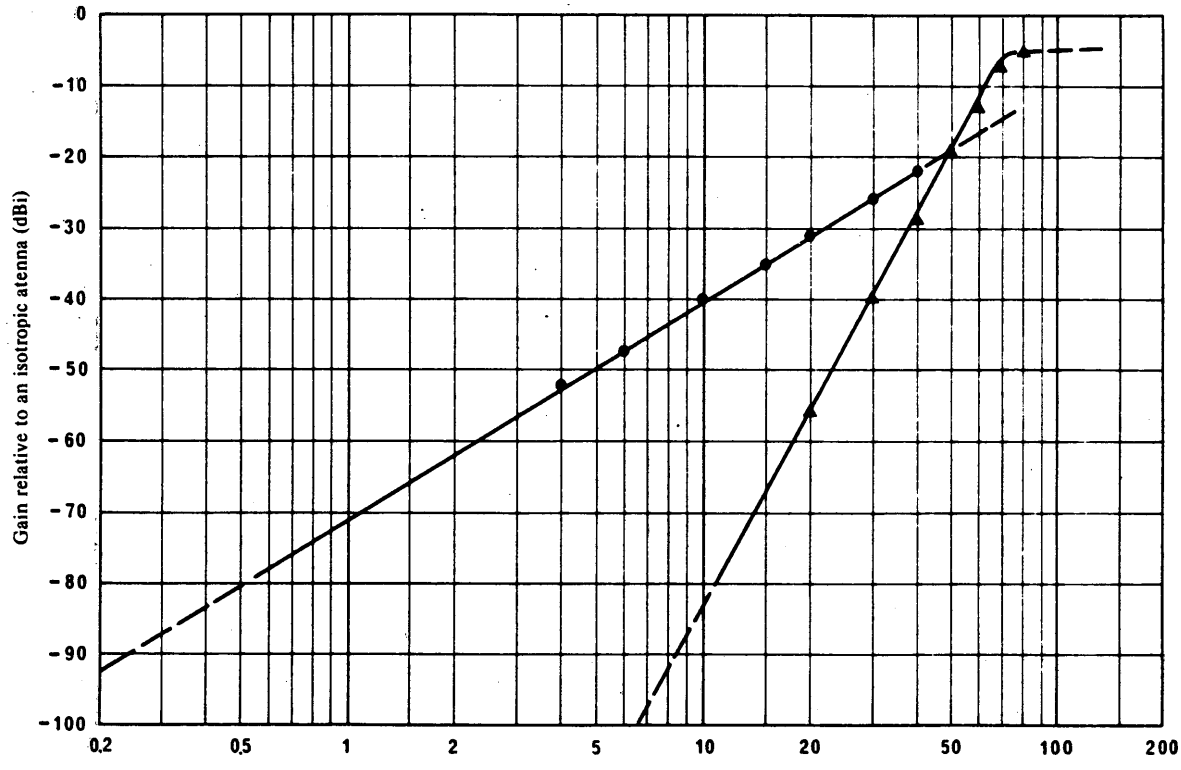


FIGURE 1 - *Antenna gain as a function of frequency*

●: data points for both the rear window defroster and the dual defroster grids for estate cars (a 94 cm horizontal spacing of vertical grid elements, mounted in a larger sized window, gave approximately a 40 MHz cut-off)

▲: data points for a parallel spaced, whip antenna system where whip length above a ground plane was 80 cm, and horizontal whip spacing was 80 cm

3. Mobile direction comparator

The output of pairs of antennas having cardioidal reception patterns are processed in the comparator. The two antenna outputs are modulated with low level, low frequency antiphase signals and are summed before passing to the receiver. The composite signal developed at the receiver IF port then contains a modulation component which varies in phase and amplitude depending upon the direction of arrival of the transmission relative to the vehicle heading. The mobile directional comparator detects this signal component and compares the phase of the component to the reference phase to develop an output indication on a direct current meter which instructs the driver to continue forward, turn left, or turn right. Deflection of the meter is approximately linear with heading relative to the vehicle so the driver rapidly learns the degree of course correction required.

4. Isolation differential combiner (IDC)

One of the antenna systems which is compatible with the homing system processor utilizes two parallel, matched vertical whip antennas which are combined in the IDC unit to produce the necessary dual opposed cardioidal patterns. The principle of combining differentially the outputs of two whip antennas (one of which is suitably delayed in time) to produce a cardioidal pattern has long been known. The purpose of the IDC unit is to simultaneously produce two opposed cardioidal patterns from two such whip antennas so that they can be individually modulated and subsequently recombined before entering the receiving equipment. This operation is performed by separating the output of each antenna, in a wideband hybrid, into two components. One component of each antenna is then delayed through a lumped tranmission line while the other is combined differentially with the delayed component from the opposite antenna. The delay line is chosen to equal the maximum propagation time between the antennas. In a practical system the resulting cardioidal pattern is useful up to the frequency at which the antenna spacing is approximately 3/8 wavelength. The useful lower frequency limit, as shown in Fig. 1, is only a function of sensitivity. The upper frequency limit is directly a function of horizontal spacing. Sensitivity is adjusted by changing whip length.

5. Rear window defroster antenna

This antenna is a special purpose unit for applications involving limited frequency ranges where concealable operations are desired. The two cardioidal patterns are developed from an antenna which is mounted on the rear window of a sedan type vehicle and to external observers appears to be a common electrical defroster grid. Note that this antenna was designed for vehicles without an actual defroster, and will not work with one superimposed. No attempt has been made to determine feasibility of combining both homing and defroster functions in a single grid. The design was derived empirically following observations that for some frequency bands, the window area of a sedan was transparent to arriving wavefronts with little distortion of the directional information. Referring to Fig. 2, the primary portions of the antenna consist of two vertical (as viewed from directly behind) elements whose outputs are combined in base mounted summing transformers with a delayed component from the opposite element which arrives by means of a multiwire-transmission line. The impedance of the multiwire line is determined by the number and spacing of the horizontal wires and this characteristic impedance is matched to the processor input impedance by the summing transformers. The patterns produced at the output ports are two cardioids which are at a maximum to the right and left of the vehicle with nulls slightly shifted to its rear. The design pictured works best for vehicles with large offset ground planes represented by the roof and trunk areas and is practically useless for glass fibre or estate-car bodies. Estate-car type vehicles can be accommodated by using two identical grids similar to that shown in Fig. 2, one on each rear *side* window. Fore/aft information is then available at the output ports of each antenna grid. By summing these side ports using delay lines, right/left information may also be derived. The useful lower frequency limit, as shown in Fig. 1, is only a function of sensitivity. The upper frequency limit is directly related to size of the window and metal framework within which the defroster antenna is mounted. If the grid were mounted on the vehicle's roof, for example, upper frequency limit is only a function of vertical element spacing.

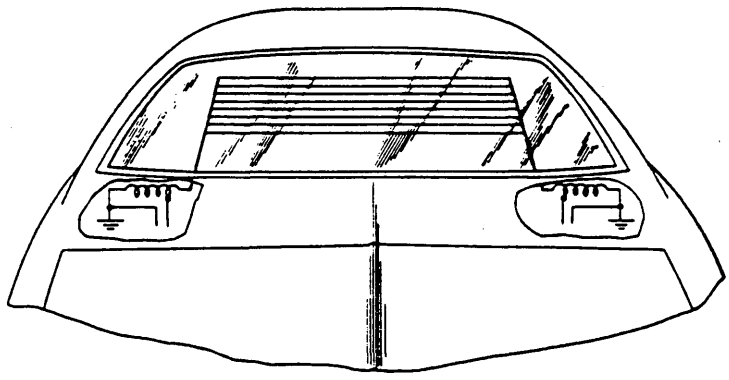


FIGURE 2

6. Antenna pattern combiner

The antenna systems previously described provides only right/left cardioid patterns and the homing system indicates only if the transmitter is to the right or left of the vehicle heading. Information on either fore or aft location is provided only as the vehicle makes a course correction and the meter is observed to determine if the deflection increases or decreases. An increase in deflection indicates that the vehicle is heading away from the transmitter site. By adding another element which has a loop type pattern to the existing antenna configuration in the rear window and suitably combining the additional output with the existing outputs, fore/aft information can be derived to supplement the right/left data. The choice of fore/aft or right/left sampling can be switch-selected by the operator to allow rapid determination of the emitter's quadrant relative to the vehicle heading. For generating the fore/aft cardioids, the right and left cardioid patterns are first combined to generate an omnidirectional pattern which is then combined either in or out of phase with the central loop type pattern to produce maximums in either forward or reverse directions. The location of the additional element and the resulting patterns at various ports are shown in Fig. 3.

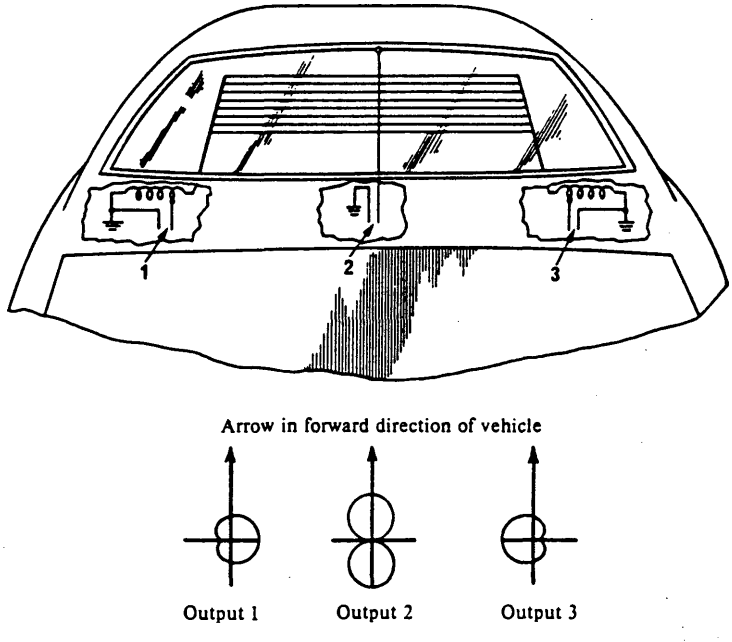


FIGURE 3 - *RF pick-up pattern*

7. Future developments

The equipment described represents the result of considerable effort in the development, production, and testing of mobile direction finding equipment to support field activities. The development effort is still continuing and further advances are expected. Present effort is directed toward providing equipment suitable for use from MF to UHF bands in a single package which can be securely installed and concealed in a medium or small vehicle. It is hoped to eventually provide continuous and accurate indication of direction of arrival to the driver throughout a full 360° range relative to the vehicle heading. Use of remotely controlled receiving equipment with the provision of a small, unobtrusive control panel, would avoid crowding of the passenger compartment and could improve operational convenience.

REPORT 1106

DEVELOPMENTS IN AUTOMATED REMOTE MONITORING

(Question 29/1)

(1990)

1. Introduction

The need to provide low cost monitoring facilities in areas that are remote from existing fixed monitoring stations had led the Canadian Administration to develop a Remote Interactive Monitoring System (RIMS). This system can be situated at a remote unattended site and operated by a personal computer (PC) from a control site. The remote control of the monitoring system results in more efficient use of monitoring staff.

The system is composed mainly of "off-the-shelf" components which are rugged and easily transportable and need only AC power and access to a telephone line. Communications with the remote site is achieved through a 300/1200 baud data/voice alternate modem.

The monitoring system software permits the remote monitoring for enforcement purposes of channel occupancy, the recording of voice transmissions and the direct communication by administration staff with the users of the frequency channel being monitored. This interactive capability has proven to be an effective means of monitoring channel activities ensuring circuit discipline and thus eliminating radio channel misuse.

The system has a user friendly menu-driven software and automatic reset in case of power outage. In addition, it is easily transportable from one site to another as it requires only $0.23 m^3$ of rack space. The present software is capable of the scanning of 20 preselected channels over a predetermined time period. Initially, the system has been used to monitor the mobile service in the VHF bands.

2. Remote Interactive Monitoring System Description

At the remote site the RIMS equipment consists of a personal computer and an interface board to control VHF transceivers and an audio tape recorder. The remote units are controlled via a telephone line using a data/voice alternate modem and a PC at the control site. Figure 1 shows the block diagram of the RIMS.

The PC at the remote site uses a self-booting multi-task operating system to ensure auto start after power interruptions and is equipped with a 10 megabyte hard disk, two parallel ports, a data/voice alternate modem and a voice synthesizer.

3. System Operation

The menu driven software permits three modes of operation, Channel Occupancy, Channel usage and Channel Interaction as described in the following sections.

3.1 Channel Occupancy Mode

The VHF Mobile transceiver can scan up to 20 frequencies every 2.5 seconds over a predetermined time period. The occupancy data collected can be accessed and the results transferred to the control site where a hard copy is produced.

Occupancy studies are made on a daily, weekly or monthly basis. If a channel is found to have a high occupancy level, then this channel can be observed directly using the Channel Usage Mode.

3.2 Channel Usage Mode

In this mode, the RIMS records the voice transmission on preselected channels. These preselected channels are continually scanned until a transmission is detected. The scanning then ceases and the transmission is recorded for a specified time period. When this period ends, a channel identification and time of day message is recorded using voice synthesis techniques. Channel scanning is then resumed.

The recording of the Channel Usage study can be transferred via the telephone line to the tape-recorder at the control site. The Channel Usage results permit the determination of the misuse/abuse of the monitored channels.

3.3 Channel Interaction Mode

This mode permits the operation of the remotely sited VHF transceivers via telephone line and the direct mobile radio interaction with the users of the bands. This direct communication with the users effectively corrects inappropriate use.

4. Conclusions

The IMS was tested at a remote site in a maritime environment, 500 km from the control site. The Interaction Mode was used to establish satisfactory communciations with the mobile radio units being monitored. This system has potential benefits for monitoring in remote locations due to its low cost, unattended operation and the capability of transporting the equipment from place to place.

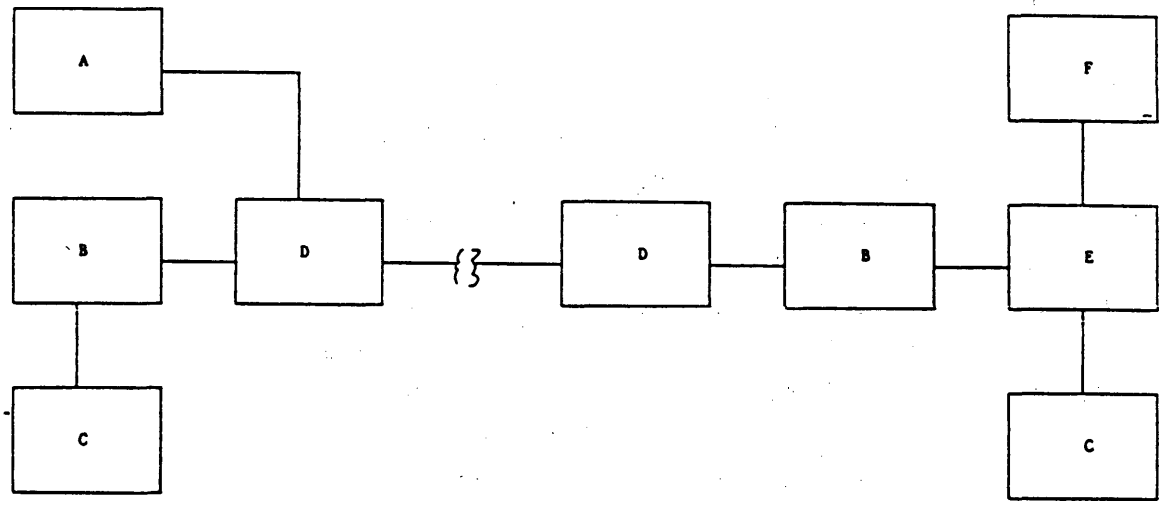


FIGURE 1 - Remote interactive monitoring system

A. TELEPHONE
B. PERSONAL COMPUTER
C. RECORDER
D. DATA/VOICE ALTERNATE MODEM
E. INTERFACE CONTROLLER
F. RADIO TRANSCEIVERS

REPORT 1107

THE DIGITAL SIGNAL PROCESSING METHODS IN SPECTRUM MONITORING

(Question 22/1)

(1990)

1. INTRODUCTION

Digital signal processing has been increasingly used in many fields of wired and wireless communications. In radio monitoring also, the obvious advantages of digital signal processing may improve the accuracy, repeatability and speed of measurements.

The current level of technology does not allow the replacament of the functions of a radio-frequency measuring system covering a wide frequency range. With the increasing speed and dynamic range, however, digital solutions can step-by-step, take over the functions of analog circuits starting at the baseband output and moving toward the front end.

By using general-purpose or dedicated hardware and software, an integrated monitoring system can be built which performs several monitoring tasks (e.g., measurement of field strength, frequency, bandwidth, modulation characteristics and direction of incidence) with the same instrumentation.

The purpose of this Report is to give a condensed overwiev of the various digital signal processing methods suitable for different tasks of radio monitoring. Methods for the sorting, evaluating and storing of monitoring data are not covered here.

In **Figure 1** the block diagram of a radio-monitoring system with digital signal processing facilities is shown. This system, in principle, can be used for any radio monitoring task depending on the actual selection of instruments.

2. SUMMARY OF TERMS

The bandlimited intermediate-frequency signal can be written in the form of

$$S(t) = A(t).\cos(\theta(t))$$

where A(t) is the amplitude function
θ(t) is the phase function

This signal may consist of the desired signal(s), interfering signal(s) and additive noise.

For digital signal processing purposes, S(t) should be quantised both in the amplitude and time domain. Usually, but not necessarily, the quantization is combined with orthogonal mixing and resampling/decimation (Jondral, 1985).

Three parameters of the input signal S(t) are usually defined in the time domain:

$$A(t) = \sqrt{S_R(t)^2 + S_I(t)^2}$$

$$f(t) = d\varphi(t)/dt$$

$$\varphi(t) = \arctan\left\{S_I(t)/S_R(t)\right\}$$

where A(t), f(t) and φ(t) are the instantaneous amplitude, frequency and phase. $S_R(t)$, $S_I(t)$ are the real and imaginary components of S(t).

For spectral-related methods, a data block of N (usually complex) samples are transformed to the frequency domain by DFT (Discreet Fourier Transform) or FFT (Fast Fourier Transform):

$$S(k\cdot\Delta t) = S_R(k\cdot\Delta t) + jS_I(k\cdot\Delta t) \qquad k=0,1,2,\ldots N-1$$

$$F(i\cdot\Delta\omega) = \frac{1}{N}\sum_{k=0}^{N-1} S(k\cdot\Delta t)\cdot\exp(-j2\pi ik/N) \qquad i=0,1,2,\ldots N-1$$

The sampling and quantization results in specific features of the spectra obtained by DFT or FFT:

- spectra are also quantized along the frequency axis ($\Delta\omega$)
- spectra are periodic for N

Also, finite sample length causes the smearing of sharp spectral peaks (leakage). The leakage can be reduced and the equivalent selectivity curve of DFT or FFT can be shaped by applying an appropriate weighting to the input signal. For more detailes, including description of spectrum of noise, sinusoidal signals and sinusoidal signals embedded in noise see (CCIR 1986-90).

2.1. Frequency measurement by interpolation of spectra

Using a data block in the frequency domain, it is possible to reduce the inherent uncertainty of obtained spectrum by interpolation. After the course search for peaks of sinusoidal signals, the spectral lines of highest levels around the selected peak are used to calculate the interpolated frequency. The ultimate improvement factor in frequency measurement is roughly proportional to the signal-to-noise ratio of the spectral peak. It is also possible to calculate the confidence range of measured frequency value from a single block of spectrum.

For formulae of interpolation see (Novak, 1987).

2.2. Bandwidth measurements

One suggested method for bandwidth measurement, well suited for digital signal processing, can be found in Recommendation 328, paragraph 4.2. In the DFT or FFT spectra, in-band and out-of-band powers are obtained by simple summation of spectral lines. The span of spectrum should be sufficiently wide to accomodate the signal to be measured. On the other hand, an unnecessarily wide span reduces the resolution of measurement. The resolution of bandwidth measurement is in the order of the resolution of spectrum, which is equal to the selected span divided by the number of spectral lines.

Since the DFT or FFT spectrum is not consistent, the variance of stochastic modulation content can be reduced by averaging.

2.3. Measurement of modulation

Measurement of modulation characteristics can be implemented in different modes: using digital demodulation techniques, the processing of demodulated amplitude, frequency and phase waveforms is possible, allowing the calculation of AM modulation depth as well as angle deviation.

Without demodulating the signal, the modulation depth of AM signals with carrier is also available from the spectrum. The peak of carrier is to be identified and its level measured, followed by calculation of sideband power. The modulation depth is equal to the square root of sideband power divided by the carrier power. In case of dynamic carrier control (DCC), the measured sideband power can be normalized to the nominal carrier power. Since the modulation depth is given by the ratio of powers, no level calibration in the receiving path is necessary.

3. PRACTICAL CONSIDERATIONS

3.1. Instrumentation

For implementing the spectral-related methods, most of the necessary hardware is available in the form of general-purpose instruments. For digital demodulation, however, dedicated hardware and software is necessary. Some of the recent signal analyzer instruments provide one or more demodulation functions, where the universality of the instruments are paid for by increased cost and reduction of speed.

Automatic measuring and monitoring equipment containing different levels of digital signal processing were developed in Japan, the Federal Republic of Germany, in the USA (Report 668) and in the U. K. (CCIR, 1988).

3.2. Sampling

The sampling characteristics of the incoming signal determines the ultimate resolution and accuracy of output data. During the course of sampling, incoming data are quantized in both the time and level domain.

The f_s sampling frequency must be chosen to obey the sampling theorem. For unknown signals an antialiasing filter is to be used to cut out incoming components above $f_s/2$. The size of frames (N) determines the frequency resolution in baseband DFT. The accuracy of frequency measurements can be further enhanced by interpolation.

3.3. Antenna, receiver

Location of measuring antenna should be selected so that it will not pick up noise from the digital equipment.

For digital signal processing purposes, the receiver of the measuring system should have synthesizer tuning with a tuning step matched to the available spans of signal processing. To retain the highest possible dynamic range, all local oscillators for heterodyne mixings should have low phase noise so that it would not mask low-level signals in the close vicinity of a large signal.

In order to be able to carry out most of the measuring and monitoring tasks with the same unit, the receiver should preferably have a calibrated signal path for measuring, parallel with a receiver section used for communications purposes (AGC, MGC, various demodulators, etc.)

4. EQUIPMENT DEVELOPED IN HUNGARY

A high-resolution monitoring system based on digital signal processing was developed and has been regularly used by the radio-monitoring service in Hungary since 1967. The system measures received signal level (field strength), frequency, AM modulation depth and bandwidth of emissions in the frequency range of 150 kHz to 30 MHz.

The system consists of a calibrated loop antenna, a high-performance receiver and a controller with signal

processor. The receiver combines the advantages of a communications and a measuring receiver. Owing to the low phase noise of the analog part, the full dynamic range of the FFT analysis is available for the operator to select, measure and identify signals in both manual and automatic operation.

The FFT analyzer section provides a panoramic display with selectable line bandwidth down to 2.5 milliHertz. Bandwidth and frequency measurements as well as evaluation of complex emissions are possible with very high accuracy. An external frequency standard can be connected to the system to lock all the instruments (receiver, demodulator and FFT analyzer) to the same reference.

Using the latest advances in science, the system is supplied with measuring algorithms which enable the automatic recognition and high-accuracy measurement of sinusoidal signals (e.g., carriers and sinusoid modulation signals). Beside signal recognition, an unprecedented accuracy in frequency measurements is also possible with a significant reduction of measuring time. An ultimate accuracy of 100 microHertz is guaranteed.

The DSPU allows the user to store digitized data or programs on floppy drive, to run system softwares in stand-alone mode or to control the measuring system through IEC bus from an external computer.

4.1. SYSTEM DESCRIPTION

The configuration of system is shown in **Figure 1.**

Receive path

The receive path of the measuring system is routed from the antenna through the receiver to the signal processing unit. In the receiver high 1st intermediate frequency (72.5 MHz), a set of electromechanical, crystal and active filters, low phase-noise synthesizer and a low-noise front end with high intercept point guarantee good reception.

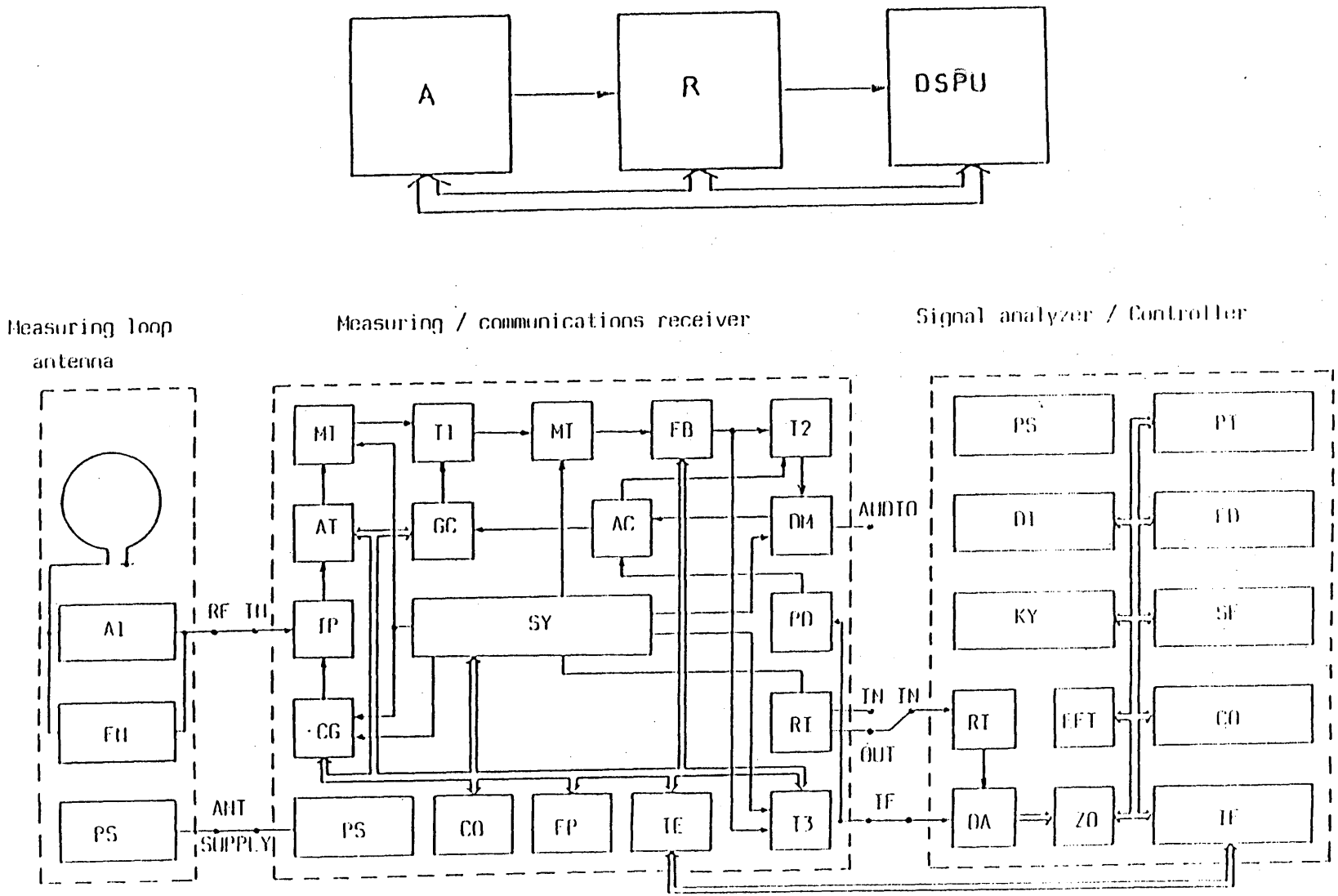


FIGURE 1

Block diagrams of radio-frequency monitoring system with digital signal processing developed in Hungary

Legend to the Figure 1

A: antenna (system)
R: receiver(s)
DSPU: digital signal processing unit with the necessary peripherals

A1: Antenna amplifier
AC: AGC circuit
AT: RF attenuator
CG: Calibration generator
CD: Controller
DA: Data acquisition
DI: Display
DM: Demodulators
FB: Filter bank
FD: Floppy drive
FFT: Hardware for Fast-Fourier Transform
FN: Feedback network
FP: Front panel
GC: Gain control circuit
I1: IF Amplifier at 72.5 MHz with AGC
I2: IF amplifier at 200 kHz with AGC
I3: IF amplifier and converter at 200 kHz/10 kHz with calibrated gain
IE: IEC 625 interface
IP: Input protect circuit and calibration switch
KY: Keyboard
MI: Mixer
PD: Peak detector
PI: Printer interface
PS: Power supply and EMC filter
RI: Interface for 5 MHz external frequency reference
SE: Spectrum evaluation firmware
SY: Synthesizer
ZO: Zoom processor

The IF signal path of a communications receiver with MGC or AGC and that of a measuring receiver with calibrated gain operate simultaneously. The band limited 10 kHz IF signal is digitized by the signal processing unit and a zoom filter sets the final bandwidth for analysis.

All conversion and sampling signals are generated from the same oven-controlled crystal oscillator which can be locked to an external frequency reference.

The spectrum of the IF signal is evaluated by dedicated algorithms.

RF IF section

The signal from the 50 ohm antenna input is connected to an input protective circuit, followed by a switch for inserting the calibration signal. RF attenuation can be varied within the range of 0...50 dB in 10 dB steps.

After conversion with a 1 Hz step synthesizer, selectivity at the 1st IF is provided by a crystal filter with 14 kHz bandwidth. The variable gain of the 1st IF stage is used in both AGC and calibrated modes.

After conversion to the 2nd IF of 200 kHz, main selectivity is provided by a set of electromechanical filters. The signal path is splitted here: in the communications section a variable-gain IF amplifier is followed by conversion to 10 kHz. With active filters at 10 kHz, the set of bandwidth values is: 150 and 400 Hz, 1, 1.4, 2, 3, 6 and 14 kHz.

In the calibrated path an aperiodic 200 kHz IF amplifier is used with 0...70 dB range of gain in 10 dB steps, followed by wide-band conversion to 10 kHz. The nominal full-scale level of low-passed 10 kHz output is 1 Vrms behind a source impedance of 1 kohm.

Gain adjustment of the calibrated path is provided by a D-A circuit, which, during calibration periods, is driven by the error signal of the peak IF signal compared to a precision voltage reference. An 80 dBμV tracking signal for calibration is generated from the synthesizer signals.

Synthesizer

The synthesizer provides the conversion signals for the IF section: a low phase-noise, 1 Hz step 1st local oscillator signal and fixed frequencies for the 2nd and 3rd conversions. Reference clock for the synthesizer is provided by an oven-controlled crystal oscillator, which can be replaced by external frequency standards. Error status of PLLs and the adequate signal level of the external reference are displayed by LEDs.

Demodulators

They convert the IF signal into audio band in all of the commonly-used transmission modes of HF range. The FSK demodulator also provides logic commands. The continuous tuning range of BFO is +- 2.5 kHz. AGC rectifiers and digitally-tuned AFC circuit is also part of the demodulator section. The tracking range of AFC is not limited.

Data acquisition, zoom

The 10 kHz IF is connected to the data acquisition module of the digital signal processing unit. After the anti-aliasing filter, quantization is done on 12 bits at a rate of 51.2 kHz. The sampling clock is crystal controlled, and can be replaced by an external frequency reference.

The quantized signal is processed by a zoom filter allowing further reduction of analysis bandwidth in 1, 2, 5, 10 times steps down to 1 Hz.

FFT

Computation of spectrum is completed within 250 msec by dedicated hardware. The 1024 point FFT provides a 400 line spectrum, which, considering the 12 bit quantization, yields about 70 dB dynamic range.

The total analysis bandwidth and the width of an FFT line can be set in 1, 2, 5, 10 times increments in the range of 1 Hz - 20 kHz and 2.5 milliHz - 50 Hz, respectively. Spectra

can also be averaged in linear, exponential and maximum hold modes.

Spectrum evaluation firmware

The computer of the digital signal processing unit evaluates spectra according to user-selectable input parameters and measurement results together with the corresponding spectra are continuously updated on the screen.

Evaluation of spectra includes automatic recognition of sinusoidal components, high-resolution high-accuracy measurement of frequencies and levels (field strengths) of sinusoids and bandwidth as well as AM modulation depth measurement of modulated signals. Owing to its unique algorithms, confidence ranges of the measured frequencies are provided for each instantaneous spectrum.

The user-friendly operation, besides of correction for the inherent errors of digital signal processing, significantly cuts measuring time by providing all selected measurement results in parallel and by speeding up high-accuracy frequency measurements with interpolation.

Control units with microprocessors

Microprocessors are used in the controllers of receiver and signal processing unit to handle keyboards, displays and peripheral interfaces (printer and IEC 625 bus).

Controller in the signal processing unit is switch selectable for both controller and talker-listener IEC 625 bus modes. It contains a microcomputer with user RAM which also offers programmability in high-level language.

Power supply and EMC filter

The receiver and digital signal processing unit can be supplied from 110-220V 50 Hz. The active loop antenna is powered from the receiver with +-12V DC. The power supplies are fitted with EMC filters. Careful mechanical design of instruments guarantee high immunity to external fields.

4.2. OPERATION

The operation of high-resolution monitoring system is explained by the help of a sample printout of screen, shown in **Figure 2.**

Both instruments of the measuring system, the receiver and the signal processing unit can be operated either manually or as an automatic system through the controller of signal processing unit or through the IEC 625 bus by an external controller.

The firmware offers easy-to-use, fast and accurate measurements. Besides of displaying living spectra of analyzed signals, the selected measuring parameters and measurement results corresponding to the spectrum on the screen are also output. Easy and fast operation is provided by selection of input parameters in simple radio-frequency terms: the optimum combination of receiver and signal processing parameters is set by the computer.

The most important input parameters can be changed in the function-list mode: actual fields are selected by an inverse highlight and their content can be changed on the screen by cursors or numeric keypad. Besides, simple dedicated commands are accepted during program operation. Finally, operators can make use of a series of commands accessible from the high-level programming language to make up complex measuring routines (e.g., channel scan with predefined parameters).

Alphanumerical information parallel to spectra on the screen of signal processing unit can be grouped into three categories:

output fields (measurement results)
input fields (measurement parameters, selectable from the function list)
auxiliary informations

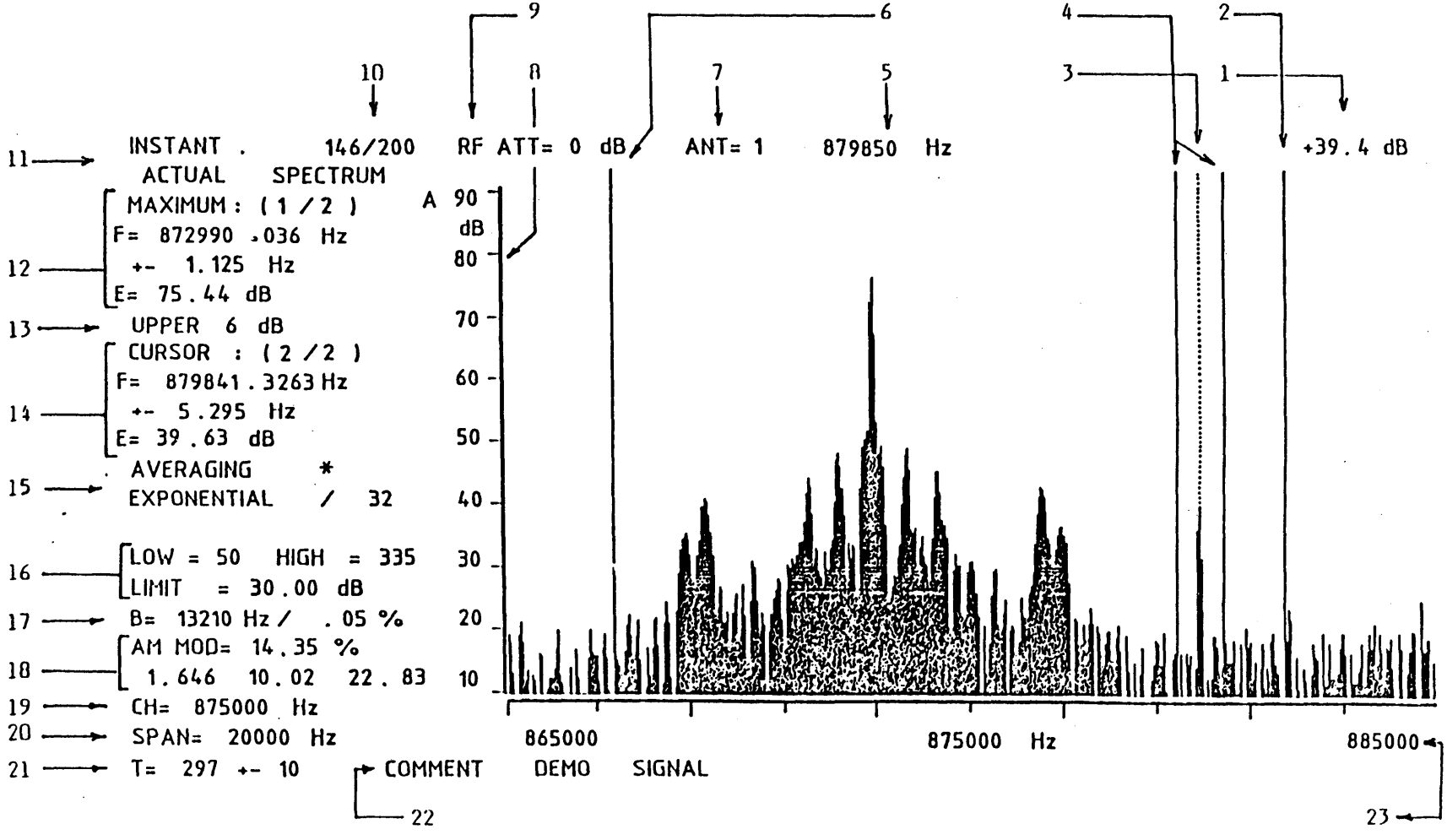

INSTANT . 146/200
ACTUAL SPECTRUM
MAXIMUM : (1 / 2)
F= 872990 .036 Hz
+- 1.125 Hz
E= 75.44 dB
UPPER 6 dB
CURSOR : (2 / 2)
F= 879841.3263 Hz
+- 5.295 Hz
E= 39.63 dB
AVERAGING *
EXPONENTIAL / 32
LOW = 50 HIGH = 335
LIMIT = 30.00 dB
B= 13210 Hz / .05 %
AM MOD= 14.35 %
1.646 10.02 22.83
CH= 875000 Hz
SPAN= 20000 Hz
T= 297 +- 10
COMMENT DEMO SIGNAL
RF ATT= 0 dB
ANT= 1
879850 Hz
+39.4 dB
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FIGURE 2

The numbering in the following list of fields refers to that used in **Figure 2.**

Output fields:

1., 5.: Cursor level and frequency: show the level and frequency of the spectrum line selected by the graphic cursor. The cursor can be set line-by-line, displayed data are not subject to error correction (interpolation).

10.: Spectrum counter: displays the number of spectra analyzed from the beginning of measurement or since the counter was cleared.

12., 14.: Maximum and cursor fields: contain the frequency, confidence interval of frequency and level (field strength) of the largest level and cursor-selected sinusoidal components found by the evaluation in the spectrum. In the headers, in brackets, the numbers separated by slash show the actual number and total number of sinusoids, respectively. All values shown are subject to corrections by interpolation and correspond to the actually displayed spectrum.

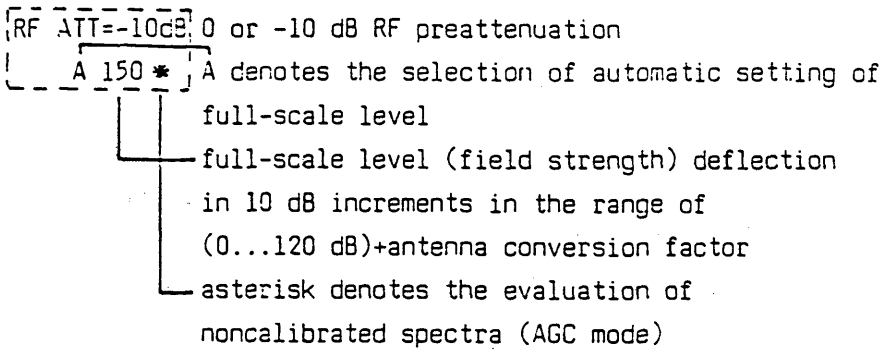
13.: Overload indicator : **indicates the too large input signals**

17.: Bandwidth: displays the bandwidth of signal in the frequency band selected by the low and high spectral line boundaries with the given out-of-band power ratio.

18.: AM modulation depth: displays the modulation depth of AMDSB/FC transmissions calculated from the level in the maximum field taken as carrier and the summed sideband lines as modulation sideband power. The instantaneous effective value in % behind the header are followed by the minimum, average and maximum values in the next line.

Input fields:

9.: Full-scale deflection:



11.: Type of spectrum: instantaneous actual
instantaneous stored
averaged actual
averaged stored

15.: Type of averaging: linear } averaging number up to
exponential } 2048
maximum hold

16.: Evaluation boundaries:
the frequencies according to the low and high spectral line boundaries select the actual frequency band of analysis for signal identification, bandwidth and modulation depth measurements. The limit contains the user-selectable threshold level below which all signal components are neglected in signal identification.

19.: Channel frequency: center frequency of the analyzed band. Selectable with 1 Hz increments in the range of 150000 and 29999999 Hz.

20.: Analysis bandwidth: bandwidth of analysis set by signal processing. Selectable in 1, 2, 5, 10 times increments in the range of 1 Hz to 20 kHz.

21.: Tracking: with the tracking mode switched on, the frequency of sinusoidal spectral line selected by the graphic cursor is tracked within +- spectral line range shown behind the actual line number.

Auxiliary informations:

2.: Upper boundary (solid line) of frequency band selected for spectrum evaluation

3.: Graphic cursor (dotted line)

4.: Boundaries of tracking range (solid line)

6.: Lower boundary (solid line) of frequency band selected for spectrum evaluation

7.: Antenna selector

8.: Level axis of displayed spectrum

22.: Comment line

23.: Frequency axis

Sample commands:

CLS: clears the spectrum counter and statistics
R: reads the actual status (channel frequency, bandwidth, full-scale level) of receiver
C: starts calibration of receiver (approx. 400 msec)
A: starts automatic setting of full-scale level of receiver
T: switches tracking on and off

4.3. Main characteristics

Main characteristics of the system are shown in Table I.

TABLE I

Main characteristics of the high-resolution monitoring system with digital signal processing

Frequency range	150 kHz-30 MHz
Sensitivity (EMF) in AGC mode A1A, Bw=400 Hz, (S+N)/N= 10 dB A3E, Bw= 6 kHz, (S+N)/N= 10 dB f_{mod}=1 kHz, m= 80% J3E, Bw=2,4 kHz, (S+N)/N= 10 dB	 0,4 µV 1,5 µV 1 µV
Full-scale level in calibrated mode	0,10,20,...100,110, 120 dBµVrms at the 50 ohm input
Synthesizer step	1 Hz
Frequency accuracy	within $1x10^{-7}$, connection for external frequency reference
Field strength accuracy	within +- 2dB
IF bandwidth (analog)	150 Hz, 400 Hz, 1 kHz, 1.4 kHz, 2 kHz, 3 kHz, 6 kHz, 14 kHz
Analysis span (signal processing)	1 Hz-to-20 kHz in 1,2,5,10 times steps
Display of spectrum	400 line FFT on a 12" screen
Quantization	12 bit, 51.2 kHz

TABLE I (continued)

Analysis:	
signal identification	5 point search for sinusoidal signals
frequency measurement	with 3 point interpolation and calculation of confidence range
modulation measurement	AM effective modulation depth
bandwidth measurement	with 0.1...49.9% out-of-band power ratio

5. MEASURING RESULTS

The hardcopy in **Figure 3**, shows common-channel interference on a LF channel, measured without external frequency reference. Owing to the 1 Hz full span, the two carriers with an only 60 milliHz frequency difference are well separated in the spectrum. The equivalent 2.5 milliHz digital channel spacing provides a very low background noise level, thus, in spite of the about 20 dB level difference of carriers, the smaller signal can also be accurately measured.

The advantage of spectrum post processing is clearly shown by the 25 microHz and 140 microHz confidence ranges of measured carrier frequencies. The same accuracy in frequency measurements by conventional means would usually require an equivalent 40000 and 7000 sec measuring time, respectively. Considering the 400 sec total sampling time for the 1 Hz span used, it means a 100 as well as 17 times reduction of measuring time.

The hardcopy in **Figure 4**, shows the modulation characteristics of a transmitter measured at a remote place. The symmetrical sidebands are clearly recognizable in the averaged spectrum. The measuring system automatically identifies and measures the level and frequency of carrier, the effective AM modulation depth and its statistics according to the selected out-of-band power ratios.

An optional user program of the measuring system extends its capabilities to on-site measurements of detailed transmitter parameters on calibrated, rectified signals taken from the transmitter. The measured parameters are: carrier power, sideband power, positive and negative peak modulation depth, effective modulation depth, bandwidth. With the exception of bandwidth, the statistics of measured parameters (minimum, average, maximum) are also available and displayed on the screen. For measuring transmitters with dynamic carrier control (DCC), all results of modulation measurements can be (optionally) normalized to a user-selectable carrier power.

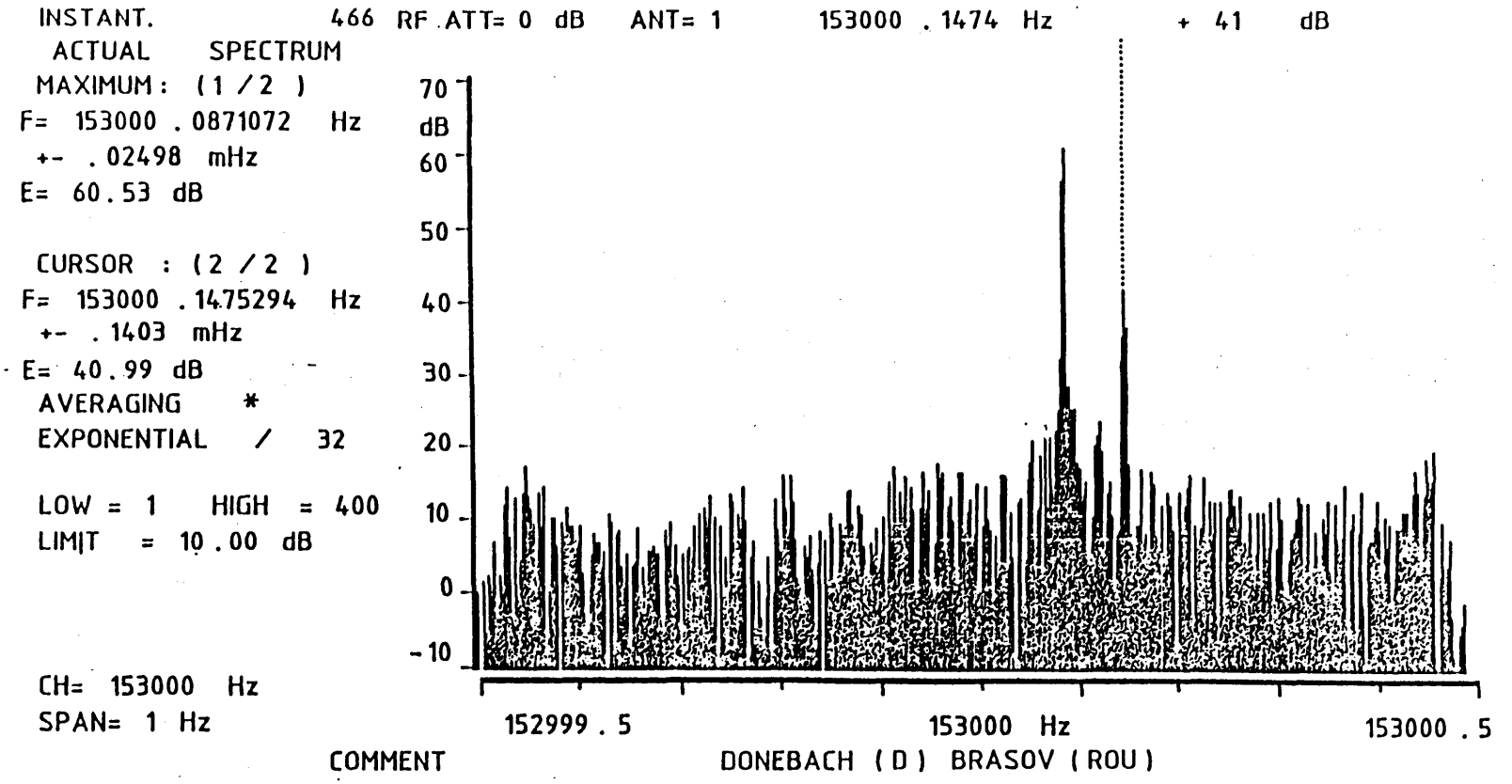


FIGURE 3

Sample hardcopy of high-resolution
common channel interference measurement

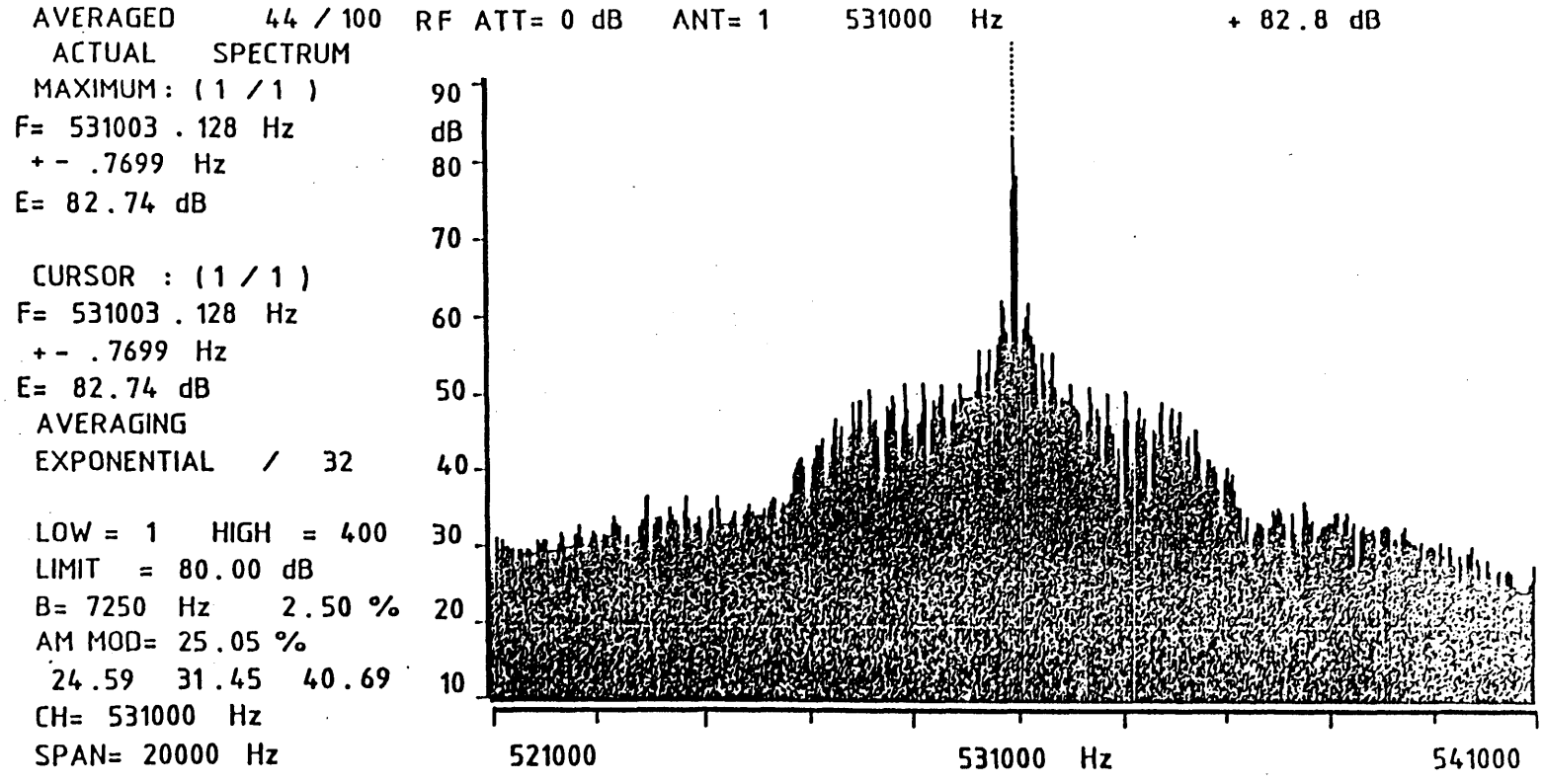


FIGURE 4

Sample hardcopy of remote measurement of transmitter characteristics

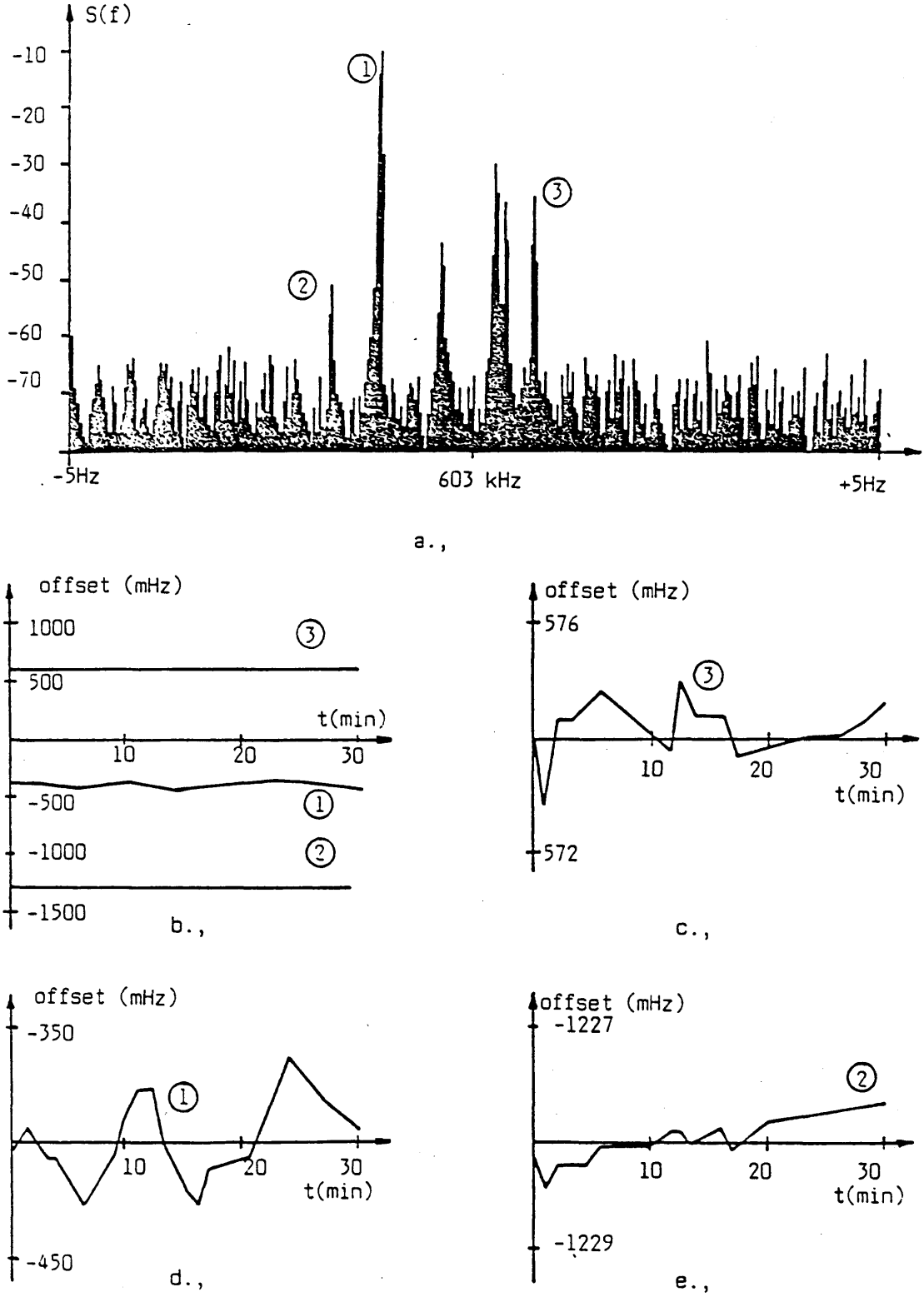


FIGURE 5

Result of repeated high-resolution frequency measurements with sample spectrum (Figure 5/a) and plotted frequency variations of the three carriers marked with numbers 1, 2 and 3 (Figures 5/b, 5/c and 5/d)

In Figure 5, the result of high-resolution frequency measurements of a MW channel is shown. The common-channel carriers were repeatedly measured over a period of time, and the frequencies of the three marked peaks together with a typical sample spectrum are shown in the figure. The plots with magnified scale clearly show the advantages of spectrum post processing: several carriers with very close frequency separation can be measured and monitored in parallel.

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[1986-90]: Doc. 1/51 (Hungary)
[1986]: Handbook for Monitoring Stations, Chapter V., paragraph 3.3

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REPORT 273-7

FIELD-STRENGTH MEASUREMENTS AT MONITORING STATIONS

(Question 24/1)

(1959-1963-1966-1970-1974-1978-1982-1986-1990)

Question 24/1 seeks information on various aspects of the measurement of field strength at monitoring stations, including the preferred methods and equipment for making such measurements in connection with propagation studies, measurement problems peculiar to certain special types of emissions (e.g., interrupted and reduced carriers and television signals), measurements in the presence of noise and interference; together with determination of the extent to which measurements at a distance can give useful information relative to harmonic levels at the transmitter itself; and a determination of the most useful programme which can be carried out to fulfil the needs of the IFRB, other Study Groups of the CCIR and other bodies.

In considering Question 24/1, reference should also be made to Report 227 which contains useful information concerning monitoring stations.

For information concerning the field strength resulting from two stable electromagnetic fields, refer to Report 516.

Further guidance on field strength measurements is contained in the Handbook for Monitoring Stations (Geneva, 1988) in chapter 7.

1. Determination at a distance of relative radiation levels of fundamental and harmonic frequencies of an emission

It would appear that this matter should be considered in two parts:

- conditions where both fundamental and the harmonic emissions to be measured arrive at the measuring point over a ground-wave or line-of-sight path;
- where either the fundamental or the harmonic emission, or both, arrive at the measuring site over a sky-wave path.

The situation as regards these two modes of propagation will be considered separately.

1.1 *Measurement of radiation levels under conditions of ground-wave or line-of-sight propagation*

Study Group 5, in response to a request for assistance on this matter, has referred Study Group 1 to Report 227. This Report contains detailed information on making and evaluating such measurements, taking into account instrument errors, differing attenuation rates at different frequencies and other variables.

1.2 *Measurements of radiation levels where sky-wave paths are involved*

Study Group 1 has considered this matter, as evidenced by Recommendation 329. Likewise, Study Group 6, in Report 252, reviews progress which has been made in determining the best method of calculating the field strength at a distance from the transmitter (ionospheric propagation), and describes plans for further study of this matter.

It is expected that the results of the further deliberations of Study Groups 1 and 6 will be of benefit to the international monitoring system in its efforts to minimize interference from spurious emissions. Pending the outcome of these deliberations, it would be highly desirable for stations in the international monitoring system to have certain guidelines which might be followed in obtaining evidence of possible excessive radiation at the harmonic frequency. While such evidence would not be conclusive, it might serve as an indication that close-in measurements should be made to verify the existence of harmonics at excessive levels. Appendix 8 to the Radio Regulations gives power limitations on spurious emissions as measured at the antenna transmission line input. Although these emissions may readily be measured at that point, there is no simple means of converting the spurious power limitations given in Appendix 8 to equivalent field strength at a distance. Since it is the actual radiated spurious field strength rather than the power into the antenna which is of primary concern as an interference source, it might be well to consider establishment of a limitation on spurious emissions, in terms of the maximum unattenuated field strength at about 1600 m (1 mile). Having once established such a limit, the maximum allowable inverse-distance field could readily be determined. Since the attenuation would actually increase more rapidly than inversely with the distance, the measured value of the spurious emission would be a conservative index of the actual radiated field, especially if the condition continues over a period of several days. This limitation would be in addition to, rather than a substitute for, the limitations on spurious power input to the antenna transmission line specified in Appendix 8. Admittedly, the results could not be compared directly with measurements made at the transmitter. However, it would appear that such standards might be of help to monitoring personnel in bringing about a reduction of interference due to harmonic radiation.

2. **Additional propagation studies that are desirable**

Opinion 68 points out the need for experimental data concerning HF sky-wave signal intensities and requests that administrations Members of the CCIR render assistance to Study Group 6 on this matter.

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REPORT 275-5

BANDWIDTH MEASUREMENT AT MONITORING STATIONS

(Question 26/1, Study Programme 26A/1)

(1953-1956-1959-1963-1966-1970-1974-1978-1990)

1. General discussion of measurement techniques, considering the definition of occupied bandwidth – No. 147 of the Radio Regulations

1.1 *Measurement of power ratios*

The present definition of occupied bandwidth suggests the principle described in Recommendation 328, §4.2. of measuring the ratio of the total power to the parts remaining outside of the bandwidth being measured. It would be necessary to locate the upper and lower edges of the band, by totalling the power in the out-of-band components on the high side until the 0.5% value is obtained and then repeating this procedure for out-of-band components below the band, starting in each case sufficiently far from the centre frequency so that no appreciable energy is omitted from measurement.

1.2 *Limitations in the accuracy of bandwidth measurements made at a distance from the transmitter*

Although a determination of occupied bandwidth of an emission, by the method of measuring total power and out-of-band power, can be accomplished when the measurements are made near the transmitter, this method is not generally applicable to measurements at a distance from the transmitter where the presence of interfering emissions or noise tend to mask the out-of-band signal components of interest. This is particularly true in the crowded MF (band 6) and HF (band 7) portions of the spectrum, regions of primary interest to international monitoring stations.

In spite of the very definite limitations which observations at a distance place upon the accuracy of measurements of occupied bandwidth, approximate determinations have been found of value when monitoring the spectrum for enforcement of bandwidth limitations. However, these non-precise measurements made at a distance, which are subject to inaccuracies for the reasons previously enumerated, should be considered as advisory only. Where greater accuracy is required, measurements at the transmitter may be desirable.

Paragraph 12 of this Report contains a comparison of the effect of noise on the accuracy achieved in the measurement of emission bandwidth by various methods.

1.3 *Bandwidth measuring method which appears to be practicable for use at monitoring stations*

It is the practice of some administrations, in evaluating spectrum occupancy of an emission from a distant transmitter, to use the method described in Recommendation 328, §4.1.1. In using this method, the x dB bandwidth is considered to include discrete components attenuated less than 26 dB below the peak level of the emissions. This procedure admittedly will not give a precise measurement of occupied bandwidth in terms of the definition in the Radio Regulations. For example, it is possible that a particular emission will have numerous low-level components on either side of the main emission such that their sum on each side would be equal to much more than 0.5% of the total mean power while having none of these discrete components exceeding the -26 dB level. In such a case, the occupied bandwidth, as determined at the transmitter by measuring power ratios, would presumably be somewhat greater than this x dB bandwidth measured at a distance by the method described in Recommendation 328, § 4.1.1.

According to [CCIR, 1966-69a], results are obtained which are in better agreement with No. 147 of the Radio Regulations by measuring the bandwidth of classes of emission at specially calculated "x dB" levels, the values of which are given in Table I below. In optimum emission conditions (Recommendation 328, § 2), the x dB bandwidth readings at these levels correspond to the necessary bandwidth values for the classes of emission concerned.

TABLE I

Class of emission	"x dB" levels	Observations
A1A, A1B	− 35	For α(1) $<3\%$ (all pulse shapes)
	− 30	For α(1) $\geq 3\%$ (all pulse shapes)
A2A, A2B	− 32	For a modulation of the order of 80 to 90%
F1B	− 25	For all signal shapes and a modulation index $2 \leq m \leq 24$
F3C	− 25	For all types of transmitted pictures and modulation indexes $0.4 \leq m \leq 3$
F7BDX	−28	For modulation indexes $9 \leq m \leq 45$ (2)

(1) The relative build-up time, α, of a telegraph signal is defined in § 1.21 of Recommendation 328.

(2) The modulation index for class F7BDX emissions was determined on the basis of a frequency deviation equal to half the difference between the extreme instantaneous frequencies of the transmitter.

Document [CCIR, 1966-69b], also states that these x dB levels for A1A, A1B and F1B classes of emission can be used to measure the bandwidth of these classes of emission in the following two cases: emission of periodic signals and of actual-traffic telegraph signals at least in stable field conditions. In the former case, the filter passband of the spectrum analyser should be several times smaller than the modulating frequency. In the second case, it should be slightly greater than the modulation rate (see Recommendation 328, §4.4). Zero level can be adjusted according to the maximum response of the spectrum analyser during three or four partial displays, i.e. according to the highest level of the maximum spectrum component during this number of partial displays.

According to [CCIR, 1974-78a], the measurement level −30 dB is the standard value applied in the USSR for different classes of emissions for checking of emissions "optimality". Report 324 contains the required relationships between the "reference bandwidth" determined at this level and the necessary bandwidth for a number of classes of emission.

2. Representative bandwidth measuring equipment

2.1 A spectrum analyzer, of the type described in Recommendation 328, §4.1.1, is suitable for use at monitoring stations. With this equipment, the emission spectrum is analyzed by passing each component successively through a narrow-band filter of fixed frequency, by heterodyning the signal with an external frequency, varied either automatically or manually.

For the measurement of narrow-band emissions, a spectrum analyser having a high degree of resolution is desirable, so that an accurate display of the emission spectral distribution may be obtained. A typical instrument has a maximum resolution of 10 Hz and provides a swept frequency range adjustable from 1 kHz to 100 kHz, together with a sweep rate adjustable from 1 to 30 sweeps per second.

For the analysis of wideband emissions, spectrum analysers are available which incorporate a complete receiver as well as those designed for use with general purpose receivers. A typical instrument covers a frequency range from 10 MHz to 44 GHz with a sweep width continuously variable from 200 kHz or less to as much as 70 MHz (at the higher frequencies). The sweep rate is adjustable from 1 to 60 sweeps per second.

A major shortcoming of spectrum analysers, which use a single filter to sweep the entire band under surveillance, is the incompatibility between high resolution and rapid sweep rates, especially where a band of considerable width must be studied. A fast sweep rate is necessary to obtain a representative display of transient components. However, as the sweep rate is increased, the resolution becomes poorer so that significant components of the emission will not be displayed accurately. A number of systems have been devised to minimize the sweep-rate versus resolution limitations. One method involves the division of the band to be observed into as many as eighteen sub-bands, each of which is individually swept. The eighteen outputs are then displayed one at a time on a single cathode-ray tube or simultaneously on separate tubes.

2.2 Document [CCIR, 1963-66], describes two methods of estimating the occupied bandwidth in terms of the power distribution of the emission. The first of the two methods is designed primarily for the HF portion of the spectrum, while the second is reported to be useful for comparatively broadband emissions in the VHF bands.

2.3 An energy spectrum recorder is described in [CCIR, 1959a]. This device makes a chart record of spectrum occupancy in terms of available power from a standard antenna.

2.4 A device for measuring frequency-shift, especially in questions concerning the monitoring of complicated emissions and those types of emissions using a synchronous high speed multi-channel system, is described in [CCIR, 1962]. This device permits simultaneous measurements of frequency shift of multiplex channels as well as a visual check on the frequency-shift signals.

2.5 Mikhailov *et al.*, [1973] describe an automatic equipment for measuring occupied bandwidths in the 150 kHz to 30 MHz range using the measuring method set out in §4.2.3of Recommendation 328. The equipment can also be used to measure x dB bandwidths by the method set out in §4.1 and4.1.1of Recommendation 328.

3. A somewhat different approach to the measurement of bandwidth is described as follows (see [CCIR, 1959b]):

Endeavours have been made to determine the bandwidth occupied by an emission, by way of the signal shape instead of by analysing the spectrum. Such experiments have been made for classes of emission A1A, A1B only, starting from the fact that the bandwidth occupied is a function of the shortest build-up (and decay) time of the emitted signals (see Recommendation 328, § 3.1.3).

As regards the effect of fading, a comparison between 60 single measurements of A1A, A1B emissions taken at the transmitter and in the far field (up to 425 km), resulted in a maximum deviation at 16%. The average value of the deviations amounted to 4.8%. The differences in the results of the measurements at the transmitter and those taken far away may be explained by the deformation due to the fading effects of the signals. Measurements made under extreme fading conditions will not yield satisfactory results.

The effect of interference was investigated by superimposing a received non-fading A1A, A1B signal with a simulated interference signal of variable frequency generated locally. It was found that the wanted signals could still be evaluated with a signal-to-interference ratio as low as 35 dB. With signal-to-interference ratios lower than 35 dB, evaluation was no longer possible, because the 10% value of the wanted signal could not then be recognized.

4. Spectrum analysers may also be valuable in the recognition and classification of emissions, particularly complex emissions. Many types of emission have certain peculiar combinations of characteristics which, when viewed on a spectrum analyser by a trained observer, may provide valuable information leading to the recognition of the emission. Photographs of spectrum analyser displays of known emissions may be kept available to the monitoring observer for reference and for comparison with questionable emissions.

Spectrum analysers have also been used to advantage in connection with frequency measurements in the presence of interference, where matching of the frequency of the measuring equipment against the unknown frequency by aural methods is difficult. By observing the two signals on the spectrum analyser, the frequency of the measuring equipment may be adjusted to that of the unknown carrier or to other discrete components in the signal.

5. Although the estimations of the bandwidths, in terms of discrete components attenuated less than 6 dB and less than 26 dB below the peak level of the emission, may have no direct relationship with the "occupied bandwidth" as defined in No. 147 of the Radio Regulations, such estimates can be of considerable value in detecting emissions occupying excessive bandwidths, and which may cause interference. The use of a standard method by all monitoring stations in reporting estimates of bandwidth to the IFRB should also assist the IFRB in evaluating reports of different monitoring stations (see Report 324).

6. Bandwidth measurements of frequency-modulation transmitters

For the efficient operation of broadcasting networks, it is useful to have a written record (recording) of the programme levels obtained, either at the programme source, i.e. in the studios, or at several points in the connecting links (cables or radio-relay systems), or at reception, i.e. after the signal has been radiated by frequency- or amplitude-modulation transmitters.

Excellent results have been obtained in Italy, with the use of fewer staff, by employing peak programme level analysers installed with paper recorders at the output of the Rome sound broadcasting studios to monitor the outgoing audio-frequency signal levels, and at the Monza monitoring centre to measure the programme level of the transmitters.

The measurements are effected at the same time and, whenever anomalies are found, the recordings are compared so as to trace their causes.

In frequency-modulation transmitters, where the bandwidth of emissions depends on the programme signal level, and when these programme signals lie in the audio-frequency band, the programme level analysers enable the bandwidth to be monitored and recorded.

The peak-programme level analysers are electronic (level) devices which store the information on the maximum level attained for a long enough time (storage time) to enable this information to be correctly reproduced on the paper strip of the recorder, regardless of the duration of the modulation peak, providing diagrams that can be easily and quickly interpreted.

The recorder of normal quality inscribes an immediate recording on paper, since the recording is independent of the measurement process which is not affected either by the inertia of the writing mechanism or by the friction of the stylus on the paper.

The characteristics of the peak programme level analyser are as follows:

- number of levels: 6

  (Taking 0 dB = 100% modulation, the level thresholds are arranged on the instruments at the studio output or transmission line input for the values: -18, -8, -5.5, -3.1, -1.4 and -0.9 dB, while on the instruments used to measure the transmitter programme level, the level thresholds are arranged at -4.6, -3.6, -2.7, -1.8, -0.9, $+0.8$ dB.)

- rectifier characteristics: $n = 1$ (see Note 3 to Report 292);

- storage time: 30 s;

- rate of advance of paper strip: 60 mm/h.

7. Another approach to the measurement of bandwidth of frequency-modulated emissions described in [CCIR, 1966-69c], is summarized as follows:

A reference (0 dB) level, representing the peak value of the emission, is sometimes difficult to establish for the purpose of bandwidth measurements in accordance with Recommendation 443 in the case of frequency-modulated emissions because of amplitude-reduction of the carrier with modulation. With some emissions (e.g. microwave emissions), modulation is continuous, rarely permitting the full power of the transmitter to return to the carrier frequency so that a reference level may be established. With frequency-modulation, the total power output is constant but is distributed over the transmitted frequency spectrum in accordance with Bessel functions. A "substitution method" of bandwidth measurement has been devised which takes advantage of the fact that if the uniform portion of a receiver's bandwidth is adequate to include all significant components of an FM signal, there is no variation with modulation in the amplitude of the total voltage seen by the detector. This bandwidth estimation procedure is valid only on a stable signal where no fading of the emission exists at the point of measurement at least over the time interval of measurement.

A receiver's relative indicating voltmeter operating from an envelope detector may be used to establish the level that represents the total output level of the FM transmitter. That level is displayed on a spectrum analyser which is connected to the receiver IF output. Also required for this measurement method is a signal source of arbitrary but variable output voltage, such as a stable heterodyne frequency meter. Some method of introducing signal attenuation is necessary. Attenuators are usually an integral part of the spectrum analyser or receiver used.

Using this method, the desired signal is carefully centred in the receiver's passband, using the receiver's voltmeter as an indicating device. Other signals within the receiver passband will invalidate the calibration procedure. The observer must be certain that the receiver response is uniform over the bandwidth required to pass all significant frequency components of the modulated emission. The indicated, arbitrary level of the voltmeter is then noted. Disconnect the antenna from the receiver and then introduce the unmodulated signal source to the receiver, varying its output amplitude so as to match the same level found for the desired emission on the

receiver's voltmeter. The unmodulated signal source, at the same voltmeter level as the modulated emission will then be displayed on the spectrum analyser with peak level equivalent to the reference level (0 dB). The desired amount of attenuation (i.e. 6 or 26 dB) is then applied, preferably using internal attenuators. If external attenuators are used, connecting cable attenuation at the radio-frequency must be added in establishing the desired level. The maximum amplitude level of the unmodulated source is noted on the spectrum analyser while the attenuation is inserted. A stopped or as slow a sweep speed as possible will best resolve the peak level, particularly if there is oscillator instability or incidental frequency modulation. After the attenuated level has been established on the spectrum analyser, the attenuation is then removed and the antenna is reconnected to the receiver so the desired emission is displayed. The observer may then determine the frequency limits at the calibrated attenuation level to estimate the bandwidth.

8. Bandwidth measurements of reduced or suppressed carrier emissions

Document [CCIR, 1970-74a], outlines a relatively simple method, using existing equipment, for the bandwidth measurements of reduced or suppressed carrier emissions at a monitoring station. Based on the concept of measuring the bandwidth at the x dB level, in relation to the peak level of an emission as indicated by a spectrum analyser, it seeks to establish the reference or 0 dB level. The method is particularly applicable to measurements of R3E and J3E types of emission. The techniques involved are briefly explained together with the equipment requirements and the measurement procedure, with reference to a block diagram.

9. Method of measurement of bandwidth of amplitude-modulated telegraph classes of emission

In accordance with [CCIR, 1970-74b], the method of establishing the reference level (0 dB) described in § 7 may be extended to measurements of the x dB bandwidth of amplitude-modulated telegraph classes of emission and pulse-modulated emissions. In this case, an oscillograph is connected to the input of the spectrum analyser connected to the output of the receiver intermediate-frequency amplifier, thus enabling the amplitude of the substitute signal to be matched to that of the radio pulses in those sections of the pulses in which it corresponds to the unmodulated carrier. Calibration of the spectrum analyser on the substitute signal and spectrum analysis in the x dB bandwidth measurements must be carried out at the same scan rate. The sinusoidal substitute signal may be introduced either at the input of the receiver or at the input of the spectrum analyser to which an oscillograph is connected. The oscillograph may also be connected to the wideband intermediate-frequency channel of the spectrum analyser itself.

10. Additional information concerning measurement of bandwidth may be found in Recommendation 328.

11. Comparative distance measurements of bandwidths by various methods

Document [CCIR, 1974-1978c] contains information on the results of comparative distance measurements of occupied bandwidth and x dB bandwidths at the x dB levels shown above in Table 1 (i.e. representative bandwidth values), of transmitters using classes of emission A1B, F1B, F3C and F7BDX. The measurement method is briefly described in Report 324, section 3.

The measurement results confirm the validity of these methods for the distance measurements of transmitter emission bandwidths, i.e. for purposes of monitoring, at least in stable field conditions.

12. Effect of noise on the accuracy achieved in the measurement of emission bandwidth by various methods

12.1 Document [CCIR, 1974-78b] contains the results of studies [Pavliouk, 1972] on the effect of noise on the bandwidth measurement accuracy achieved with the following methods:

12.1.1 Direct method of occupied bandwidth measurement by power ratio (Recommendation 328, § 4.2);

12.1.2 Direct method of measuring the x dB bandwidth (Recommendation 328, § 4.1);

12.1.3 Indirect method of occupied bandwidth measurement based on the out-of-band spectrum decay rate (Recommendation 328, § 4.3.1).

The noise was assumed to be evenly distributed over the receiver passband. In the study of the first method, account was also taken of the noise concentrated in the occupied bandwidth and in the measuring filter (i.e. in the out-of-band spectrum).

The spectrum envelope of the measured emission was approximated by power and exponential functions as it is shown in Report 324.

The calculation results are shown in Figs. 1 to 3 as the relations of the relative measurement errors δij (i: index of measurement method, j: index of spectrum envelope approximation) to the ratio Z of signal spectral power density at the limit of the measured bandwidth to the spectral power density of noise evenly distributed over that bandwidth. The out-of-band spectrum decay rate is shown alongside the curves in dB/octave (for the second type of approximation, for the first octave). The parameter ε is equal to the ratio of the measurement filter passband (first method) to the measured bandwidth and the parameter h is equal to the ratio of the spectrum analyser filter passband to the measured bandwidth. $\delta_3(k)$ characterizes the error obtained with the third measurement method when the spectrum slope is assessed on the out-of-band part, while $\delta_3(p)$ refers to the in-band part at the limit of the measured bandwidth. The experimental data are indicated by the dashed lines.

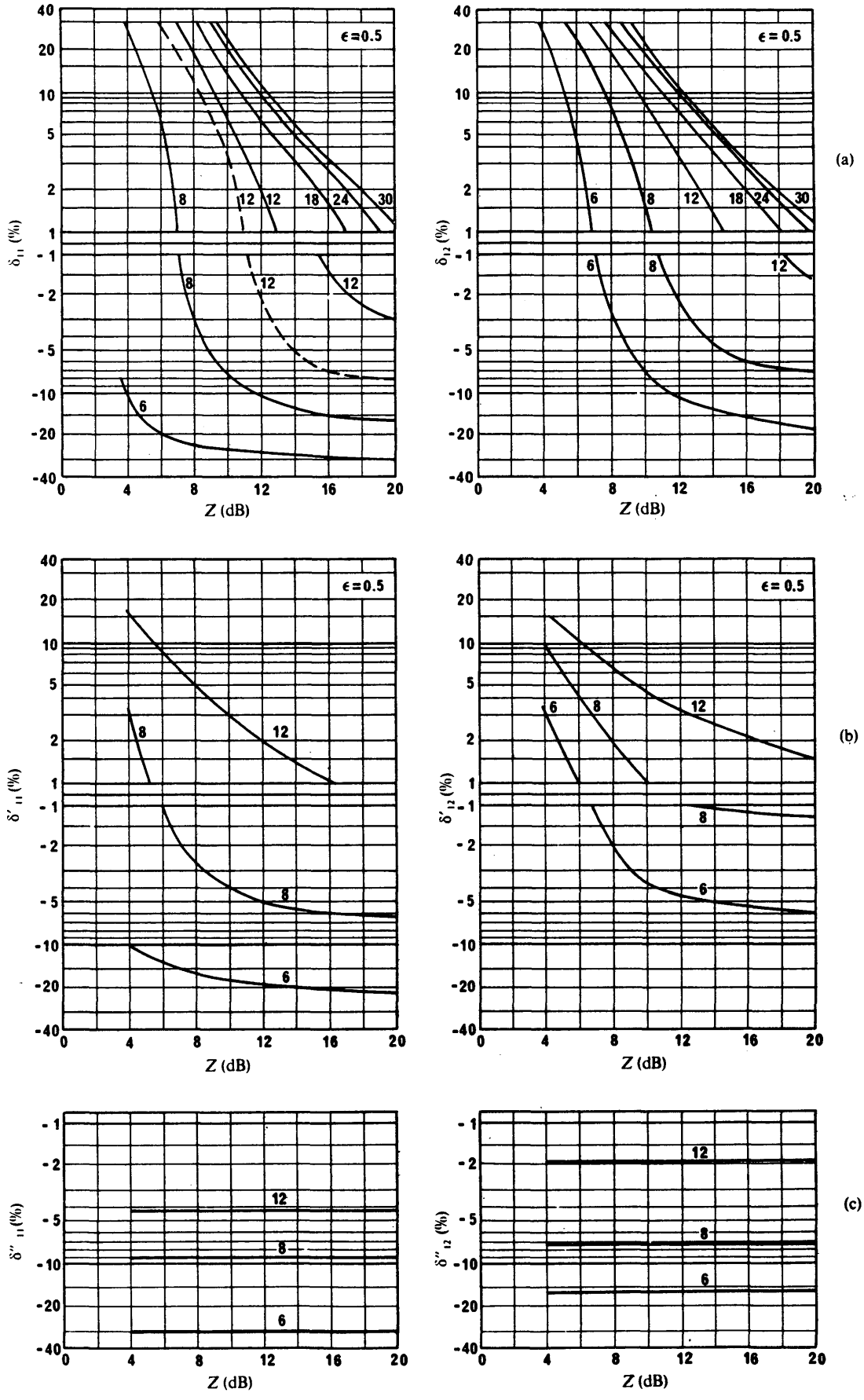


FIGURE 1 — *Relationship between the percentage error (δ) in measuring the occupied bandwidth by the direct method and the signal-to-noise ratio (Z), for different values of N*

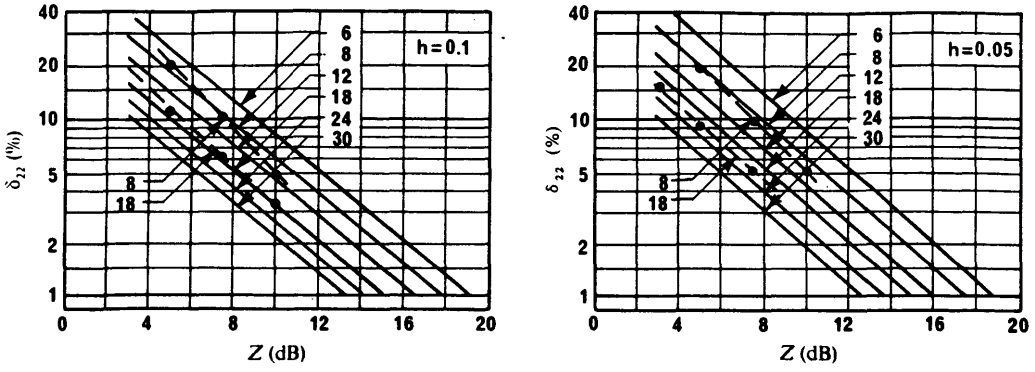


FIGURE 2 — *Relationship between the percentage error (δ) in measuring the x dB bandwidth and the signal-to-noise ratio (Z), for different values of N*

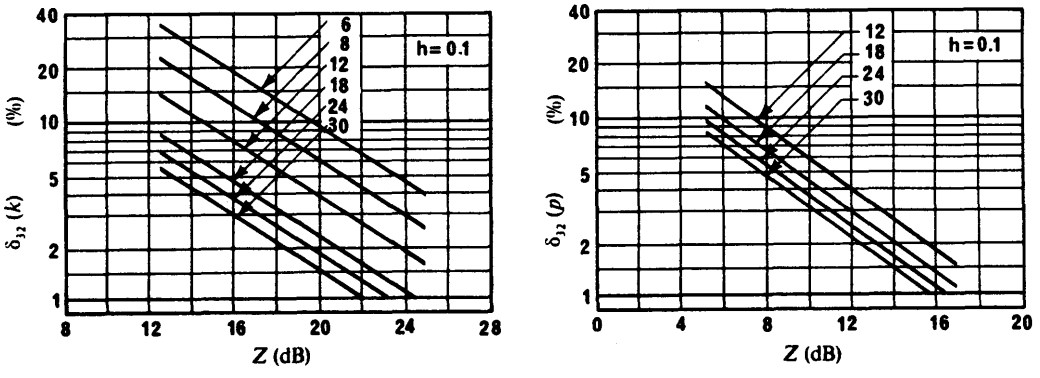


FIGURE 3 — *Relationship between the percentage error (δ) in measuring the occupied bandwidth by the non-direct method and the signal-to-noise (Z), for different values of N*

12.2 The following conclusions may be drawn from these data:

12.2.1 In the presence of noise, greater accuracy is achieved with the second method than with the first. The measurement error is particularly high when the noise occurs within the passband of the measuring filter and concentrates in that part of the out-of-band spectrum having the lower level.

12.2.2 In the presence of noise, the accuracy obtained with the third measurement method depends on which part of the spectrum (in-band or out-of-band) is used to assess the bandwidth in which the spectral power density is reduced by 10 dB. If this assessment is based on the in-band part of the spectrum, virtually the same measurement accuracy is achieved with the third method as with the second method.

12.2.3 The errors calculated from the various types of spectrum envelope approximation show only minor differences.

12.2.4 With the first method, the noise-induced error can be offset by an appropriately selected measuring filter passband. In practice, however, this compensation is complicated by the extent to which the sign and magnitude of the error depend on the spectrum slope and the relative noise power.

12.2.5 The analysis filter passband has only a slight effect on the measurement accuracy achieved with the second and third methods. A halving of the filter passband increases the accuracy by only 2-3%.

13. Conclusions

In order to assess the accuracy of measuring bandwidth which can be achieved at monitoring stations, studies are needed to compare such measurements with those made in conformity with the definition of "occupied bandwidth" in No. 147 of the Radio Regulations. Comparison should include measurements on various classes of emission under good reception conditions and in the presence of interference. In addition, comparative results using any new method of measuring bandwidth will be helpful to Study Group 1 in evaluating such methods.

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REPORT 272-5 *

FREQUENCY MEASUREMENTS AT MONITORING STATIONS

(Question 22/1)

(1956-1959-1963-1966-1970-1974-1982-1986)

Question 22/1 seeks information on the desirable ratio of the error of frequency measurement to the permissible tolerances of emission and the accuracy required for frequencies extending up to, in certain cases, at least 10.5 GHz.

Information on equipment, accuracy of measurement and methods, is contained in Docs. [CCIR, 1962a, b, c, d, e; 1963-66; 1966-69a, b, c, d, e, f, g; 1970-74a, b and 1982-86]. The information available at present may be summarized as follows:

1. General

1.1 When an individual frequency measurement of a radio station is made, its accuracy can be estimated from a knowledge of all of the factors (except perhaps that of Doppler effect), which produced the error. The sources of error which should be known to the measurer are: the maximum error in the frequency standard of reference; the maximum instability error of the transfer oscillator and the error of setting that oscillator to zero-beat with the signal. Inasmuch as the direction of errors, other than that of the reference standard, will not usually be known, the errors of measurement should be provisionally estimated, until an answer to Question 22/1 is obtained, as if they were all in the same direction after allowing for any known error of the reference standard.

1.2 It is considered that a conservative ratio of error to tolerance as applied to the majority of frequency measurements is to be preferred over a ratio difficult to obtain. With a conservative ratio, the measuring equipment may be operated well within its capabilities with a greater degree of assurance that all measurements meet the desired minimum accuracy.

1.3 The ratio of 1 to 10 between the error of measurement and the tolerance appears to be a reasonable criterion to be retained at least until such time as conclusions are reached in the work of Question 22/1. In this study, it may be desirable to adopt a statistical method of arriving at the accuracy of measurement of an unknown frequency (see Doc. [CCIR, 1962c]).

2. Accuracy of measurement

With regard to § 1 of Question 22/1, on the accuracy of frequency measurements which can be accomplished at monitoring stations, the following statements may be made:

2.1 The accuracy of the secondary standard *P* (see Table I) of a frequency measurement installation is determined by calibration. A stability of $\pm$ 1 part in 10^9 or better can be achieved with currently available standards, including those designed for mobile use. The accuracy may not exceed $\pm$ 1 part in 10^7 when using standard-frequency transmissions received over a sky-wave path. However, by selecting a time of day for the comparison during which sky-wave propagation conditions are most stable (usually around noon of the path mid-point), and by continuing the comparison over a period of several minutes, accuracy of a few parts in 10^8 is readily achievable. Where ground-wave HF or usable VLF/LF emissions from standard-frequency stations are available, the secondary standard accuracy may be maintained to better than $\pm$ 1 part in 10^8.

Where greater accuracy is required, either for fixed or mobile installations, use may be made of an atomic frequency standard. Atomic standards weighing less than 20 kg, requiring 50 W or less of operating power, and having an accuracy of $\pm$ 1 part in 10^9 or better, are available.

2.2 The methods of measurement usually employed result in the determination of the instantaneous frequency as received. Therefore, when measuring distant stations, due regard must be given to frequency-shifts caused by varying propagation conditions (e.g. Doppler effect).

Doc. [CCIR, 1966-1969c], gives, as a result of measurements, some values of the influence of the transmission path on frequency measurements. In the HF band, this document reports that these frequency deviations may be as much as $\pm$ 3 parts in 10^7 at night. If both the transmitter and receiver are in the daytime region, frequency variations in most cases are less than $\pm$ 3 parts in 10^8. With VLF and LF, transmitter frequency variations of the order of about 1 part in 10^9 have been noted.

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* This Report is the result of merging Reports 272 and 365. Report 365 is hereby cancelled.

2.3 Doc. [CCIR, 1966-1969f], proposes that uniform rules should be established for determining the measurement error and for its mode of presentation, especially in the "List of International Monitoring Stations". Taking also into account Doc. [CCIR, 1966-69c] the proposal is made that, for each class of emission the value of the error in frequency measurements at monitoring stations be presented in the following form:

$$\text{Error} = P \times f + x \qquad \text{Hz} \tag{1}$$

where:

P: the accuracy (in parts per million) of the frequency standard used for the measurements, which is obtained by calibration of the standard;

f: the frequency measured (Hz);

$x = y + z$;

y: the error (in Hz) dependent on the harmonic order used in the frequency measuring apparatus, excluding the error of the standard;

z: the error (in Hz) introduced by the receiving part of the equipment (e.g. phase variations or transient phenomena), which is independent of the harmonic number used in the measuring equipment.

Additionally, it is proposed that the error attributable to the method be taken as the maximum value (expressed as an absolute value) of the Gaussian error distribution for which, with a confidence level of $1 - \alpha = 95\%$, there would be a 1% probability of errors (in absolute value) in excess of the value thus reckoned; and that the method described in Report 422 (Geneva, 1982) should be adopted for determining the values of x.

2.4 For emissions with a discrete component at the assigned frequency, which can be identified, the accuracy is practically the same as for unmodulated emissions. For emissions having discrete components which can be identified at other than the assigned frequency, the accuracy is further limited by the accuracy with which the displacement is known.

2.5 Under conditions of fading, and where the measurement is made using a pattern on the screen of an oscilloscope, the measurement is performed only when there is sufficient signal strength, for example, A1A and A1B stations are measured only during on-keyed conditions and not during a field-strength minimum period. Under these conditions the accuracies listed in Table I can be obtained.

TABLE I - *Accuracy of frequency measurement in Hz attainable under optimum reception conditions at monitoring stations with current measuring methods*

Class of emission	Measuring method		
	Frequency synthesizer	Transfer oscillator and electronic counter	
		Count period: 10 s	Count period: 1 s
N0N, A2A, A2B, A3E, B8E	$P \times f + 0.4$	$P \times f + 0.5$	$P \times f + 1.7$
A1A, A1B	$P \times f + 0.7$	$P \times f + 0.6$	$P \times f + 1.8$
F1B, F7BDX (not greater than about 75 bauds)	$P \times f + 2.5$	$P \times f + 2.8$	$P \times f + 2.8$

Note 1. - Assuming a synthesizer resolution of 0.1 Hz.

Note 2. - *P:* accuracy of the frequency standard used. In the case of an atomic frequency standard, $P = 1$ part in 10^{10} or better; when using standard frequency transmissions in the HF bands received over a sky-wave path, the accuracy may not exceed $P = 1$ part in 10^7.

Note 3. - *f:* frequency in Hz.

Note 4. - In the frequency measuring equipment, the number of harmonics should not exceed a value of 3 or 4.

Note 5. - The above values are the errors not exceeded in more than 99% of the cases (see Report 422 (Geneva, 1982)).

Note 6. - The method described in § 2 of Report 273 provides an accuracy of frequency measurements of stable A3E signals of $P \times f + 0.01$ (Hz).

2.6 Under conditions of interference between two amplitude-modulated stations, the frequency of the weaker station can still be measured with the accuracy indicated in Table I, if the carriers of the wanted and unwanted stations are spaced by at least 40 Hz. With a field-strength ratio between the wanted and the unwanted stations of 1 to 1, good measurements may be made with the accuracy indicated in Table I, when the carriers are spaced by only 12 Hz. For a field-strength ratio of 2 to 1, the minimum frequency spacing at which good results are otained is 2 Hz. These high accuracies of measurement are made possible, even under conditions of interference by other stations, by visual observation of the Lissajous patterns produced on the oscilloscope screen by combining the heterodyne beat between the received signal and the comparison frequency and a fixed standard-frequency of 1000 Hz. By receiving mutually interfering stations on a direction-finding equipment, their ratio of signal strengths at the receiver input can be improved, thus making possible measurements with the accuracies shown in Table I at smaller carrier spacings.

The method and instrumentation described in § 2 of Report 273 provides the accuracy given in Note 6 of Table I for stable A3E signals with at least 1 Hz frequency spacing and level differences up to 60 dB.

2.7 Two interfering frequency-modulated stations can be measured for unfavourable field-strength ratios when the frequency separation between their carriers is at least 2 to 3 kHz. The measurements are taken during periods of no modulation by heterodyning the wanted carrier with the comparison frequency and observing the results on a panoramic display screen. The accuracy will then be within ± 500 Hz. Here too, by employing a direction-finding technique, the received signal strength of the unwanted station may be sufficiently attenuated to obtain the accuracies indicated in Table I.

2.8 Measurements of carrier instability can be made, either as a series of measurements of instantaneous frequencies employing the oscilloscope method or, with rapid variations of the carrier frequency, the centre frequency can be determined by means of a discriminator equipment, or a panoramic adaptor, resulting in an accuracy of measurement of the extreme frequencies of about ± 50 Hz.

2.9 Keyed carriers can be measured by the oscilloscope method with the accuracies listed in Table I.

With F7BDX classes of emission, each new wave train of the shifted frequencies begins with an undetermined phase position. A frequency measurement can therefore be made only after the change in phase position during one oscillation train. Thus the possible accuracy is limited, especially in the case of frequency shift keying with high speed (e.g. about 100 bauds or higher).

2.10 For suppressed carriers, individual frequencies of the types listed under § 2.5, 2.6, 2.7, 2.8 and 2.9, which are contained in the emission spectrum, can be measured with the accuracies listed in those paragraphs. For a carrier not completely suppressed, the measurement can be made with the accuracy indicated in § 2.6.

2.11 Determination of the centre frequency of complex emissions, where the spectrum of the emission is unknown, presents a difficult problem, particularly when a signal is encountered which has no discrete carrier and where all the tones of the system (in the case of a multitone system) are not being transmitted. From the point of view of the monitoring stations, the measurement of independent-sideband (ISB) and single-sideband (SSB) emissions would be facilitated if a carrier were transmitted, but this is often not practicable when considering the requirements of many present-day systems.

In many ISB emissions a prior knowledge of the system can give guidance in locating the carrier frequency, so facilitating assessment of the centre frequency. This emphasizes the need for all transmissions to be clearly described when notifying assignments. Particularly difficult cases are ISB emissions employing two-tone or frequency-shift sub-carriers, since they may be indistinguishable from pure F1B emissions. To facilitate the measurement of the more complex classes of emission, some operating organizations have adopted designation systems which identify separately the sidebands and sub-carriers. Similarly, a system of designation may be used to indicate the position of the centre frequency with respect to the assigned frequency.

2.12 In those instances where a characteristic frequency (defined in No. 143 of the Radio Regulations) has been established, e.g. a reduced or incompletely suppressed carrier or a pilot frequency (see § 2.2.2 of Recommendation 379), it is often convenient to measure the characteristic frequency by adjusting the frequency of a stable variable radio-frequency oscillator to coincide as exactly as possible with the characteristic frequency as observed on a spectrum analyzer of high resolving power. The precise frequency of the radio-frequency oscillator may then be determined by the usual methods and the deviation from the assigned frequency calculated. Using this procedure for measurements in band 7 (3 to 30 MHz), it is desirable that the spectrum analyzer have a maximum resolving power (see Recommendation 182, § 2), at reduced sweeps, of 10 Hz and that the short-time (10 s) stability of the variable radio-frequency oscillator be 5 parts in 10^7, or better.

2.13 Doc. [CCIR, 1963-66], describes a method of determining the approximate frequency of the suppressed carrier of J3E emissions, by measuring the frequency of the receiver demodulating oscillator when the oscillator is adjusted for optimum speech articulation and understandability. According to tests conducted under varying conditions of signal level and noise, the accuracy of this method in the HF band in many cases may be within 30 Hz.

Doc. [CCIR, 1970-74a], describes a method whereby under normal operational conditions, the frequency of the suppressed carrier of a J3E emission can be determined to an accuracy of the order of ± 4 Hz. The technique used is based on the fact that the spectrum of a pure vowel sound produced by the human voice contains only the pertinent fundamental frequency and its harmonics. In this method, the receiver oscillator frequency is first adjusted to within 30 Hz of the frequency of the suppressed carrier, by means which have already been described, after which further adjustment to within 4 Hz can be achieved by observing on an oscilloscope a Lissajous figure formed during reception of the vowel content of an emission.

2.14 In multi-tone transmissions, where the characteristics of the emission are known, determination of the relationships between the various components of the emission with a spectrum analyzer and a variable frequency oscillator to serve as a reference frequency, may permit a reasonably accurate determination of the centre frequency. Another similar method is to photograph the spectrum display, together with known frequency markers.

3. Equipment and methods

3.1 With regard to measurements above 50 MHz, the frequency measuring equipment of the individual countries differs so much that it is only possible to comment generally upon them. Administrations usually prefer to make measurements wih nearly the same equipment as that which they use for frequency measurements below 50 MHz. The extension to a higher frequency range is effected by the use of harmonics of either a secondary standard or a stable oscillator or by the use of a frequency synthesis method providing a single tunable frequency. All equipment uses direct reading dials in conjunction either with electronic counters or with visual indicators such as pointer instruments or oscilloscopes.

3.2 With regard to preferred equipment and methods for the measurement of frequency in frequency-modulated emissions, the carrier frequency may be measured by any of the usual methods during intervals of no modulation. With modulation, it is common practice to measure the average frequency by means of a frequency counter. In many types of programme material, the peak excursions of a frequency-modulated carrier in the positive and negative directions may be considerably different while, at the same time, the average frequency will remain constant. This non-symmetrical programme material may result, for example, from certain combinations of a fundamental tone and its various harmonics. Then the average over a period of several cycles at the lowest modulating frequency will approximate the unmodulated carrier frequency. Therefore, the averaging technique is valid under all normal conditions, provided the period of the frequency count is long compared with one cycle at the lowest modulating frequency. Instabilities in the transmitter may, of course, result in a shift in the average frequeny of the carrier, so that the frequency indicated by the counter would differ somewhat from the unmodulated carrier frequency.

3.3 *MF/HF frequency measurement methods*

A number of different techniques are used to evaluate the frequency of signals received in these bands. Since these signals are often intermittent in nature, direct counting is not suitable and a comparison with a locally generated signal is required. A common feature is the comparison of a locally generated signal of high stability with the receiver IF signal in an oscilloscope to obtain a 1 : 1 Lissajous pattern. One or more of the conversion oscillator frequencies is usually synthesized from a high stability source so the frequency reading often can be obtained simply by measuring the frequency of a single VFO.

If the receiver is completely synthesized to sufficient resolution (1 Hz or less), the frequency measurement process is simplified to one of providing a stable frequency at the IF value and comparing this with the IF in a Lissajous pattern on an oscilloscope. A variety of beat indicators may be used for comparison of continuous signals but the oscilloscope is preferred for use on intermittent signals.

3.4 The importance and potential usefulness of statistical analyses of frequency measurement data are described in Doc. [CCIR, 1962c] This document also suggests a method of arriving at the accuracy of a measurement of an unknown frequency by obtaining the standard deviation of a number of measurements made of a known frequency.

3.5 In Doc. [CCIR, 1962d] the overall accuracy of frequency measurements, attainable in the range 1 to 10 GHz, was indicated using a typical spectrum analyser with the following characteristics: varies from 8 kHz (for a 250 kHz sweep) to 100 kHz (for a 70 MHz sweep), equivalent to ± 8 parts in 10^6 and ± 1 part in 10^4 respectively at a frequency of 1 GHz. Thus, an accuracy of ± 1 part in 10^5 would appear possible, in practice, for the range 1-10 GHz for the condition where the emission contains a known discrete component whose position relative to the unmodulated carrier frequency is accurately known. Where it is possible to obtain a beat frequency output, the accuracy of measurement may be increased to ± 1 part in 10^6 or even better.

3.6 According to Docs. [CCIR, 1966-69a and b] the introduction of precision offset for television transmitters and the narrowing of frequency tolerance to promote better use of the frequency spectrum has made it necessary to acquire frequency measuring equipment providing accuracy of the order of 1 part in 10^{10}. According to Doc. [CCIR, 1966-69b], it is possible to make measurements with this accuracy using synthesizers that can be switched in steps of 10 Hz or 1 Hz and with an adjustment accuracy of the interpolation scale that can approach values of ± 0.03 Hz or 0.003 Hz, depending on the final decade used. An accuracy of about 1 part in 10^{11} in the frequency standard is assumed.

It is further stated that checks of the division of the interpolation scale using the method described in Doc. [CCIR, 1966-69b] have shown that after several years of operation, no deviation in excess of the error limit 0.03 or 0.003 Hz had been experienced.

3.7 Doc. [CCIR, 1970-74b] describes an improved method of frequency measurement of frequency-shift emissions by phase recording. If the phase positions of the individual signals on the same frequency do not depend upon one another, a greater accuracy is obtainable than with the Lissajous-figure method.

3.8 *Remotely-controlled frequency measurement method*

HF receiving stations at present tend to be located in remote places since they require a large area for antenna farms and quiet receiving conditions free from man-made noise. Accordingly, HF receiving station and frequency monitoring facilities may be controlled from remote stations.

A typical configuration of frequency measuring equipment is shown in [CCIR, 1982-86] and consists of the following:

– antenna switch – ISB receiver – adapter to the receiver – remote control apparatus	in the unattended receiving station
– measuring console – remote control apparatus	in the control station

The ISB receiver is remotely controllable, designed for fixed communication circuits, and with a receiving frequency range from 4 to 28 MHz. The adapter to the receiver is provided in order to use the ISB receiver without any modification.

The remote control apparatus not only controls and monitors the antenna switch and receiver, but also sends data from the adapter. The remote control equipment units are connected to each other through voice grade channels of a microwave link.

The accuracy of the measurement is within ± 1 Hz. Calibration was performed by receiving the signals transmitted from a standard frequency station at a distance of 60 km from the receiving station.

The principles of measurement and frequency measuring procedures are given in [CCIR, 1982-86].

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CCIR Documents

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[1963-66]: VIII/78 (Japan).

[1966-69]: **a.** VIII/4 (Germany (Federal Republic of)); **b.** VIII/53 (Germany (Federal Republic of)); **c.** VIII/54 (Germany (Federal Republic of)); **d.** VIII/6 (Japan); **e.** VII/44 (Belgium); **f.** VIII/45 (Portugal); **g.** VIII/46 (Portugal).

[1970-74]: **a.** 1/108 (Japan); **b.** 1/12 (Germany (Federal Republic of)).

[1982-86]: 1/9 (Japan).

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REPORT 978-1

AUTOMATIC IDENTIFICATION OF RADIO STATIONS

(Question 34/1)

(1986-1990)

1. Introduction

The Radio Regulations require that all transmissions shall be capable of being identified either by identification signals or by other means.

2. An identification method for HF services

2.1 The variations in the types of automatic radioteletype systems, methods of keying and associated radio-frequency transmitting equipment are so numerous and complex that systematic methods are necessary to permit proper identification of the basic systems employed.

2.2 The identification of a radioteletype signal by conventional means is time-consuming and requires elaborate equipment. In the Federal Republic of Germany, an FSK analyzer operating on the basis of modern micro-processor technology is available. This type of analyzer enables the identification of FSK signals to be made more easily and rapidly. Supported by a microprocessor, the FSK analyzer automatically analyzes all the parameters of teletype signals of F1B and F7B emissions with high accuracy and simultaneously displays the measured results together with all the essential information. Radioteletype signal identification is thus greatly facilitated.

2.3 The analysis result is indicated on the alphanumeric display of the front panel. If there is a decode program, tests appear in real time.

2.4 For connection to the IF or AF output of a conventional VLF/HF receiver, the FSK analyzer is fitted with a demodulator which is automatically adjusted for F1B or F7B signals. Connectors for external demodulators (VF telegraph transmission systems) and for operation from the line current of a teletype line are also provided. In these cases, measurement of the frequency shift is not possible and the associated displays remain unlit.

2.5 The analyzer performs the following four main tasks:

- measurement of shift,
- measurement of baud rate,
- continuous analysis of transmission code,
- text output for standard-code signals.

3. An identification method for VHF/UHF services

3.1 In the land mobile service in the VHF/UHF bands identification is mainly by voice signals. This form of identification is becoming increasingly more difficult due to the complex nature of the emissions, digital techniques and the use of encryption systems.

3.2 Current technology has made it possible to automatically identify stations using advanced analogue and digital techniques. In Japan a system described below is being adopted.

3.3 Modulation system for automatic identification

3.3.1 In analogue communications systems, the sub-carrier MSK modulation is used for modulation of automatic identification codes. In addition, the binary FSK modulation that can be economically configured will be used as a digital direct modulation system that is suitable for high-speed transmission, in consideration of the future development of the communications technology.

3.3.2 In digital communication systems in which digital data signals or digitally-converted voice signals are transmitted, GMSK modulation, 4-level FM modulation, PLL-QPSK and binary FSK are widely used for radio facilities. Therefore, it is possible that the same type of modulation as that used for the digital data will be used for automatic identification.

4. Identification signal format

It is also desirable that the identification code format permits reliable demodulation and inexpensive system configuration and that it is as uniform as possible for various modulation systems. Considering these requirements, the following signal format will be adopted:

Bit Synchro-nization	Frame Synchro-nization	Data	Error Correction

In bit synchronization, a bit length is assigned for each modulation system, taking into account the synchronization lead-in time of the receiver. In frame synchronization, a code type and a code length are assigned so that faulty operation due to noise and voice is reduced.

Data length including identification codes is assigned depending upon the given identifiable number of stations and various types of information necessary for identification.

Error correction codes will be inserted for improving the error rate in the radio transmission of identification codes.

5. Transmission rate for identification code

The transmission rate for identification codes is determined for each modulation system and communication bandwidth, taking into account the effective utilization of radio spectrum, communication systems and transmission reliability.

REPORT 276-6*

MONITORING OF RADIO EMISSIONS FROM SPACECRAFT AT FIXED MONITORING STATIONS

(Question 27/1)

(1963-1966-1970-1974-1982-1986-1990)

Question 27/1 seeks information concerning methods and equipment desirable for measurement at fixed monitoring stations of emissions from spacecraft and practical means of identification by monitoring stations of emissions from specific spacecraft.

Information concerning these matters is contained in Docs. [CCIR, 1962a and b; 1966-69a and b; 1970-74; 1978-82 and 1982-86a, b and c]. The information in these documents may be summarized as follows:

1. Tasks and measurements to be performed

In principle, the tasks entrusted to a radio monitoring station for space services do not differ from those of a terrestrial radio monitoring station. However, they are performed under dissimilar conditions which result from the fact that a space station revolves as a satellite round the Earth and a basic knowledge of its orbit is a prerequisite for any observations and measurements. The functions of the radio monitoring station for space services can be outlined as follows:

- regular and systematical observation of the radio-frequency spectrum with the aim of detecting and identifying space station emissions;
- determination of occupancy and percentage use of transponders;
- measurement and recording of the characteristics of space station emissions;
- investigation and elimination of harmful interference caused by space stations, if appropriate, in cooperation with terrestrial and other space monitoring stations;
- investigation and elimination of harmful interference caused by up-link transmitters by measuring the signal as re-transmitted by the transponder. This measurement will be performed in the same way and under the same conditions as those described for space stations;
- performance of measurements and recordings in the context of technical and scientific projects;
- detection of illicit use of transponders.

* This Report should be brought to the attention of Study Groups 2 and 4.

2. Techniques of measurement

2.1 *General*

The main factors influencing the necessity for different techniques of monitoring observation and measurement of emissions from spacecraft as contrasted with observations and measurements of emission originating from fixed or mobile radio stations on or near the earth are, with reference to spacecraft:

- the difference between received and transmitted frequency, and the varying nature of the received frequency, caused by the Doppler effect;
- the generally weaker field strength at the earth receiving point, due to distance and low transmitter power;
- relatively short time that a signal from a near-earth orbiting satellite is receivable at a fixed monitoring point;
- the continual pointing that is required of the highly directional earth-station antennas used to receive emissions from spacecraft.

Administrations launching spacecraft will, of necessity, continue to provide for the required accurate tracking and telemetry reception to meet their special needs. However, monitoring stations responsible for the enforcement of domestic laws and regulations and engaging in international monitoring, pursuant to Article 20 of the Radio Regulations, are now participating in space monitoring as a natural and necessary extension of their regular monitoring facilities, techniques and operations.

2.2 *Frequency measurements*

When there is a relative velocity between the spacecraft and the monitoring station, a difference of frequency proportional to the relative velocity arises between the transmitted and received signals owing to Doppler-shift effect. The apparent frequency of the transmitter in the case of an active satellite, as measured by a fixed monitoring station, is higher than the source frequency when the satellite is approaching the monitoring station and lower when it is receding. A measurement gives the source frequency only at the instant when the relative velocity between the source and the monitoring station is zero.

The accuracy of determining a frequency emitted from a satellite depends on the type of orbit of the satellite, the propagation path, the measuring equipment and the method of evaluation.

To determine the source frequency and the extent of Doppler shift, the basic requirement is a rapid frequency-measuring technique and a means of accurately timing each measurement. With such measurements it should be possible to determine with the necessary accuracy the time of closest approach of the spacecraft and the maximum frequency-shift (see Fig. 1). If frequency measurements are taken every few seconds and plotted against time, a curve is obtained from which the frequency at the point where the rate of change is greatest can be estimated with reasonably good accuracy, possibly to within $\pm$ 1 part in 10^6.

To increase the accuracy of measurements such as these, which are made in series, automatic measuring methods are to be preferred. For frequency measurements, automatic auxiliary equipment should be designed to record the frequency characteristic in analogue or digital form. It should be possible to print and also read the digital values measured. Using this automatic method of measurement, a better accuracy (of the order of $\pm$ 1 part in 10^7) can be achieved than that given in the paragraph above.

Document [CCIR, 1970-74], describes a modified method intended to obtain a better accuracy. By a single differentiation of the Doppler frequency curve with respect to time, a parabola is obtained, the vertex of which indicates the time of closest approach (TCA) and thus also the transmitter source frequency of the satellite. For the construction of the parabola it is sufficient to utilize the individual measured values in the range of $\pm$ 20 s about the TCA. The time interval between the measured values must be chosen so that the shape of the curve is clearly defined, for example, in intervals of 5 s. With a time resolution of 10 mm/s, an accuracy is obtained of $\pm 7 \times 10^{-9}$. In this case, it is assumed that a satellite with an approximately circular orbit is to be measured. Fig. 2 shows the results of frequency determination performed in this manner.

Document [CCIR, 1962b], describes a frequency measuring method which uses a discriminator, the object of which is to represent on a cathode-ray oscilloscope the wave form obtained by mixing the receiver output with a low-frequency saw-tooth wave (e.g. 25 or 50 Hz) from a separate source. The discriminator is used to determine the transmitted frequency by measuring the constantly changing frequency of the Doppler shift. When the figure

remains stationary on the oscilloscope, the receiver output frequency is an exact multiple of the standard low-frequency. From the value of this exact multiple and the local oscillator frequency in the receiver, the exact frequency can be determined. By measuring the frequency each time the figure on the oscilloscope becomes stationary and plotting this against time, the curve of the Doppler shift may be drawn.

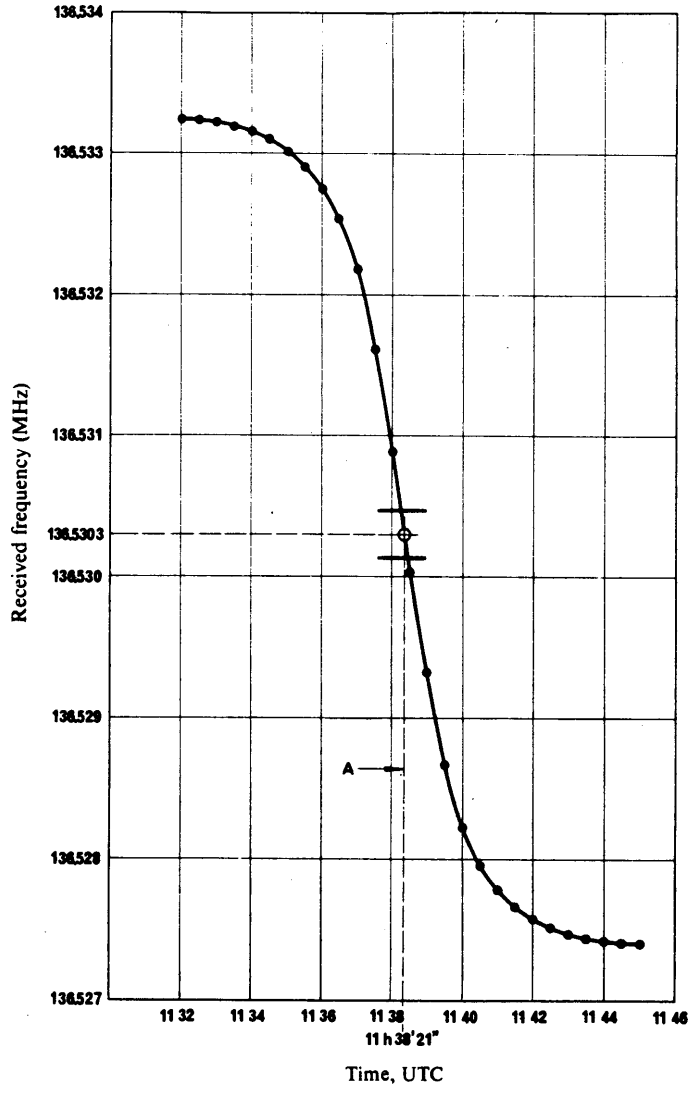


FIGURE 1 - *Variation of the received frequency as a function of time due to the Doppler effect, and time of closest approach*

A: Time of closest approach

2.3 *Bandwidth measurements*

The apparent transmitted bandwidth of a near-earth satellite, as measured at a fixed monitoring station, varies because of the Doppler-shift effect in the same manner as described for the carrier frequency. Although not of practical significance at present from an interference standpoint, it should be realized that the frequency-shift is slightly greater for signal components near the upper edge of the spectrum of the emission than for those near the lower edge. This difference could amount to hundreds of hertz for wide bandwidths. This effect causes the apparent bandwidth, as observed at the fixed monitoring point, to vary slightly.

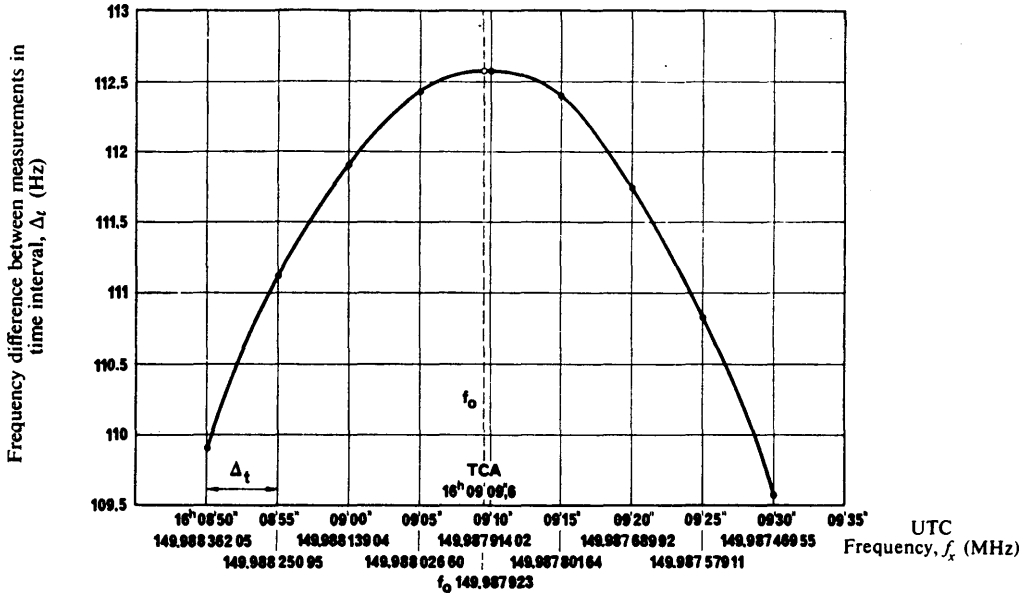


FIGURE 2 - *Differentiation of the Doppler frequency curve with respect to time*

Δ_t: time difference between measured values f_0: evaluated transmitter source frequency
TCA : time of closest approach

If the emitted field strength from the spacecraft is sufficient, the Doppler frequency-shift of the emission can be compensated by automatic frequency control of the monitoring receiver. In this case, the normal measuring methods used to determine the bandwidth in monitoring stations can be applied without radical changes. If the received signal is very weak, it is possible to ensure automatic correction of the receiver oscillator's frequency by a reference signal of the spacecraft emission (carrier or pilot) filtered in an extremely narrow band. If the measuring station does not possess appropriate receivers with automatic frequency control, account must be taken of the frequency shift of the spacecraft during the measurement, if necessary by making a simultaneous Doppler frequency measurement when determining the bandwidth. It will in any case be necessary to make a simultaneous recording of the field strength so that the effect of field-strength variations occurring during the analysis of the spectrum may be eliminated during the calculations.

The filters for the bandwidth analyses should have a fairly narrow band, particularly for the analysis of wide transmitter spectra, so that good separation of the spectrum components, and a good sensitivity limit, may be obtained. The duration of the analysis should be chosen between 30 s and 10 min so that transient phenomena may be taken into account.

2.4 *Measurements of power flux-density*

For determination of power flux-density (pfd) at the surface of the Earth, of emissions from spacecraft, two methods are used:

- by measurement, using specialized equipment;
- by calculation, based upon measurement of field strength converted to power flux-density as follows:

$$F = -25.77 + 20 \log E \qquad (1)$$

where

F: total power flux-density (dB(W/m^2)),

E: field strength (V/m).

Measurement of power flux-densities (pfd) requires a knowledge of the figure of merit (G/T) of the receiving system at the frequency of interest. Direct measurement of G/T is preferred to taking the ratio of separately measured G and T, because the opportunity for error is reduced. Separate measurements of G and T also requires the use of a signal generator, which unnecessarily introduces an additional factor of uncertainty. The Sun, rather than a radio star, is often used for G/T measurements for pfd calibration purposes at monitoring stations because of its much stronger signal. Should, however, the receiving system be sufficiently sensitive, use of radio stars would be preferable.

2.4.1 *Calculation of G/T*

The figure of merit (G/T) is usually determined at 5° elevation and expressed in dB(K^{-1}) units i.e., G/T dB(K^{-1}) = 10 log (G/T numeric).

G/T(dB(K^{-1})) = antenna gain (dBi) − 10 log (system noise temperature (K))

$$G/T(\text{numeric}) = \frac{8\pi k r_1 r_2 f^2 (y_{sun} - 1)}{sc^2 y_x} \qquad (2)$$

where:

k: Boltzmann's constant (1.38×10^{-23} J/K),

r_1: correction factor for atmospheric attenuation; for angles ⩾ 5°:

$$r_1 = \text{antilog}\ \frac{R_1\ (\text{dB})}{10}$$

where:

$$R_1 = \frac{A}{\sin \theta}$$

and A is the one-way atmospheric absorption in decibels for a vertical path and θ is the Sun's elevation angle at the time of measurement. Constant A which varies with frequency and humidity is found by reference to Fig. 10 of Report 390;

r_2: correction factor for receiving antenna half power beamwidth relative to the angular diameter of the Sun where

$$r_2 = 1 + \frac{401{,}4}{\theta_h^2}$$

and θ_h is the antenna half power beamwidth (min);

f: frequency (Hz);

y_{Sun}: measured values, expressed in numeric units, where:

$$y_{Sun} = \text{antilog}\ \frac{Y_{Sun}\ (\text{dB})}{10}$$

s: Sun flux-density obtained from national standards laboratory; if the Sun flux-density, s, at the frequency (f) of direct interest is not available, the following interpolation equation should be used, instead of linear interpolation, to obtain greater accuracy:

$$s = \left(\frac{s_1}{s_2}\right)^{R_3}$$

where:

s_1: flux at lower frequency (f_1), (J/m^2),

s_2: flux at higher frequency (f_2), (J/m^2),

$$R_3 = \frac{\log (f/f_2)}{\log (f_1/f_2)}$$

c: light velocity (3×10^8 m/s);

y_x: measured values, expressed in numeric units, where:

$$y_x = \text{antilog}\ \frac{Y_x\ (\text{dB})}{10}$$

2.4.2 *Measurement procedures*

A receiver of the type usually found at monitoring stations, having an IF output voltage indicator e.g. voltmeter or oscilloscope, is required. It is highly desirable that the indicator should have a voltage resolution of 0.1 dB (1%) or better. The receiver must be stable and have no significant gain changes during the measurement period.

For the measurements:

- the receiver automatic gain circuit should be switched off;
- the antenna should be pointed towards the Sun and maximum signal obtained. The Sun should have an elevation angle greater than about 30° to avoid atmospheric effects and to ensure that the correction factors r_1 and r_2 are minimally affected;
- the antenna should then be slewed (in azimuth only) away from the Sun e.g. more than a few degrees. The IF voltage level should be noted. This voltage corresponds to the cold sky reference figure;
- the antenna should then be returned in azimuth towards the Sun and the voltage noted. The difference in readings equals Y_{Sun} (dB);
- the antenna should then be slewed (in elevation only) downwards from the Sun to 5° ($\triangleq x°$) elevation and the voltage noted. The difference between this voltage level and the cold sky level is Y_x (dB) for $x°$ of elevation. It should be noted that 5° of elevation ($x°$) is a common reference standard.

The figure of merit (G/T) can then be calculated using the measured values of y_{Sun} and y_x and applying the correction values r_1 and r_2. The Sun flux-density, s, can be obtained from the national standard laboratory.

Using the equation (2), the root-sum-square uncertainty of the measurement is of the order of < 0.5 dB.

It is also necessary that the measurement procedures should be carried out on a bright sunny day.

2.4.3 *Determination of pfd*

Once the figure of merit, G/T, is known, the satellite power flux-density (pfd) at the monitoring station can be determined using the following equation:

$$pfd\ (\mathrm{W/(m^2 \cdot \text{measurement bandwidth})}) = \frac{4\pi k B\,(y_{sat} - 1)}{\lambda^2 (G/T\ \text{numeric})} \qquad (3)$$

where:

G/T: numeric figure of merit of the receiving system at the frequency of interest and at the appropriate elevation angle for the satellite;

k: Boltzmann's constant (1.38×10^{-23} J/K);

B: measurement bandwidth (Hz);

y_{sat}: numeric difference factor between cold sky and satellite emissions where:

$$y_{sat} = \text{antilog}\ \frac{Y_{sat}}{10}$$

λ: wavelength of signal (m).

As in the previous method, a stable receiver with a logarithmic IF output voltage indicator with a resolution of the order of 0.1 dB, is required.

For the measurements:

- the associated satellite receiving antenna is pointed towards the signal of interest and maximum indication with the antenna feed polarization optimized, is obtained. This voltage is noted;
- the antenna is slewed in azimuth to obtain the cold sky reference figure;
- the difference in levels between the two measurements is Y_{sat} (dB). The measurements, if necessary, should be repeated to ensure consistency of results.

The pfd can then be calculated using the known G/T of the receiving system and the equation (3). The pfd value obtained should be corrected for the polarization efficiency (usually unity for adjustable feed systems) and for the appropriate atmospheric attenuation using the published tables.

Using this procedure and with a stable receiver the root-sum-square uncertainty of the measurement is < 1 dB.

The conversion from the measured bandwidth to another e.g. a 4 kHz reference bandwidth, assuming uniform pfd, is carried out as follows:

$$pfd_{converted}\ (\text{dB(W/(m}^2\text{. reference bandwidth)))} = pfd_{measured} + 10 \log \frac{\text{(reference bandwidth)}}{\text{(measured bandwidth)}}$$

However, the assumption that the pfd is uniform across the spectrum is valid only in the ideal case. When a measured pfd is to be converted to another, the power distribution of the signal should be taken into account, as far as possible, so as to obtain a better approximation of the converted pfd.

3. Equipment and facility requirements

3.1 *General*

In general, as in more conventional monitoring surveillance, equipment for monitoring signals from spacecraft must have adequate flexibility to cover a wide range of frequencies, in contrast to the spot frequency coverage that suffices for the needs of a research or operating space agency. The location of the monitoring station should be such that interference from man-made signals and noise is at a minimum.

An electronic counter, providing printed frequency and time records, has been found useful to meet the requirement of taking high-speed measurements to follow the Doppler shift.

For bandwidth measurements, it is desirable to employ a camera to photograph the spectrum display for later analysis. The exact time of taking the picture should be recorded on the film at the time of exposure.

Although good results have been obtained by tracking the satellite signals by manual adjustment of antenna azimuth and elevation, and by manual adjustment of the receiver tuning to follow the Doppler frequency-shift, automatic tracking of both is suggested, if practicable. Further studies should be undertaken to determine practicable means to provide such tracking capabilities. Automatic antenna and receiver tracking is critical for satisfactory monitoring of low-orbit satellites at frequencies above 1000 MHz.

3.2 *Antennas*

Antennas ranging from fixed dipoles to log-periodic types capable of being rotated in azimuth and adjusted in elevation are applicable below 1 GHz. The latter type has been employed for frequencies between 50 and 5000 MHz and a single antenna can provide good general coverage over a 10 to 1 frequency range. Other antennas used successfully include a multi-bay Yagi, single and twin helical, conical helix, dipole with corner reflector, folded crossed doublets with half-wave phasing stubs between the doublets, 4-element folded dipole (in-phase array), and a series of uni-directional "V" antennas with a selective switching arrangement to change orientation. Parabolic reflectors of up to about 13 m (40 ft) in diameter, or horns, are generally used above 1 GHz. Because of the quite sharp directivity of such high-gain antennas, considerable care is required in acquiring and tracking the signals from the spacecraft.

Variations in the polarization of the received field occur owing to the rotation of the spacecraft and to effects on the propagation path (Faraday effect). Antennas not dependent on the polarization, or with linear polarizaton according to the measurement involved, must therefore be used. Antennas capable of receiving with circular polarization both to the left and to the right are very useful for reception. For the purpose of studying the Faraday effect, linear-polarization antennas are more suitable, while for spectrum recordings antennas with vertical polarization are to be preferred.

The antenna gain should be as high as possible to promote a good sensitivity limit for the measuring equipment. For frequencies below 100 MHz, the antennas already existing at monitoring stations can generally be used for the reception of emissions from spacecraft. Helical antennas are very appropriate for the 100 to 500 MHz frequency band. As individual antennas, they provide a gain of about 12 to 16 dB (reference: isotropic antenna with circular polarization). For frequencies above 1 GHz, the use of parabolic mirrors using circular polarization is recommended. For the 1 to 10 GHz band, one parabolic mirror with 1 to 3 interchangeable primary sources using circular polarization is adequate. For a mirror diameter of 3 metres (55% illumination is taken as a basis), antenna gains ranging from 28 dB at 1 GHz to 48 dB at 10 GHz may be obtained.

The direction of antennas may be constantly regulated by manual or automatic means and, if necessary, by programmed adjustment. While it is possible to do without adjustment based on a given programme, automatic adjustment controlled by the received signal itself is essential if the main lobe of the antennas has a mean width of less than about 15°. In any case, automatic adjustment of the antennas is preferred in order to obtain greater measurement accuracy.

Additional information is contained in Report 373 (Kyoto, 1978).

3.3 *Receivers*

For economic reasons and because a general coverage receiver is required at monitoring stations, the extremely low noise figures possible for fixed-frequency receivers used for space research and operational purposes are not equalled by available monitoring receivers. However, the receiver should have a relatively low noise figure and be capable of receiving signals in the order of a few hundredths of a microvolt *or less* at the receiver input. If an adequate signal-to-noise ratio is to be obtained, the noise factors of the antennas with receiver pre-amplifiers should not, depending upon the bandwidth of the amplifier, exceed the following limits:

3 dB up to 250 MHz,
5 dB from 250 to 500 MHz,
7 dB from 1 to 18 GHz.

For frequencies up to 500 MHz, these limits are available with modern equipment. For frequencies above 1 GHz, use is made, with present techniques, of gallium-arsenide solid state preamplifiers in addition to cooled parametric and maser amplifiers, the use of which is usually limited to exceptional cases in fixed monitoring stations. It is noted that since, for nearly all measurement purposes the signal-to-noise ratio can be improved by narrow-band filtering of only part of the emission spectrum, the signal-to-noise ratio in monitoring station measurements may be lower than that admitted in commercial operation measurements.

In general, when observing low-orbit satellites, receivers should be equipped with an automatic frequency tuning device to avoid distortion of the bandwidth and field-strength measurements due to Doppler shifts in the signal frequency.

Receivers should have a wide range of adjustable bandwidth to permit reception of both wide- and narrow-band transmissions from spacecraft. Bandwidths should be so chosen that the whole emission spectrum may pass. Intermediate-frequency bandwidths chosen from 25 kHz, 150 kHz, 300 kHz, 1 MHz, 10 MHz, and 50 MHz have proved to be necessary.

The following receiver outputs should be provided to facilitate measurements: wide-band and narrow-band intermediate-frequency outputs video output and audio-frequency outputs. The intermediate frequency should be the same for all the receivers of a measurement installation so that the same auxiliary equipment may be used with all the receivers.

3.4 *Additional equipment*

To facilitate "acquisition" of the signal from a spacecraft, it is desirable to be able to sweep a narrow band of frequencies automatically and to record the received signals both aurally and on a field-strength recorder. A sweeping device has been developed for this purpose, using a semi-conductor diode as a voltage variable capacitor applied to the frequency-determining circuits of the receiver. In operation, the sweeping device is set to vary the receiver frequency over a band greater than the amount of predicted Doppler shift.

A transmitting spacecraft simulator for making regular checks of the monitoring installation to ensure that it is perfectly suitable for the work is considered important.

Doc. [CCIR, 1966-69b], discusses the use of automatic radio-frequency spectrum recording equipment to permit reception of emissions from a number of spacecraft. The document reports that the results obtained over a period of several days using this equipment in the 136 to 137 MHz band permitted the determination of the expected time of reception, the computation of the period between consecutive passages, and the angle of inclination of the orbit with good accuracy. These factors, together with frequency and bandwidth calculations, facilitated the identification of the spacecraft.

4. Identification of transmissions from spacecraft

Considerable thought has been given to the extent of the need by space administrations and by monitoring stations for a call-sign or other special identifying signal to facilitate identification of emissions from spacecraft and of the effects of installation of equipment for that purpose. It has been concluded that neither need is great enough at present, nor likely to be in the near future, to warrant pressing for the use of an identifying signal.

The techniques employed for space research and operations in satellite and other spacecraft tracking, and foreseen for communications and other uses of space emissions, are such that special identifying signals are normally not essential. There normally being no requirement on the part of the tracker and user for identifying signals, to require their use to facilitate monitoring observations could impose limitations adversely affecting reliability, weight, cost and operation of the system involved.

It has been found possible to identify emissions from particular satellites from signal characteristics and ephemeris data. It would thus appear that the provisions of Article 25, No. 2069 of the Radio Regulations, relating to recognized means of identification other than call-signs, will be met if administrations launching satellites and other spacecraft will continue to make such information currently available to monitoring services. The following information will assist in identification:

- carrier frequencies;
- orbit time;
- angle of inclination of orbit to equator;
- equator crossing time and longitude of crossing;
- perigee and apogee distances;
- signal chacteristics, such as type of modulation and bandwidth for each carrier frequency.

In addition, data on the transmitter power, antenna characteristics including antenna polarization and other related parameters will be useful in the prediction of reception.

From the ephemeris data, monitoring stations have been able to calculate when a particular satellite will be above the horizon for a particular monitoring station and the distances involved, to predict reception and assist in identification. A considerable amount of published information is available, regarding the techniques of satellite and spacecraft tracking and the problems associated therewith (see Bibliography).

4.1 *Special aids for determining satellite positions and expected reception times*

Graphical and slide-rule aids have been found useful for determining reception time for spacecraft based on past reception data (see Doc. [CCIR, 1962a]).

4.2 *Direction finding as an identification aid*

To supplement determination of the exact time of closest approach of a satellite to a monitoring station, by noting the centre point of the maximum rate of the Doppler shift of the carrier frequency, a curve can be plotted to show the change in direction of arrival of the signal with time, as determined by direction-finder bearings or orientation of a highly directional receiving antenna. The maximum angular rate of change will occur when the satellite is nearest to the monitoring station during a particular pass and the information obtained by this method should agree closely with the information obtained from the frequency/time curve.

4.3 *Future possibilities for monitoring procedures involving transmissions from spacecraft*

Procedures have been discussed herein for identification of transmissions from spacecraft, based on past experience of comparing the measured and observed signal characteristics with published information, and by comparing the time of closest approach to a monitoring station – as determined from the measured Doppler shift of the carrier frequency and from varying azimuth (bearing) – with ephemeris data. With the expectation of

increased use of operational satellites, immediate positive identification by monitoring stations of all intercepted spacecraft emissions by these methods may become too time-consuming. Therefore, the following additional procedures may be necessary:

- a communications system involving either passive or active satellites could be identified by the identifying signal of the ground station reflected from or relayed by the satellite, even if this did not immediately identify the particular satellite of a multi-satellite system, and
- in enforcement-type monitoring of spacecraft emissions, particularly when noncompliance with Radio Regulations or cases of interference are observed, monitoring stations could log all possible information concerning frequency and bandwidth measurements, direction of signal arrival, type of modulation and other signal characteristics, together with accurate times of observation and measurements, and request identification, based on these data, from identification and tracking centres.

The procedures should also require the identification and tracking centres to establish the identity of the spacecraft in question on the basis of the above-mentioned data. The time that the satellite is closest to the monitoring location on a given pass, as determined from Doppler-shift measurements, and the carrier frequency as measured for this time, will be of utmost importance to the space centre in effecting identification.

However, further consideration might be given to the incorporation of an appropriate identifying signal.

5. System developed in the United States of America

Within the United States of America [CCIR, 1978-82] the system described below has been developed as a prototype.

5.1 *Capability*

5.1.1 Direct measurements and observations have been found essential in order to determine whether emissions from spacecraft are in compliance with the Radio Regulations. These measurements and observations include:

- carrier frequency,
- occupied bandwidth,
- power flux-density,
- purity of emissions,
- station-keeping accuracy,
- identification of signal source,
- detection of unauthorized transponder use.

5.1.2 Capability to perform additional measurements and observations beyond those indicated above is desirable for purposes of interference modelling and transponder loading studies. The addition of this capability increases the effectiveness of the monitoring facility as a spectrum management tool. The additional parameters that may be measured or observed depend on the type of emission and are as follows:

FDM-FM	– frequency and amplitude of baseband signals – number of voice or other channels – peak deviation of the carrier – frequency and amplitude of the energy dispersal signal, if used
TV-FM	– frequency and amplitude of baseband signals – peak deviation of the carrier – frequency and amplitude of the energy dispersal signal, if used – pre-emphasis characteristics utilized
SCPC-FM	– peak deviation of carriers – highest modulating frequency – carrier spacing – number of carriers per transponder – channel occupied bandwidth – relative power per carrier
Digital (PSK, FSK SCPC, etc.)	– bit rate – number of phases – number of amplitude levels – number of carriers per transponder – carrier spacing – occupied bandwidth of channels – relative power per carrier.

5.1.3 If all types of spacecraft are to be observed, it is necessary that the antenna system be capable of tracking "low- and eccentric-orbit" satellites as well as having the ability to accurately point at any geostationary satellite in the visible segment of the arc.

5.2 *Equipment and facility requirements*

5.2.1 *General*

There are a number of factors in providing the above-mentioned capabilities which must be balanced against cost: frequency coverage, system sensitivity, antenna slewing rate, antenna drive gear backlash (pointing accuracy), ease of changing antenna feed hardware, receiving bandwidth capability, degree of sophistication of signal analysis instrumentation, and degree of automation of measurements. A highly automated and sophisticated spacecraft monitoring system, fully steerable, with continuous frequency coverage across the 1 to 18 GHz spectrum for example, and sensitive enough to give carrier-to-noise ratios of at least 26 dB on all signals of interest would be ideal. However, it will be found, for example, that costs increase nearly exponentially for a several decibel increment of sensitivity improvement. Each administration must, therefore, analyze its own priorities and internal demands for spectrum management information and decide what capabilities it must sacrifice.

5.2.2 *Antenna systems*

In the United States of America, priorities dictated the development of a prototype installation providing maximum flexibility and adaptability of operation, while sacrificing ultimate sensitivity in the interests of saving cost. Because of rapidly escalating costs and complexity for steerable reflector antennas of diameters greater than 5 m, this was the size chosen. A maximum slew rate of 15° per second was specified for the elevation-over-azimuth type pedestal to provide the mechanical capability to track any low-orbiting spacecraft, except perhaps within a few degrees of zenith. Surface accuracy specified for the reflector is sufficient for use up to 18 GHz without appreciable gain degradation. Starting with this versatile basic antenna structure, the most expensive single item, allows addition of equipment on a "building block" basis, as knowledge and requirements grow. An increase in monitoring capability without the need to build new complete specialized systems or to discard inadequate ones is also possible.

The addition of interchangeable feed assemblies of octave or semi-octave bandwidths enables continuous frequency coverage, albeit at some sacrifice in reflector illumination efficiency compared to antennas designed for a single, narrow band. This continuous coverage is important, however, to allow detection of out-of-band emissions. If feeds utilizing dual orthogonal linear elements are used, horizontal, vertical, or circular polarization may be selected using appropriate switches and hybrid networks. If optimum sensitivity on a particular frequency is needed at a future date, a dedicated feed assembly can easily be designed. Similarly, GaAs FET low noise amplifiers of bandwidths equal to those of the feed assemblies may be permanently mounted within the housing of the assemblies, along with narrower-bandwidth amplifiers which may be connected for more optimal noise performance on specific bands.

5.2.3 *Receivers*

The portion of the microwave spectrum from 1 to at least 18 GHz may be adequately covered with a single receiver if that receiver is a quasi-continuously tunable, surveillance-type receiver with a maximum RF and IF passband of at least 50 MHz. Such receivers are now available due to the utilization of synthesized local oscillators and yttrium-iron-garnet technology in the RF stages. A receiver in which the first down-conversion stage is housed in a separate cabinet is very desirable to allow location of the down-converter as close as possible to the antenna, while permitting the remainder of the receiver to be used at a more convenient location. A stable log IF amplifier driving a large faced voltmeter with a dynamic range of at least 40 dB, along with calibrated IF attenuators, is useful for making bandwidth measurements. Proper interface circuitry to allow computer control of the receiver's functions is also desirable.

5.2.4 *Spectrum analysis equipment*

Some sort of capability for spectrum analysis is essential for monitoring emissions from spacecraft. Obtaining the spectrum display by connecting the analyzer to an IF output on the receiver is to be preferred for most purposes as it eliminates the need for placing the analyzer very close to the antenna, as would be necessary if the signal were not first down-converted. This avoids the relatively high transmission line losses that would occur at the microwave frequencies. If a spectrum display exceeding the IF bandwidth of the receiver is desired (to display all the active transponders on a satellite, for example) it is desirable to have the capability to automatically sweep the receiver's local oscillator over a specified range of frequencies. The detected output of the receiver would then be used to drive the vertical axis of a cathode-ray tube display, while the horizontal axis is driven by the sweep generating voltage. The IF bandwidth control of the receiver is then used to specify the swept bandwidth, thus generating an extremely wide-scan spectrum display. A spectrum analyzer is also useful for displaying the baseband of a received signal as described in § 5.3.4.

5.2.5 *Frequency counters*

Frequency counters selected for spacecraft monitoring should cover the frequency range of the receiver's local oscillator and possess adequate sensitivity to allow counting of the frequency of the local oscillator, even at attenuated levels encountered because of long cable runs from the down-converter at the antenna, to the control point. If very large cable losses render it impossible to use a counter because of insufficient sensitivity, the counter may be located with the down-converter, and the display remotely installed at the control point. Counters should have an FM tolerance compatible with the maximum bandwidth of signals to be measured, and time base stability sufficient to allow frequency measurements to an accuracy at least ten times greater than the authorized frequency tolerance for the signals to be measured.

5.2.6 *Additional equipment*

For monitoring emissions from spacecraft other than geostationary satellites, some sort of antenna tracking ability is obviously necessary. Relatively low speed step-tracking may be used in quasi-stationary cases where the angular velocity as perceived at the monitoring point is low, but automatic program tracking is needed for low-orbiting craft. Monitoring systems described here do not have the luxury of a tracking beacon with feedback loop through the satellite that automatically keeps the receiving antenna pointed correctly. A computer of relatively high speed is necessary to generate tracking data in real time, especially if the spacecraft is to be followed from horizon to horizon, which may require elevation angles near 90° as it approaches zenith. A microcomputer of rather limited capacity may be used, however, if observations over only a portion of a single pass are necessary, and an incremental table of elevations and azimuths may be established in advance, to be sent out later to the antenna command system during the desired orbital pass. Ephemeris data is, of course, required, and must be obtained by some means such as published data. A microcomputer can also be used to help automate the collection of monitoring data and calibration procedures if the equipment involved has a suitable instrumentation bus interface.

Signal generators, especially those of the synthesized, bus-controllable variety, are necessary for generating calibrating signals. The frequency range of these generators should cover the tuning range of the primary receiver, as well as all intermediate frequencies used.

Video monitors, television waveform analyzers, and oscilloscopes are all useful in analyzing the characteristics of the demodulated signal. General coverage communications receivers with a good variety of IF bandwidths are also useful, when connected to the demodulator output of the primary receiver, for demodulating intelligence mutliplexed onto sub-carriers.

5.3 *Measurement techniques*

5.3.1 *Frequency measurements*

The frequency measurement methods utilized by the Administration of the United States of America are representative of techniques usable with double-conversion superheterodyne receivers. The receiver in the example described below has intermediate frequencies of 250 and 45 MHz, and the measurement system is shown in Fig. 3.

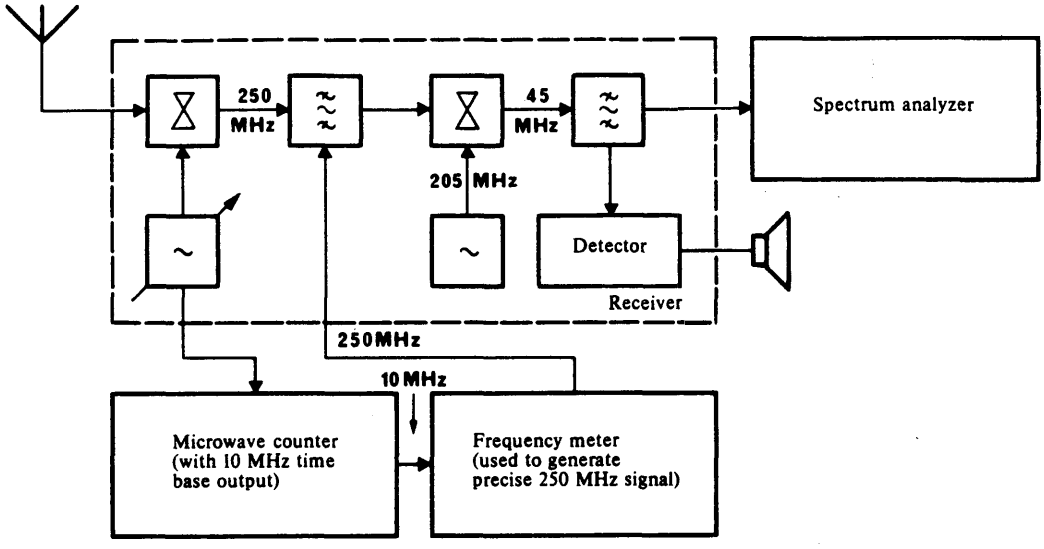


FIGURE 3 - *Equipment configuration for freqency measurements by "zero-beating" the signal component with a reference signal in the IF*

A spectrum display is obtained by sweeping the 45 MHz second IF bandwidth and is used as a tuning aid in the procedure. The time base of the frequency counter serves as the measurement standard, and a precise CW marker signal at the nominal first intermediate frequency of 250 MHz is synthesized from this standard and fed into the first IF. While the observer watches the spectrum display, the signal to be measured is tuned using the receiver's manual tuning control, until the signal component apparently coincides with the 250 MHz reference signal. By expanding the analyzer display and aurally monitoring the audio output of the receiver, it is possible to fine-tune the receiver to zero beat the signal component with the reference. The local oscillator signal of the receiver, which is thus always precisely 250 MHz above the frequency of the signal component to be measured, is sampled and counted. The counter is pre-programmed with a 250 MHz negative offset and therefore provides a direct readout of the measured frequency. The accuracy of this method, when using a receiver with a synthesized local oscillator is, of course, limited by the tuning increment of that oscillator.

An alternative method, as indicated in Fig. 4, may be employed where only one symmetrical FM or AM emission is present in the passband of the receiver. In this method, the receiver's 205 MHz second local oscillator and 45 MHz second IF are sampled and mixed in an external mixer and filtered to regenerate a relatively high-level representation of the 250 MHz IF. This is done to provide adequate levels for the counter (greater than -30 dBm) while allowing control of bandwidth by the IF filtering of the receiver, which is located in the second IF stage. The receiver is then tuned to centre the signal to be measured in the IF passband, and both the nominal 250 MHz high-level signal and the receiver's first local oscillator sample are counted either simultaneously on separate counters, or consecutively on the same counter using a coaxial switch. In either case, the nominal 250 MHz frequency is then subtracted from the local oscillator frequency to obtain the measurement. Receiver frequency drift between consecutive counter readings is generally negligible if the time interval is kept reasonably short.

Successful employment of this method on broadband FM signals requires a counter with adequate FM tolerance. This method will not work on asymmetrical signals or on AM signals where the modulation depth is such that instantaneous carrier levels have excursions below the sensitivity threshold of the counter. The accuracy of the centre frequency determination by this method is limited only by the counter's time base stability, plus or minus one count of the least significant digit.

The frequencies of emissions from non-geostationary spacecraft which are either approaching or receding from the monitoring station will be subject to Doppler shifts.

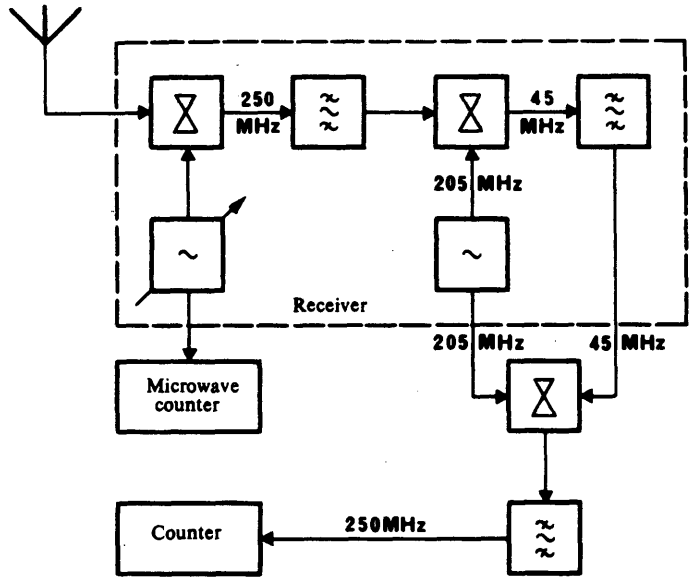


FIGURE 4 - *Equipment configuration for frequency measurements on symmetrical emissions*

5.3.2 *Bandwidth measurements*

In the United States of America, bandwidth measurements of frequency-modulated spacecraft emissions are made according to Recommendation 443 using the x dB method at levels of 6 dB and 26 dB below the reference level, provided that the signal levels are adequate and provided that the passband of the receiver is adequate to contain all significant modulation components of the signal in question.

The reference level is established using the method described in § 7 of Report 275. An accurate method of determining the proper reference level for signals employing digital modulation techniques remains a particular topic for further study.

5.3.3 *Power flux-density measurements*

Power flux-density can be calculated using measured power delivered by the antenna to the receiver according to the formula:

$$pfd = \frac{4\pi p}{g\lambda^2}$$

where:

pfd: power flux-density (W/m^2),

p: power delivered to a receiver by a matched antenna (W),

g: power gain of the antenna over an isotropic radiator,

λ: wavelength in metres at the frequency of interest,

assuming the gain of the antenna at the frequency of interest is accurately known. Antenna gain calibration may be performed by a manufacturer or separately by the user on a properly designed antenna range at incremental frequencies, but this could be an expensive procedure.

Alternatively, gain calibrations may be accomplished using celestial bodies of known power flux as source emitters. With larger antennas ($\geq$ 10 m) having very low noise receiving systems, calibration using radio stars may be practicable. For smaller antennas, such as the one described, however, the Sun is a more useful source of calibrating signals. Values of solar flux-density valid to a specified degree of

accuracy for a specified time, at various frequencies, may be obtained from national standards organizations or observatories. Solar flux-density values for frequencies other than those provided will, of necessity, have to be obtained through interpolation. Accuracy of the calibration will, therefore, be limited by both the accuracy of the solar flux data and the vagaries inherent in interpolation.

5.3.4 *Baseband measurements and analysis*

For FM signals, a spectrum display of the modulation baseband may be obtained by connecting a spectrum analyzer to the demodulator output of the receiver. By calibrating the vertical axis of the analyzer display using standard modulating signals, the contribution to the carrier deviation of individual baseband channels may be determined directly from the analyzer. The Bessel function null method is used to determine the frequency and amplitude of the calibrating signal.

General-coverage communications receivers may also be connected to the detected output of the main receiver and used to tune and demodulate intelligence multiplexed onto sub-carriers. Using these receivers and conventional HF and VHF frequency measurement techniques, the frequencies of baseband component signals may also be measured.

5.4 *Interference resolution and emission identification*

5.4.1 Because of spatial considerations, the capabilities of a fixed, earth-bound, spacecraft monitoring system with respect to interference resolution are, of course, primarily limited to analyzing interfering signals in the down link. However, even though interference to an up-link signal may occur from a terrestrial source, observation of the manifestation of the interference on the down-link signal may give experienced monitoring observers sufficient clues to aid in identifying the offending signal.

5.4.2 As spacecraft are not required to emit a special identifying signal, identity must be established by comparing observed orbital parameters with ephemeris data, comparison of frequencies in use with published frequency assignments, analysis of the characteristics of the signal, and analysis of demodulated message content. Identification of transmissions by such methods is generally successful, and there appears to be little need for signal identification requirements for spacecraft at this time.

6. System developed in the Federal Republic of Germany

6.1 *Capability*

The tasks as indicated in § 1 have led to the inclusion of the following measurement possibilities into the system:

- frequency band recordings,
- frequency measurements,
- power flux-density measurements,
- polarization measurements,
- determination of the position of geostationary space stations,
- class of emission and modulation characteristics,
- bandwidth and spectral composition of emissions.

6.2 *Requirements for measuring equipment and measurement accuracy*

6.2.1 *Frequency band recordings*

In the terrestrial radio monitoring service, frequency band recording systems have been used successfully for a long time. Such systems are well suited for investigating frequency bands as to their occupation by space station emissions. It is also possible to record emissions from satellites on which no other information is available. In the case of low-orbiting satellites, particular advantages are offered by orbit-specific relationships, which yield a characteristic recording and allow the following basic data to be derived to a good accuracy:

- transmit frequency,
- period,
- time of return into the coverage area of the monitoring station.

Since the system has to be operated with non-directional receiving antennas, the sensitivity loss must be compensated for by a reduction of the analyzing filter bandwidth.

The noise-limited sensitivity of the system, which can be operated between 100 MHz and 2.5 GHz, is about −138 dBm, relative to 1 kHz bandwidth.

6.2.2 *Frequency measurements*

Methods for measuring the transmit frequency of geostationary space stations do not differ from those used for measuring the transmit frequency of terrestrial radio stations. However, the situation is different for satellites moving in near-Earth orbits, since their emissions are subject to Doppler effect at the receiving site. Provision has therefore to be made for a frequency measuring device which covers Doppler-shifted frequencies. For the method of measurement, which is fully computerized, see § 2.2 and in particular [CCIR, 1970-74].

Taking into account the present state-of-art, provision was made so that:

- the individual frequency measurement error does not exceed 10^{-9};
- the measured individual frequency values are based on a time-frame which is in full synchronism with UTC.

6.2.3 *Power flux-density measurements*

The concept of power flux-density measurement is based on the following considerations:

- the power flux-density values specified in the Radio Regulations are average values. For averaging, use can be made of a thermo-electric power meter connected to a suitable IF output of the receiver or analyzer;
- at the IF output, the satellite signal should not be affected by a frequency shift caused by the Doppler effect, so as to prevent the spectral components, during the measurement period, exceeding the measuring filter bandwidth;
- the polarization of the measuring antenna should be matched to that of the signal to be measured, in order to avoid degrading the measurement results;
- as long as the power flux-density in the sub-band under consideration can be regarded as uniform, it is not necessary to measure within a 4 kHz bandwidth. If the effective bandwidth of the filter is known, easy conversion is possible.

It is considered that the desired measurement accuracy, independent of frequency band, should be such that the total error does not exceed ± 2.5 dB.

6.2.4 *Polarization measurement*

Space stations make use of right-hand circular, left-hand circular and linear polarizations. The decision as to the extent to which the radio monitoring station for space services should be equipped with polarization measurement facilities was based on the following:

- space stations in near-Earth orbits alter their position continuously relative to the observer. For this reason and due to the fact that the satellite normally rotates, polarization at the receiving site is subject to constant change;
- the orientation of the polarization plane of a linearly polarized signal transmitted by a geostationary space station relative to the horizontal plane of the observation point depends on the geographical location of the latter relative to the satellite.

In view of the above and the fact that a radio monitoring station for space services should be capable of observing both geostationary and orbiting space stations, provision has been made for a polarization measurement system which allows continuous measurement and recording of the polarization parameters of the received signal.

Since polarization measurements in connection with frequency re-use methods are not regarded as a task of the radio monitoring service, no stringent requirement for measurement accuracies was imposed.

6.2.5 *Determination of the position of geostationary space stations*

Space stations on board geostationary satellites which use any frequency band allocated to the fixed satellite service or to the broadcasting satellite service, have to be kept within ± 0.1° of the longitude of their nominal position (see Radio Regulations, Article 29, Section III). The duties of a radio monitoring station for space services should therefore include the measurement of the nominal positions of geostationary satellites. The question as to what extent the resultant measurement requirements are economically justifiable is dealt with below.

A tolerance of ± 0.1° of the longitude of the nominal position defines an angular segment of the equatorial orbital plane. In the limiting case where a monitoring station is operated on the equator, position-keeping would have to be checked by angle measurements in the elevation plane only of the antenna. When the monitoring station is moved to the South or North, a rotation takes place into the azimuth plane of the monitoring antenna, which in the case of 50° latitude (Federal Republic of Germany) means that the longitudinal station-keeping tolerance of a geostationary space station is predominantly measured as an angular difference in the azimuth plane of the antenna. It reaches the value ± 0.13° for a longitudinal difference of 0° between the monitoring station and sub-satellite point and drops to ± 0.085° for a longitudinal difference of 60°. From this, a measurement error can be derived which should be about ± 0.01°, i.e. smaller by a factor of 0.1.

However, the above considerations do not justify the abandonment of measurements in the elevation plane of the positional accuracy of the satellite. This is not possible as the antenna must also be used for tracking and taking angle measurements on, low-orbiting objects. Other information which can be derived from the actual non-stable orbital behaviour of a geostationary satellite requires a measurement accuracy corresponding to that of the azimuth plane.

There exists a relationship between the half power beamwidth of an antenna and the maximum achievable direction-finding accuracy. When use is made of the monopulse method (see § 6.3.2), the following approximation holds:

$$R \simeq \pm n \cdot \theta$$

where:

R : angle measurement error,

n : improvement factor,

θ : half power beamwidth.

For optimized narrow-band antennas $n = 0.01$. For a broadband antenna of the kind generally used by a radio monitoring station for space services, a factor of 0.1 seems to be realistic. The remaining variable θ cannot be reduced without limit. For example, a parabolic antenna with a diameter of 12 m and an efficiency of about 55%, has the following half power beamwidths and angle measurement accuracy values:

TABLE I

Frequency (GHz)	1.65	2.2	11.4
θ (degrees)	0.86	0.65	0.13
Angle measurement accuracy (degrees)	± 0.09	± 0.08	± 0.02

A comparison of the values indicated in Table I with a measurement accuracy of 0.01° (corresponding to a distance of approximately 7 km in the geostationary orbit) shows that the ability to mount the nominal position of geostationary satellites under the measurement conditions that were achieved by the Federal Republic of Germany's radio monitoring station for space services is limited, at least in the lower frequency range.

6.2.6 *Class of emission and modulation characteristics*

The cost of the equipment that would be needed in a radio monitoring station for space services if class of emission and modulation characteristics of a space station had to be identified and measured completely could be extremely high because the configuration of such a system is often very complex. For the time being, provision has therefore been made only for conventional equipment for demodulation of narrow-band (maximally 300 kHz) amplitude, frequency and phase-modulated emissions.

6.2.7 *Bandwidth and spectral composition of emissions*

The differences between the requirements for making bandwidth measurements of terrestrial stations and of space stations are dealt with in detail in § 2.3. Care was taken so that any Doppler shift in the spectrum could be eliminated by a phase or frequency shifting control built into the measurement receiver (see also § 6.2.3) and which provides a frequency-stable spectrum for analysis. The two conditions which had to be met were a Doppler frequency shift $\leqslant$ 80 kHz and a spectrum bandwidth $\leqslant$ 300 kHz i.e. maximum IF bandwidth of receiver.

6.3 *Technical concept of the receiving equipment*

6.3.1 *Antenna systems*

Two antenna systems allow full coverage of the frequency range 130 MHz to 13 GHz. They are mounted on separate pedestals, at distances of 100 m and 200 m from the station building. Below 1300 MHz, the cross-over point between the two systems, multiple feed array composed of four log-periodic antennas is used; above 1300 MHz, a 12 m parabolic antenna of the Cassegrain type is used.

A special feature of the parabolic antenna is a motor-driven system for changing frequency of which the initial construction stage has now been completed. The system allows switching between two frequency bands. The functional principle of this system is illustrated in Fig. 5. It uses a rotatable deflecting mirror arranged directly behind the main reflector apex on the main antenna axis, in such a way that the receive power is exactly reflected into the elevation axis. On the elevation axis, one exciter horn is accommodated to the right and left of the mirror apex in each of the cabins 1 and 2. Frequency band changes can therefore be performed simply by turning the deflecting mirror by 90° which requires about 15 s. The mirror system can be extended to include additional frequency bands. Possible other solutions are shown in Fig. 6. Whatever system is a finally chosen, it can only be accommodated in cabin 1 for space reasons.

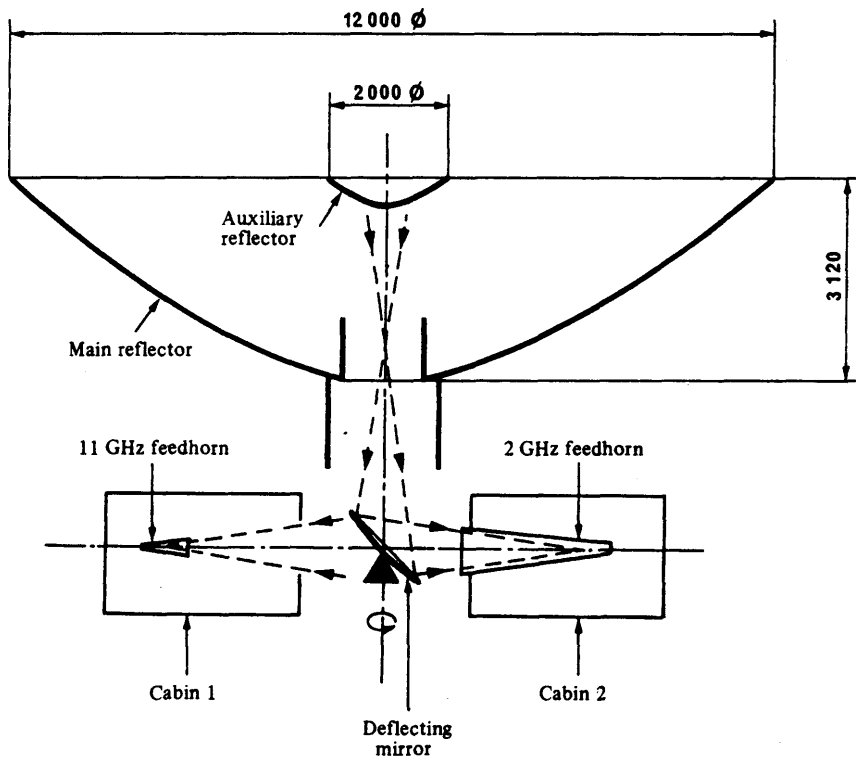


FIGURE 5 – *Geometric-optical configuration of the 12 m Cassegrain antenna (dimensions in mm)*

Ø : diameter

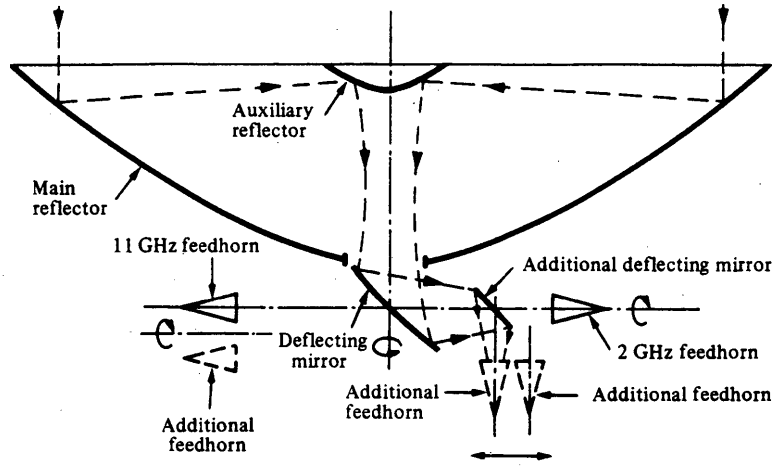


FIGURE 6 – *Possible future extension to additional frequency bands and lay-out of the actual beam path*

As the half power beamwidth of the parabolic antenna is relatively small, it can be increased by a factor of 3 by using a so-called defocusing device, in order to facilitate the acquisition of satellites whose exact position is not known. This is accomplished by a sub-reflector which is divided into two concentric sections, the central part of which can be adjusted with respect to the fixed outer ring along the antenna main axis.

Since the parabolic antenna is also used for reception of low-orbiting satellites, comparatively high tracking speeds are necessary.

Either antenna provides two linear and, relative to each other, orthogonal polarization planes. In the lower frequency range, they are obtained by use of crossed dipoles, and in the upper range by means of a polarization diplexer arranged behind the mode coupler (see § 6.3.2), by which the received signal is split up into two power components (ΣA, ΣB) – the prerequisite for measurement of the polarization parameters (see § 6.3.4).

6.3.2 *Antenna tracking system*

In auto-tracking operations of the two antennas, the monopulse method is used. Pre-programmed and manual control methods are also possible. Unlike other techniques, the monopulse method offers the advantage that information on the extent and direction of the antenna deviation from the target is simultaneously and continuously available for the two tracking planes (azimuth and elevation). A characteristic feature of the monopulse method is that, in addition to the sum pattern (in the present system one each for polarization planes A and B corresponding to ΣA and ΣB), direction-finding patterns are produced for the tracking planes of azimuth and elevation, which exhibit a null on the antenna main axis. The error signals $\Delta_{azimuth}$ (ΔAz) and $\Delta_{elevation}$ (ΔEl), obtained from the direction-finding patterns, enable conditions to be drawn on the extent and direction of the angular misalignment.

The direction-finding patterns can be produced by means of the two different systems, the multi-feed and the multi-mode system.

In the lower frequency range (< 1300 MHz) there is a multi-feed array. The direction-finding pattern is obtained by means of a monopulse comparator by summation and subtraction of the voltages supplied by the four-feed array.

In the upper frequency range, the direction-finding pattern is produced by means of a multi-mode system. The latter works according to the principle that in a waveguide with an adequately large cross-section, higher-order wave modes are excited in addition to the fundamental mode whenever the direction of incidence of the electromagnetic wave does not coincide with the waveguide axis. As a result, two difference signals are available at the output of the mode coupler.

To reduce the number of receiver channels, one difference signal is added to the other with 90° phase shift after the comparator for the upper range it is already included in the mode coupler, so that only one difference signal channel is needed for further processing.

6.3.3 *Restriction of frequency range to individual sub-bands*

The antennas are followed by low-noise pre-amplifiers for the frequency bands indicated in Section A of Table II. In the lower frequency range, the pre-amplifiers are connected directly to the receivers which are tunable from 130 MHz and 1300 MHz, whereas for the upper range, broadband converters are inserted. The restriction to individual sub-bands means the abondonment of the wide frequency range provided, in principle, by the antennas. However, this disadvantage is outweighted by a considerable improvement of the intermodulation behaviour and noise characteristics of the receiving system within these sub-bands.

Three receivers with equal measurement characteristics are provided. Two of them are designed for three channels (Σ *A*, Σ *B*, Δ) and equipped with a tracking unit. A fixed assignment exists between these receivers and the antennas. The third receiver is of the two-channel type (Σ *E*, Σ *B*) and can be operated alternatively in the lower and upper frequency range. Additional frequency counters and d.c. recorders allow the measurements described under § 6.2 to be made. The storage of the measured values in the station computer is eventually envisaged but is only partly realized at the present time.

6.3.4 *Technical realization of the polarization measurements*

The polarization of the received signal is obtained electronically in each receiver and the ellipticity and orientation of the major axis are indicated as angular values. The method employed requires the use of an antenna system with two orthogonal polarization planes and the separate processing of the power components of the received channels (Σ *A* and Σ *B*) which must not exhibit any phase or amplitude differences. To achieve this, every frequency change is followed by a calibration process. The method uses a dynamically operating evaluation unit which adapts itself continuously to polarization changes. Obtaining the polarization in the receiver, as opposed to obtaining it in the antenna feed system, allows independent polarization measurements to be made on two frequencies of a satellite at the same time.

6.3.5 *Significant features of the system concept*

Some of the significant performance characteristics of the radio monitoring station for space services are listed in Table II; Fig. 7 shows the basic configuration of the receiving equipment. Although the figure is largely self-explanatory, mention should be made of the following features:

- one antenna allows parallel reception on two frequencies of a space station. Below 1300 MHz, the choice of frequencies is unrestricted. They must only lie within the individual sub-bands. Above 1300 MHz, parallel reception is possible only in the frequency band selected. It also allows parallel reception for auto tracking according to the monopulse method on one frequency and polarization measurements on two frequencies. All converters must be designed for three (Σ *A*, Σ *B*, Δ) or two channels (Σ *A*, Σ *B*);

- the receive frequency is obtained on the second converter by means of synthesizers and the desired receive frequency is entered via the station computer;

- after conversion to the IF of 60 MHz, another conversion takes place so that the sum channels are offset by $\pm$ 1.8 MHz relative to the difference channel. The multiplexing of channels achieved in this way allows their combined signal and to be transmitted to the station building over a common cable. This method offers the advantage of an increased long-term stability of the receiving paths;

- the measurement receivers have a maximum IF bandwidth of 300 kHz. In the upper frequency range (1300 MHz-13 GHz) this is not sufficient. Therefore, a broadband path is provided on which the total IF spectrum available at the output of the first converter can be transmitted to the station building;

- in the lower frequency range (< 1300 MHz) the broadband path serves the purpose of by-passing the four optimized individual sub-bands, thus enabling reception on other frequencies. This requires individual tuning of the combined filter and amplifier unit inserted for improvement of the intermodulation behaviour. The noise characteristics of this path are, however, worse than those of the individual sub-bands;

- in the 2 GHz band, the antenna feed system is extended by a compensation network which increases the direction-finding accuracy of the antenna.

TABLE II – *Most important performance characteristics of the radio monitoring station for space services of the Federal Republic of Germany*

	130-1300 MHz	1.3-13 GHz
A. *Antennas and frequency bands*		
Antenna type	Log-periodic dipole array	
Antenna gain	13 dBi	1.65 GHz: 44.3 dBi 2.2 GHz: 47 dBi 11.4 GHz: 60 dBi
Half power beamwidth	$\leq 45°$	1.65 GHz: 0.86° 2.2 GHz: 0.65° 11.4 GHz: 0.13°
Frequency bands	136 -138 MHz 149.9-150.05 MHz 399.9-402 MHz 460 -470 MHz	1.5-1.8 GHz 2.1-2.3 GHz (combined in one band unit) 10.95-11.8 GHz
Figure of merit, G/T		1.65 GHz: 17 dB(K^{-1}) 2.2 GHz: 19.5 dB(K^{-1}) 11.4 GHz: 35.5 dB(K^{-1})
Noise figure	$\leq$ 5.8 dB (at comparator input for all frequency bands indicated)	
B. *Other features*		
Tracking speed and acceleration of the antennas:		
– azimuth	6°/s, 10°/s²	16°/s, 6°/s²
– elevation	6°/s, 10°/s²	3,5°/s, 6°/s²
Doppler frequency compensation (kHz)	± 40	± 40
Dynamic range (dB)	90	90
Measurement accuracy		
Power flux-density (dB)	± 2.5	± 2.5
Frequency	1×10^{-9}	1×10^{-9}
Polarization:		
– ellipticity (e) (arc tan of the axial ratio)	± 8°	± 4°
– orientation of the main axis of polarization	$\pm \frac{8°}{\cos 2e}$	$\pm \frac{4°}{\cos 2e}$
Azimuth and elevation angles		0.1 × half-power beamwidth

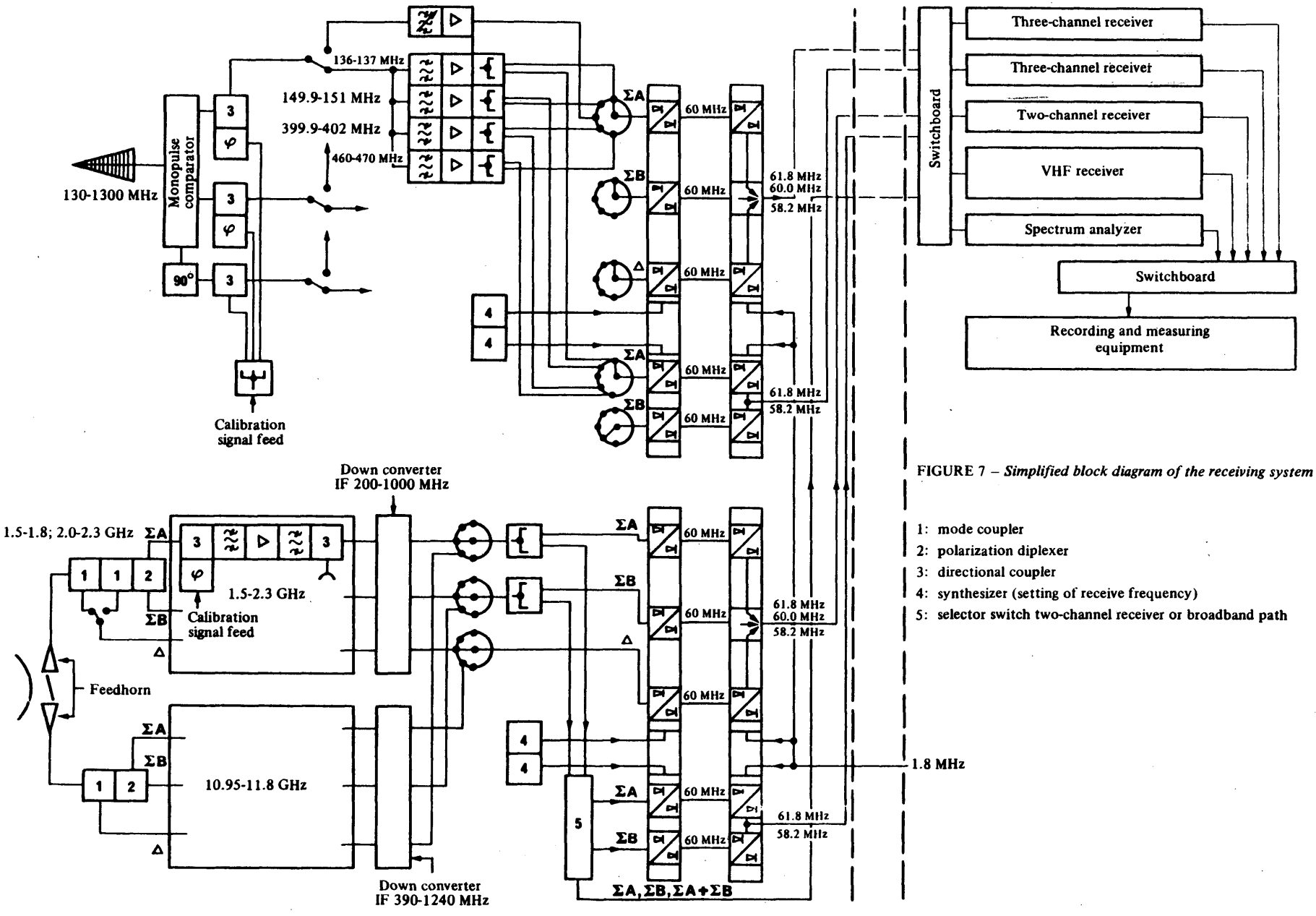


FIGURE 7 – *Simplified block diagram of the receiving system*

1: mode coupler
2: polarization diplexer
3: directional coupler
4: synthesizer (setting of receive frequency)
5: selector switch two-channel receiver or broadband path

7. Simplified small earth monitoring station

The monitoring stations described above can undertake full scale monitoring of communication satellite emissions, including assessment of channel quality. However, to assess the basic characteristics of emissions by satellites registered with the IFRB, the practical efficiency of the use of frequency assignments and field strengths in accordance with the Radio Regulations, it is possible to use simplified small earth monitoring stations with a lower figure of merit and a limited range of standard earth-station receiver equipment as shown by experimental study conducted in the USSR [Yastrebstsov, 1990]. Another advantage of such stations is that they can be transportable.

Setting up such earth stations for technical monitoring purposes does not involve heavy outlays, expensive development of additional special equipment or high operating costs and is justified in the context of intensive use of the geostationary orbit.

The accuracy of these measurements should be determined by an error analysis of the system which considers, among other parameters, the intermodulation distortion of the low-noise amplifier, the frequency stability of the local oscillator and the port-to-port isolation of the circulator/mixer.

7.1 Parameters to be determined:

- orbit occupancy
- simple orbit data
- observed polarization
- measured frequency
- occupied bandwidth
- deviation from assigned frequency
- class of emission
- field strength or power flux density

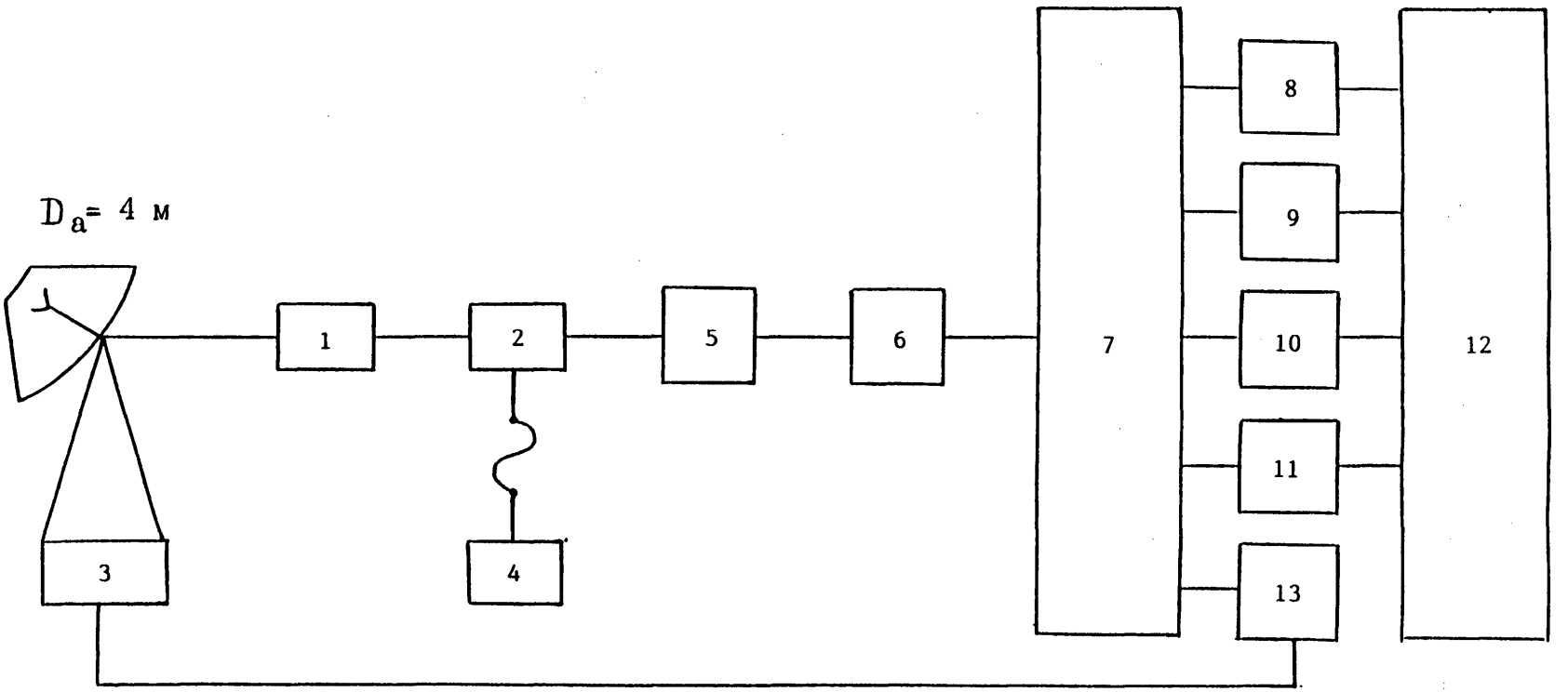
7.2 Measuring apparatus

A block diagram of a small earth monitoring station is shown in Figure 8. It gives some parameters of the equipment when standard instruments in the 4 GHz range are used. The sensitivity of such equipment makes it possible to evaluate the characteristics of a transmission from a satellite operating in the 4 GHz band, with a spectral density at the output of the satellite antenna in the direction of the point of measurement under clear sky conditions of -53 dBW/Hz.

7.2.1 Antenna system

The antenna system can consist of parabolic reflectors having a diameter of 2.5 - 4 m with a simplified support and rotating device and a pointing system.

The antenna system can be based on a single-reflector or double-reflector design. To maximize the antenna gain it is advisable to use a range of interchangeable feeds operating in the appropriate frequency bands allocated for different kinds of satellite services within the 1.5 - 12 GHz band.



1 - Polarization unit
2 - Directional coupler
3 - Electric drive of the support and rotating system
4 - Oscillator
5 - Low-noise amplifier
6 - Step-down frequency converter
7 - Power divider
8 - Spectrum analyser
9 - Oscillograph
10 - Selective voltmeter
11 - Demodulator
12 - Signal recording and processing equipment
13 - Pointing apparatus

FIGURE 8

Block diagram of small earth monitoring station

The antenna can be pointed in azimuth within the limits $\pm 90^{\circ}$ and in elevation within the limits 0° - 90°, roughly to start with, by means of the antenna's manual system and then by remote control for fine setting. The satellite can be tracked automatically from the signal received if the signal-to-noise ratio is high enough. Figure 9 shows the maximum available accuracy of satellite coordinate reading as a function of frequency for antennas with a diameter of 4 m when ordinary methods of measurements are used.

Interchangeable polarization units make it possible to receive signals with circular and linear polarization.

7.2.2 Receiver

The earth monitoring station receiver consists of a low-noise amplifier and step-down frequency converter connected in series, such as are used at earth stations. This considerably reduces the cost of the measuring apparatus.

When the working frequency band is changed, the receivers and frequency converters are replaced at the same time as the feed system and the corresponding waveguide components.

The signals received at the output from the step-down converter at intermediate frequency, e.g. 70 MHz, can be fed for analysis, processing and recording to a spectrum analyser, oscillograph, etc.

The figure of merit for the measurement apparatus with an antenna diameter of the order of 4 m may attain 7 dB/$^{\circ}$K at a frequency of 1.5 GHz and 27 dB/$^{\circ}$K at a frequency of 12 GHz. The sensitivity of such an earth monitoring station is adequate for analysing the emission characteristics of most existing and notified satellites.

The measurement error will be determined by the design characteristics of the equipment used.

7.2.3 Measuring equipment

Various spectrum analysers, oscillographs, cameras and computers can be used to observe, evaluate and record the signals received.

To read the levels of the signals received and the frequency values for comparison purposes a control signal can be fed to the input of the low-noise input amplifier through a directional coupler from a separate oscillator.

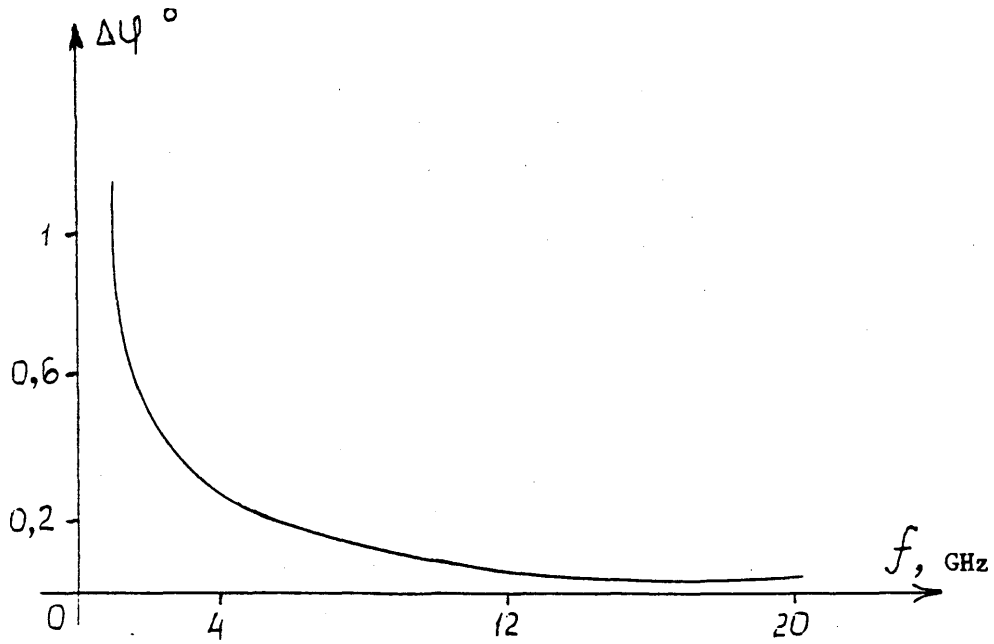


FIGURE 9

Maximum available accuracy of satellite coordinate reading for antennas with diameter of 4 m

8. Experimental study in Japan

8.1 Purpose

The Japanese Administration is interested in space radio monitoring. Prior to the establishment of operational monitoring facilities, basic experiments are being conducted in order to clarify the image for the function and operation of the monitoring system.

8.2 Scope of the study

The object of the monitoring at present day is classified into the two classes of geostationary and low altitude satellites. Considering the importance in space telecommunications, geostationary satellites are taken as the primary object of the monitoring.

The points that have to be clarified by the experiment are as follows:

- with what procedure the system equipment should be operated;
- in what way the measurement data should be acquired, recorded, processed and displayed.

According to the ultimate purpose of the monitoring, the system can be classified into the following two types/grades:

- quick survey - to make an outlined observation of the current status of the use of orbit and frequency in a short time;

- detailed monitoring - to make a precise measurement of orbital position and radio spectrum for a specified satellite.

The quick survey will be performed for the first step, and whenever a possibility of trouble is found the detailed monitoring should be activated as the second step. The detailed monitoring is likely to be supported by the system introduced in section 6. Reported in the following is a study aiming at developing the quick survey concept for the geostationary orbit.

8.3 Quick survey experiment for the geostationary orbit

8.3.1 Experimental system

In this experiment, 2 GHz signals are surveyed for basic study, considering the ease of the acquisition of satellite signals owing to a wider beamwidth.

The experimental system is shown in Figure 10. The antenna, 11 meters in diameter with a front-feed, has the 3 dB beamwidth of 0.8°, which gives the degree of resolution of the satellite position survey. The system G/T is 25 dB/K.

The receiver passband is 2.20 GHz - 2.32 GHz, which is divided into six equal sub-bands of 20 MHz width. Each of the sub-bands is measured by the spectrum analyser, by successively tuning the frequency of the first local oscillator (a frequency synthesizer).

Satellite signal detection is made by peak detections for the observed spectrum. If a peak appears above a specified threshold level, its frequency, orbital longitude and the peak level are recorded onto a disk to make the survey data. The frequency measurement error is ±400 kHz, which is specified by the resolution of the peak detection in the spectrum analyser. The detection threshold is 15 dB above the noise level (bandwidth of the spectrum analyser is 100 kHz), which means that a satellite signal is detected if its maximum power flux density is greater than -171 dBW/m^2.

8.3.2 Operation procedure

The survey site (Kashima station, Communications Research Laboratory, Minstry of Posts and Telecommunications) is located at 36°N/141°E, at which the arc of 70°E - 210°E of the geostationary orbit is visible with elevation angle higher than 5°.

Along the visible arc, the orbital longitude is scanned by the antenna beam at the constant increment of 0.8° (equals to the antenna beamwidth), using the calculated Az/El for the antenna pointing. The Az/El calculation is based on the nominal radius of the geostationary orbit. Six sub-bands are measured and the survey data are recorded at each orbital longitude. Reception polarity is unchanged during one scan. One scan of the survey takes about one hour.

As is obvious from the description, the survey procedure requires no information in advance such as the ephemeris data of the satellites.

8.3.3 Experimental results

Examples of the survey data are shown in the form of longitude-frequency distribution in Figure 10. Two cases of scanning survey are displayed, where several Japanese satellites are recognizable as noted. BS-2a and BS-2b are

broadcasting satellites making a pair of primary and back-up satellites at the same position. The signals at 71° and 194° are also considered geostationary signals as they are observed recurrently in different scans. The figure indicates that the current status of the use of the geostationary orbit in a specified frequency band is clearly outlined by the longitude-frequency survey.

It is possible, however, that a part of the survey data are undesired signals from ground radio stations which came into the antenna side lobe or from a low altitude satellite which accidentally crossed the antenna beam. The signals appearing at 81° and 182° in Figure 11 are likely to be such non-geostationary signals. Care should be taken to identify such signals.

Among the satellites in the geostationary orbit, some satellites have large orbital inclinations so that they undergo considerable north-south motions, and some satellites are operated intermittently with occasional radio emissions. It is thence possible that some satellite signals are missed in the scanning survey while they are desired signals. Reliable acquisition of geostationary signals and exact discrimination of non-geostationary signals are thus the problems requiring further study.

8.4 Summary

From the experimental study of the monitoring for the geostationary orbit, the following results were obtained:

- it is probable that the outlined status of the orbit/frequency use is observed by the quick survey procedure, without requiring *a priori* data such as the ephemeris of the satellites;

- the form of longitude-frequency distribution is effective for the display of the survey data.

The exampled cases of the survey indicate, however, that the following problems are still under study:

- how to eliminate the non-geostationary signals;

- how should the signals from quasi-geostationary satellites or from intermittently operated satellites be recognized as valid signals?

On the basis of the presented experiment and further study for the problems, the quick survey at higher frequency bands is to be investigated.

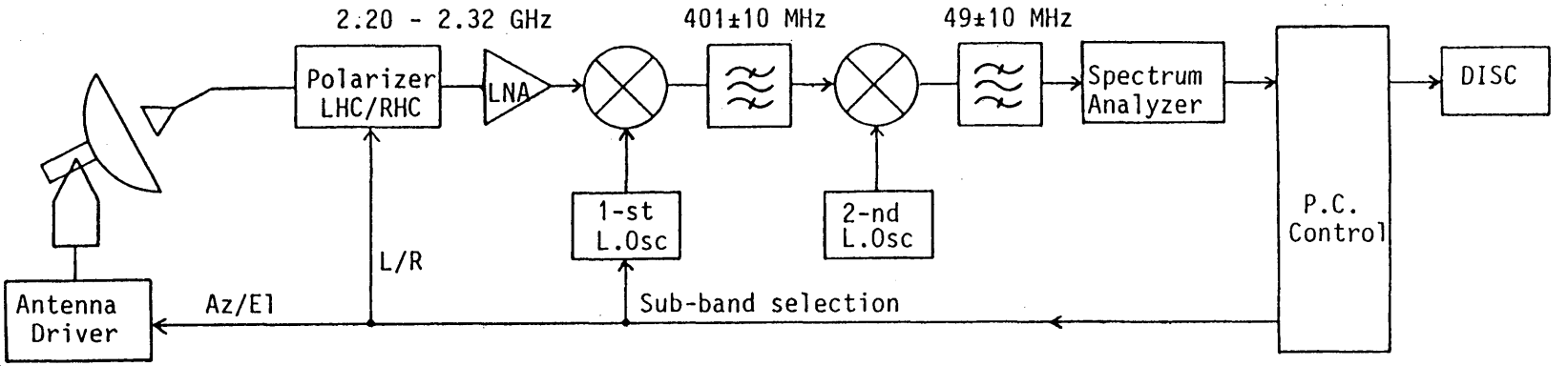
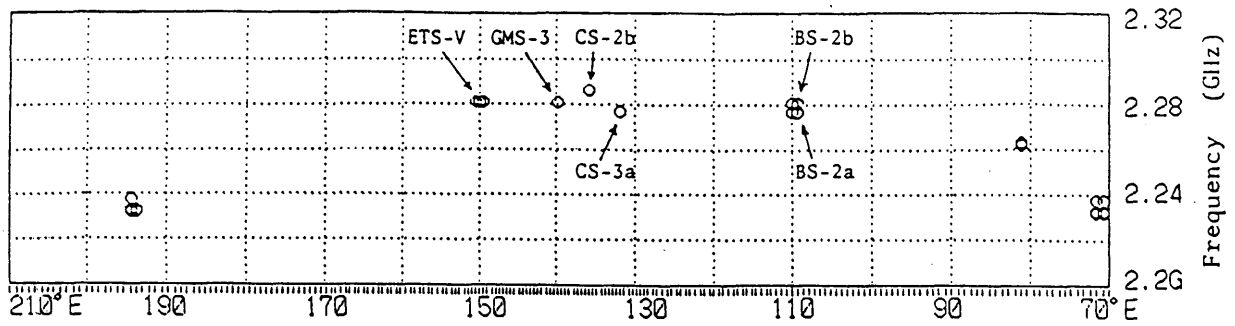
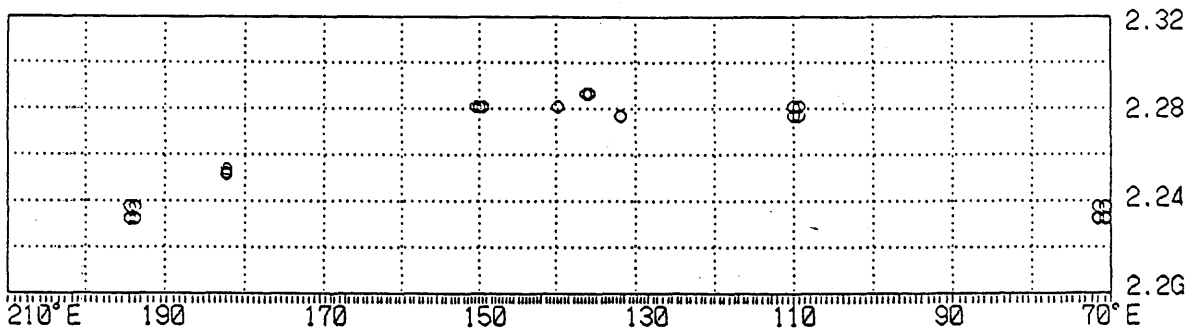


Fig. 10 Experimental 2 GHz quick-survey system



22 July, 18H - 19H



23 July, 18H - 19H

FIGURE 11

Survey data examples (1988)

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REPORT 372-6

DIRECTION FINDING AT MONITORING STATIONS

(Question 28/1)

(1966-1970-1974-1978-1982-1986-1990)

Study Group 1, in response to Question 28/1, seeks to furnish:

- information concerning fixed monitoring stations;
- information concerning mobile monitoring stations;
- a summary of information contained in various documents submitted concerning the accuracy of direction finding bearings and of the subsequently determined positions of electromagnetic sources.

Further information in response to this Question is contained in Chapter 12 of the Handbook for Monitoring Stations.

1. Fixed direction finders (MF and HF)

Methods of direction finding (DF) and procedures for improving the accuracy of bearings at fixed monitoring stations are generally directed towards producing an acceptable degree of accuracy on waves propagated via the ionosphere. Ionospheric propagation exerts pronounced effects on the characteristics of radio waves, which seriously affect the results obtainable with the different methods of direction finding. The simple rotatable or crossed loop method of direction finding on ionospheric waves fails at the outset to meet the requirements of accuracy and consequently need not be considered. Since sky-waves throughout the entire HF range are of primary importance in direction finding associated with international monitoring, we may also disregard methods that are limited to either very narrow ranges of frequencies or to frequencies lying above the HF spectrum. Considering also that it may often be necessary to locate unidentified sources of interference or unauthorized transmitters, we may dismiss from discussion those methods of direction finding which depend upon the control or cooperation of the transmitter whose location we are trying to determine.

1.1 *India*

1.1.1 Direction-finding equipment with 8-element Adcock antennas, rotating goniometer and automatic visual indicator is being used at the monitoring stations of the Indian Administration. By the use of a suitable pre-amplifier after the goniometer at the centre of the antenna system and using simple remote control technique, it is possible to locate the antenna system at a distance of 300 m from the main reception/indicator equipment.

1.1.2 For the type of direction-finding equipment indicated, it is possible to have the remote control facility at a moderate extra cost. The availability of the direction-finding equipment alongside other monitoring equipment has proved to be most convenient in the monitoring work. The same operator who is working on a monitoring channel can quickly take bearings of any station he desires. Selection of site with respect to the main monitoring building does not become very critical, as the antennas can be located up to a distance of 300 m from the main building. The effect of other antennas in the field also can be thus reduced.

1.1.3 The commercial direction-finding equipment has been modified to use a monitoring receiver which has better frequency stability (calibration at every 1 kHz interval right up to the highest frequency of 30 MHz). This again provides a great facility in tuning quickly and correctly the wanted station for the direction-finding bearing.

1.1.4 The above direction-finding system provides a bearing accuracy between ± 1° for stable signals exceeding 50 μV/m and ± 2° for stable signals exceeding 3 μV/m. Considering that, in the HF spectrum, the bearing inaccuracies contributed by fading and instability of the signals and other factors could be of this order, it is felt that the basic accuracy of the equipment as mentioned above should be adequate for monitoring purposes.

1.1.5 The direction-finding system employing a rotating goniometer has certain disadvantages as compared with one using phase amplitude balanced amplifiers etc. With the rotating goniometer type, bearings are difficult to take in the presence of interference. Separate calibrations are required when selectivity of the receiver is changed and it is not always possible to choose the optimum selectivity to avoid interference. Most systems can, therefore, offer either one or two selectivities such as 1.2 kHz or 3 kHz. Proper adjustments of tuning and gain control settings, etc., are required for taking good bearings. However, usually, trained operators are employed at monitoring stations, hence these drawbacks can be overcome to some extent. This type of direction-finding equipment is fairly cheap.

1.1.6 It is the experience of the Indian Administration that operational personnel for taking bearings in the HF bands require considerable training. This involves training in proper adjustments of various controls as also training in taking bearings especially when the signals are fading or the bearings are swinging. Untrained operators tend to read bearings with certain bias towards one particular value, thus tending to give a false indication of class of bearings.

1.2 *United States of America*

1.2.1 *Adcock system*

1.2.1.1 *Description*

An Adcock direction finder in use in the United States of America, having a useful frequency range of 2 to 40 MHz, utilizes two vertical dipoles each about 6 m (20 feet) long and spaced about 6 m (20 feet) apart. The centre line of the dipoles is approximately 6 m (20 feet) above the ground. Provision is made for manually or automatically controlled rotation with a maximum rotational speed in the latter condition of 20 to 30 revolutions per minute. Provision is made for aural or visual readout of the bearing; also for automatic averaging of the bearings over a period of up to 20 or more rotations of the array with a visual display of the averaged bearing.

1.2.1.2 *Bearing accuracy*

In a series of tests over a period of several months in 1962, Adcock bearings were taken on 60 W transmitters in locations unknown to the observers. The distances involved were random, to simulate the conditions of actual problems, and varied from a few hundred to more than 1600 km (1000 miles) in the various tests. Thirteen Adcocks in widely separated locations produced bearings having an accuracy of 2° or better in 82% of the tests. For further information on this subject of bearing accuracy, see Report 93.

1.2.2 *Wullenweber system*

1.2.2.1 *Description*

The Wullenweber direction finder in common use in the United States by the FCC employs 60 antenna elements deployed in a circle approximately 140 m in diameter. The reflector screen normally used in a true Wullenweber system is not implemented, therefore sense information is not present and bidirectional patterns are produced. Lack of sense information is not normally a disadvantage since other units in the network can provide information to resolve the sense ambiguity. A rotating capacitive goniometer provides selectable null or lobe patterns utilizing 25, 50, or 100% of the maximum aperture available. Both manual and automatic rotational modes are provided and a visual display device allows the integration of multiple bearing samples as well as storage of the bearing after the signal goes off the air.

1.2.2.2 *Site calibration tests*

At the time of installation, each direction finder is tested by taking a number of bearings on readily identified signals whose positions are known. An observed bearing is compared with the calculated great circle bearing and errors are analyzed to detect possible site and survey errors. In an attempt to reduce the sources of error, only strong signals with stable, displayed bearings are used for data in these tests.

Tables I and II give data from two such calibration tests. Site 1 is a very good direction finder location (flat terrain, minimum of interfering structures, and remote from industrial activity) while Site 2 is considered less than optimum as it is located near an urban area with many man-made structures and terrain discontinuities. The influence of site effects in increasing the direction finder standard deviation can be readily seen.

TABLE I - *Site 1*

Frequency band (MHz)	Number of samples	Average error (degrees)	Standard deviation (degrees)
< 2	40	+ 0.4	1.15
2-10	39	+ 0.43	0.77
10-25	48	+ 0.64	0.68

TABLE II - *Site 2*

Frequency band (MHz)	Number of samples	Average error (degrees)	Standard deviation (degrees)
0.5-1.5	27	+ 1.02	3.85
1.5-5.0	22	+ 0.43	1.55
5.0-10	23	- 0.41	1.25
10-15	59	- 0.15	1.11
15-20	21	- 0.43	0.97
20-30	7	+ 0.04	0.49

It can be seen that the characteristics of Site 1 are reflected in considerable improvements in standard deviation, particularly at low frequencies. The influence of antenna capture area to wavelength ratio in reducing standard deviation is shown in both sets of data.

1.2.2.3 *Analysis of performance*

Performance data of the direction-finding network has been kept for approximately five years on actual cases in which the subject transmitter was definitely identified and therefore the exact transmitter coordinates could be determined. The error performance of each monitoring station participating in the exercise was calculated and used to analyze the accuracy of bearing information in comparison with other stations in the network. As might be expected, there was a considerable difference in performance found early in the program which was subsequently improved by one or more of the measures of improved equipment maintenance, operator education, and correction of site defects. The data presented below is offered as an indication of the performance which might be expected from the equipment described.

Bearing distribution data from about 1000 bearing samples is shown in Fig. 1 with normalized bearing deviation as the abscissa, and number of bearings submitted as the ordinate. The distribution could be expected to be Gaussian in nature but it is seen that half degree increments are favoured. This is thought to be a consequence of the fact that the direction finder scale is calibrated in one degree increments and the operators tend to make only rough interpolations. It is apparent that the majority of the bearings are concentrated within a 3° error range. The bearing error probability may be estimated from the curve of Fig. 2 which is a cumulative error curve beginning with the value of $-5°$ and extending to $+5°$. Circled data points on Fig. 2 represent true Gaussian distribution for the sampled data. The confidence interval for a particular error limit may be easily read from this curve. It is seen that less than 1% of the reported bearings exhibited errors of more than 5°. Approximately 60% of all bearings reported fell within 1.5° of the true bearing.

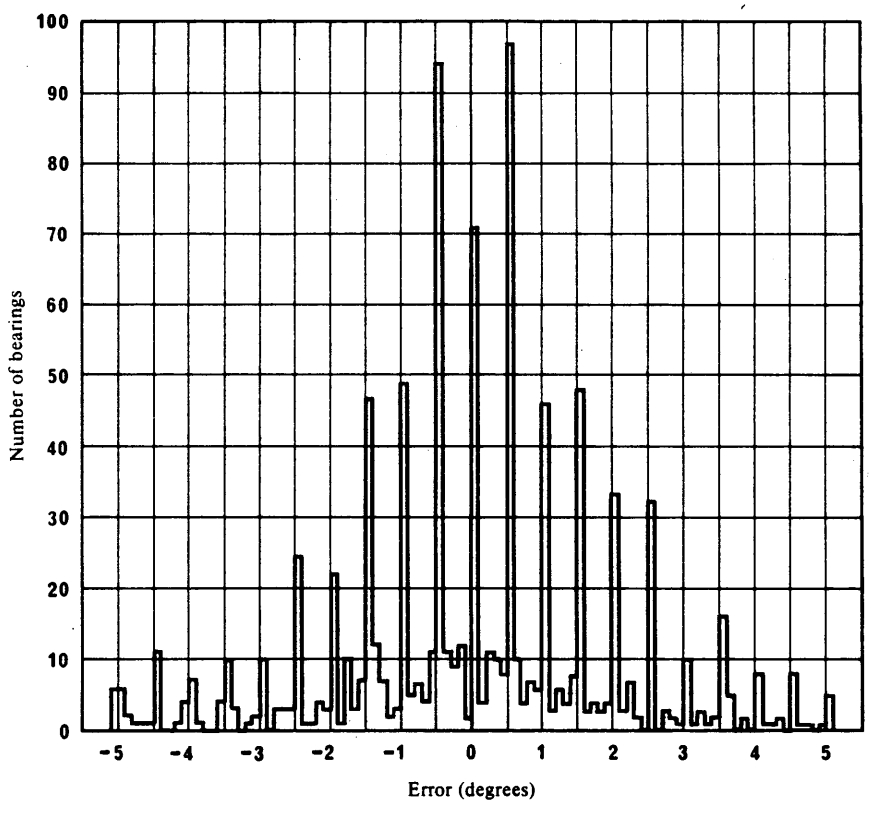


FIGURE 1 - *Bearing error distribution (932 samples)*

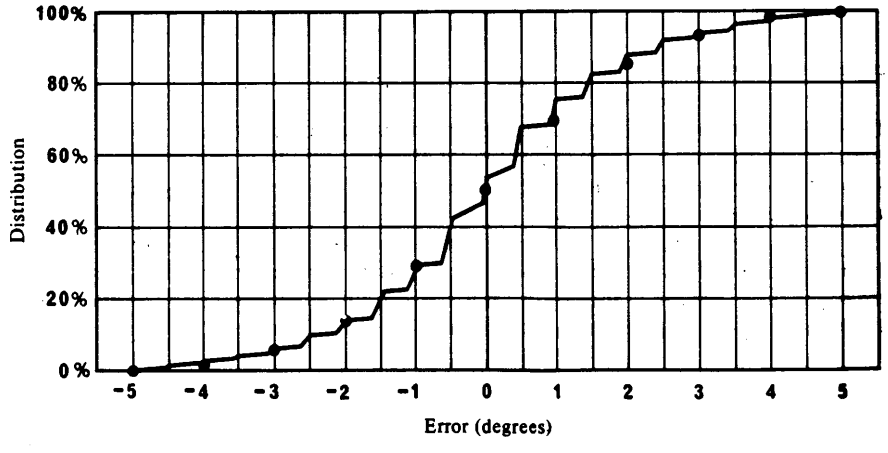


FIGURE 2 - *Bearing error - cumulative distribution*

In an attempt to improve the position-fixing performance of the network, the bearing errors from 277 sample reports were analyzed as a function of the auxiliary data reported to determine if a correlation existed between the class of bearing and the standard deviation expected.

The results of this analysis (see Table III) shows that the bearing error is reduced for bearings of higher classification. This suggests that overall performance of the system will be improved if bearings are classified accurately and if more importance be given to highly-classified bearings in performing the position-fixing calculation.

TABLE III - *Bearing performance by classification and errors of measured values*

Bearing classification (1)	Number of samples	Average error (degrees)	Standard deviation (degrees)
A	28	0.5	1.6
B	135	0.2	2.1
C	114	0.2	3.4

(1) For an explanation of bearing classification, see the Handbook for Monitoring Stations, Chapter 12, page 14, Table II.

One of the factors which affect bearing classification is the signal strength of the received signal. Although even weak signals may exhibit stable bearings, it seems logical to assume that the accuracy to which a signal null may be determined would depend somewhat on the ratio between the signal and the background noise. Table IV gives the results of a sorting of the bearing errors by signal strength reported and shows a positive correlation between the strength of the signal and a reduction in bearing error.

TABLE IV - *Bearing performance by signal strength and errors of measured values*

Signal strength	Number of samples	Average error (degrees)	Standard deviation (degrees)
S1	2	2.5	4.0
S2	54	0.5	3.4
S3	151	0.1	2.7
S4	63	0.3	1.9
S5	7	- 0.3	1.1

It can be inferred from the results of the above data that the best direction-finding performance is obtained when strong and class A signals are present.

Direction-finding measurements must be made for all qualities of the signal, therefore it is useful to determine the average operating conditions for direction finding and the performance of the systems under the constraints of signal parameters encountered. The performance data for approximately 1700 samples has been sorted and analyzed in terms of both signal quality and strength and is given in Table V. The typical bearing is taken under less than ideal conditions, which in turn affects the accuracy obtained. Given the realities of operating conditions, it is imperative that operators accurately determine the parameters of the subject signal so that the quality of the bearing may be adequately judged.

TABLE V - *Distribution of bearings by class and signal strength*

Signal strength		S1	S2	S3	S4	S5
Class A	Number of samples	1	1	40	103	27
	Average error	- 0.5°	- 1.8°	- 0.2°	+ 0.1°	- 0.7°
	Standard deviation	-	-	2.1°	2.1°	2.6°
Class B	Number of samples	1	65	527	309	15
	Average error	+ 2.0°	+ 0.3°	+ 0.1°	+ 0.1°	- 0.5°
	Standard deviation	-	3.2°	2.7°	2.3°	2.3°
Class C	Number of samples	40	216	301	51	-
	Average error	0	- 0.3°	+ 0.3°	+ 0.1°	0
	Standard deviation	3.1°	4.3°	3.1°	3.7°	-

From analyses of bearing performance data just presented, it is apparent that the errors exhibited by a direction-finding system will depend directly on the parameters of the signal being examined. One of the most crucial factors in evaluating the bearing information reported by a station in a direction-finding network should be the assurance that an accurate description of signal parameters accompanies the bearing submission. It should be possible to significantly reduce the size of the location areas if more importance is given to the better classified bearings submitted.

1.3 *Japan*

1.3.1 *Doppler system*

A remotely controlled direction finder using the Doppler method has been in practical use in Japan since late 1979. The antenna system for this direction finder, located at an unattended monitoring station, consists of 24 antennas which are precisely arranged on the circumference of a circle of 36 m diameter. One additional antenna is at the centre of the circle with switching circuits. The height of an antenna is 9 m.

This direction finder is effective for amplitude-modulation and angle-modulation emissions in the frequency range of 300 kHz to 30 MHz.

Loss of sensitivity at the lower frequencies caused by switching noise can be avoided by sequentially sampling the outputs of the antennas at a rate of 130 times per second, with two triangular-shaped switching signals that have a phase difference of 180°.

The instantaneous values measured at the unattended monitoring station are transmitted via a voice grade circuit to the manned monitoring station and stored in memory. The results of measurement are indicated on a CRT display unit and printed out after averaging the instantaneous values.

1.3.2 *Accuracy of bearing*

Emissions from known radio stations which are located at about 50 km to 3000 km from the direction finder have been measured at random. Test results are shown in Tables VI and VII. The errors of measurement are within a permissible range for practical use.

TABLE VI - *Measured frequency bands and errors of measured values (all samples were averaged over a period of ten seconds)*

Frequency band (MHz)	Number of samples	Average error (degrees)	Standard deviation (degrees)
0.3-2	118	+ 0.47	2.81
2-5	50	+ 0.73	2.36
5-10	108	+ 1.54	2.74
10-15	131	+ 0.72	2.65
15-20	109	+ 0.05	2.50
20-25	72	+ 2.48	2.45
25-30	13	+ 4.28	2.63

TABLE VII - *Bearing performance by classification and errors of measured values*

Bearing classification(1)	Number of samples	Average error (degrees)	Standard deviation (degrees)
A	269	+ 1.20	2.66
B	106	+ 0.71	3.21
C	21	+ 0.61	4.29

(1) For an explanation of bearing classification, see the Handbook for Monitoring Stations, Chapter 12, page 14, Table II.

1.4 *Federal Republic of Germany*

1.4.1 Doppler system

A new kind of shortwave direction-finder system, which works according to the Doppler principle, has been developed and tested in the Federal Republic of Germany. As a result of the special antenna system, using crossed-loop antennas as the individual radiators, it is possible to take bearings not only on ground waves and sky waves with a small angle of incidence but also, and without any restrictions, on high-angle sky waves with an accuracy being of the order of 1° (without propagation effects). The elevation angle of the incident sky wave is also indicated in this system. Through the use of the Doppler principle, bearings can be performed automatically and remotely controlled.

The direction finder is fitted with a microprocessor controller, which enables direct exchange of data with a computer system. The use of magnetic antennas means that the direction finder can be set up on practically any subsoil.

Neither special earthing arrangements nor special cables between the detached antenna and the instrumentation are necessary.

1.4.2 Detection of extremely short signals using HF/VHF/UHF direction finding

Modern telecommunications make increasing use of transmission techniques that ensure that emissions are difficult to intercept and resistant to ECM. One such method is frequency hopping.

Special signal processing methods are required for the interception of these signals with a dwell time of between a few milliseconds and some hundreds of milliseconds per frequency.

A novel integrated search and DF system permits the interception of signals in the range from 2 to 512 MHz with a dwell time of about 1 ms. The system operates on the Adcock/Watson-Watt principle with fully automatic, digital signal processing. It enables any frequency subrange to be scanned with variable step width and adjustable thresholds. During scanning, the bearing of each significant emission is determined. Scanning of discrete frequencies entered in a scan table is also possible. The bearings are presented as a bearing-versus-frequency display on a colour CRT with co-ordinate grid.
It is, of course, also possible to use this DF system in the conventional way, i.e. on a fixed frequency with the bearing being indicated on the screen as a straight line or an ellipse.
The system is operated via a menu-driven, touch-sensitive control panel, if required also over long distances. Integration in large-scale monitoring and radiolocation system is also possible.

The system accuracy is $\pm$ 1° (1σ) for signals of 10 ms duration and $\pm$ 2° (1σ) for signals of 1 ms duration.

1.5 *United Kingdom*

1.5.1 *Circular array system*

This system, developed in the United Kingdom, is designed for use at various latitudes. It is effective for all modulation types. It provides direction finding facilities over the frequency range 1.5-30 MHz. It can also support any number of monitor receivers, all operators having independent access simultaneously at any frequency in the 1.5-30 MHz range, to any of 24 directional beams (with performance similar to rhombics) which, between them, cover the full 360° of azimuth. The beams provide about 10 dB of gain relative to isotropic in the forward direction. Depending on frequency, the signal gain exceeds the noise gain by 5.4 dB to 9 dB.

1.5.2 *System design*

The antenna system consists of two concentric rings, each of 24 elevated feed vertical monopoles. The outer ring of 12 m monopoles has a diameter of 150 m and is usable for DF purposes over the range 1.5-15 MHz. The inner 50-m diameter ring of 6 m monopoles is usable for DF purposes over the range of 6-30 MHz. The design is such that the DF system is generally externally noise limited. A goniometer receives the signals from each antenna of the outer ring via an amplifier and splitter, the latter also driving the beam-forming apparatus of the monitoring facility. The rotating broadside sum and difference beam outputs from the goniometer are formed at any instant from the combined outputs of eight adjacent monopoles. Another goniometer is provided for the inner ring, thus providing overall DF coverage of the 1.5-30 MHz range. Up to eight DF operator positions can be operated simultaneously and independently from the pair of goniometers. In the manual system, the sum and difference outputs of a goniometer are received and displayed. The difference beam display null is used to determine the signal bearing, the sum display being used to resolve ambiguities. Self-test facilities are built in.

The system is effective for elevation angles of arrival from about 3° (dependent on site and also frequency) up to 60° or more, and also for ground waves. The wide aperture (0.6 of a wavelength at 1.5 MHz up to 4.3 wavelengths at 30 MHz) provides high accuracy on both ground and sky-wave signals.

A recently developed computerized enhancement of the system, called "auto DF", provides for full automatic and remote control of the DF function. This operates solely on the sum beam output of the goniometer, thus obtaining improved performance at low signal-to-noise ratios. The auto DF system samples data over a 0.5 s interval and calculates the bearing corresponding to the greatest degree of symmetry of the peak, interpolating to the nearest 0.1°; this is called a snap bearing. A further algorithm processes the results from sixteen snap bearings to produce a final bearing and an associated confidence factor, derived from the variance of the snap bearings.

1.5.3 *Accuracy*

As is well known, an Adcock array, with its four antenna elements, exhibits cyclic errors in each octant of azimuth. The system as described, exhibits a pattern of 144 cyclic errors in azimuth. The magnitude of these errors is dependent on frequency and bearing and is generally less than 0.5°. Other electrical and mechanical errors do not exceed 0.2°, resulting in a typical instrumental accuracy of around one half of 1°. Tests of ground-wave and sky-wave performance produced the following results:

- on five sets of ground-wave calibration trials, carried out at a distance of 32 km, an average standard deviation of 0.9° with a range of ± 0.3° was obtained;
- using known sky-wave targets, the results obtained from eight sets of trials gave an average standard deviation of 1.5° with a range of ± 0.5°.

1.5.4 Transportable direction finders

A transportable direction finder covering the frequency range 1.5 - 30 MHz has been developed in the United Kingdom. In man-portable packages weighing a total of 300 kg the complete system can be erected from a vehicle in 30 min. Working on the interferometer principle, it has 6 antenna elements arranged in a pentagon with a central reference element. The use of crossed loop type elements enables both low-angle and high-angle sky waves to be received.

The processor operates by measuring the phase difference between the signals on the central reference antenna and the signals on each of the outer elements. This phase information is then used to calculate both the azimuth and elevation angles of the signals. The processor uses an FFT algorithm so an instantaneous frequency spectrum of the received signal can be displayed simultaneously with the DF information. Using the frequency selectivity of the FFT, the system automatically computes the direction of the maximum energy in the passband. This enables bearings to be taken on any type of modulation in the presence of interfering signals or noise. The equipment has typical on-site accuracies on a class A site of:

1.5 - 2 MHz a standard deviation of 2 degrees

2 - 10 MHz a standard deviation of 1 degree

10 - 30 MHz a standard deviation of 0.5 degree

As the system uses modern technology it is capable of computer control and remote operation with the operator located away from the DF site. The system can also be erected as a permanent installation.

1.6 *France – direction-finder system using the interferometry principle*

A direction finder, using interferometry, determines the direction of arrival of a radio source from the phase differences of signals received by an array of N vertical antennas. The antennas can be arranged in different geometrical configurations – square, triangular or pentagonal – according to the frequency range covered and the type of installation of the direction-finder system. The number N of the antennas can be adapted to the accuracy required and, by using interferometry, it is possible to operate with antenna apertures which are much larger than the wavelength in order to minimize multipath effects. The phase differences can be obtained directly with two phase-matched receivers or indirectly with normal receivers after treatment of the amplitude-modulated signal whose modulation is used to obtain the phase information. The sampled phase differences are processed by a DF processor using azimuth determination algorithms and also the angle of elevation of the electromagnetic wave. In the HF region, this opens up new possibilities for location when the interferometer is used in association with ionospheric characteristic measurement techniques.

1.7 Australia - HF direction finding using interferometric techniques

A very accurate HF direction-finding system has been developed in Australia. Based on a simple 7-antenna interferometer, the system provides full single station location capability, as well as the ability to act as a network of line-of-bearing instruments at remote locations. A series of trials on long-range targets demonstrated the system's ability to provide bearing estimates with RMS errors of 0.8 degrees, and range estimates with median absolute accurancy of around 7% of range. The system can operate successfully down to -5dB S/N ratio in a 1 kHz bandwidth, and employs spectral processing to discriminate against co-channel interferers. Descriptions of the system and its capabilities are given by Jeffrey et al. (1988), and by McNamara (1988a,b).

2. Remote control of direction finders

2.1 *Remote control of HF direction finders*

Remote control of direction finders presents the equipment designer with many challenges. Control algorithms must be written to perform all the tasks normally handled by the operator. These include initializing the equipment, tuning the receiver, taking and analyzing the bearing. In addition, new tasks will have to be accepted and scheduled, and the data from completed tasks made available on demand.

A number of communication options are available for satisfying the various requirements both at the DF site and among the several possible control points. These include, but are not limited to, hardwire landlines, terrestrial microwave links, fibre optic waveguides, and conventional HF radio. Each method offers both advantages and disadvantages when all the factors such as bandwidth capability, freedom from interference, and economic requirements are considered. It is probable that a mixture of options will be found desirable to fully satisfy the various requirements within a complete system.

It should be expected that a remotely controlled direction finder will not always be as efficient as having an operator on site. Difficulties may be experienced when transmissions are of brief duration, or where two or more signals share the same band. For these reasons, the remote operator must have real-time audio information available as well as control capability for assuring that the proper signal is being monitored and its direction found.

More information can be found in Chapter 12 of the Handbook for Monitoring Stations.

2.2 *Remote control of VHF and UHF direction finders*

2.2.1 *Outline of the system*

A remotely controlled direction-finding system using Adcock antennas has been developed and put into operational use in Japan. This system consists of two remotely controlled direction-finding stations and a manned monitoring station.

AM and FM emissions in the frequency range 25 MHz to 1 GHz can be measured by this system.

The measured data such as bearings, signal strength, and quality of bearings, together with the received audio signal are sent from the two direction-finding stations to the monitoring station by means of a VHF radio circuit.

These data are displayed on the liquid crystal display (LCD) of the monitoring equipment as three-digit figures. Since the bearing lines and their cross-point are also indicated on a CRT map display, the operator can easily find the estimated position of the emission. In the case where the emission is made from a mobile unit, the operator can estimate the moving direction of the mobile unit by watching the cross-point on the CRT display.

2.2.2 *Accuracy of bearings*

A field test was carried out for about six months from March, 1984. For the test, a mobile station was used as a target and the station made emissions in the 150 MHz band from various places in the urban area of Tokyo as well as its suburbs.

From the test results, it can be considered that the measurement errors of the system are within a range permissible for operational use.

The monitoring area in which the required accuracy of measurement is obtained, can be determined by taking account of the geometrical relationship, including distance, between the direction-finding stations, the monitoring area and the target stations.

The true positions were found with a probability of 70% to be within 3 km of the estimated positions obtained by using this system. The antenna heights of the direction-finding stations were as follows:

Station A: 20 m (120 m above sea level)
Station B: 30 m (30 m above sea level)

In the case where the target emissions were located in the urban area of Tokyo, the measured bearings were found with a probability of 68%, to be within ± 9° of the true bearings. Since the antenna height of station B was lower than that of station A, it is considered that the measurements by station B were more influenced by obstacles in the propagation path (see Table VIII).

TABLE VIII – *Error of measured bearing (degrees)*

Environments	*n*	Station A		Station B	
		μ	σ	μ	σ
Urban area	55	−0.6	7.4	−0.1	9.4
Suburbs	33	−0.2	4.5	1.5	10.7

n: number of samples
μ : average value of errors (degrees)
σ: standard deviation (degrees)

2.3 Remote control of VHF and UHF mobile direction finders

2.3.1 Outline of the system

A remotely controlled direction finding system using Doppler principle has been developed and put into practical use in Japan. This system consists of two mobile DF stations equipped with a Doppler direction finder and a monitoring equipment respectively. AM, FM and CW emissions in the frequency range 25 MHz - 1 GHz can be measured by this system.

Two mobile DF stations are connected by VHF radio data links and operated in cooperation with each other. Either one of two stations can be assigned as the master station and the other the slave station, or vice versa. The master station transmits a DF instruction by which both direction finders are controlled to perform DF measurement on the same target at the same time. The bearing data obtained at each station is evaluated according to "grade" based on the bearing swing, then sent to each other with evaluation data and field strength. The data processor on board each station processes the bearing and evaluates data into bearing beams. The bearing beam is made of two lines of which opening angle is decided (single line when the evaluation is the best) by the evaluation data. Position of the emitter is then estimated from the cross section of two bearing beams.

2.3.2 Special features of the system

(1) Telescopic pole
The telescopic pole is installed at the center of vehicle. One of three Doppler antennas is mounted on the top of pole. Height of the DF antenna is set by the up-down controller. Since the control is of pneumatic type, the controller functions with an air-compressor and pole controller.
The pole can take one of three positions by the control switch.

LOW	The pole is completely lowered.
MID	The pole is set to midway for DF while the vehicle is moving (height above ground: approx. 3m)
HIGH	The pole is set to the maximum height for DF by stopping the vehicle. (height above ground: approx. 6m)

(2) A magnetic compass sensor
On the front of roof, a magnetic compass sensor is mounted to acquire the refernece azimuth information for DF measurement.

(3) A color CRT display
A color CRT display is provided in the data processor for indicating the map, bearing beam and other information. There are three kinds of display selected by "DISPLAY MODE" switch on the control terminal board, "AREA 1", "AREA 2" and "OPTION".

a) When the switch is set to "AREA 1", the CRT displays the bearing beams of master and slave stations, covering the range of 130 km in east-west direction and 100 km in north-south direction, centered on the metropolitan area.

At the same time, data such as date, time, frequency, measured data with evaulation, and position of each vehicle are also displayed on the left hand column of the CRT screen.

b) When the switch is set to "AREA 2", the CRT displays bearing beams, covering the range of 9.6 km in the east-west direction and 7.5 km in north-south direction, centered on the vehicle location. All the data of left hand column at "AREA 1" is also indicated.

c) When the switch is set to "OPTION", the CRT displays the area 2 bearing beams. On the left hand column of the CRT screen, various "touch panel control" items are also displayed.

From the touch panel, various man-machine interface such as selection of control function, scrolling of map and manual setting of data etc. can be performed.

(4) Function of correction
Various functions of correction are provided. For example, with respect to errors caused by vehicle body effect in the High and Mid positions of the antenna height, the correction data for 32 bearings and up to 64 frequencies are respectively stored in memory, and the errors can be directly corrected. Moreover, these correction data can be read out and altered by operating the keyboard.

3. Mobile and portable direction finders (MF, HF, VHF and UHF)

3.1 *General*

Because of the requirements of portability, mobile-unit direction finders normally used in automobiles are necessarily limited to types that are physically small and easily manipulated. For the HF range one practical choice is a rotating loop. Since mobile units operate primarily within a pre-determined fix area and generally within the ground wave of the unknown target transmitter, quite satisfactory results have been obtained with a simple rotating loop. In general, the requirement of a boom for either spaced loops or spaced vertical elements practically eliminates these types from consideration for mobile units.

The accuracy of mobile-unit loop bearings on ground-wave emissions can be improved by averaging multiple bearings, as with long-range direction finders. The advantage of a calibration curve, covering the full 360° of azimuth, when the direction finder is mounted on an automobile, should not be overlooked. Check bearings should be taken on known transmitters to confirm the proper operation of the equipment.

Portable DF systems are normally used at temporary sites that are not in congested environments. Best locations for portable units are at the top of tall buildings overlooking the vicinity (particularly above 30 MHz), or in a clear, obstruction-free area where greater accuracy and reception range are found than with mobile units.

The accuracy of bearings taken by a mobile direction finder in the VHF/UHF range is affected by reflecting objects in the vicinity of the direction-finding vehicle. In a moving vehicle, an assessment of the bearing accuracy may be accomplished by simultaneous graphical display on a video monitor of a large number of successively measured bearing and received-signal level values. When the direction-finding vehicle moves through an electromagnetic field disturbed by adjacent reflecting objects, strongly varying successive bearing and received-signal level values can be observed. If relatively constant received-signal levels occur together with nearly stable bearings, the bearings taken may be assumed to be correct. A microprocessor-controlled mobile receiving and direction-finding system working according to this principle has been developed and used in the Federal Republic of Germany since 1980 [CCIR, 1982-86a].

3.2 *Doppler direction finder*

Mobile direction finders using the Doppler principle in the frequency range 1 to 30 MHz and the VHF-UHF ranges 20 to 500 MHz and 500 to 1000 MHz have been developed and tested in the Federal Republic of Germany. They have produced good bearing results in practical conditions. These mobile direction finders use virtually the same instrumentation as their stationary, Doppler direction-finder counterparts. One Doppler DF system for the HF range works with a crossed-loop antenna array and uses a new kind of phase-modulation principle. The VHF/UHF equipment is a Doppler direction finder with two antenna systems, one for 20 to 500 MHz and another for 500 to 1000 MHz. The antenna systems are accommodated in compact, suitcase-like radomes. In the VHF/UHF ranges in particular, additionaly bearing errors caused by the vehicle body work are astonishingly low, with this antenna system. The average system bearing accuracy is within 2 to 3° at an obstacle-free calibration site. This high accuracy is, of course, not achieved in congested or disturbed areas, due to reflections.

3.3 *Adcock direction finder*

A newly developed direction finder with 4 pairs of Adcock antennas for the frequency range of 25-1000 MHz is in operational use in Japan. This equipment has no sense-antenna and employs an electronic goniometer which consists of balanced-modulators switched with a 135 Hz square wave signal in place of a mechanically rotating goniometer in order to get the very wide frequency coverage required. It also utilizes a microcomputer to make accurate, automatic and fast measurements. The average system bearing accuracy is around ± 3°, and it has the advantage that it can be used with almost any receiver which covers the required frequency band.

3.4 *Portable equipment*

Besides mobile direction finders accommodated in vehicles, easily portable and battery-operated devices are required for locating signal sources within buildings or other places that are not easily accessible. A portable direction-finding equipment to be used in conjunction with a manual DF antenna is available [CCIR, 1982-86b]. The antenna consists of a pistol-like handle with integrated low-noise antenna electronics and indicator. This handle can be equipped with three antenna modules for the sub-ranges 20-200 MHz, 200-500 MHz and 500-1000 MHz operating as loaded loop antennas and featuring a cardioid radiation pattern which is practically frequency-independent. Maximum or minimum values are indicated on an extended S-meter or by means of an acoustic tone, the pitch of which varies with the signal level.

Furthermore, a portable/mobile direction finder similar to the Adcock system has been developed in Japan, which covers a frequency range 25 MHz - 1 000 MHz employing three antennas with subranges 25 MHz - 150 MHz, 150 MHz - 500 MHz and 500 MHz - 1 000 MHz.

Each antenna consists of orthogonally arranged two butterfly shaped planar elements with a cardioidal directivity, the height of which is 3/4" above the ground with a surge resistor. This configuration yields a cardioidal field pattern, i.e. to be overlapped with a figure eight pattern and sense determining non-directive pattern, enabling it to operate with travelling wave mode.

The output of each element is switched to another in turn and the envelope of the resultant signal is processed to yield the bearing.

The receiving part and DF processing part are assembled in a carrying case, and the measured bearing is indicated by 8 LEDs circularly arranged with 45° step. Therefore, it is used both for bearing measurement and for homing. Its rugged structure and small size with low profile make it suitable for either portable or mobile use even while walking or driving, and allows camouflaged operation especially when the antenna is embedded in the car-roof.

4. VHF and UHF direction-finding techniques

Direction finding in the VHF and UHF bands is a desirable capability for many of the same reasons as in the lower frequency bands, e.g., location of transports in distress, resolution of harmful interference, identification of unauthorized operations, etc. Propagation conditions in these bands usually limit reception to about line-of-sight, requiring that the direction finder enter the local area of the transmitter to obtain bearings. The equipment is thus constrained by circumstances, to be either mobile or portable. Ease of transporting the equipment, speed with which it can be placed in service, the nature and operating requirements of the equipment, and the characteristics of the signal of interest will often determine whether a portable or mobile unit will be most suitable for a particular task.

Two basically different direction-finding strategies exist, stand-off bearings and homing, each of which has its particular advantages.

4.1 *"Stand-off" plotted bearings*

This strategy is generally characterized by the availability of accurate direction-finding equipment that can obtain bearings from a suitable baseline. Lines of position are then plotted on an area map to indicate the fix location. An advantage of this method is that it can be used when for operational reasons it is undesirable to enter the immediate vicinity of the transmitter. Disadvantages of "stand-off" bearings from a fixed position are that sites must be well chosen to eliminate error contributions from surrounding structures, equipment must maintain a high degree of accuracy, and multiple DF sets must be used if almost simultaneous bearings are to be obtained. The latter is imperative if the transmitter is moving. Local site DF calibration verification, and true north DF orientation are required.

The "time of arrival" technique, one variation of "stand-off" direction finding, measures the propagation time of arrival of a signal using at least three cooperating sites that also "stand-off" from the transmitter. For this system to work, the signal as received at all sites must be a direct or a ground wave, i.e., not a sky wave or reflected signal. If the locations of the cooperating reception sites are precisely known, the locus of all possible transmitter sites which can produce the measured differential times of arrival can be plotted on a map or computed for their intersection, which indicates the transmitter "fix". A characteristic of this technique is the rapid acquisition of the fix data on relatively short transmissions. Accuracy of time of arrival systems is generally excellent, compared for example, to other types of stand-off plotted bearings. Disadvantages are the necessity for multiple, cooperating sites, the precision required of the time delay measurement equipment, and the extensive computations required for correlating the received signal components and determining the intersection of the loci. This equipment is also expensive.

VHF/UHF direction finders based on the Doppler principle and working in the frequency range 20 to 1000 MHz are available. The Doppler principle enables wide-aperture antennas to be implemented, and these direction finders can consequently be used, without any major restrictions, in areas where strong reflections occur, e.g. in towns and cities.

This kind of direction finder may be operated automatically, achieving high accuracy depending on its location and the presence of reflections, and can be remotely controlled in all its functions, meaning that a direct link with a computer system can be established. The Doppler principle also obviates the need for special cables, attuned in amplitude and phase between the antenna system and the instrumentation.

4.2 *Mobile homing system*

This second strategy makes use of mobile equipment (usually in an automobile or aircraft) of low complexity and relatively less accuracy which can be rapidly moved into the vicinity of the transmitter by following successive approximations of the true line of bearing. As the distance to the transmitter is decreased, the magnitude of the absolute error also decreases to a manageable value even if the relative error in degrees remains constant. Advantages of this strategy include equipment which is lower in cost and complexity, the fact that adequate results can usually be obtained in a reasonable time, the additional fact that site error problems are much less important because the unit is mobile and error contributions can be "averaged" over a long baseline, and finally, the DF information can be derived from sub-audible and locally induced homing system modulation which does not disrupt the information being transmitted. Widely available homing systems have been found to operate either in a "homing" configuration without the ability to clearly hear normal modulation components, or in an "audible" configuration when the homing circuit is switched off. This limitation is inherent with design of equipment that superimposes audible, local modulation proportional to the direction of signal arrival. Only one DF homing system is required to locate a signal source, although simultaneous use of others may speed transmitter location. Disadvantages are that inconspicuousness of stand-off capability is sacrificed, and multiple transmissions over a reasonable time period are required for adequate location of the transmitter.

5. Accuracy improvement using a computer

It is possible to apply a computer program to improve the accuracy of a radio direction finder "fix" at any point on the Earth. Programs are available to enable a modern digital computer with permanently stored memory capability to compute a "fix" in a few seconds, using reported bearings from many different fixed locations of direction-finding stations.

The mathematics required for such a program are well known and established, consisting of consecutive solutions of spheroidal triangles. The automatic solution is rendered somewhat complicated by ambiguity, since two intersecting points are to be found on the globe for any two great circles, assuming the circles are not coincident. This ambiguity is resolved in the computer program by making assumptions from available information. In the absence of propagation data to the contrary, the assumption of great circle lines of bearings is also made.

There are many ways and techniques of evaluating the most likely point or area of a transmission source taking into account factors such as:

- propagation conditions;
- non-spheroidal Earth;
- statistical averaging;
- particular performance of a direction-finder including possible inherent errors;
- experience of the operator taking the bearing, etc.

It is possible to make use of a computer to compensate automatically for some of these factors.

Use of a computer exclusively for this purpose sometimes leads to limitations in accuracy by eliminating a human element from the determination. Depending upon their ability and experience, operators who plot bearings manually, are, from their acquaintance with maps, generally aware of population concentrations or other pertinent geographical features and they may possibly also have a personal knowledge of certain transmitter locations.

Provided ample observation time is available, a small computer can be programmed to regularly sample the bearings of a given signal in order to accurately evaluate the statistical distribution of sky-wave propagation conditions. Reduction of data gathered over a significant period of time may provide more accurate predictions of the direction of arrival when compared to instantaneous observations. It can also provide a classification of the bearing according to the magnitude of the standard deviation of the data observed. The computer system can be programmed to evaluate the bearing data to any desired degree of resolution within the capability of the system. Resolution and accuracy are therefore more predictable and are not dependent on the experience of the operator. More accurate evaluation and classification of bearing data from the various stations participating can in turn lead to significant reductions in the size of the most probable area of location of the transmitter. For more information, refer to Chapter 12, Handbook for Monitoring Stations.

It may be desirable to use the computer program to supplement or to refine manually-plotted bearings so as to provide additional knowledge which will enable the operators to make the decisions as to the most likely location, on the basis of an accumulation of known facts.

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REPORT 979-1

MONITORING TECHNIQUES FOR DETECTING AND LOCATING SOURCES OF INTERFERENCE LIKELY TO DISTURB THE SATELLITE EMERGENCY POSITION-INDICATING RADIO BEACON SYSTEM OF THE MOBILE-SATELLITE (EARTH-TO-SPACE) SERVICE IN THE BAND 406 - 406.1 MHz

(Question 63/1)

(1986-1990)

1. General

The COSPAS-SARSAT search and rescue programme is the fruit of international cooperation between the SARSAT partners (United States, Canada and France) and the USSR, whose COSPAS system is compatible with the SARSAT system.

Its dual purpose is to:

- facilitate search and rescue operations through satellite locating of emergency position-indicating radio beacons (EPIRBs) operating on 121.5 and 243 MHz;
- introduce a new emergency detection identification and location system using radio beacons operating on 406 MHz.

During the engineering evaluation phase, the COSPAS-1, 2 and 3 and SARSAT-1 and 2 (NOAA-8 and 9) satellites revealed the presence in certain regions of the globe of many interference sources in the band 406 - 406.1 MHz (see Figure 1).

The presence of these sources, even for transmitted powers of less than 5 W, leads to significant deterioration in the system's performance. The purpose of the two methods outlined in this report is therefore to detect and localize interference sources with an intent to prevent in-band interference.

Another method that can also be used is to derive Doppler and Doppler-rate measurements from interference signals observed by a satellite.

2. Principle of the 406 MHz system

Every 50 s, the 406 MHz satellite EPIRBs transmit by pulses a digital message that lasts 0.5 s. The messages are transmitted independently of each other on a random time basis.

The frequency of every received message is measured by the receiver processor aboard the satellites and the information is stored to provide complete global coverage.

The position of a satellite EPIRB can be calculated by using the time/frequency pairs which translate variations in the Doppler effect as a satellite passes overhead.

The satellite receiver processor has an input bandwidth of 25 kHz, centred on 406.025 MHz.

3. Processing of unauthorized 406 MHz signals

3.1 COSPAS-1 satellite

In the past, when a source of interference was detected by the COSPAS-1 satellite, the frequency measurement and the date in the information retransmitted by the on-board processor was usually correct. The remaining data consisted of a random succession of bits. These "pseudo-messages" did not appear in a uniform manner during the orbit but in a characteristic fashion when the satellite passed over certain regions of the globe (see Figure 1).

3.1.1 Method of location

A study of the "pseudo-messages" received from COSPAS-1, according to their time-frequency pairs, indicated the presence of coherent signals with clear Doppler curves. Taking this characteristic into account, the Centre National d'Etudes Spatiales (CNES) developed a data processing chain for locating certain unauthorized transmitters. Similar algorithms have been developed by other SARSAT partners. The principle was the same as that employed for locating the satellite EPIRBs, i.e. least squares fit of a theoretical Doppler curve to the measurements. Unlike the processing of satellite EPIRB emissions, the grouping of messages from one and the same source could not be effected on the basis of the satellite EPIRB's identification but had to rely on shape-recognition techniques. Although COSPAS-1 is no longer operative, future SARSAT spacecraft using a commandable function, would be capable of performing measurements similar to those formerly performed only by COSPAS-1.

3.1.2 Validation and use of the method

The method described above has been used since 1983 and has led to the location of about 30 sources of interference. Wherever possible, the reliability of these locations has been checked through bilateral contacts. As a result, 17 sources were eliminated in southern Europe and Africa, as well as two sources located in North America.

3.1.3 Limitations of the method

Work carried out in 1983-1984 showed that the locations obtained were perfectly reliable and that satisfactory correlation always existed between the results of the satellite location method and the real locations.

However, it was not possible to locate all the sources by this method. Only sources of interference emitting a stable frequency signal could be located.

3.2 SARSAT repeater (NOAA-9, 10 and 11 and subsequently NOAA-12 and 13)

The NOAA-9, 10 and 11 satellites have a repeater for retransmitting emissions from Earth received in the band 406 - 406.1 MHz.

The time/frequency pairs of emissions can then be measured at an earth station by applying principles similar to those of the on-board equipment. The coverage area for processing unauthorized emissions is thus limited to the mutual earth station-satellite-interference source visibility area (see Figure 2). At present, France, Canada and the United States have units for processing relayed emissions.

In the United States, a station which detects interference sources in North America, has been operational since 1986; and since May 1987, the United States, in response to Resolution No. 205 of WARC MOB-83, has been providing the IFRB with monthly reports of detected interferers. These reports include a listing (typically 30 to 40) of the locations of interferers that have been detected by 50 or more satellite passes. Since initiation of the monthly reports, a total of 14 interfering stations have been turned off in the western hemisphere (as of April 1989). In the future, other United States stations will also be equipped to receive and process interference data from the SARSAT repeaters, thus expanding the interference monitoring coverage area.

Similarly, since January 1987, France operates a station located in Toulouse that detects interference sources in Europe, North Africa, and the Middle East. Monthly reports are also sent to the IFRB in response to Resolution No. 205 of WARC MOB-83. As a result of IFRB actions and bilateral contacts, 12 interfering stations have been turned off. This activity will be continued in the future because the permanent monitoring of the 406 - 406.1 MHz band is considered in France to be an integral part of the COSPAS-SARSAT system operation.

3.3 Auto-correlative method to derive Doppler and Doppler-rate measurements

The Doppler frequency shift and its change with acceleration (Doppler-rate) can be derived by auto-correlative processing of an interference radio signal observed by an Earth-orbiting satellite. These measurements can be obtained even in the case of fairly poor S/N ratios and can be used to locate the source [Shiomi and Takahashi, 1985].

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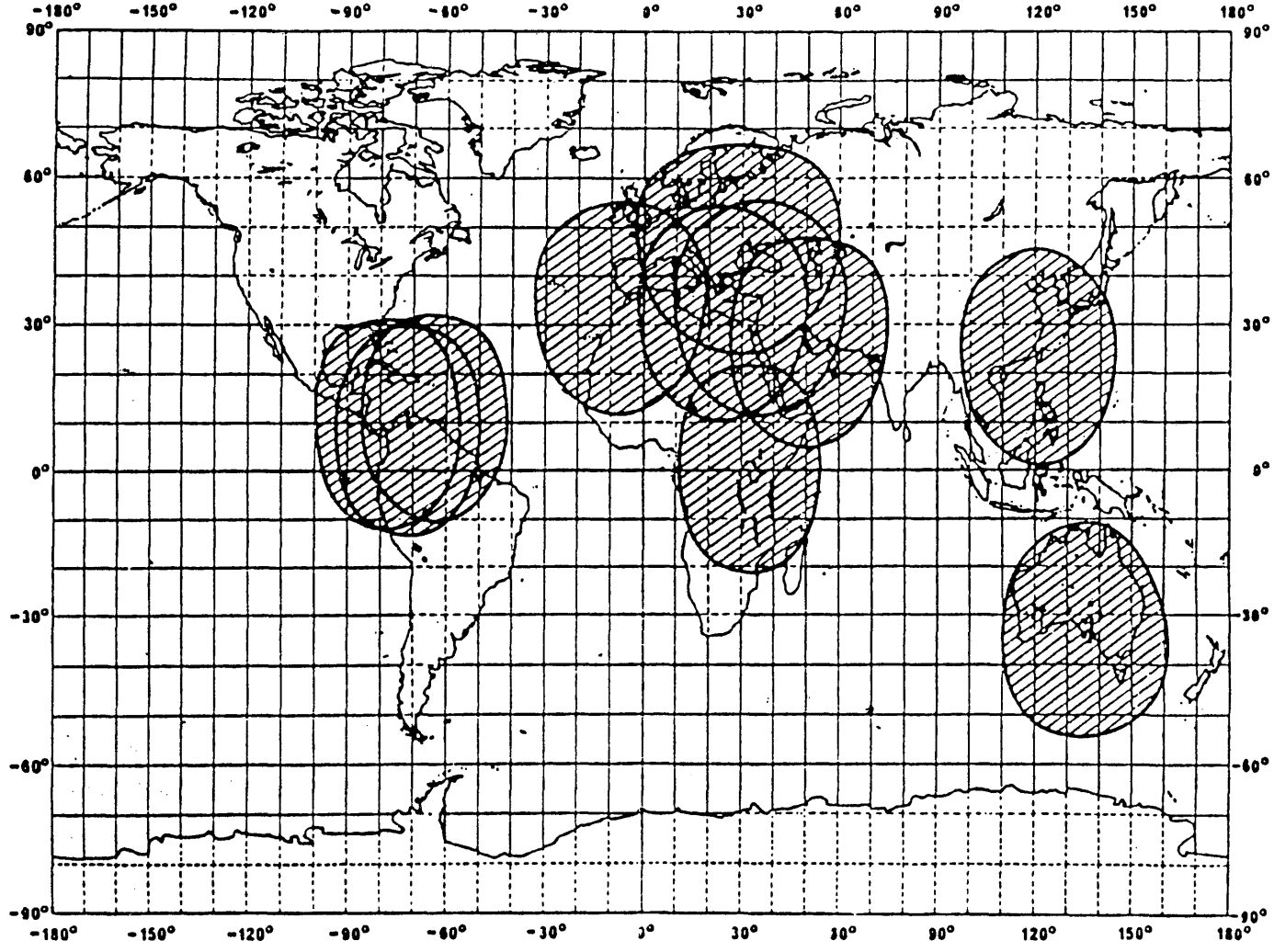


Figure 1 - **Primary Interference area of the COSPAS-SARSAT system measured by COSPAS I as of 1987 (See Section 3.1.1 of this Report) Frequency 406.010-106.040 MHz**

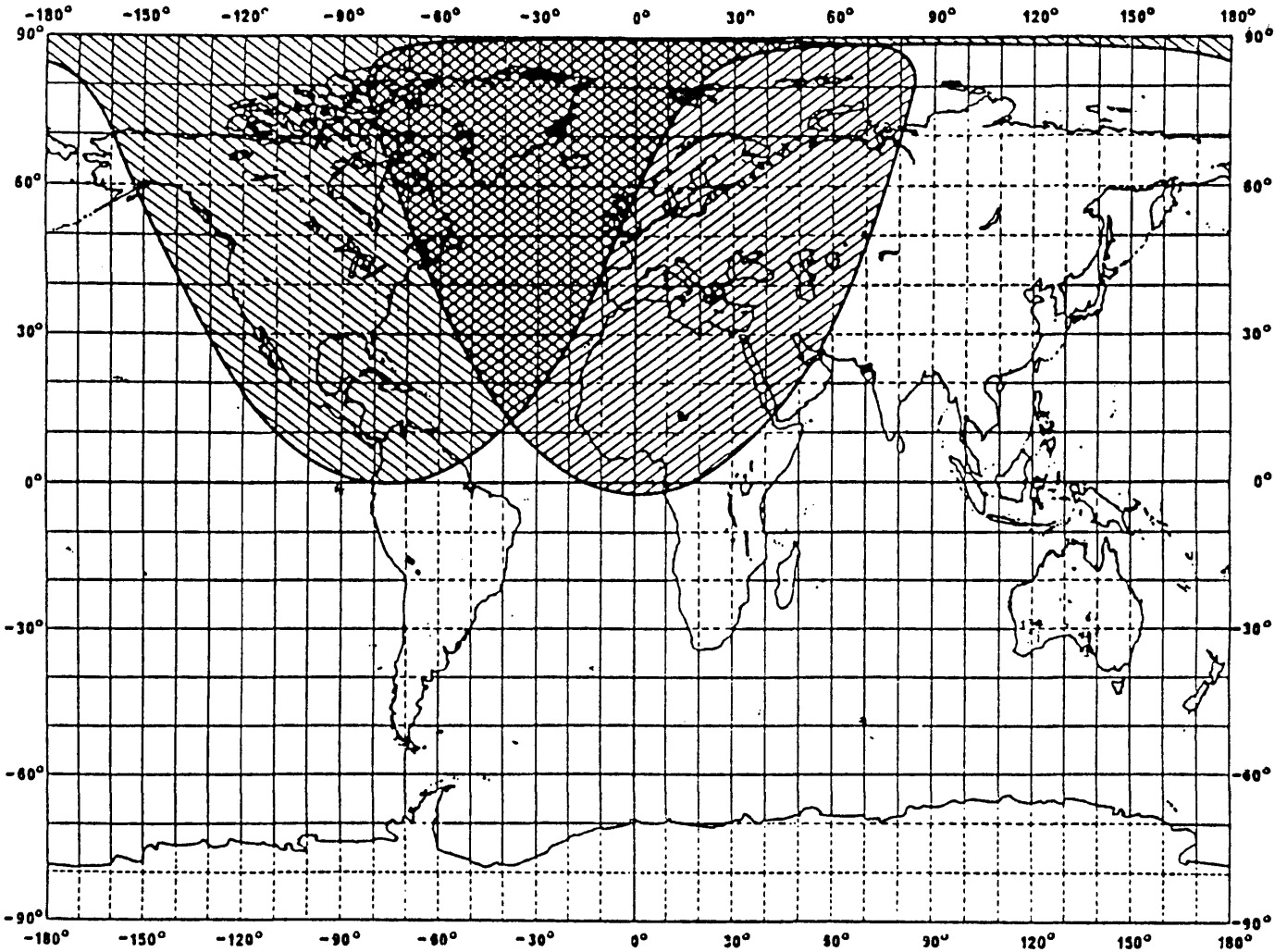


Figure 2 - Coverage area of COSPAS-SARSAT interference monitoring stations as of 1988

REPORT 835-2

THE AUTOMATIC ACQUISITION OF RADIO-FREQUENCY SPECTRUM OCCUPANCY DATA AND ITS USE FOR SPECTRUM MANAGEMENT PURPOSES

(Question 29/1)

(1982-1986-1990)

1. Introduction

Automatic monitoring of the occupancy of the radio spectrum has previously been reported by the United States of America [CCIR, 1966-69], Canada [CCIR, 1974-78] and Australia [CCIR, 1982-86].

In designing an automated system to gather occupancy data for use in spectrum management, the first problem to be faced is what parameters are to be measured and how often measurements have to be taken to ensure the data is statistically significant.

Spectrum occupancy data can provide information on the actual level of usage of individual frequency channels, e.g. in land mobile bands, such that additional mobiles could be licensed for existing systems or new systems added up to the maximum channel loadings compatible with acceptable levels of service. Additionally, it can provide information on the average level of usage of these bands so that the reserve capacity of the spectrum can be estimated. Such an estimate can be used for the planning of additional spectrum allocations and as a means of assessing the impact of new technologies on spectrum usage.

A spectrum occupancy system consists of three essential parts:

- the monitoring equipment required to collect the raw statistical data on the levels of signals on individual frequency channels;
- the software necessary to extract statistically meaningful occupancy data from the monitoring data;
- the software required to use the spectrum occupancy data in the frequency assignment process, and to provide global trend data on the level of spectrum usage in the land mobile bands.

This Report focuses on three main topics:

- acquisition of spectrum occupancy data and statistical considerations;
- utilization of spectrum occupancy data for frequency assignment.

The Annex I gives some practical results and briefly describes a way of monitoring trunked systems.

2. Acquisition of spectrum occupancy data and statistical considerations

A radio message is defined, for the purposes of this analysis, as one complete exchange of information between a base station and its associated mobile or mobiles. Therefore, it consists of a series of base station transmissions and mobile transmissions separated by transmission gaps.

Since a channel is normally unavailable to another user during the entire length of such a message, the occupancy that characterizes the level of utilization of a channel is the message occupancy. This is defined as the percentage of the time that a channel is occupied by such messages, over a given monitoring period.

The message occupancy O_m of a particular channel is the sum of the base-station transmission occupancy O_b, mobile transmission occupancy O_{mob} and the transmission gaps O_g. A breakpoint T_B is used to differentiate between transmission gaps and message gaps. Any gap less than T_B s is considered a transmission gap and therefore part of an on-going message.

$$O_m = O_b + O_{mob} + O_g \tag{1}$$

Proper spectrum management decisions can only be made if it is known what degree of confidence can be placed in any spectrum occupancy estimate. In designing measurement procedures, consideration must be given to the time required to obtain sufficient data for estimating the occupancy to a given accuracy and confidence level. Because of the probabilistic nature of the time and the amplitude structure of signals appearing in a communications channel, the statistical nature of the measured parameters must be considered in estimating occupancy.

In considering statistical factors, it is useful to define channel transmission occupancy as a two-state random process. In this process, the first state is labelled occupied and is defined as an event such that, during observation, the signal strength of a given emission at a monitoring receiver is above a given threshold. The second state is when the signal is below this threshold. This definition is depicted in Fig. 1 for a given channel, and for the sake of clarity, only a portion of the monitoring done during the total time T_t is shown.

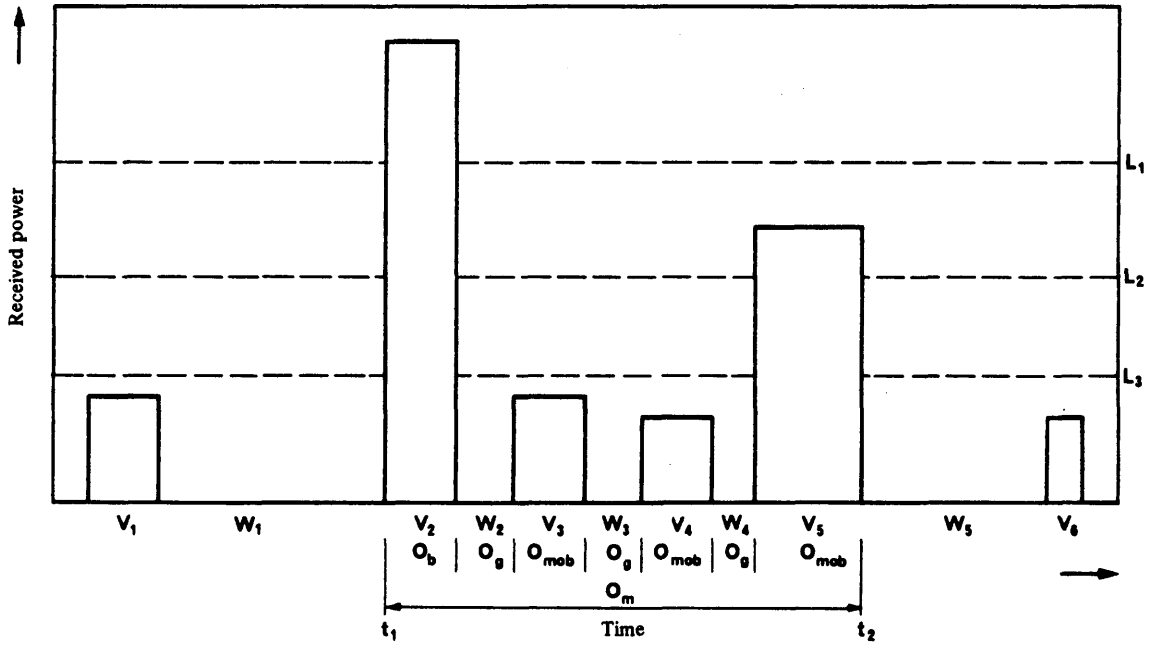


FIGURE 1 – *Received power on a channel*

V: transmission time above a given threshold L

W: length of time between transmissions (transmission gap)

T_t: total time during which measurements are made

L_1: threshold power level for channel occupancy

L_2 and L_3: arbitrary receiver powers below threshold power level L_1

The channel transmission occupancy percentage would be, in this case, the transmission time of V_2 over the T_t, the total time of observation. Either a base or mobile transmission that exceeds the L_1 threshold would be included in the calculation of the channel transmission occupancy.

In terms of the definition of equation (1), the actual message occupancy percentage (O_m) of a base station and its mobiles, used in this example, consists of the times of the base transmission V_2 (O_b), three mobile transmissions V_3, V_4 and V_5 (O_{mob}) and three transmission gaps W_2, W_3 and W_4 (O_g) over the time period $t_2 - t_1$ observed during some total period of monitoring T_t. The values W_1 and W_5 may be gaps between messages or part of such gaps. Also, V_1 and V_6 may be part of the area noise or part of other messages.

It is instructive to consider the effect of both independent sampling, i.e. assumed instantaneous measurements having no relation to each other and of dependent sampling, i.e. sampling of finite messages with fixed sampling time intervals, on the monitoring time required to achieve the desired relative accuracy and confidence level.

By using two limit theorems and considering the use of independent sampling, equation (2) can be derived for the calculation of the number of trials M needed to reach a required confidence level C, for a given occupancy, p.

$$M = \frac{\log [1 - C]}{\log [1 - p]} \tag{2}$$

Table I gives the number of independent samples required at various occupancy levels to achieve a 99% confidence level.

TABLE I – *Number of required measurements to achieve a 99% confidence level at various occupancies*

Occupancy ($p \times 100$)%	M required measurements
50	7
20	21
10	44
1	459
0.1	4 603
0.01	46 050

For example, Table I indicates that for 4603 independent measurements of a given channel and for which no signals are detected, then one is 99% confident that occupancy is below 0.1%.

Although of interest, these considerations are not applicable to actual monitoring since messages have finite lengths and the rate of sampling can be variable. In this case, a statistical scheme can be developed for dependent sampling by a first order Markov chain. The sampling time to achieve a given confidence level is then a function of the transmission length statistics and the sampling rate.

To examine the effect that independent and dependent sampling has on the number of measurements required to achieve a desired relative accuracy and confidence level, let us consider the case of a channel that has forty transmissions per hour and an average transmission length of 6 s, which is a 6.67% transmission occupancy. To achieve a value of p to a relative accuracy of ± 10% and a 95% confidence level, the number of observations required assuming independent sampling would be 5850. For dependent sampling where a measurement is taken every 4 s and the supposed transmission lengths are exponentially distributed with a mean value of 6 s, the number of measurements to obtain the same relative accuracy and confidence level is 18 166.

Table II compares the number of measurements required to achieve a relative accuracy of ± 10% and a 95% confidence level for the above described conditions and different occupancy percentages for both independent and dependent samples.

TABLE II – *Number of dependent and independent samples required to achieve ± 10% relative accuracy and a 95% confidence level at various occupancy percentages*

Occupancy (%)	No. of required independent samples	No. of required dependent samples	Required hours of sampling
6.67	5850	18 166	20.18
10	3900	12 120	13.47
15	2600	8 080	8.98
20	1950	6 060	6.73
30	1300	4 040	4.49
40	975	3 030	3.37
50	780	2 424	2.69
60	650	2 020	2.24
70	557	1 731	1.92
80	488	1 515	1.68
90	433	1 346	1.49
100	390	1 212	1.35

The required hours of sampling assumes a 45 s sampling period. For channels with 20% occupancy, it would be possible to achieve the ± 10% relative accuracy at a 95% confidence level during a 6.73 h monitoring period. If the receiver measures a channel every 0.002 s, then 2000 channels could be monitored.

In designing a measurement programme, it is important to consider the statistical implications of demanding high accuracy and confidence levels, since the measurement times needed to obtain these values quickly become prohibitively long.

The other basic consideration for monitoring is what is to be measured. In general, the base-station transmissions received by the monitoring equipment are easily detected signals, while the mobile signals have lower and varying signal levels that are often buried under ambient radio noise.

Two methods for the derivation of occupancy values from the from the monitoring data have been investigated. In the threshold occupancy method, a threshold is set that eliminates the contribution of radio noise and the occupancy above threshold O_t is multiplied by a factor K_t that takes into account the transmission gaps as well as the undetected mobile signals:

$$O_m = K_t\, O_t \tag{3}$$

The drawback of this method is that the amount of undetected mobile signals varies from one radio system to another, depending on the distance to the monitoring station.

In the base-station occupancy method, this uncertainty is removed when the occupancy is derived from the base-station signals only, since in this case, the contribution of all mobile signals is systematically removed on all channels:

$$O_m = K_b\, O_b \tag{4}$$

where K_b is the factor derived for the base-station occupancy method.

For unshared radio channels, the experimental values for average peak hour occupancy have indicated that the average value of the factor K_b is equal to 2.74 and the average user dependent variation on individual channels is 18% which is considered sufficiently reliable for frequency assignment purposes.

It was also found that the threshold occupancy method is much less accurate since the factor K_t varies from a minimum average of 1.52 when most mobile signals are above threshold, to a maximum average of 2.74 when all mobile signals lie below threshold. The factors K_t and K_b have been estimated, using linear regression with a breakpoint T_b equal to 6 s.

An example of the problem of using the factor K_t can be seen in Fig. 1 where, if the threshold was set at L_2, only transmission V_2 and V_5 would be detected, whereas if the threshold was set at L_1, only V_2 would be detected. The actual monitoring site selected can therefore easily alter the number of detectable mobile signals.

This example indicates the advantage of the base-station occupancy method, where only the base-station signals are considered. Furthermore, this method also allows the selection of monitoring sites in suburban areas, where monitoring is more reliable due to the absence of strong unwanted radio signals that could cause desensitization of intermodulation.

3. Accuracy of the message occupancy estimates using the factor K_t or K_b [De Couvreur *et al.*, 1981]

In the original Canadian experiment, Burke and Coyne [1980] used a breakpoint T_b to differentiate between transmission gaps and message gaps. Any gap length of less than T_b seconds was considered a transmission gap and part of an on-going message. Using this breakpoint, the message occupancy could then be determined.

Experiments by the United States Administration have indicated that, on average, a value of T_b equal to 7 s is appropriate to distinguish transmission gaps from message gaps, based on subjective audio tests of 21 single user channels [Dayharsh *et al.*, 1974].

Until the results of further on-going research are available, it will be considered that the most suitable average value of T_b should be in the range of 5-7 s on shared channels.

The results of Canadian experiments designed to determine K_b are based on a sample size of 52 unshared simplex channels. The K_b factors used to obtain average daily message occupancy and average peak hour message occupancy were compared together. It was found that the value of the factor K_b, for the peak hour message occupancy was smaller and more accurate, in terms of smaller variations from one channel to another, for the same types of users and also in terms of smaller maximum errors regardless of the type of user. Therefore, it was concluded that the factor K_b, for the average peak hour message occupancy was preferable, as compared to that for average daily message occupancy for evaluating the channel message occupancy in the shared frequency assignment process used for obtaining an improved spectrum utilization.

An investigation of the data for the 52 channels indicated that the K_b factor was a function of the different types of users. The method of grouping the users into categories will need to be further developed in order to achieve the best possible estimation accuracy with a statistical model based on K_b. Another investigation for average peak hour message occupancy indicated that the average variation of the probabilistic mean value K_b was -16% to $+18\%$. Part of the variation could have been caused by imperfect mobile coverage since there were no audio monitoring data that could be used to check the experiment.

One must also consider that in practice, for K_b the actual value of T_b will be in the range of 5-7 s due to user dependent variations, i.e. the busier the channel, the shorter the gap between messages and *vice versa*. In this case, as shown in Fig. 2, a combined average error of $\pm$ 19% will result. For cases where the mobile coverage is known to be very good, the average error of the K_t factor is $+8\%$ and -7% and when the user dependent variation in T_b is considered, one obtains a combined average error of $\pm$ 9%. However, in the case of Canada, this is seldom experienced.

Even if the intrinsic variations of the K_b factor are greater than those of the K_t factor, the potential error caused by the unknown mobile coverage makes the base-station method superior.

In the Canadian situation, the desired loading criteria for effective spectrum utilization corresponds to an average peak hour message occupancy of 75% for multi-user channels.

However, until the accuracy of the factor K_b can be improved, a loading criteria of 60% is recommended to avoid overloading. The relative accuracy of the Canadian monitoring at this level of occupancy introduces a small error caused by the uncertainty in K_b.

4. Utilization of occupancy data for frequency assignments

Three methods for the utilization of monitoring information in the frequency assigment process have been investigated.

The direct method of using monitoring information (see experiments described in Annex I) uses the message occupancy at a monitoring site as an approximation of the message occupancy that would be observed at the site of a proposed station. This approximation has two main drawbacks. Firstly, the quality of the approximation is affected by the distance between the monitoring site and the proposed station site, which can be as large as 25 km in practice. Secondly, complex calculations are involved in the assessment of the incremental message occupancy perceived by an existing receiver as a result of a new shared frequency assignment.

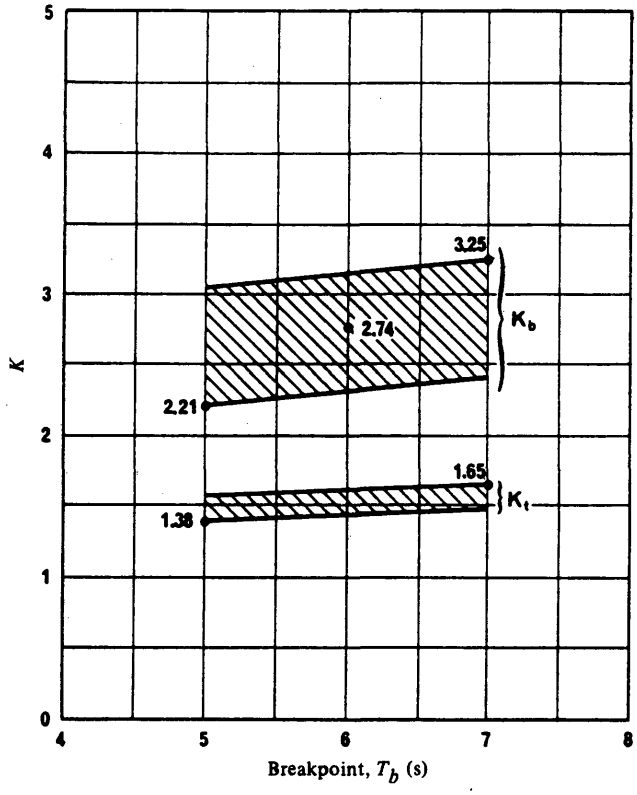


FIGURE 2 – *Variation of the average peak hour occupancy factors K_b and K_t as a function of the breakpoint time T_b for all users*

The indirect method uses monitoring information to compute the message occupancy of each station which is in the vicinity and shares the channel. This method is more complex in that it requires the evaluation of data from all stations contributing to the occupancy at a proposed site. This additional complexity is offset by a higher accuracy, in particular for the evaluation of the effect of a new station on its environment in terms of incremental message occupancy.

The third method is to evaluate message occupancy by relating it to the number of mobiles for each user category, by using a statistical model. The computations are similar to those of the indirect method but tend to be more numerous. The validity of the statistical model depends on the availability of a large sample of monitored data in order to assess user-dependent patterns of spectrum usage. Although monitoring would be less exhaustive and less frequent with this method, further monitoring would still be required to keep the model up-to-date as spectrum usage patterns evolve with the introduction of new technologies such as digital signalling.

5. Conclusions

Current automated monitoring systems are being used by the Australian and Canadian Administrations to gather occupancy information on a routine basis. In designing these systems, the consideration of the statistical nature of monitoring has been paramount in achieving an acceptable relative accuracy and a given confidence level in a reasonable monitoring time.

It has been the experience of the Canadian Administration that the base station occupancy method provides a more reliable means of determining occupancy than the threshold occupancy method.

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[1982-86]: 1/37 (Australia).

ANNEX I

RESULTS OF THE EXPERIMENTS CONDUCTED IN CANADA

1. Utilization of occupancy data for frequency assignments

The direct method was used during extensive field trials held in Montreal in 1979, where the complete land mobile computer-assisted licensing system was tested. This method has been in use by the Canadian Administration since then.

2. Level of spectrum usage

The results given in Table III are expressed in terms of average daily message occupancy, namely the message occupancy during the period 0800 h to 1700 h, averaged over five consecutive days.

TABLE III – *Spectrum usage in the unrestricted channels capable of being shared*

Band (MHz)	Percentage of channels, the occupancy of which is less than		Average occupancy (%)
	30%	40%	
27- 50	84	90.2	20.3
138-174	78	84	19.3
410-470	88.4	92	15.2

The results indicate the availability of a significant reserve of spectrum capacity for shared frequency assignments, which can be exploited with the knowledge of the occupancy on individual channels provided by the spectrum occupancy sub-system.

The reserve of spectrum capacity is readily evaluated for the channels which are used for non-emergency service and could be shared. In this case, an acceptable level of service is maintained as long as the occupancy during congested hours does not exceed 70% on a shared channel. Assuming Poisson statistics, this means that in the case of a channel shared by five users, the waiting time is three times the message length. This corresponds to an average daily occupancy of 40%, as shown by a detailed analysis of short-interval occupancy patterns.

Consequently, all channels that are operating at present at an average daily occupancy below 40% could be loaded up to that level by further shared-frequency assignments.

3. Measurement of traffic characteristics in trunked mobile radio systems

In a conventional dispatch mobile radio system, users are assigned individual frequencies. In a trunked radio system, Report 741, a set of frequencies are grouped together and dynamically assigned on demand to individual users. Trunked systems hold the promise of increased spectrum efficiency over conventional systems, better grade of service and good privacy conditions. In order to assist spectrum planning a monitoring system was developed to obtain occupancy data for trunked systems on individual frequencies and at the same time provide a consolidated measure of overall trunked system utilization. The indicator selected is the frection of the time on an hourly basis that the trunked system is busy and not available to the user; that is, the percentage of time all individual frequencies are busy.

The system used is a mobile spectrum monitoring system which is also used to measure occupancy data in the land mobile frequency bands for individual frequencies (Vaccani, 1985). The system has been augmented using a special software algorithm to measure the traffic characteristics of trunked systems (Vaccani, 1987). The system consists of a vehicle, associated antennae sub-system, and a custom designed computer controled receiver.

The major characteristics of the system are:

- Measurement of signal amplitude of each individual frequency, over 1 or 2 msec intervals. This includes the complete cycle time of the computer and receiver. Status data of active (above threshold) or non-active channels, are stored by the computer.

- The storage capacity of the system can accomodate more than one week of data.

- Data are processed in the computer memory on an hourly basis prior to storage to disk. This minimizes the over-head required to write to disk and facilitates the production of reports.

- Up to 1000 frequencies for trunked systems can be selected with up to 10 individual trunked systems.

For each trunked system the following measurements and parameters are provided on an hourly basis, based on a user provided threshold level. The threshold level is used to determine if an individual frequency is active or non-active.

- Percentage busy time is the percentage of time that all frequencies making up a trunking system are active;

- Longest busy period is the longest time period in seconds that all frequencies are active;

- Distribution of busy periods gives the distribution of the busy periods in 10 steps ranging from 2 msec up to greater than 80 sec.

For each individual frequency the following measurements are provided:

- Percentage busy time is the percentage of time that an individual frequency is active;

- Number of transitions is the total number of changes from active to non-active status for each frequency.

This monitoring system has been used effectively in Canada and has provided useful information for loading trunked systems.

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VACCANI, P. (March, 1987) A method of obtaining occupancy data for trunking systems in the radio frequency spectrum. 7th International Zurich Symposium and Technical Exhibition on Electromagnetic Compatibility, Zurich, Switzerland.

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REPORT 279-4

VISUAL MONITORING OF THE RADIO-FREQUENCY SPECTRUM

(Question 30/1)

(1963-1970-1978-1986-1990)

1. Technological advances in the capabilities of wideband frequency domain spectrum analysers, having large dynamic ranges of displayed signal amplitude, have steadily resulted in increasing importance of visual monitoring operations. Spectrum analysis provides a means for rapidly recognizing and classifying various types of complex emissions. Visual monitoring can increase efficiency of monitoring operations by defining areas of current activity worthy of further examination. Service to users of radio spectrum is speeded by use of visual display techniques to synchronize the occurrence of interference with the activity of the emissions causing the interference.

1.1 Analysis of radio spectrum in the frequency domain is provided by two broad classes of equipment.

1.1.1 Wideband analysers are available which may display selectable portions of the entire radio frequency spectrum. However it should be noted that the usable frequency range will depend upon the frequency range of the receiving antennas.

1.1.2 Panoramic display modules, connected to receiver intermediate frequency outputs, are available which show a limited portion of spectrum surrounding the tuned receiver frequency.

2. Spectrum analysers have been developed into automatic equipment and are nowadays very useful instruments for visual monitoring.

The application of microprocessor technology and digital signal processing techniques to spectrum analyser instrumentation has improved both the performance characteristics and ease of use of these instruments.

It should however be noted that analogue analysers have advantages in some cases, compared with digital equipment.

2.1 **Performance characteristics of a microprocessor controlled spectrum analyser might typically include the following specifications: a frequency range for any single sweep of up to 2.5 GHz or if an internal harmonic mixer is available up to 60 GHz and up to 300 GHz, if external mixers are used;**

measurement range from -138 to +30 dBm with an 90 dB dynamic range displayed logarithmically; automated calibration to reduce measurement uncertainty of the order of ± 2 dB; display of frequency of

cursor position with accuracy of 11 Hz ——————————; display of either absolute or relative value of components identified by cursor with allowance made for attenuator settings; automatic selection of sweep time/bandwidth to produce optimum display for frequency band selected; digital storage of display information for long term stability and elimination of display flicker.

2.2 Incorporation of a microprocessor in the spectrum analyser allows a variety of new procedures to be used which can improve both the ease and efficiency of making measurements with the equipment.

2.2.1 *Spectral averaging*

Making bandwidth measurements with conventional instrumemts on signals with time-varying components previously required operators to either "average" the occupied bandwidth mentally or to build up a pattern on a storage oscilloscope. The automated spectrum analyser typically incorporates an algorithm which can average a specified number of sweeps and display a composite bandwidth for analysis.

2.2.2 *Multiple display capability*

The use of digital techniques for presentation of the display allows a number of useful manipulations of the displayed information to be carried out. A signal may be captured and stored for later comparison with a standard signal. A spectral "mask" which indicates maximum allowable limits for spurious emissions as a function of offset frequency from the assigned frequency may be stored and recalled at will to compare with the signal under analysis. Use of arithmetic functions can rapidly normalize a signal under analysis and indicate where limits have been exceeded.

2.2.3 *Remote interfacing*

The use of local intelligent software in the spectrum analyser facilitates incorporating it into a network of other equipment for automatic monitoring of the spectrum. Using the capabilities mentioned above, a controller can rapidly scan the spectrum, test each signal found for the presence of possible spurious emissions, and record the location and the relative level of each suspected spurious component. Performing this detailed analysis in a matter of seconds provides the possibility of greatly increasing the accuracy and efficiency of spectrum monitoring personnel.

3. In the Federal Republic of Germany, equipment combining the functions of a panoramic adapter and a spectrum analyser is available. This is effected by using a multi-purpose VHF/UHF receiver operating from 20-1000 MHz together with a panoramic adapter with a 10 × 8 cm CRT screen using 8 × 10 dB logarithmic calibration, the degree of accuracy being ± 1 dB. In addition to IF analysis with sweep widths of 20 kHz, 200 kHz and 2 MHz, representation of the spectrum over a segment of 500 MHz is also possible. For this purpose, the local oscillator (LO) of the VHF/UHF receiver is seprated from the synthesizer system and swept by the saw-tooth generator of the panoramic adapter.

A "blinking" marker shows the tuning position of the receiver. Any sector of the total range can be extended to any size and displayed on a second line. By means of the various displays available on the panoramic screen, unknown signal activities can be recognized, analysed and identified via the receiver more quickly. The spurious-free dynamic range of the panoramic adapter exceeds 70 dB. As a result, this combination results in a spectrum analyser suitable for application in dense large-signal areas. Spectrum occupancy recording is possible together with a radio-monitoring recorder (YT recorder).

4. Conclusions

Technical developments have resulted in the availability of a wide variety of equipments for visual monitoring operations. Units available for use range from relative, uncalibrated displays to direct readout, absolutely calibrated and stored displays. ——————————————————————————————. Advantages of visual spectrum monitoring have already been shown in increased service to radio users and increased efficiency of monitoring efforts. The large dynamic range of known calibration of amplitude allows ——————————— records of significant observations to be made, which contain a great amount of non-ambiguous spectrographic data.

REPORT 668-3

AUTOMATIC MONITORING AND MEASUREMENT IN THE RADIO-FREQUENCY SPECTRUM

(Question 29/1)

(1978-1982-1986-1990)

Question 29/1 seeks information on various aspects of automatic monitoring of the radio-frequency spectrum, including measuring methods and equipment for conducting such measurements and monitoring.

1. Accuracy of automatic monitoring equipment

1.1 *Determining bandwidth*

In considering the accuracy of automatic monitoring equipment in determining bandwidths of emissions, due consideration should be given to relating measurements of narrow-band emissions of the order of 2 kHz or less to the characteristics of the particular instrument in use. It has been determined, for example, that in some instruments which sweep a frequency range and resolve the bandwidth presentation by selective circuits, the shape factor of the selective circuits is such as to complicate bandwidth determinations at other than accurately known comparison levels. The variation of received signal levels due to propagation conditions, together with other factors, impose restrictions on the accuracy of indicated signal bandwidth of narrow-band signals. Consideration should therefore be given to the need for an analysis of the relative distribution of the signals level and the resolving capability of the instrument at several input levels. For signals with bandwidths of several kHz, the errors contributed by the above-mentioned equipment limitations become less important. Nevertheless, it may be desirable, in the interest of correlating data obtained with different instruments, to take into account their resolving power and the repetition rate and duration of discrete components of the observed signals.

1.2 *Determining field strength*

In determining the accuracy of field-strength measurements made by automatic devices, the polarization and directivity of the antennas are important factors. In the 3 to 30 MHz range, where signals received over sky-wave paths contain both horizontally and vertically polarized components to a significant extent, a decision must be made as to the component to be measured. Since the vertical angle of arrival is most commonly less than 45°, less error will usually be introduced by measuring the vertically-polarized component rather than the horizontally-polarized component. Where greater accuracy is required, an appropriate correction may be applied, based upon a computed value for the vertical angle of arrival of the wave front.

Measurement of the vertically-polarized component may readily be accomplished by using a vertical receiving antenna, the electrical length of which is appreciably less than one-quarter wave, for the frequency band under study. To reduce errors due to non-homogeneous earth in the vicinity of the antenna, a ground screen, together with radial ground wires extending out at least 30 m (100 ft) in all directions has been found beneficial. The level of signal input to the recorder may be measured by comparison with a calibrated standard radio-frequency signal generator. The antenna effective height is determined by comparison with measurements made with an accurate field-strength meter using an antenna of like polarization.

In evaluating the field-strength data thus obtained, due allowance must be made for errors introduced by the various instruments. Allowance must also be made for fluctuations in signal-level, since, during the sweeping process, the receiver is tuned to a particular signal for only a small proportion of the time.

2. Measurement of relative field strength

A variable attenuator is added to the receiver unit of the equipment, the attenuation is automatically increased step-by-step in succession at each specified number of times of sweeping, and the signal output is recorded. The relative field strength of each signal is determined by the attenuation inserted immediately before the recorded line fades out.

The absolute field strength is obtained as follows: the absolute value of the wanted signal among those recorded is measured by means of a second field-strength meter, and, on the basis of this value, the absolute value of other signals is derived from the relative field strength.

3. Capability of determining signal-to-noise ratio

Measurement of signal-to-noise ratio primarily involves a comparison between signal and no-signal conditions making use of the same techniques as for field-strength measurements. In this instance too, variations in signal-level due to propagation must be taken into account. In the lower portion of the HF range, the problem is somewhat complicated, especially during the warm months of the year, by the generally high level of atmospheric noise, which is subject to wide variations in level, both instantaneous and over significant periods of time. The effect of receiver bandwidth and of the bandpass characteristics will also need to be considered, so that satisfactory correlation between measurements made with different equipment may be accomplished.

4. Analysing and evaluating records

The preferred means of analysing and evaluating automatic monitoring records depends primarily upon the information desired. In some instances, it may be desirable to obtain information concerning the direction of arrival of the various signals in the frequency band under study; this would provide data relating to the determination of channel occupancy with respect to the different azimuths of origin of the emissions. For example, a long-range direction finder may be associated with the automatic recorder in such a way that the azimuth would be advanced a small increment (such as one or two degrees) at the start of each sweep of the automatic recorder. Signal nulls and maxima could readily be determined by inspection of the data. Additionally, data could be extracted from such recordings which would indicate occupancy in a given azimuth at regular intervals of time. In congested bands, the use of a directional antenna would also aid in reducing overlapping of signal traces.

Identification of the type of emission, by analysis of automatic recorder indications, is subject to shortcomings with narrow-bandwidth emissions, because of limitations in the resolving power of the instrument. However, those emissions occupying a significant bandwidth, such as multi-element tone printers, multiple-component multiplex, wide-shift simplex or multiplex and certain transient types of emission, can frequently be catalogued by inspection of recorder indications, together with precise frequency increment chart calibration.

5. Automatic analysis

The degree to which records may be analyzed by automatic means is an economic rather than a technological question. Because of the variety of data normally available from the automatic recorder, equipment for complete automatic analysis can be complicated.

6. Types of equipment

6.1 *Equipment for the automatic aural monitoring of occupancy of specific frequencies*

This refers to an apparatus for the automatic monitoring of operation and occupancy in time of a given frequency. The apparatus is used to monitor occupancy of specific frequencies (ascertaining the number of stations working on the same frequency and the occupancy rate), for making automatic recordings of the monitored transmissions for the duration of the working signal, subsequently to identify transmissions, if necessary, to warn the monitoring cabin of an expected transmission, to make time-signal recordings upon the same tape of the end of the transmission, using the talking clock of the telephone exchange.

The apparatus is started and stopped by a working signal on the monitored frequency.

Any receiver and a standard tape recorder, or other transmission recording instruments, can be connected to the apparatus. It can operate either as a permanent monitor or for the period of the watch, scheduled in advance, on the frequency concerned.

The working signal of the receiver from the intermediate-frequency terminal, the audio-frequency terminal or the d.c. terminal of the automatic gain control is used as triggering signal. After amplification, amplitude limitation and signal detection, a polarized relay is used and its sensitivity can be regulated.

To prevent unwanted triggering of the apparatus by atmospherics, variable time-constant elements are incorporated in the detector stage to obtain a delay effect.

The criterion adopted to avoid mis-triggering is the duration of the control signal.

At the end of the transmission, the polarized relay clears after a slight delay, which is necessary to avoid an unwanted break in the recording due to the pause between the letters or words in the case of an A1 emission.

A stage, with a time constant of 20 to 30 s after the end of the transmission, ensures continued operation of the apparatus while the telephone number of the talking clock is being dialled to record its time signals on the same tape. The telephone line is occupied only while the time is being recorded.

In case this number is busy, a special selective stage is provided which uses the characteristics of the busy signal and sets the whole switching process in motion again.

Recording of the working time with this apparatus is very accurate, with a tolerance of ± 10 s.

6.2 *Automatic monitoring of known radio emissions at monitoring stations*

[CCIR, 1970-74], describes equipment for the automatic monitoring of known radio emissions at monitoring stations. This equipment automatically and systematically measures the frequency, field strength and occupied bandwidth of known and pre-selected transmissions. The equipment is controlled by a recorded prepared programme. Some details are given of the principal performance characteristics of the equipment together with block diagrams of the equipment. [Mikhailov *et al.*, 1973] describes similar equipment for automatic measurement at monitoring stations of the occupied bandwidths of emissions of known and pre-selected transmitters operating in the 150 kHz-30 MHz range.

6.3 *Automatic radio-frequency spectrum-occupancy recording equipment*

Document [CCIR, 1978-82] describes equipment for the automatic recording of spectrum-occupancy in the frequency range 10 kHz to 30 MHz or 10 kHz to 1300 (option: 2500) MHz. The equipment comprises a digitally controlled receiver with incorporated synthesizer, a programme control unit with microprocessor and up to six YT (occupancy) recorders in simultaneous operation. The controller provides the possibility to select up to six independent frequency bands for strip recorders. An XY recorder may be optionally added.

The controller independently computes the optimal stepping width and resolution bandwidth according to the resolution of the recorder, and can accommodate approximately 1000 increments between selected start and stop frequencies. The receiver, governed by the controller, will successively scan up to six frequency ranges with the maximum possible speed. All frequency readings for active channels will, corresponding to the lines on the chart, be written into a semi-conductor memory. Depending on the time compression desired, several lines can be combined in such a way that no signal is lost or the paper feed rate can be adapted to the duration of observation. The records produced from five scans in one frequency band for example, can be printed in one line so that observations lasting 24 hours can be recorded on a single A4 page.

System specifications include:

Frequency range:	1st receiver: 10 kHz to 30 MHz with digitally controlled receiver; 2nd receiver: 10 kHz to 1300 (2500) MHz using a search receiver.
Analysis bandwidths (in the controller):	± 12.5 Hz (four parallel channels); ± 25 Hz (two parallel channels); ± 50 Hz, ± 150 Hz. *Note.* – Additional bandwidths are available in the associated receiver.
Dynamic range:	⩾ 60 dB with constant receiver gain.
Frequency resolution:	two adjacent signals 200 Hz apart with a difference in amplitude of ⩾ 40 dB, using the analysis filter ± 12.5 Hz, are clearly distinguishable in the recording.
Scanning speed:	1st receiver: 5 to 30 s per scan with analysis filter bandwidth ⩾ ± 12.5 Hz; dependent on the selected sweep width; 2nd receiver: from approximately 1 s per scan for receiver filter bandwidth ⩾ ± 12.5 kHz to approximately 30 s per scan for analysis filter bandwidth ⩾ ± 12.5 Hz.
Number of recorded sub-ranges:	six; a recorder is assigned to each sub-range, and the six recorders operate in quasi-parallel fashion in as much as permitted by the scanning speed of the receiver; to reduce paper consumption the activities of up to eight lines can be displayed in one line by means of an OR gating.
Receiver control:	standard IEC bus.
Frequency lines:	electronically superimposed; maximum ten lines per sub-range.
Time markers:	built-in digital clock with display in hours and minutes; continuous line on the full hour in the recording.
Power-outage protection:	program secured in a nonvolatile memory.

7. Equipment developed in Japan

Multifunctional monitoring equipment developed and installed in major monitoring stations in Japan facilitates the simultaneous measurement of frequency, occupied bandwidth and field strength.

Successful field operations have established that this equipment has proved efficient and reliable.

7.1 *Outline and composition*

This equipment which consists of a receiver, a digital frequency synthesizer and a micro-computer provides "panoramic reception" in the frequency range of 25 to 1525 MHz. The receiver is a quadruple super-hetrodyne type in order to improve intermodulation and spurious response performance. The micro-computer controls the measuring system including the receiver.

A digital frequency synthesizer which is used as the first local oscillator provides frequency scanning with minimum increments of 1 Hz.

7.2 *Frequency measurement*

A counter is used to measure the frequency of the signal at the intermediate frequency (F_{I4}, nominally 3 MHz) derived from the fourth local oscillator. Variations in the frequency of local oscillators F_{L2} and F_{L3} are cancelled by feed-back to the first local oscillator. The frequency at the input of the frequency counter is given by the following formula:

$$F_{I4} = F_x - (F_{L1} \pm \Delta F_{L1}) + (F_{L4} \pm \Delta F_{L4})$$

As F_{L4} is derived from a stable crystal oscillator (1×10^{-8} or better), the high accuracy of frequency measurement is attributable to the accuracy of the standard oscillator frequency (F_{L1}) in the synthesizer (5×10^{-7} or better). The frequency at the receiver input is obtained by conversion of F_{I4} in the micro-computer.

7.3 *Field strength measurement*

Field strength measurement is achieved by measuring the voltage at the input of the receiver, with automatic correction by the micro-computer for the antenna coefficient factor and feeder loss.

Field strength is measured as follows: the received signal at the 4th intermediate frequency stage passes through a logarithmic compressor (dynamic range of more than 70 dB), and is quantized by an analogue to digital converter (8 bits, 0.5 dB steps) after demodulation. Due account is taken of the RF and IF attenuators settings inserted by the micro-computer.

7.4 *Bandwidth measurement*

The x dB method is used to measure occupied bandwidth. The frequency spectrum of the received emission is observed on a CRT. The display is very accurate and stable since the resolution bandwidth, frequency sweep, scan step, and sweep time of the CRT are controlled by the micro-computer.

7.5 *Main characteristics*

The main characteristics of the equipment are shown in Table I.

TABLE I - *Main characteristics of integrated radio monitoring equipment*

Items	Performance
Frequency range	25 MHz - 1525 MHz
Receiver type	Quadruple super heterodyne
Control method	Micro computer
Sensitivity	Less than 0.2 μV
Frequency accuracy	Within 1×10^{-6}
Field-strength accuracy	Within ± 1.5 dB
Receiver characteristics:	
RF attenuator	0 - 50 dB, 10 dB step
IF attenuator	0 - 40 dB, 10 dB step and 10 dB continuously variable
Dynamic range	70 dB
Resolution bandwidth	600 kHz, 300 kHz, 100 kHz, 30 kHz, 10 kHz, 3 kHz, 1 kHz, 300 Hz and 100 Hz
Digital scan characteristics:	
Scan mode	Auto, single, stop
Scan width	30 MHz, 10 MHz, 1 MHz, 300 kHz, 30 kHz, 10 kHz, 3 kHz, 1 kHz, 300 Hz 9 switch selectable
Scan step	100, 500
Initial frequency setting	10 MHz step, thumb-wheel switch selectable
Panoramic display	5 inch (12.7 cm) storage CRT

7.6 Automatic radio-frequency spectrum analysis system

An improved economical monitoring system with automatic analysis software to assist in effective spectrum management has been developed in Japan in early 1989.

This system automatically collects radio-frequency spectrum data detected within specified swept frequency ranges from 100 kHz to 2 000 MHz by using a spectrum analyser that is under the control of a personal computer in accordance with a preset schedule for a long-term period. The spectrum data stored in memory are analysed by the personal computer into the following output formats:

- variation in time of spectrum occupancy;
- signal amplitude variation;
- amplitude spectrum;
- channel occupancy variation in frequency domain;
- single channel occupancy variation;
- channel occupancy spectrum.

These data, which are displayed on a CRT and/or hard-copied, offer effective means for spectrum management.

8. Equipment developed in the Federal Republic of Germany (10 kHz to 960 MHz)

8.1 *System No. 1*

An electronically controlled receiving and measuring system has been used by the radio monitoring service in the Federal Republic of Germany since 1974. This system measures received signal level, bandwidth, frequency error, FM frequency deviation, modulation factor, phase angle and channel occupancy as a function of time, in the frequency range 10 kHz to 960 MHz. The measured values (except the phase angle), which are displayed digitally, can also be plotted in the analogue mode by an XY or YT recorder. They are also available at a printer outlet. The system can be used as a digital high-precision spectrum analyzer to investigate spectra of individual transmitters or for analyzing wide frequency spectrum ranges.

The system was developed for fixed and mobile use to simplify and speed up the measurement of emissions at great distances from the transmitter. It is capable of carrying out a large number of automatic and semi-automatic measuring and control functions and has been provided with simple calculating circuitry. The equipment has capacity and flexibility for new measuring tasks and for full computerization (see Table II).

A block diagram of the system is given in Report 668-1 (Geneva, 1982).

The attenuator following the antenna allows up to 90 dB attenuation to be inserted, so as to avoid overloading due to strong input signals.

TABLE II – *Main characteristics of the receiving and measuring system*

Frequency range	10 kHz - 960 MHz
Noise figure	9 dB (−165 dB (mW/Hz))
Field strength accuracy	±2 dB
Frequency uncertainty	5×10^{-9}/24 h
Input attenuation	0 to 90 dB in 10 dB steps
IF bandwidth	0.5 kHz - 1 MHz
Measurements:	Measuring ranges:
Input level	dynamic range: 75 ... 85 dB
Frequency error	0.01 ... 9.99 kHz / 0.1 ... 99.9 kHz
Frequency deviation	0.01 ... 19.99 kHz / 0.1 ... 199.9 kHz
Modulation factor	0 - 99.9%
Phase angle	±180°
Channel occupancy	0 ... 99.9%
Measuring time	6, 12, 30, 60 min
Analysis:	
Number of steps	1 ... 1000
Step magnitude	1 Hz ... 10 MHz
Repetition rate of frequency steps	0 ... 10 000/s
Power consumption	425 VA

The received frequency is pre-selected by 14 successive band-pass filters that cover the frequency range 10 kHz to 960 MHz. The appropriate filter is selected automatically as the synthesizer frequency changes. The noise figure is 9 dB, (−165 dB(mW/Hz)). The conversion frequency is generated in the master oscillator (synthesizer) to produce a 60 MHz IF.

In the IF part of the central measuring unit the signal first passes IF filters with different bandwidths, selected optionally by the user. Individual transmission channels, for example, are preferably monitored by employing highly selective channel filters (12.5, 20, 25, 50 kHz bandwidths), whereas Gaussian filters with a good transient response (0.5, 1, 3, 10, 30, 100, 300, 1000 kHz) are best suited for the purpose of frequency analysis.

A number of analogue IF measuring circuits allow the following parameters to be determined:

– *input level* in the digital display range 0.00 ... 99.9 dB, relative to a reference level of +30 ... −60 dBm, that can be pre-selected by means of the attenuator. Owing to the noise level of the preceding receiver stages, the 100 dB inherent dynamic range of the logarithmic amplifier is limited to 75-85 dB, depending on the bandwidth chosen;

– *frequency error* of an AM (with carrier) or FM signal as compared with a nominal frequency which can be pre-set in the two ranges 0.01 ... 9.99 kHz and 0.1 ... 99.9 kHz. The digital readout includes the sign. The measuring time is pre-set to 0.1, 1 or 10 s;

- *frequency deviation* of an FM signal which is the average of the positive and negative peak deviation absolute values within a measuring period over the measuring ranges 0.01 ... 19.99 kHz (narrow-band FM) and 0.1 ... 199.9 kHz (wide-band FM);

- *modulation factor (depth)* in the digital display range 00.0 ... 99.9%;

- *phase angle* in the measuring range ± 180° (analogue display and recording), allowing among other things, the built-in reference oscillator to be calibrated against a standard-frequency transmitter (e.g., a precision offset tv transmitter to $< 2 \times 10^{-9}$).

Using the printer outlet of the system, the measured digital data enumerated above can be printed out by means of an external printer. An audio amplifier with a built-in loudspeaker is also available.

The digitally controlled master oscillator (synthesizer) tunes the receiver in steps of down to 1 Hz resolution. Received frequency is displayed to 9 digits to this resolution. In the case of manual frequency selection, the received frequency is set to the desired value by means of code switches or varied by means of the quasi-continuous tuner. In frequency analysis and continuous channel changing, tuning is effected automatically by the master oscillator.

Frequency changes of the synthesizer take place very rapidly. The frequency changes of the system as a whole depend, however, on the decay time of the IF filters, required during the transition from an occupied to an unoccupied channel. In the worst case, this is 4 ms for the narrow and highly selective channel filter with 12.5 kHz bandwidth.

The frequency inaccuracy of the quartz crystal oscillator is specified to be 5×10^{-9}/day at ambient temperature after 1 hour warm-up time. If a higher frequency accuracy is needed, the built-in crystal oscillator can be synchronized with the aid of an atomic standard.

The digital programming and calculating unit internally coordinates the system functions with simple calculating circuitry. When used as an analyzer, for example, it sends commands for 1/2/4/10 times 1/10/100 steps of 1 Hz ... 10 MHz each. The repetition rate of the frequency steps is 0/1/2/5/10 times per second.

For the control of an external XY recorder, an X voltage (with a maximum value of 10 V) proportional to the frequency, is supplied.

In frequency analysis the coverage of wide frequency ranges with a high resolution may necessitate more than 1000 steps. Therefore, the digital programming and calculating unit has been provided with switching facilities that allow several adjacent frequency bands of equal width to be swept sequentially. The sweep can be directed alternatively to higher or lower frequencies.

In addition to frequency code switches, a control knob enables "quasi-continuous" setting of the received frequency. During one turn of this control the unit performs 18 changes in frequency the step magnitude of which is adjustable from 1 Hz to 10 MHz.

Moreover, the digital programming and calculating unit compares the measured frequency deviation (FM) and the frequency error of an emission when these are outside the permitted and pre-set tolerances. The number of signals beyond tolerances can be counted and evaluated.

To determine the degree to which a frequency channel is utilized, the channel occupancy is determined as the ratio of the holding time to the total measuring time, where the latter can be set to 6, 12, 30 or 60 minutes. Channel occupancy is displayed digitally in percentage to three significant figures with 0.1% resolution.

With the aid of the recording panel and scope monitor a frequency range to be analyzed can be displayed visually, where digital memory and comparator circuits enable one minimum and maximum level distribution curve each to be represented.

It is possible to show a maximum of 100 occupied adjacent channels in the form of vertical bars with a resolution of up to 1000 points (corresponding to 1000 samples). The percentages of occupancy, frequency error and frequency deviation beyond pre-set references can also be represented for a maximum of 10 channels.

Peripheral equipment, such as oscilloscopes, XY recorders, YT recorders, digital printers, magnetic tape and data recorders, can be connected to this receiving and measuring system for further processing and storage of the results of measurement.

8.2 *System No. 2 with increased field strength measuring capability*

CCIR [1982-86a] describes a monitoring system consisting of two automatic measuring receivers, the first one for the frequency range 9 kHz to 30 MHz (in use since 1980) and the second one for the frequency range 20-1300 MHz (in use since 1983). Antennas with calibrated antenna factors, an XY and YT recorder, up to ten occupancy recorders, five controlled by one receiver (for automatic occupancy monitoring of the radio-frequency spectrum), a panoramic adapter (IF spectrum analyzer) and finally a bus controller with high-resolution graphics and a printer for hard copies and digital print-out may be added for full use of all the capabilities of the receivers.

The system measures received signal level, field strength, channel occupancy as a function of time, frequency and frequency offset, FM frequency deviation, AM modulation depth and band occupancy as a function of time. Furthermore, it fully meets the requirements of CISPR Publication 16 for the measurement of radio interference.

8.2.1 *System characteristics*

Main characteristics of the two receivers:

	1st receiver	*2nd receiver*
Frequency range:	9 kHz to 30 MHz	20 to 1300 MHz
Level measurement range:	−30 to 137 dBμV	−20 to 137 dBμV
Errors: input level:	< 1 dB	
field strength (including antennas):	< 2 dB	
frequency:	$< 1 \times 10^{-9}$/24 h	
Overload capacity (3rd order intercept point):	+25 dBm	+20 dBm
Dynamic selectivity (values 20 kHz apart from carrier):	Typically 110 dB (B_{IF} = 0.2 kHz)	Typically 85 dB (B_{IF} = 7.5 kHz)
IF bandwidths (6 dB):	0.2, 0.6, 2.6, 9.5 kHz	8.3, 13, 120, 1000 kHz
Swept frequency range:	Start and stop frequencies may be arbitrarily chosen within the frequency range of each of the two receivers; step width (integral multiples of 100 Hz and 1 kHz for the first and second receiver, respectively) is normally set to increments less than, or equal to, the resolution bandwidth; number of steps per sweep should not be greater than 1024 when operated with associated recorder (10 bits D/A converter).	
Number of sweeps per time unit:	Determined by number of steps; one step approximately 30 ms. *Example:* sweep with 250 steps yields 8 sweeps/min.	
Resolution bandwidth:	See IF bandwidth	
Further measurements:	Frequency deviation, AM depth, offset	
Demodulator functions (for aural monitoring):	CW, AM, FM, SSB (USB, LSB)	
Power supply:	from a.c. or d.c. (+24 V)	

Additional characteristics:

- 50 Ω input
- RF attenuator 0-140 dB in 10 dB steps with a 1 dB pad for linearity test
- tracking generator for calibration and insertion loss measurements
- receiver 9 kHz to 30 MHz: 16 pre-selector ranges (switch selected)
- receiver 20-1300 MHz: 10 tuned pre-selectors
- weighting circuits for average, peak and CISPR quasi peak
- linear detectors for true averaging.

8.2.2 *Signal level and field-strength measurement*

High level measuring accuracy is obtained by using internal tracking generators with a typical level error of less than 0.2 dB, together with high precision RF attenuators and a built-in automatic calibration procedure. Verification of level accuracy is easily done with a power meter at the tracking generator output of the receivers. In addition to the manual setting of the RF and IF attenuators, auto-ranging can be carried out by a microcomputer by controlling, in steps, the RF/IF attenuators, to give priority to low distortion or low noise. Measuring time (i.e. time for digital level averaging or the peak hold time) may be chosen between 5 ms and 100 s. With a measuring time of 5 ms and a sufficiently fast bus controller, with small frequency steps using fixed attenuators, a measuring rate of 20 per second may be achieved.

Measurement of intermittent emissions is facilitated in the VHF/UHF range with internal and external trigger capability.

Field strength is indicated and recorded directly in dB(μV/m) using the available active antennas which have constant antenna factors in the frequency range of 9 kHz to 30 MHz. In the frequency range of 20-1300 MHz, the dependent antenna factors of the available antennas are taken into account. The antenna factor of any other calibrated antenna (e.g. omnidirectional) may be entered into the receiver covering the range 20-1300 MHz. Minimum measurable field strength is given in Fig. 1.

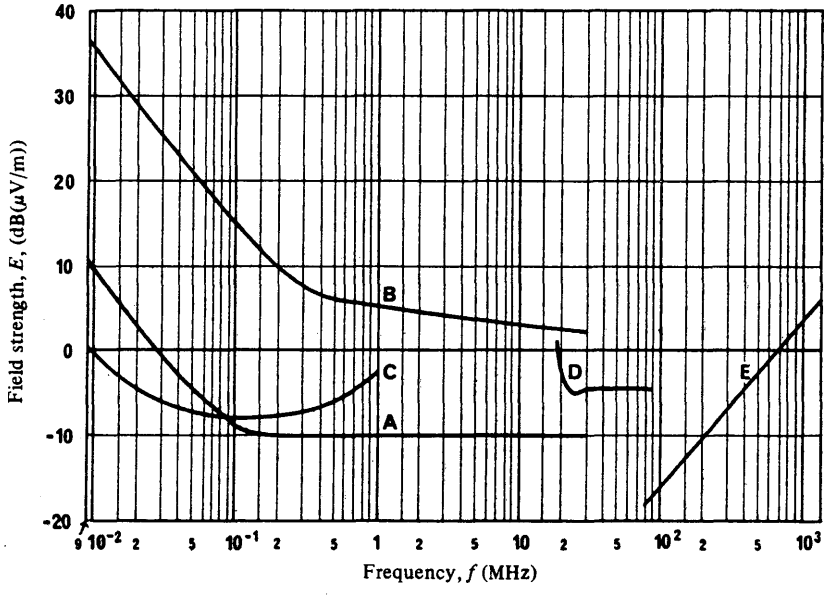


FIGURE 1 – *Minimum measurable field-strength level (for S/N = 1) for different types of antenna*

A: active rod antenna, 1 m (9 kHz to 30 MHz)	receiver 9 kHz to 30 MHz B_{IF} = 0.2 kHz
B: active loop antenna, 0.6 m diameter (9 kHz to 30 MHz)	
C: active loop antenna, 2.4 m side length (9 kHz to 1 MHz)	
D: CISPR broadband dipole (20-80 MHz)	receiver 20-1300 MHz B_{IF} = 7.5 kHz
E: log periodic antenna (80-1300 MHz)	

8.2.3 *Frequency setting, measurement of input frequency and automatic frequency scans with recording of results on recorders and printers*

All internal synthesizer and oscillator frequencies are derived from one master oscillator. Automatic frequency scans can be programmed for start and stop frequencies, step widths and threshold levels. Signal levels that are above the threshold level interconnection are recorded together with the associated frequency on a printer connected to the interconnection bus. AM depth, FM deviation and offset can also be printed if these functions are chosen.

Either of the two receivers can use two D/A converters for the control of XY and YT recorders and up to five occupancy recorders of the type mentioned in § 6.3, to provide calibrated amplitude versus frequency records and channel and band occupancy records as a function of time.

For frequency measurement of AM, FM and PM signals, the filtered and limited IF signal is reconverted to the input frequency using the same oscillators as for down conversion. Thus the receiver is used as a tunable filter, and measurement uncertainty is determined by the frequency counter connected to the tracking generator output, even for low to medium signal-to-noise ratios. This method is also described in Report 272. In addition, the VHF/UHF receiver has a built-in IF frequency counter for either measurement of input frequency or offset from receiver frequency with resolutions of 0.1, 1, 10 and 100 Hz. This measurement accuracy requires a signal-to-noise ratio of equal to, or greater than, 14 dB.

8.2.4 *Extended monitoring capabilities with a bus controller*

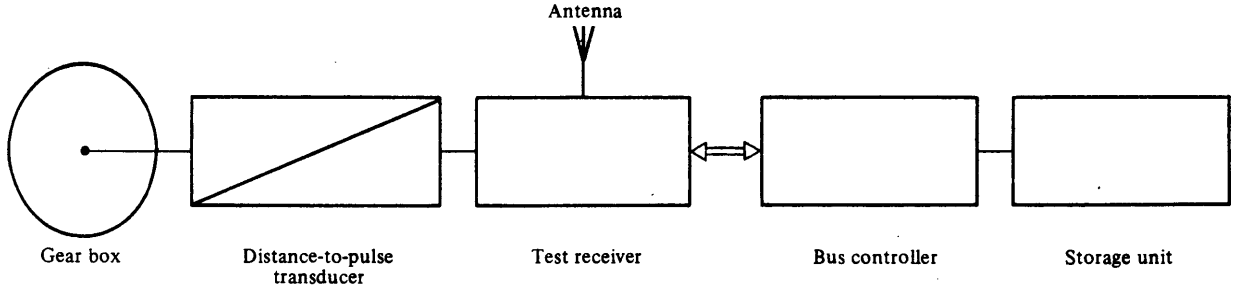
All settings can be entered into the receiver and results obtained by the use of a bus controller (IEC 625/IEEE 488). Special attention is given to ease of programming and reduction of data stream. Thus, receivers and controller form a system with distributed intelligence. Results are presented with high-resolution graphics and with printer hard copy.

8.2.5 *Field-strength measurement in a moving vehicle in the frequency range above 20 MHz*

A method for judging the receiving conditions in the mobile radio bands is needed which gives valid statistical data of the field-strength distribution at the receiving antenna of mobile radio stations. This applies particularly in built-up areas where the line-of-sight to the transmitter is blocked, and the field strength measured at any point is made up of the contributions of many indirect waves of relatively low power that reach the antenna by refraction, scattering and reflection. The field-strength distribution thus varies strongly with the location.

Field-strength statistics are used world-wide for the planning of mobile-radio networks in the VHF/UHF bands. The second receiver (20-1300 MHz) of system No. 2 has the capability of fast field-strength measurement with up to 1000 measurements per second.

In order to ensure that the field-strength measurements along the route are not affected by the speed of the vehicle, a distance-to-pulse transducer is used which generates pulses, each of which triggers an A/D converter process in the test receiver.



There are two ways of evaluating the A/D converter outputs:

- the binary data output of all samples can be sent to a fast IEC bus controller for data storage and further processing;

- an internal classification of n samples (n: integer multiples of 100; n = 100-65 000) can be made and the output of the field-strength distribution sent to the bus controller or a printer. Fifteen field-strength values corresponding to the exceeding percentage of the distribution are output after every n samples and from these values an immediate plot of the distribution can be drawn on a Rayleigh grid.

9. System and equipments developed in France (20-1 350 MHz)

An automatic radio network monitoring system has been in use with the frequency monitoring service since 1985.

The system is based on a fixed infrastructure comprising six computerized monitoring centres and automatic outlying stations (measuring sensors) and provides the following functions:

- systematic monitoring of the technical characteristics of known transmitters in order to detect certain anomalies;
- measurement of the activity rate of known transmitters (per time slot) in order to improve frequency assignment and sharing;
- tracing of unknown or illegal transmitters;
- monitoring of networks having a standard identification system (selective calling) to ensure conformity of the network structure with the initial specifications.

Defined initially mainly for monitoring private communication networks, the system is to be extended in 1990 to cover public networks, inter alia for locating mobile stations of the Radiocom 2000 network with a view to evaluating traffic density per geographical area.

9.1 Description of monitoring centres

All licensed monitoring centres receive full descriptive data on known networks from the administrative centre. The data are translated by the technical data system supplying the outlying stations.

Each monitoring centre is operated by two persons responsible for:

- defining and initializing the automatic measuring and monitoring programmes of each outlying station;
- identifying transmitters by manually remote-controlled use of the outlying stations and implementation of the location function;
- processing real-time alerts generated by the outlying station upon automatic detection of anomalies in the networks monitored;
- processing the information generated by outlying stations upon completion of a monitoring programme.

Each monitoring centre is equipped with the following facilities:

- a data base for storing initial data and the results of measuring and monitoring operations;

- a graphic system displaying:
  - the geographical map,
  - the known networks,
  - the location results,
- a system for supervising and monitoring all the tasks.

9.2 Description of outlying stations

Outlying stations are fully automated and remotely-controlled from the monitoring centre.

They can thus be set up on sites selected for their radio coverage.

Each outlying station comprises sensors for taking basic measurements of received transmitters.

The basic measurements are:

- direction-finding measurements;
- frequency measurements;
- signal level and field-strength measurements;
- maximum frequency deviation measurements;
- decoding of standard selective calls.

A computer and its associated real-time software are used for implementing the automatic monitoring functions.

The functions are:

- scanning of a frequency band or a discrete frequency list;
- initialization of basic measurements upon detection of power in a channel;
- comparison of the measurements taken with the local file describing known transmitters;
- generation of a real-time alert to the monitoring centre upon detection of anomalies during a monitoring exercise;
- generation of a results file listing all the measurements taken. The file is transmitted to the monitoring centre once the function is completed;

- automatic management of a recorder for deferred transmitter identification;

- processing of a direction-finding request from the monitoring centre during a location exercise.

9.3 Identification of transmitters monitored

The monitoring functions implemented in the system are based on the identification of transmitters received.

Identification based chiefly on the position of the transmitter requires great precision in the direction-finding measurement and a measured azimuth acquisition time compatible with the mean communication times of the networks monitored.

After comparative measuring campaigns conducted in an urban environment, the administration has adopted a broad-based interferometer-type direction finder, which is relatively insensitive to the reflection and multipath situations encountered in urban environments.

When the azimuth angle is not sufficient to identify a transmitter without ambiguity, a field-strength measurement is taken as well.

Some functions exploit closer transmitter identification based on selective call decoding.

9.4 Results generated by outlying stations

Outlying stations generate real-time alerts upon detection of anomalies and consolidated results based on all the measurements taken.

Real-time alerts are generated when:

- the frequency offset is exceeded;

- the maximum modulation deviation is exceeded;

- a specified occupancy rate is exceeded;

- an unknown transmitter is detected;

- an illegal transmitter is detected.

The consolidated results give for each transmitter:

- the overall and/or time-slot activity rate;

- the measured anomaly rate.

9.5 Physical architecture of outlying stations

Each outlying station is equipped with a single receiving channel covering the frequency range 20-1 350 MHz.

The channel is used sequentially for:

- scanning a frequency band and detecting power in a channel,
- direction finding;
- measuring and decoding.

Depending on the processing sequence, the receiving channel is automatically connected either to an array of omnidirectional antennas or to the direction-finding antenna.

The interferometer radiodirection finder is of the single-channel type.

A set of switches and delay lines is used to obtain differential phases from the dipole pairs of the broad-based direction-finding antenna.

This technique obviates multiple reception channel pairing problems.

The measuring and decoding function comprising measurements of frequency, signal level, field-strength and maximum frequency deviation. The decoding of standard selective calls is periodically tested by injecting a reference signal from a test generator at the input to the reception chain. The generator is fitted with a 10-MHz high-stability pilot tone, which is fed to the station equipment.

Field strength is measured by applying a coefficient K to the signal level measurements. Coefficient K includes the antenna factor and the link losses between the antenna system and the measuring receiver. The table of K values is specified for each station.

The outlying stations are designed to allow for the radio environment encountered in an urban environment. A rack that can be offset at the foot of the antennas contains the antenna switch, a "tracker" filter automatically tuned to the tuning frequency of the reception channel (diode tuning), a switchable amplifier, a variable attenuator and, depending on the site, a set of strong-signal rejection filters. The rack is used to match the characteristics of the outlying station to local reception conditions. The measuring receiver itself has a "tracker" filter identical to the one upstream.

10. **Automatic monitoring studies**

A development project for a computer-assisted spectrum management system (SMS) was initiated in Canada in 1975. This project identified a requirement for channel occupancy data for channel-sharing purposes in the frequency assignment process. A sophisticated mobile computer-controlled scanning receiver and data acquisition system was deployed to determine practical monitoring procedures which would require the minimum number of monitoring sites for a given area. The monitoring units scan 246 channels twice per second and record time, frequency and amplitude of signals detected on each scan. Software has been developed which reduces the monitoring data to give a measure of occupancy for each base or fixed station (signal detected on each channel during a monitoring period). Software and methods to associate the measured occupancies to the corresponding station licence records have not yet been completed [CCIR, 1974-78].

The Australian Administration has developed a simple and relatively inexpensive computer-controlled system to allow automatic gathering and processing of data and provide an indication of channel occupancy in the VHF and UHF land-mobile bands. The system comprises a low-cost domestic quality scanning receiver, a medium priced desk-top computer complete with floppy disc mass storage, plotter and printer, and some custom-built interface electronics which allows the computer to control the scanning receiver [Bray, 1983]. The system is capable of scanning up to 30 channels over a period of 7 days, each scan requiring 5 s with an individual channel sampling time of 1 ms [CCIR, 1982-86b].

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Note. – The above items are sold by the US Department of Commerce, National Bureau of Standards, Clearing House for Federal Scientific and Technical Information, Springfield, Virginia, 22151, at a cost of $3.00 each, post paid. The following numbers have been assigned to the reports and should be used when ordering from the Clearing House:

Part A – PB 182-792
Part B – PB 182-918

The *Final Reports* are also available from the same source for those who may desire information on the results and conclusions of the land mobile spectrum utilization study made at certain cities within the United States of America.

However, the details concerning the spectrum occupancy data accumulation equipment system, and method of analysis, are found in the *Interim Reports* listed above.

SECTION 1D: SPECTRUM UTILIZATION AND APPLICATIONS

REPORT 830-1

USE OF INTERFERENCE CANCELLERS IN ORDER TO INCREASE SPECTRUM USAGE

(Questions 18/1 and 44/1)

(1982-1986)

1. Introduction

One of the most important tasks of modern radio engineering is to design equipment enabling the maximum number of radio systems to be operated.

One method is to construct interference cancellers for interactive communication systems using various types of modulation. In a number of cases where the traditional methods of ensuring the trouble-free operation of radio systems by geographical or frequency separations are inapplicable, interference cancellers are one means of enabling interactive radio links to operate simultaneously. Another important application of interference cancellers occurs in systems in which the main communication channel is multiplexed with additional information transmitted in the same bandwidth as the main information. In this case, interference cancellers allow the main and additional channels to be separated with a negligible level of crosstalk and a proportionate increase in channel capacity.

The most universal type of interference cancellers are devices which form a replica of the interference and subtract this replica from the received signal, made up of the wanted and interfering components. Such cancellers may be used to combat a wide range of different interfering signals, including signals modulated in the same way as the wanted signal.

In many cases, however, such as for interference caused to multi-channel radio-relay and satellite systems by radiolocation and radionavigation equipment, when the pulse signal becomes powerful enough to block the receiver, these methods prove ineffectual. More effective in such cases are methods which eliminate the impulsive noise and restore the affected parts of the wanted signal by extrapolation or interpolation. Since in practice, the interfering pulses are often extremely short in comparison with the wanted message correlation time, have a very low repetition frequency and have sufficient level for reliable detection, it is possible in many cases, to considerably simplify the design of the pulse interference cancellers.

Two categories of interference cancellers are described in this Report:

- single-channel devices which process the additive mixture of the wanted and the unwanted signal reaching the receiver input;
- multi-channel devices, in which at the input of each of the channels, either one or several unwanted signals from other receiving channels arrive in addition to the wanted signal.

Pulse interference cancellers are divided into two categories:

- devices using spectrum transformation and signal limitation,
- devices which restore the affected parts of the wanted signal by extrapolation or interpolation.

2. Single-channel interference cancellers designed for the separation of AM and FM signals

2.1 *Results of theoretical studies*

Applying the theory of optimum reception of analogue types of modulation, papers [Bykhovsky, 1979a; Bykhovsky and Dorofeev, 1978; Sundresh *et al.*, 1977] select the optimum devices for discriminating AM and FM signals. The devices in question extract the message $a_i(t)$ from the composite signal $W(t)$ at the receiver input:

$$W(t) = A\left[\sum_{i=1}^{N} \mu_i S_i(t, a_i, \bar{\lambda}_i)\right] + \eta(t) = S(t, \bar{a}, \bar{\lambda}) + \eta(t) \quad (1)$$

where:

$S_i(t, a_i, \bar{\lambda}_i)$: i^{th} received signal modulated by message $a_i(t)$;

$\bar{\lambda}_i$: its parameters;

$\mu_i A$: the amplitude of the i^{th} signal;

$\eta(t)$: white Gaussian noise with one-sided spectral density N_0.

For AM: $S_i(t, a_i, \bar{\lambda}_i) = [1 + a_i(t)] \sin(\omega_i t + \varphi_i)$

For PM: $S_i(t, a_i, \bar{\lambda}_i) = \sin[\omega_i t + a_i(t) + \varphi_i]$

For FM: $S_i(t, a_i, \bar{\lambda}_i) = \sin\left[\omega_i t + \int_{-\infty}^{t} a_i(t)\, dt + \varphi_i\right]$

Here, ω_i and φ_i are the carrier frequency and phase of the i^{th} signal. If $N = 2$ in equation (1), an analysis based on the theory of optimum reception results in the following equations describing the operation of the optimum equipment:

$$a_i^*(t) = \mu_i H(\rho)\left\{[W(t) - S_i(t, a_i, \bar{\lambda}_i)]\frac{\partial S_i(t, a_i, \bar{\lambda}_i)}{\partial a_i}\right\} \quad (2)$$

where $a_i^*(t)$ is the optimum estimate of the process a_i, $H(\rho)$ the operator representation of the linear filter transfer characteristic determined by the correlation function of the process $a_i(t)$. Figure 1 is a block diagram of a tracking interference canceller modelling equation (2) for $N = 2$. This device has at its input two reception branches, each of which estimates one of the messages a_i. These estimates are sent to the modulators where they are converted into signals.

$$\mu_i S_i(t, a_i^*, \bar{\lambda}_i) \quad \text{and} \quad \mu_i \frac{\partial S_i(t, a_i^*, \bar{\lambda}_i)}{\partial a_i}$$

the first of which are summed, forming a replica of the signal received $W^*(t)$, while the second are sent to multipliers in each branch. The replica $W^*(t)$ is sent via the feedback circuit to the receiver input, where it is extracted from the signal received. For certain types of signals $S_i(t, a_i, \bar{\lambda}_i)$ (AM, FM, etc.), the receiver described in Fig. 1 may be greatly simplified.

Bykhovsky [1979a] describes the optimum device designated in accordance with the diagram in Fig. 1 for AM signals and is represented in Fig. 2. Narrowband filters extract the carrier frequency from the received signal. In one branch, the carrier, after limitation in a band limiter, is transmitted to synchronous detectors of the corresponding channels and in the other branch it is modulated in amplitude modulators by wanted message estimations. The signals from the synchronous detector output are filtered by transfer function filters $H(\rho)$. The subtractors in the device in Fig. 2 are intended to cancel inter-channel crosstalk.

Bykhovsky and Dorofeev [1978], Sundresh *et al.* [1977] describe the optimum interference canceller for two FM signals. The diagram adapted from Fig. 1 is shown in Fig. 3. It incorporates two subtractors, one at each channel output and two synchronous phase detectors consisting of a phase detector, a low-pass filter and a voltage-controlled oscillator. The attenuators are so adjusted that the FM signal amplitudes are equal at their output to the amplitudes of the compensated signals.

An analysis of the operation of the optimum AM signal receiver with an interference canceller [Bykhovsky, 1979a] showed (Fig. 2):

- that it effectively separates two AM signals occupying a common bandwidth;
- that it enhances interference immunity on reception compared with the normal synchronous receiver, the improvement being the greater the larger the amplitude of the interference due to the receiver's interference suppression;
- that in the presence of unwanted signals at the input, the receiver has a considerably improved (25-28 dB) signal-to-interference ratio at the output, compared with the normal synchronous receiver, with a detuning between the carriers equal to the upper modulating frequency, with a wanted-to-unwanted signal power ratio at the input of 0 dB.

The theoretical study in [Bykhovsky, 1979b] of the potential interference immunity of two FM signals using an interference canceller (Fig. 3) showed:

- that FM signals can be completely separated without crosstalk between them, when they have carriers of the same frequency and amplitude, with a 90° phase shift between them, which doubles the capacity of systems using FM without any increase in occupied frequency bandwidth;
- that when the phase shift between the separated FM signals is random and may assume any value, the carrier frequency separation has to be increased for effective separation to obtain low crosstalk;
- that optimum receivers equipped with interference cancellers should also be used where there is wide frequency separation, since this type of receiver, unlike the standard types, has a demodulator output interference level which is independent of the level of the input unwanted signal.

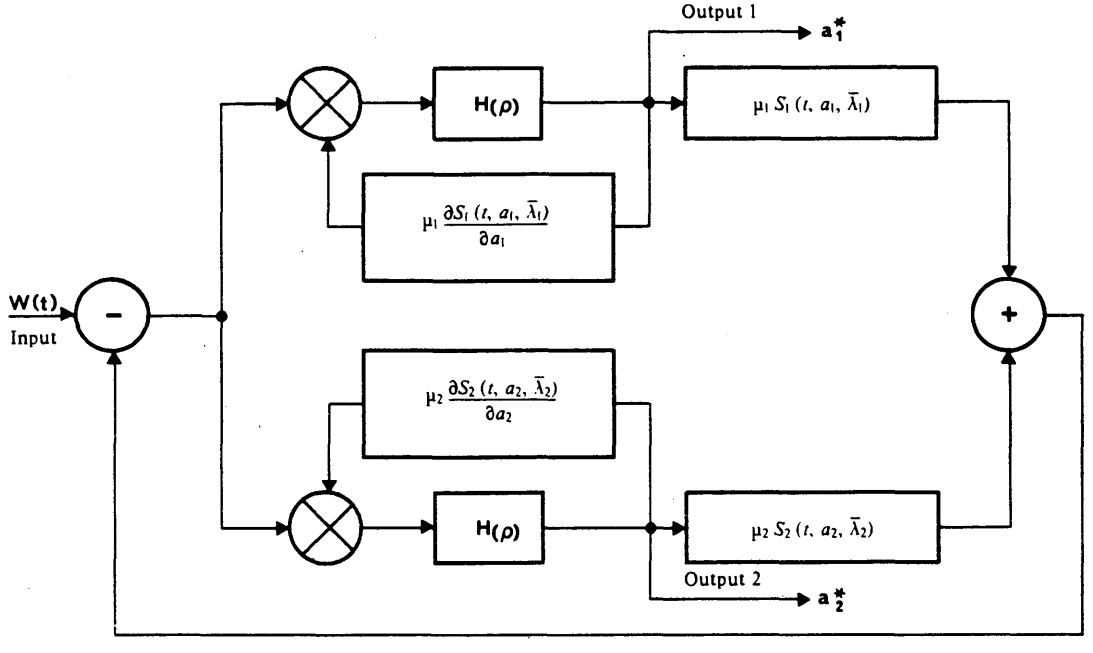


FIGURE 1 - *Optimum signal separation receiver with analogue modulation*

$H(\rho)$: linear filter

$\mu_i S_i (t, \bar{a}_i, \bar{\lambda}_i)$, $\mu_i \frac{\partial S_i (t, \bar{a}_i, \bar{\lambda}_i)}{\partial a_i}$: units forming the corresponding signals

2.2 *Test results*

Sundresh *et al.* [1977] gives the test results for the receiver in Fig. 3. The test conditions were as follows:

- the frequency of carriers was 450 kHz;
- modulation was produced by signals with a frequency of 100 Hz and 200 Hz;
- peak deviation 7 kHz;
- synchronous phase detector bandwidth of 200 kHz;
- low-pass filter bandwidth of 20 kHz.

Effective separation was evaluated as the ratio (χ) of the two modulating signals at the demodulator output.

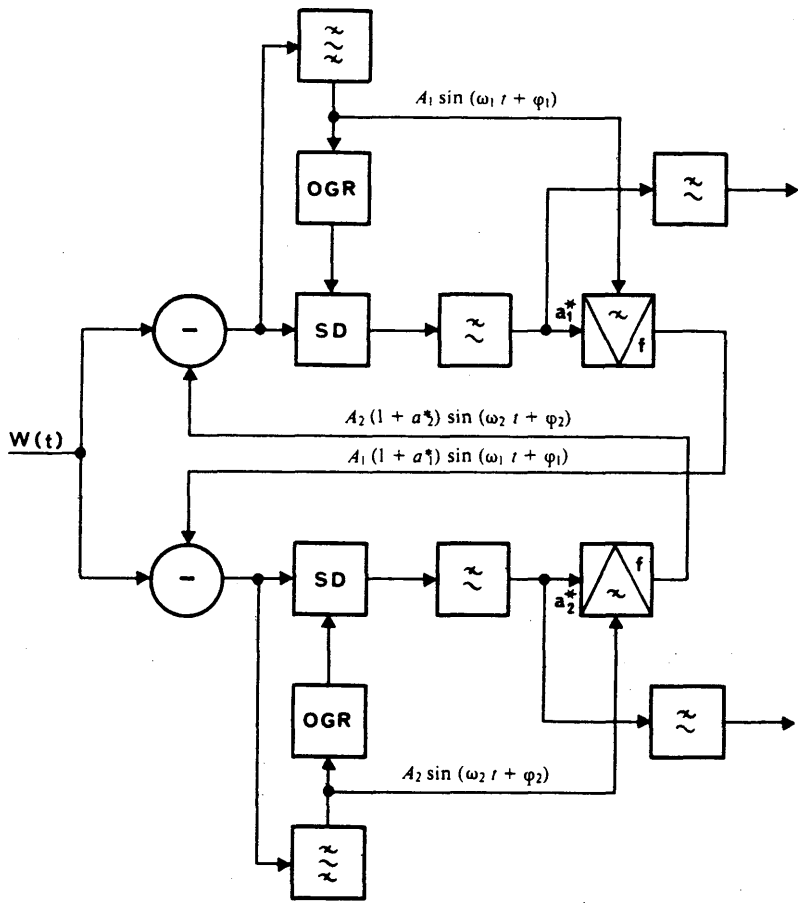


FIGURE 2 - *Optimum receiver with separation of 2 AM signals*

OGR: limiter
SD: synchronous detector

The test showed that the device reliably separates two FM signals if

$$\eta = \frac{\min_{i=1,2} \mu_i A}{\max_{i=1,2} \mu_i A} \in [0.1; 0.9]$$

The improvement in (χ) for the tracking interference canceller compared with a standard frequency detector separating a powerful FM signal was about 21 dB. Tests were conducted on the separation of two FM signals with modulation by voice signals and transmission of the signals from the outputs of the tracking interference canceller to the loudspeaker. The test showed the possibility of separating the signals at $\eta = 0.5$.

Bykhovsky [1980] proposes a device capable of separating two FM signals that employ synchronous orthogonal carriers. It contains a device (Fig. 3) supplemented by circuits for the mutual synchronization of the voltage-controlled oscillators forming part of synchronous phase detectors SPD_1 and SPD_2.

Only optimum single-channel AM and FM signal separators are discussed above. Numerous publications are now available which propose various non-optimum devices for the separation of AM and FM signals in a common bandwidth. Theoretical and experimental studies of these devices have shown that the use of an interference canceller can lead to a 20-50 dB reduction in interference. The use of such equipment enhances the electromagnetic compatibility of different radio systems operating in the same bandwidth.

3. Multi-channel noise cancellers

Multi-channel interference cancellers for communication systems in which, at each channel input, one or more sources of radio interference are present in addition to the wanted signal, are discussed below.

Report 828 discusses the advantages gained through the use of blanking in FDM systems.

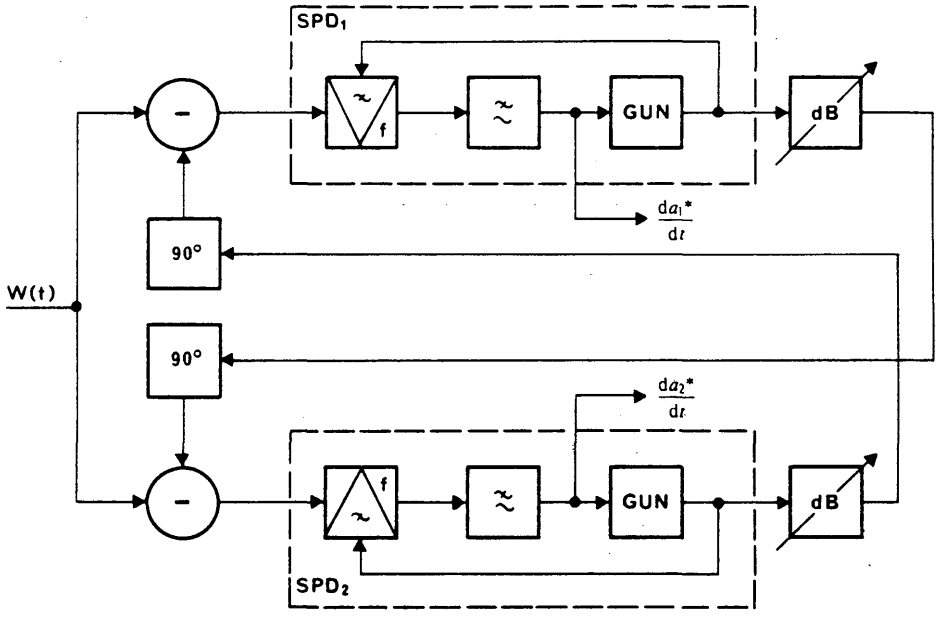


FIGURE 3 - *Optimum receiver with separation of 2 FM signals*

GUN: voltage controlled generator
SPD: synchronous phase detector

3.1 *Wanted and interfering signals at one input and interfering signal at second input*

Interference cancellers of this type are applied in satellite communication systems for protecting earth stations against interference from radio relay systems in the 4 GHz range [White *et al.*, 1975]. They substantially widen the freedom of choice of earth station sites, in many cases making it possible to use sites in the neighbourhood of towns. This considerably reduces the cost of installing lines linking the earth stations to communication centres in the town. In interference cancellers for earth stations, an auxiliary antenna pointed towards the source of interference and to the additional receiver connected with it is used. The auxiliary antenna's gain in the direction of the wanted signal must be insignificant, so that, at this input, the unwanted-to-wanted signal ratio is as large as possible (over 50 dB). This is due to the fact that the value of this ratio restricts the possibility of cancelling the unwanted signal on the main path [Widrow *et al.*, 1975]. Technically, an interference canceller may be designed either for RF or for IF or for a combination of both: partly for RF and partly for IF. It is obviously simplest to construct RF interference cancellers because it is easier to ensure the identity of the frequency and phase characteristics of the main and auxiliary receiving paths in a wide band of frequencies, which is essential for effective cancellation. Furthermore, the RF interference canceller is more easily inserted in the receiving paths of already operating earth stations.

Figure 4 is a block diagram of an RF interference canceller [White *et al.*, 1975]. In this system, low-noise (parametric) amplifiers are connected to the antenna outputs. Further along the main path, a directional coupler is connected which acts as a subtractor; the cancelling signal is fed into its second input from the automatic weight control device. This device contains a quadrature hybrid which forms four signals from the input signal: a co-phased and a counter-phased signal and signals at ± 90° to the input signal. Each of the branches of the quadrature hybrid is fitted with a pin diode attenuator. The attenuator outputs are summed, forming the outputs of the automatic weight control unit. Control voltages are produced at the attenuators by means of control

circuits. When the voltage at the synchronous demodulator output is positive (or negative), the first (or second) control circuit output signal is proportional to it; in the opposite case, it is zero. Signals reaching the synchronous demodulator direct or with a 90° phase shift are pre-amplified by RF and IF amplifiers and frequency converted by means of a general reference oscillator. The auxiliary path is equipped with a limiter required to reduce the dynamic range of voltage variation and to rule out the possibility of exciting the interference cancellers.

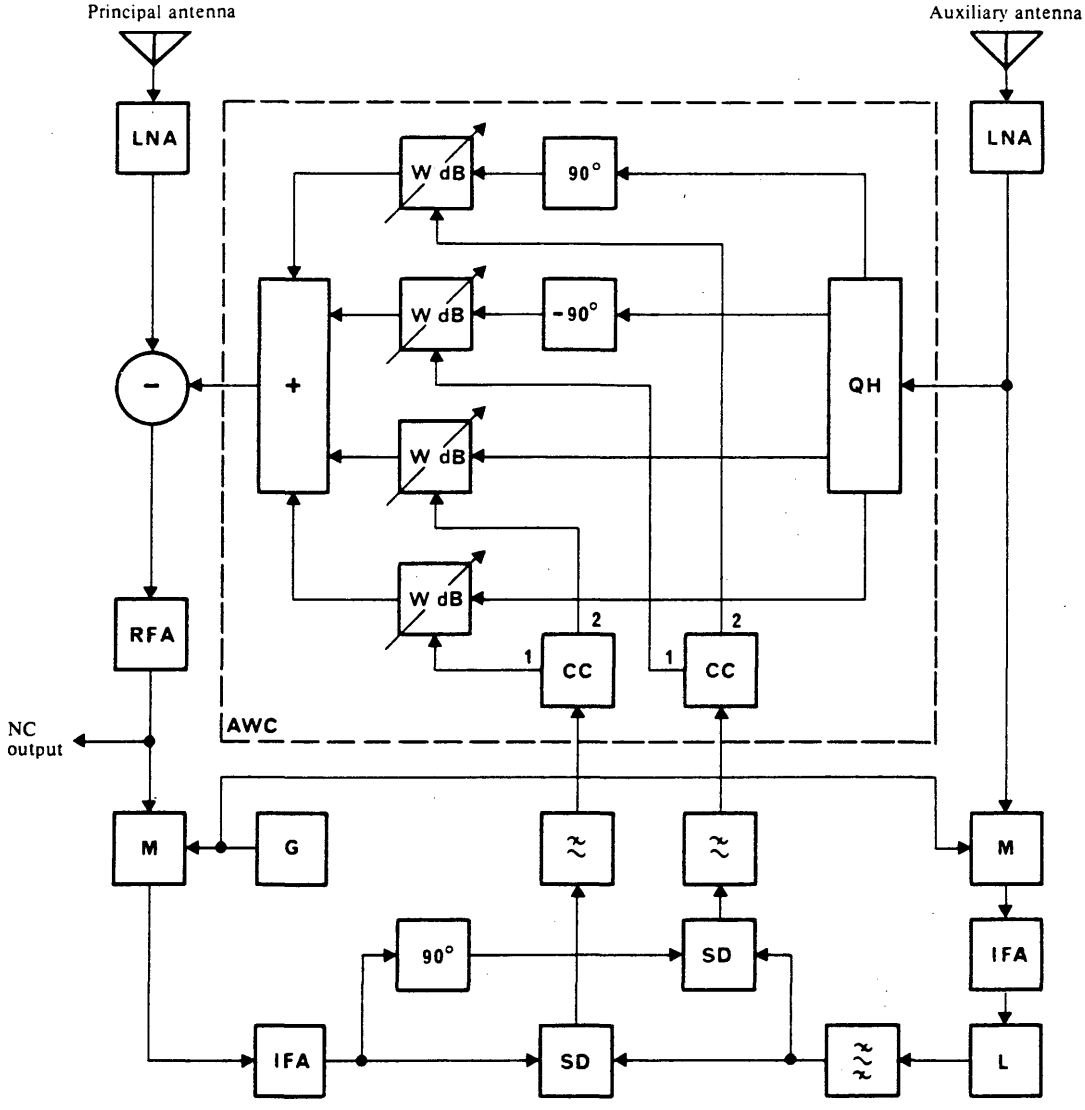


FIGURE 4 – *Two-channel RF interference canceller with reference input*

AWC: automatic weighting control unit
LNA: low-noise amplifier
L: limiter
QH: quadrature hybrid
SD: synchronous demodulator
M: mixer
RFA: RF amplifier
IFA: IF amplifier
CC: control circuit
NC: noise canceller
G: generator
W dB: variable attenuator

When the wanted and unwanted signals have the same carrier frequency, the interference canceller described above suppresses the interference by 40 to 50 dB. Detuning of the frequency of the unwanted signal by ± 10 to ± 15 MHz reduces its suppression to 25 dB. This reduction is due to the non-identity of the amplitude-frequency and phase-frequency characteristics of the amplifiers used and the difference in the lengths of the feeder lines of the main and auxiliary paths.

Interference can be suppressed by 40 dB within the occupied bandwidth when very strict criteria are imposed for the identity of the receiving path characteristics, i.e. the amplitude-frequency characteristics differ by not more than 0.1 dB, and the phase-frequency characteristics by not more than 0.6°. The effectiveness of the suppression of wideband interference is also affected by the frequency-dependence of the antenna gain in the direction of the interference.

3.2 *Wanted and interfering signals at both inputs*

It is advisable to use an interference canceller of a second type, for example, at the intermediate stations of radio relay systems. Owing to the difficulty involved at these stations in obtaining a sufficiently low level of the antenna lobes in the backward direction, it is extremely difficult to obtain decoupling between the RF channels received from opposite directions, if they are transmitted on the same frequency. In a number of cases (for example, when using periscope antennas) this involves the use of a four-frequency plan for radio relay. The use of an interference canceller on such radio relays makes it possible to switch to a two-frequency plan, thus doubling the number of RF channels on which wanted information can be transmitted in either direction.

The possibilities of separating two FM signals by means of optimum or sub-optimum interference cancellers are considered in [Bykhovsky, 1979c], which shows that there is practically no increase in interference when FM signals are separated by means of an interference canceller if the crosstalk attenuation between the channels is not more than 10 to 15 dB. With separate demodulation of the signals in each channel, an interchannel crosstalk attenuation of more than 50 to 60 dB is normally required.

3.3 *Signals with orthogonal polarization*

In this case transmissions use orthogonal polarization, where each of the signals is modulated independently by its wanted message; these signals have to be separated at the receiving end.

The capacity of radio relay and satellite communication systems can be doubled without increasing the occupied bandwidth by using orthogonal polarization to transmit the two RF channels. Propagation studies in the 4 to 12 GHz band have shown that radio waves are depolarized by precipitation on the path, which causes considerable interchannel crosstalk. Since the level of crosstalk fluctuates, its elimination calls for the use of adaptive interference cancellers of a third type.

A wide variety of such devices have been proposed for satellite systems [Baird and Rassweiler, 1977]. In these systems, the particular harmonic pilot signal needed to adapt the interference canceller to the changing level of interchannel crosstalk is transmitted in each channel.

Figure 5 shows the diagram of a matrix-type interference canceller [Baird and Rassweiler, 1977] in which a polarization selector for separating radio waves with orthogonal polarization is fitted to the antenna output. In addition to the wanted signal, crosstalk occurs in the channels at the polarization selector outputs. After amplification in a low-noise amplifier, the signals in both channels reach the matrix unit inputs equipped with automatic weight control (AWC) units and subtractors. The AWC units comprise controlled attenuators and phase inverters. The control signals for these devices are produced in a special control unit connected to the interference canceller output. The control unit is shown in Fig. 6. Narrowband filters Φ_i for separating the pilot signals are connected to its inputs. Owing to the interchannel crosstalk, the residual pilot signal of the first (second) channel will act on the second (first) interference canceller output. The control unit regulates the controlled attenuators and the phase inverters in such a way as to ensure minimum pilot signal power in the second channel at the interference canceller output. For this purpose, the Φ_i filter outputs are connected to the synchronous detectors producing the control voltage R_i and Q_i either directly or through the phase inverter at 90°. Kreutel *et al.* [1977] describes other possible interference canceller and control unit assemblies. One disadvantage of the matrix-type interference canceller is the comparatively narrow frequency band within which a significant cancellation of crosstalk is achieved. This is due to the difficulty of obtaining identical amplitude-frequency characteristics of the low-noise amplifier in a wide frequency band. It has been shown in the literature that when the group delay in the low-noise amplifier approaches ± 0.5 ns, 20 dB interference suppression can be obtained only in a 30 MHz band, and in a 160 MHz band only by 6 dB.

Since the RF channel bandwidth in satellite communication systems is 500 MHz, wideband interference cancellers in which all the necessary operations are performed by waveguide components, have been developed in order to suppress crosstalk between channels in a wide frequency band using a single interference canceller.

Research with an experimental interference canceller in the 12 GHz band [Baird and Rassweiler, 1977] has shown that with an initial crosstalk attenuation between separated channels of 20 dB, the interference canceller increases the attenuation to 36 dB in a frequency band of 700 MHz. In the experimental system, the pilot signals were located at the edges of the usable bandwidth (f_1 = 11.7 GHz and f_2 = 12.5 GHz). For frequency channels separated by not more than 250 MHz, the interference canceller increased the crosstalk attenuation by 30 dB.

The design of adaptive antenna arrays (AAA) is discussed in a number of works [Kreutel *et al.*, 1977; Gabriel, 1976]. Two types of AAA have been used in communication systems.

In the first type of AAA, all K elements form K fixed and orthogonal radiation patterns by means of matrix-type pattern-forming circuits (e.g. Butler circuits, Davis circuits, etc.). The peculiar feature of these patterns is that in the direction of the maximum of any of the patterns all the others have a zero value. The canceller circuit is shown in Fig. 7. One of the outputs of the pattern-forming circuit, which is the main output and which corresponds to the radiation pattern having a maximum in the direction of arrival of the wanted signal is used as a reference output for adaptive control in the AWC devices by means of complex weighting factors which compose the other outputs and form a replica of the interference present in the main receiving channel. This replica is fed to the subtractor for interference cancelling. A high-gain amplifier is inserted in the cancelling circuit to minimize the interference level at the interference canceller output.

Results of an experiment with a similar type of interference canceller have shown the possibility of increasing the signal-to-interference ratio at the interference canceller output by 40 dB.

The second type of AAA does not exploit previous information about the direction of arrival of the wanted signal. The AAA is tuned by a special pilot signal introduced during transmission; this signal is separated at the AAA output and tunes its radiation pattern to the maximum in the direction of arrival of the wanted signal. These interference cancellers are described [Hansen, 1977; Compton, 1978]. The experimental results reported in these works show the possibility of suppressing interference by 30 to 50 dB.

The interference cancellers described in this Report are therefore fully capable of solving electromagnetic compatibility problems in cases where, for one reason or another, the traditional methods of solving these problems by means of frequency or geographical separation between mutually interfering systems are either not feasible or ineffective.

Interference cancellers provide one possible solution to the problem of protecting earth stations from radio-relay interference. On radio relay links, this means substantially less strict criteria for the siting of receiving antennas. In particular, in radio relay systems with periscope antennas, which normally use a four-fold frequency plan, interference cancellers make it possible to switch over to a two-fold frequency plan, thus doubling the number of channels available for the transmission of wanted information.

Interference cancellers also make it possible to transmit information on two channels occupying a common frequency bandwidth, using orthogonal polarization, owing to the substantial reduction in interchannel crosstalk. This means doubling system capacity without increasing the occupied bandwidth.

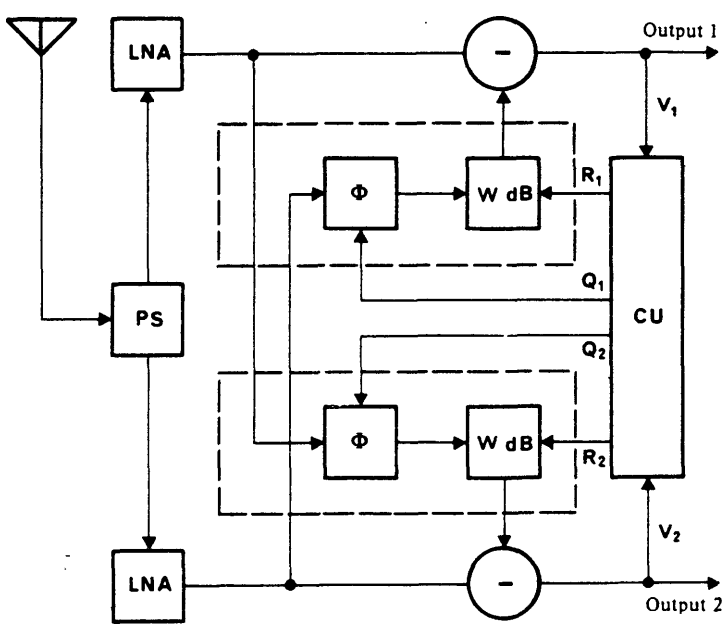


FIGURE 5 - *Polarization interference canceller*

CU: Control unit
LNA: low noise amplifier
PS: polarization selector

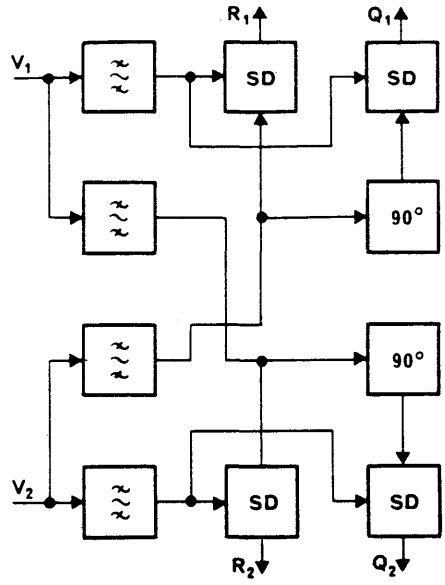


FIGURE 6 - *Control unit*

SD: synchronous demodulator

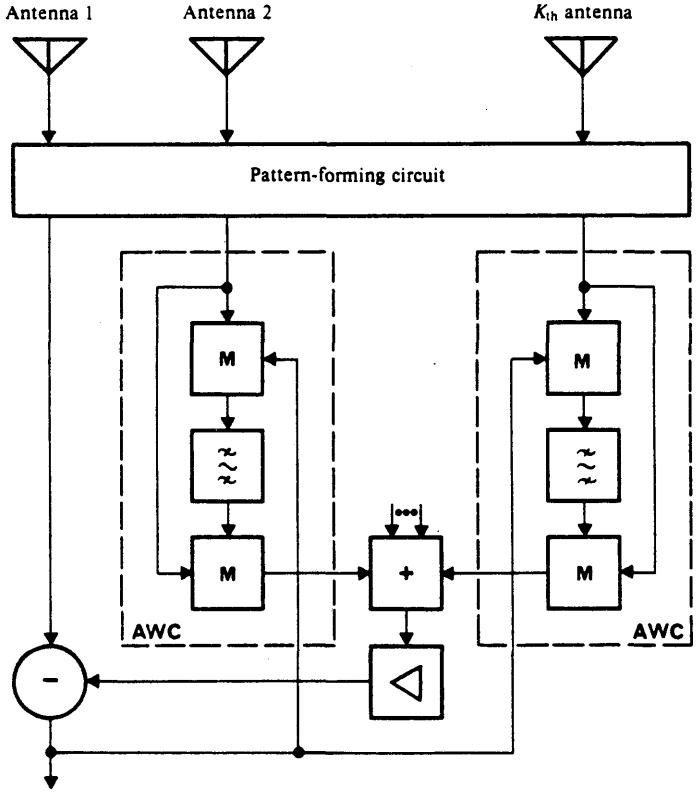


FIGURE 7 - *Interference canceller with adaptive antenna array*

AWC: automatic weighting control unit
M: mixer

4. Evaluation of spectrum efficiency of interference cancellers

Interference cancellers may be used in terrestrial radio services, such as radio-relay links operating in the SHF band, to reduce the geographical or angular separation between interacting stations of radio-relay links and also to reduce the permissible angular separation between geostationary satellites operating in systems of the fixed-satellite service.

The effect of the degree of interference suppression obtained by cancellers (Δ dB) on geographical separation may be approximately evaluated by the coefficient of the reduction of necessary geographical separation $K_r = R_\Delta / R_0$, where R_0 is the geographical separation necessary between the stations without cancellers and R_Δ the distance between these stations when a canceller, with a degree of suppression Δ, is used. To determine K_r, an expression has been used linking R_0, R_Δ and Δ:

$$L(R_0) = L(R_\Delta) + \Delta$$

where:

$L(R_0)$: losses in signal propagation on a path of length R_0,

$L(R_\Delta)$: propagation losses on a path of length R_Δ.

Figure 8 shows curves plotted from the data in Kalinin [1979] and representing L (R), the signal propagation loss over a smooth spherical Earth surface on a path of length R, as a function of various frequency bands. The solid curves relate to median losses and the broken curves to losses observed for 99.9% of the time [Bykhovsky and Timofeev, 1983].

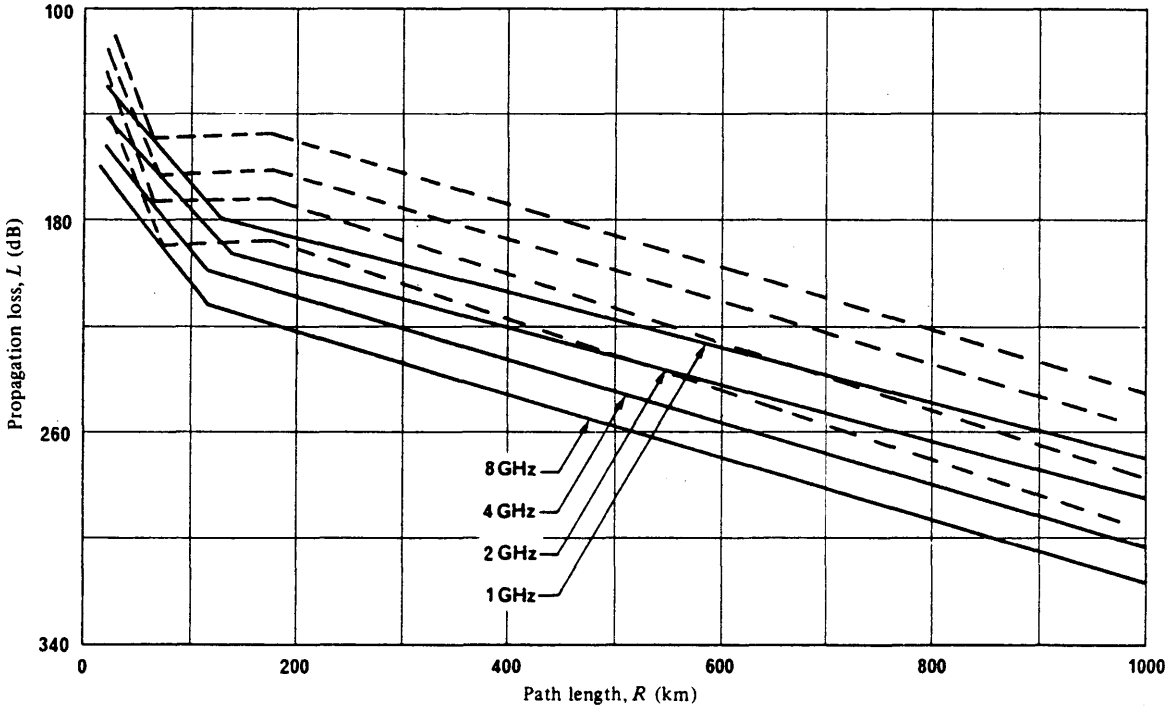


FIGURE 8 – *Propagation losses as a function of path length and frequency*

The steep slope of the curves in Fig. 8 in the range of low R values is due to propagation by diffraction and the shallow slope for large R values to the tropospheric scatter mechanism.

The rise in the broken curves at $80 \leq R \leq 200$ is due to the fact that in this range there is a maximum dispersion of propagation losses, owing to the combined effect of both diffraction and tropospheric scatter propagation mechanisms.

On the basis of the solid curves in Fig. 8 for frequency 1 GHz, the coefficient K_t for various values of R_0 are plotted against Δ in Fig. 9. These relations may be used to evaluate the effectiveness of cancellers in radio-relay links operating in the 1-8 GHz range. As is clear from Fig. 9, the dependence of K_t on Δ is non-linear with the result that the efficiency of a canceller providing a given degree of suppression Δ will differ according to the initial conditions.

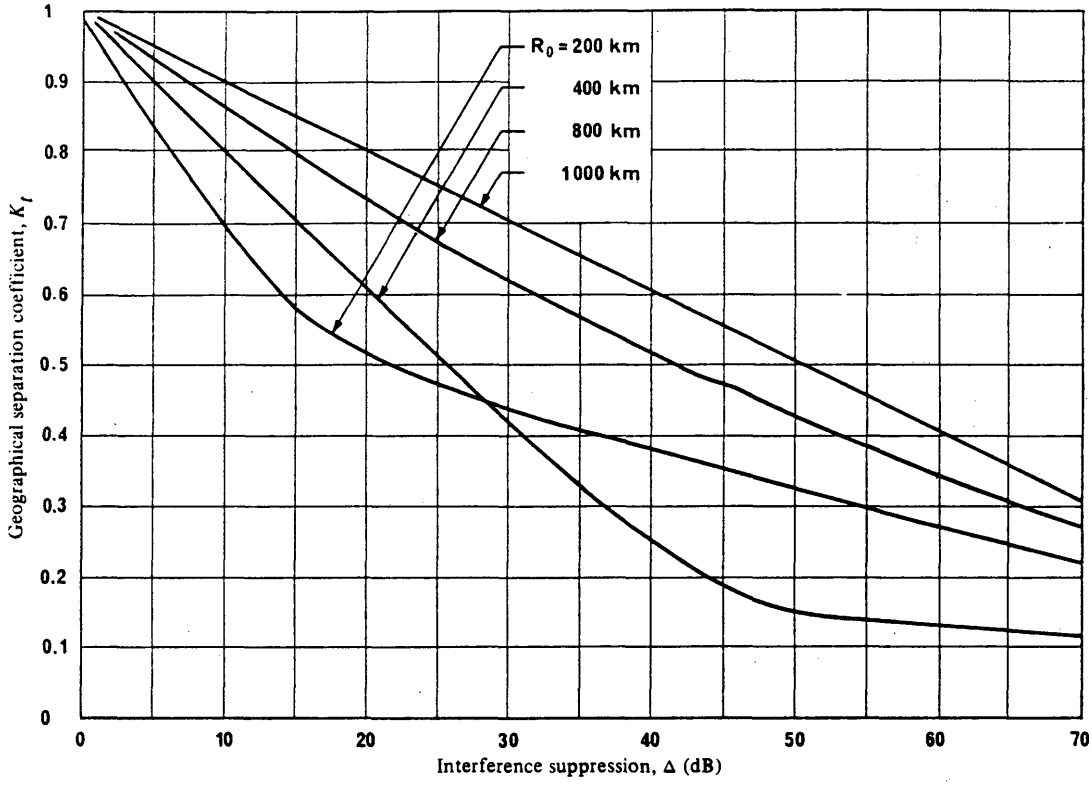


FIGURE 9 – *Coefficient K_t as a function of Δ for f = 1 GHz*

The curves in Fig. 9, for example, show that with low R_0 values a substantial increase in the degree of interference suppression, Δ, by cancellers serves no useful purpose, since it does not appreciably reduce the geographical separation required. This is due to the fact that at low R_0 values, the interference propagation path lies in the diffraction zone, in which propagation losses are very quickly reduced by decreasing the path length.

The effect of the degree of interference suppression, Δ, on the necessary angular separation is evaluated by calculating the coefficient $K_\theta = \theta_\Delta / \theta_0$, where θ_0 and θ_Δ are the angular separations required to ensure electromagnetic compatibility of the interacting systems with and without interference cancellers.

The antenna gain coefficient may be characterized generally by the expression $G(\theta) = \beta - 25 \log \theta$. The values θ_0, θ_Δ and Δ are linked by the obvious correlation $G(\theta_0) = G(\theta_\Delta) - \Delta$, hence $K_\theta = 10^{-\Delta/25}$.

It follows from the above expression that the use of a canceller with $\Delta \approx 8$ dB allows the angular separation to be halved and with $\Delta = 25$ dB the separation can be reduced to a tenth.

The expression for K_θ may also be used for the approximate evaluation of the extra capacity of a geostationary orbit resulting from the use of interference cancellers. In this case, a model is used which is represented by homogeneous satellites evenly distributed over the orbit, at angular separations of θ_Δ with interference cancellers and of θ_0 without. An increase in the number of satellites causing interference naturally complicates the design of the cancellers. The value K_θ therefore caracterizes only the theoretical evaluation of the maximum improvement offered by interference cancellers. Clearly, however, even a relatively slight degree of interference suppression, e.g. $\Delta = 8$ dB, leads to a considerable increase (doubling) in orbit capacity.

The above analysis shows that interference cancellers may serve as a very effective means of ensuring electromagnetic compatibility in radio-relay links and satellite systems.

5. Pulse interference cancellers using spectrum transformation and signal limitation

The simplest device of this type [Gorbachev, 1958] is illustrated in Fig. 10. The first linear spectrum transformer LT_1 has a transmission factor which increases with frequency and essentially constitutes a differentiating circuit for the spectral components of the wanted signal. It attenuates the level of the wanted signal reaching the input of the limiter L. At the same time, its parameters are selected such that short impulsive noise passes through it without any noticeable attenuation. The limiter has the following response:

$$f(X) = \begin{cases} X & \text{for } |X| \leqslant U_0 \\ U_0 & \text{for } |X| > U_0 \end{cases}$$

The level of limitation U_0 is selected such that the attenuated wanted signal is not limited. The second linear spectrum transformer LT_2, which has parameters selected to be the opposite of the first, essentially constitutes an integrating circuit. When interference occurs, this circuit gives way to limitation by the limiter L and integration by the second linear transformer LT_2.

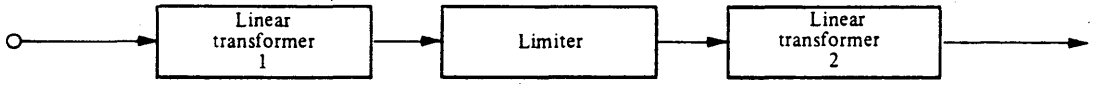


FIGURE 10 – *Pulse interference canceller*

Atakhanov *et al.* [1971] describe another device designed to suppress impulse noise of significantly shorter duration than the wanted signal correlation time. A block diagram of the device is given in Fig. 11.

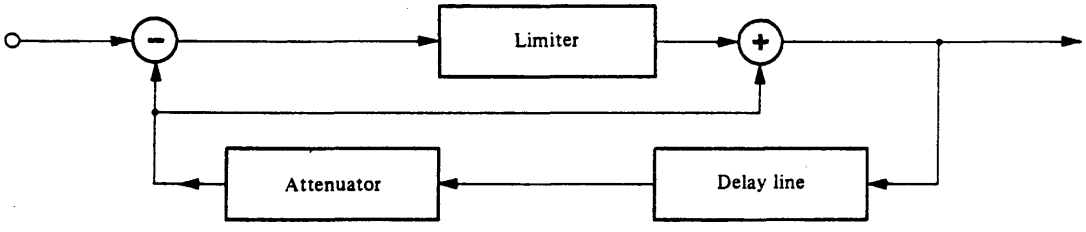


FIGURE 11 – *Short-term pulse interference canceller*

The device contains a negative feedback circuit incorporating the delay line DL at a time, τ, equal to the duration of the interfering signal pulse. The limiter L has the same response as the previous device. Owing to the negative feedback circuit, the output signal, delayed by τ and weighted by the attenuator A, is extracted from the input signal in the first adder. Because of the significant correlation of the instantaneous values of the wanted signal at the moments of time t and $t + \tau$ in the absence of impulsive noise, the difference signal reaches the limiter L at a level substantially lower than the level of the wanted signal reaching the input stage of the device. If the limitation threshold selected is higher than the level of this difference signal, then the signal proceeds to the output without limitation. Under the influence of short impulsive noise on the other hand, it reaches the limiter unattenuated and is limited. This makes it possible to restore the wanted signal with a relatively high degree of accuracy.

The device described in Fig. 11 can be altered as shown in Fig. 12. In this case, an amplifier with a sufficiently high gain and a narrow-band filter are connected to the output of the limiter L. When impulsive noise reaches the limiter input, it goes into the saturation region and disconnects the negative feedback circuit. Thus the passband of the device changes suddenly to match the passband of the narrow-band filter, the output voltage of which does not change significantly owing to the lack of response time during the interference effect.

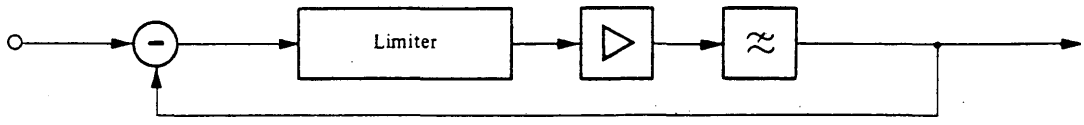


FIGURE 12 – *Short-term pulse interference canceller*

The advantage of the devices in Figs. 11 and 12 over the first device is that they do not require precise correlation of the frequency and phase responses of the two spectrum transformers over a wide band of frequencies.

These devices are the simplest known and may be used to eliminate short-term impulsive noise in communication systems with a small number of channels. As pointed out by Bykhovsky [1985], they are far superior to the well-known "blanking" method which is widely used in single channel systems, owing to their effectiveness in suppressing short-term impulsive noise. They are also simpler to construct since they do not require a special impulsive noise detector or a high-quality blanking switch.

6. Pulse interference cancellers with extrapolation or interpolation of the affected parts of the signal

Interference cancellers with extrapolation or interpolation of the affected parts of the signal are even more effective than the devices described in the previous section.

Figure 13 shows the block diagram of a device [Kataev and Polykovsky, 1955] in which an extrapolator E is inserted on a delay line DL with several outlets. The extrapolated signal $a_e(t)$ is formed by summing the signals $a(t - i\tau_d)$, acting at the outlets of the delay line at the preceding moments in time and weighted by means of attenuators A, i.e.:

$$a_e(t) = \sum_{i=1}^{n} \alpha_i \, a(t - i\tau_d)$$

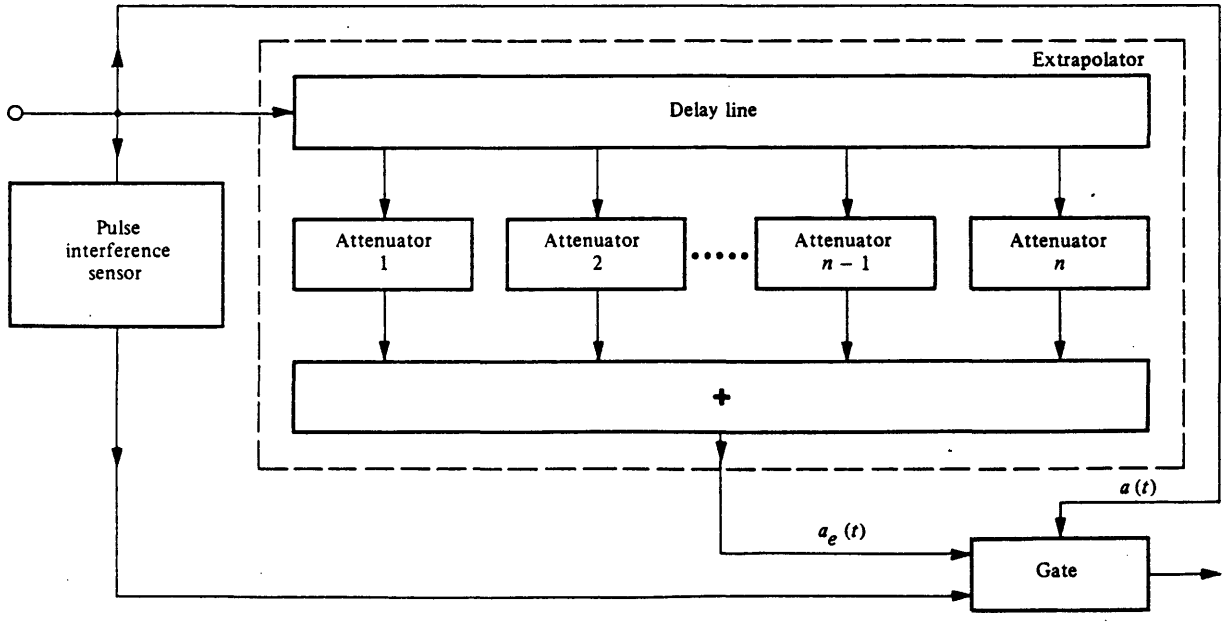


FIGURE 13 – *Pulse interference canceller with signal extrapolation*

The delay τ_d between the outlets of the delay line should just exceed the duration of the impulsive noise τ. The input $a(t)$ and extrapolated $a_e(t)$ signals reach the gate G, which sends the substitute extrapolated signal to the output stage during the impulsive noise effect. The gate is controlled by an impulsive noise sensor IS. By selecting appropriate transmission factors α_i for the attenuators A_i, a high degree of correlation between the input $a(t)$ and extrapolated $a_e(t)$ signals may be obtained. The input signal interpolation device (Fig. 14) also uses a delay line with outlets DL, but, as opposed to the extrapolators, the substitute signal is formed in the interpolator I as the weighted sum of not only the preceding values of the signal $a(t)$ but also the subsequent values, i.e.:

$$a_I(t) = \sum_{\substack{i=-n/2 \\ i \neq 0}}^{n/2} \alpha_i \, a(t - i\tau_d)$$

Interpolators restore the affected parts of the wanted signal more accurately than extrapolators. It should be pointed out, however, that a high degree of accuracy in restoring the wanted signal may be obtained only at a sufficiently low frequency of repetition of the interfering signal.

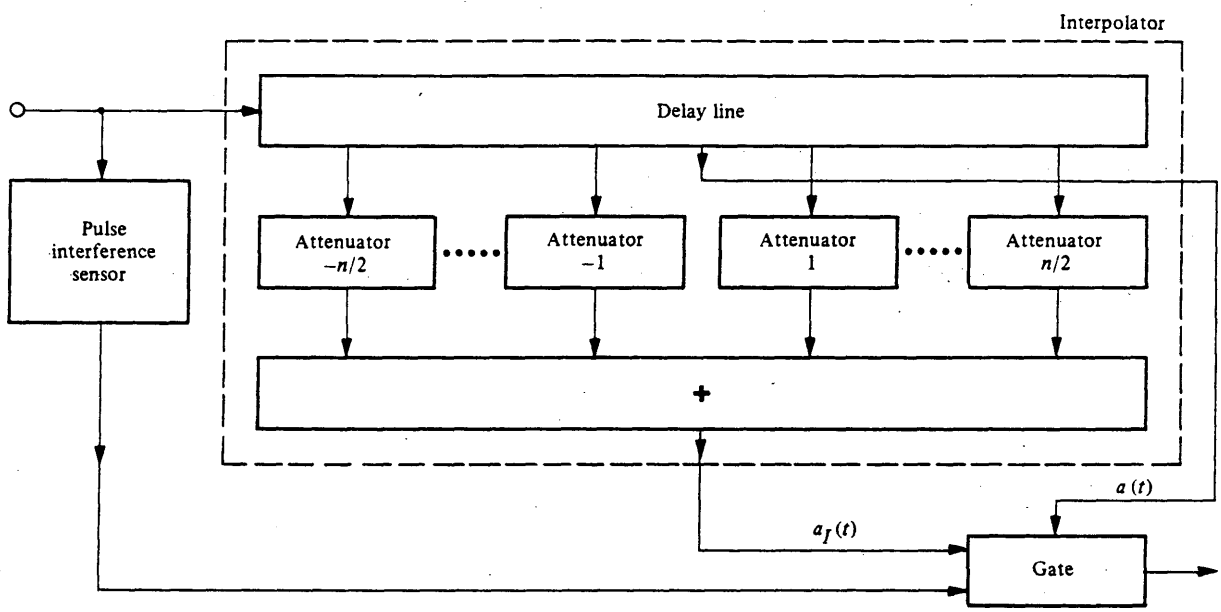


FIGURE 14 – *Pulse interference canceller with signal interpolation*

Figure 15 illustrates the block diagram of a device [Gorbachev and Vinogradov, 1961] using a different method to extrapolate the signal. The input to this arrangement is made up of n series-connected differentiating circuits DC. The output of the last of these circuits is connected via a gate G controlled by an impulsive noise sensor IS to the input of the extrapolator E which is made up of n series-connected integrating circuits IC with opposite frequency responses to those of the differentiating circuits DC. It may be demonstrated that this arrangement extrapolates the wanted signal to the n^{th} partial sum of a Taylor series.

The shortcoming of this device is the complexity involved in matching the transmission characteristics of the differentiating and integrating circuits over a wide frequency band.

The effectiveness of the methods described above has been considered by Bykhovsky [1984].

Madan *et al.* [1978] propose an optimum impulsive noise canceller for reception of signals with phase modulation, the block diagram of which is shown in Fig. 16. An impulse noise probability counter PC, which controls the gain in the feedback circuit, is inserted in the usual automatic frequency control circuit comprised of an integrator E, an amplifier, a switchable generator SG and a phase detector PD. The integrator E, which is, in this case, an extrapolator, maintains the voltage at its output if it receives no signal at its input. The transmission factor of the negative feedback circuit is controlled by the PC, which deletes any detected impulsive noise, and the extrapolator E ensures that the substitute voltage is delivered to the output.

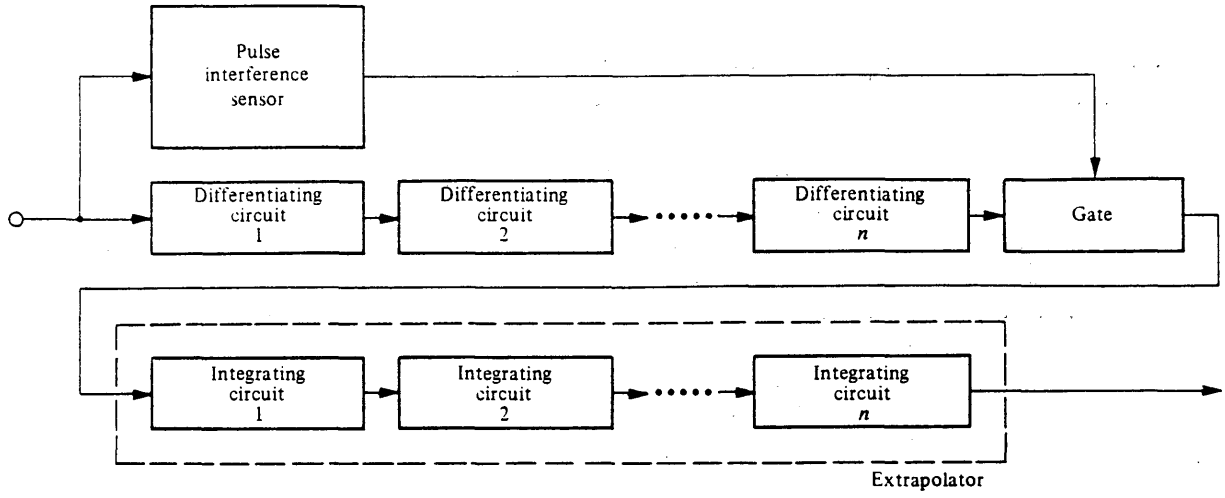


FIGURE 15 – *Pulse interference canceller with signal extrapolation*

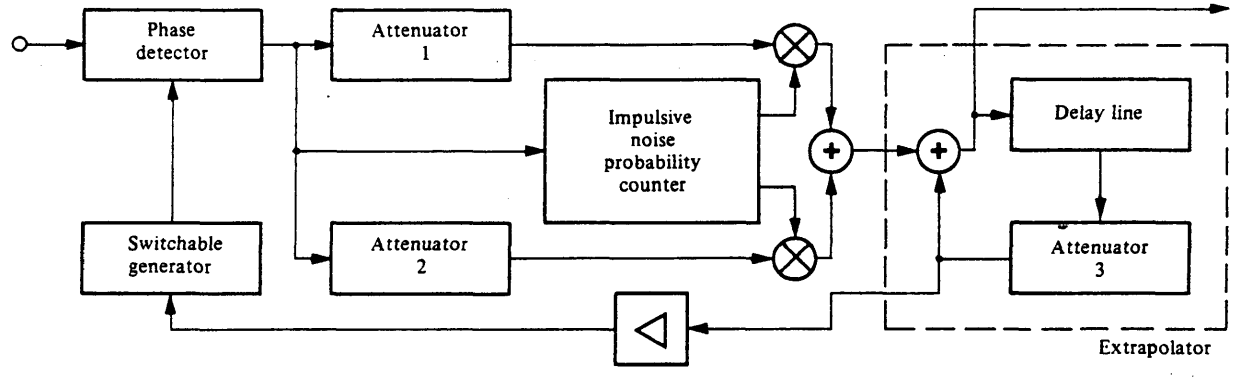


FIGURE 16 – *Optimum pulse interference canceller for phase modulation*

Various types of impulsive noise cancellers may be designed according to the specific features of the communications systems involved. In particular, Bernoskuni *et al.* [1984] propose an effective method of eliminating impulsive noise in communication systems using frequency-diversity reception, in particular for tropospheric systems. The method, which is simple to apply, restores the affected signal for virtually any length of interfering pulse and in many cases almost completely eliminates the effect of interference on the signal reception noise immunity.

A simplified block diagram of a device of this type is given in Fig. 17. The same multi-channel message is transmitted over n radio channels (which by convention are shown in Fig. 17 as n transmitters T and n receivers R) connected to two transmission and reception channels operating with two space-diversity antennas. A time delay τ_d exceeding the maximum possible impulsive noise duration is introduced into one transmission channel by the delay line DL_1, and on the receiving side a similar delay τ_d is inserted in the second channel by the delay line DL_2. This means that in the absence of impulsive noise, the transmitted message may be restored by combining all the n received channels coherently. When impulsive noise is detected, for the duration of the noise, the interference sensor IS either immediately or with a delay τ_d produced by delay line DL_3, isolates the receiving channel in which the message is affected by the interference by means of gates G_1 and G_2. The transmitted message is then restored by coherently combining the signals only in the receiving channel where there is no impulsive noise at that time. The effectiveness of this method, which is fundamentally similar to the extrapolation and interpolation methods of restoring the affected signal described above, is discussed in Report 828.

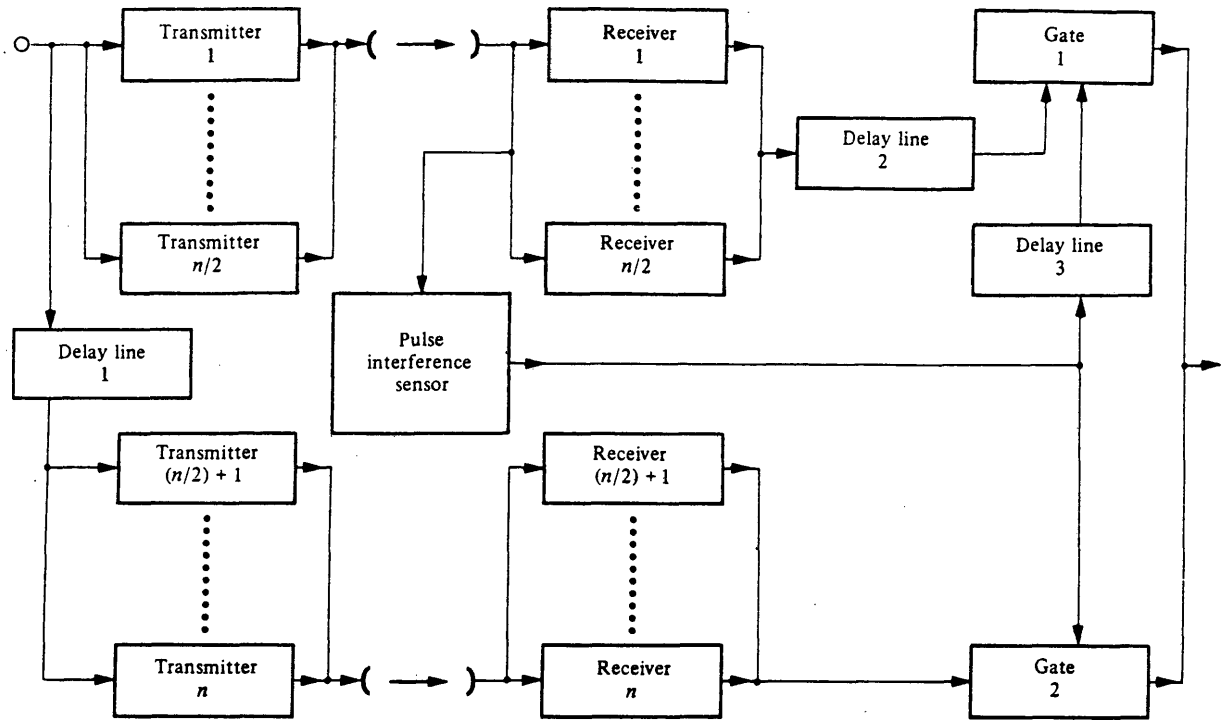


FIGURE 17 – *Pulse interference canceller for frequency-diversity systems*

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REPORT 831-1*

USE OF SPECIAL SCREENS TO IMPROVE INTERFERENCE IMMUNITY IN RADIO LINKS

(Question 18/1)

(1982-1990)

1. Introduction

While radiating the wanted signal in the main lobe, an antenna also radiates part of the energy in all other directions. This side-lobe radiation is a source of interference to other radio receivers.

This Report discusses the possibilities for reducing mutual interference between the antennas of different links operating in the same frequency bands by suppressing the interfering signal with electromagnetic radiation screens installed at a given distance from the antenna to be protected (outlying screens) or mounted directly on the antenna.

The Report also discusses the problem of improving the interference immunity of antennas in radio links in the rear back space by using devices mounted directly on the antenna.

2. Screens for protection against interference from low-directivity antennas

The method applied in the design of small high-performance screens is based on the definition of the shape of the screen edge for which field zero at the calculation point is obtained in the Kirchhoff approximations [Melnikov, 1976].

In using screens, care must be taken that the energy scattered by the screen does not cause interference elsewhere.

The field of source S at point P is calculated by summation of the fields from the unblanked portions of the concentric rings (Fig. 1). For the ring radius r and its aperture angle φ' we have the parametric equation which assures the field is zero at the calculation point.

* This Report should be brought to the attention of Study Group 9.

$$r = \sqrt{\frac{\lambda_0 r_0 R_0 \arctan \dfrac{\rho' \sin\psi + \rho\cos\psi}{\rho'\cos\psi - \rho\sin\psi}}{\pi (r_0 + R_0)}}$$

$$\varphi' = 2\pi \frac{(\rho^2 + \rho'^2)^{3/2}}{\rho^2 + 2\rho'^2 - \rho\rho''} \cdot \frac{1}{\rho_{max}} \qquad (1)$$

The parameters r, R_0, r_0, φ' in the parametric equation (1) are given in Fig. 1, and λ_0 is the wavelength of the incident wave.

Inserting the functional relation $\rho = f(\psi)$ into equation (1), we obtain the corresponding shape of the screen contour.

Calculations and experiments were carried out for two types of function $\rho = f(\psi)$: $\rho = \sin a\psi$ and $\rho = 1 - e^{-a\psi}$. Figure 2 shows the screen configurations when a = 1.0, 1.4 and 2.0. The shaded region is the shape of the screen and the circle is the first Fresnel zone.

The screen for $\rho = \sin a\psi$ when $a = 1$ has the simplest configuration. The smallest value results from screens with $a = 2$.

Figure 3 shows the radiation patterns of a screen irradiated by a plane wave in the suppression area for $\rho = 1 - e^{-2\psi}$. The patterns were calculated for the planes XOZ and YOZ and are designated as $F(V_1)$ and $F(V_2)$ respectively, where:

$$V_1 = \frac{x}{\sqrt{\lambda_0 r_0}}, \qquad V_2 = \frac{y}{\sqrt{\lambda_0 r_0}} \qquad (2)$$

λ_0 : mean wavelength of working bandwidth,

r_0 : distance from screen to calculation point with a zero field value.

The values of $F(V_1)$ and $F(V_2)$ are expressed in dB relative to the unperturbed field level. The experimental data are indicated by the circles on Fig. 3.

Curves of $F(V_1)$ and $F(V_2)$ are plotted on Fig. 4 to permit evaluation of the waveband properties of the screen. The hatched sections indicate the region in which the suppression is satisfied in the range ± 10% about λ_0.

To obtain a wider region of suppression and increased screen bandwidth, two axes of symmetry must be used [Frolov, 1974]. As an example, we can take a screen with an edge shape described by the equation $r(\varphi) = \sqrt{a^2 + \dfrac{4\varphi}{\pi} r_0 \lambda_0}$ (Fig. 5), where φ is the angular coordinate in radians and a is a constant with a value selected in the range (1.5-2.0) λ_0.

The theoretical radiation patterns are given in Fig. 6.

These screens are suitable for the protection of low-gain antennas (20-25 dB) or for the feeds of large antennas, such as those used in transhorizon or space links.

The "ring" screen (Fig. 7) [Melnikov, 1975a; Lokshin *et al.*, 1975] is proposed for the protection of higher-gain antennas (around 30 dB). The annular shape permits further reduction in screen dimensions.

In accordance with the principle of the "ring" screen compensation of the diffraction fields from the outer edge of the ring, by a field passing through the aperture, is provided.

The radii of the ring are defined by the formulae:

$$\rho_1 = \sqrt{\frac{1}{3}}\rho, \qquad \rho_2 = \sqrt{\frac{2}{3}}\rho, \qquad \rho = \sqrt{\lambda_0 r_0} \qquad (3)$$

where:

λ_0 : theoretical wavelength,

r_0 : distance from screen to calculation point with zero field value (i.e. up to protected antenna).

Figure 8 contains radiation patterns of the ring in the suppression region for the band $(0.9\text{-}1.1)\lambda_0$. The graph is symmetrical relative to the axis $F(V)$, so that the figure shows only the right-hand side of the graph. As in the previous case, $V = \frac{x}{\sqrt{\lambda_0 r_0}} = \frac{y}{\sqrt{\lambda_0 r_0}}$. The screen creates an axisymmetrical suppression region around the calculation point. The greater the distance from the calculation point, the lower the suppressive properties of the screen.

The dimensions of the ring (ρ_1, ρ_2), its distance from the calculated suppression point (r_0) at which the antenna to be protected is located, and also the radius of the suppression region (r) at the edge of which a given level of suppression is assured, are interrelated by:

$$\rho_1 = \sqrt{\frac{1}{3}}\sqrt{\lambda_0 r_0} = 0.576\frac{r}{V}, \qquad \frac{r}{V} = \sqrt{\lambda_0 r_0}$$

$$\rho_2 = \sqrt{\frac{2}{3}}\sqrt{\lambda_0 r_0} = 0.815\frac{r}{V}, \qquad r_0 = \frac{1}{\lambda_0}\left(\frac{r}{V}\right)^2 \tag{4}$$

In this case, r is practically the radius of the antenna to be protected. The value of r and the level of the necessary field suppression at the antenna edge are known in advance. The value of V is first determined on the basis of the suppression pattern for a given suppression level, after which ρ_1, ρ_2 and r_0 are calculated by formulae (4).

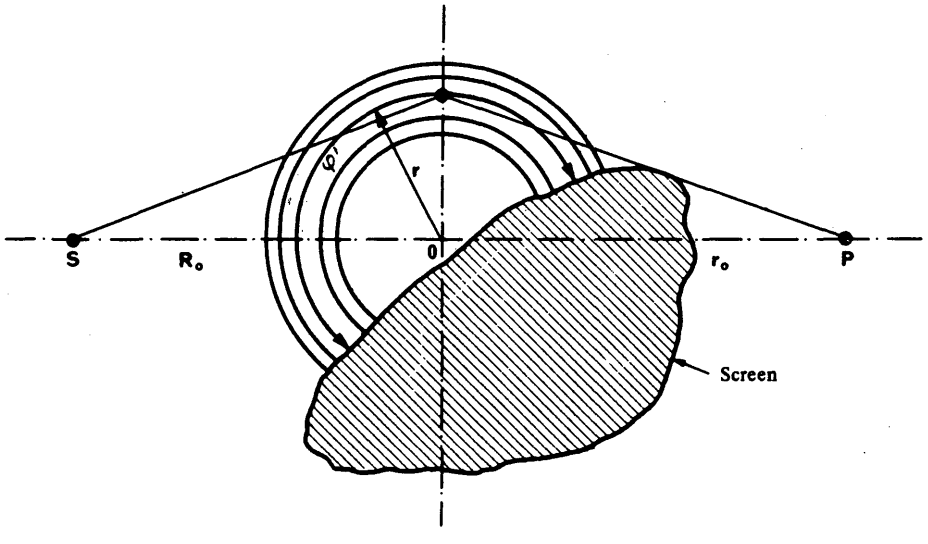


FIGURE 1

3. Screens to protect high-gain antennas

The screens discussed above have the relatively narrow suppression pattern. Therefore, such screens must be generally larger than the antenna and located at a considerable distance from the antenna.

To widen the screen suppression region and reduce r_0, the "double-ring" screen may be used [Melnikov, 1978] (Fig. 9), the dimensions of which are given by the formulae:

$$\rho_1 = 0.388\ \rho,\quad \rho_2 = 0.851\ \rho,\quad \rho_3 = 1.111\ \rho,\quad \rho_4 = 1.198\ \rho,\quad \rho = \sqrt{\lambda_0 r_0} \tag{5}$$

Figure 10 shows the radiation patterns of the "double ring" in the suppression region for the range $(0.93\text{-}1.07)\lambda_0$. The dashed line on this figure indicates the pattern of the single ring screen. The wide suppression region of the "double ring" (about three times wider than with the single ring screen) renders it particularly suitable for the protection of high-gain antennas (40 dB and up).

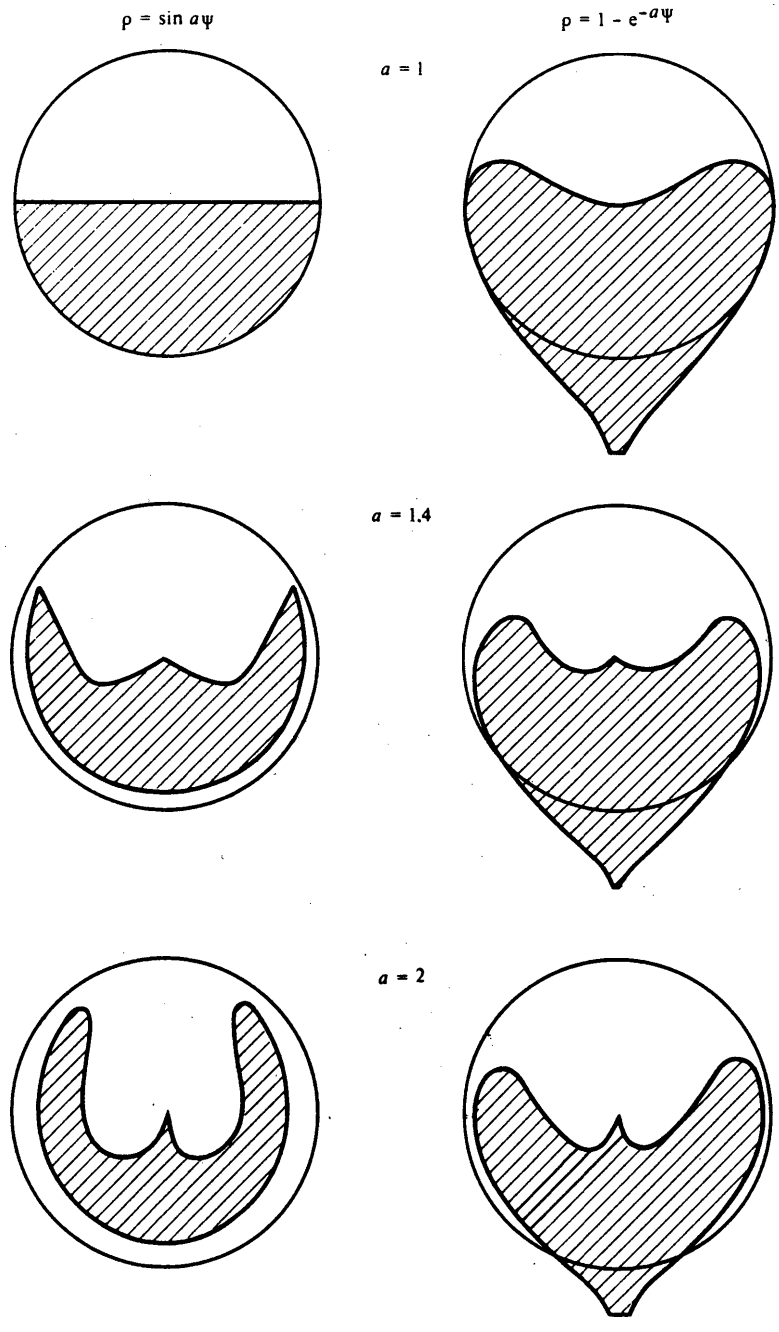
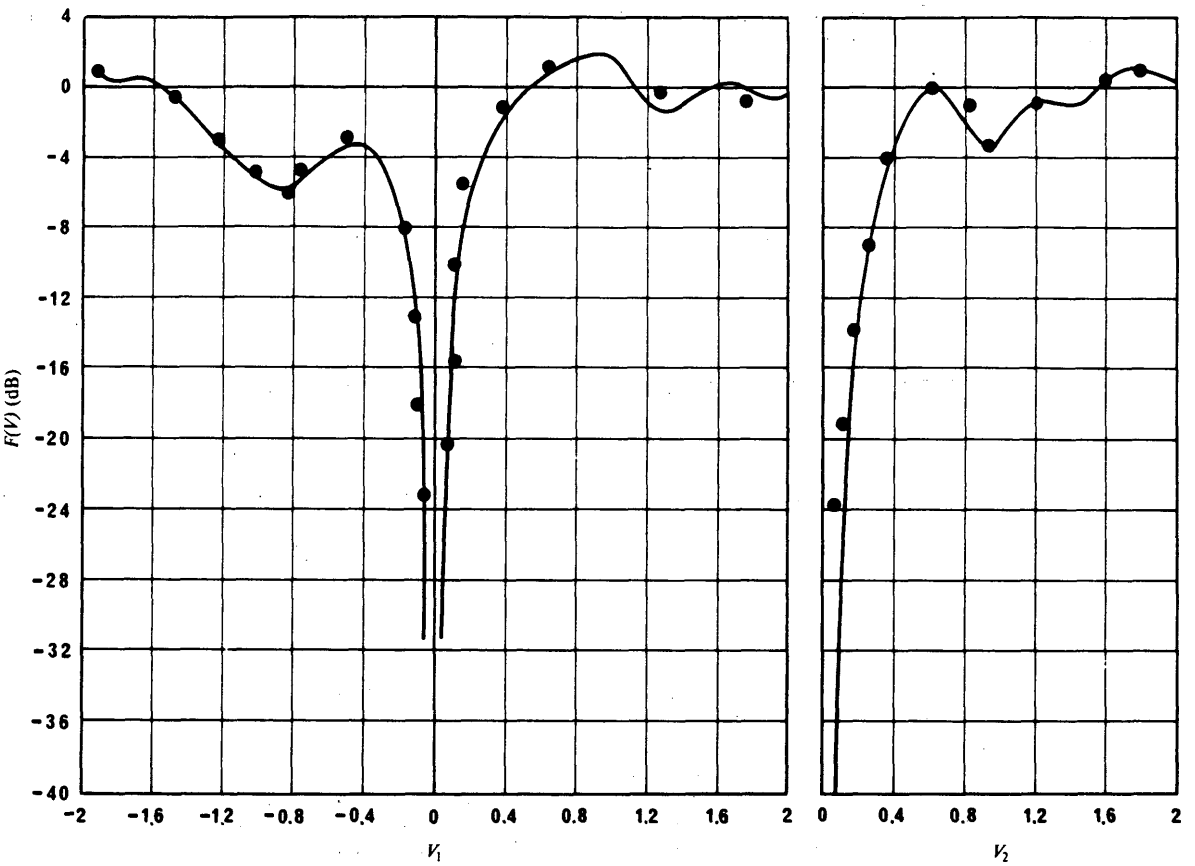


FIGURE 2 - *Screen configurations*



$\rho = 1 - e^{-2\Psi}$; $a = 2$

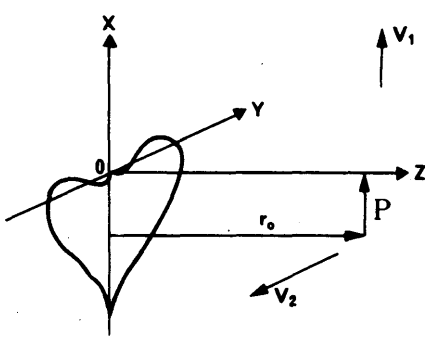


FIGURE 3 - *Radiation patterns of a screen in the suppression area for* $\rho = 1 - e^{-2\Psi}$

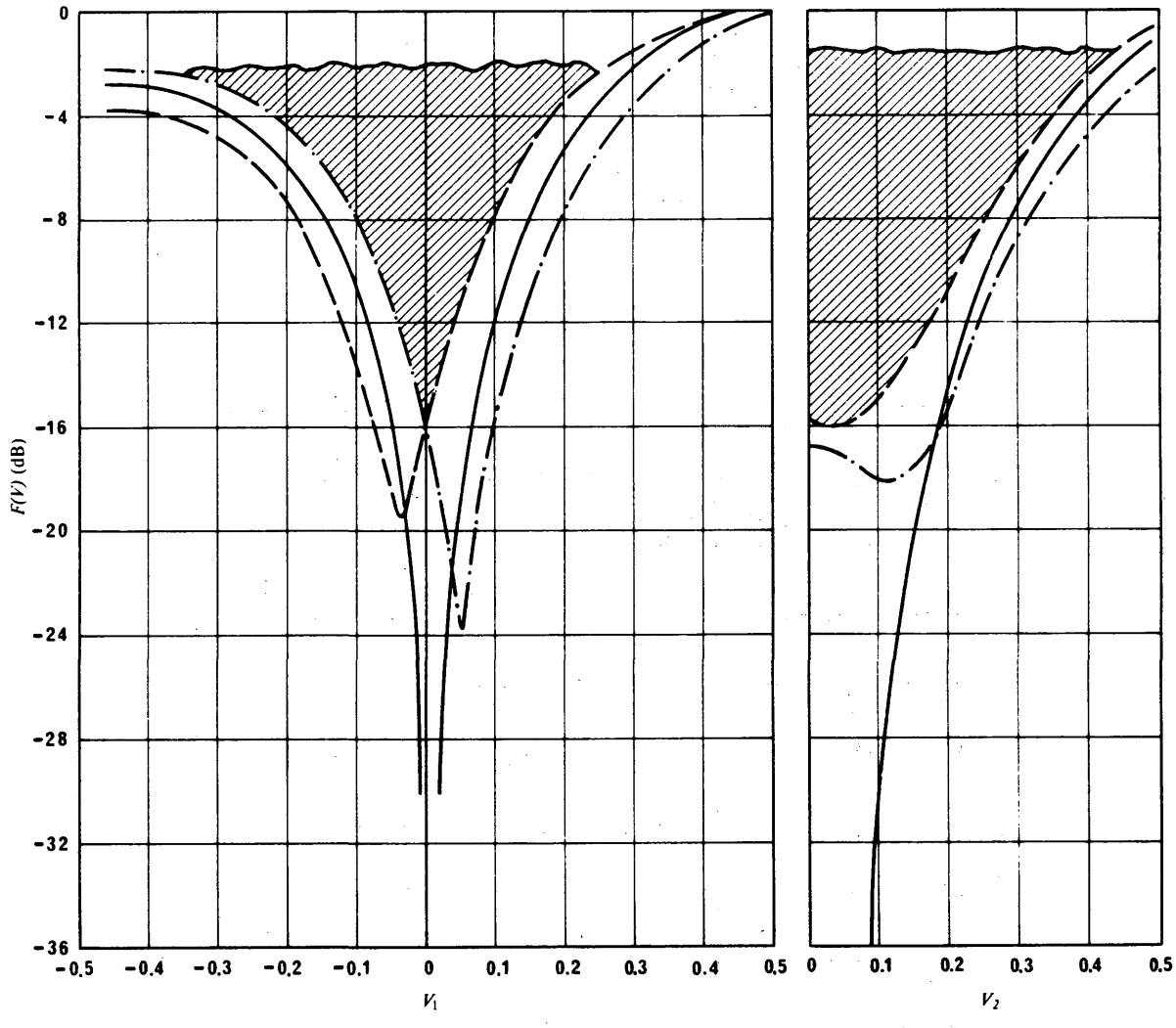


FIGURE 4 - *Radiation patterns in the range λ_0 (1 ± 0.1)*

——— : calculated for $\lambda = \lambda_0$

—·— : calculated for $\lambda = 0.9\ \lambda_0$

— — : calculated for $\lambda = 1.1\ \lambda_0$

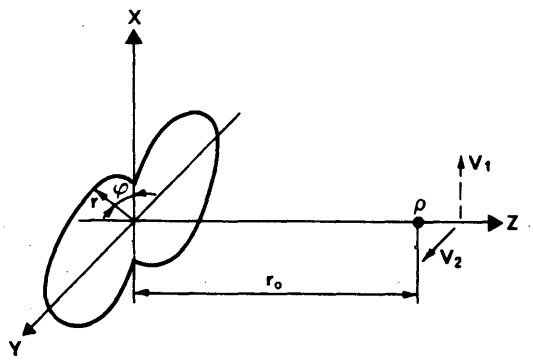


FIGURE 5 - *Configuration of a screen with two axes of symmetry*

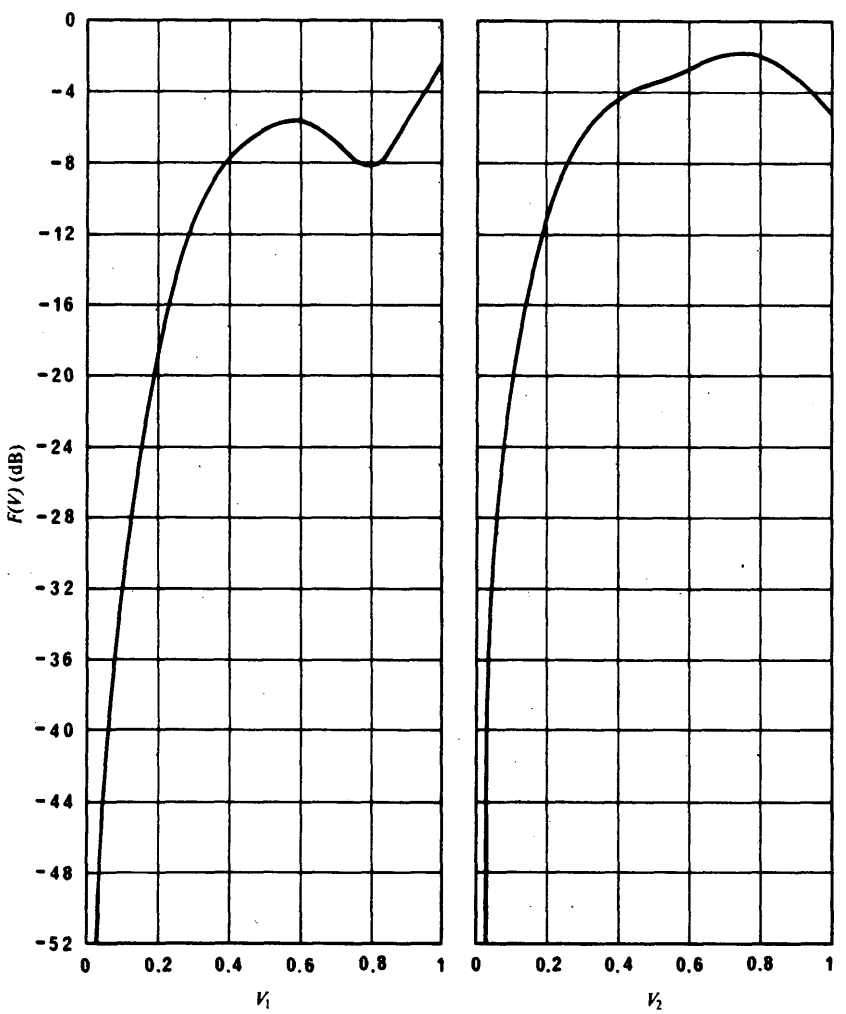


FIGURE 6 - *Radiation patterns for a screen with two axes of symmetry*

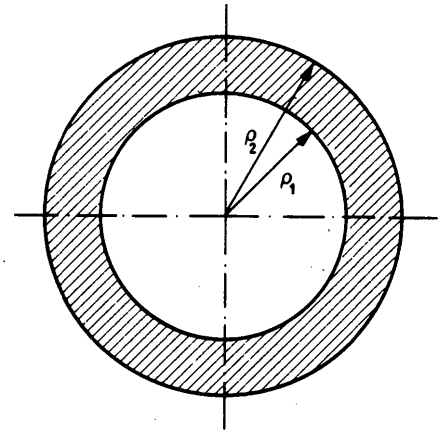


FIGURE 7 - *"Ring" screen*

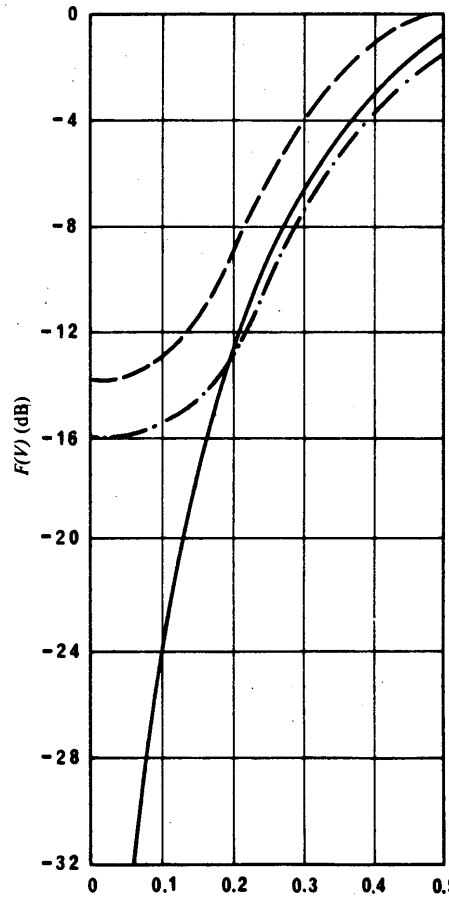


FIGURE 8 - *Radiation patterns of a "ring" screen in the suppression region for the band λ_0 (1 ± 0.1)*

——— : calculated for $\lambda = \lambda_0$

—·— : calculated for $\lambda = 0.9\ \lambda_0$

— — : calculated for $\lambda = 1.1\ \lambda_0$

If the "double ring" screen is suspended directly from the antenna mast, the mast and the supporting elements reduce the efficiency of the screen in the region of greatest suppression. The "double square" screen [Melnikov *et al.*, 1980] (Fig. 11), in which the antenna supports are blanked by the screen curtain, does not suffer this disadvantage. The screening efficiency is only slightly reduced by substituting a square for an annular screen. The "double square" screen possesses a nearly axisymmetrical suppression pattern.

The screen dimensions are given by the formulae:

$$x_1 = y_1 = 0.3\ \rho, \quad x_2 = y_2 = 0.81\ \rho, \quad x_3 = y_3 = 0.99\ \rho, \quad x_4 = y_4 = 1.145\ \rho \tag{6}$$

Formula (4) is used to determine the positioning of the screen.

The "double ring" and "double square" screens may be used to protect antennas from interference arriving from any direction, including the front halfspace, outside of angles $\theta \simeq 0.7\ \theta_0$, $2\ \theta_0$ being the null beamwidth of the antenna to be protected.

In some cases, it is useful to have a screen which is considerably widened, even if only in one plane. This situation arises when antennas have to be protected from a number of sources lying in a single plane from directions in which the main field contribution originates. In this case, screening is best achieved by one or more strips of radio-opaque material which are large compared to λ_0 [Melnikov, 1975b; Melnikov *et al.*, 1981].

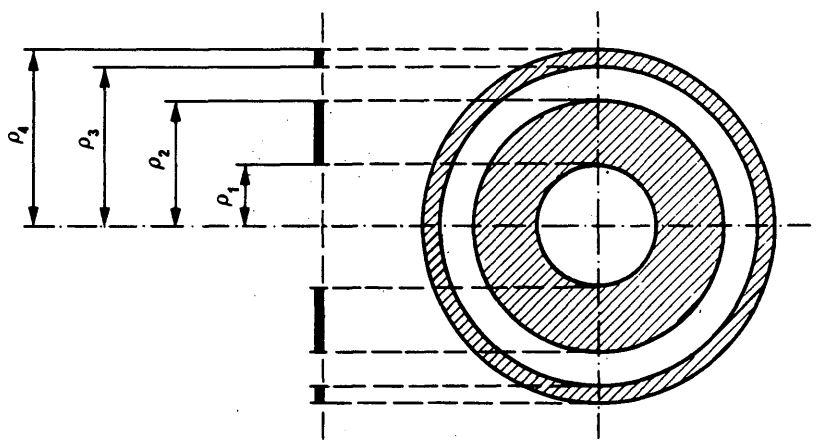


FIGURE 9 - "Double ring" screen

4. Antenna-mounted screens

Screens mounted on the antenna reduce the side-lobe radiation in particular sectors. One of the simplest types is the disc screen mounted on the rearside of the reflector [Gross *et al.*, 1969] at a certain distance from the edge of the aperture (Fig. 12a). With the dimensions indicated in the figure, the screen increases the immunity of the antenna to interference by 8-10 dB in the sector 180° ± 35°.

The greatest suppression (15-18 dB), but in a much narrower sector around the direction 180°, is produced by the slotted screen (Fig. 12b) [Melnikov and Yampolsky, 1980]. Installed at a distance of $2\lambda_0$ from the edge of the aperture, the slotted screen is smaller than the disc.

The slotted screen may be successfully used to suppress the overflow lobe field of antennas employed in trans-horizon links and designed for use with an offset reflector having a square or rectangular aperture (Fig. 12c). The gap (slot) between the antenna reflector and the screening strip Δ is $0.19\sqrt{\lambda_0 r_0}$, with a strip width $\delta = 0.57\sqrt{\lambda_0 r_0}$, r_0 being the distance from the phase centre of the radiator to the bottom edge of the antenna. The radiation pattern with and without screen is shown in Fig. 13.

To reduce the horizontal radiation of a horn-parabolic antenna with horizontal polarization of the field in the sector 90° ± 15°, a screen is used which dephases the radiation from the lateral edge of the antenna. The screen shape (Fig. 14) [Yampolsky and Petrova, 1976] is such that the maximum dephasing rate is reached at the point where the field distribution at the edge is maximum. The screen is mounted as an extension to the back walls of the aperture antenna. The antenna radiation pattern with and without screen is shown in Fig. 15.

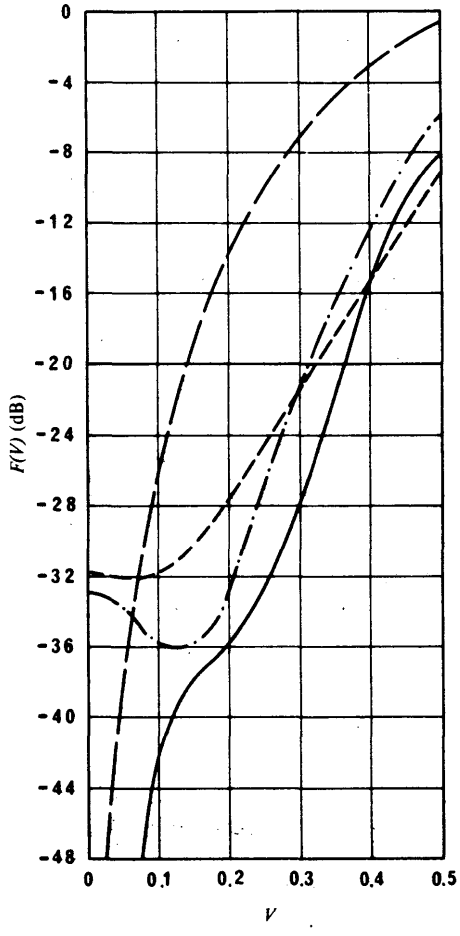


FIGURE 10 - *Radiation patterns of the "double ring" screen in the suppression region λ_0 (1 ± 0.07)*

— — : for single "ring" screen for $\lambda = \lambda_0$

——— : for "double ring" screen for $\lambda = \lambda_0$

– – – – : for "double ring" screen for $\lambda = 1.07\,\lambda_0$

— · — : for "double ring" screen for $\lambda = 0.93\,\lambda_0$

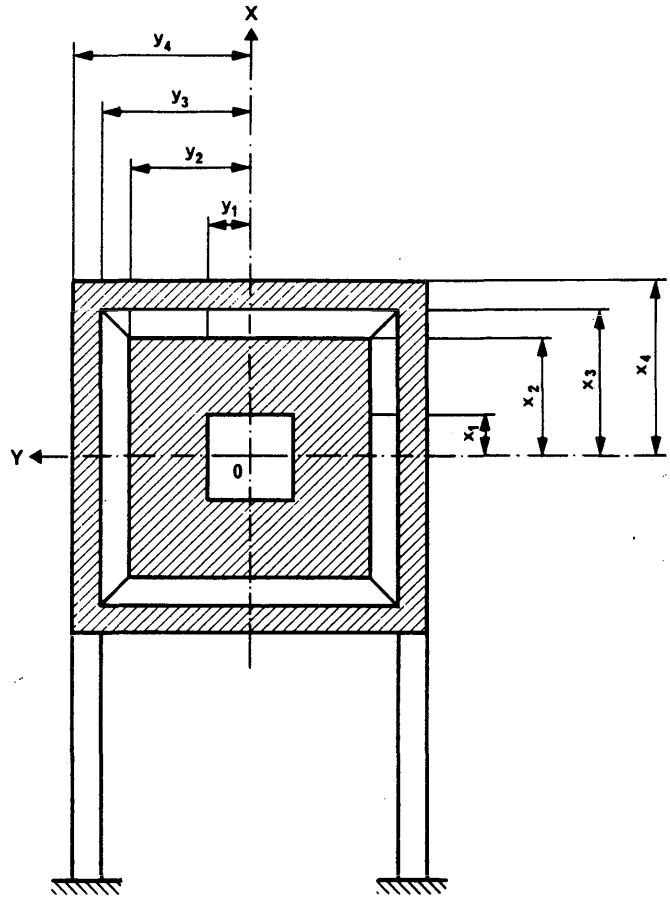
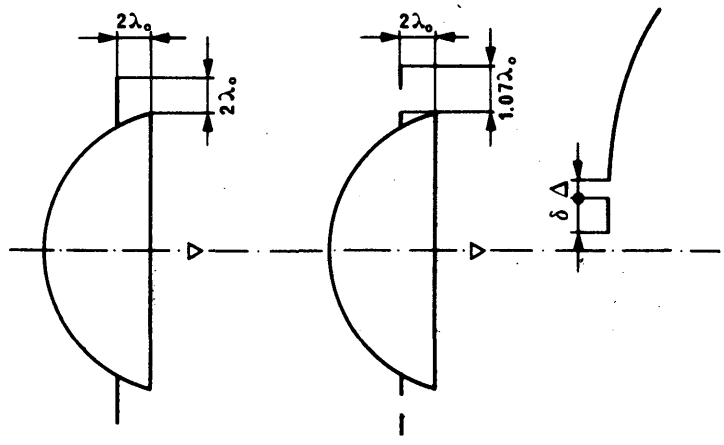


FIGURE 11 - *"Double square" screen*



(a) Disc screen on axisymmetrical antenna

(b) Slotted screen on axisymmetrical antenna

(c) Slotted screen on non-axisymmetrical antenna

FIGURE 12 - *Antenna mounted screens*

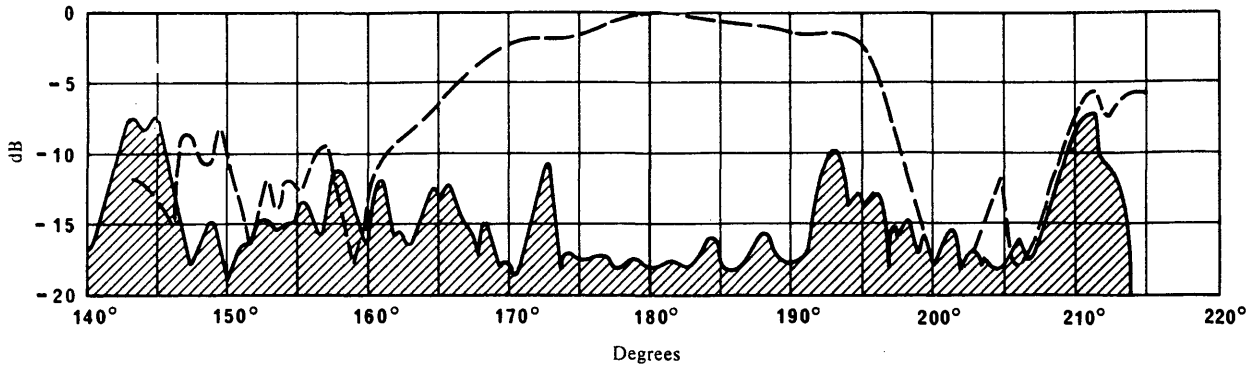


FIGURE 13 - *Measured radiation patterns for offset reflector antennas with and without slotted screens*

——— With screen

— — — Without screen

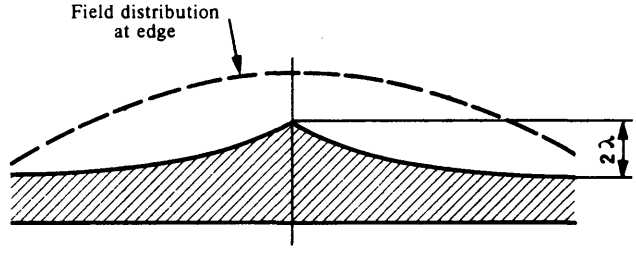


FIGURE 14 - *Screen shape for horn-parabolic antenna*

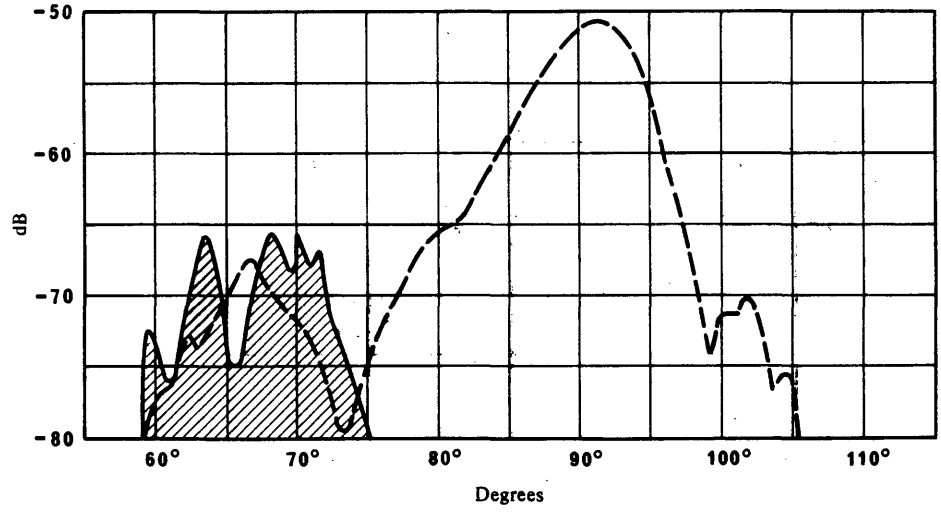


FIGURE 15 - *Measured radiation patterns for horn-parabolic antenna*

——— With screen

— — — Without screen

5. Screen as a polygonal star

The radiation level of antennas used in radio-relay links operating in the band 4 GHz with two-frequency operation within a sector of 180° ± 50° should not exceed -60 to -70 dB. The natural interference immunity of axisymmetrical antennas in the shadow part of the space near the shadow axis is relatively low. This is due to cophasal excitation by the feed of the edge of the main dish. The radiation level immediately behind, normally has a value of about -5 to -15 dB relative to isotropic radiation.

In this connection a number of devices have been developed and have been described in the literature which make it possible to improve the protection of circular aperture antennas, i.e. to reduce their radiation level in the rear half-space. The simplest are various types of flat screens (solid, slotted, etc.) mounted at the rear of the antenna, or cylindrical blinds with bevelled or spiralled edges.

A description is given below of the results obtained in studies of a more effective screen whose edge is shaped to form a regular polygonal star. The screen should be fastened directly to the antenna flange (Figure 16). The operation of the screen is based on the strong and periodically repeated dephasing of the edge currents. The field which is diffracted at the edge of the screen is mainly dispersed in the front half-space. Since the radiation level of the antenna determined by other factors is fairly high in the front half-space, the total interference immunity is practically unchanged.

If the screen is shaped as a regular or truncated polygon, it is not possible to achieve any substantial suppression of the field within a wide sector because of the slight phase difference in the fields along the edge of each projecting tooth; far better results can be obtained if the screen is designed as a polygonal star (Figure 17) which ensures that the phase of the field changes rapidly along the edge of each projecting tooth.

Figures 18-24 show characteristics of this antenna diagram. The antenna diagram is symmetrical relative to axis CC', i.e. $G(\theta) = G(-\theta)$ (Figure 18). Figure 19 shows calculated antenna diagrams for an antenna with a star-shaped screen in a plane passing through the tip of the teeth (plane A'OA in Figure 17) for N = 12 (N - number of sides on the star-shaped screen, N/2 - number of teeth), $h = 2.19\lambda$, $R_1 = 10.97\lambda$ for a straight-edge tooth (curve 1) and a spiral-edge tooth (curve 2). The diagram in the E plane is shown in Figure 19a, and in the H plane in Figure 19b. All the diagrams are referred to the radiation level of an antenna with a circular cophasal edge. It is clear that the star-shaped screen results in intense suppression of the antenna radiation in the shadow half-space, better results being obtained with small values of θ with spiral-edged teeth (linear phase taper). The angle θ characterizes the position of the observation point in the suppression plane. However, when $\theta = 160°$ in the H plane, there is a considerable degradation in the protection obtained in comparison with the teeth with rectilinear edges. In the plane passing through the hollows (plane B'OB) the picture is similar. Figure 20 shows calculated diagrams for the side radiation of an ADE type antenna with a star-shaped screen having a piecewise-spiral edge and the following geometry: N = 32, $h = 2.19\lambda$, $R_1 = 10.97\lambda$ (E plane - solid line, H plane - dotted line; the calculated plane passes through the projections of the flange). As the experimental data show (Figure 21), the efficiency of the screen at first increases when the number of sides is increased, but subsequently it decreases. The dotted line in Figure 21 shows the averaging curve for the experimental points. This is apparently due to the joint action of the diffraction fields at the tips and in the hollows around the screen, the change in the polarization structure of the field in the screen, etc. Figure 21 shows the variation in the suppression of the field at $\theta = 180°$ in the E plane as a function of the number of teeth where $h = 2.5\lambda$. The dots indicate the experimentally measured level of suppression when the screen is arranged with the teeth in the suppression plane, and the asterisks with the hollows.

The table below gives the results of studies made of the effectiveness of the operation of a star-shaped screen where θ = 180° with N = 32, as a function of the height of the teeth h, for various positions of the screen in the plane of suppression.

Position of screen relative to plane of suppression	Orientation of field strength vector in plane of suppression:	Level of suppression as a function of height of teeth (dB) 2λ	2.5λ	3λ	3.5λ	4λ
Tips	E plane	22.5	18.5	28.5	18.5	23
	H plane	18.5	16	24	18	23
Hollows	E plane	20.5	19.5	27	18	24.5
	H plane	17	16	22	16.5	20.5

The table shows that when h/λ is an integer, the suppression is most intense. The best results with the given antenna dimensions are obtained with $h = 3\lambda$.

Figures 22 and 23 show experimental antenna diagrams for the ADE antenna with a gain coefficient of G = 35.5 dB using a polygonal star-shaped screen with 16 teeth, each with a height of 3λ for two positions of the screen in the plane of suppression: through the tips and through the hollows, respectively. The antenna diagram without a screen (angle θ = 180°) is indicated by dots. It will be seen that the screen provides a high degree of suppression (15 - 28 dB) and a broad sector of suppression (180° ± 50°).

Experimental studies have also shown that a polygonal star-shaped screen suppresses the cross polarization component of the field just as intensely. The cross polarization components of antenna fields without a screen (dots) and with a screen (solid line) are shown in Figure 24.

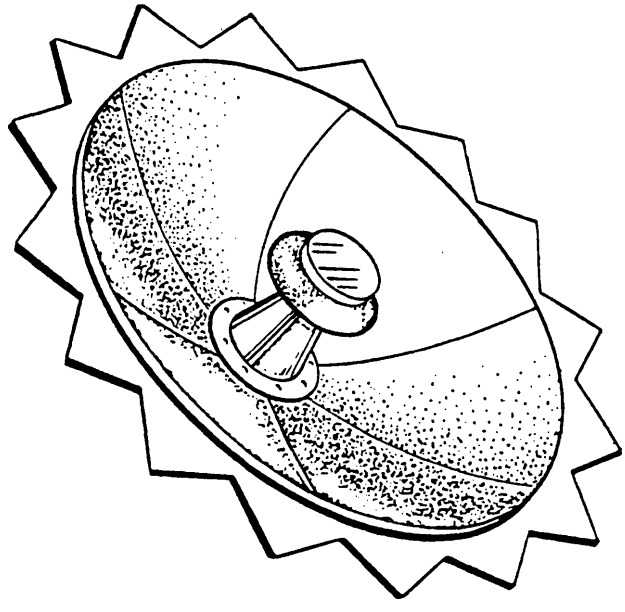


FIGURE 16 - Antenna mounted screen

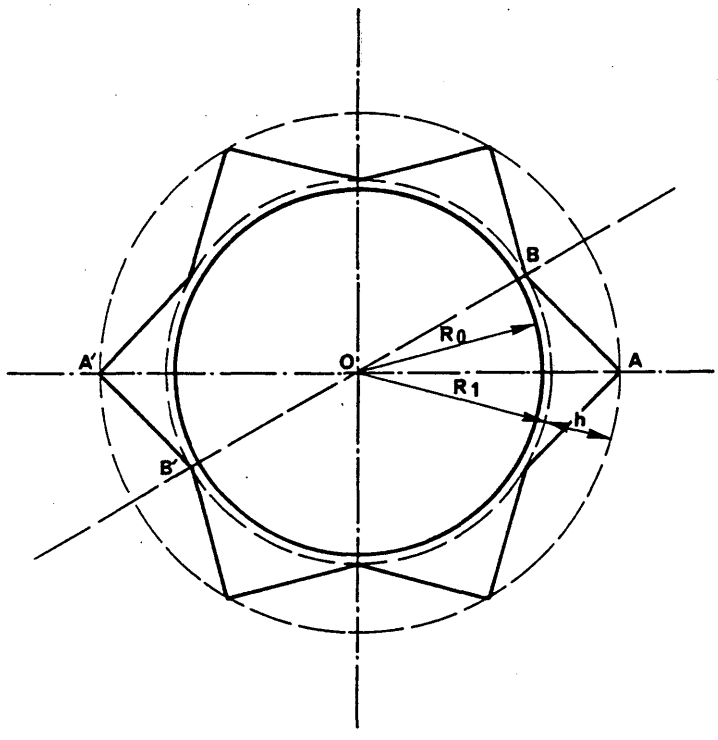


FIGURE 17 - Screen as a polygonal star

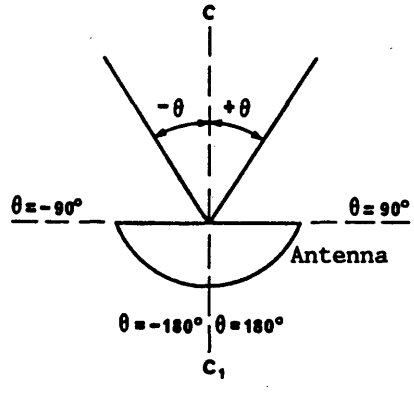
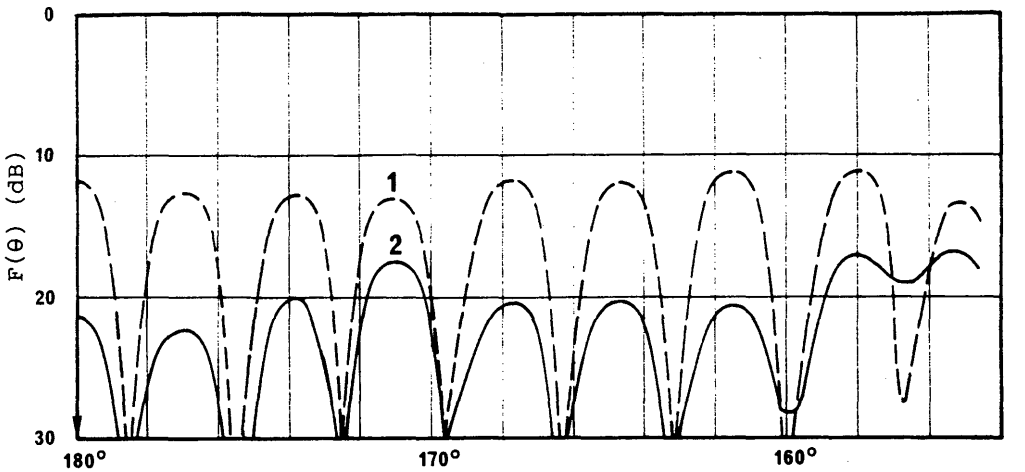
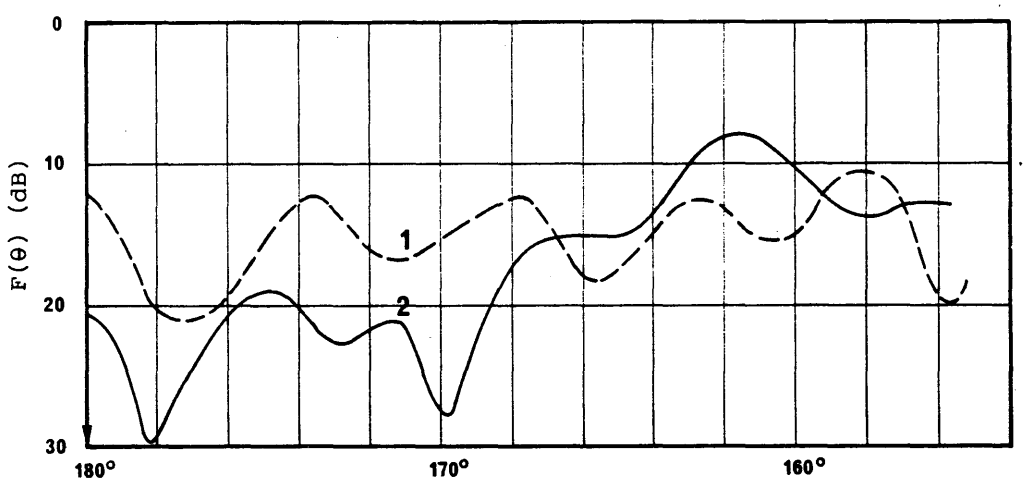


FIGURE 18 - The definition of an angle θ



(a) E plane



(b) H plane

FIGURE 19

Calculated antenna diagrams in a plane passing through the tip of the teeth

N = 12; $N/2$ = number of teeth; h = 2.19λ; R = 10.97λ.

Curve 1 - for a straight-edge tooth; Curve 2 - for a spiral-edge tooth.

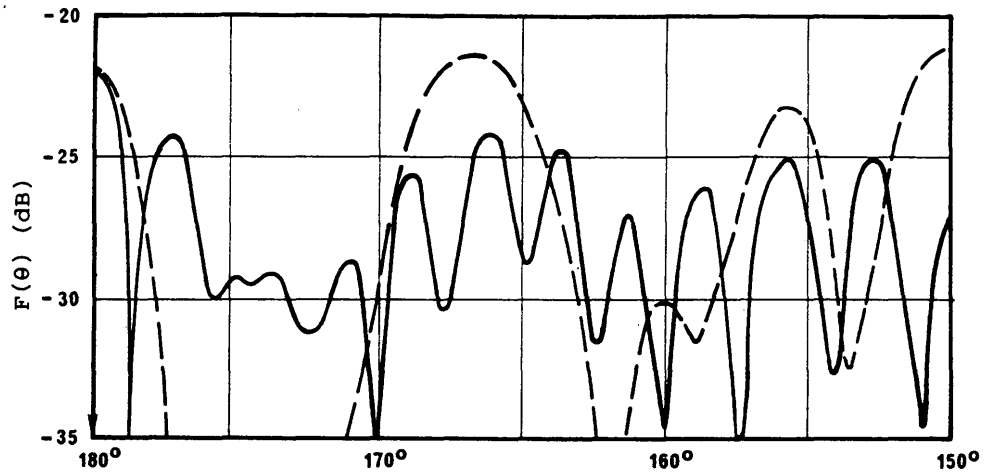


FIGURE 20

Calculated diagrams for the side radiation of antenna with a star-shaped screen having a piecewise-spiral edge

N = 12; h = 2,19λ; R_1 = 0,97λ

—— E plane --- H plane

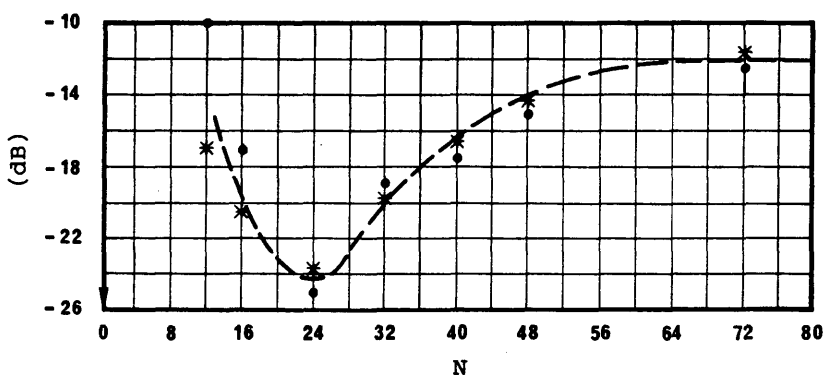
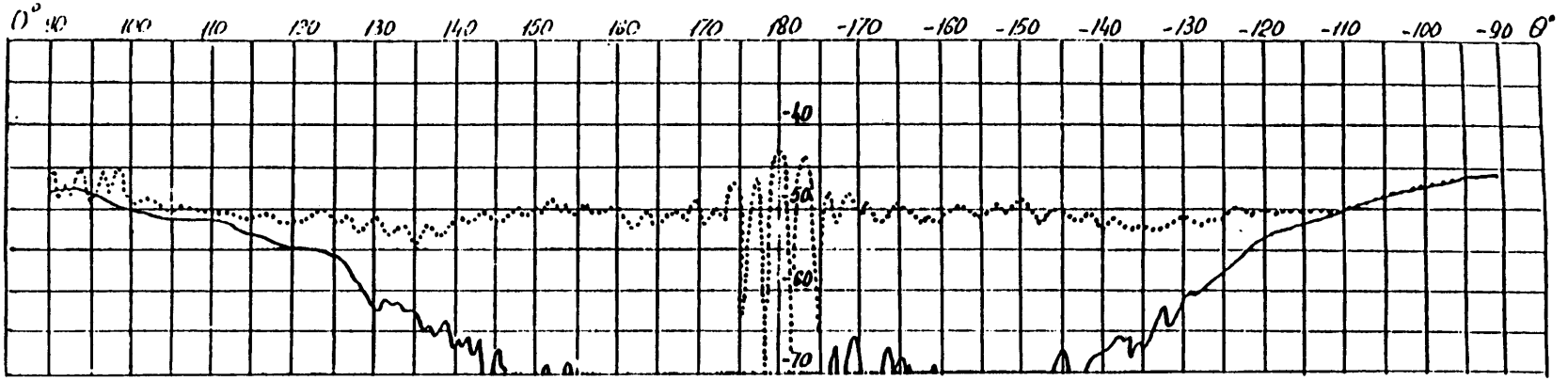


FIGURE 21 - Dependence of the efficiency of the screen on the number of T, N

(experimental data)



E plane

H plane

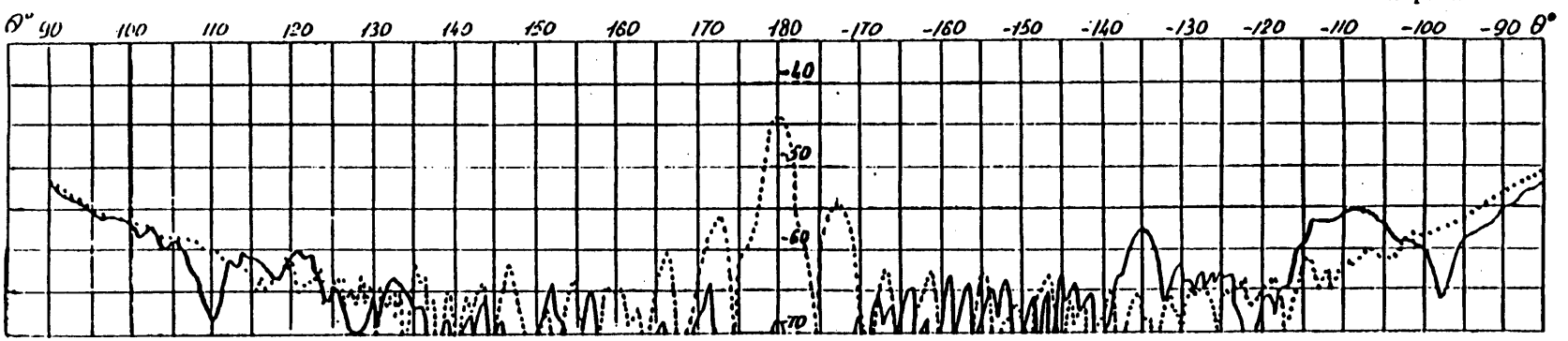


FIGURE 22

Experimental antenna diagrams for ADE antenna with a gain coefficient of G = 35 dB using a polygonal star-shaped screen ($N/2$ = 16; h = 3λ).

The position of the screen in the plane of suppression through the tips.
The antenna diagram without a screen is indicated by dots.

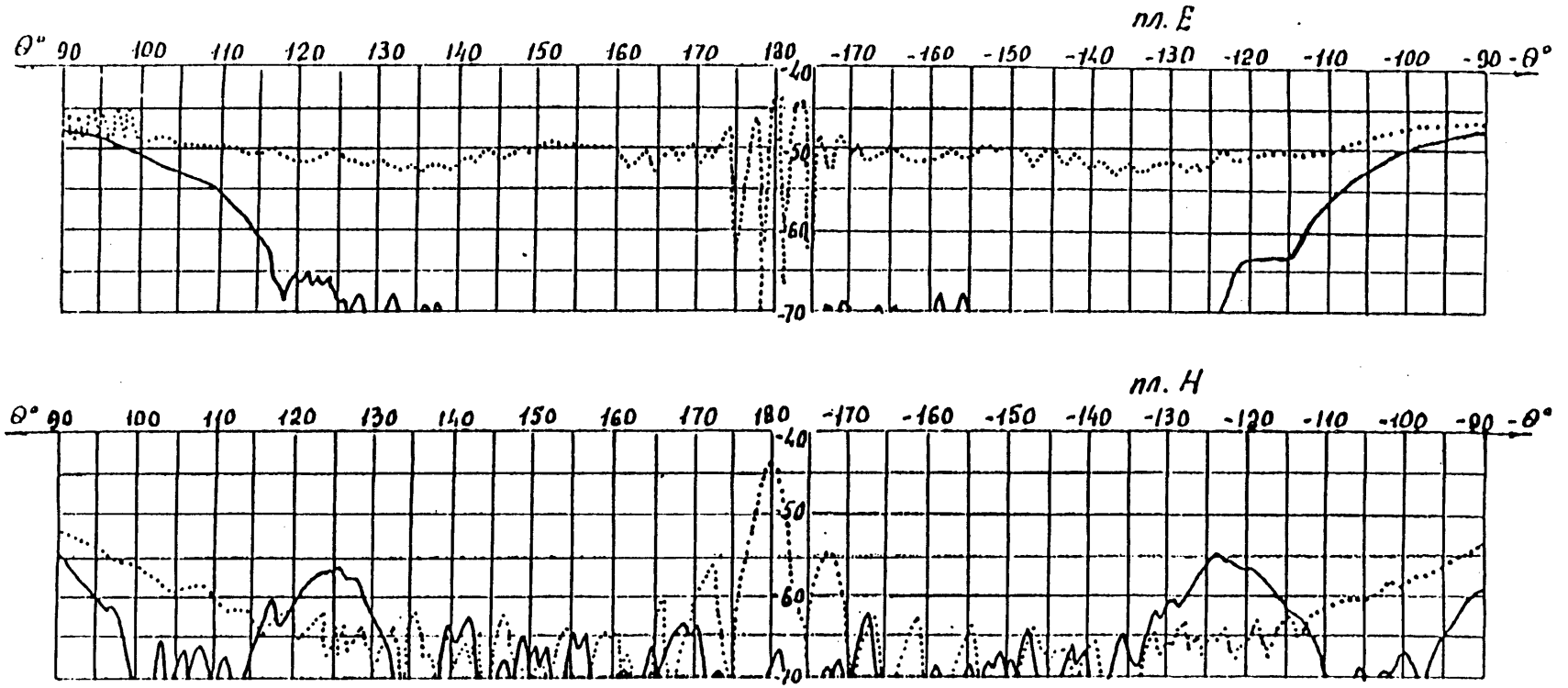


FIGURE 23

Experimental antenna diagrams for ADE antenna with a gain coefficient of G = 35 dB using a polygonal star-shaped screen ($N/2$ = 16; h = 3λ).

The position of the screen in the plane of suppression through the hollows.
The antenna diagram without a screen is indicated by dots.

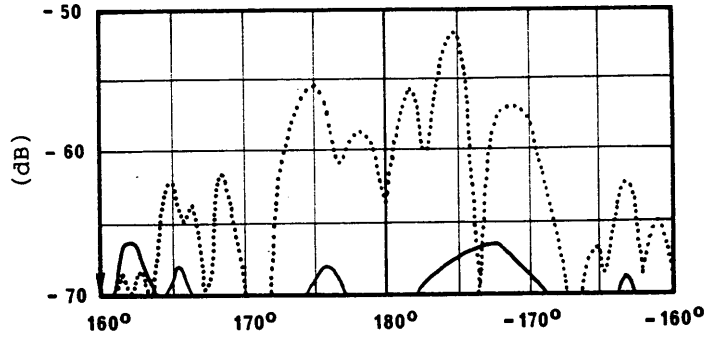


FIGURE 24

The cross polarization components of antenna fields

(experimental date)

—— with screen ··· without screen

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REPORT 973

ADAPTIVE ANTENNAS FOR LAND MOBILE APPLICATIONS IN REDUCING CO-CHANNEL INTERFERENCE

(Question 18/1)

(1986)

1. Introduction

Adaptive antenna systems hold promise for application in commercial mobile radio systems by reducing co-channel interference on shared frequencies, thereby allowing closer base or repeater station spacing with an effective increase in the number of available channels. Question 18/1 considered, in particular, that the radio spectrum is used more efficiently when more users can operate (efficiently) at the same time and also, that advances in technology, including "interference cancellers", are making practicable new sharing schemes that offer advantages for increasing the efficiency of spectrum sharing and band utilization. The Question asks how factors such as signal processing and antenna design can be employed for this purpose. In the case of adaptive antennas, the "trade-offs" for increasing the availability of channels are the additions of an electronic complexity and increased cost. More than one antenna is also required, but for mobile units the adaptive system can for example, very effectively use two simple whip antennas to reduce interference from a co-channel signal. The additional complexity and cost introduced will be substantially affected as integrated circuit technology proliferates, as production of these units increases, and as competitive forces operate in the market-place. The cost/benefit ratio for adaptive antenna systems has been improving remarkably in recent times.

2. Application

The purpose of an adaptive antenna system is to introduce a sharp null in the receiving antenna pattern in the direction of interference on a real-time basis. The system responds so as to rapidly adapt the direction of the null to meet changing interfering conditions. The system also acts faster than the rate at which desired information is being received. For speech transmission, which is the usual mobile radio case, the required response times are relatively slow and are measured in tens of milliseconds. Current electronic switch speeds used to adapt the antennas can be orders of magnitude faster than the rate of received information.

The adaptive antenna system continuously samples the wanted received signal, working on a known, previously coordinated signal component voltage, and then automatically adjusts antenna phasing to optimize output signal-to-noise ratio using a specific algorithm. In many modern land mobile installations, particularly those sharing a frequency with other stations, a continuous tone coded squelch system is used to help minimize interference and noise. This tone voltage is ideal for use with adaptive antennas. The tone modulation of the signal is filtered from the speech channel to enable the receiver processor to discriminate between the wanted signal and other unwanted signals modulated by different tone frequencies. In the same manner, it could become the reference signal for operation of the adaptive antenna system.

Mobile radio systems designed for communications purposes are universally equipped with omnidirectional (in the horizontal plane) whip antennas so the mobile radio will have equal capability no matter what the direction of travel. This configuration also allows, unfortunately, introduction of noise voltage and radio-frequency interference from any azimuth without discrimination. Degradation of normal mobile radio communications occurs as well from natural multipath reception or sway of antennas on a vehicle. Adaptive antenna systems have the additional benefit of working automatically to reduce the undesirable effects of multipath.

Adaptive antenna systems can be used to provide nulls in the frequency domain as well as in antenna patterns, but this Report will discuss only adaptation of antenna patterns as the simplest, most effective case, under the assumption that two (land mobile) co-channel signals approximately coincide in spectral distribution.

Adaptive antenna systems for commercial purposes are particularly suited to VHF and UHF land mobile systems and frequencies where spectrum congestion is noted. Most of the on-going work in this field is today addressing the problem of minimizing interference to mobile radio systems because of their inherent need to be able to communicate well in any direction. Figure 1 shows the block diagram of an adaptive antenna system.

3. Antenna system

In the usual mobile radio configuration, one-whip antenna of the adaptive system is employed to receive the wanted signal along with any unwanted RF voltage or interference that may be present. A second or auxiliary whip antenna associated with the system, is used to receive only a function of the unwanted interference. The RF voltage from the second whip is operated upon by a single complex gain function and then subtracted from the signal derived from the first antenna to produce a pattern null. The adaptor circuits recursively determine the gain value for operating on the wanted voltage.

In a multi-frequency communications system, the phase relationship of the two-whip antennas must be adjusted for proper pattern production as the frequency of operation changes. This is a complicating factor that applies only as the channel frequency changes. It is not a factor in most land mobile communications systems as they predominately receive on one frequency. All that is needed for proper pattern production is careful installation of the two-whip antennas.

The production of pattern nulls by phasing two parallel antennas is a well-known phenomenon. Less well-known, however, is the variation of null width as a function of phased spacing (base line) between the two antennas. The null width is, of course, also a function of angle of interest with respect to the base line between the two antennas. This matter is of importance when deciding how far apart to mount two antennas on a mobile unit. Null width becomes sharper with increased spacing and so the receiving system performance becomes more critical with vehicle motion. A sharper null, however, decreases the probability of cancelling the wanted signal should both the wanted and unwanted signal lie close to the same azimuth measured from the mobile unit. In a defined 1° null width system using two-whip antennas spaced $\lambda/2$ apart, the maximum probability that the wanted signal will be cancelled at any instant is about 2/360 or 0.55%.

The number of nulls is a function of base-line length and therefore, longer base-lines tend to increase the probability (albeit small) of nulling a wanted signal. Up to four nulls may be obtained for each wavelength of base-line distance, although only one of them is controllable for interference reduction purposes in a two-whip antenna system. This figure is a maximum number assuming the wanted and unwanted transmitters are not co-sited. In a moving mobile system, the two-dimensional geometry between mobile unit, wanted transmitter, and unwanted transmitter is constantly changing (except for the almost trivial case where the motion vector is identically towards both transmitters, i.e. all three points remain in one dimension).

Figure 2 shows null width related to antenna spacing as a function of angle of arrival of unwanted signals. Note that null width is always inversely proportional to base-line length. As null width increases, the combined antenna pattern approaches omnidirectional (i.e. the unwanted signal wavefront is incident perpendicularly to the whip base-line).

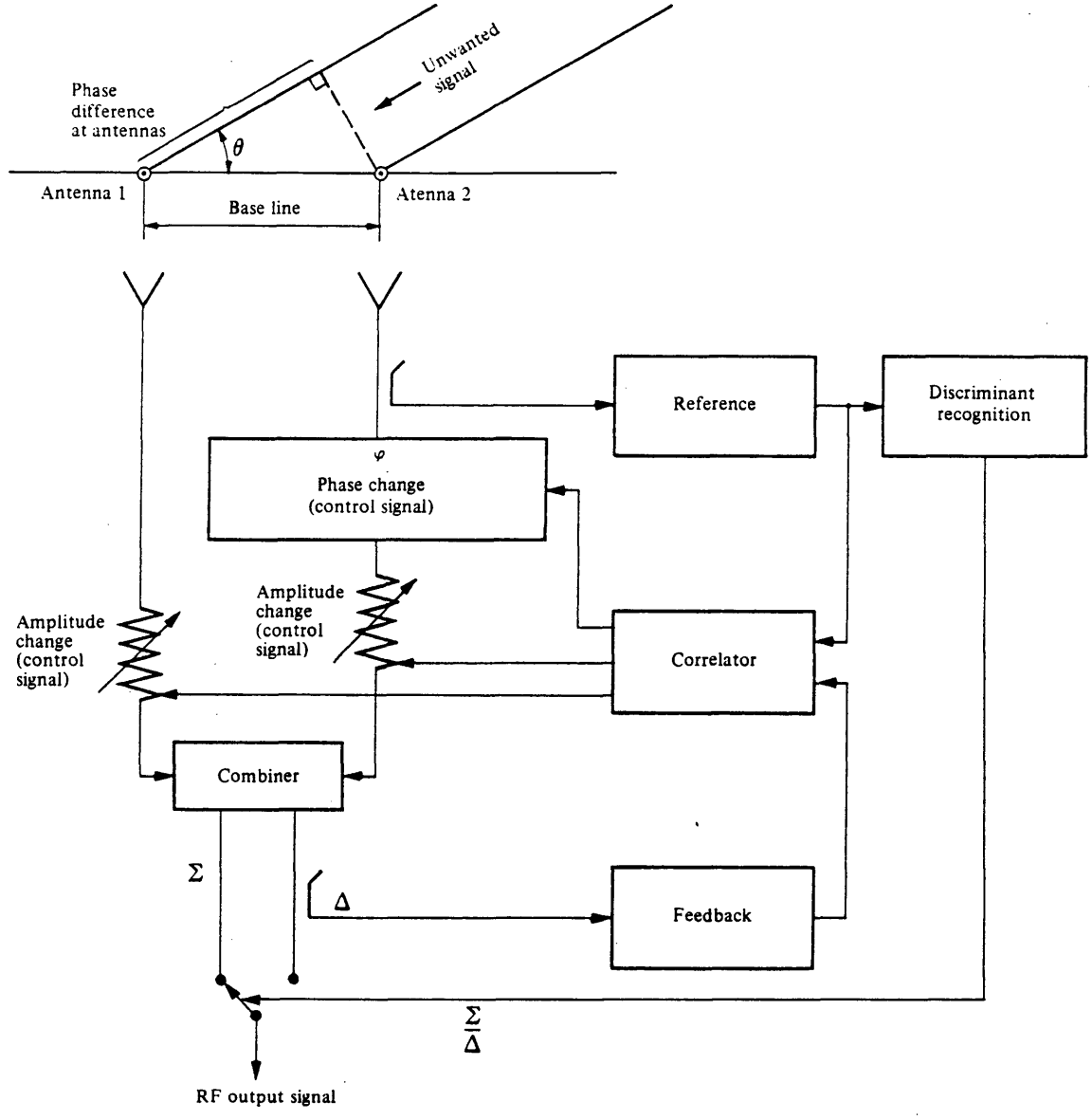


FIGURE 1 – *Block diagram – Adaptive antenna system*

Σ: sum signal
Δ: difference signal
Σ/Δ: control signal

Figure 3 shows null width as a function of angle of arrival of the unwanted signal, calculated for three null depths, with antenna spacing of one wavelength in each case. Again, as the null width increases, the resulting pattern approaches omnidirectional.

Note from Figs. 2 and 3 that geometric relationships give boundary limits. Propagation characteristics and vehicle motion will act to reduce performance of an adaptive antenna system. Null width, as developed for Figs. 2 and 3, is the azimuthal angle difference included between two points on the skirts of a curve from an infinite null.

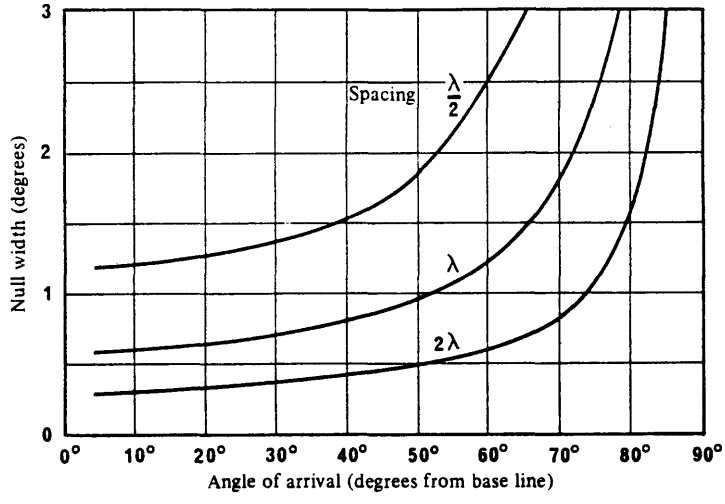


FIGURE 2 – *Null width as a function of angle of unwanted signal arrival, θ, for three whip antenna spacings (for 30 dB minimum interference reduction)*

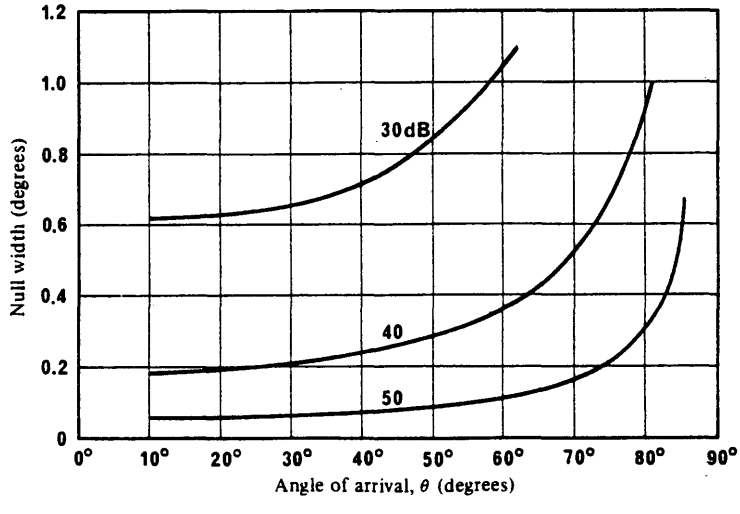


FIGURE 3 – *Null width as a function of angle of unwanted signal arrival, θ, for three calculated values of interference reduction (with an antenna base-line length of one wavelength)*

4. System integration

The two-whip antennas may be identical, both matched to the input impedance of the receiver through their control circuits, or they may be dissimilar. There is no requirement for the second antenna to be a good receptor for the wanted signal. The chance of inadvertently nulling the wanted signal becomes small if the second antenna is a poor receptor for the wanted signal. However, dissimilar antennas will require that an amplifier or attenuator is used in order to satisfy the equal signal amplitude requirement for optimum nulling. Gain choices must also be made with consideration of the potential for amplifier self-oscillation at frequencies which provide greater coupling between the two antennas.

Receiver front-end pre-selection or band-bass filtering is required in the auxiliary antenna circuit for the same reasons that it is required in the receiver. Amplifiers in the device need to be linear at the level of all received signals, including interference signals from the auxiliary antenna. Linearity and band-pass filtering are especially important when there is interference from a nearby transmitter on an adjacent frequency. Since null depth is limited by correlation of the waveforms from the two antennas at the combination point, it is important to ensure that signals from the two antennas pass through nearly identical filters.

Optimum interference reduction takes place when the signal voltages received from the direction of the unwanted transmitter, are of equal amplitudes and of opposite phase at the combining point of the two antennas. The following limiting expression can be derived when both voltages are of unequal amplitude:

$$\text{Interference reduction} = 20 \log \left(10^{x/20} - 1\right)^{-1}$$

where x is the difference in level of two voltages that will cancel when they are of equal amplitude. For example, if the adaptive system will allow precision of matching to 0.1 dB, then maximum interference signal reduction is calculated at 38 dB, assuming the signals are of opposite phase. Phase consideration is relatively simple in land mobile systems which usually operate on only one frequency.

5. Results

Experimental results of an adaptive antenna system using two-whip antennas at VHF to reduce the level of an interference source have been obtained [Widrow *et al.*, 1975]. Performance has been evaluated under both laboratory conditions and in the field. For unmodulated signals, a cancellation ratio of 50 to 60 dB was achieved over a dynamic range of 120 dB. For modulated signals, cancellation ratios of the order of 40 dB have been measured across a typical land mobile channel bandwidth of 30 kHz. Multipath cancellation with a stationary vehicle, resulting in poor performance by the adaptive antenna system rarely occurred.

Tests with a moving vehicle showed more susceptibility to multipath effects than for a stationary vehicle. Multipath reflections do not appear to the antenna system as a new unwanted signal, but rather as a change in the apparent amplitude and azimuth of the signal. As a mobile unit moves, the wanted and unwanted field strength will randomly vary and their apparent azimuth will vary from the real azimuth. Although adaptive antennas assist in reducing multipath interference, there are limits to what can be accomplished by these systems and wanted to unwanted signal ratios may sometimes exceed the capability of the adaptive processor. This was found to be a result of limitations in the feedback loop in responding fast enough to very rapid signal fluctuations. System improvements were possible by trading response time with null depth. In open areas having no multipath with a vehicle moving at up to 50 km/h, 40 dB of interference cancellation was maintained most of the time.

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REPORT 832*

ISSUES IN PACKET RADIO COMMUNICATIONS

(Question 18/1)

(1982)

1. Introduction

Packet switching was originally designed to provide efficient network communications for "bursty" traffic, and to facilitate computer network resource sharing. Packet radio [Kahn, 1977] is a technology that extends the original packet switching concepts, which evolved for networks of point-to-point communication lines, to the domain of broadcast radio networks. Packet radio offers a highly efficient way of using a multiple access channel, particularly with mobile radio service users and large numbers of "bursty" traffic sources.

The advantages of multiple access radio channels for information distribution and computer communications have been well established and several experimental networks are now in operation [Kahn *et al.*, 1978]. The basic concept in packet radio is that data is transmitted in packets, of fixed or variable length, containing a few thousand bits. A packet includes all the addressing and control information necessary to correctly route it to its destination terminal possibly via a nearby repeater. In the case where packets are routed via repeaters, only one repeater will typically be identified to relay them, while other repeaters will discard them. In some network configurations, packets can be relayed through several repeaters until they reach their final destination, where their reception is acknowledged if they have been correctly received.

The various types of random access protocols that have been proposed [Tobagi, 1980] fall essentially into two groups, namely ALOHA and Carrier Sense Multiple Access (CSMA), with the latter providing under some circumstances maximum throughput that can reach 90% of the channel speed capacity.

In using mobile channels, factors such as fading, man-made noise, etc., have to be taken into account in the context of the analysis of packet radio. This Report looks at these aspects and presents some preliminary solutions to these problems.

* This Report is brought to the attention of Study Groups 8 and 9.

2. Packet length considerations

In packet radio communication networks, the unit of data information to be transmitted is typically a message of variable size consisting of a number of constant length packets, each containing the data information as well as a small amount of overhead. However it is also possible [Kahn *et al.*, 1978] to transmit variable length packets. In the former case, in order to minimize the end-to-end average message delay, the packet length has to be optimized. Factors affecting this optimization include:

- multipath fading,
- man-made noise,
- vehicle speed,
- signal-to-noise ratio.

Using a scheme where packets in error are retransmitted, Figs. 1 and 2 [DaSilva *et al.*, 1980] show how the optimal packet size in a fading channel is related to the bit rate and the average message length. As the signal-to-noise ratios increase we observe (Fig. 1) that the optimal packet length increases as well. For example, for an average message length of 10 000 bits and a signal-to-noise ratio of 40 dB, the optimal packet length varies from 240 bits at 1 kbit/s to 750 bits at 16 kbit/s. In Fig. 2 we notice that for average message lengths in excess of 4000 bits, variations in the optimal packet length are minimal. Variations in the vehicle speed and signal-to-noise ratio as a function of the geographical location have an impact on the optimal packet size. Using the classical stop-and-wait protocol, [Hafez, 1980] a relation between the optimal number N of packets per cycle, as a function of the packet size in bits and the minimum wasted time is displayed in Fig. 3. It is clearly indicated that N has to be adjusted as a function of received signal level at the receiver. Figure 4 [Leung, 1980] shows that further improvement on the stop-and-wait approach is possible with the use of forward error correction.

3. Synchronization considerations

For systems of moderate carrier stability operating in the UHF range, carrier acquisition represents a problem which is further compounded in the case of "bursty" packet transmissions in a varying signal-to-noise ratio environment. Recently, however, [El-Tanany and Lam, 1981] a COSTAS loop circuit designed specifically for a packet radio system, has provided a substantial decrease in acquisition time as shown in Fig. 5.

For a TFM (tamed frequency modulation) modulated signal [Muilwijk, 1979], the results indicate that under low signal-to-noise ratios, (S/N = 6 dB) locking condition is achieved in 8 ms. These figures were derived for channel bandwidths of 25 kHz, over which practical transmission bit rates are of the order of 16 kbit/s. Different figures would be obtained for other channel widths and transmission bit rates [Kahn *et al.*, 1978].

4. Discussion and conclusions

The underlying objective of random access packet radio schemes is to maximize the channel or system information throughput of a large number of geographically scattered "bursty" traffic sources to some acceptable level of service. To realize this objective, a number of interrelated system constraints and variables have to be examined. In § 2 the problem of selecting an optimal packet length for the base station to mobile station link, was considered. To do so, the multipath fading nature of such channels as well as the influence of the error control procedures were considered. It remains, however, to re-examine the problem in the context of man-made noise which will affect the optimal packet length and to study the improvements that can be obtained by the use of bit interleaving and space diversity.

One other important random access issue is the so-called network stability problem [Kleinrock and Lam, 1973]. Temporary increases in traffic loads can drive the random access protocols into an unstable region where, if no control action is taken, the actual system's throughput will be reduced to zero. It is, therefore, crucial to develop the appropriate stability control mechanisms which themselves are only a subset of the signalling, error control and link protocol design issues. Such issues are particularly relevant when random access systems cease to be configured as local one-hop networks to become larger networks, possibly involving several stations, as well as multiple-hop transmissions. In this regard, it is important to consider not only the spectrum efficiency of various systems' architectures including single versus multicell networks, but also the inter-work user protocols.

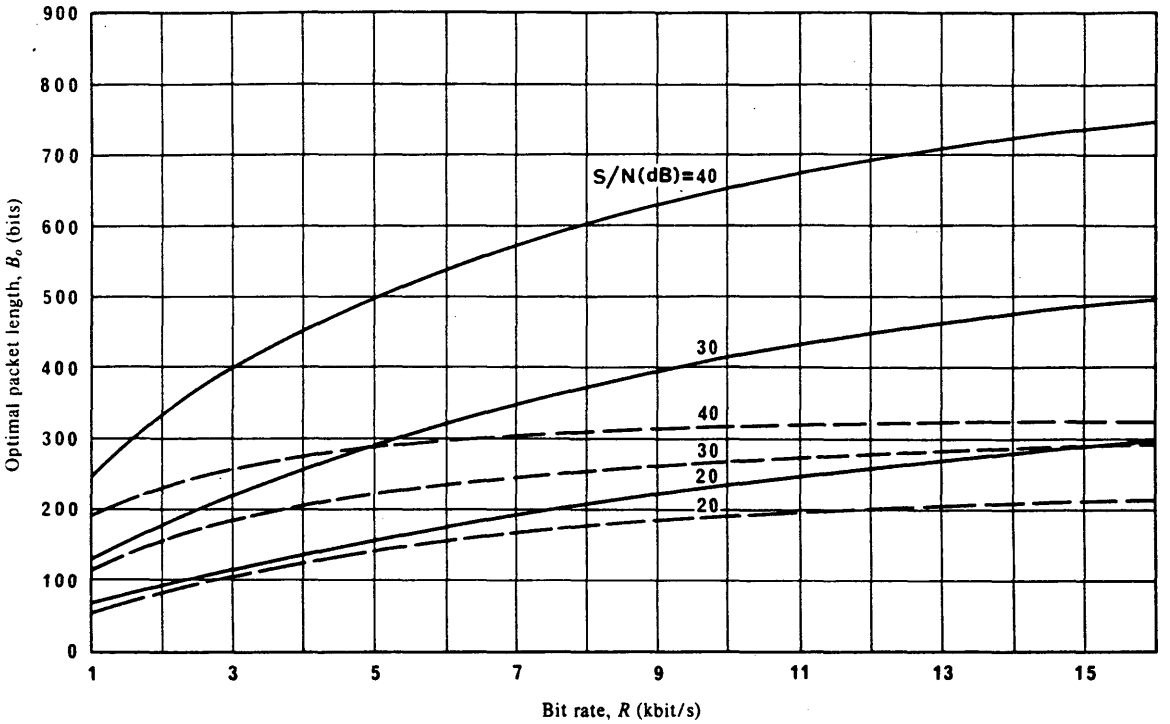


FIGURE 1 - *Optimal packet length versus bit rate*

Frequency: 850 MHz
Average message length: ——— 10 000 bits
— — — 1 000 bits

Vehicle speed: 50 km/h (30 mph)

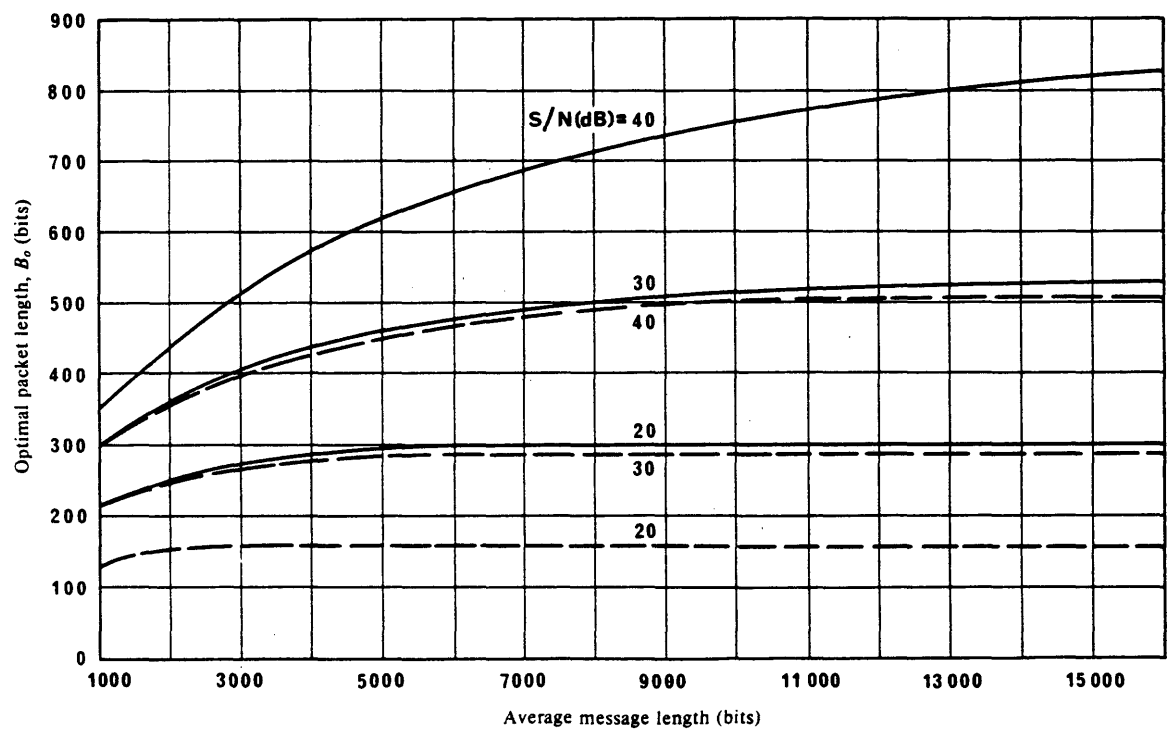


FIGURE 2 - *Optimal packet length versus average message length*

Frequency: 850 MHz
Bit rate: ——— 16 000 bit/s
— — — 4 800 bit/s

Vehicle speed: 50 km/h (30 mph)

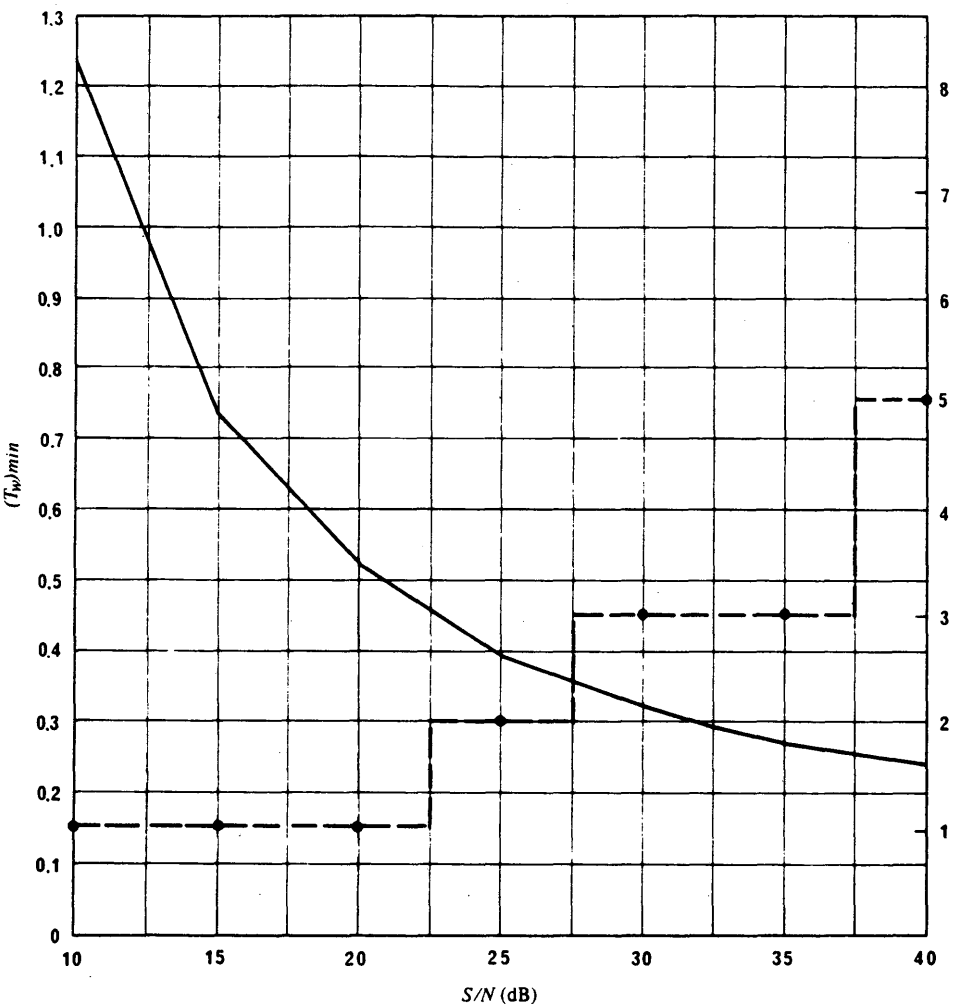


FIGURE 3 - *Minimum wasted time versus signal to noise ratio*

——— $(T_w)min$ = minimum wasted time relative to the message transmission time

— — — *Nopt* = optimal number of packets per transmission cycle

Packet length: 200 bits
Overhead length: 32 bits
Bit rate: 16 000 bit/s
Frequency: 850 MHz
Vehicle speed: 50 km/h (30 mph)

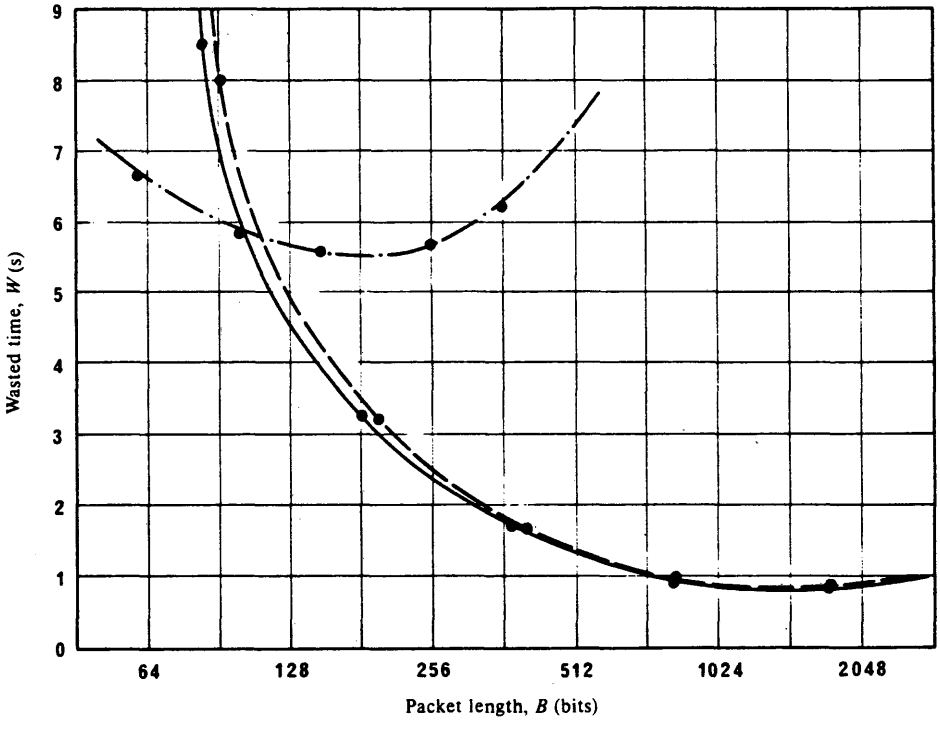
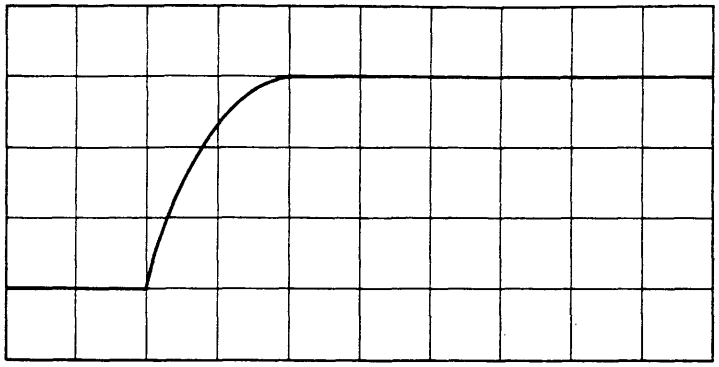


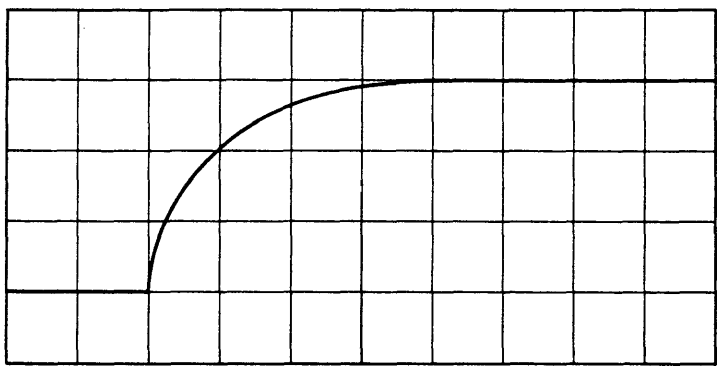
FIGURE 4 - *Wasted time versus packet length with and without error correction*

———: with error correction (P_r = 0.1)
— —: with error correction (P_r = 0.05)
—·—: with no error correction

Average bit error ratio: 10^{-2}
Bit rate: 4000 bit/s
Average message length: 2000 bits
Retransmission time out: 0.2 s.
Packet overhead: 30 bits
P_r: probability of retransmission
Doppler frequency: 25.5 Hz, frequency: 850 MHz, speed: 30 km/h (20 mph)



(*a*) No noise



(*b*) *C/N* = 6 dB

FIGURE 5 - *Response time of the modified COSTAS loop*

Horizontal axis: 2 ms/div.
Vertical axis : voltage controlled oscillator error signal

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REPORT 833

MULTI-FUNCTION TELECOMMUNICATION SYSTEMS

(Question 18/1)

(1982)

1. Introduction

Recommendation No. 65 of the WARC-79 recognizes that advances in technology are making practicable new sharing schemes that offer technological advantages for increasing the efficiency of spectrum sharing and band utilization. Use of recent developments in microprocessors, signal processing techniques, and digital modulation processes will allow more efficient use of the spectrum. Such developments permit the system designer the ability to develop equipment that can routinely support numerous functions using the same frequency band and waveform.

There are numerous areas in which systems that support multiple functions can be expected to appear. They encompass space applications, avionics, position location, telemetry, security, and others. Several services already exist which offer multiple functions. These include, but are not limited to, TACAN, the air traffic control radio beacon system (ATCRBS), the discrete address beacon system (DABS), and the cellular mobile radio system. In this latter system, in addition to the basic function of voice communications, other functions are performed including vehicle location and hand-off. As such, the location and hand-off functions have waveforms very much different from the primary function of the system.

These systems that support multiple functions can, and do, employ fully integrated waveforms which make it difficult to explicitly identify portions of the various signal structures. For example, a given signal structure can be used to convey information pertaining to identity, address, synchronization, etc. as well as the various message forms. This array of functions could be designed into the signal structure to utilize previously unused time slots.

The advantages to be achieved by multi-function systems are not attained without raising questions regarding electromagnetic compatibility (EMC) with existing systems since the new functions do not always fall into the same radio service.

2. Multi-function EMC considerations

In agreement with DECIDES 5 of Question 18/1, for systems that support functions which cannot be segmented and uniquely identified with specific portions of the waveform, new approaches have to be considered in sharing the spectrum with other users. It is, however, possible to identify the sharing potential of the composite waveform with other users in a common band based on an EMC analysis. Hence, frequency assignment support for systems having composite waveform characteristics could be sought based on the analysis showing that the additional services do not increase the systems interference potential. This approach complements current spectrum management practices, according to which traditional systems do not automatically receive frequency support (the right to operate on a specific frequency or over a band of frequencies) even though it is in accordance with national or international allocation tables.

3. Summary

The concepts employed in new systems that support several functions are extrapolations of multiplexing which have been in use for many years. Increased interest in these concepts have been fostered through technological advances in solid-state devices, microprocessor capabilities, the economic advantage of combining telecommunication functions into single equipments, requirements for more accurate and reliable capabilities, and the increased congestion being encountered in the electromagnetic environment. One approach to dealing with this new generation of systems has been presented. This approach, addressing the multi-function system, is based on the assessment of the systems' electromagnetic compatibility showing that the additional services do not increase the interference potential. This approach continues to provide protection to existing services while allowing the advantages of new technology to be realized.

REPORT 664-1

REVIEW OF SPECTRUM UTILIZATION BETWEEN ABOUT 40 AND 3000 GHz

(Question 64/1)

(1978-1986)

1. Introduction

Question 64/1 seeks information on spectrum usage and sharing criteria at frequencies above 40 GHz. Utilization of radio waves from about 40 to 3000 GHz has been severely limited due to the large atmospheric attenuation and the cost of hardware. Systems operating above 40 GHz will contribute to alleviating the present congestion in the use of radio waves. This analysis considers propagation characteristics, current usage, and technical problems in the realization of systems at these frequencies.

2. Atmospheric effect on wave propagation above 40 GHz

Attenuation of radiation by the Earth's atmosphere plays an essential role in the spectrum utilization above 40 GHz. Figure 1 shows the specific atmospheric attenuation (dB/km) for the paths near the ground as a function of frequency, which is given in Report 719. Gaseous absorption and rain attenuation are the two major effects on wave propagation at frequencies above 40 GHz.

Calculation techniques of the specific absorption produced by either the oxygen or water vapour components of the atmosphere are given in Report 719 up to the frequency of 350 GHz. Measured values up to 430 GHz [Liebe *et al.*, 1985] are generally within several percent of predictions, and the location of the maxima and minima are well known [Liebe, 1983]. The minor atmospheric constituents have also many absorption lines in the frequency band from about 40 to 3000 GHz and the values of attenuation due to these constituents are negligibly small. These lines, however, are quite important for radioastronomy and passive remote sensing. Therefore some line frequencies are selected for radioastronomical measurements to secure adequate protection as given in Recommendation 314.

The specific values of attenuation due to rain and water cloud are given in Report 721 up to 1000 and 200 GHz, respectively. Above 40 GHz, rain attenuation is primarily due to energy scattering.

Small-scale variations of the refractivity in the troposphere result in several propagation effects such as amplitude and angle-of-arrival variations due to multipath conditions, turbulent fluctuations, and tropospheric scatter. These effects are described in Report 718.

There is a lack of experimental data above 40 GHz, especially in the frequency range of decimillimetre wavelength (300-3000 GHz) and experimental studies of wave propagation in this range are urgently required.

3. Current and projected status of millimetre wave hardware

The difficulty in producing economical hardware has limited the utilization of the millimetre wave spectrum. Intensive research and development work, however, has been made to obtain practical millimetre wave hardware. Microwave and optical techniques have been extrapolated into the millimetre wave region, and also new ways of obtaining millimetre wave components are being introduced to overcome the problems inherent in conventional technologies.

Major barriers to hardware availability above 40 GHz are:

- lack of adequate transmitter power sources in practical packages;
- availability of reliable, low-noise receiver front ends;
- availability of RF circuit components – switches, filters, and transmission-line components – with low loss and sufficient bandwidth and power handling capability;
- high cost of all hardware;
- no capacity for production of large numbers of standardized components.

In the 40-100 GHz band, cost is the major problem. However, components are available from a large number of manufacturers world-wide and are utilized for system development, instrumentation, and for services and applications requiring small numbers of custom components, such as radioastronomy and specialized government applications.

Large investments have been made in development of systems for the mobile-satellite service using the frequency allocations above 40 GHz [Manz, 1984]. The increasing demand for radiolocation systems for such applications as collision avoidance, aircraft landing and adverse-weather tracking and imaging has stimulated development of components and manufacturing technologies which promise low-cost mass-produced hardware useful for all services above 40 GHz. New materials and devices, computer-aided design and manufacturing, the development of monolithic microwave integrated circuits (MMIC), and other advances in technology and understanding indicate that for the 40-100 GHz band, the hardware needed for many applications will be available at acceptable cost by 1990. Above 100 GHz, it can be expected that performance will improve markedly in the same time-frame, but hardware will continue to be available only in small numbers and at high cost. This situation is unlikely to change until needs equivalent to those below 100 GHz develop. Between 300 and 3000 GHz, it can be expected that only laboratory devices, radioastronomy equipments and instrumentation hardware will be available for the foreseeable future.

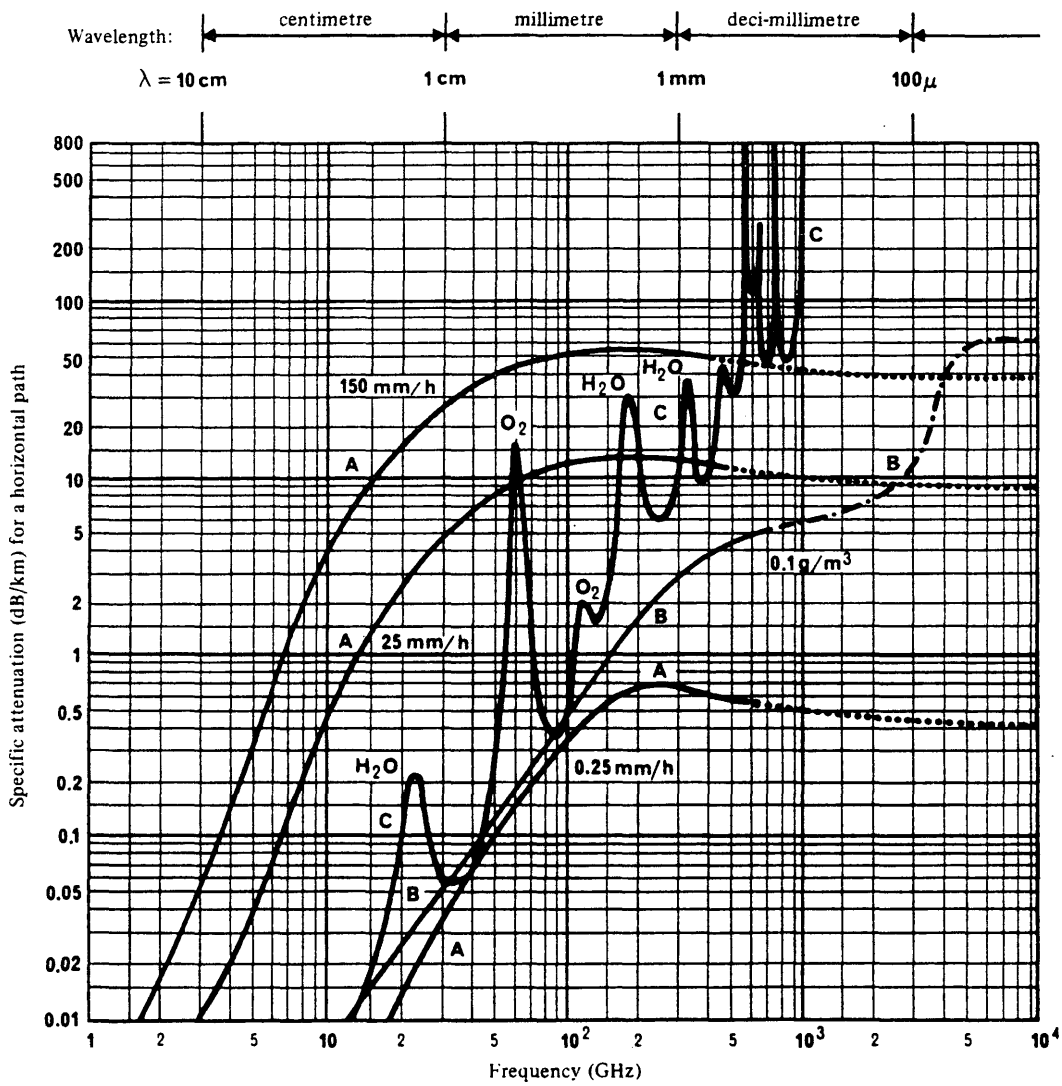


FIGURE 1 – *Attenuation due to gaseous constituents and precipitation for transmissions through the atmosphere*

Temperature: 20° C
Pressure: sea level: 1 atm
Water vapour: 7.5 g/m³

A: rain
B: fog
C: gaseous

3.1 *Transmitters*

3.1.1 *Tube and laser*

Figure 2 compares projected performance limits of various tube and laser technologies. In practice, these powers are generally not achieved at the higher frequencies. Advances in tube manufacturing technology and design will provide sufficient power for fixed services and many mobile services. Tubes generally fall into two classes, slow wave and fast wave. The TWT, klystron, magnetron, backward-wave oscillator (BWO) and extended interaction oscillator or amplifier (EIO or EIA) are slow-wave tubes. These tubes require an electron beam to pass closer than a fraction of a wavelength to a fragile periodic wave structure (scaled to the wavelength) or impinge either on the structure or on a delicate cathode. The tolerances required, the heating caused by stray electrons and high power density of the beam at millimetre wave frequencies limit the power output; but tubes with sufficient power for many services above 40 GHz have been developed. For example, a TWT developing 500 W over a 3 GHz bandwidth with a centre frequency of 44 GHz has been developed for mobile satellite up links [Glass, 1983]. Other TWTs have produced 50 W over a bandwidth of 93-95 GHz with a gain of 50 dB [Hieslmair *et al.*, 1983].

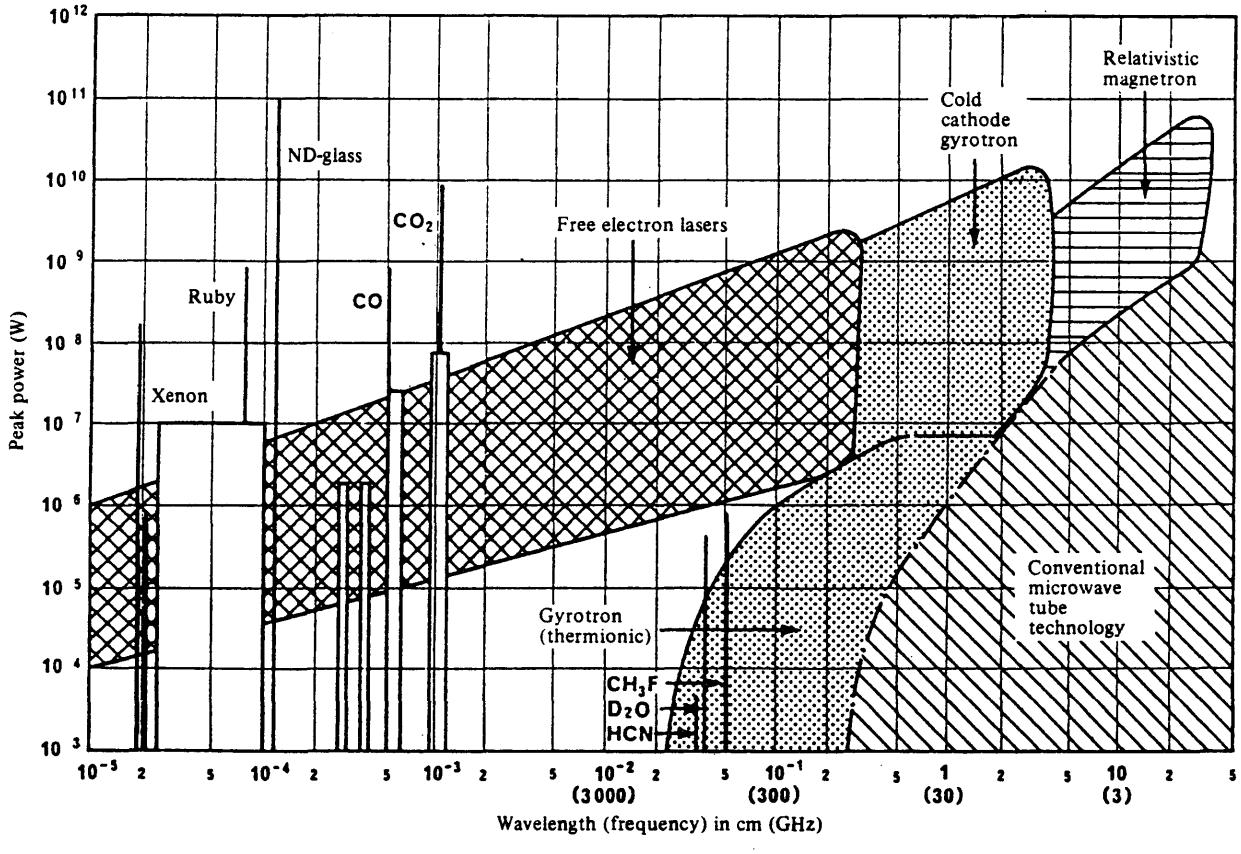


FIGURE 2 – *A comparison of technologies available for the generation of high-power coherent radiation*

The extended interaction oscillator (EIO) and amplifier (EIA) have the advantages of relatively small size and capability for operation above 100 GHz. These tubes have been used at 94 GHz, 140 GHz and 220 GHz, operating in both CW and pulsed mode. Typical peak pulse power is 1.5 kW at 94 GHz and 60 W at 220 GHz [Hieslmair *et al.*, 1983].

Gyrotron oscillators have been developed in the USSR, the United States of America and Japan for nuclear-fusion research and custom packages are available commercially. The large size of the superconducting magnets required for their operation has discouraged telecommunications applications, but development of suitable configurations is proceeding. Also under development are amplifiers, such as gyro-klystrons and gyro-travelling-wave tubes, which are more suited for telecommunications applications, although the high-power output of gyrotron oscillators makes them difficult to modulate.

A 60 GHz gyrotron with 200 kW of continuous output has been successfully tested. Under development is a 140 GHz continuous-wave device, available in late 1986 and a 120 GHz, 1000 kW device is expected to be available by 1990 [Loring, 1984]. Both are designed for fusion research. Powers in excess of 1.5 kW peak at 300 GHz have been reported.

Current gyrotron production is small, but even without taking into account communications applications, it is expected to grow 25% annually into the 1990s [High Technology, 1984]. Expected during the same period is the development and production of communications packages for fixed and some mobile services, such as maritime mobile, where the bulk of the magnets and power supplies can be accommodated. Figure 3 illustrates the state of the art of tube performance [Hieslmair *et al.*, 1983].

The gas lasers in current use are of the gas discharge and optically pumped types. An example of the former is the 890 GHz radiation from the HCN laser. Many oscillation lines in the 300-3000 GHz range, generated by optically pumped lasers, have been observed. Several milliwatts of output power are easily obtained in the laboratory from gas lasers and the D_2O laser is being developed as an ultra-high power source. Some difficulties exist, however, in the frequency tunability of these lasers.

3.1.2 *Solid state*

Solid-state sources offer the greatest potential for low-cost, mass-production, but despite large investment in their development, additional development is still required to reduce cost, increase power, and mass-produce reliable solid-state hardware. Impact avalanche transit time (IMPATT) diodes have produced the most useful power to date. Such oscillators have been operated up to 225 GHz. A peak power of 63 W has been reported at 93 GHz and 9.2 W peak at 140 GHz. These powers were obtained by combining the output of several diodes [Ying, 1983].

IMPATTs have undesirable FM and noise characteristics for applications such as local oscillators and coherent radars. Gunn oscillators (also called transferred electron oscillators) have the desired signal characteristics but produce far less power, e.g. 1 mW at 120 GHz. They are commonly used for local oscillators and low-power, short-range radars.

Development of the gallium arsenide (GaAs) field effect transistor (FET) is receiving a great deal of attention. Development of monolithic microwave integrated circuits (MMIC) using these devices would make possible low-cost, mass-produced millimetre wave systems; however, there is considerable uncertainty as to their commercial feasibility and advantages above 40 GHz. Above 100 GHz, development is focusing on indium phosphide (InP) devices. Figure 4 [Ying, 1983] illustrates the state of the art of solid-state sources.

3.2 *Receivers*

Low noise receivers are important for the development of millimetre wave systems. Receiver detectors in the 40-3000 GHz range are classified into three types according to their physical mechanism:

- diode detectors: Schottky barrier diodes, point contact diodes, MIM (metal-insulator-metal) devices;
- thermal detectors: Golay cells, pyroelectric detectors, bolometers;
- quantum detectors: Josephson junction devices, photoconductive detectors.

Widely used above 40 GHz are heterodyne receivers employing GaAs Schottky barrier diodes. Development of beam lead configurations has yielded much more rugged and reliable devices than point contact diodes. The beam lead devices also lend themselves to mass production by integrated circuit techniques. Low-noise preamplifiers preceding the mixer are not commonly used above 40 GHz because of lack of hardware and its cost and bulk. Signals are amplified at lower IF frequencies, using GaAs FET, parametric and other types of low-noise amplifiers. Cryogenic cooling is utilized in some applications. Cryogenically cooled GaAs Schottky mixers with noise figures of approximately 0.22 dB up to 150 GHz have been reported. Development of diodes for use above 300 GHz is reported, and conversion losses of 17.1 dB at 692 GHz and 18 dB at 2520 GHz have been obtained.

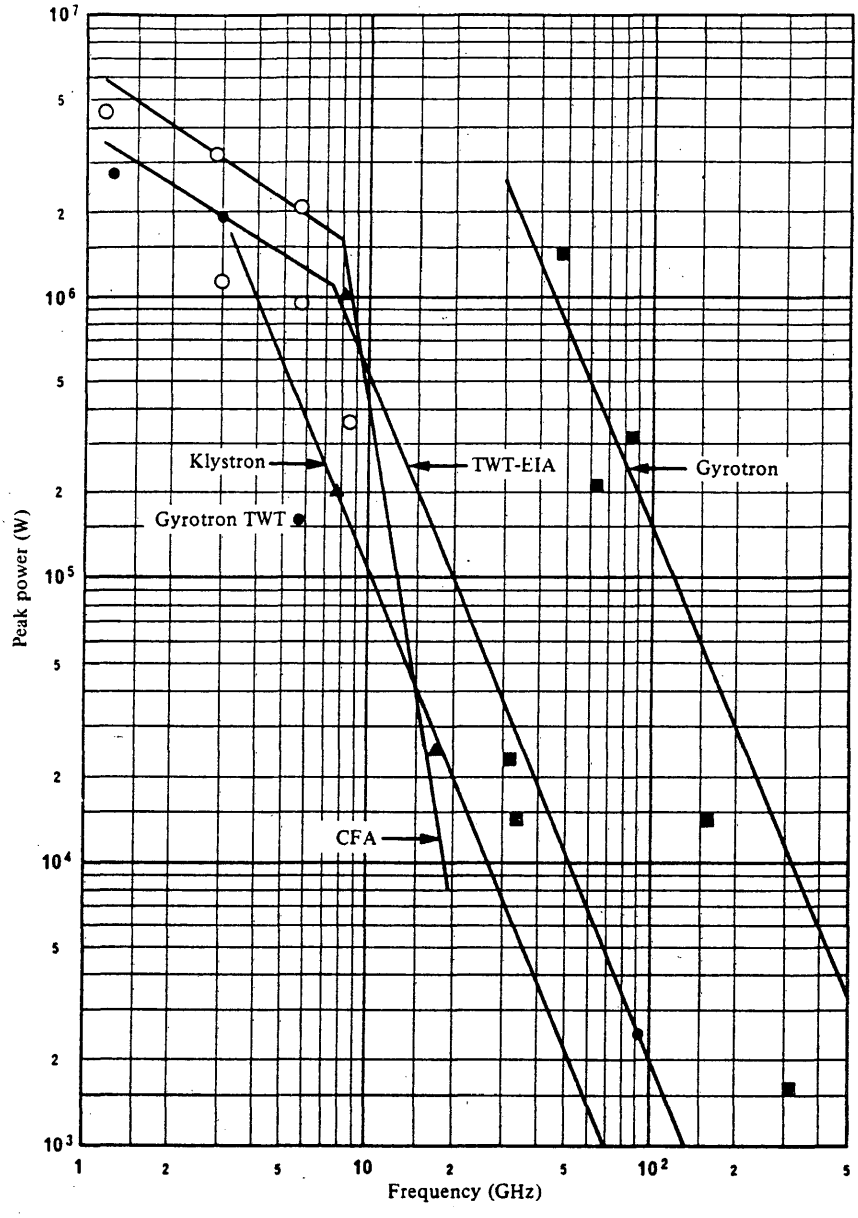


FIGURE 3 – *Tube performance data*

○ cross-field amplifier (CFA)
● travelling wave tube (TWT)
▲ klystron
■ gyrotron
EIA: extended interaction amplifier

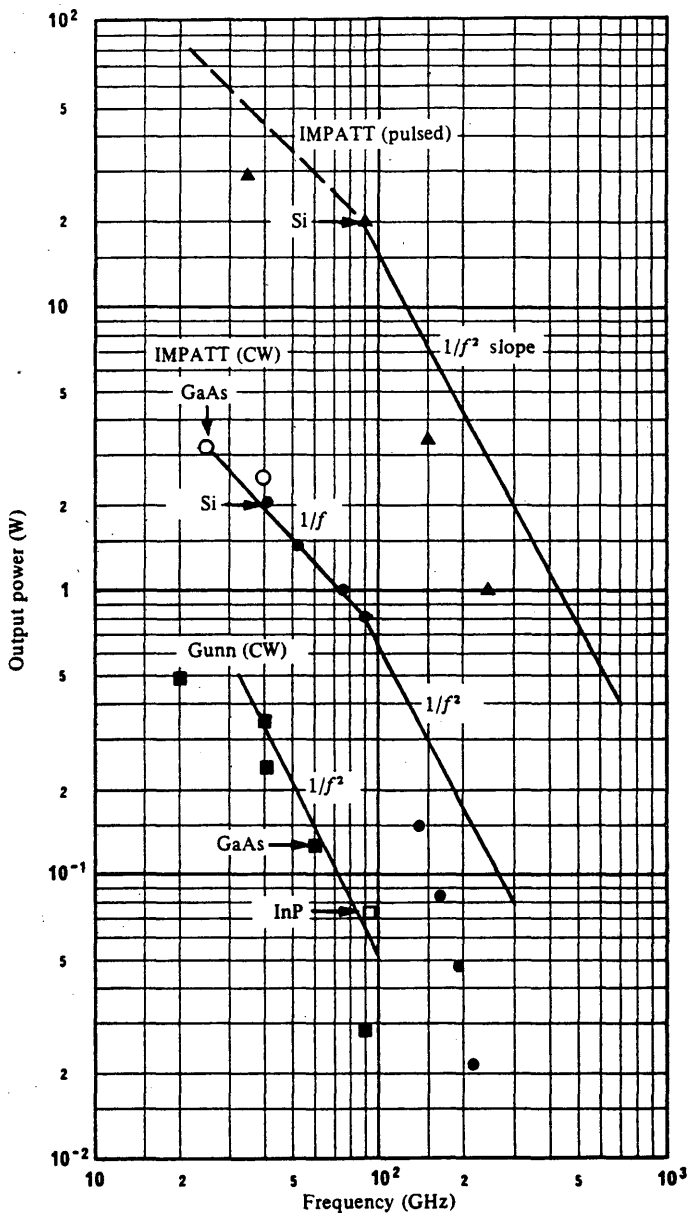


FIGURE 4 – *Solid-state performance data*

GaAs: gallium arsenide FET
Si: silicon devices
IMPATT: impact avalanche transit time diodes
InP: indium phosphide devices

Thermal detectors have advantages of wide frequency coverage, but their response time is slow and sensitivity is relatively low. Quantum detectors for use at cryogenic temperatures are being developed. Josephson junction devices with mixer noise temperatures of 54 K at 36 GHz, 120 K at 115 GHz, and 823 K at 300 GHz have been obtained.

The state of the art of receiver front-end performance is summarized in Fig. 5. Receiver performance is currently adequate for most applications, but cost, producibility, and reliability are problems to be overcome if widespread use of the spectrum above 40 GHz is to take place.

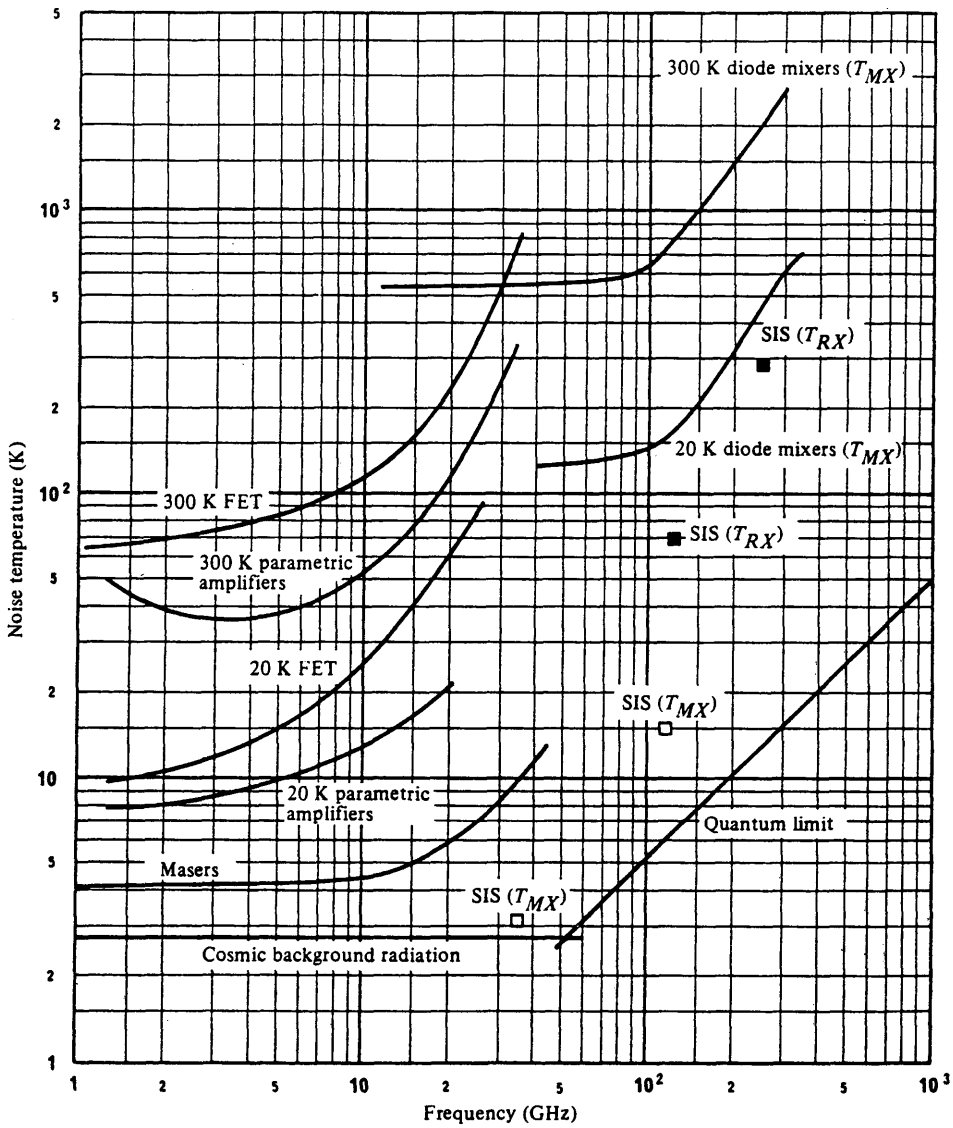


FIGURE 5 – *Noise performance of SIS (superconductor-insulator-superconductor) junction mixers*

T_{MX} and T_{RX}: noise temperatures of mixer only and whole receiver, respectively

(Performance of other types of detectors are shown for comparison)

3.3 *Antennas*

Present technology allows antennas to be made with a precision of δ (surface r.m.s. error)/D (diameter) ratio of 3×10^{-5} and makes it possible to use large antennas for this frequency band and high gain and small beamwidth are easily obtainable in antennas of small size. Table II gives examples of large millimetre and decimillimetre antennas.

Recent developments in active array antenna technology could reduce the requirement for production of high power from a single device. A low-power source (1-10 W) can be provided for each radiator (Fig. 6) to enable the array to produce a large effective radiated power (e.r.p.). Feasibility of such antennas depends on development of low-cost, mass-produced MMICs which include output power devices in the 1-10 W range.

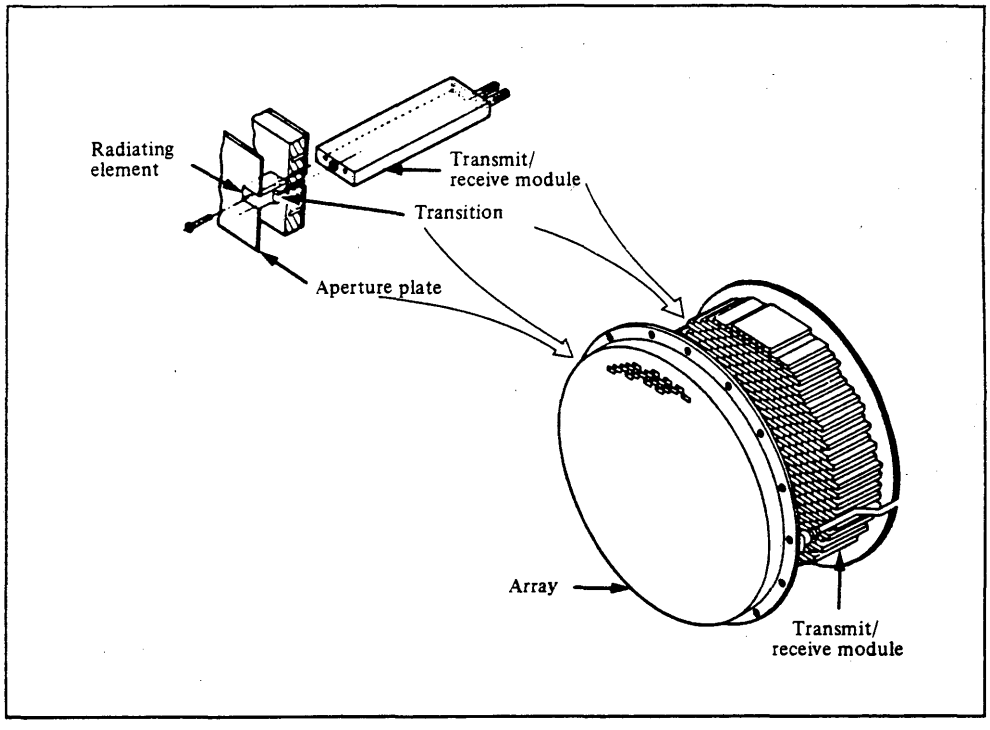


FIGURE 6 – *Active array concept*

4. Usage above about 40 GHz

Atmospheric attenuation and unavailability of equipment with adequate cost and performance, limit the usage of frequencies above 40 GHz. A number of systems and services such as terrestrial radiocommunication systems, fixed and mobile satellite systems, radioastronomy, space research, radiolocation and radionavigation services, use these frequencies. The radioastronomy service uses this frequency band extensively. Frequencies above 300 GHz especially were pioneered by radioastronomers and are now being widely used.

4.1 *Terrestrial radiocommunication systems*

It can be expected that improvements in industrial manufacturing techniques will increase spectrum utilization above 40 GHz. Although the problems concerned with large-scale production of equipment above 100 GHz have not been resolved, it can be anticipated that small numbers of specialized quasi-experimental systems will be built for radioastronomy and research purposes above 100 GHz.

The expanded use of the millimetre-wave spectrum for terrestrial fixed services has been slowed by competition from fibre-optics communications. Fibre optics also offer wide bandwidths and are suitable for the same short range relay systems as millimetre-wave systems. Extensive, long-haul, fibre-optics systems are being installed in the United States of America, Europe and Japan. The use of fibre optics in local area networks, another application where millimetre waves have advantages, is also expanding.

Millimetre-wave mobile services for telephone and low data rate transmission have to compete with cellular radio which uses low power and modern digital switching equipments to provide the same sharing and freedom from interference obtained by operation at millimetre wavelengths. For systems requiring high data rates

such as video and high-rate digital data transmissions, use of the spectrum above 40 GHz can be expected to expand, once affordable equipment becomes available.

Currently, radiocommunication systems in the frequency band from 38 to 40 GHz, as mentioned in Report 783, are operationally used for communications between governmental organizations for disaster prevention, data transmissions between a host computer and terminals, and some other systems. Some experimental tests have already been conducted in this frequency band for high-definition television transmission and for field pick-up (FPU).

In the 50 GHz band, simple communication systems have been proposed for the transmission of video, audio, and other data signals. Several hundred of these systems have been used in Japan where alternative methods of signal transmission are not readily available [Furuhama and Shimanuki, 1984].

A short-range still-image transmission system using the 60 GHz frequency band where strong absorption lines due to oxygen exist, has been produced as part of a communication system with vehicles in an overall system of traffic control. Major specifications of the above-mentioned system are listed in Table I. Both 57 and 96 GHz terrestrial systems are operating at the Institute for Telecommunication Sciences, National Telecommunications and Information Administration, Boulder, Colorado, USA.

TABLE I – *Examples of terrestrial radio systems at frequencies from 40-60 GHz bands*

Frequency (GHz)	36.5-39.5		50.4-51.4		59-64	
Output power (dBm)	10-17	26	11	16-24	5	17
Modulation	FSK PSK SS-FM	FM	FM	PSK FSK ASK	PSK-AM	FM-AM
Transmission capacity	1.544 Mbit/s 6.313 Mbit/s 300 channels	Colour TV: 1 channel (30 MHz)	Colour TV: 1 channel Audio: 1 channel	30 Mbit/s 6.3 Mbit/s 1.544 Mbit/s	2 Mbit/s	Audio: 8 channels
Antenna	30-40 cm diameter parabola	40 cm diameter parabola	30 cm diameter parabola 3.6 × 3.0 cm horn	30 cm diameter parabola 3.6 × 3.0 cm horn	Horn	10 cm diameter bicone antenna

4.2 *Satellite communication services*

Allocations currently exist above 40 GHz for the mobile-satellite, fixed-satellite, inter-satellite and broadcasting-satellite services. Large investments are being made in the development of mobile satellite systems using the 43.5-47 GHz allocation. Currently, the use of this band is only planned for up links as space qualified hardware for the band is not yet sufficiently mature. The associated down links will use the 30/20 GHz bands. This usage is expected to stimulate the development of low-cost hardware for other bands and services above 40 GHz [Manz, 1984].

A 3 GHz transmission bandwidth is available in the 50/40 GHz band and the investigation of an experimental communication satellite using this frequency band has been started [Shimada *et al.*, 1984].

Frequencies for inter-satellite transmissions currently under consideration are 60 GHz (5 mm), 28 300 GHz (10.6 μm), and 566 000 GHz (0.53 μm); operation at 60 GHz will minimize interference due to the peak in atmospheric absorption from oxygen, whereas the optical frequencies make good advantage of existing technology [Pritchard, 1977]. The first test of an inter-satellite relay took place in the 36-38 GHz band.

4.3 *Radioastronomy service*

Over one thousand molecular spectral lines have been observed in the 40-300 GHz range and many more above 300 GHz. Observations of these spectral lines provide vital information on the cool gas matter in the universe, which is not accessible by astronomical observations using other techniques. Continuum emission is radiated from cool dust and compact high energy quasars sources which are of unique astrophysical importance.

Many observatories operate millimetre and decimillimetre wave telescopes, examples of which are shown in Table II (single dish telescopes) and Table III (interferometers).

TABLE II – *Millimetre and decimillimetre wave radio telescopes (Single dish telescopes)*

Organization and location	Diameter (m)	Surface accuracy (mm, r.m.s.)	Maximum frequency (GHz)
Commonwealth Scientific and Industrial Research Organization, Australia	4	0.09	200
University of British Columbia, Canada	4.5	0.13	150
Bell Telephone Laboratory, United States of America	7	0.1	190
Owens Valley, United States of America	10.4	0.035	600
National Radio Astronomical Observatory, United States of America	12	0.07	430
Itapetinga, Brazil	13.7	0.3	100
Onsala, Sweden	20	0.18	150
Nagoya, Japan	4	0.05	450
Nobeyama, Japan	45	0.18	120
Institute Research Astronomy Millimetric (IRAM), Federal Republic of Germany, France and Spain	30	0.09	350
Science and Engineering Research Council (SERC), United Kingdom (1)	15	0.07	500
Caltech, United States of America (1)	10.4	0.015	1000

(1) Under construction.

TABLE III – *Millimetre and decimillimetre wave telescopes (Interferometer)*

Organization and location	Element antennas: diameter (m) × number	Maximum spacing (m)	Frequency (GHz)
Owens Valley, USA	10.4 × 3	430	88-120
Hat Creek, USA	6 × 3	300	88-115
Nobeyama, Japan	10 × 5	560	22 and 115
IRAM, France (1)	15 × 3	390	70-375

(1) Under construction.

4.4 *Space research service*

The space research service concerns itself mainly with deep-space and near-Earth satellite research. Telecommunication links are used to communicate with and control both manned and unmanned satellites. Recommendations 364 and 578 specify the protection required for near-Earth and deep-space research satellites.

Passive measurements of the Earth and its atmosphere are made at millimetre wavelengths by ground-based instruments and instruments carried by satellites, aircraft and balloons. Frequencies as high as 60 GHz have been used in ground- and satellite-based equipments and measurements using frequencies as high as 300 GHz may occur within the next five years. The TIROS-N system became operational in 1978. Measurements at frequencies as high as 220 GHz have been made from aircraft. Preferred frequency bands for the passive services have been identified up to and above 300 GHz. However, the use by these services of frequencies above about 60 GHz is highly experimental and of a limited nature.

4.5 *Radiolocation and radionavigation services*

Radiolocation has stimulated much millimetre-wave hardware development, particularly that of solid-state components. Millimetre-wave radiolocation systems obtain high resolution with small apertures. Their small beamwidths and atmospheric attenuation reduces the possibility of interference. Above 40 GHz, radiolocation development is concentrated in specific spectral regions of relatively low atmospheric attenuation, such as the 92-100 GHz band. The systems in these bands are useful for applications requiring high resolution in distance and direction, such as short-range airport surveillance radar. In addition to this use, a mobile Doppler radar system for measuring speed using the 50-60 GHz band has been developed to monitor vehicle skidding. The system is very useful because of its high accuracy and compactness. An experimental imaging radar has been tested at 94 GHz. Exploratory development work is in progress in the 225 GHz band.

Navigation systems for mobile use in urban areas have been proposed for resolving heavy traffic congestion by using a 60 GHz transmission link, in which the repeated utilization of the frequency at each cross road is possible without any interference. Cost reduction is expected to stimulate expansion of radiolocation and radionavigation services using these frequencies.

5. Factors affecting spectrum sharing above 40 GHz

Usage of the 40-100 GHz spectrum is accelerating, but the number of systems put into service is not expected to be large enough to make sharing a significant problem until the year 2000. The small beams of systems above 40 GHz, the use of relatively low power, and atmospheric attenuation will provide sharing possibilities that exceed requirements during this period. In the spectrum above 100 GHz, the values of attenuation and high directivity will reduce the possibility of interference even further.

Additional knowledge is required in the following areas:

- effects of the structured atmospheric spectrum,
- effects of rainfall, fog, anomalous propagation, and other attenuation mechanisms on spectrum usage,
- achievable radiation patterns of antennas,
- achievable power levels and bandwidths,
- achievable receiver characteristics.

6. Conclusions

Developments leading to utilization of the spectrum above about 40 GHz are important in order to alleviate increasing congestion in the spectrum below this region; and, although low-cost equipment for frequencies above 40 GHz is only now becoming available and achievable performance levels are not yet satisfactory, a number of advantages resulting from utilization of these frequencies can be predicted. These include:

- large available bandwidths,
- narrow or multi-beam antenna patterns from physically small apertures, and
- unique applications which may result from the highly structured atmospheric attenuation.

Atmospheric attenuation and adverse weather effects above 40 GHz are significant and highly dependent on the specific operating frequency. A number of phenomenological questions require deeper investigation for better utilization of the spectrum above 40 GHz, and to encourage its use.

Use of the 40-100 GHz bands is likely to expand rapidly with all services being well represented by the 1990s. Investment and the current state of the art indicate that space services, particularly in the 40-50 GHz band will make the first and most heavy use of these frequencies. The radiolocation service can also be expected to make increased usage of the 92-100 GHz band in the same time period. The usage by terrestrial communications services of frequencies above 40 GHz will be affected by competition from fibre-optics systems and technologies such as cellular radio. However, as video and computer communications needs increase, the use of these frequencies is also likely to increase for mobile and remote site communications services for which other technologies are not suitable. Hardware for frequencies greater than 100 GHz will not be available on a large, commercial scale in the near future. However, these frequencies will be used extensively for radioastronomy.

For frequencies above 40 GHz, particularly those above 100 GHz, spectrum sharing will be much easier than at lower frequencies because of the large available bandwidth, increased atmospheric attenuation, and high beam directivity at these frequencies.

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REPORT 666

SYSTEMS FOR TELECOMMUNICATIONS USING THE ELECTROMAGNETIC SPECTRUM ABOVE 3000 GHz

(Question 53/1 and Study Programme 53B/1)

(1978)

1. Introduction

Many kinds of useful systems for telecommunication using the infrared and visible spectrum already exist or are being developed. In order to implement some of these systems, electromagnetic wave sources of infrared and visible light radiation, e.g. lasers, are necessary. Many kinds of lasers have been developed. This Report contains a brief outline of laser system characteristics and applications and of the other systems using the spectrum above 3000 GHz.

2. Components

2.1 *Laser sources*

Laser oscillators of various types have been found to generate electromagnetic waves in the frequency spectrum extending from millimeter waves through the infrared, the visible, the ultraviolet, and approaching the soft X-ray region.

The radiation can be generated from some eight basic laser types including the dye laser and the semi-conductor laser, and electromagnetic waves of many frequencies from the ultraviolet up to and including the infrared regions can be obtained by using the non-linearity of materials, involving techniques for frequency multiplication, light mixing, parametric oscillation, etc.

In order to utilize lasers for some practical applications (especially telecommunication), CW laser oscillators with good stability, high efficiency and long life are required. Only a fraction of the lasers comply with these requirements. Table I shows the wavelengths and characteristics of typical laser sources.

TABLE I – *Laser sources and their characteristics*

Classification	Material	Oscillating wavelength (μm)	CW output (dBW)	Pulse output (dBW)	Efficiency (%)
Semiconductor laser	Ga Al As	0.85 - 0.9	−20	10	10
Solid state laser (1)	Nd: YAG	1.06	20		1
	Nd: glass	1.06			
	Ruby	0.6943		80	0.1
Liquid laser	Dye	0.33 - 1.1		20	0.1
Gas laser	N_2	0.3371			
	He-Ne	0.6328	−10		0.1
	CO_2	10.6	20		15
	Ar	0.5145 0.4880	10		0.02
	He-Cd	0.4416	−10		0.03
LED (2)	Ga Al As In P/GaAs In Ga As	0.85 - 0.9	−30		3

(1) There are NdP, LNP which are excited by LED and which have a high efficiency.

(2) LED is not a laser, but as it is used frequently as a light source, the characteristics are listed here for reference.

2.2 *Modulator*

There are two methods of modulation. In the one method, the laser source, which may be a semiconductor laser, light emitting diode etc., is directly modulated. In the other method, external modulation is achieved using electro-optical, magneto-optical or acousto-optical effect, or by the use of semiconductor modulators.

2.3 *Detector*

Table II shows various items of detectors with typical characteristics.

2.4 *Transmitting and receiving antennas for laser*

Both refractors and reflectors are used as optical antennas in laser systems. Usually, a telescopic system is convenient for both transmitting and receiving antennas.

Keplerian telescopes have converging lenses and Galilean telescopes have diverging lenses. For reflectors, the Cassegrain type and its modifications with off-axis antennas (for example, Herschelian type) are more conveniently used than Newtonian type.

The aperture of a receiving antenna is mainly determined by the required antenna gain.

To avoid vignetting effect, the aperture D for the transmitting antenna should be determined by the following equation:

$$D \geqslant d \frac{\textcircled{H}}{\theta} \quad (1)$$

where

d: aperture of laser source

$\textcircled{H}$: beam divergence of the laser source

θ: required antenna beamwidth.

For a high power laser, it may be necessary to take precautions against high field strengths occurring which may result in dielectric saturation or breakdown. Also, the surface reflection may interfere with laser oscillation.

The second reflector of a Cassegrain type optic increases the optical loss by shading the central portion of the laser beam. Therefore, the off-axis type reflector is preferred.

Fresnel lenses and mirrors are also applicable for light weight and low cost systems.

TABLE II – *Detectors and their characteristics*

Classification	Material	Wavelength (μm)	Detection bandwidth	Sensitivity $[D_\lambda^*]$ max $(cm(Hz)^{1/2} W^{-1})$
PD (Photo diode)	Si, Ge	0.7 - 2	10 GHz	
APD (Avalanch PD)	Si, Ge	0.4 - 5	1 GHz	
PV (Photo-voltaic)	InSb	1.5 - 5.5		7×10^{10}
PT (Photo transistor)	Si	0.7 - 2	1 MHz	
PC (Photo conductor)	PbS	1.5 - 3.2	10 MHz	1×10^{11} (λ = 2.5 μm)
	InAs	1.5 - 3.8		5×10^{9} (λ = 3.5 μm)
	PbSe	1.5 - 4.5		4×10^{9} (λ = 3.7 μm)
PE (Pyro-electric detector)	$BaTiO_3$, $LiTaO_3$	0.7		
PM (Photo multiplier)	GaAs, Sb·Cs, Ag·Cs	0.1 - 1.1	100 MHz	

$$^* D_\lambda = \frac{S/N}{P_\lambda} \cdot \frac{\sqrt{\Delta f}}{\sqrt{A}} \quad (cm\ (Hz)^{1/2}\ W^{-1})$$

S: Output signal voltage (V)
N: Output noise voltage (V)
P: Input power of light having wavelength λ ($W \cdot cm^{-2}$)
A: Detector area (cm^2)
Δf: Bandwidth of amplifier (Hz)

3. Systems

3.1 *A colour television system*

A prototype atmospheric transmission system using a diode laser (near infrared light) which can simultaneously transmit 4 MHz colour television signal and a voice signal using Pulse Interval Modulation (PIM) was developed and field tested over a period of one year [Ueno and Yasugi, 1976].

The principal characteristics of this system are shown in Table III.

TABLE III – *Characteristics of an existing telecommunication system*

Modulation	PIM-IM
Transmitting power (peak value)	−4 dBm to +6 dBm (automatic power control)
Optical pulse width	7 ns
Laser diode	GaAlAs laser (0.83 μm wavelength)
Photo detector	Avalanche photo diode (silicon)

Figure 1 shows the lightwave propagation characteristics in Tokyo (2.7 km distance, during 3 months in autumn) and in Niigata (0.6 km distance, during 2 months in snow season).

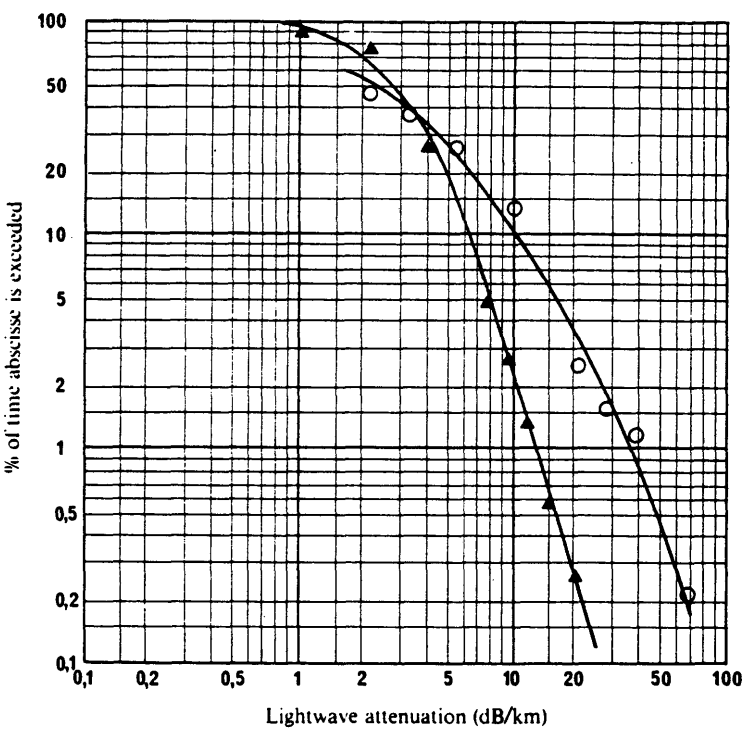


FIGURE 1 — *Cumulative lightwave attenuation distribution*

▲ in Tokyo

○ in Niigata

Assuming an allowable S/N for colour television signal transmission of 30 dB, circuit availability values of 99.4% in Tokyo and 99.7% in Niigata were obtained, respectively. Based on data obtained, it is estimated that the lightwave atmospheric transmission availability of the circuit at a distance of 2 km in Tokyo is 99.9%.

3.2 *A system for detection of air pollutants*

Several schemes for detection of air pollutants using lasers have been developed. Interactions of light with particles, which are termed

(a) Raman scattering,

(b) resonance fluorescence, and

(c) resonance absorption, are utilized.

Scattering cross sections due to (a) and (b) are so small that detection schemes can be useful only for high particles concentration which, for example, occurs near a smokestack. The differential absorption scheme due to (c) is the most promising in order that the ordinary concentrations of pollutants in the atmosphere (below 0.1 ppm) can be detectable [Inomata and Igarashi, 1975].

This detection method of air pollutants using differential absorption has been implemented. For example, Fig. 2 shows one of the experimental results obtained with CO_2 laser. In this figure, the solid line corresponds to the result with a commercially available ozone meter placed nearby the laser system. There is good agreement between the results by CO_2 laser and those by ozone meter [Asai and Igarashi, 1975].

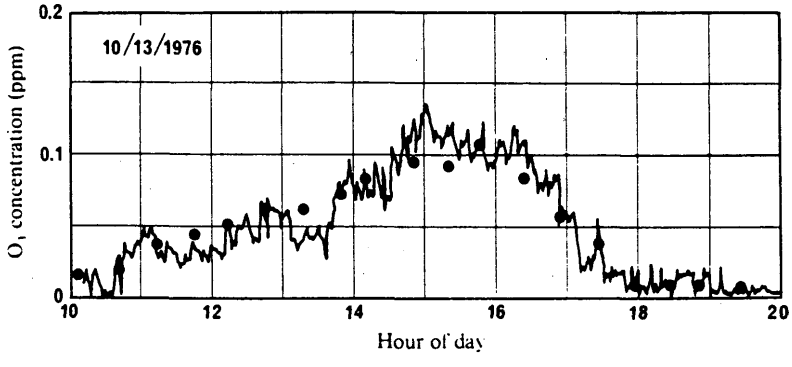


FIGURE 2 — *Long path ozone monitoring with differential absorption method on 13 October 1976*

Other laser systems for measuring pollutants, trace gases, meteorological and aerodynamic properties make use of such lasers as the ruby/dye, helium neon and CO_2 lasers. These determine the location, amount, velocity, etc. of such substances as H_2O, CO, SO_2, N_2O, SF_6, HCL, NH_3, C_2H_4, freon, methyl alcohol, acetone, aerosols, as well as NO_2 and O_3 [ILE, 1975].

3.3 *A system for a range-gated laser oscilloscope (under-water detection)*

When light travels in water from a light source to a target, strong backscattered light come back to the receiver and tends to degrade the target definition. Backscattering of light increases the difficulty of under-water detection.

If these light components can be rejected, the target can be detected very clearly as a self-illuminated subject. By using very short duration and high intensity laser pulses and an ultra-fast gated receiver, the backscattered light (in the time domain) can be rejected.

3.4 *A system for space-determination*

Practical systems for determining the distance from the Earth to the Moon or a spacecraft measure the travelling time of laser pulse which is reflected from a corner cube located on the Moon or spacecraft.

The characteristics of these systems are shown in Table IV.

TABLE IV – *Characteristics of systems for space-determination*

	A system for determination of the Moon	A system for determination of a spacecraft
Wavelength and laser oscillator	0.6943 μm ruby	0.6943 μm ruby
Output energy (power)	8 joules (400 MW)	0.5 joule (120 MW)
Type of signal	20 ns 0.2 pulse per second	4 ns 0.1 pulse per second
Transmitting optics antenna	(1) ø = 50 cm	ø = 10 cm
Receiving optics antenna	ø = 3.8 m	ø = 50 cm
Accuracy	$\pm 4 \times 10^{-9}$	$\pm 4 \times 10^{-8}$

(1) ø: Diameter of antenna.

3.5 *Aircraft wake and wind profile system*

A scanning laser Doppler velocimeter system has been used to detect and measure the velocity structure and movement of aircraft wake vortices as well as vertical wind profiles. The coherent emission beam is split with part of it going for use as a local oscillator. A Bragg cell is used to shift the frequency for Doppler shift observation. Beam polarizations and mirror scanning are utilized. The main beam returning from the atmosphere is detected and passed through a signal processor, resulting in a Doppler velocity distribution for the aerosols tested. Such signals for all involved space points are further processed for appropriate display. Characteristics are listed in Table V [Jeffreys and Bilbro, 1975].

TABLE V – *Aircraft wake and wind profile system characteristics*

Laser source	CO_2 (10.6 μm)
Output power	20 W (CW)
L.O. shift	24 MHz
Detector	Hg Cd Te
Detectable Doppler velocity	61 m/s (max.)

3.6 *Other than laser systems*

There are a number of systems operating beyond 275 GHz that do not include laser sources. Many of these are passive systems. These have receivers sensitive primarily to the infrared and/or visible regions, with some extending into the ultraviolet and X-ray areas. These systems are used and planned for use in meteorology, earth exploration, space research, radio astronomy, navigation, etc. Examples such as those in the next two sections are plentiful, including proven operational systems as well as those in experimental phases.

3.7 *Forest fire location system*

An aircraft is used to carry a thermal imaging system to locate the position of a forest fire and to survey the area, day or night. This small Forward Looking Infrared (FLIR) system uses a multifaceted scan mirror and a nodding mirror to scan the image from the forest across the detector arrays. The signals from the detector arrays feed a TV display. Characteristics of the system are seen in Table VI [DeLangre, 1975].

TABLE VI – *Forest fire location system characteristics*

Source	Infrared from forest
Field of view	30° × 40°
Spatial resolution	2 milliradians
Bandwidth	5 MHz
Detectors	Hg Cd Te (8-12 μm)
Power	200 W

3.8 *Low light level astronomy*

A recent example of a radio astronomy operating in the visible spectrum is the use of Charge Injection Devices (CIDs) with the USA Kitt Peak National Observatory telescopes. In one case, a silicon array of CIDs was used with the 4 meter Mayall Telescope at the Cassegrain focus for low light level study of a planetary nebula. The CID device using non-destructive readout was found to be extremely promising for astronomical purposes. The system operates with a Varian 620F computer [Aikens *et al.*, 1975].

4. System considerations beyond 3000 GHz

Systems operating in this region can be classified generally as passive or active. The passive systems have a longer history, with some being centuries old, and many others just on the horizon. Capabilities and limitations are better known for some of the passive systems. Active systems in most cases involve lasers, which first became available in 1960. Some lasers are useful now and others need further development. However, capabilities and limitations are not as well known for the active systems.

Among considerations which vary with frequency in this region are power output, beam shape, bandwidth, effects of atmospheric gases, effects of particulates, aerosols, humidity, turbulence, background radiation, detector sensitivity, etc. These and other parameters determine system practicality.

5. Minimal problem nature

The previous examples of the systems operating above 3000 GHz are exemplary of the many systems now operating and expected to operate in the decades ahead. These systems appear to have relatively few interference problems. There have been practically no cases of interference cited for the new laser systems primarily because of the high beam directivity and low side and back lobes. Careful design has resulted in essentially interference-free and, in some cases, worldwide operation of the passive systems, and this trend is expected to continue. In the centuries-old case of optical astronomy, there appear to have been no unmanageable interference problems; and in the developing laser case of fibre optics transmission, the spectrum system will be guided within the fibres while operating effectively interference-free.

6. Applications and related services interest

The foregoing applications, representing a type with potential for future improvement, have been correlated with possible interest of some of the existing services in Table VII. It is significant to emphasize that while telecommunications includes the entire electromagnetic spectrum, radio telecommunications currently is limited to the spectrum below 3000 GHz.

TABLE VII – *Potential applications and related services interest*

Applications \ *Services*	Fixed	Aeronautical fixed	Earth exploration satellite	Meteorological aids	Meteorological satellite	Location	Navigation	Space research	Astronomy	Broadcasting
Color television	x									
Pollutants Trace gases Meteorological Aerodynamics		 x	x x	 x	 x	x x		x x		
Underwater imaging						x	x			
Space determination						x	x			
Wind profile Acoustic wake						x x				
Other than lasers			x		x	x	x	x	x	
Forest fire location						x	x			
Low light level astronomy									x	

7. Summary

A large number of systems exist and are being developed in the spectrum above 3000 GHz. Whereas there already is a large investment, the potential exists for many more systems. Some useful passive systems are operational, while many laser systems are still being developed. These systems are of possible interest to several services. Applications such as those in this Report are contributing, and will continue to contribute, to the alleviation of congestion in the spectrum below 3000 GHz.

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REPORT 667

INTERFERENCE OF OPTICAL COMMUNICATION LINKS DUE TO DIRECT AND SCATTERED RADIATION

(Question 53/1)

(1978)

1. Introduction

This Report discusses the modes of interference which can exist between optical communication links, and which are caused by direct and scattered radiation effects. In this regard, twenty-four distinct interference situations are identified and analyzed. Drawing upon available information on relevant optical phenomena, worst-case interference estimates are established for most of the modes. Further, coordination restrictions on transmitter power and link geometry are investigated with the intent being to eliminate interference in the sense that it is reduced to tolerable levels under practically all meteorological conditions.

Because of the extremely narrow beamwidths employed in optical systems, few interference problems are expected. Those few which do arise are most simply resolved by coordination of the optical links operating in a given geographical area. The interference problem then becomes one of geometrical relationships – that is, angles, fields of view, and site separations. In general, geometrical coordination is easily obtained, and so no great burden, technical or financial, is placed on any of the administrations involved.

2. Interference modes in optical communications systems

The modes of interference which may arise in a general situation involving space and terrestrial optical systems can be classified systematically by considering four types of communication links: these are: space-to-space, space-to-ground, ground-to-space and ground-to-ground. There is a total of twelve distinct interference modes which arise from the fact that each of the three types of receiving stations (space, terrestrial and earth station), can experience interference from transmitters used in the four types of links. Figure 1 illustrates the various communication links which may be involved in the general situation. The interference modes to be studied are summarized in Table I, and are arbitrarily numbered for convenience.

For each of the interference modes, there are two sources of interfering signals:

- *direct illumination* of the receiver by the main beam or side lobes of the interfering beam, and
- *scattered radiation.*

Thus, in reality, a total of twenty-four interference situations must be considered.

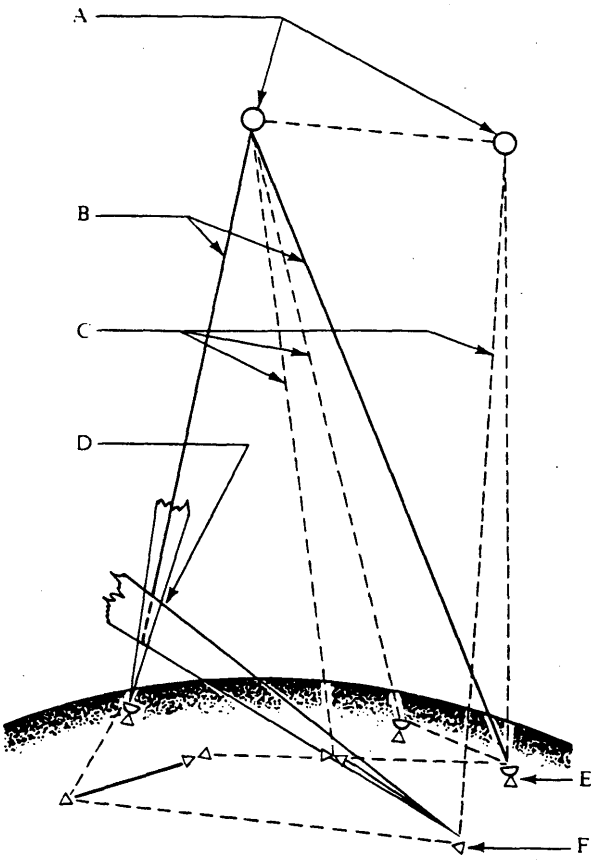


FIGURE 1 — *Interference modes with shared frequency operation of satellite and terrestrial systems*

A: Space stations
B: Wanted signals
C: Unwanted signals
D: Scatter from common volume
E: Earth station
F: Terrestrial station

TABLE I – *The possible modes of interference*

For each mode, interference from direct and scattered illumination must be taken into account. Thus, 24 interference situations must be considered.

Transmitter type / Receiver	Space to space	Earth to space	Space to Earth	Terrestrial to terrestrial
Space	1	2	3	4
Earth	5	6	7	8
Terrestrial	9	10	11	12

3. Direct interference situations – General

One method for protecting an optical communication link is to ensure that the signal-to-interference (S/I) ratio in the receiver's field of view is above a specified tolerance level. Hult and Reinhart [1971] have shown that unless the interfering signal is similar to the wanted signal (i.e., a similar modulation scheme, a similar or smaller bandwidth expansion ratio, and a carrier frequency close to the carrier frequency of the wanted signal), the characteristics of the wanted signal permit the carrier-to-noise ratio to be increased by factors of 10 to 50 dB. They have also shown that, for two television signals on the same carrier frequency, the interference effects are barely perceptible if the power in the wanted signal is 30 dB or greater than the power of the interfering signal. Choosing this 30 dB figure as a conservative tolerance level, the receiver's antenna gain will be less than -30 dB if [Hult and Reinhart, 1971]:

$$\frac{D}{\lambda} \times \frac{\theta}{100} \geqslant 3 \tag{1}$$

where:

D: is the aperture size, in centimetres (cm);

λ: is the wavelength, in centimetres (cm);

θ: is the half-angle from the main beam axis, in degrees.

For a small aperture size ($D = 30$ cm):

$\theta \geqslant 5 \times 10^{-4}$ deg for $\lambda = 0.5$ μ (visible);

$\theta \geqslant 0.01$ deg for $\lambda = 10.6$ μ (CO_2).

For a satellite at synchronous altitude (35 000 km), these half-beam-widths translate into circular fields of view on the ground of above 300 m in radius for a laser operating in the visible spectrum, and roughly 7 km in radius for a CO_2 laser. Thus, if earth stations are placed more than 7 km apart, even if adjacent earth-space links operate on the same wavelength with precisely the same modulation technique and polarization, direct interference between links should not be a problem. Actually, since we have completely neglected the transmitter antenna gain patterns which would be equally narrow at the distances involved, the protection ratio would be 60 dB or better, depending on the wavelength. For this reason, the distance between ground stations could be reduced even further.

4. Scatter interference situations – General

While the treatment of direct interference modes is relatively straight-forward, even with the meteorological uncertainties of turbulence-induced beam spreading, reliable, quantitative estimates for scatter situations are more difficult to establish. Each of the scatter interference situations is discussed in some detail below.

4.1 *Interference to a space receiver from a space-to-space transmitter (Mode # 1)*

For fixed position links involving geosynchronous satellites, the geometric coordination restrictions discussed above will resolve direct interference problems.

As regards scattering, atmospheric scattering is irrelevant in this case. Only energy which is scattered from another satellite which lies within the field of view of the receiver is of concern. In this case, the same method used to resolve direct interference problems should resolve this problem. Specifically, taking cognizance of the 30 dB S/I protection ratio, the space-borne antenna pattern need only be smaller than the minimum angular separation of adjacent synchronous satellites.

If the satellites involved are not fixed relative to one another – that is, for satellites not in geo-synchronous orbit – then, although transient interference is possible, it should be of short duration, and as such, would probably be tolerable.

4.2 *Interference to a space receiver from an Earth-to-space transmitter (Mode # 2)*

If the angle between satellites is greater than 0.01 degree, direct interference should not be a problem for any optical wavelength.

For problems involving scattering phenomena, interference can be reduced by placing restrictions on transmitter power p_T and on distance Z_T from the transmitter to the intersection region. The pertinent relationship between the parameters is given as follows [Kershenstein and Horrigan, 1977]:

$$\frac{p_T\,(\text{watts})}{Z_T\,(\text{km})} < \frac{3780}{\lambda\,(\text{microns})} \tag{2}$$

Assuming a minimum separation between earth stations, we must have $Z_T > 7$ km, and hence:

$$p_T \text{ (watts)} < \frac{26\,500}{\lambda \text{ (microns)}} \tag{3}$$

Evidently, even with a CO_2 laser system (i.e., $\lambda = 10.6\mu$), several kilowatts of transmitter power are required before interference problems are encountered.

4.3 *Interference to a space receiver from a space-to-Earth transmitter (Mode # 3)*

Geometric considerations suggest that direct interference is unlikely to occur.

With regard to scattering problems, one case of interest occurs when the field of view of the space receiver coincides with the interfering transmitted beam spot on earth. For energy which is scattered from earth in a Labertian distribution, no interference is expected to occur if the power of the space-borne transmitter is less than, or of the order of, 400 watts.

4.4 *Interference to a space receiver from a terrestrial transmitter (Mode # 4)*

This interference situation is essentially the same as that of Mode # 2. Even with beam crossovers, no direct or scattered interference problems are anticipated.

4.5 *Interference to an earth station receiver from a space-to-space transmitter (Mode # 5)*

This interference case is identical to the corresponding case of a space receiver and a space-to-space transmitter (Mode # 1) and offers no problems.

4.6 *Interference to an earth station receiver from a ground-to-space transmitter (Mode # 6)*

Direct interference is not a significant problem.

As regards scatter interference, this case is almost identical to that to be discussed under Mode # 7, with the important distinction that the scatter is primarily backscatter, rather than forward scatter. Thus, the multiple interference problem should be reduced by the ratio of back-to-forward-scattering cross sections. This ratio can be expected to vary anywhere from 5 to 1000, or so; hence, resolving scatter in Mode # 7, which is discussed below, automatically resolves scatter in the case of Mode # 6.

4.7 *Interference to an earth station receiver from a space-to-ground transmitter (Mode # 7)*

As noted earlier, for a satellite at synchronous altitude (35 000 km), typical antenna patterns translate into circular fields of view on the ground of about 300 metres in radius for the visible ($\lambda \approx 0.5$ μm) and about 7 km in radius for the CO_2 laser wavelength ($\lambda \approx 10.6$ μm). Thus, direct interference between earth stations separated by more than 7 km should not be a problem.

There remains the possibility that a neighboring ground-to-space link, with its ground station situated 7 km or more from the receiver of interest so as to avoid direct interferences, can deliver unwanted power to an earth station receiver via multiple scatter effects.

Many of the space receiver considerations for scattered interference situations discussed above apply, with the important exception that the earth station receiver is located quite near the scattering regions, and hence, it does not benefit as much from the inverse distance squared ($1/R^2$) free-space propagation losses as does the space receiver. If the earth receiver is using heterodyne detection, scatter is irrelevant [Kershenstein and Horrigan, 1977; Giannario *et al.*, 1976; Mooradian *et al.*, 1976]. This is so because the scattered beam travels a longer path, looses coherency with the unscattered beam, and is not processed by the coherent detection, heterodyne receiver. Direct detection systems, however, may experience interference problems.

If the interfering beam crosses through the field of view of the earth station receiver somewhere within the Earth's atmosphere, then there exists a possibility of scatter interference. However, site coordination can ensure that no beam intersections occur for links to geosynchronous satellites, and interruptions due to other satellite links will necessarily be of a temporary nature. Thus, this type of interference can be minimized, if not eliminated, in most cases.

If a beam of light is propagating vertically downward through the atmosphere, multiple scatter will cause it to spread beyond that predicted by geometric considerations. There is a predicted maximum in the spread as the optical thickness of the scattering medium varies; in the worst case, multiple scatter spreads the beam a radial distance approximate to the physical thickness of the scattering layer [Bucher, 1973]. Evidently if the separation between adjacent sites was increased by this amount, in addition to the distance required by the 30 dB direct interference criteria, neither link would experience interference due to any direct or scattered energy. In the

unlikely situation of a large cloud or fog which completely fills the space from the ground to an altitude of 12 km, with a worst-case total optical thickness of 2 or 3, [Kershenstein and Horrigan, 1977; Bucher, 1973] then an additional separation distance of 12 km must be added to the previously-determined upper limit (for direct illumination) of 7 km in order to obtain the minimum, safe separation distance. For most conditions, scattering will be confined to cloud layers or to regions which are close to the ground. In these cases, the thickness of the scattering layers would be of the order of 2 km or less, and so, a 19 km separation distance would be very effective.

4.8 *Interference to an earth station receiver from a terrestrial transmitter (Mode ≠ 8)*

Direct interference is readily avoided by geometric coordination of sites.

If the terrestrial link employs scatter propagation, scattered energy could be received by the earth station. Keeping the main beam of the scatter link out of the field of view of the earth-space link will resolve most problems. Then too, the high bandwidth of the space-to-ground link, and the necessarily low bandwidth of the scatter link, make serious mutual interference doubtful as the information processing gains, which have been neglected to this point, could easily produce 20 to 50 dB of additional protection.

4.9 *Interference to a terrestrial receiver from a space-to-space transmitter (Mode ≠ 9)*

There is little possibility of direct or scatter interference if grazing angles are excluded.

4.10 *Interference to a terrestrial receiver from an Earth-to-space transmitter (Mode ≠ 10)*

If the terrestrial link is a direct link (that is, line-of-sight), considerations of interference from a second transmitter of a ground-to-space link are the same as that for the case of the earth station receiver and a direct terrestrial link. In this case, the beams are narrow, and the geometric coordination required to prevent their intersection is easily implemented. If the terrestrial link is a scatter link, the analysis of Mode # 8 applies with the roles of transmitter and receiver reversed.

4.11 *Interference to a terrestrial receiver from a space-to-Earth transmitter (Mode ≠ 11)*

Geometric coordination to eliminate interference problems should be implemented.

Scatter interference is of primary concern in this case. This is especially true for terrestrial links employing scatter propagation since the scatter link necessarily operates with a large field of view. Fortunately, large processing gains (of the order of 50 dB), together with the use of continous emission on the space-to-ground link, can reduce the possibility of interference to the terrestrial receiver [Kershenstein and Horrigan, 1977]. However, if the space-to-ground link is a pulsed system which employs unusually high power levels (on the order of megawatts), interference problems cannot be ruled out by the simplistic arguments employed above, and additional efforts to resolve the interference problems are required.

4.12 *Interference to a terrestrial receiver from a terrestrial transmitter (Mode ≠ 12)*

It is possible to eliminate direct or scatter interference to coherent or heterodyne receivers, for they are insensitive to everything except the unscattered portions of their intended transmitter beams [Kershenstein and Horrigan, 1977; Giannario *et al.*, 1976; Mooradian *et al.*, 1976]. Direct detection receivers, however, will be influenced by direct and scattered radiation. Thus, geometric considerations, together with scatter analyses, may be required to resolve interference problems on a case-by-case basis.

5. Summary

Most interference problems involving optical links are easily resolved through geometric coordination of all links in a given geographic area. In fact, extremely conservative analyses indicate that 21 of the 24 possible interference situations involving direct and scatter interference can be controlled, if not eliminated, by the application of relatively simple geometric coordination procedures. No limitations are placed on the frequency bands employed.

The three unresolved interference situations involve scatter-interference from space-to-Earth, Earth-to-space and terrestrial transmitters into the receiver of a terrestrial, scatter-propagation link. Even in these cases, geometric coordination may resolve conflicts, but such coordination may be more restrictive in nature because it is more difficult to implement than for other modes. Scatter links necessarily use antennas with larger fields of view than

do line-of-sight systems, and so, they render large volumes of space unusable by other optical links. When beam intersections do occur, serious interference problems can be expected. This is particularly true for the case of two terrestrial scatter-propagation links which do not benefit from the large processing gains which characterize the other two cases. However, the seriousness of these potential interference problems can only be assessed on a case-by-case basis.

In summary, analyses of direct and scatter interference situations for optical communication systems suggest that most interference problems can be resolved by geometric coordination within the region of operation.

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REPORT 663-2*

CHARACTERISTICS OF LARGE MILLIMETRE AND SUBMILLIMETRE WAVE REFLECTING ANTENNAS

(Question 64/1)

(1978-1982-1986)

1. Introduction

This Report will examine characteristics, achievable main beam radiation patterns, and frequency performance with the far and back side lobes of large reflector-type millimetre and submillimetre wave antennas. An antenna is generally considered large if it has a diameter to wavelength ratio greater than 100. A discussion of the effects of surface tolerances of these antennas and of their susceptibility to interference from other transmitters is presented.

2. Gain and side-lobe characteristics

The requirements that a given millimetre wave antenna collect energy from a preferred direction, while suppressing terrestrial or galactic noise and signals from interfering transmitters received from off-axis directions, place constraints on its entire radiation pattern. Thus, a full 2π-steradian radiation pattern must be characterized at millimetre frequencies to enable predictions of interference levels between systems.

The radiation pattern of an antenna can be broken into roughly four regions, which outline the mechanisms responsible for the pattern (Fig. 1) [Ratnasiri *et al.*, 1970]. The main lobe and near side lobes, Region I, are the diffraction pattern caused by the feed illuminating the reflector. The far side lobes, Region II, are the sum of feed energy not intercepted by the reflector (spillover) and the interference pattern created by the illumination of the reflector edge. The back lobes, Region III, and the rear lobe, Region IV, are due to the reflector edge interference pattern alone since the reflector obscures the feed in this area. Specialized mathematical techniques have been developed for each of these regions; the resulting formulae can be used to make predictions concerning millimetre wave antennas.

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* This Report should be brought to the attention of Study Groups 2, 4 and 9.

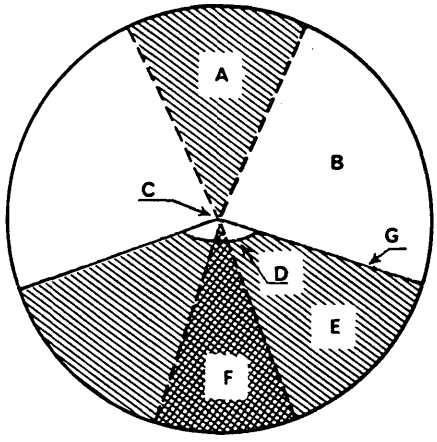


FIGURE 1 — *Regions used for analysis of radiation pattern*

Two features of millimetre wave antennas distinguish them from lower frequency antennas:

- physically small reflecting surfaces can be very large with respect to the wavelength in question;
- physically realizable surface tolerances can be a significant fraction of a wavelength.

The large amount of aperture available makes possible, in principle, the design of antennas with very sophisticated patterns. The relatively severe surface tolerance limitations create phase errors in the reflected radiation, however, and can scatter significant amounts of energy out of the intended or designed radiation pattern.

2.1 *Main beam and near side lobes*

Silver [1965] has given a formula that describes the far field radiation pattern created by illuminating a perfect reflecting surface with electromagnetic energy. The expression involves mathematical approximations to the field coordinates which make it valid near the forward axis of the antenna only. It does not include effects of feed blockage or tolerance errors on the surface of the reflector.

Gravitational loading, thermal stresses, aging, and the limitations of fabrication processes result in deviations from the desired reflecting surface. These surface tolerance limitations result in errors in the desired phase distribution across the aperture. The effects of perturbations of the aperture phase distribution function have been examined [Ruze, 1966; Scheffler, 1962]. Ruze [1966] modified the equation of Silver [1965] by introducing a phase perturbation function. Scheffler [1962] derived a closed form expression for the scatter pattern which is valid for uniform illumination.

The equations derived by Ruze and Scheffler are valid for uniform illumination and describe the following characteristics of reflector antennas as they are used at higher frequencies:

- the radiation pattern taking into account the reflector tolerance errors is the sum of the pattern in the absence of tolerance limitations, G_p, which has been degraded by an exponential factor $e^{-\overline{\delta^2}}$ and a scatter pattern, G_r,

where:

$\overline{\delta^2} = \left(\frac{4\pi\varepsilon}{\lambda}\right)^2$: mean square value of phase perturbations caused by surface tolerances,

ε: r.m.s. tolerance error across aperture,

λ: wavelength of radiation;

- as the frequency increases the main axis gain of a given antenna increases to some maximum value and then decreases. The initial increase is due to the growing number of wavelengths across the aperture; the decrease is due to the exponential factor that predominates at higher frequencies. This limiting behaviour in attainable gain is illustrated for several large antennas (Fig. 2) [Ruze, 1966];
- the energy lost from the principal pattern appears in the scatter pattern. The magnitude of the scatter pattern increases with frequency for a given reflector and illumination function until the magnitude dominates the principal pattern.

Thus, as a given reflector is used at increasingly higher frequencies, the energy received from or transmitted to a given target will suffer a loss described by $e^{-\overline{\delta^2}}$, and the received noise level will be increased due to the larger scatter pattern. The increase of noise will be a function of the environment in the pointing direction of the antenna.

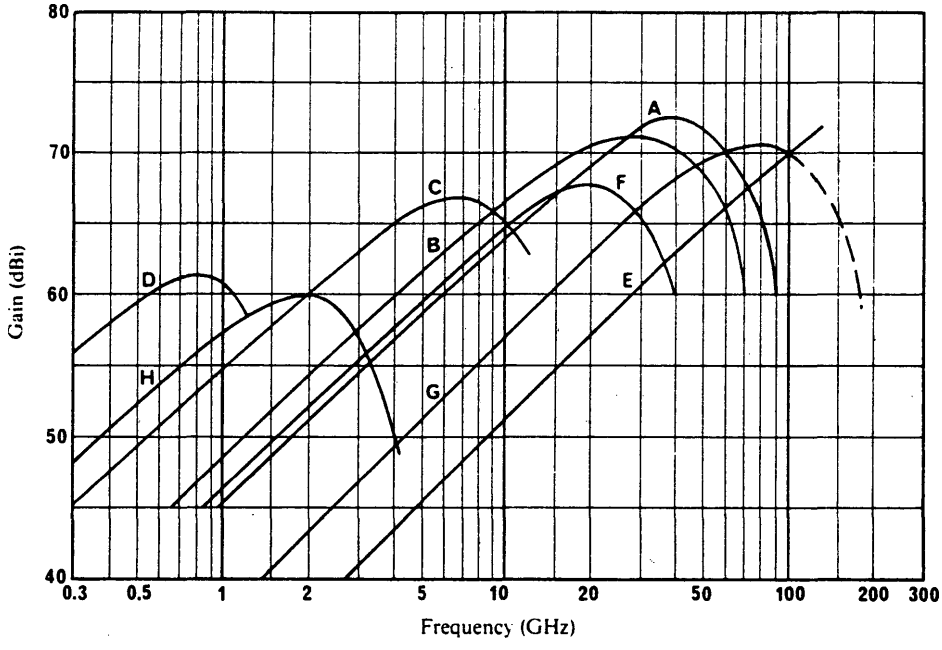


FIGURE 2 — *Gain of large paraboloids (based on published estimates)*

A: Lebedev – 72 ft (22 m)
B: Haystack – 120 ft (36.5 m)
C: Parkes – 210 ft (64 m)
D: Arecibo – 1000 ft (305 m)
E: Aerospace – 15 ft (4.6 m)
U. Texas – 16 ft (4.9 m)
F: Good – 85 ft (26 m)
G: Spun cast – 28 ft (8.5 m)
H: Greenbank – 300 ft (91.5 m)

The effects of surface errors on the main axis gain of the primary and scatter radiation patterns are plotted for a 10 metre reflector with assumed efficiency of 0.8 (Fig. 3). ε = 0.5 mm is a worst case, 0.1 mm is typical, and 0.05 mm is attainable [Cogdell *et al.*, 1970]. ε = 0.01 mm is included in Fig. 3 to illustrate the advantages available from improved fabrication techniques. Figure 3 indicates that greater precision allows operation at higher frequencies and shows that for a given precision the principal pattern is dwarfed by the scatter pattern at sufficiently high frequencies.

The on-axis scatter pattern gain reaches a limiting value at large values of $\overline{\delta^2}$. This behaviour is seen in the curve for ε = 0.5 mm in Fig. 3, noting that $\overline{\delta^2}$ increases with frequency for fixed ε.

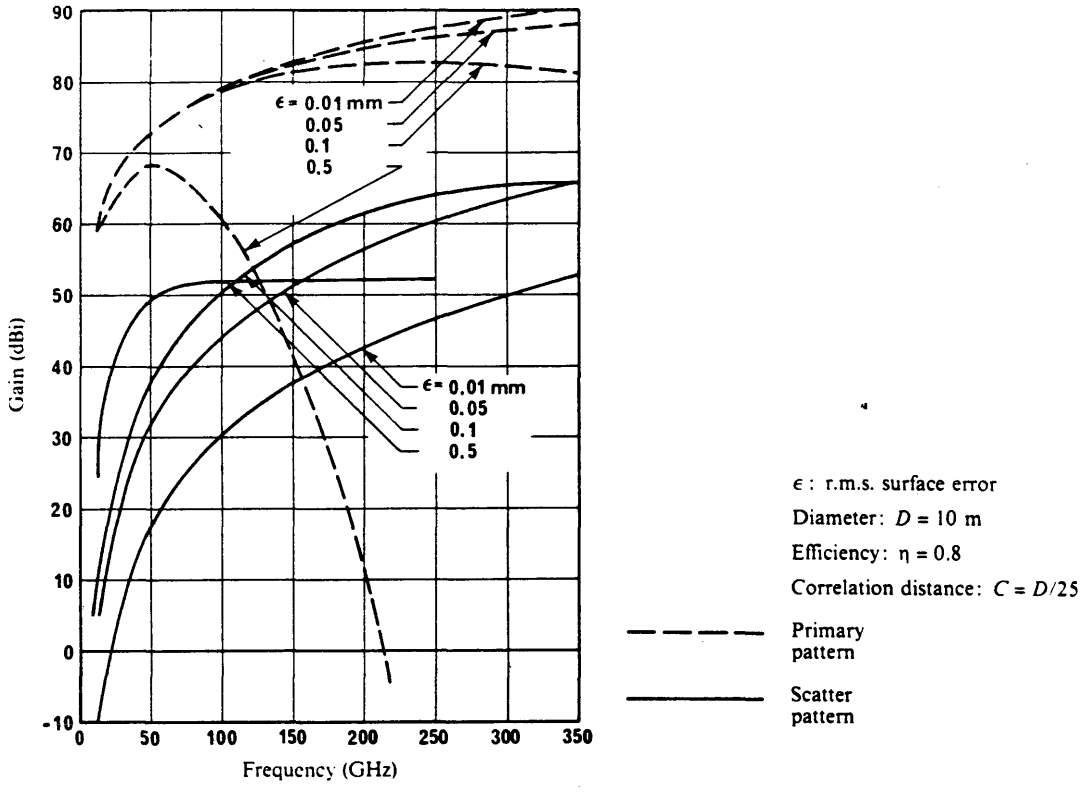


FIGURE 3 — *Main axis gain for a 10-meter reflector (primary and scatter patterns) as a function of mechanical tolerance*

A computerized analysis of the gain of large reflecting antennas using infinite series expressions for the scatter pattern has yielded on-axis behaviour with increasing frequency similar to that presented here [Zucker, 1968].

The angular spread of the scatter pattern is of great importance for the prediction of interference between equipments. Therefore the 3 dB and 10 dB points of the scatter pattern are calculated for $\varepsilon = 0.1$ mm as a function of frequency for a family of reflector diameters (Fig. 4). The scatter pattern narrows with increasing frequency until it reaches some limiting value. Table I presents the limiting values reached for a cross section of tolerance values. Note that the scatter patterns are very narrow, even for the worst case of 0.5 mm tolerance and a 1 m diameter antenna.

2.2 *Far side lobes*

The far side lobe radiation pattern (Region II in Fig. 1) for a focal-fed reflecting antenna is the power sum of electromagnetic energy radiated by three mechanisms:

- energy radiated directly from the feed,
- reflected energy which is scattered by the feed and its supports,
- energy radiated by diffraction at the reflector edge.

Tolerance limitations of the reflecting surface do not have any direct effect on the far side lobe patterns.

Ratnasiri *et al.* [1970] have derived expressions for the diffracted field using the geometrical theory of diffraction. Using their results, it is possible to write an expression for the radiation pattern in the far side lobe area, valid everywhere in Region II except near its boundaries [Katzenstein, 1977]. This expression ignores energy scattered by the feed.

This analysis shows that the far side lobes of a focal-fed antenna will diminish as a given reflector is used at higher frequencies if the illumination $f(\bar{r})$ is assumed constant with frequency.

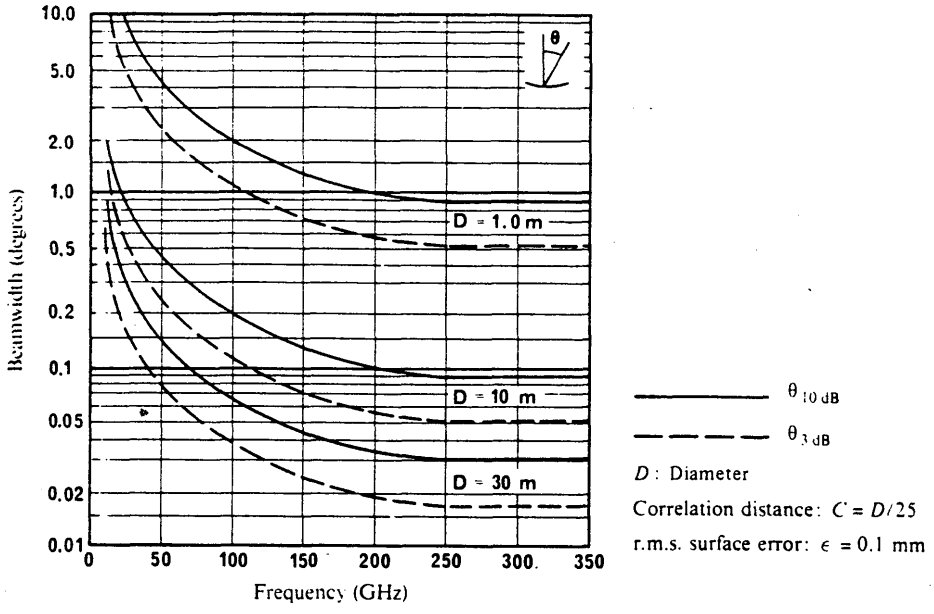
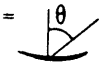


FIGURE 4 — *Scatter pattern beamwidth for* $\epsilon = 0.1$ *mm*

TABLE I – *Limiting beamwidth scatter pattern*

ε (mm)	Diameter (m)					
	1.0		10		30	
	Beamwidth (deg) (1)					
	$\theta_{3\ dB}$	$\theta_{10\ dB}$	$\theta_{3\ dB}$	$\theta_{10\ dB}$	$\theta_{3\ dB}$	$\theta_{10\ dB}$
0.01	0.048	0.087	0.0048	0.0087	0.0016	0.0029
0.03	0.143	0.261	0.014	0.026	0.0048	0.0087
0.05	0.239	0.435	0.024	0.043	0.0080	0.0145
0.10	0.477	0.869	0.048	0.087	0.0159	0.0290
0.20	0.954	1.74	0.095	0.174	0.0318	0.0580
0.30	1.43	2.61	0.143	0.261	0.0477	0.0869
0.40	1.91	3.48	0.191	0.348	0.0636	0.116
0.50	2.39	4.35	0.239	0.435	0.0795	0.145

(1) Beamwidth = 

2.3 *Shadow boundary*

An important far side lobe occurs near the boundary separating Regions II and III, due to the in-phase addition of direct feed radiation with energy diffracted by the reflector edge. Ratnasiri *et al.* [1970] have derived expressions for the radiated fields at the boundary for a focal-fed antenna.

The power near the shadow boundary is the sum of direct feed energy, diffracted energy, and a fine structure resulting from the interaction of these two. It can be shown [Katzenstein, 1977] that the peak in the far side lobe pattern at the shadow boundary will tend to remain constant as a given focal-fed reflector with a fixed illumination function is used at higher frequencies.

2.4 *Back lobes*

The radiation pattern in Region III of Fig. 1 results from edge diffraction alone since the feed is blocked by the reflector.

Using the geometrical theory of diffraction, it can be shown [Katzenstein, 1977] that the backlobes are modulated by an envelope proportional to $1/k$ at millimetre and submillimetre frequencies. Thus, the rear sidelobes will become smaller with increasing frequency, if all other parameters are assumed constant.

2.5 *Rear axis pattern*

The geometrical theory of diffraction cannot be applied on the rear axis since it is a caustic for the diffracted rays [Ratnasiri *et al.*, 1970]. Ratnasari *et al.* [1970] have derived an expression for the pattern maximum which occurs on the rear axis using the ring current method. This expression is frequency independent. Thus, a given reflector will tend to demonstrate no change in the magnitude of the rear pattern maximum as it is used at higher frequencies with fixed illumination.

3. Cassegrain antennas

Many antennas used at microwave and millimetre frequencies are of the Cassegrain geometry, as illustrated in Fig. 5. The far and back side lobe patterns of the Cassegrain antenna will differ somewhat from those of the prime focus-fed antenna discussed above. The mechanisms responsible for these patterns are:

- energy from the feed reflected by the subreflector onto the main reflector creates the desired or principal pattern;
- blockage by the subreflector and its supports degrades this pattern somewhat;
- feed energy not intercepted by either reflector (spillover) is radiated into space;
- a diffraction pattern is created by the edges of the subreflector;
- a diffraction pattern is created by the edges of the main reflector, which is illuminated by the subreflector or its diffraction pattern.

An analysis of the Cassegrain antenna would differ in several respects from that of the prime focus-fed antenna:

- the spillover radiation from the feed becomes inverted with angle since the feed is pointed along the main reflector axis instead of against it. Thus, the main feed sidelobes will be closer to the antenna main forward axis;
- the diffraction energy due to the subreflector being illuminated by the feed will be introduced;
- the main reflector edges will now be illuminated by energy either diffracted or reflected from the subreflector, instead of by energy directly from the feed. Efficient use of the aperture requires that most or all of the aperture be illuminated. Since the fields are continuous at the boundary between diffracted and reflected energy [Kouyoumjian and Pathak, 1974], the main reflector edge will generally receive illumination comparable to that of a prime focus-fed antenna for either reflection or diffraction illumination.

Thus, the primary differences between prime focus and Cassegrain-fed antennas will be the inversion of the feed spillover pattern and the addition of diffraction energy from the subreflector.

The far side lobes, rear side lobes, and back lobe will show the same general improvement with increasing frequency for fixed illumination function for a Cassegrain-fed antenna as for a prime focus-fed antenna. The principal pattern degradation with increasing frequency, due to surface tolerance limitations, will be the same for either feed type.

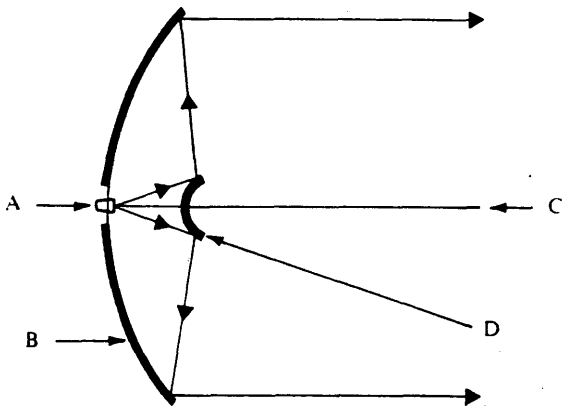


FIGURE 5 — *Cassegrain antenna geometry*

A: Feed
B: Main reflector
C: Main axis
D: Subreflector

4. Reference antenna pattern

It can be concluded that the reference radiation pattern for large terrestrial parabolic antennas operating between 2 and 10 GHz presented in Report 391 and Recommendation 465 is useful for calculations of interference between services at millimetre wavelengths, since the far and back side lobe levels of large millimetre antennas will generally be equal to or lower than those specified in the reference pattern. This reference pattern can be replaced with measured data when this data becomes available for millimetre wavelengths and should be regarded as provisional for these frequencies.

5. Validity of analysis at high frequencies

The tolerance analysis is valid to optical frequencies, since Scheffler's analysis [Scheffler, 1962] of the scattered pattern was done for optical devices. The diffraction analysis does not include irregularities in the diffracting edge or edge thickness which may become more important at higher frequencies.

6. Main beam cross polarization characteristics

In designing antennas with high polarization purity and/or estimating the main beam cross polarization characteristics of existing antennas, it is necessary to take into account not only the standard deviation of the reflector surface errors but also the correlation distance of the errors and the feed horn characteristics.

For instance, the main beam cross polarization characteristics of antennas can be improved by using corrugated conical horns (balanced feed type) which have better performance than conventional conical horns (unbalanced feed type). In the case of the corrugated horn, better cross polarization isolation is attained for antennas with a larger correlation distance under given r.m.s. surface errors. On the other hand, for a conventional conical horn, better cross polarization isolation is attained for a smaller correlation distance under given r.m.s. surface errors [Mizuguchi, 1982].

7. Summary and conclusions

The degradation of the principal pattern due to tolerance limitations of the reflecting surface is the main obstacle to utilization of large reflectors above 100 GHz. As a given reflector is used at higher frequencies, assuming fixed illumination, its gain reaches a limiting value and then diminishes. This lost energy appears in a narrow scatter pattern, which becomes larger than the principal pattern at sufficiently high frequency.

The appearance of the scatter pattern due to tolerance limitations of the reflecting surface will result in increased probability of receiving undesired emissions from threat emitters. This pattern is narrow, however, as shown in Table I and good system design can preclude interception of unwanted emissions.

Far side lobe levels will generally decrease as parabolic reflectors are used at higher frequencies. Tolerance limitations in the reflecting surface do not impact on far or back lobe levels. Thus, immunity to energy received through the far side lobes will increase.

The side lobe near the shadow boundary caused by in-phase addition of feed spillover and diffracted energy will tend to remain constant at higher millimeter wave frequencies.

Back lobe levels will generally decrease with increasing frequency, causing greater interference immunity.

Rear axis lobes will tend to remain constant as parabolic antennas are used at higher frequencies, resulting in no increased susceptibility to undesired emissions from this direction.

The reference radiation pattern presented in Recommendation 465 can be used for calculations of interference at millimetre wavelengths.

It should be noted that the cross polarization characteristics depend largely on the polarization characteristics of the feed horn, the standard deviation of reflector surface errors and the correlation distance of surface errors.

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REPORT 665-2

NEW TECHNOLOGIES FOR OFF-SET REFLECTOR MULTIPLE-BEAM ANTENNAS

(Questions 18/1, 64/1)

(1978-1982-1986)

1. Introduction

Question 18/1 invites studies on how factors and "trade-offs" including antenna design and antenna beam pattern structure would maximize the efficiency and utility of spectrum use. Report 558 discusses the need for investigation of multiple beam antennas (MBA). Question 64/1 invites studies on what practical kind of systems are feasible for communications above 40 GHz and what technical problems are encountered to enable the realization of these systems and how these systems can contribute to alleviating the present congestion in the use of radio waves. This Report discusses the state-of-the-art on multiple beam antenna systems, how multiple beam antenna design can increase the efficiency of spectrum use and trade-offs for applications below 40 GHz. Some comments are made concerning systems operating above 40 GHz.

In early satellites, a global beam was used for communication between different countries. In the Intelsat-IV-A series satellite, "east" and "west" hemispheric beams are being used. Newer generation satellites make use of "spot" beams for communication between different countries re-using the same frequency bands. The most representative MBAs in use presently are of the offset-fed reflector type similar to those employed by Intelsat satellites. Note that a true MBA for spatial-frequency re-use produces many separate beams from one common aperture.

The present (and also the next planned) generation of Intelsat satellite MBAs achieve at least 27 dB isolation between beams. Beams radiating orthogonally polarized energy provide satisfactory frequency re-use (i.e., different users operating at the same frequency within the same beams) provided the polarization purity is adequate and not significantly reduced by rain or other particulate scattering. To achieve satisfactory beam isolation, the unwanted power has to be reduced by the design of low side-lobe antennas. The general requirements of isolation between beams, as well as the total coverage provided by the beams, invoke certain restrictions in the design of a multi-beam antenna. These generally fall into the following areas [Mathews *et al.*, 1976]:

(a) Antenna beamwidths and beam separation angles must be chosen to provide the required isolation as well as coverage.

(b) Characteristics of the individual beams, including both beam shape and side-lobe levels, must be properly selected. Side-lobe level discrimination of the order of 35 dB below the main beam is currently required. The steeper the side-lobe decay, the smaller the angular spacing between individual beams that can be achieved.

(c) Use of circular polarization diversity requires that axial ratios of the individual beams be maintained under 0.5 dB to prevent excessive leakage into the cross-polarized channel.

(d) Feed networks that provide excitations for individual beams to achieve the desired shapes and low side lobes, as well as necessary switching, should be as simple as possible to avoid excessive losses.

Limits on interbeam isolation may be established by any of the following coupling criteria:

- main-beam coupling,
- side-lobe levels, or
- cross-polarized levels.

These levels must ordinarily be below the desired isolation level between beams, over the entire coverage area of each beam, with sufficient margin to allow for all contributions to the total isolation from other beams. The minimum beam separation to meet typical isolation requirements is usually established by the null-to-null beamwidth of the down-link (broader) beam.

2. Fundamental performance

The use of an MBA instead of a phased-array, or other antenna, requires analysis in order to focus on the fundamental characteristics that are unique to specific and general applications. For example, communication systems which postulate a single hopping beam (e.g. satellite switched time division multiple access, TDMA) can be served equally well by an MBA or a phased-array. The antenna complexity, etc., is very nearly the same for both devices; designer preference is usually the predominating factor determining the ultimate choice. However, if more than one beam must be formed simultaneously, the MBA at present appears to be the best antenna system. The preformed beams provide an extremely attractive radiation characteristic as prior to launch, the antenna radiation characteristics can be verified. After launch, such antennas can be expected to maintain the same radiation characteristics.

It is important first to point out the special features which define an MBA. An MBA consists of a radiating structure with more than one input port and an equal number of corresponding beams each pointing in a different direction; the total beam arrangement defines a composite coverage area usually called the antenna field of view (FOV). The MBA can be in the form of a lens illuminated by an array of feed horns, each horn generating a corresponding beam. A similar feed-horn array could illuminate a paraboloid reflector or the sub-reflector of a Cassegrain antenna. In fact, virtually any single, dual, or multiple lens, or reflector system illuminated by an array of feeds represents a *bona fide* MBA with all the concomitant characteristics.

An array of similar radiating elements of arbitrary size forming a "filled", as opposed to a "thinned", aperture can be excited by a Butler beam-forming network (BBFN) and result in an MBA. In this case, the lens function is carried out by the unique arrangement of transmission lines in the BBFN.

Currently one of the worst constraining problems in the realization of practically useful multi-beam, multi-re-use antenna systems, especially those using a single reflector, is the containment of interference in any one beam due to the cumulative effects of the side lobes from other beams. An array of X beams, each isolated against the other by an average Y dB, will allow no better than about $Y - 10 \log X$ dB isolation for any one beam. Administrations are invited to provide additional technical information on this problem.

In summary, the MBA can be a reflector, lens, or array-type antenna. It will probably perform as good as or better than a comparable phased array. The desired side lobe, etc., performance can usually be achieved through careful design.

3. Earth-station antennas

Multible beam antennas can be used in either a satellite or earth-station ground terminal; however, the special requirements of earth stations do not necessarily warrant the use of an MBA. Specifically, earth-station antennas must be designed to prevent interference to neighbouring satellites. This requires a directional discrimination property and the maintenance of adequate link margin and usually dictates an antenna gain greater than 35 dB.

3.1 *Paraboloid reflectors*

Economic considerations usually indicate a preference for a reflector antenna, and a paraboloid type is most commonly used in single beam applications where only a very few beamwidths of scanning are needed. The paraboloid reflector is usually illuminated by a single feed horn producing a single beam. The beam is pointed toward the satellite by mechanically steering the reflector-feed assembly. A paraboloid can also be illuminated by a feed-horn array and as such is a *bona fide* MBA. However the paraboloid is a single point focus system, and those feeds located off the focal point produce radiated beams whose polarization purity and associated side lobes are degraded. The degree of degradation depends on the focal length-to-diameter ratio (F/D) of the paraboloid and the angle between the focal axis and the beam axis. Generally, a reflector with an $F/D \approx 0.5$ will suffer a 0.5 dB loss in gain and $\approx$ 5 dB increase in side-lobe level when the beam is pointed or scanned about five half-power beamwidths from the boresight direction. Increasing F/D will decrease the degradation in performance as the beam is scanned; however, the paraboloid reflector does not have wide-angle scanning properties as good as those of toroidal or spherical reflectors. They are also, in general, inferior (i.e., with regard to scanning) to lens type antennas.

A paraboloid with as much as 60 dB gain can be designed to have a 1° diameter FOV ($\pm$ 0.5° scan) by the use of MBA techniques. This is usually adequate for earth terminals with fixed reflector antenna that can track or scan their beams to accommodate satellite motion, atmospheric-induced diffraction, and thermal distortion effects.

3.2 *Toroids*

Whenever a larger field of view is required, as in those cases where the earth terminal is either on a movable platform or the earth terminal is required to communicate with more than one satellite, toroidal reflectors can often provide the desired performance. The rectangular torus antenna uses a reflector surface generated by rotating a parabola or ellipse about an axis perpendicular to the focal axis of the parabola and in the plane containing the parabola. The axis of rotation is usually located about $2F$ from the vertex of the paraboloid, where F is the focal length of the paraboloid. The feed is usually located slightly ahead of the focal point.

The torus is not a perfect focusing surface. With respect to the on-axis beam or a paraboloid, it possesses degraded characteristics, such as increased side lobes, decreased gain and lack of polarization purity. These effects can be compensated through careful, but complex feed design, or by the use of a Cassegrain aberration correcting sub-reflector. The increased cost and complexity of the feed design is often justified by the extremely large scan angles that can be realized. The property of scanning large angles occurs only in the plane perpendicular to the axis of rotation (i.e., used to generate the surface); and due to the inherent circular symmetry, the radiated beam shape, side lobes, etc., are invariant with scan, over angles up to about 120° using small equivalent apertures (less than 10 m).

A conical torus, used in a fixed antenna installation and fed by a multiple-feed horn array, was described by Hyde *et al.* [1974]. This antenna has a pointing accuracy of 0.05° and has a gain of ≈ 40 dB. Although higher gain is possible, a complex feed-horn array or a Cassegrain aberration correcting sub-reflector (Hyde, patent, and presentation at URSI XVIIIth General Assembly, Lima, Peru, 1977) would be required to compensate for spherical aberration introduced by the circular symmetry. Analysis has shown that good aperture efficiencies are obtained for D/λ less than 220. The antenna uses a thermoelectrically cooled parametric amplifier with a 75 K noise temperature. The G/T of the antenna is 30.1 dB(K^{-1}). This antenna has been satisfactorily tested with 612 voice channels [Grossman, 1974]. Because of the overlapping of illuminated regions on the conical torus, less antenna area is required for multiple beams. Theoretical comparison has been made [Hyde *et al.*, 1974] between a 9.6 m torus and a typical 9.6 m paraboloid. For only two beams, the torus antenna surface is significantly smaller than that of the paraboloid for fields of view approaching 20°. For three or more beams, the area of the torus antenna is smaller for field of view less than 30°.

Toruses are not limited in scan angle by electrical considerations. In principle, with an offset focus, an angle of 360° can be scanned. In practice, for a conical torus, scanning the surface of a very flat cone which can be used to approximate the geostationary arc, when seen from an earth station, the limits that are encountered are either structural or due to approximating the desired scan surface with a right circular cone (and limited deviations to the surface which governs scanning in the orthogonal direction). It would seem that 40° would be about the practical structural limit for equivalent apertures (10 m or larger), and significantly larger for the smaller equivalent apertures.

3.3 *Spherical reflectors*

A spherical reflector can be used to provide wide angle beam-scanning over a spherical surface. The feed must be designed to compensate for the spherical aberration introduced by the reflector. The scanning properties are similar to those of the toroidal antennas with the exception that the wide-angle scan can be achieved in two orthogonal planes containing the focal axis. Spherical reflectors have been used in radar and radioastronomy applications but they are not generally used in communications systems. The 305 m (1000 foot) diameter spherical reflector used at the radioastronomy site in Puerto Rico is an example of the versatility of the device.

An offset spherical reflector antenna has been developed with a combination of auxiliary and sub-reflectors which are useful for compensating not only for spherical aberration but also for cross-polarization degradation [Watanabe and Mizugutchi, 1983].

A scale-model antenna of 2 m × 3 m (D/λ = 200) was tested at 34 GHz. Experimental results show that the side-lobe level is significantly lower than that of Recommendation 465 and that the cross-polarization characteristics are almost equal to those of the high performance Cassegrain antenna. Excellent beam scanning characteristics have also been obtained e.g., the gain reduction is less than 0.4 dB at an off-axis angle of 8° (30 × half power beamwidth).

4. Space station antennas

Among various types of antennas, the following have been considered for satellite usage: spherical-reflectors, two reflectors, off-set and centre feed reflectors, phased arrays, horns, bootlace lenses and zoned lenses. The fundamental principles described in § 1 represent a general consideration of these devices as compared with the specific discussions which follow.

4.1 *Reflectors*

Reflector antennas are undoubtedly the most commonly used high-gain antennas because they are usually lighter, simpler to design and build, less costly, and probably yield the best performance. Given an adequate aperture, the maximum path gain achievable is limited by the accuracy of the reflecting surface and the precision with which the antenna can be pointed to the intended terminal.

Most reflector antennas are symmetrical dish-shaped paraboloids illuminated by a feed-horn located at the focal point. Arrays of feed-horns are used in multiple-beam antennas using reflectors and, unfortunately, give rise to aperture blockage and a concomitant degradation of radiation performance (i.e., reduced gain, increased side lobes, etc.). This can be ameliorated by offset-feeding the reflector (i.e., using an asymmetric section of the paraboloid as the reflector). This can eliminate blockage but could increase energy radiated in the orthogonal polarization. Cross-polarized energy reduces the antenna gain and decreases beam isolation; this latter phenomena often leads to reduced performance in a frequency re-use antenna.

Offset-fed paraboloid reflectors can be designed for less than −30 dB side lobes and wideband operation. In many ways, they compete favourably with virtually any other type of high-gain antenna, with the possible exception of the centre-fed MBA. Most satellite MBAs use the feed-horn array offset paraboloid with considerable success.

4.1.1 *Spherical reflectors*

The spherical reflector using an appropriately designed feed can perform nearly as well as a paraboloid. Moving a correcting feed on the surface of a concentric, but smaller (approximately one half of the radius) spherical surface, permits wide-angle scanning over a circular field of view with very little, if any, change in the antenna radiation characteristics. Without a correcting field, the spherical reflector introduces spherical aberration and the concomitant degradation in performance (i.e., lower gain, increased side lobes, increased cross-polarization radiation, etc.). In fact, the use of a simple feed and $F/D \approx 1$ limits the gain of a spherical reflector to ≈ 50 dB.

A laboratory model 1.5 m diameter spherical reflector has been fabricated with an r.m.s. surface accuracy of 0.05 mm and has been tested [Turrin, 1975]. The feed horns were designed to operate both at 20 and 30 GHz bands with linear orthogonal polarizations. By effectiveiy controlling the feed aperture sizes, optimum illumination of the spherical reflector was obtained at both the frequencies. The maximum beam pointing error was about one-half beamwidth. For maximum scan angle, the degradation was 1 dB at 19 GHz and extrapolated degradation was 2 dB at 30 GHz.

4.1.2 *Dual reflectors*

Several forms of dual reflector systems have been studied, analyzed, and evaluated experimentally and operationally. They fall into three main classes: Cassegrain, Gregorian, and other. Cassegrain systems use a hyperboloid reflector between the main reflector focus and its vertex. Classical Gregorian reflector systems use an ellipsoid located on the main reflector focal axis, outside of the area between the vertex and focal point of the main reflector. The feed horn is located between the focal point and the vertex of the main reflector and illuminates the Gregorian secondary, or sub-reflector. This minimizes the length of the feed waveguide.

The Cassegrain system has the primary advantage of placing the feed near the vertex of the main reflector which effectively increases the antenna system focal length, resulting in a reduction of the overall dimensions of the antenna. The aperture blockage introduced by the sub-reflector is negligible if the antenna gain is greater than 40 dB, and the feed-horn array (i.e., the aperture of the feed-horn array) is smaller than the diameter of the sub-reflector. The scanning corrected Cassegrain antenna is a dual reflector unit with partial correction, by the design of reflector profiles, to achieve beam scanning with a minimum loss in performance. Cassegrain systems can compete favourably with centre-fed paraboloid reflectors and in some instances have better radiation characteristics. However, they have the same limitations as paraboloid reflector antennas in most MBA applications.

Gregorian reflector systems offer little advantage over the Cassegrain configuration and, consequently, are not used to any great extent.

The third class of dual reflectors use surfaces of revolution synthesized to meet or maximize specific performance characteristics. For example, Galindo [1964] has shown that the main and sub-reflectors can be shaped to produce nearly uniform distribution over the radiating aperture resulting in increased gain and aperture efficiency and −30 dB side lobes. Using similar design procedures, one could optimize the reflector surfaces for improved side-lobe level, polarization purity, etc.

An offset bifocal antenna which consists of dual reflectors with specially designed curvatures has been developed [Mizugutchi and Watanabe, 1983]. It can radiate two independent main beams with no aberration. The measured results at 11/14 GHz for an experimental antenna of 1.3 m in diameter show that the antenna has good polarization characteristics as well as low loss properties over a wide range of beam scanning angles.

4.1.3 *Offset fed reflectors*

Offset feeding has been referred to in several of the previous sections. Fundamentally, it has the same properties as a centre-fed paraboloid except that the aperture blocking can be reduced or eliminated which enables multiple feeds to be used. The offset reflector is larger than its centre-fed counterpart; for example, the area of a full offset-fed reflector is approximately 50% greater than that of a centre-fed reflector with the same gain. Offset feeding can be used with any of the foregoing reflector systems and

will, in general, provide the same benefits (i.e., less aperture blockage). Offset reflectors can be fed by various means including a Butler matrix, phased array, lens, etc.

4.2 *Lenses*

Bootlace and zoned lenses have received attention for satellite applications. In these lenses, two surfaces give the designer more flexibility for optimizing the performance of the system.

4.2.1 *Bootlace lens*

In a Bootlace lens, elements on two surfaces are connected by transmission lines. As the phasing network is inherently broadband, this type of antenna has a bandwidth of about 20%. The shapes of the two surfaces are determined by the Abbe condition.

Development of a wideband Bootlace lens has been reported [Mathews *et al.*, 1976; Scott *et al.*, 1976] to provide simultaneous transmit-receive dual-polarized multiple beams from a single aperture for operation in the 4 and 6 GHz bands.

There were 19 elements in the feed. To produce low side lobe beams, a symmetrical cluster of 7 elements at 6 GHz and a cluster of 13 elements at 4 GHz were used for illuminating the lens. The phase and amplitude excitations of the centre and six elements in a ring around it were 0 dB and 0°, and -8 dB, and $+12°$ respectively. Higher mode generation and scattering [Scott *et al.*, 1976] within the lens elements were believed to be the cause of high experimental side lobes. The beamwidth is reasonably independent of the scan angle.

4.2.2 *Zoned lens*

The use of a multiple beam antenna in a communications satellite has been introduced, and analytic techniques and some experimental results have been presented [Ricardi, 1977]. The inner surface of the lens is spherical. The length of the individual waveguides in the lens is adjusted to produce the phase advance necessary to convert the incident "spherical" wave into a "plane" wave. The length of the waveguides which introduce a phase advance greater than 360° is reduced so as to decrease the phase advance by n 360° where $n = 1$ or 2. This results in a "stepped" or "zoned" lens as characterized by the outer surface of the lens.

4.2.2.1 *Beam-forming networks*

The multiple feeds can be connected to a transmitter by many networks to obtain variable radiation patterns. This network, which is called a beam-forming network (BFN), largely determines the versatility of the antenna system. A combined feed network made of continuously variable power dividing junctions has been used as a BFN. With such a network and by properly dividing the power among several adjacent horns, pencil beams may be formed in any desired direction within the field-of-view.

4.2.2.2 *Recent experiments*

The design of a three-step lens has been reported [Scott *et al.*, 1976]. The feed array contained 61 horns located on a spherical surface using a nearly triangular grid with a capability of producing 61 beams. To produce a shaped beam, a variable beam forming network (VBFN) was used connecting the 61 feeds to a single port. Capability was provided to excite any one of the 61 pencil beams alone or any subset. When all 61 beams are equally excited a flat-topped beam with a width of over 18° could be produced.

4.2.2.3 *Phased array fed microwave lens*

Results [McGahan, 1975] have been reported of a microwave lens fed by a phased array containing fourteen $\lambda/2$ spaced monopoles located on a circular arc. One passive reflector was located $\lambda/4$ behind each feed element to reduce the radiation toward the centre of the feed. A waveguide lens with a planar front face was used in the experiment. The back face of the lens is not circular and is designed to meet the required scanning properties of the lens. A circular wavefront incident on the back face of the lens has to be transformed into a plane radiating wave from the front face. The lens provides two degrees of freedom: one is the shaping of the front face and the second is the selection of the medium between the two faces.

The phase paths were equalized by inserting dielectric phase shifters in each waveguide such that the phases across the front face were made equal to modulo 2π. Experimental measurements gave a gain loss of 2.2 dB at a scan angle of 7.3°. The half-power beamwidth changed from 1.8° at broadside to 2.4° at a scan angle of 7.3°.

4.3 *Arrays*

Array antennas have always, and are expected to continue to be, the preferred choice for many applications. Their biggest virtue is, however, their greatest disadvantage. Although the design theory of the array is simple and gives a unique possibility for controlling the radiation pattern, the need to control the radiation of each element of the array usually increases the cost, weight, etc., of most array antennas. In fact, if one sets a limit N on the number of feed points or phase shifters or power dividers, etc., that will be used in an antenna design, the antenna gain is limited to:

$$G_{max} \approx 10 \log N + G_0 \qquad \text{dB} \tag{1}$$

where G_0 is the gain of the individual elements of the array. Geostationary satellites using an array designed to cover the Earth FOV will have a maximum gain:

$$G_{max} \approx 10 \log N + 17 \qquad \text{dB} \tag{2}$$

Clearly N must be within practical bounds (because of complexity, weight and power requirements) e.g. less than 100, hence the gain of the array is limited to ≈ 37 dB. This figure should not be taken as a hard limit, but certainly an array gain of ≈ 50 dB seems highly unlikely to be achieved.

4.3.1 *Phased arrays*

When each element of the array incorporates a phase shifter, the beam can be scanned over an angular sector as large as 60°. Scanning over the FOV from geosynchronous orbit can be done with very little degradation in performance. However, without phase and amplitude control of the radiation from, or signal received by, each element, the radiation pattern cannot be shaped arbitrarily as with an MBA. Satellite switched-beam systems could use a phased-array antenna as a normal means of meeting the basic requirements for beam scanning. Side-lobe levels less than -30 dB can be achieved, and bandwidth $\approx$ 15% are well within the state-of-the-art. Note that this antenna can theoretically be designed for optimum side-lobe levels. A significant disadvantage is the limited bandwidth available from a phased array.

Some successful examples of phased-array antennas designed for satellites include:

- an electronically scanning 300 MHz wide microwave radiometer which has been tested in the Nimbus-5 satellite;
- an electronically scanning 250 MHz wide dual-polarized microwave radiometer which is operating in the Nimbus-6 satellite.

Investigation has been carried out [Kummer, 1975] on the design of a 3 channel multifunction circular phased array for operation in the 20 and 30 GHz bands. It has been estimated that with 12 dB gain elements and 5 mW of transmit power per element, the e.i.r.p. of the array would be 25 dBW. A communications link could be established at 20 GHz from a 400 km orbit using a 4 m diameter earth-station antenna and receiver with a 1500 K noise temperature.

An engineering model of a retrodirective or a self-steering array has been evaluated [Kummer and Villeneuve, 1968]. The system had two independent 125 MHz FM channels each transmitting in the 7.5 GHz and receiving in the 8 GHz band and covering a $\pm$ 15° cone. The beam directions are controlled by a CW pilot signal transmitted by the earth station. Two arrays were used to avoid beam squint, one for transmitting and one for receiving.

4.3.2 *Array antennas as an MBA*

Planar arrays of identical elements excited through a Butler beam-forming matrix (BBFM) [Cummings, 1978], a Blass beam-forming matrix, or equivalent networks, have MBA characteristics similar to a lens or reflector illuminated by feed-horn array. The design of these beam-forming networks operating

at frequencies up to 10 GHz has now matured and the networks are in common use. These systems are highly competitive with their lens, or reflector, counterparts. A choice of one over the other is often a case of designer's choice, and this choice is often based on personally-limited knowledge and experience.

The time delay, as opposed to phase shift, converging phenomena characteristic of the reflector type antennas makes the reflector system inherently capable of operating over a wider frequency band. Many lens type MBAs have an operating bandwidth comparable to a reflector system; but in most designs the structural and fabrication techniques required to make a lens design practical, decrease its bandwidth. Generally, the inherent operating band is largest for a reflector system and smallest for the planar array ($< 5\%$). Most lens systems have an operating bandwidth greater than 5%.

5. Systems operating above 40 GHz

The previous discussions apply generally to antennas operating below 40 GHz. Although many of the comments, etc., apply to antennas operating at frequencies above 40 GHz, practical consideration usually rules out the use of arrays because the desired aperture size exceeds one hundred wavelengths. A very large number of elements, more than is considered practical, would be required to provide the desired gain. The lens is also rather difficult to fabricate at these higher frequencies. This leaves the reflector systems as perhaps the only, or at least the best, antenna type for use at frequencies above 40 GHz.

Beam-forming networks operating at frequencies up to 60 GHz can, in general, be readily fabricated. However, at higher frequencies, known fabrication techniques do not appear to be adequate. At these higher frequencies, phase shifters, switches, and variable power devices use ferrites exclusively. All other agile devices (e.g., diode phase shifters, switches, power dividers, etc.) have high-loss, low-power handling capability, or are physically impossible to build.

6. MBAs for increased efficiency of spectrum use

Directive beams, such as those generated by an MBA, permit simultaneous operation on the same frequency via the same satellite by separate users transmitting different information. This frequency re-use characteristic can permit the same frequency band to be used over many separate areas within the FOV. Specifically, more users can transmit and receive signals over the same frequency band with tolerable interferences from one another.

An MBA can be designed with prescribed beam shape(s) and corresponding input (output) ports. For example, the Intelsat-IV satellite series uses an MBA whose radiation patterns were determined prior to launch. On the other hand, an MBA that is excited by a variable beam-forming network (e.g., DSCS-III) can change its radiation pattern after launch. Thus the radiation characteristics, over the FOV, can be varied when the satellite is in orbit.

In both of the above cases, the MBA could be the same device but the beam-forming network (BFN) would be considerably different. For this reason, the BFN is often considered as the most important sub-system of an MBA system. The system described by Ricardi [1977] is perhaps the most versatile and possibly the most complex BFN available. It is capable of exciting the associated MBA to produce virtually any radiation pattern that is physically able of being produced by the MBA aperture.

Spectrum use efficiency can also be increased by the use of dual-polarized antennas. Generally any MBA can be designed to radiate and receive orthogonally polarized signals. Recent results indicate that polarization isolation greater than 27 dB can be achieved in systems requiring many spot beams. Isolation between beams (i.e., spatial isolation) is also greater than 27 dB. It follows that four-fold (i.e., spatial and polarization) frequency re-use can be realized with an MBA, and in fact, for Intelsat-VI, up to eight frequency re-uses are being considered.

Lenses are among the antennas that are most attractive for use in an MBA system. At the lower frequencies (i.e., 4 to 6 GHz), the "bootlace" or transverse electromagnetic (TEM) lens is probably most attractive because of the increased number of design parameters available. However, at the higher frequencies (i.e., $\approx$ 10 GHz), the constrained waveguide or equivalent lens is most attractive. A solid dielectric lens is usually much heavier than other lenses and, consequently, it is seldom used.

7. Applications

A number of satellite services are planning to use the multiple beam antenna technology. Investigations are being carried on to determine the feasibility of using multiple beam antennas for some other services.

7.1 *Tracking and data relay satellites*

The tracking and data relay satellite (TDRS) will function as an integral part of the spaceflight tracking and data network (STDN). The TDRS consists of two geosynchronous relay satellites 130 degrees apart in longitude and a ground terminal located in the United States of America. The purpose of the TDRS is to provide telecommunications services which will relay communications signals between low earth orbiting user spacecraft and the user control and/or data processing facilities. The TDRS west has been placed at 171° west longitude and the TDRS east has been placed at 41° west longitude. One multiple access forward channel of the TDRS uses a phased array containing 30 helical elements. The transmit left-hand circular polarized antenna, having an e.i.r.p. of 34 dBW, will be steerable by ground command over a ± 13° conical field of view at a frequency of 2106.41 MHz with a 2.5 dB bandwidth exceeding 6 MHz.

7.2 *Broadcasting satellite*

One broadcasting satellite for either the USA or Canada could cover more than one time zone, requiring the need for multiple beam capability. In some places of Southern and Central America and in Europe, a multi-national satellite could be envisaged providing separate beams to different coverage areas.

7.3 *Maritime and aeronautical satellites*

For communications with a large number of mobile earth stations (aircraft, ships), it may be desirable to use low cost, low gain antennas. In such cases, the possibility of using multiple beam capability for the satellite has been investigated [CCIR, 1974-78].

8. Conclusions

Various types of multiple beam antennas have been and are being investigated for satellite communications. Reflector systems offer the benefits of simplicity; however, the side-lobe levels at the higher scan angles could be unacceptable, depending upon the application. Phased-array designs for spacecraft use are limited by the weight of beam-forming networks and prime power allocations for beamsteering and linear transmit operation. Lens designs offer good electrical performance. Waveguide lens designs are lightweight but have limited bandwidths. Bootlace lens designs are capable of broadband operation. The use of multiple beam antennas for satellite communications would increase the efficiency of frequency use; however, additional work is required to determine the degree of efficiency improvement.

The major sub-system in an MBA is the beam-forming network. There are two basic types of BFN; one provides a fixed distribution of radiated power to each beam port of the MBA and the other varies the distribution of power to accommodate the instantaneous communication requirements. State-of-the-art techniques permit the fabrication of systems operating at ≈6 GHz with four-fold to eight-fold frequency re-use. Relatively straightforward extension of these techniques will produce systems operating at ≈12 GHz. A plausible extension of most of the fabrication techniques can result in systems operating up to at least 40 GHz.

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REPORT 827*

RADAR SPECTRUM UTILIZATION

(Question 45/1)

(1982)

1. Introduction

Radar systems perform many functions in radiocommunications including radiolocation, radionavigation, altimetry, meteorology, radar astronomy and earth resource sensing. It is appropriate, for background purposes, to review the rationale for radar system needs, to examine some of the sharing problems created by the existence of these systems, and to discuss some aspects of interference reduction techniques and other matters germane to sharing (see Bibliography).

2. Radar requirements

To perform the above functions, radar system characteristics vary greatly with the application. It is not the intent of this section to examine requirements for all systems, but rather concentrate on those which tend to show the range of values which can be expected for such radar system parameters as frequency, bandwidth, power, antenna gain and waveform.

2.1 *Frequency*

The frequency needs for radar can be divided into three classes: necessary bandwidth, tuning range and operational band. Even though radar frequencies extend from HF to optical frequencies, the operational bands considered in this Report will be from 200 MHz to 40 GHz.

2.1.1 *Necessary bandwidth*

The system requirement for necessary bandwidth is inversely proportional to the system requirement for range resolution. Range resolution affects such system performance aspects as: classification through resolution of multiple objects and object imaging, improved tracking through improvements in maintaining tracks during merges and splits (i.e. during converging and diverging target manœuvres), and improved detection in clutter (i.e. undesired signal returns) through reduction in total clutter cross section in a resolution cell. Wide bandwidth for precision range measurement is also an important aspect of altimetry systems for use in low-level aircraft operations, and for precise altitude measurements from satellites. At the present time, the necessary bandwidth for radars is largely a function of the needs of the surface navigation community where 50 m resolution is quoted, but where 15 to 30 m resolution (10 to 5 MHz bandwidth) is common. There is a trend in those radars where operation in highly cluttered environments is essential to operate with very short pulses to reduce the clutter background. These systems, which can have a necessary bandwidth of 100 MHz or more, will continue to be a small but significant segment of the radar population. Radar altimeters, which can operate either in a pulse or FM-CW mode, currently have a necessary bandwidth in excess of 100 MHz for operation at low altitudes. As automated landing systems come into use, it is anticipated that a bandwidth of at least 200 MHz will be needed for the final landing approach.

2.1.2 *Tuning range*

The tuning range of a radar system derives largely from the factors of interference avoidance, elevation scanning and target decorrelation. Target decorrelation (i.e., the smoothing of target fluctuations) can be obtained because of the frequency sensitivity of complex targets. Pulse-to-pulse variations in the transmitted frequency have the effect of improving the detection continuity. The amount of bandwidth required is a function of several factors, but a frequency spread of 10 to 100 MHz is generally used. There is a class of radars which use frequency as a parameter to scan the antenna beam in elevation. The bandwidth required is a function of the necessary radar performance features and the design of the antenna, and may be as high as 10% of the basic system frequency. However, the need for tunable bandwidth derives primarily from the need for interference avoidance. Frequency flexibility within a band

* This Report should be brought to the attention of Study Group 8.

is an important aspect of radar design if the system is to have the capability, through frequency management, to avoid other transmitters which are causing interference. At the present time, the tunable bandwidth for the majority of radar systems is about 10% of the operating frequency. The proliferation of radar systems coupled with the use of wider necessary bandwidths has resulted in significant spectral crowding in a number of bands. This crowding has provided, in part, the motivation for the development of broader band components, a new technology which will permit the growth of tunability bandwidth from 20 to 25% of the operating frequency.

2.1.3 *Operational bands*

The many portions of the electromagnetic spectrum used by radars are determined by both the radar's operational requirements and the platform carrying the system. Frequency needs based on the operational requirements are:

2.1.3.1 *Long range air surveillance*

Radio frequencies in the VHF and UHF bands are used to take advantage of reduced propagation losses and reduced clutter cross sections. As the frequency is reduced below 200 MHz the reduced propagation losses begin to be offset by increases in sky noise temperatures. The upper frequency limit for long range detection is less well defined in terms of propagation losses, but it is heavily influenced by the clutter backscatter (especially from precipitation) as the frequency is increased above 1 GHz.

2.1.3.2 *Navigation*

This operation encompasses aspects of both surface and airborne navigation. Frequencies are needed which can provide sufficient bandwidth for good range resolution, measurement accuracy for low flying aircraft, narrow antenna beams from relatively small antennas to provide good angular resolution, and reliable detection of precipitation for airborne weather avoidance radars. Frequencies from 3 to 35 GHz meet these requirements. Since this mission is essentially for short range purposes, propagation losses are not of prime importance.

2.1.3.3 *Height finding*

Height finder systems which combine azimuth and range along with elevation angle determination require azimuth and elevation beamwidths which are narrower than those used for long range detection. 3 GHz is a good compromise frequency for this application since the propagation losses are reasonable for the typical operating range and since the required beamwidths can be obtained from an antenna with dimensions suitable for operation in mobile applications. Weather clutter remains a problem which must be resolved by other means. In this regard, height finder radars in the 3 GHz band are often used as weather radars. The frequency is ideal since good signal levels are returned from precipitation clutter without significant absorption losses. In addition, the height finding capability permits the determination of an elevation profile of the clutter.

2.1.3.4 *Tracking*

Tracking systems are required to provide highly accurate coordinate information on both terrestrial and extraterrestrial objects. For terrestrial tracking from mobile platforms, the frequency range from 3 to 10 GHz provides acceptable propagation losses, low sky noise, and antenna sizes which result in the narrow beamwidths and pedestal mechanics required for accurate tracking of manœuvering targets. For extraterrestrial tracking from fixed ground stations, the useful frequency range can be lowered to the VHF-UHF region where sky noise is still an acceptably low value. Further, the technology exists to construct and control the large, steerable antennas needed to provide narrow beams at these low frequencies. Trackers used for control of aircraft landing operations may use 35 GHz in order to obtain better angular resolution.

2.1.3.5 *Earth resource sensing*

The use of radar for mapping and land use studies has been demonstrated in the RADAM project conducted in Brazil. Similar application from space platforms is contemplated as a follow-on to the optical and radiometric data being collected by LANDSAT and other space systems. The intent of this active sensor program is to obtain data on such diverse phenomena as crops, ice, soil moisture, wind, waves, geological information, snow, and land use. Because of wide variations in surface reflectivity and in sensor measurement errors, measurements must be made at more than one frequency if measurement ambiguities are to be resolved. Further, it has been shown that there is an optimum range of frequencies for each of the above items. In order to make meaningful measurements of earth resources, selected frequency bands from 1 to 75 GHz need to be utilized.

2.2 *Power*

Peak power output from radars range from a fraction of a watt (for certain FM-CW altimeters) to megawatts (for long range detection, height finding and tracking). There are no generalities which can be applied to the power aspects for all cases. There are, however, some practical limitations which can give some guidance when considering power levels. For instance, surface navigation radars need only "see" to the horizon. For this application, peak transmitter powers will be in the range from a few kilowatts to a few hundred kilowatts depending on the size of the ship. On the other hand, transmitter peak power in aircraft can vary from a few watts (e.g., altimeters) to over a megawatt. This latter figure applies only to aircraft which perform long range surveillance. Other aircraft with search and tracking requirements of more moderate ranges operate with peak powers up to a few hundred kilowatts. Land and sea based search and track radars have peak power values ranging from several hundred kilowatts to more than 10 MW. The latter value applies to a class of stacked beam height finders used in air traffic control where simultaneous transmissions in each elevation beam sum to the total power indicated above. In space, the power requirements for space navigation functions are modest, and should have little impact on terrestrial sharing. Of greater concern are spaceborne radars used for earth resource sensing. Initial peak power requirements for this purpose are about 1 kW. Future systems, operating in bands where atmospheric absorption is significant, may require peak powers of tens of kilowatts.

Megawatt levels are generated in systems which operate up to 6 GHz. From 10 GHz up, power level tends to decrease with increasing frequency.

2.3 *Antennas*

Radar antennas can be of almost any form; spirals, horns, reflectors, arrays, or phased arrays to name a few. In airborne systems used for navigational purposes (altimetry, station keeping, Doppler navigation, etc.), antenna gain is generally between 3 and 10 dBi. Systems used for surface navigation, air traffic control and long range surveillance have antennas which are usually rotating reflectors or arrays. Here, antenna design is such that the azimuth beamwidth is much narrower than the elevation beamwidth with, in some cases, beam shaping applied in the elevation plane to approach range independent detection capability for aircraft flying at a constant altitude. Antenna gains for these systems are generally in the range of 25 to 35 dBi. Height finders and trackers have the requirement of resolving targets in both azimuth and elevation and, in the case of the latter, providing accurate coordinate information. The narrow beamwidths needed to meet these requirements imply the need for higher gain antennas. Antenna gains of 35 to 45 dBi are common in this application. From the point of view of radiolocation operations, antenna gains in excess of 50 dBi are seldom used primarily because of the difficulties encountered in acquiring targets with such narrow beamwidths. Antennas with gains of more than 60 dBi are used for radar astronomy.

Basic to radar operation is the fact that the beam generated by the antenna moves in space. In some cases, such as airborne navigation devices, the beam is fixed with respect to the vehicle and its motion is prescribed by the aircraft's flight path. Search and height finder systems generally move the beam through 360° in azimuth by mechanically rotating the antenna. Rotation rates from 5 to 30 rpm are common with the lower rates being associated with long range search systems. There are, however, short range search systems with rotation rates in excess of 100 rpm. Such systems are expected to increase because of their utility in navigation and collision avoidance applications on high speed ships. In addition to azimuth scan, some of the height finder radars scan the beam in elevation. Elevation coverage to 45° is not uncommon. The rate of elevation scan is made compatible with the azimuth beamwidth and rotation rate in order to insure complete volume coverage. Tracking radar antennas may point to almost any angle with respect to the platform. When tracking, the antenna motion is

primarily a function of the target motion with respect to the radar. Thus, track antennas do not repetitively cover the total volume surrounding a site, but rather they dwell for a relatively long time in a more restricted angular region, which, as pointed out above, may be at any azimuth or elevation angle with respect to the radar. Note that the scan characteristics of phased array radars which combine the search and track functions are not significantly different from the characteristics of mechanically scanned antennas. Beam position in the track mode is again dependent on the target; in the search mode, however, the volume to be covered is divided into discrete beam positions with each position being examined at intervals which generally correspond to the rotation rates given above.

Antenna design generally involves a compromise between antenna efficiency and side-lobe level. Measurements on existing radar antennas show a median gain of -10 dBi which implies that only a few of the side lobes exceed 0 dBi. However, when the antenna is placed at its operational site, the local terrain, buildings and other obstructions will have significant effect on the *in situ* side-lobe levels. As the rotating beam sweeps around the site, side lobes can, and will be generated in arbitrary directions with gains which may be only 5 dB below the main beam gain.

The antenna polarization is generally fixed although some antennas exist which have a variable polarization capability. It should be noted that the polarization ascribed to a given radar antenna usually only applies to the main beam region around the pointing axis.

2.4 *Waveforms*

As with other radar characteristics, the waveforms used can be extremely diverse. Not only do the waveforms vary greatly from system to system, they can generally be varied within any given system. Thus, the range of pulsewidths, repetition rates and duty factors will be addressed rather than specific system parameters.

2.4.1 *Pulsewidth*

Transmitted pulsewidths range from CW to about 10^{-8} s. The simple CW signal is used in some Doppler systems which are concerned only with measuring velocity. In other systems, notably altimeters, a frequency modulated CW (FM-CW) signal is employed where the necessary bandwidth is established by the accuracy required. Simple pulsed radars have pulsewidths ranging from 10^{-5} to 10^{-8} s. The shorter pulses provide acceptable range resolution, while the longer pulses are needed where detection range is more important than resolution. A more sophisticated form of pulse radar permits the transmission of pulses which can exceed 10^{-3} s. By pulse compression techniques, these long pulses, transmitted with a known phase or frequency progression, can be processed in the receiver to yield a pulse whose compressed width is commensurate with the bandwidth of the transmission. This method of synthesizing short pulses has the advantage of illuminating the target with sufficient energy for long range detection while obtaining a pulse from the receiver which provides good range resolution.

2.4.2 *Repetition rate*

To a large degree, repetition rate is determined by the need for unambiguous range measurements to the object. Thus, for long-range surveillance radars, the repetition rate is in the range of 100 to 300 Hz. For short range navigation radars, the rate may increase to about 2 kHz. In height finder radars, the rate varies as a function of elevation angle, being greater at the higher angles. The repetition rate for tracking radars is in the range from several hundred to several thousand hertz, while repetition rates in the range of tens to hundreds of kilohertz are typical of pulse Doppler radars. In the latter, the high repetition rate is needed in order to see targets in clutter. The range ambiguities inherent at the repetition rates used can be resolved in a variety of ways. In many radar systems, the repetition rate can be jittered or staggered as an aid in interference reduction or to identify second-time-around echoes.

2.4.3 *Duty factor*

The duty factor of a radar is defined as the product of the repetition rate and the pulse width. It can vary from 1.0 (CW) to about 0.0001. The duty factor for most pulse radars will fall in the range 0.01 to 0.0001, while the duty factor for pulse Doppler systems may approach 0.5. It is expected that some new radars will employ solid state amplifiers as part of their transmitter. For these devices, efficiency considerations will dictate duty factors of 0.15 to 0.3.

3. Interference problems

By themselves, the diversity of radar characteristics, in terms of frequency, power, antenna properties and waveforms, define an extremely complex electromagnetic environment. When coupled with the fact that these systems are operated from fixed and mobile land sites, aboard ships and aircraft, and in space vehicles, the potential for interference with other radiocommunication services seems inescapable. In the following sections, various aspects of the radar interference problem will be discussed. A factor common to all cases is that sky and ground wave components of propagation are insignificant for the frequencies under consideration. The relevant propagation factors will include line-of-sight, weather scatter, tropospheric scatter, ducting and other anomalous modes likely to occur.

3.1 *Land sites*

The potential for line-of-sight interference at land sites will be most heavily influenced by the local geography; hilltop sites being the most prone, valley sites offering more protection. Of almost equal importance to the geography is the existence and distribution of buildings and other obstructions. Reflections from such obstacles will have the same effect as unwanted side lobes; i.e., an increase in the amount of time that interference will exist. Over-the-horizon interference can arise by means of a number of anomalous propagation modes. Forward scatter from hydrometeors is one of these modes. Because of the particle size-frequency dependence on scattering cross-section, it is expected that the coupling via this mode will increase with frequency up to about 10 GHz, where path losses from attenuation will tend to compensate for any increase in cross section. Tropospheric scatter will be another mode which must be considered for systems operating at 6 GHz or below. Super-refraction, or ducting can also occur at land sites given the proper atmospheric conditions, noting that such ducting can occur at all frequencies of concern. Further, propagation in overland ducts is subject to significantly higher losses than in over water ducts. Some idea of the distances involved can be obtained using the techniques of Appendix 28 to the Radio Regulations. This Appendix is specifically for interference to an earth terminal in the fixed satellite service. However, by properly establishing interference levels and probabilities of occurrence, interference to other services can be estimated. It should be noted that the most pessimistic results will be obtained since the procedure is based on the most unfavourable case assumptions as regards interference.

3.2 *Shipboard sites*

All of the comments noted above for land sites apply to shipborne sites as well. However, super-refraction becomes more of a problem since ducting conditions are more prevalent at sea and, once established, permit propagation over much longer distances than observed over land.

3.3 *Airborne sites*

Line-of-sight coupling is the major interference problem with radars located aboard aircraft. Without considering any of the scatter or anomalous propagation modes, a single aircraft flying at a height of 10 to 15 km can affect ground services over almost a million square km area. Increases in the area affected can occur from rain scatter and, to a lesser degree, from ducting.

3.4 *Space sites*

Radars in space pose more of a problem than space communications systems. The reason, of course, is that the power flux density (pfd) limits applied to communications systems are not compatible with the radar function. Signal levels significantly higher than those used in communications services must illuminate the Earth if useful signal-to-noise ratios are to be obtained. However, these high levels will not exist over the same large areas associated with the communications systems since the radars will be in lower, non-synchronous orbits, and the antenna beamwidths will be narrower. It is not expected that the space systems used for navigation and docking will significantly impact terrestrial systems.

3.5 *Communications services* (fixed, mobile, fixed-satellite, amateur, broadcasting)

The diversity of radar characteristics and their locations suggest that there is a large potential for interference to communications services. In general, this interference will be most severe when line-of-sight propagation conditions prevail, since the transmissions from both the side lobes and main lobe of the radar antenna must be considered. In this case, interfering pulses can be expected to be almost continuous. As the distance increases, however, the effect of the side lobes will decrease and the interference will become a function of the particular scanning in use by the radar. Interference is also a function of the type of antenna used by the

communication service. For omnidirectional or wide beamwidth antennas, a high probability of intercepting interference signals will be offset by a low level of interference due to the low value of antenna gain. The converse is true for highly directive communication antennas, but only when the antenna has been designed for very low side lobes.

Information transfer in communications systems can be either in analogue or digital form. Radar interference in analogue systems can vary from barely discernable to extremely annoying. Since the radar pulses typically interfere with a small percentage of the information being transmitted, the analogue information will not be completely destroyed. For example, it is shown in Hatch *et al.* [1970] that voice intelligibility, as measured in terms of articulation score is not appreciably lowered in amplitude modulated communications systems until the radar duty factor approaches 0.50.

With respect to interference to radars, the level of interference which radars can tolerate from communications systems must be handled for specific systems, since performance degradation will be a function of the type of interference, the function of the radar, and the interference reduction capability of the radar. Narrowband AM and FM signals, as well as CW signals can gradually desensitize the radar receiver as the interference signal level increases beyond the sensitivity level. For some receiver-signal combinations, the operator may not be aware of the resulting performance degradation. Wideband FM signals will appear as noise to the radar receiver and an interference criterion may be established by the relation of this noise to the receiver thermal noise. Pulse interference will, in general, require some other criterion depending on the density and distribution of the interfering pulses on the radar display. An existing qualitative criterion contains elements of pulse density, pulse level, and the extent of the angular regions of the display showing interference. Lustgarten and Grigg [1977] describe a more quantitative criterion for pulse interference.

3.6 *Radiodetermination services* (radiolocation, radionavigation, Earth resources sensing)

Radars do interfere with other radars and other systems of the radiodetermination services. The mechanisms noted above apply here also.

3.7 *Non-radio systems*

The high peak power, fast rise and fall time pulse characteristics, pulse repetition rate and operation frequency characteristics of some radars suggests that there is a potential for interference to electronic equipment, such as computers, electronic signal and data processors, computer data transfer systems and measuring equipment. In general, this interference will be of the same nature as that indicated above for communications services.

4. Radar sharing with other services

Service using radars do share frequency bands with other services. It is appropriate, therefore, to note some of the procedures and techniques which promote this sharing.

4.1 *Frequency management*

The allocation and assignment process is, and will remain, an important tool in the solution of the sharing problem. The fact that the Table of Allocations provides a given band for a service using radars allows frequency managers to assign frequencies so as to reduce the mutual interference between similar systems. Cohen and Mayher [1979] show how computer aided techniques should provide significant benefits in this area. However, frequency management is more than just an allocation and assignment procedure. Although tuned to their assigned frequencies, radars can emit signals which depart significantly from the assignment. In addition to the spectrum broadening to be expected from pulse emissions, spurious and harmonic frequencies are often generated in the transmitter. This is particularly true of magnetron type transmitters. The problem is, of course, magnified when it is considered that higher frequencies can be radiated from the antenna via higher order modes (note that the direction of radiation from these modes is not necessarily coincident with the principal radiation axis). Procedures now exist to promote spectrum efficiency through better waveform design and more stringent controls on spurious emissions. Efforts to implement these procedures should continue as well as efforts to strengthen them.

It should be clear that frequency management cannot solve all the problems of sharing. The wideband operation required from many radars can interfere with one or many channels of a communications system which has been allocated a sharing role. The frequency agile capability of some of the wideband radars should reduce the amount of time that interference exists in systems which are narrower in bandwidth than the tuning, or agility bandwidth of the radar.

4.2 *Geography*

A second method for encouraging sharing is the judicious siting of systems which are allocated the same frequency space. The mobility of radars, however, has a tendency to nullify some of the advantages that geographic siting can provide. This is particularly true when airborne and spaceborne radars are involved. Here the location of these systems surmounts most geographical obstacles and provides a direct, line-of-sight link between systems. For the case of surface radars, especially shipborne, geographic siting can still play a part. In this case, land based services can be located far enough inland or be protected by topographical features. The use of pits, fences, vegetation and other shielding techniques should be considered to augment the effects of natural terrain.

4.3 *Interference reduction techniques*

Even when interfering signals are incident upon some system, interference rejection is possible. There are a number of system and ciruit techniques which can be used to remove or reduce the effects of interference. These techniques may be grouped as follows:

- *prevent* the interference from entering the system (§ 4.3.1);
- *avoid* the interference by changing a system parameter (§ 4.3.2);
- *neutralize* the interference by signal processing (§ 4.3.3);
- *ignore* the interference which does not conform to the information format of the system (§ 4.3.4).

It should be clearly understood that not all the techniques apply to all services or systems. Also most of the techniques involve a certain degree of performance degradation over that which would be obtained without the use of the technique and without interference signals being present. Finally, the effectiveness of some of the techniques may be limited in terms of the number and level of interfering signals which can be handled.

4.3.1 *Prevent the interference*

The geographical aspects mentioned above are a form of prevention. However, from a systems point of view, antenna techniques are of primary importance. An obvious candidate for this area is antennas with low side lobes. Since close-in side-lobe levels are closely tied to the aperture distribution and therefore the system mission, the biggest payoff in low side-lobe techniques would be those directed toward reducing the back lobe and far out side lobes. With reflector antennas this means continuing attention to hoods or shrouds. With array antennas, maintenance of dimensional tolerances to insure close control of the amplitude and phase distribution across the aperture is an important consideration.

Polarization discrimination provides another way of preventing unwanted signals from entering the antenna. Both reflector and array antennas will respond to cross-polarized signals, with the angular distribution (antenna pattern) of this response being different from that of the desired polarization. This facet of antenna behaviour represents another avenue for the entry of unwanted signals. Some antenna types can benefit from a polarization screen across the aperture which effectively shorts out the orthogonal polarization component before it can be focused by the antenna. Such a screen is of value only when the antenna operates with a single, fixed polarization.

A form of prevention, albeit not at the antenna, is image suppression. Unless positive steps are taken, an interfering signal at the image frequency can be just as damaging as one which is on tune. Techniques do exist which can suppress the image at the mixer or place it far enough out of the band where it can be attenuated by filters.

Lastly, interference to electronic equipments, such as those indicated in § 3.7, can be minimized by a combination of several techniques. Attenuation of the radar signals by geographical features, such as pits and vegetation, is one approach. A complementary approach is to build integral shielding (i.e. screen room) into the walls, floor and ceiling of the room containing the electronic equipment. In addition, for certain land radar sites, blanking sectors could be established in azimuth and/or elevation to prevent the radar from radiating towards the electronic equipment.

4.3.2 *Avoid the interference*

Interfering signals can be avoided in some cases through the use of one or a combination of several techniques.

- *Frequency agility* – Frequency agility, the ability to rapidly shift from one frequency to another, is practical for some types of services and can provide a quick way of avoiding interference. As mentioned above, agile systems operating in the same band as non-agile systems can reduce the total time over which interference occurs.
- *Changes in polarizations* – In this technique, system circuitry seeks to null the interfering signal by adjusting the polarization of the antenna. The danger, of course, comes from the possibility that the interfering signal has the same polarization as the desired signal.
- *Diversity* – Interference can be avoided through the use of alternate links of diverse characteristics, such as space and frequency.

4.3.3 *Neutralize the interference*

A variety of signal processing techniques are available to reduce or eliminate the effects of interference. One of these techniques is the side-lobe canceller which samples the electromagnetic environment by means of auxiliary antennas, suitably weights the interfering signals in adaptive loops to obtain the proper level, and performs a coherent subtraction in the main receiver channel. The principal benefit of this technique is that the cancellation is done on the interfering signal only, without disturbing the desired signal. The technique requires a separate auxiliary antenna and loop for each interfering source. The technique works well when the interference is from a point source. Interference which arrives as a result of forward scatter from rain may appear as a distributed source. The efficacy of the technique under this condition is still under study. A related technique is side-lobe blanking. As before, an auxiliary antenna and receiver are used. In this case, however, instead of coherently subtracting the unwanted signal, the main receiver is blanked whenever the signal level in the auxiliary channel exceeds a selectable threshold. The technique is well suited to eliminating pulse interference having a low duty factor. The effect of such blanking on the information content is a function of the information format, the number of interfering sources and the duty factor of the interferers. A third technique involves transmitting a coded waveform and matching the receiver for the particular code. Pulse compression is an existing example but the concept can be expanded to include other spread spectrum techniques. While such techniques do provide some reduction in non-matched signals, the benefits of this as a spectrum conservation method for radars have yet to be demonstrated.

4.3.4 *Ignore the interference*

It is possible to recognize signals at the output of a system which do not conform to the format of the desired information. A simple example of such a pattern recognition technique is the video integrator. The technique utilizes a recirculating delay line whose delay is equal to the interval of time between transmitted pulses. Valid signals will add in this loop while unsynchronized interference will not. The output is then free from the unwanted pulses. Other algorithms can be implemented in a data processor to recognize not only pulse interval but also pulse width.

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REPORT 651-2

SPREAD SPECTRUM TECHNIQUES

(Study Programme 18B/1)

(1978-1982-1986)

1. Introduction

Study Programme 18B/1 calls for descriptions of typical bandwidth expansion techniques and an analysis of what advantages for spectrum use can accrue through their development. One such technique is known as "Spread Spectrum". This Report describes broadband "Spread Spectrum" techniques and the interference rejection capabilities of these systems. Report 652 (Geneva, 1982) addresses the subject of frequency sharing criteria identified in Study Programme 18B/1.

A Spread Spectrum (SS) system can be defined as one in which the average energy of the transmitted signal is spread over a bandwidth which is much wider than the information bandwidth (the bandwidth of the transmitted signal is wider than the information bandwidth by at least a factor of two for double sideband AM and typically a factor of four or greater for narrowband FM and 100 to 1 for a linear SS system). These systems essentially trade the wider transmission bandwidth for a lower power spectral density and increased rejection to interfering signals operating in the same frequency band. They, therefore, have the potential of sharing the spectrum with conventional narrowband systems because of the potentially low power that is transmitted in the narrowband receiver passband. In addition, high levels of interference will be rejected by SS receiving systems. These systems should therefore be examined to identify how efficiently they use the spectrum.

Two distinct types of bandwidth expansion SS techniques need to be discussed. These are the techniques that provide either linear or non-linear interference signal rejection. The classical FM approach typifies non-linear techniques because there is only an increase in the output (S/N) ratio when the input (S/N) is greater than the first or noise capture ratio. This means that the input (S/N) must be typically greater than 10 dB in order to obtain a linear enhancement against noise. In contrast to the FM type of system, the SS systems described in this Report are linear so that the improvement remains constant even if the input wanted-to-unwanted signal ratio is negative. The output wanted signal-to-interference signal ratio $[(S/I)_{out}]$ is increased over the input wanted signal-to-interference ratio $[(S/I)_{in}]$ and is defined as the Processing Gain (PG) of the system. This PG might typically be 100 to 1, or larger, since, by definition, the PG is found to be:

$$(S/I)_{in} = (S/I)_{out} - 10 \log PG \qquad \text{dB} \tag{1}$$

A system with a PG of 100 (and no loss due to non-ideal signal processing) and a minimum output (S/I) of 10 dB requires that the input (S/I) is, at least

$$(S/I)_{in} = 10 \text{ dB} - 10 \log (100) = -10 \text{ dB}$$

A linear SS system that can operate with an input (S/I) of -10 dB is extremely desirable since with an unwanted signal 10 dB higher than the wanted signal, the system can still be effectively used. For conventional systems with an input (S/I) of -10 dB, the wanted signal would be suppressed or "captured" and no information would be transferred.

A second major feature of commonly used SS techniques is that the resulting transmitted signal is a wideband low-power density signal which resembles noise. Therefore, the transmitted signal is not readily detected by a conventional receiver. Recovery of the baseband information from the wideband transmitted signal can be accomplished only through correlation or Matched Filter (MF) signal processing. Because of this property, the unintended listener does not detect the baseband information and because of the low power density, it may not cause any significant interference effects to other users of the spectrum. SS inherently provides a degree of message privacy to nonspread spectrum systems as well as other SS systems using different codes and no special signal processing. The coding also provides a selective addressing capability. Multiple users using different codes (code division multiple access) can simultaneously transmit in the same frequency band with a minimum amount of cross interference (codes that are used should have a low cross-correlation function).

A third advantage of SS techniques over conventional modulation techniques is increased transmission reliability in the presence of selective fading and multipath effects. This advantage can be significant for typically encountered fading transmission mediums, e.g. in tropospheric scatter systems. Increased resistance to multipath is a direct consequence of spreading the transmitted bandwidth. As a first approximation, improvement is directly proportional to the ratio of transmitted bandwidth to information bandwidth. Receivers built to detect SS signals typically generate, prior to their final demodulation, a cross-correlation function between a replica of the transmitted signal and the signal received from the antenna. The correlation function of the wanted signal is always chosen to be as "good" as possible, i.e. maximum output at the centre and the signal falling to near zero in a time period equal to the reciprocal of the transmitted signal bandwidth and staying at near zero at all other times. Multipath degrades link performance when it combines with the direct signal in such a manner as to degrade the correlation function of the detected signal by reducing its peak value. The introduction of false trailing peaks into the correlation function due to simple multipath is typically not a problem. The receiver will detect and process the first peak of adequate amplitude, either the direct signal if it is strong enough or the first reflected signal of adequate amplitude if the direct signal is interfered with. In the latter case, timing becomes synchronized to the multipath return and it is processed in lieu of the direct signal. Consequently, for multipath to be destructive, it must occur with a differential delay less than the duration of the peak of the correlation function, with a phase that causes destructive interference (cancellation rather than enhancement) and with an amplitude adequate to prevent the peak from exceeding detection thresholds. As the transmitted bandwidth is increased, the duration of the correlation function peak is proportionally decreased, and the multipath differential path delay that can affect performance is also proportionally decreased.

Finally, the management of frequency bands that encompass only SS systems could conceptually be simpler than the management of bands occupied by conventional narrowband and bandwidth expansion equipments. Once a band has been designated for SS use and the modulation technique determined, the primary concern is to insure that the same spreading codes (i.e. direct sequence (DS) and/or frequency hopping (FH) as described in the next section) are not being used by more than one user. This could be accomplished through "careful" code assignment techniques and/or record keeping procedures.

One of the techniques (non-sinusoidal carrier) has the additional advantage of permitting the employment of SS in systems requiring a larger processing gain than can be attained by the other techniques. This technique provides the additional processing gain by spreading the carrier as well as the modulation.

In summary, the advantages of SS modulation techniques include:

- large interference rejection capabilities, for both SS and conventionally modulated signals;
- low-power density transmissions, resulting in minimum co-channel interference effects;
- message privacy;
- selective addressing;
- selective fading rejection;
- simple frequency management procedures.

These SS properties have been extensively discussed in literature on the subject. General references describing SS systems can be found in the bibliography. An extensive bibliography describing SS can also be found in [Dixon, 1976].

The following describes typical SS techniques, and the interference rejection capabilities of these systems. Report 652 (Geneva, 1982) describes the interference effects that low-power density transmissions have on other systems. Discussions of the other advantages or disadvantages, including detailed discussions of how efficiently narrowband and broadband expansion systems use the spectrum, need to be addressed in future reports.

2. SS techniques

2.1 *General*

Definitions for various types of spread spectrum techniques/signal structure are as follows:

- *Direct sequence (DS) spread spectrum:* a signal structuring technique utilizing a digital code sequence having a chip rate much higher than the information signal bit rate. Each information bit of the digital signal is transmitted as a pseudo-random sequence of chips.

- *Frequency-hopping (FH) spread spectrum:* a signal structuring technique employing automatic switching of the transmitted frequency. Selection of the frequency to be transmitted is typically made in a pseudo-random manner from a set of frequencies covering a band wider than the information bandwidth. The intended receiver would frequency-hop in synchronization with the code of the transmitter in order to retrieve the desired information.
- *Hybrid spread spectrum:* a combination of frequency-hopping spread spectrum and direct-sequence spread spectrum.
- *Chip rate:* an encoding bit rate; coding rate.

Two other types of spread spectrum modulation exist. The first uses pulsed frequency modulation or "chirped modulation" in which a carrier is swept over a band of frequencies. Radar systems, in particular, may use a sweep rate that is a linear function of time (see Report 837 for methods used to calculate the emission spectrum for such radars). The second spread spectrum type employs a non-sinusoidal carrier to provide additional processing gain. The following discussion does not specifically discuss the hybrid or chirped spread spectrum types.

2.2 *DS modulation system*

In general, amplitude modulation (AM), frequency modulation (FM) or phase modulation (PM) could be used to modulate a baseband DS SS signal on an RF carrier. A common modulation method is biphase phase shift keying (PSK). This is implemented as shown in the transmitter and receiver block diagram in Fig. 1 which is the basis for the following discussion.

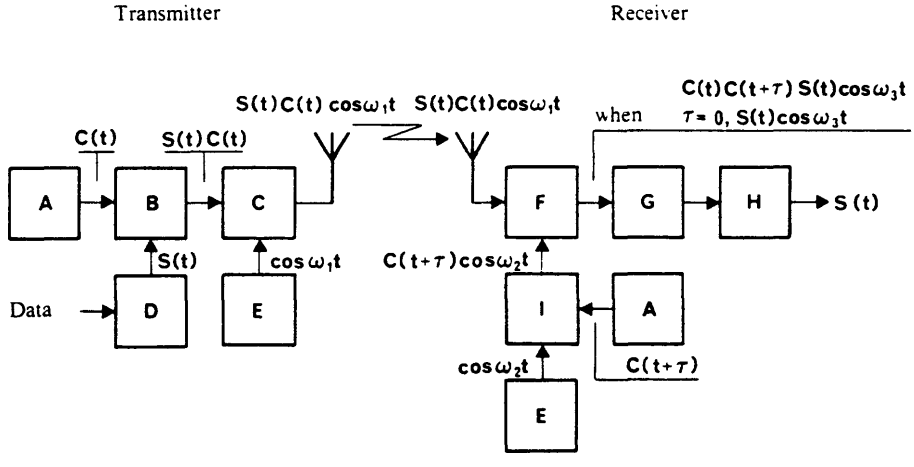


FIGURE 1 — *Basic direct sequence transmitter and receiver block diagram*

A: Pseudo noise sequence generator
B: Binary modulator
C: Modulator
D: Binary encoder
E: LO
F: Correlator
G: I F
H: Demodulator
I: Balanced modulator

Figure 1 indicates that in the DS technique which uses a correlation or time domain approach, the desired digital data signal $S(t)$ is multiplied by a coded signal (pseudo noise) $C(t)$ to provide the SS signal $[S(t) \cdot C(t)]$. In a frequency domain analysis of this operation, this multiplication would be equivalent to convolving the wideband spectrum of $C(t)$ with the narrowband spectrum of $S(t)$. In most situations, the bandwidth of $C(t)$ is 100 to 1000 times that of $S(t)$. This effectively results in producing the wideband spectrum of $C(t)$ or the SS as the transmitted signal. The bandwidth between first nulls in the envelope of the spectrum of $C(t)$ is equal to two times the code rate. In the receiver, this signal is multiplied by a synchronized coded signal $[C(t + \tau)]$ and then filtered to match the information bandwidth. This filtering reduces the interference signal level because it has been spread to the bandwidth of $C(t)$ and produces the PG of the SS system (see equation (1)). The baseband signal

representations for the encoded data and the coded information in the time domain are shown in Fig. 2. The binary data is shown for illustration purposes as a square wave. Frequency domain representations of the narrowband and SS signal are shown in Fig. 3.

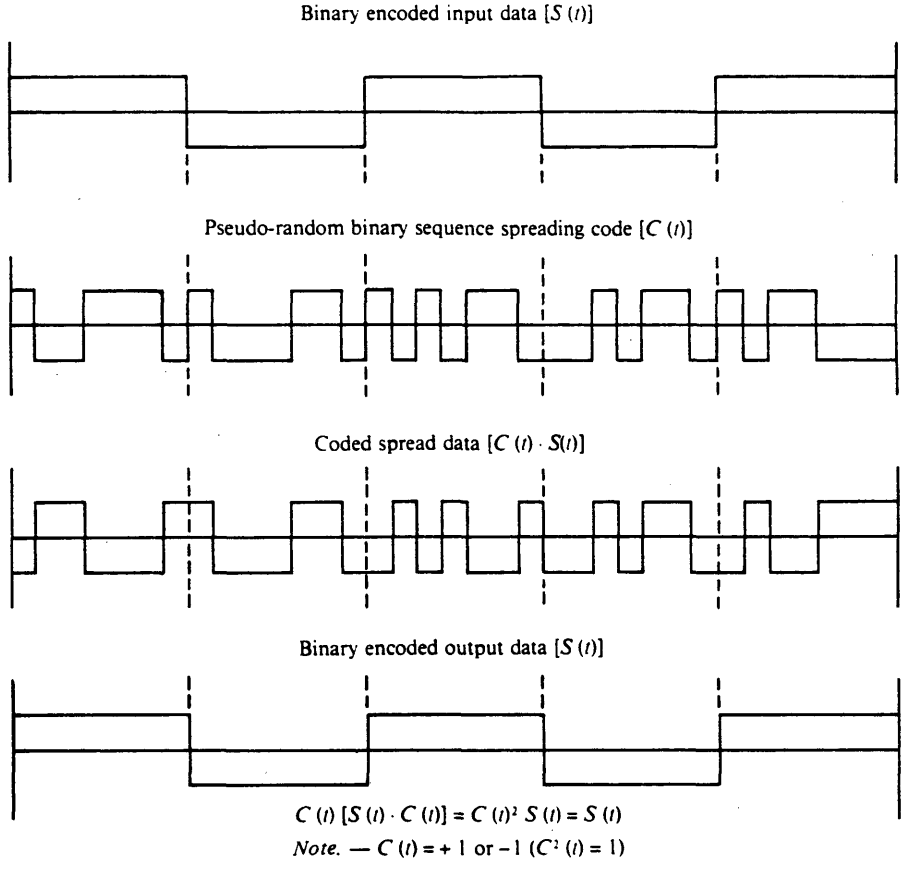


FIGURE 2 — *Time domain signal representations*

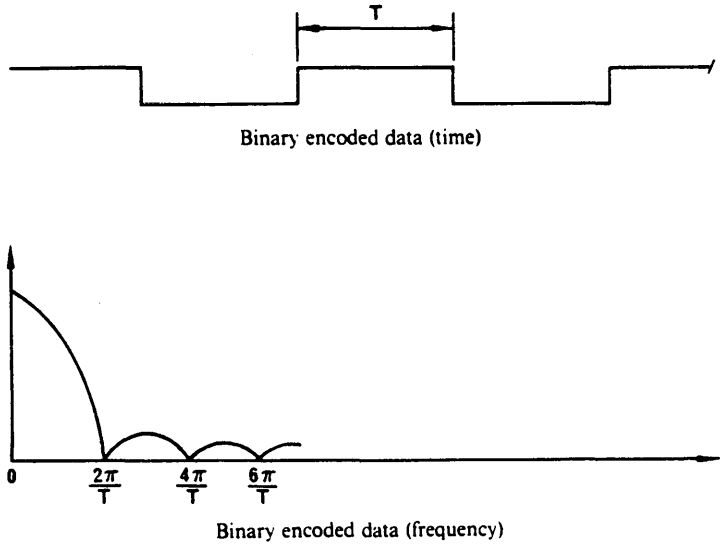


FIGURE 3a — *Unspread time frequency relationships*

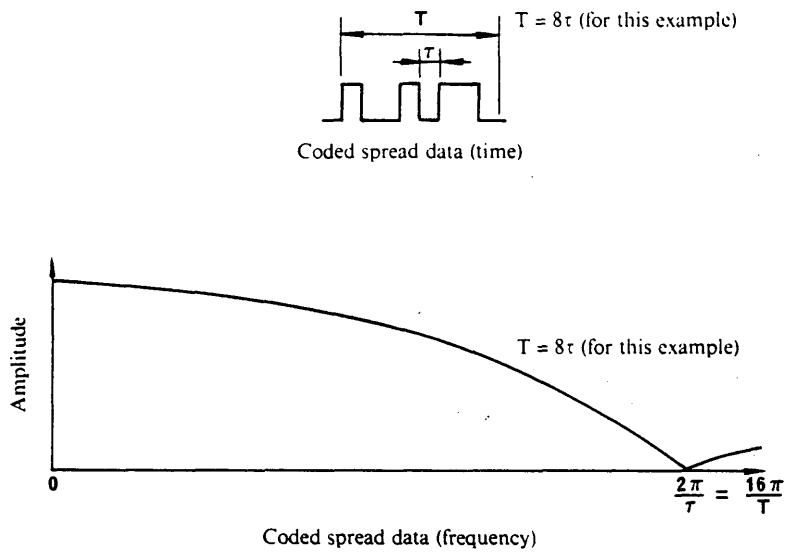


FIGURE 3b — *Direct sequence spread spectrum time frequency relationship*

A second method of implementing the DS SS technique would be to use a structure derived from the Fourier transform of the correlation function. This would employ a frequency domain or matched filter (MF) receiver signal processing structure. One method of generating the transmitted signal would be to employ an impulse response of a filter $H(\omega)$ to represent the mark and space of the transmitted signal. This impulse response could be equal to the coded signal $[C(t) \cdot S(t)]$. In the receiver, this signal is passed through a matched or conjugate filter which obtains the same signal $S(t)$ and PG as the correlator by operating in the frequency domain. The block diagram of this typical MF system is shown in Fig. 4 and has usually been employed with a fixed mark and space filter structure. Since this produces a repetitive coded signal, most systems that employ SS techniques use the correlator approach that allows for easy generation of coded signals which do not repeat themselves for long periods. Various types of hardware implementations of the MF technique have been employed. Surface acoustic wave (SAW) devices offer tremendous potential for high time bandwidth product delay line MFs [Bristol, 1974]. SAW devices have also been used in DS applications. All fixed MF systems have the disadvantage of "fixed" repetitive codes while having the advantage of relative simplicity.

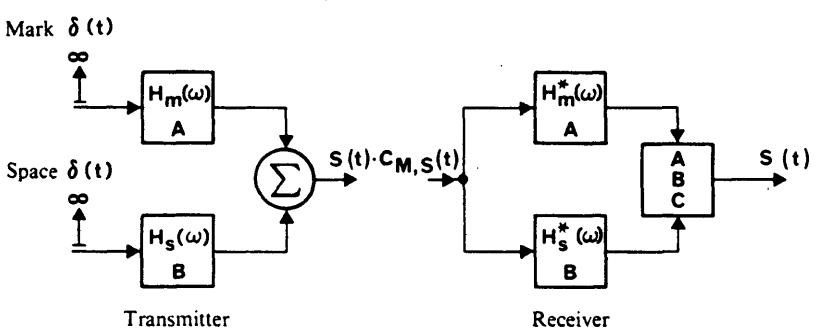


FIGURE 4 — *Ideal frequency domain matched filter signal processing*

A: Mark
B: Space
C: Comparator

2.3 *FH modulation system*

In the basic FH system, the carrier frequency is pseudo-randomly changed in discrete increments over a range of frequencies at a rate typically close to the information rate. In a general sense, an FH system is a frequency shift keying (FSK) system in which a large number of code selected frequencies are used to transmit the information carried by one narrowband complement of frequencies in an FSK system. A block diagram of a basic FH system is shown in Fig. 5. In the FH transmitter, a code generator controls a frequency synthesizer or local oscillator (LO) which pseudo-randomly changes frequencies. The transmitted output spectrum of the FH signal is quite different from the DS spread signal in that the instantaneous spectrum of the FH signal ideally looks like a narrowband or single frequency signal. In reality, the instantaneous frequency hopper output may be composed of sidebands and spurious frequencies created by a non-ideal transmitter. The long-time average spectrum has a power density which is spread and approximately equal to the power density in the bandwidth of the direct spread system as shown in Fig. 6.

An FH modulator may also be combined with the DS modulator to produce a DS modulated spectrum at each of the specified hopping frequencies. The power spectrum at each of these frequencies is equal to the DS spectrum of approximately a $(\sin X/X)^2$ spectrum for a rectangular pulse transmitted signal (i.e. for the typical case in which the bandwidth of $C(t) \gg S(t)$).

The FH receiver mixes the input signal with an LO signal controlled by a synchronously operated code generator. This produces a constant output desired signal in the IF passband. Interference which is narrowband relative to the FH bandwidth and not synchronized with the hopping LO only occasionally appears in the passband of the IF. The IF passband, therefore, rejects the undesired signal power in the same way as in the DS system. For wideband undesired interfering signals, the IF passband that is effectively hopping about only intercepts an amount of interference proportional to the IF bandwidth and the *PG* is again the same as for the DS system.

Table I summarizes the characteristics of typical SS systems.

TABLE I – *Typical spread spectrum system*

Name	Type	Required RF bandwidth	Processing gain (dB)	References
RAKE	Direct sequence (FSK)	10 kHz	22	Price and Green, 1958
CHEROKEE	Direct sequence	1 MHz	16	Corneretto, 1961
RACEP	Frequency hopping Time hopping	4 MHz	-	Corneretto, 1961
(Magnavox)	Direct sequence	1 MHz	25	Judge, 1962
GPS	Direct sequence	25 MHz	47	Global positioning system

2.4 *Non-sinusoidal carrier (NSC) system*

In the NSC system, the waveshape of the carrier is designed to provide a large spread of sinusoidal frequencies. Large processing gains can be realized by making the information bandwidth small compared to the radiated bandwidth. Obtaining the spectrum spread in this manner allows the radiated bandwidth to be nearly as wide as the highest frequency component of the radiated signal. The theoretical foundations for these systems have been explored by several authors; most extensively by Harmuth [1981]. An into-the-ground radar system, first proposed by Cook [1960], is currently being used for construction surveys, airborne ice surveys, probing during mining operations, etc.

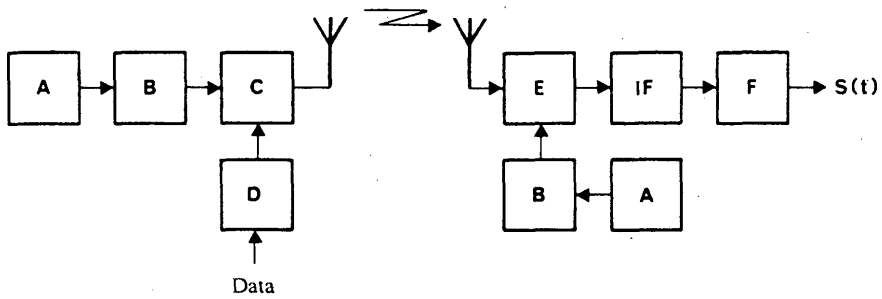
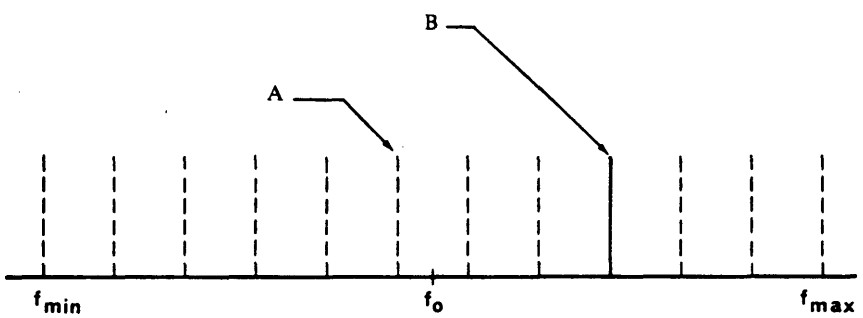


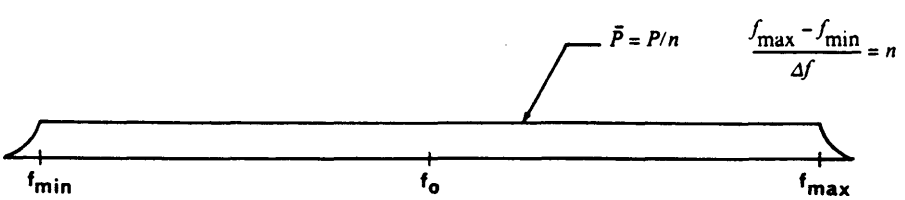
FIGURE 5 — *Basic frequency hopping system*

A: Pseudonoise sequence generator
B: Frequency synthesizer
C: Modulator
D: Binary encoder
E: Mixer
F: Demodulator



(a) *Ideal instantaneous FH spectrum*

A: Alternate frequencies
B: Ideal sinusoidal FH signal at time t of peak power P



(b) *Average power spectral density*

FIGURE 6 — *Frequency hopping spectral representations*

In the literature, the use of non-sinusoidal electromagnetic waves is usually connected with the Walsh functions. This is a system of two-valued functions that has the widely accepted notation wal(ν, t/T), is mathematically simple, and has much similarity to sinusoidal functions. For these reasons, the Walsh functions are preferred by researchers working on non-sinusoidal electromagnetic waves. However, many other two-valued functions as well as sawtooth and other functions have been used [Harmuth, 1977].

The principles involved are explained with the aid of Figs. 7 to 9. A basic block diagram of a spread spectrum radio-frequency transmitter is shown in Fig. 7. The large bandwidth carrier to be radiated is generated by exciting the antenna with two-valued currents as shown in Fig. 8. The trapezoidal antenna current shown on Fig. 8a produces the electromagnetic far-field variation shown in Fig. 8b. The Fourier transform $a(\nu)$ of the field strengths according to Fig. 8b is shown by the solid line in Fig. 9 for a ratio $\Delta T/T = 1/40$ [Harmuth, 1977; Schreiber, 1976]. If the pulse pair is repeated periodically, one obtains the discrete Fourier transform shown by the dashed-dotted lines in Fig. 9. Hence, the sinusoidal components have the frequencies $\nu = kT/40\,\Delta T$, with $k = 1, 3, 5, \ldots$ or $f = \nu/T = k/40\,\Delta T$.

For example, if the first component frequency is chosen to be $f_0 = 100$ MHz, then this calls for a value $\Delta T = 1/40\,f_0 = 0.25$ ns. The second component frequency will then be $f_1 = 3/40\,\Delta T = 300$ MHz, the third will be $f_2 = 500$ MHz, etc. The frequency $\nu = T/\Delta T$ in Fig. 9 becomes $f = 1/\Delta T = 4$ GHz, and the period T in Fig. 8 equals $40\,\Delta T = 10$ ns.

If in Fig. 7 it is assumed that the binary spreading carrier has spread the baseband signal over a band of 100 MHz width, and the non-sinusoidal radio carrier is a periodic sequence of pulses according to Fig. 8a with a period $T = 10$ ns and a transient time $\Delta T = 0.25$ ns, then the radiated signal will be spread over a bandwidth of about 4 GHz located in the band $0 < f < 4$ GHz. The power density has been reduced by nominally 10 log (4 GHz/100 MHz) = 16 dB. However, something more important has also been achieved: the used band is located between 0 and 4 GHz.

One possible application of these principles to spread band sharing can be illustrated by assuming that it is desired to band-share a 100 Hz teletype signal with signals in a band of approximately 1 MHz width. The power density required for the teletype signal can be reduced by approximately 40 dB (10 log ($10^6/10^2$)), permitting the transmission of this signal with a very low probability of causing detectable interference.

Radar applications such as the in-the-ground radar mentioned above arise because the resolution required dictates that emission bandwidths must be used that approach the highest frequency which can be radiated without unacceptable attenuation.

Short-range communications applications have been proposed where the direct transmission of ultra-narrow pulses provides a low probability of interference from digital data to other systems because the signal is rapidly attenuated with distance as well as spectrum spreading over a large band.

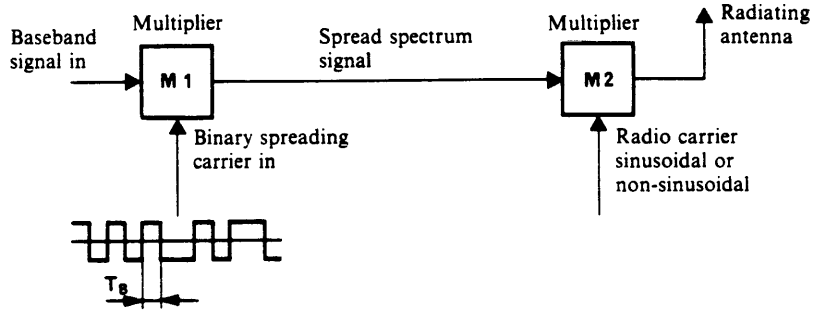


FIGURE 7 — *Frequency spreading of a baseband signal by means of a binary spreading carrier, and radiation of the spread signal by means of a non-sinusoidal radio carrier*

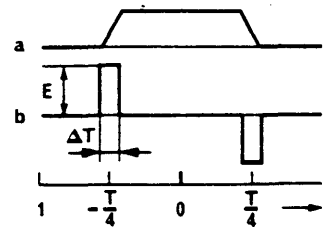


FIGURE 8 — *A two-valued current with finite transition time ΔT (a), and its first derivative (b)*

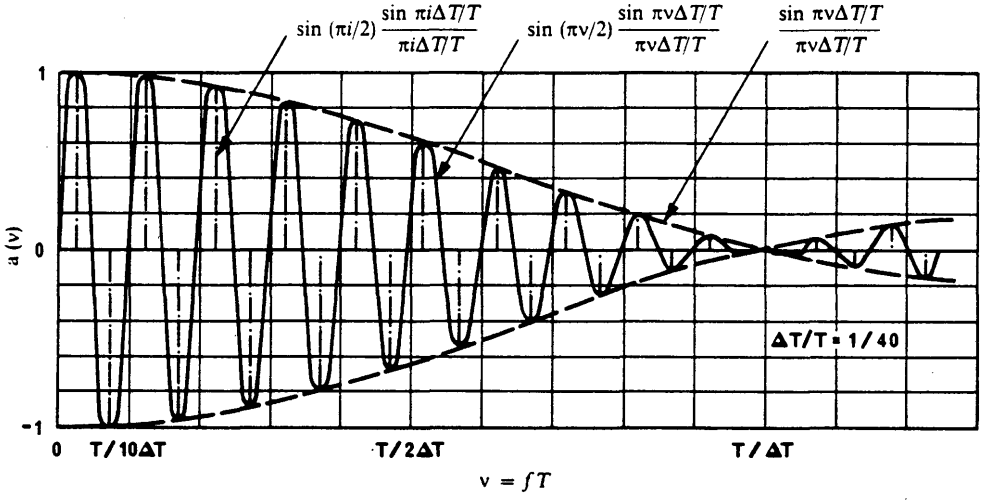


FIGURE 9 — *The Fourier transform of the function of Fig. 8b in the infinite interval (solid line) and the coefficients of the Fourier series of the periodically repeated function of Fig. 8b (dashed-dotted lines)*

3. Signal-to-noise (S/N) performance of DS and FH systems

The (S/N) performance of a linear DS spread spectrum system in the presence of Gaussian noise is described by [Dixon, 1976 and Mayher, 1967] and applies to receiver system noise and external noise with Gaussian characteristics. For this condition, DS performance is given by:

$$PG = \frac{(S/N)_{out}}{(S/N)_{in}} = 2B_{sin}\,T_s \qquad (2)$$

where:

PG: processing gain of the system

$(S/N)_{out}$: output signal-to-noise power ratio

$(S/N)_{in}$: input signal-to-noise power ratio

$2B_{sin}$: bandwidth of RF input signal power density spectrum at first nulls

T_s: time duration of input signal information.

The processing gain (equation (2)) is considered the most important parameter of an SS system. Its implications have been discussed previously.

It is also reasonable to assume that the peak of the code rate signal autocorrelation function at $u = 0$ will have a duration of the order $1/B_{s_{in}}$ (which is also approximately equal to reciprocal of the code rate). The ratio of the duration of the signal information (T_s) to the main peak response is thus given by $T_s/T_{s_{in}}$. Thus the large $B_{s_{in}}T_s$ case affords a "pulse compression" effect whereby the signal energy in a relatively long pulse is "compressed" into a high level short pulse. The result is a high detection probability at the intended receiver with no loss in time resolution.

The same results can also be obtained in the time domain by taking the inverse Fourier transform of the respective functions and using equivalent time operations.

The FH system basically consists of a narrowband filter matched to the information bandwidth pseudo-randomly shifting in frequency over the SS bandwidth. The noise out of the system is therefore governed by the bandwidth of the narrow filter. When an analysis is made of noise or an unwanted interfering signal which occupies the full hopping bandwidth, a decrease in the output unwanted signal power is obtained which is equal to the ratio of the bandwidths. The *PG* for this case is, therefore, equal to:

$$PG = \frac{B_{FH}}{B_s} \tag{3}$$

where:

B_{FH}: FH bandwidth

B_s: wanted information bandwidth.

If the FH system utilizes the same RF bandwidth as the RF bandwidth of the DS system to the first nulls and transmits the same information rate, then B_{FH} and B_s in equation (3) are respectively equal to $2B_{s_{in}}$ and $1/T_s$ in equation (2) so that the processing gain of both systems is the same, neglecting second order effects. It should be noted that this is only for the case of a noise or an unwanted signal spread over the wide bandwidth and not for a narrowband signal. The sensitivity of the FH system does not contain the *PG* improvement of the direct sequence system and is simply proportional to the noise temperature of the system and the information bandwidth.

4. Signal-to-interference performance

The following describes the S/I performance of the DS and the FH SS systems.

For the case in which the bandwidth of the input interfering signal ($B_{I_{in}}$) is less than or equal to the bandwidth of the input wanted signal ($B_{s_{in}}$), the *PG* has been obtained as a function of the off-tuning ($\Delta\omega$) as [Mayher, 1967, equation (3-33), Table I]:

$$\frac{(S/I)_{out}}{(S/I)_{in}} = 2B_{s_{in}}T_s \left[1 \Big/ \frac{\sin\left(\Delta\omega/2B_{s_{in}}\right)}{\Delta\omega/2B_{s_{in}}} \right]^2 ; \quad B_{I_{in}} \leqslant B_{s_{in}} \tag{4}$$

where:

$\Delta\omega$: the radian frequency difference between the carrier of the wanted and unwanted signal;

$B_{I_{in}}$: input bandwidth of the RF interfering signal

and all other terms are as defined in equation (2).

When the bandwidth of the interference is greater than the wanted signal, the *PG* has been obtained as [Mayher, 1967, equation (3-47), Table I]:

$$\frac{(S/I)_{out}}{(S/I)_{in}} = \frac{2\left(\dfrac{B_{I_{in}}}{B_{s_{in}}}\right)\left(T_s B_{s_{in}}\right)}{\left[\dfrac{\sin\left(\Delta\omega/2B_{I_{in}}\right)}{\Delta\omega/2B_{I_{in}}}\right]^2} \tag{5}$$

This clearly shows that the gain is proportional to the bandwidth filtering ratio and the time bandwidth product as would have been expected.

From the point of view of the output power ratios, the wideband SS system overcomes interference to the same degree that it overcomes noise.

In the FH SS system, the frequency hopper reinserts the correct carrier frequency for the wanted signal and mixes it to the IF centre frequency. Any interfering signal, entering at a fixed frequency relative to the centre frequency of the FH SS system, is converted after the FH to a signal of reduced amplitude and random off-tuned frequency due to the FH mixer and IF filter action.

The interfering signal is then filtered so that effectively only those signals that fall within the same frequency channel as the desired signal are transferred through the system as interference. Since there are n possible frequency channels, this happens on an average only $1/n$th of the time.

The interference at the input to the detector is, therefore, similar to a Poisson process. The degradation in the receiver output depends upon the signal processing structure. Since this structure heavily affects the degradation analysis, no simple generalization can be given for required input (S/I) ratios. However, the determination of what degradation results, depends upon the structure of the receiver and the input (S/I) ratio.

If the interference consists of a uniform wideband amplitude modulated (AM) signal that equals the FH bandwidth, the hopper output signal consists approximately of a constant amplitude signal with a random frequency offset term. For this case, the degradation analysis consists of essentially analyzing an unwanted FM signal, reduced in power by the ratio of the bandwidths (equation (3)), against whatever subsequent type of signal processing structure is being employed by the frequency hopper. If the detection structure consists of a frequency modulated (FM) or phase modulated (PM) structure, the *PG* will be similar to that obtained against noise except for low (S/I) ratios where an FM capture effect will be introduced.

If the interference consists of a wideband FM signal that equals in bandwidth the FH bandwidth, the hopper output signal consists of a somewhat random amplitude signal with a random frequency offset. The effect of this signal would be very similar to the noise interference case so that the *PG* is given by equation (3) and the required input (S/I) ratio is the same as required for the Gaussian noise case.

5. Spectrum efficiency

Since SS systems use more bandwidth to transmit a given amount of information than a conventional narrowband system, the question of how efficiently the spectrum is being used should be considered as specified in Study Programme 18B/1. Report 662 indicates that spectrum use efficiency is generally a product of bandwidth, geometric space and time. The problem which therefore requires examination from the viewpoint of spectrum efficiency is how many narrow-band and SS systems can transmit simultaneously in the same frequency band and the same geographical area. This problem has been preliminarily investigated in a paper by [Cooper and Nettleton, 1977]. This paper concluded that although SS is clearly not the best answer to all problems, it may prove to be the only solution for high-density mobile communication problems. Relating to the problem, field tests of FH mobile radio equipment in the frequency band of about 800 MHz with Reed-Solomon coding and soft decision decoding have been carried out in Japan [Mizuno *et al.*, 1985]. The result shows that the reduction of bit errors by the frequency-diversity effect is remarkable, for instance a 17 dB improvement in BER of 10^{-3} in a laboratory experiment by using a fading simulator. Similar results were obtained in a field test in urban areas, and in addition, the irreducible errors which are often encountered on digital transmission in mobile radio communication do not occur. Thus it has been experimentally verified that the FH technique can tolerate significant fading in urban areas, and this feature will give a significant impetus to the design of SS mobile radio systems with high spectral efficiency. The spectrum efficiency issue should be addressed in detail in future reports on SS systems.

6. Summary

SS systems have been defined and descriptions of the DS, FH and NSC techniques have been given. The *PG* has been shown for various interference cases to be proportional to the spread bandwidth divided by the information bandwidth. For many SS systems the *PG* can be a large positive number which allows the operation of the system when the input interference is much greater than the wanted signal. This is the case in which the wanted output information would be lost for a typical narrowband system. The large signal bandwidths required for many SS applications result in low power spectral density signals. These low power spectral density signals potentially cause negligible performance degradation to systems using conventional modulation techniques in the same frequency band.

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REPORT 975

APPLICATION OF SPREAD-SPECTRUM TECHNIQUES TO RADIOLOCATION AND RADIO COMPATIBILITY

(Question 18/1)

(1986)

1. Introduction

Spread-spectrum techniques have been tested successfully in the MF band (700 kHz bandwidth centred on 2 MHz) for high-precision maritime radiolocation applications. At ranges of the order of 800 km a location accuracy by day and night of between 10 and 20 m has been obtained.

These techniques allow full separation of ground-wave signals, the only ones useful for radiolocation, from those produced by ionospheric waves or other natural reflectors, the effects of which have to be completely eliminated.

By using transmissions of relatively very great spectral width (of the order of 30% of the central frequency), it is possible to improve the accuracy and reliability of radiolocation data considerably with unambiguous measurements of group delay and signal phase. The ratio between the spectral width of transmitted signals and the final passband of the processing equipment, sometimes referred to as "correlation processing gain", can have values of the order of 100 million (10^8).

Direct signal coding may be used. The use of maximum length 4-phase modulated pseudo-random binary sequences which change state at given speeds has proved particularly worthwhile and efficient. (Modulation characteristics comparable to those of so-called minimum phase-shift modulation.)

Transmissions require only very low power spectral densities and are produced inside the coastline, several dozen kilometres away from the coast. As a result, the signals are practically undetectable within the maritime area concerned, which means that *a fortiori* they cannot give rise to harmful interference for other spectrum users.

A very careful analysis, which has proved very favourable, has been made of electromagnetic compatibility with the transmissions of the other services using the band.

Note. – See also Reports 651, 826 and 1041 related to some aspects of the same subject.

2. Transmission characteristics applicable to the introduction of wide spectrum transmissions for radiolocation

Many radiolocation or radionavigation systems using transmitters having a terrestrial base-line have been tried in the past. The most commonly used technique has been to evaluate the arrival time of signals, from which the distances covered can be measured by measuring the phase of an unmodulated carrier wave.

This technique has two major difficulties:

(a) phase measurement is totally ambiguous close to $2\ k\pi$, which leads to an ambiguity in the measurement of distances or distance differences at $k\lambda$ or $k\lambda/2$;

(b) the type of signal allows no possibility of time discrimination, so that measurements are fully exposed to the effects of unwanted waves or propagation paths, particularly those produced by ionospheric reflections.

As a partial remedy to the first difficulty, frequency and phase related A_0 waves are transmitted on a number of frequencies. This leads, in practice, to crowding of the spectrum by many transmissions which, though each is emitted over a very narrow spectral width, in fact render a large overall bandwidth unusable for other services, considering the minimum operational passband required by receivers. Depending on the systems, between three and six frequencies are generally used and sometimes up to twelve, without even then completely resolving the ambiguity problem and without in any way alleviating the effects of difficulty (b).

A more radical solution to these two problems has been provided by Loran systems, and in particular by Loran-C. This technique consists of transmitting short pulses with very high peak power. The form and duration of the rise and fall times and the carrier frequency used, are related in a particular way, so that in theory, problems (a) and (b) can be resolved in all cases. In actual fact, ambiguity is sometimes difficult to resolve for long distances, especially at night.

Even though the Loran technique has contributed considerable operational improvements, it remains limited owing to the simplicity of the receivers used and their limited processing capability. Because of these limitations, the system has to transmit with very great power, which necessarily restricts it to a specific frequency

band, since any radio compatibility with other users in the same frequency band is virtually impractical. As a result, the spectral band of Loran transmissions has had to be limited to the absolute minimum, which also means that the system's performance has virtually no room for any future improvement.

The example of the Loran system illustrates very clearly the value of improved receiver processing on systems performance, as far as their operational qualities and capacity to coexist with other services in the same frequency band are concerned. Taking into account the transmitting pulse form factor, the correlation processing gain of Loran receivers (i.e. the ratio between the spectrum width and the final passband used for position data) is of the order of 1000-5000.

For applications that require high positioning accuracy and for important emissions, due to the progress of technology it is possible to increase the complexity of receivers and to considerably improve their processing performance. Thus processing gains of the order of 100 million have been obtained for real practical applications.

This order of magnitude of processing gain opens up spectacular possibilities with regard to radiolocation performance such as:

- elimination of distance measurement ambiguities;
- complete elimination of ionospheric reflection effects;
- a much better definition of ground-wave propagation delay.

A detailed explanation of these possibilities is given elsewhere. The main point to be made here is that these processing gains can open the way to total compatibility with other services using the same frequency band, since the field-strength levels produced by these emissions are equal to or of the same order as those of atmospheric noise in areas of common interest to the services concerned.

3. Transmitting characteristics of the tested system

The first experiments were conducted over a period of more than two years near the west coast of France.

The source of the transmissions was three low-powered transmitters, about 300 km apart. Each transmission was based on a signal of very high frequency stability, whose phase is modulated according to pseudo-random binary sequences. Every binary state corresponds to a phase state, but the speed of transition between states is constant and limited, in a similar way to MPSK techniques. There is no amplitude modulation, and signals are emitted continuously, i.e without pulses.

The sequence generating shift register clock has a frequency of 2/3 MHz. Binary sequences are repetitive, with a periodicity of the order of 25 s. The spectrum of the transmitter signal is therefore practically similar to that of noise, when observed in an analysis bandwidth of a few hertz to a few kilohertz, and its width is of the order of 700 kHz, centred on 2 MHz. The power may vary from a few tens of mW to a few watts and is automatically adjusted. The adjustment is controlled by means of a processor combined with a special receiver, which, on a local and continuous basis, estimates the average level of local atmospheric noise as accurately as possible.

For a given transmitter location, and depending on local ground conductivity, a transfer coefficient may be calculated to establish a proportionality factor relating transmitted power level to observed local noise level [Gounon, 1984]. The normal rules for calculating this factor are such that field-strength levels produced by the transmissions, whether propagated by ground-wave or by ionospheric reflection, are of the same order of magnitude as that of atmospheric noise when the signals reach the geographical area of interest, i.e. the maritime area.

A broadband antenna is used for emission, with omnidirectional horizontal radiation and a gain of around 0 dB.

4. Field-strength levels related to atmospheric noise as a function of distance and frequency

The power radiated by the transmitters is continuously adjusted so as to follow automatically changes in the level of atmospheric noise. For this purpose, the transmitting station is equipped with a frequency scanning receiver and a processor which establishes the distribution function of observed reception levels every 15 min.

If this distribution function is automatically analyzed, it is easy to distinguish spectral components produced by narrowband emissions and to determine what may be attributed to wideband natural noises.

As indicated earlier, transmitter power is determined by applying a proportionality factor to the observed level of atmospheric noise (K_p). However, for the sake of safety, provision is made for setting an upper radiated power limit, which may not be exceeded, even though the noise level rises well above normal values or for any other reason.

This proportionality factor establishes a constant ratio between the power spectral density level of the radiated signal E_s^2 and the spectral density level of natural noise E_n^2.

According to the electrical properties of the local ground and to the distance between the transmitter and the coastline, the value of the proportionality factor K_p is chosen in order to satisfy the following conditions:

- the value of $\frac{E_s^2}{E_n^2}$ derived from ground-wave radiation should be less than or equal to 0 dB at the coastline;
- the value of $\frac{E_s^2}{E_n^2}$ derived from ionospheric reflection should be less than or equal to -4 dB at the point where it reaches a maximum.

For example, for average ground conductivity ($\sigma = 5 \times 10^{-3}$ and $\varepsilon = 13$), using a transmitter located 20 km inside the coastline, the standard values of the quantity 20 log E_s/E_n (dB) are those given in Table I (as a function of the distance between the transmitter and the observation point and as a function of frequency):

TABLE I – *Values of E_s/E_n (dB)*

Distance (km) \ Frequency (MHz)	1.6	1.65	1.7	1.8	1.9-2.1	2.2	2.3	2.35	2.4
0.3	−1	+27	+52	+57	+62	+57	+51	+32	+12
1	−13	+17	+40	+45	+49	+45	+39	+20	0
5	−37	−7	+15	+20	+25	+20	+15	−4	−24
10	−47	−17	+5	+10	+15	+10	+5	−14	−34
20	−63	−33	−10	−5	0	−5	−11	−30	−50

5. Power spectral density levels

For the above type of ground, at a distance of 1 km from the transmitter, the signal level is about -4 dB in relation to free-space propagation. Therefore, at 1 km distance, the E_s/E_n ratio corresponding to free-space propagation $(E_s/E_n)_{fs}$ will be:

$$20 \log \left(\frac{E_s}{E_n}\right)_{fs} = +49 + 4 = +53 \text{ dB}$$

The power radiated by the transmitter P_r is defined by:

$$P_r = \frac{E_n^2}{30} \left(\frac{E_s}{E_n}_{fs}\right)^2 D^2 B$$

where:

D = 1000 m

B = 666 kHz

and for:

$$20 \log \left(\frac{E_s}{E_n}\right)_{fs} = +53 \text{ dB}$$

$$P_r = -15 + 53 + 60 + 28 + 20 \log E_n$$

$$P_r = +126 + 20 \log E_n$$

In terms of the experimental example described, the average value of natural noise levels in the region, as defined in Report 322, will be equivalent to:

- by day: $E_n^2 = -12$ dB(μV/(m · kHz)), or -132 dB(V/(m · kHz))
- by night: $E_n^2 = 0$ dB(μV/(m · kHz)), or -120 dB(V/(m · kHz)).

The values of radiated power and radiated power spectral density will therefore be as follows:

- by day: $P_r = +126 - 132 = -6$ dBW, or 0.25 W

  Spectral density: $\frac{0.25}{666} \simeq 0.00057$ W/kHz
- by night: $P_r = +126 - 120 = +6$ dBW, or 4 W

  Spectral density: $\frac{4}{666} \simeq 0.005$ W/kHz.

6. Practical results

The results obtained with respect to radiolocation performance have been given in [Nard, 1984, 1985; Nard *et al.*, 1984].

Results obtained with respect to mutual compatibility with other services and measurements of radiated field strength have also been given in [Nard, 1985].

The readings obtained by French official services have yielded E_s/E_n values in line with those described in Table I [DTRE, 1985]. Additionally, under the same conditions, the feasibility of using spread-spectrum signals with the characteristics given above for carrying out radiolocation functions has been verified.

7. Conclusions

The trials have shown the great efficiency of spread-spectrum techniques in applications relevant to the radiolocation service.

They have also shown that there can be satisfactory compatibility between existing telecommunication systems and the operation of accurate maritime radiolocation systems using spread-spectrum techniques in the MF band (700 kHz bandwidth centred on 2 MHz).

There is no doubt that the implementation of these techniques can lead to more efficient use of the radio spectrum.

At present there are no clear rules for the convenient management of this kind of transmission. The trials have demonstrated that the operation of a limited number of such systems within the same geographical area is no longer a problem.

Since this type of system is likely to become increasingly widespread, it would seem advisable to study and introduce rules for an equitable sharing of the spectral density level usable within the same geographical area between different spread-spectrum systems.

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REPORT 671-1

SAFETY ASPECTS OF RADIO-FREQUENCY RADIATION

(Question 52/1)

(1978-1986)

1. Introduction

In considering hazards arising from the exposure of human beings, equipment and flammable or explosive materials to electromagnetic radiation, the CCIR should be guided by the recommendations of organizations competent to specify the appropriate limits. These limits enable the areas in which hazardous levels may exist to be identified.

This Report deals with four subjects. Section 2 presents the results of an investigation carried out on laboratory models and in the vicinity of LF, MF and HF transmitters.

Section 3 describes a method of assessing power flux-densities from transmitters operating at frequencies above about 1 GHz.

Section 4 gives information about the impact of electromagnetic fields on persons and the measurement of the incident field strength.

Section 5 presents some suggested design and operational procedures intended to prevent human exposure to hazardous radio-frequency radiation.

2. Hazards arising from induced voltages in metallic structures

The Federal Republic of Germany has investigated a number of technical parameters near long-wave, medium-wave and short-wave transmitters as well as on laboratory models. The structures which acted as receiving antennas in these investigations, had more or less elongated conductor configurations. The effects of electromagnetic radiation on loop-like structures were not investigated.

2.1 *Hazard parameters occurring at motor vehicles*

It is not only the field strength but also the geometric dimensions of a receiving structure and possibly its ambient conditions that may present indirect hazards, i.e. by the induction of dangerous contact voltages in the structure or by the extraction of dangerously high power from it.

2.2 *Symbols used*

C: capacitance,

E: absolute value of field strength,

f: frequency,

h: height of a vertical conductor,

h_{eff}: effective height,

h_M: height of the antenna mast of the radio test vehicle,

h_m: maximum height of a structure capable of receiving electromagnetic waves,

l_m: maximum length of a structure capable of receiving electromagnetic waves,

P_w: real power,

$P_{w\,max}$: maximum extractable real power,

R_L: load resistance,

U_{ind}: induced voltage,

U_K: terminal voltage,

$|\underline{Z}_i|$: absolute value of internal impedance,

λ: wavelength.

2.3 *Effective height and induced voltage*

For receiving antenna installations, the induced voltage is calculated by means of the formula:

$$U_{ind} = h_{eff} \times E \qquad (1)$$

Under given polarization conditions of the RF field, the effective height, h_{eff}, depends exclusively on the characteristics of the structure acting as a receiving antenna. The effective height may therefore be used to assess the extent of the hazard from an antenna-like structure. In practice, the effective height is calculated by the measurement of E and U_{ind} and the insertion of these values in formula (1). Table I contains values found in this way for motor vehicles (see Fig. 1). The values show that the effective height does not exclusively depend on maximum height h_m and maximum length l_m.

TABLE I – *Measured values of h_{eff} for test arrangements according to Fig. 1*

Test arrangement in Fig. 1	Maximum height	Maximum length	Orientation in relation to transmitter	
			Radial	Tangential
No.	h_m (m)	l_m (m)	h_{eff} (m)	h_{eff} (m)
1a	2.5	5.2	0.6	–
1b	10.00	5.2	1.15	–
2	3.5	7.8	1.45	1.30
3	2.5	7.2	0.88	0.75
4	3.5	16.0	1.40	1.05
5	3.5	22.0	1.52	1.12

f = 971 kHz

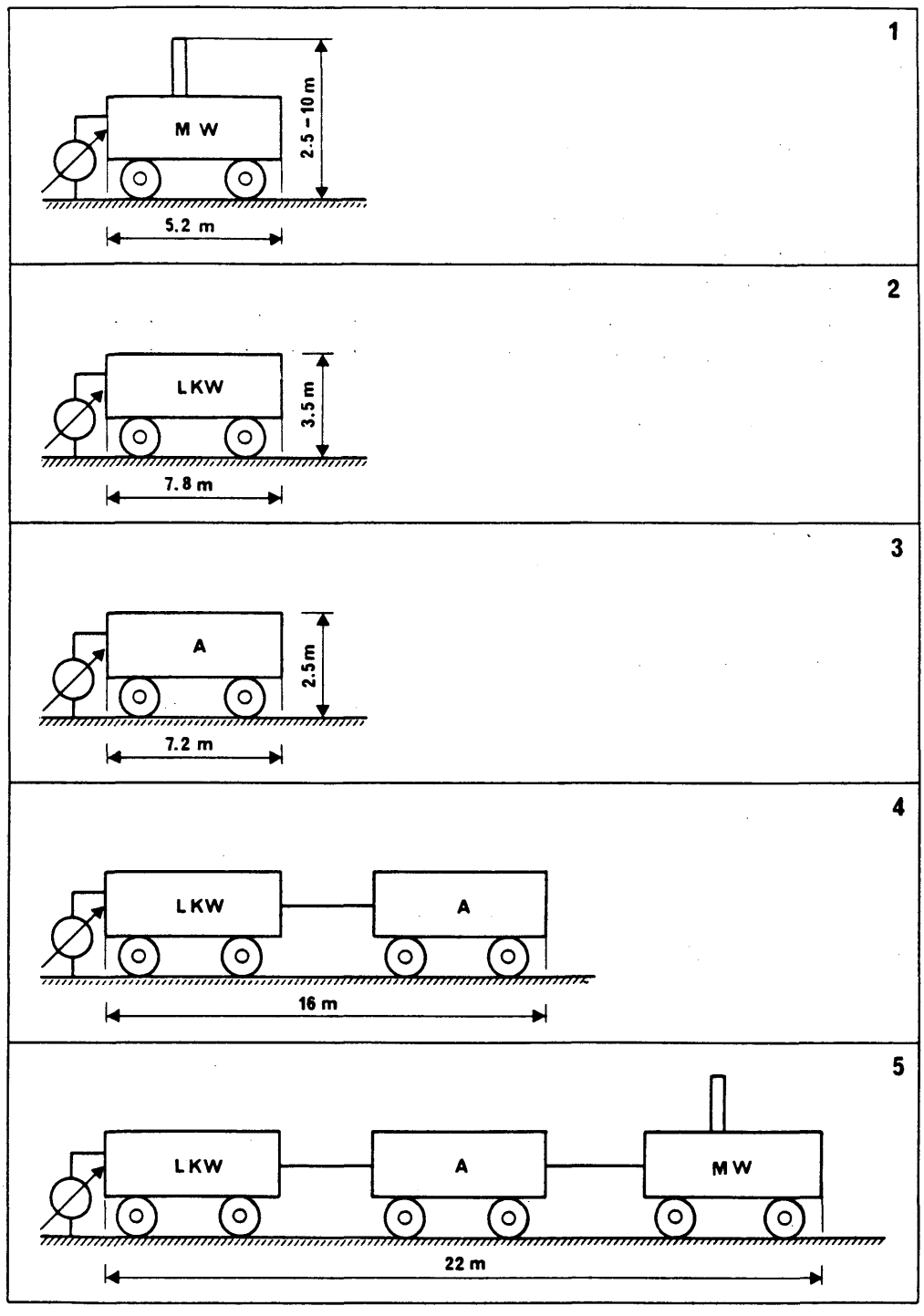


FIGURE 1 — *Types and combinations of vehicles used in the tests*

MW: radio test vehicle
LKW: lorry
A: trailer

2.4 *Internal impedance and extractable real power*

The extent of the hazard results primarily from the maximum real power, $P_{w\,max}$, that can be extracted. It is this real power which determines whether contact with the antenna-like structure will lead to a dangerously high current or to sparking with an energy high enough to ignite explosive mixtures. In addition to the induced voltage, the internal impedance of the antenna-like structure, which can be regarded as a generator, plays an important role. $P_{w\,max}$ is extracted from the antenna-like structure when the real power,

$$P_w = \frac{U_K^2}{R_L} \tag{2}$$

dissipated in load resistance R_L, reaches its maximum. This is known to be the case if the absolute value of the internal resistance

$$|\underline{Z}_i| = R_L \tag{3}$$

$|\underline{Z}_i|$ can be measured by determining the maximum value of U_K^2/R_L from measurements of terminal voltage U_K at different values of load resistance R_L.

In Fig. 2 the measured values of U_K as a function of R_L are shown for a radio test vehicle with antenna masts of different heights. These values were substituted in equation (2) to calculate the curves for P_w. The maximum ordinate values and the associated abscissa values of these curves yield $P_{w\,max}$ and $|\underline{Z}_i|$, respectively.

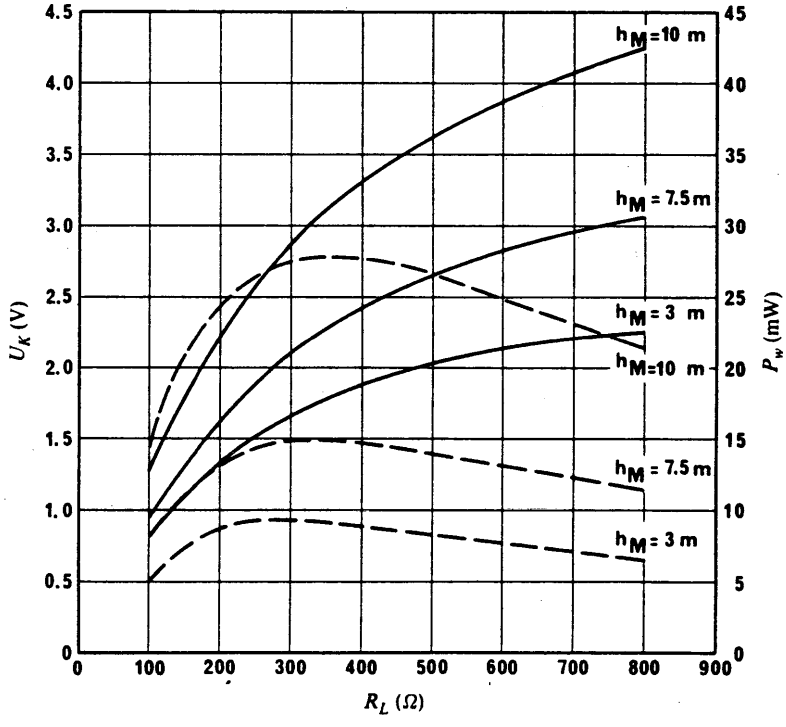


FIGURE 2 — *Terminal voltage U_K and real power P_w dissipated in load resistance R_L for a radio test vehicle*

f = 971 kHz, E = 5.2 V/m

——— U_K

— — — — P_w

Another governing factor of the maximum extractable real power and internal resistance is the ground conductivity prevailing at the place of the antenna-like structure. Figure 3 represents the results of measurements performed on a passenger car under conditions of relatively low natural ground conductivity and with extremely high ground conductivity, the latter simulated by means of a wire netting spread over the ground. In the case of the low natural conductivity, it is remarkable to note that the maximum extractable real power depends on the angle at which the car is oriented to the transmitter. The same effect was found to exist for the effective height with different combinations of vehicles (compare Table I).

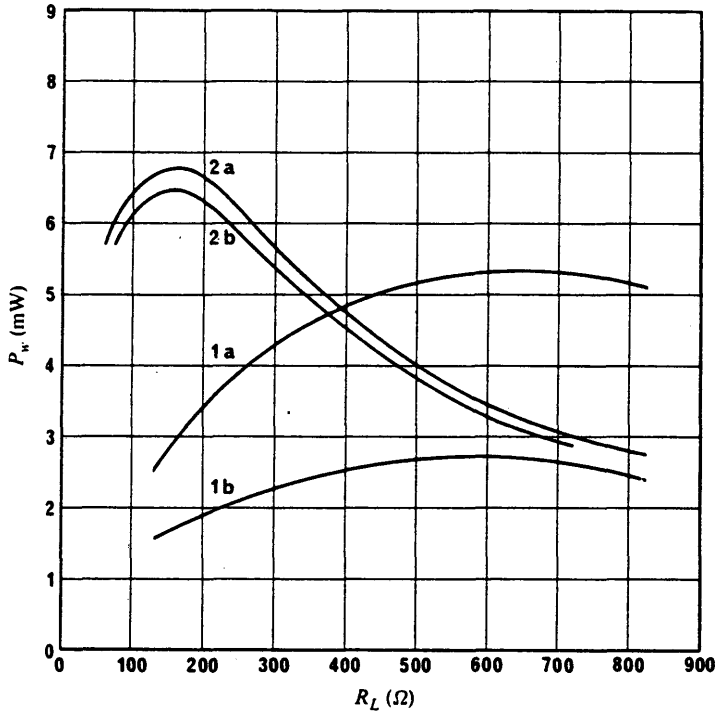


FIGURE 3 — *Influence of the ground conductivity on the extractable real power and internal resistance in the case of a passenger car*

f = 1268 kHz, E = 8 V/m

1: natural ground conductivity $\sigma < 1$ mS/m

2: simulated ground conductivity $\sigma = \infty$

a: car aligned radially to transmitter

b: car aligned tangentially to transmitter

2.5 *Dependence of the hazard parameters on frequency*

Figure 4 shows hazard parameters h_{eff}, $|\underline{Z}_i|$ and $P_{w\,max}$ for a radio test vehicle, measured at different frequencies. In addition, the absolute value of the capacitive reactance of an 800 pF capacitor is plotted in a dashed line. It can be seen that, at least at low frequencies, the internal impedance of a car body acting like a receiving antenna resembles practically that of the capacitance. The deviations occurring at higher frequencies indicate that, compared to the capacitive component of the internal reactance, the component deriving from the earth resistance is too great to be neglected.

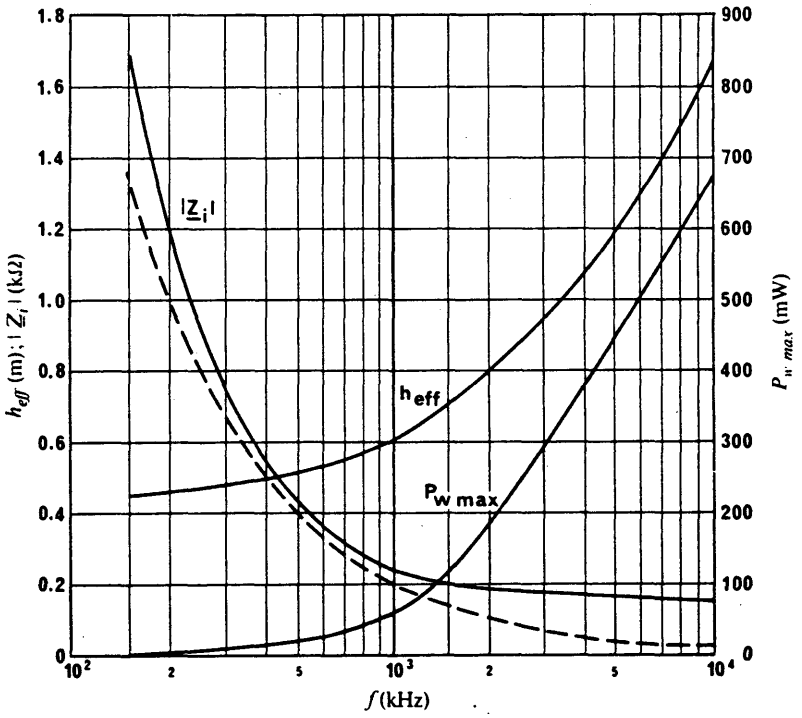


FIGURE 4 — *Dependence of effective height, internal impedance and maximum extractable real power on frequency in the case of a radio test vehicle*

E = 10 V/m

——— measured values

— — — — values calculated for C = 800 pF

Models built on a scale of 1 : 150 were used to investigate the effects of frequency on the hazard parameters in the resonant range of the receiving structure. Figure 5 gives the results obtained for a vertical conductor insulated from earth. Since the reactive components are compensated for at a conductor length of $\lambda/4$, the internal resistance is reduced to a minimum and the real power extractable from the conducting structure is by two orders of magnitude higher than for a conductor length of $\lambda/10$.

2.6 *Significance of the test results for the potential hazard presented by electromagnetic radiation*

The described tests have revealed that in the LF and MF ranges the effective height to be expected for individual motor vehicles or combinations thereof is not in excess of 2 m. This means that, in the case of a reference field strength of 10 V/m, RF voltages above 20 V do not occur. The internal resistance is several hundred ohms so that the dissipated real power is 1 watt maximum, again referred to a field strength of 10 V/m. In the HF range, resonance phenomena must be taken into account to an increasing extent. In general, this applies to any frequency if the length of the conducting structure concerned is about $\lambda/4$. The real power extractable may then be some orders of magnitude higher than that of an electrically shorter structure. The ground conductivity is of great influence on the maximum extractable real power, which also depends largely on the angle at which the conducting structure involved is oriented in relation to the direction of the wave-front.

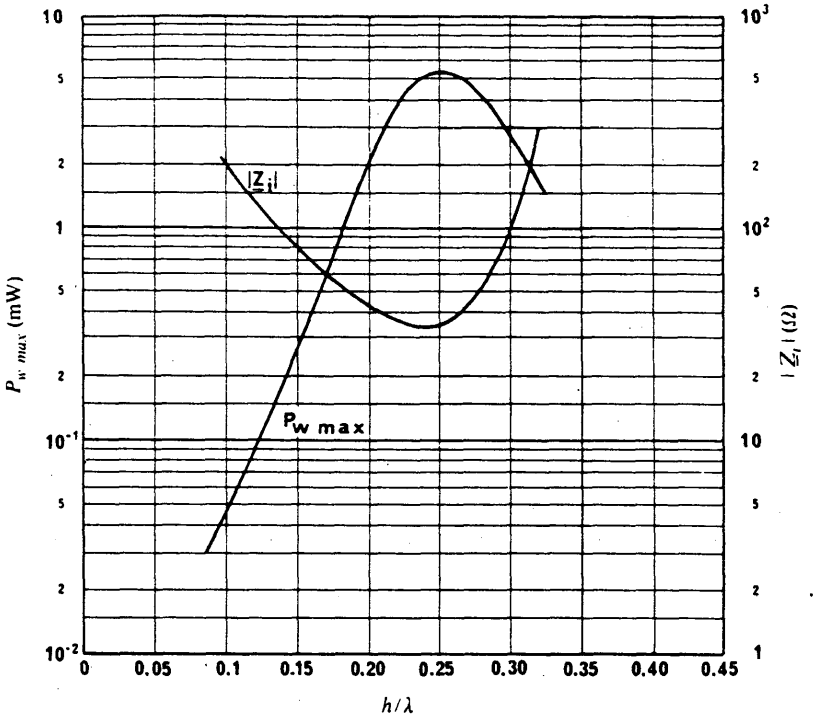


FIGURE 5 — *Maximum extractable real power and internal resistance for general models of vertical conductors differing in height*

3. Radio-frequency radiation hazard at frequencies above about 1 GHz

This section of the Report provides a method of assessing the power flux densities to be expected from transmitting stations which employ antennas having circular apertures and is therefore applicable mainly to systems operating at frequencies above about 1 GHz. The method is limited to on-axis power flux densities and is based on theoretical derivations. Adequate safety factors are incorporated to allow its application to the determination of hazard distances.

Although derived for circular apertures the method given in this Report may also be applied to non-circular apertures without serious error by assuming an effective circular aperture equal in area to the actual aperture.

3.1 *Method for calculating power flux density*

The mathematical derivation of the power flux densities on the axis of a circular transmitting aperture [Bickmore and Hansen, 1959; Hansen, 1975; Hansen 1976] has been simplified into the consideration of two zones. It should be noted that the method is applicable only to on-axis distances exceeding the diameter of the aperture.

In the first zone the power flux density is assumed to be constant and equal to the maximum value of the power flux density that can occur. The maximum on-axis power flux density will depend upon the distribution of aperture illumination and for the purpose of deriving the formulae below a distribution which results in a ratio of main beam to first side-lobe peak power flux density of 35 dB has been assumed. This side-lobe ratio is typically achieved with a taper-on-pedestal form of illumination and higher power flux densities are unlikely to result from other forms of aperture illumination in current practice. In the second zone, it has been assumed that the power flux density varies inversely as the square of the distance from the aperture, and that uniform aperture illumination which gives the highest power flux density in this zone is used.

In order to make some allowance for the addition of direct and reflected signals which could arise in the practical situation, the theoretically derived power flux densities in both zones have been multiplied by a factor of four.

3.2 *Method*

The method assumes two zones, one in which the power flux density is considered to be constant and the other in which an inverse square law applies. The formulae for these zones, based on the assumptions given in the preceding section, are:

$$pfd = \frac{39.3\ P}{D^2} \quad \text{for} \quad D < d < d_1 \text{ (for the first zone)}$$

$$pfd = \frac{\pi P D^2}{\lambda^2 d^2} \quad \text{for} \quad d > d_1 \text{ (for the second zone)}$$

where $d_1 = 0.283 \frac{D^2}{\lambda} = 0.94\, fD^2$

and pfd: power flux density (W/m^2)

P: power input to the antenna (W)

D: physical diameter of antenna aperture (m)

λ: wavelength (m)

f: frequency (GHz)

d: on-axis distance from antenna (m)

d_1: the distance in metres at which the pfd is assumed to change from a constant value to an inverse square law.

The formulae above and the nomograms given later assume an antenna efficiency of 100%. The derived values of power flux density may be multiplied by the actual efficiency of the antenna, if this is known. For multiple feeds at different frequencies and/or polarizations the input power should be treated as the sum of the individual input powers.

The formulae are presented in the form of nomograms in Figs. 6, 7 and 8 and these should be used as follows:

To determine the distance for a specified power flux density Fig. 6 should be used to determine whether or not a specified power flux density limit may be exceeded in the first zone. If not, then it may be assumed that this limit is not exceeded anywhere but in the immediate vicinity of the antenna. If Fig. 6 indicates that the limit may be exceeded, then Fig. 8 should be used to determine the distance within which the power flux density can be greater than the specified limit.

To determine the power flux density at a given distance Fig. 7 should first be used to determine whether the distance is greater or less than the change-over distance d_1. If the distance is less than d_1, Fig. 6 should be used to determine the power flux density. If the distance is greater than d_1 then Fig. 8 should be used.

3.3 *Limitations of method and practical interpretation*

There will be a number of practical situations for which the method may not be strictly applicable or when it may significantly overestimate the hazard. For example in the case of line-of-sight radio-relay transmitters it may be unlikely that there will be any efficient reflecting surfaces within the solid angle illuminated by the main beam. It is therefore suggested that in individual cases in which the method is found to produce unrealistically large hazard distances the results should be treated only as a guide and the particular situation should be examined in greater detail.

At distances greater than d_1, the off-axis power flux density may be assessed by applying to the on-axis power flux density at the equivalent distance a correction for the radiation pattern of the antenna.

The only areas where the off-axis power flux density may exceed the on-axis level is near the feed point and in the case of a Cassegrain antenna, in the space between the primary and secondary reflectors. Within the volume contained in an imaginary cylinder based on the aperture and of length equal to the aperture diameter, the distribution of power flux density depends on the feed and reflector arrangement and does not lend itself to a general theoretical treatment. In this region, the greatest hazard is found at the aperture of the feed where the average power flux density is given by dividing the power fed into the antenna by the area of the feed aperture. The peak power flux density, generally at the centre of the feed, may be up to three times the average value.

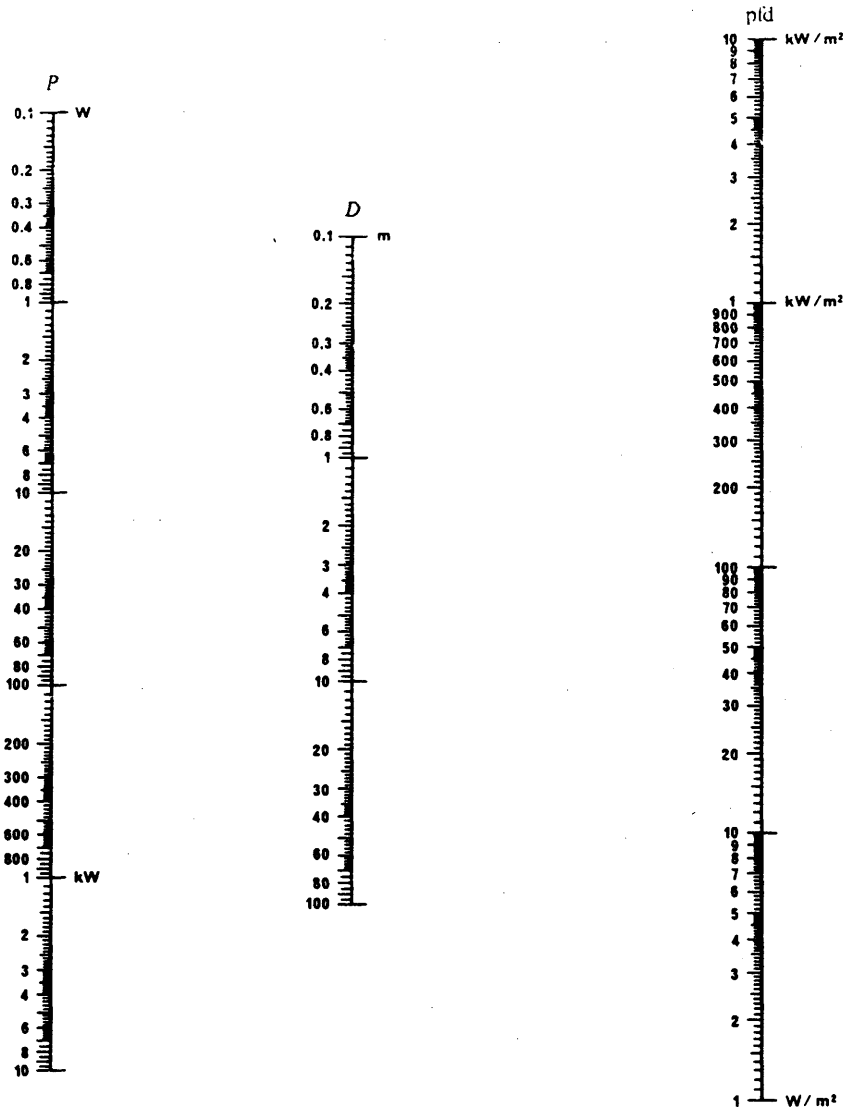


FIGURE 6 — *Power flux-density in the first zone*

The nomogram of Fig. 6 is based on the formula for the on-axis peak power flux-density:

$$\text{pfd} = \frac{39.3\ P}{D^2}$$

Symbols: pfd: on-axis power flux-density (W/m^2)

P: input power to antenna (W)

D: physical diameter of antenna (m)

Use: Join the appropriate points on the P and D scales with a straightedge, and the point of intersection of the straightedge with the pfd axis gives the peak power flux-density.

Note. — The value of power flux-density obtained may be multiplied by the efficiency of the antenna (typically between 0.5 and 0.8) if this is known. The derived value will also include a factor of 4 to allow for the addition of direct and reflected rays. In cases where it can be safely assumed that no reflection is possible on or near the axis of the antenna, this factor may be considered to be unwarranted in an assessment of the actual hazard.

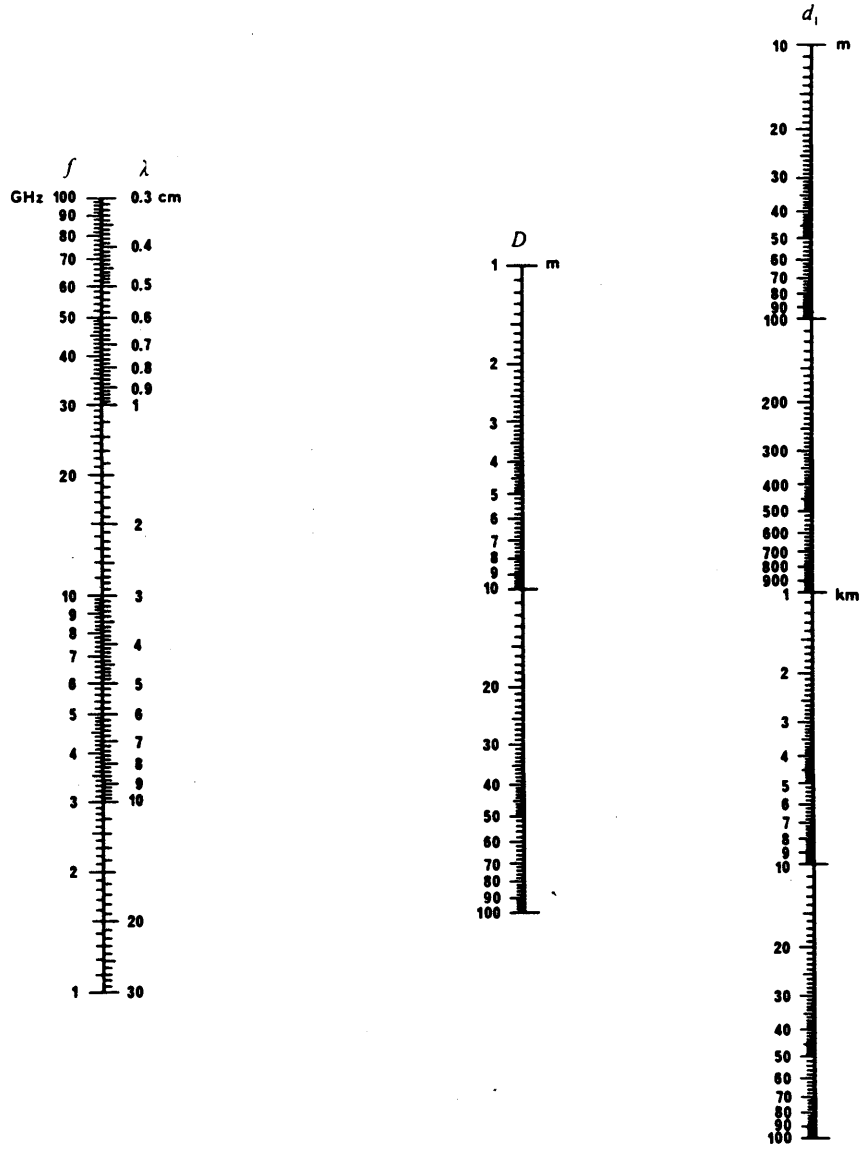


FIGURE 7 — *Change-over distance* d_1

The nomogram of Fig. 7 is based on the formula $d_1 = 0.283 \frac{D^2}{\lambda} = 0.94 f D^2$

Symbols: d_1: the distance in metres at which the power flux-density is assumed to change from a constant value to an inverse square law

D: physical diameter of antenna (m)

f: frequency (GHz)

λ: wavelength (m)

Use: Join the appropriate points on the f and D scales with a straightedge, and the point of intersection of the straightedge with the d_1 axis give the change-over distance.

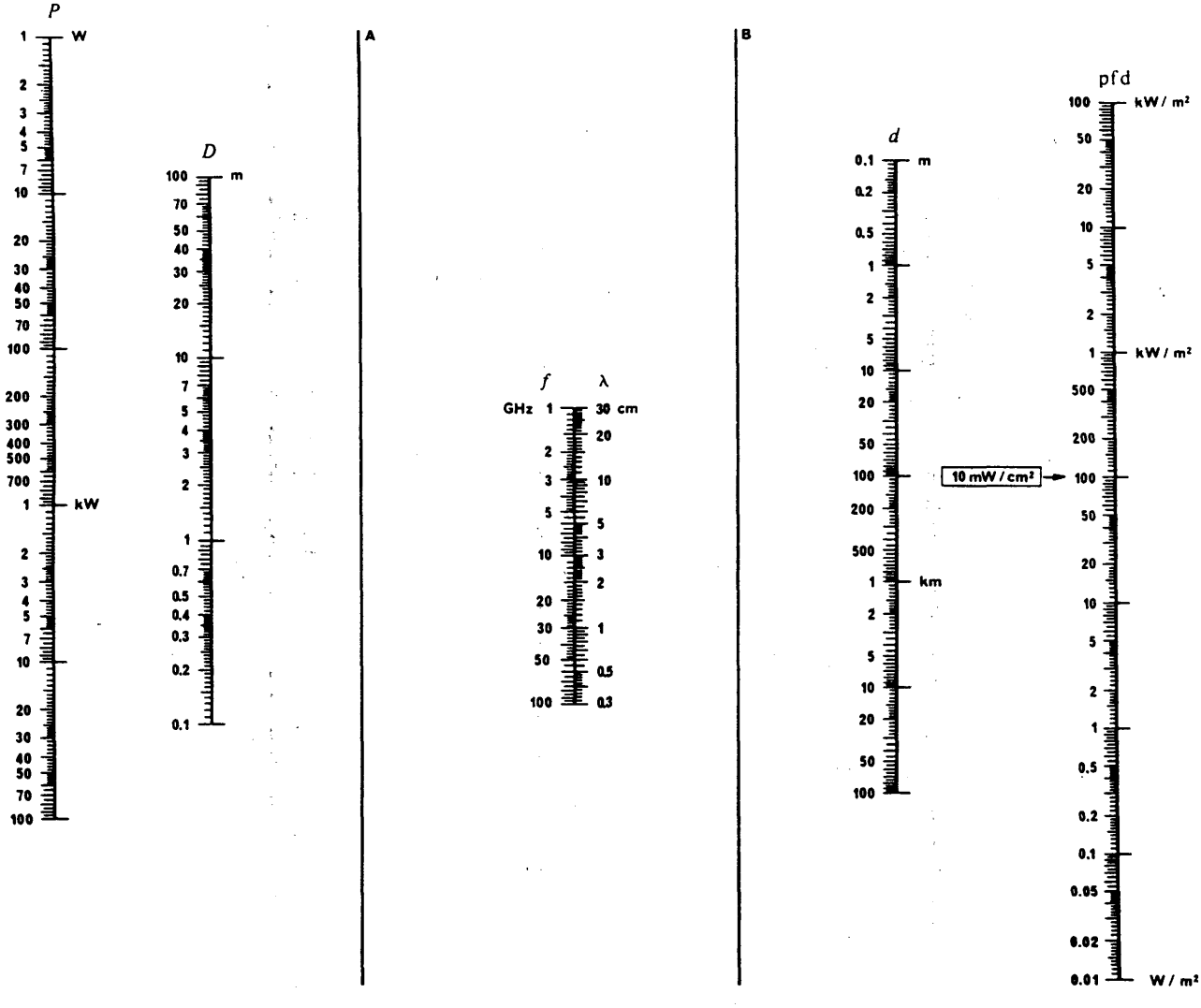


FIGURE 8 — *Power flux-density in the second zone*

The nomogram of Fig. 8 gives the power flux-density in the second zone on the axis of a uniformly illuminated circular transmitting aperture and includes a safety factor of four to allow for reflected signals.

The nomogram is derived from the formula $\text{pfd} = \frac{\pi PD^2}{\lambda^2 d^2}$

Symbols: pfd: power flux-density (W/m²)

P: input power to antenna (W)

D: physical diameter of antenna (m)

d: on-axis distance from antenna (m)

λ: wavelength (m)

Use: Join points on the P and D scales with a straightedge to produce a point on line A. The point on line A is then joined through the appropriate point on the f axis to produce a point on line B. This point on line B is used with either a point on the d axis to give the pfd or a point on the pfd axis to give the distance.

Stray radiation fields outside the collimated tubular beam emerging from the aperture may comprise some 15% of the total power radiated. The largest contributions to the stray fields are spillover past the main and sub-reflectors and scatter from blockage elements such as the feed and/or sub-reflector supports. Diffraction effects to the rear of a solid main reflector are so small in magnitude as to involve no hazard.

In particular cases where the method is not fully applicable (such as off the axis of the main beam in the near zone, or close to the feed) it is recommended that measurements of the power flux density should be made.

4. Direct impact of electromagnetic fields on human beings

In the Federal Republic of Germany, the thermal effect of electromagnetic fields on human beings has been considered and it is thought that the absorption of energy from the electromagnetic field by the whole body or by parts of it, may lead to overheating and consequent health hazards. The specific absorption rate (SAR) for the absorbed power per kilo of body weight (W/kg) has proved to be a useful approach to determine whether hazards are involved [Durney, 1980; Gandhi, 1980].

Various researchers have shown that specific absorption depends on the geometric characteristics of the part of the body in question, the wavelength and the polarization of the field (for an overview, see [Gandhi, 1979 and 1980; Bernhardt *et al.*, 1983]).

Many of the investigations that have been carried out apply only to the far field region. In the case of far fields, it is sufficient to measure the electric field or the magnetic field in order to determine the power flux-density. Fields with an intensity which may lead to health hazards can in principle only be found in the vicinity of transmitting stations. Due to the complex field structure in the near field region, there is no readily usable method for determining the power flux-density. Thus, the two field components E and H, in three space directions, must be considered when evaluating possible health hazards.

The root-sum-square field strengths E_f and H_f are useful in this context because they are a measure of the maximum electromagnetic energy $\left(\frac{1}{2}(\varepsilon E^2 + \mu H^2)\ \frac{\text{W/s}}{\text{Volume}}\right)$ available in the field (where ε is the permittivity and μ is the permeability of free space).

$$E_f = \sqrt{E_x^2 + E_y^2 + E_z^2} \qquad \text{(V/m)}$$

$$H_f = \sqrt{H_x^2 + H_y^2 + H_z^2} \qquad \text{(A/m)}$$

E_x, E_y, E_z, H_x, H_y, H_z are the root-mean-square values of the amplitudes of the E- and H-field in three mutually orthogonal directions, taking modulation into account.

5. Operational precautions to prevent exposure to hazardous radiation

Personnel should not gain access to locations where radio transmitting stations produce hazardous electromagnetic fields. The following suggested design and operational procedures are intended to minimize the hazard from exposure to radio-frequency radiation.

5.1 The radio frequency power flux-densities or field strength that can be generated by a transmitting station should be calculated. If the results indicate that a hazard level may exist in an area of interest, a measurement of the power flux-density or the field strength especially in the near field region will establish the actual hazard in that area.

5.2 Metal supporting masts or other metallic structure (e.g. framework, window frames) must be grounded if dangerous RF-voltages due to induction are anticipated. The length of the grounding wire must be short compared with the wavelength of the incident electromagnetic field. Care must be taken in the use of metal tools because of the risk of electric shocks caused by induced voltages. Flammable substances or explosives should not be stored in the vicinity of a transmitting station. The nozzles of tankers and the filling holes of storage tanks should be short-circuited in advance, so that no RF voltages can occur while petroleum products or other combustible liquids, gases or dusts are being introduced.

5.3 Personnel must be restricted from gaining access to transmitting equipment or antenna structures when dangerous fields may exist. Examples of possible precautions are:

- entrance restrictions (signs, bars, etc.),
- temporary reduction of transmitter-power,
- shielding.

The procedures to be employed will depend upon the nature of the radio station.

5.4 Access to elevated platforms, maintenance towers, etc., in hazardous areas should be controlled by interlocked mechanisms. In the case of mobile platforms which cannot be interlocked with the transmitter controls, procedural instructions should prohibit use of the platform unless a safe indication is given by means of a warning system which is interlocked with the transmitter controls.

5.5 In the case of steerable antennas, hazardous radiation may be avoided by automatic restriction of the transmitted power whenever the elevation of the antenna beam is below a pre-determined angle (e.g. 3° above the visible horizon) or whenever the azimuth bearing of the beam at higher angles could place humans at risk. The control devices for restricting the power should be "fail-safe".

5.6 Waveguides should be identified if the power flux that would be radiated if they were cut could reach a dangerous level. (Desirably, the identification should be by an internationally agreed colour code.) Prominent warning notices should be displayed in rooms where such waveguides are present to stress the danger of looking into open-ended waveguides. All flanged joints in high power waveguides should be checked periodically for RF leakage.

6. National standards on radiation hazards

Several countries have fixed limits for non-ionizing radiation hazards. These standards are derived from theoretical and experimental considerations: values of power flux densities or field strengths used as limits in various countries and further details, for example with regard to exposure time and the work being done in this area by other international organizations, such as the World Health Organization, are contained in [CCIR, 1974-78].

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REPORT 1117

DETERMINATION OF THE ELECTRIC FIELD STRENGTH IN THE NEAR FIELD ZONE OF HIGH-POWER LF/MF ANTENNAS

(Question 52/1)

(1990)

1. Introduction

The increase of the radiated powers in the lower frequency bands (LF, MF, HF and even VHF and UHF) during the last three decades has called for the study of the harmful effects on the human body of the radio-frequency energy in these frequency bands. The accurate determination of the near field-strength in the vicinity of high power LF/MF antennas is necessary to allow administrations to assess any possible radiation hazard.

In contrast to microwave frequencies, where the electromagnetic field has the characteristics of the field in the far zone even at very short distances from the radiation source, and where the concept of the radiated power density (Poynting vector intensity) is very useful in defining the protection safety standards, in the LF/MF frequency bands the field in the antenna vicinity is very complex. In fact, in the near field zone, the simple relationship between the electric and magnetic fields no longer exists: the two fields are not in phase and their ratio is not 377 ohms.

The purpose of this Report (CCIR, 1986-1990) is to provide one analysis for determining the electric field strength in the near field zone of MF and LF mast antennas. This zone is of interest because the associated electric field strengths, which are dependant upon the electrical length of the mast antenna, can have large values. The procedure is derived from theoretical approach based on the Hallen's integral equation for the antenna current distribution function, the integral equation being solved numerically by the point-matching method. The procedure is limited to the simple mast antenna. Caution should be used in trying to apply the procedure in the complex zones, where the physical supporting structure is complicated, when long guy wires are close, and where extensive ground systems are used.

Although in this Report the electric field is found, the magnetic field also can be determined by using the same method. However, it is to be noted that the high magnetic fields near the base of a mast antenna can be of great concern in the presence of large inductors which comprise tuning or matching networks, and which are not part of the present analysis. The base region of a practical monopole antenna is usually not geometrically simple. The base capacitance and the currents

flowing in the feeder system can modify the currents and therefore the fields substantially. As a result of this situation, both point-matching and moment-method solutions (and probably other methods) are useful for currents in the radiating elements, but are less precise for prediction of base impedance or base region currents.

After a brief outline of the theory, which is presented in sections 2 and 3, some measured results, concerning the electric field strengths in the close vicinity of the mast antenna of a high-power MF transmitter (Beograd 1) are given in section 4. These results are compared with the theoretical ones.

2. Field strength in the near zone of mast antennas: theoretical approach

The most common type of transmitting antenna for use in the medium frequency band comprises a vertical steel mast, insulated and driven at its base and kept in an upright position by means of guys. The height of masts is between 0.1 and 0.65 of the transmission wavelength, depending on the desired vertical radiation pattern and the radiated power.

In analyzing the electromagnetic field of the antenna, the influence of the stays can be neglected if they are divided into sections much shorter that the wavelength, by means of special insulators. A further simplification, concerning the ground, can be introduced: with a properly designed ground system, consisting of at least 90 radially buried wires, 0.4 - 0.5 wavelength long, the losses in the earth can be neglected and the ground can be considered as a perfectly conducting plane.

Therefore, a typical MF transmitting antenna can be treated as a cylindrical monopole above a perfectly conducting plane. The masts having triangular or square cross section can be replaced by the equivalent circular cylinder of radius a.

According to the equivalence theorem, the problem of the monopole antenna above the perfect conducting plane and fed at the base by a voltage V, is equivalent to the problem of a symmetric dipole antenna (Fig. 1), fed by a voltage 2V. The leg of the dipole has the same geometrical dimensions as the monopole.

Supposing that the antenna conductor is perfect, an exact solution to the problem of the current distribution along the antenna conductor and, consequently, of the electromagnetic field in the near far zone, can be obtained by solving the well known Hallen's integral equation [2]

$$\int_{-h}^{+h} I(z')\, K(z,z')\, dz' = C \cos kz + (2V/j60) \sin k|z| \quad , \quad (1)$$

which must be satisfied on the dipole surface.

In this equation $I(z')$ is the unknown current distribution function along the antenna axis. $K(z,z') = \exp(-jkR)/R$ is the kernel of the integral equation, where $R = \{a^2 + (z-z')^2\}^{1/2}$ is the distance of the field point z on the antenna surface from the current element $I(z')dz'$ on the antenna axis; h is the height of the monopole; and a its equivalent radius. $k = 2\pi/\lambda$ is the free space propagation constant and C is the unknown integration constant, which will be determined in the process of solving the equation (1).

The integral equation (1) cannot be solved directly and a solution for $I(z)$ in a closed analytical form is not known. However, an approximate solution can be obtained by using one of the known numerical methods. In the present analysis the so-called point matching method, which is the simplest variant of the moment method [3], has been used. According to this method, the current distribution function $I(z)$ is approximated by a functional series of convenient functions with unknown complex coefficients. These coefficients are then determined by stipulating that, with this series substituted for $I(z)$, equation (1) be satisfied at a sufficient number of points along the antenna.
In Ref. [1], the basis for this Report, a simple functional series in the form of a low order polynomial was used [4]

$$I(z') = \sum_{m=0}^{M} A_m (z'/h)^m \qquad 0 \leq z \leq h \qquad (2)$$

with $I(h)=0$.
Substituting (2) into equation (1) and applying the point matching procedure, a system of linear equations with the unknown coefficients $A_0, A_1, \ldots, A_M$ is obtained.
By solving this system for A_m, the current distribution function is solved.

With the known function $I(z)$ the field strength in the near and far zone of the monopole (dipole) can be easily calculated. The input impedance may also be characterized, but with lesser accuracy.

The input admittance of the monopole, the real part of which, G, is necessary for calculating the field strength as function of the input power, is

$$Y = G + jB = I(0)/U \qquad (3)$$

Since we are particularly interested in the electric field strength at the surface of the earth (the plane z=0), only the E_z component of the field will be calculated. Note that electric and magnetic field in this region are not orthogonal to the direction of propagation. In the field point M with the cylindrical coordinates r, z (see Figure 1)

$$E_z = - j \frac{\omega}{k^2} \left\{ \frac{\partial^2 A_z}{\partial z^2} + k^2 A_z \right\} \qquad (4)$$

where A_z is the magnetic vector-potential,

$$A_z = (\mu_o/4\pi) \int_{-h}^{h} I(z') \, K(z,z') \, dz' \quad ,$$

$$K(z,z') = \exp(-jkR)/R \quad \text{and} \quad R = \{r^2+(z-z')^2\}^{1/2} \quad .$$

ω is the angular frequency and μ_o the magnetic permeability of the free space . On the surface of the earth z=0 .

By developing (3) and introducing the input power of the monopole ($U=\sqrt{2P/G}$), the expression for the field strength E_z at the ground surface is obtained. It is the function of the following parameters: h/λ, r/λ, h/a and λ .

In order to have the results for E_z in a generalized form, which does not contain the wavelength λ explicitly but does only the ratios of the geometrical dimensions to the wavelength, we shall consider the modulus of the product $E_z \cdot \lambda$,

$$|E_z| \, \lambda = f(h/\lambda, \; r/\lambda \; , \; h/a) \quad .$$

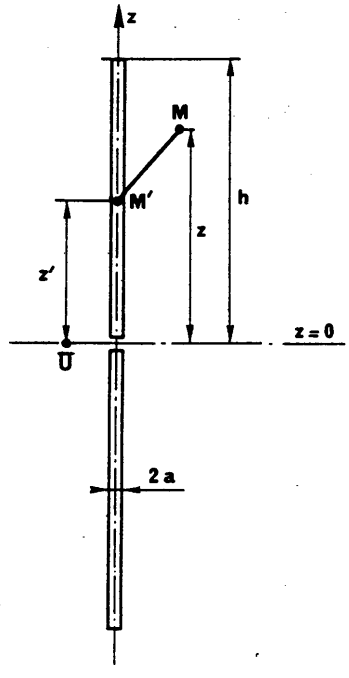


FIGURE 1

3. NUMERICAL RESULTS

The present theory was applied for computing the electrical field strength at the surface of the earth in the near field zone of MF monopole antennas, having different normalized heights, h/λ , and cross-section equivalent radii, a/λ .

Since the ratio h/a (the slenderness of the antenna) at typical MF mast antennas has the values between 150 and 250, first the influence of this ratio on the field strength in the near zone was investigated. In contrast to the antenna input impedance/admittance, which is rather sensitive to the ratio h/a, it has been found that the ratio h/a , in the mentioned limits, has a very slight influence on the field strength in the near zone. That was the reason why only the results referring to the antennas having slenderness $h/a = 200$ have been presented here.

Figs. 2 and 3 show dependence of the product $E\cdot\lambda$ (E is the electrical field strength on the surface of the earth at a distance r from the antenna, λ is the wave-length) versus the normalized distance r/λ , for the five normalized antenna heights: $h/\lambda = 0.25;\ 0.5;\ 0.525;\ 0.55;\ 0.625$. Since the product $E\cdot\lambda$ varies between the large limits, the diagram is divided in the two parts, covering the ranges $0.004 \leq r/\lambda \leq 0.05$ (Fig.2) and $0.05 \leq r/\lambda \leq 1$ (Fig.3). The field strength, i.e. the product $E\cdot\lambda$ is calculated for the input power of 1 kW; for other powers the values on the ordinate axis should be multiplied by $P(kW)^{1/2}$. It is interesting to note that the diagrams reffering to the antenna heights $h/\lambda = 0.5;\ 0.525;\ 0.55$ practically coincide.

Starting from the previous numerical results a very convenient analytical expression for the field strength can be developed. It has the form of the rational fraction

$$E = P^{1/2} \left\{ \sum_{m=0}^{M} B_m \cdot (r/\lambda)^m \right\}^{-1} / \lambda \quad .$$

The coefficients B_m for the different normalized antenna heights, h/λ , are presented in Table 1. ——The preceding formula enables easy and precise calculation of the field strength. The results obtained by the formula differ less than 1 percent from the true values. As it can be seen from the Table, the approximation within 1 percent can be attained with six coefficients B_m (M=5) even for the extremely small values of r. For longer distances the number of coefficients B_m decreases from 5 to 2.

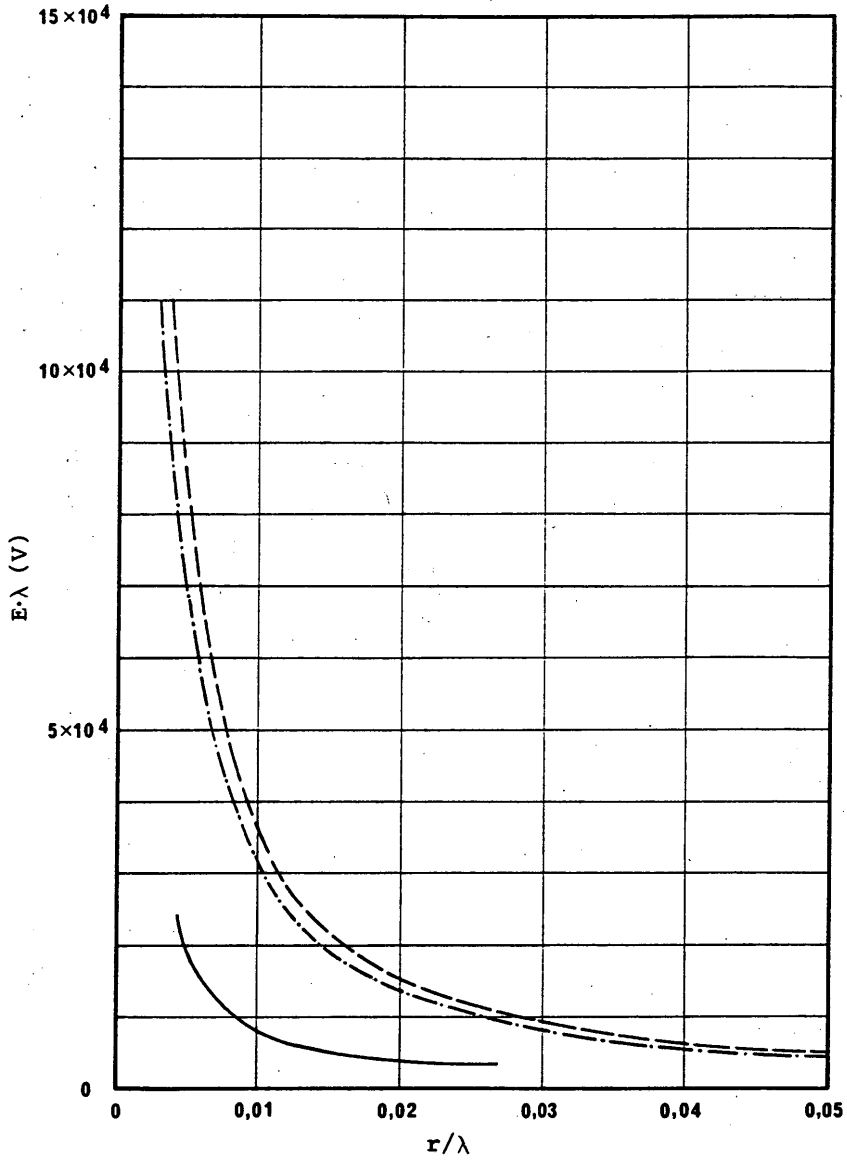


FIGURE 2 - Product E·λ versus the ratio r/λ
(0.004 ⩽ r/λ ⩽ 0.05)

——— h/λ = 0.25

– – – h/λ = 0.5; 0.525; 0.55

–·–·– h/λ = 0.625

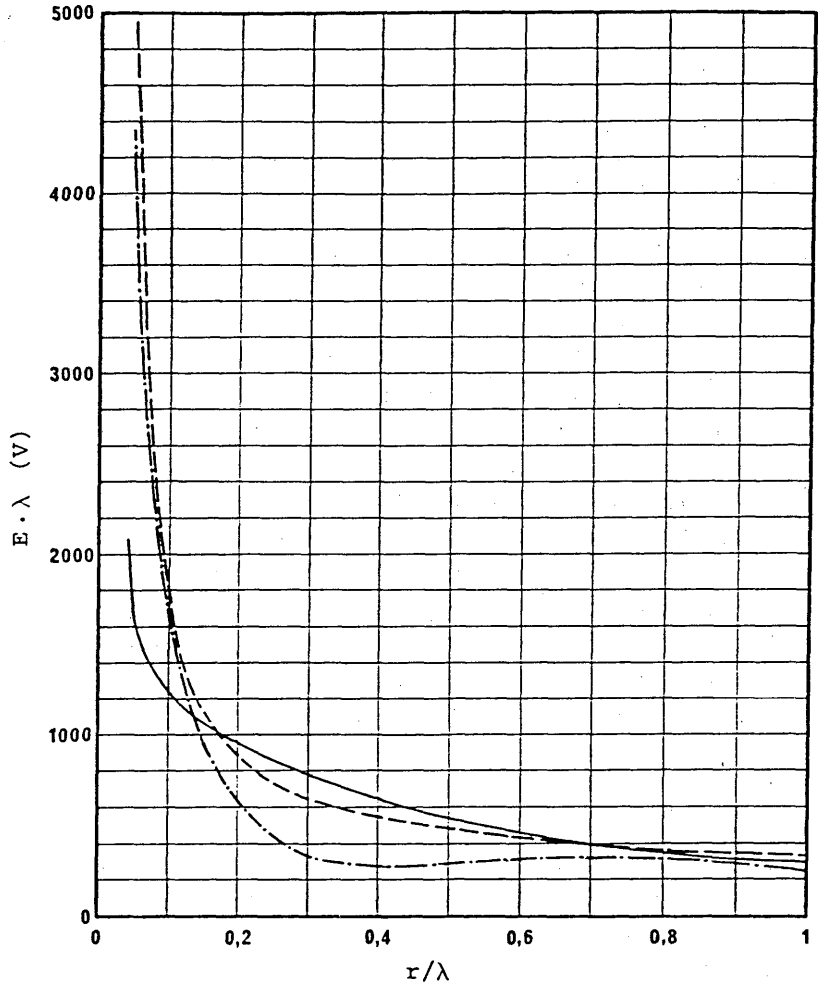


FIGURE 3 - Product E·λ versus the ratio r/λ
(0.05 ⩽ r/λ ⩽ 1.0)

——— h/λ = 0.25

— — — h/λ = 0.5; 0.525; 0.55

—·—·— h/λ = 0.625

Table 1

h/λ	0,004<r/λ<0,1	0,1<r/λ<0,4	0,4<r/λ<1,0
0,250	B_o=-0,6374 E-5	B_o= 0,5915 E-3	B_o= 0,3170 E-3
	B_1= 0,1103 E-1	B_1= 0,1964 E-2	B_1= 0,2950 E-2
	B_2= 0,2340 E 0	B_2= 0,8004 E-3	B_2= 0
	B_3=-0,7564 E+1	B_3= 0	B_3= 0
	B_4= 0,7482 E+2	B_4= 0	B_4= 0
	B_5=-0,2566 E+3	B_5= 0	B_5= 0
0,500	B_o=-0,8643 E-6	B_o=-0,2520 E-3	B_o= 0,1019 E-2
	B_1= 0,2419 E-2	B_1= 0,7976 E-2	B_1= 0,2041 E-2
	B_2= 0,4834 E-1	B_2=-0,4415 E-2	B_2= 0
	B_3=-0,3318 E 0	B_3=-0,6302 E-2	B_3= 0
	B_4= 0,1050 E+1	B_4= 0	B_4= 0
	B_5= 0	B_5= 0	B_5= 0
0,525	B_o=-0,8233 E-6	B_o=-0,2003 E-3	B_o= 0,1194 E-2
	B_1= 0,2386 E-2	B_1= 0,6688 E-2	B_1= 0,1868 E-2
	B_2= 0,4646 E-1	B_2= 0,3705 E-2	B_2= 0
	B_3=-0,3145 E 0	B_3=-0,1779 E-1	B_3= 0
	B_4= 0,1029 E+1	B_4= 0	B_4= 0
	B_5= 0	B_5= 0	B_5= 0
0,550	B_o=-0,7962 E-6	B_o=-0,1168 E-3	B_o= 0,1442 E-2
	B_1= 0,2387 E-2	B_1= 0,4884 E-2	B_1= 0,1642 E-2
	B_2= 0,4535 E-1	B_2= 0,1487 E-1	B_2= 0
	B_3=-0,3043 E 0	B_3=-0,3333 E-1	B_3= 0
	B_4= 0,1031 E+1	B_4= 0	B_4= 0
	B_5= 0	B_5= 0	B_5= 0
0,625	B_o=-0,8668 E-6	B_o= 0,8664 E-4	B_o= 0,4472 E-2
	B_1= 0,2830 E-2	B_1= 0,4446 E-2	B_1=-0,1040 E-2
	B_2= 0,5069 E-1	B_2=-0,1520 E-1	B_2=-0,2576 E-2
	B_3=-0,3453 E 0	B_3= 0,2452 E 0	B_3= 0,2782 E-2
	B_4= 0,1317 E+1	B_4=-0,4425 E 0	B_4= 0
	B_5= 0	B_5= 0	B_5= 0

4. MEASUREMENTS

In order to verify the theoretical results, obtained by the point-matching theory, the electric field strength measurements in the near zone of the MF anti-fading antenna (230 m high) of Radio Beograd 1 (684 kHz) were performed. The measurements were effectuated by using special field strength meter, developed by the IRT and described in [5] . The sketch of the instrument self-contained with three orthogonally positioned short dipoles, is shown in Fig.4. Because of the limited measuring range of the instrument, the transmitter power was reduced from 2000 to 500 kW. The disturbing influence of the operator was avoided by attaching the measuring instrument to the end of an insulating rod.

The electric field strength was measured along a radial direction at a height of 1.5 m above the earth surface. The measured results are presented in Fig. 5 as a function of the distance from the antenna mast. For the sake of comparison the theoretical results are given on the same diagram, as well. It should be noted that the agreement between the theoretical and measured values is quite good, except at the very short distances (shorter than about 8 m),

As noted in the Introduction, near field characterization (both electric and magnetic) close to the base of a mast antenna is less precise.

5. CONCLUSIONS

A simple but accurate theoretical procedure for calculating the electric field strength in the near field zone of the MF and LF mast antennas (except very close to the antenna base) is presented here. The proposed procedure is based on a rigorous approach to the linear antenna problem.

The theoretical results, obtained for a particular antenna, are compared with the measured field strengths and a satisfactory agreement was found.

Validity of the point-matching theoretical procedure is established, but its accuracy in the region within several meters of the antenna base is somewhat less.

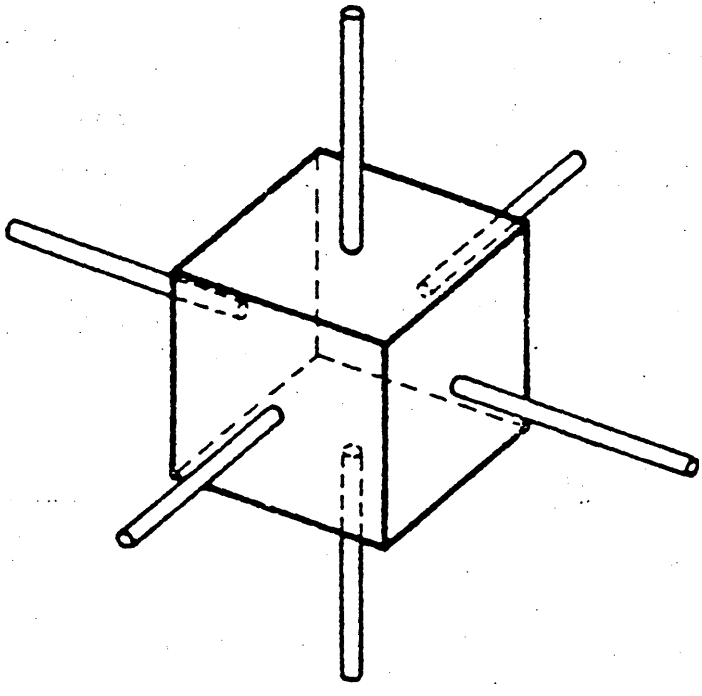


FIGURE 4

Sketch of the special field strength meter

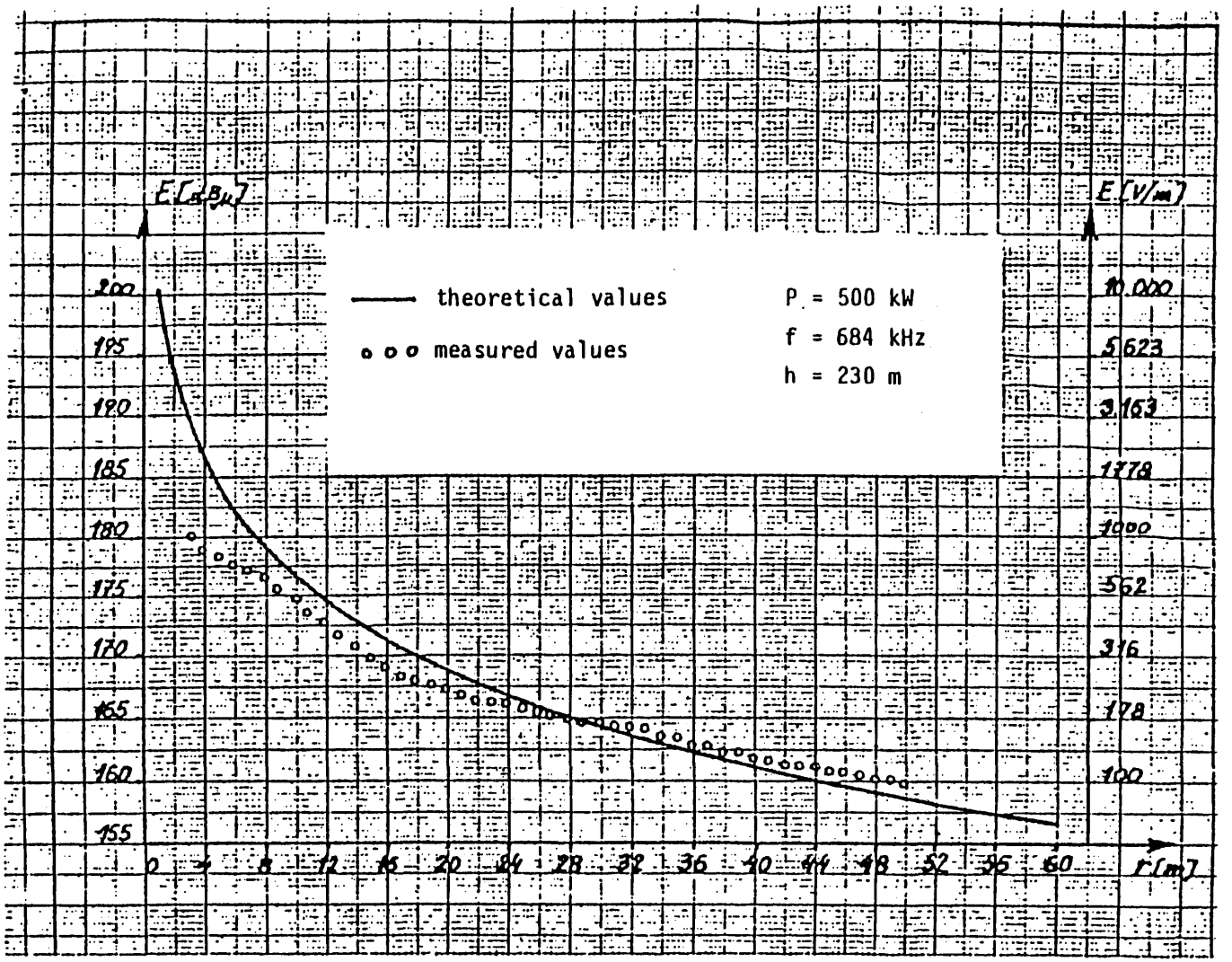


FIGURE 5

Comparison of the results obtained by the theory and measurements

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REPORT 1100

IMPACT OF ATMOSPHERIC ATTENUATION ON SPECTRUM MANAGEMENT ABOVE 20 GHz
(Question 64/1)

(1990)

1. Introduction

The magnitude of atmospheric attenuation above 20 GHz makes it potentially a useful tool for spectrum management and planning for these frequencies. To date, atmospheric attenuation has been treated integral to EMC analyses and interference evaluation but not as a primary spectrum management and planning tool. The 1979 World Administrative Radio Conference (WARC-79) revised the International Table of Frequency Allocations above 20 GHz to reflect a high level of interest and activity in this portion of the spectrum. The new table of allocations was created with the objectives of stimulating development of this spectrum resource by providing guidance and protection to users and of providing each potential user with suitable bands in all parts of the spectrum. Propagation phenomena played a maior role defining the new table. This report discusses the impact of atmospheric attenuation on spectrum management and planning for frequencies above 20 GHz; outlines potential approaches to attenuation-based management and planning; identifies existing data bases, models, and studies for the frequencies above 20 GHz useful for attenuation-based spectrum management and planning; and identifies additional data bases, models, and studies needed.

2. Spectrum management above 20 GHz

The major feature of the spectrum above 20 GHz is the frequency-dependent structure of atmospheric attenuation. The region is characterized by windows, which are bands of relatively low attenuation, and absorption bands, where attenuation is very high, as shown in Fig. 1 of Report 719. Further attenuation, even in the window areas, is much higher than for the lower frequencies.

For spectrum management and planning purposes, the spectrum may be divided into bands corresponding to windows and absorption bands [Katzenstein, et al., 1981]. These are defined primarily by the absorption characteristics of oxygen and water vapour; but for total attenuation band shapes, centre frequencies, etc. are affected by hydrometeors such as rain and snow. For the purposes of this paper, 20 GHz to 40 GHz will be treated as one band and the spectrum above 40 GHz discussed in terms of the bands defined by Table I. Table I [Katzenstein, et al., 1981] defines band limits of atmospheric windows and absorption bands for the frequencies above 40 GHz. The band edges correspond to the International Table of Frequency Allocations.

TABLE I - Atmospheric window and absorption band limits

Window	Absorption Band	Limits (GHz)
W_1		39.5 - 51.4
	A_1	51.4 - 66
W_2		66 - 105
	A_2	105 - 134
W_3		134 - 170
	A_3	170 - 190
W_4		190 - 275

The spectrum allocations above 20 GHz reflect both requirements and the projected ability to utilize these frequencies. For example, a very small portion of the spectrum between 40 GHz and 100 GHz is allocated to broadcasting. Broadcasting requires a high degree of reliability which will be difficult to achieve due to the propagation characteristics of this portion of the spectrum.

2.1 Band 20 GHz to 40 GHz

The atmospheric propagation in the region between 20 GHz and 40 GHz is dominated by the water absorption peak at 22.3 GHz. Allocations in this region reflect the sharing potential provided by the relatively high attenuation in the band, particularly between 20 GHz and 30 GHz. But unlike the allocations above 40 GHz, there are allocated a number of exclusive bands, such as the Radio Location allocation in the band 33.4 GHz to 34.2 GHz, which reflect the lower attenuation in the band 30 GHz - 40 GHz and thus increased potential for interference.

2.2 Window W_1 (39.5 to 51.4 GHz)

The window W_1 is actually a continuation of the window roughly centered at 35 GHz. The lower band edge is defined by the International Table of Frequency Allocations.

Attenuation and the limited transmitter power attainable at these frequencies will play a major role in spectrum management and planning for this window. The utility of the allocations above 47.2 GHz, which lie on the edge of an absorption band, have yet to be proven useful for services requiring high link reliability and immunity from weather degradation. Further, current state of the art cannot provide space-qualified transmitters with sufficient power to make practical satellite downlinks. Planned systems must use lower frequencies for downlinks, such as the allocation in the band 20.2 GHz to 21.2 GHz.

2.3 Absorption band A_1 (51.4 to 66 GHz)

Services in the absorption band A_1 (51.4 to 66 GHz), such as intersatellite communications and radiolocation for ground traffic control, can take advantage of the high attenuation to provide the high degree of protection from interference needed by these services. However this absorption band is not uniform in terms of attenuation, but consists of a complex fine structure at higher elevations which could affect sharing [Report 719].

Airborne mobile and radiolocation services could possibly interfere with the Intersatellite service. Adequate sharing studies are required to determine the extent of this problem, and such a study program has been requested of the CCIR. The Radio Regulations take note of this absence of sharing criteria and limit activities of the Aeronautical Mobile and Airborne Radar services to those not causing harmful interference to the Intersatellite service operating in the same bands, and frequency assignments could be affected by results of the CCIR studies.

2.4 Window W_2 (66 to 105 GHz)

The attenuation in this window is considerably higher than in W_1, but is somewhat offset by the narrow beamwidths that can be achieved by small antennas Report 663-1 . These narrow beamwidths make it particularly attractive for the Radiolocation Service. Propagation characteristics weather effects. equipment state of the art, and desire to minimize antenna size make the 90 GHz to 100 GHz portion of this band the most attractive, particularly for the Radiolocation Service [Report 664]. Studies [Katzenstein, 1980] have concluded that sharing involving the Radiolocation Service and the other services allocated to the band 90 GHz to 100 GHz will represent no significant problem.

The window W_2 is the highest window in which allocations to the Broadcasting-Satellite and Broadcasting services have been made. Operation of broadcasting systems above these frequencies cannot be achieved with good reliability using todays technology because of attenuation due to hydrometeors, and the feasibility of this service in the W_2 region has not been established.

2.5 Frequency bands above 105 GHz

The frequency bands between 105 GHz and 275 GHz (the limit of allocations made at WARC-79) generally follow the pattern of allocations below 105 GHz. The large amount of spectrum available and the breadth of the windows and absorption bands lessen the frequency assignment problem.

3. Availability of propagation data and models for spectrum management and planning above 20 GHz.

The ability to perform spectrum management at frequencies above 20 GHz depends heavily on the availability of propagation data and models to employ it. Calculation of attenuation and thus coordination distances, interference levels, etc differ markedly from those at lower frequencies due to both the large, highly frequency dependent attenuation encountered and differing phenomenon. For example at frequencies above about 30 GHz water-vapour attenuation is quadratic with frequency instead of linear as at the lower frequencies [report 719]. These calculations also depend heavily on adjustments and coefficients derived from empirical data which is in short supply. The following sections provide an overview of the availability of data and models addressing the major factors in path attenuation at these frequencies: gaseous attenuation; attenuation by hydrometeors; scattering due to precipitation; and ducting.

3.1 Gaseous attenuation

Molecular absorption experienced during radio-wave propagation through the atmosphere at centimetre or millimetre wavelengths is due primarily to atmospheric water vapour and oxygen. However, trace gases can contribute significant attenuation in the absence of water vapour at frequencies greater than about 70 GHz [Report 719].

The calculation of the specific absorption produced by either the oxygen or the water vapour components of the gaseous atmosphere is complex. The dependence of these calculations on empirical data [Report 719 .] emphasizes the need for additional data, particularly in the above 40 GHz region.

Report 719 provides a detailed description of attenuation by atmospheric gases and provides equations and data for calculation of specific and path attenuation for frequencies below 350 GHz.

Liebe [1983] and Liebe et al. [1985] provide detailed models of atmospheric millimeterwave propagation and comparison of model and path data.

3.2 Attenuation by hydrometeors, in particular precipitation, and other atmospheric particles

Precipitation, especially rain, causes absorption and scattering of radio waves. These effects combine to produce attenuation. Reports 721 and 882 provide extensive treatment of this subject.

Calculations of the specific attenuation at a given frequency versus rainfall intensity strongly depend on the assumed microstructure of rainfall (size distribution, temperature, terminal velocity and shape of the raindrops) (see Report 563). The calculations are hampered by lack of empirical data and the difficulty of modelling rain microstructure.

To predict rain attenuation Report 721 recommends the semi-empirical approach defined in Reports 338 and 564 . It uses very simple relations only, as these suffice to describe the limited data available and is considered to have practical applicability in system planning. This approach emphasizes the need for additional data.

Attenuation statistics are a major consideration in establishing path and site diversity required to overcome the severe weather degradation experienced in these bands, but more information on the statistical structure of rain intensity over an area is required to develop models for predicting path diversity improvement from statistics of rain rate measured at a point. Report 721 , 564 , 563 and Ippolito et. al [1983] provide detailed data and analytical methods needed to plan for site and path diversity for space services. Some of this information is applicable to terrestrial services.

3.3 Ducting

Report 563 calls attention to ducting, particularly over the sea surface. These ducts allow both communication and interference beyond the line of sight limit normally associated with frequencies above 20 GHz. This is a significant consideration for the Maritime Mobile service. There exists no body of ducting statistics for the bands above 20 GHz adequate for accurate assessment of the impact of ducting on telecommunications. As the number of maritime systems operating above 20 GHz increases, such statistics will be needed.

3.4 Aspects relative to the terrestrial broadcasting and mobile services

Report 562 is the only CCIR report that deals with the aspects of attenuation relative to terrestrial broadcasting and mobile services operating above 20 GHz, but does not provide adequate data or methodology. Data applicable to the terrestrial broadcasting and mobile services is inadequate.

3.5 Aspects relative to terrestrial fixed services

Report 338 provides information useful for planning path diversity up to 40 GHz, but is primarily concerned with frequencies below 12 GHz. The other reports of the CCIR dealing with the Terrestrial Fixed service do not provide data adequate for spectrum planning and management for the bands above 20 GHz. However, reports dealing with coordination and interference for space services, such as Report 724 , contain some information applicable to Terrestrial Fixed services operating above 20 GHz. Measurements are lacking over adequate path lengths which are made for sufficient time to encompass a representative range of meteorological conditions, including precipitation.

3.6 Aspects relative to space telecommunications systems

There exists data, studies, and models for space telecommunications systems operating above 20 GHz, including satellite data. Most of this data has been summarized by Ippolito et al. [1983]. Report 564 also provides a good treatment of space telecommunications systems operating above 20 GHz and discusses the additional data needed. Report 565 addresses broadcast satellites, covering bands up to 106 GHz.

The CCIR Reports for the mobile satellite services, 884 and 1009, do not provide information useful for the bands above 20 GHz and data and studies applicable to these mobile services is needed.

4. Conclusions

Spectrum management above 20 GHz requires of the manager much more extensive understanding of the state of the art and the physics of propagation phenomenon than is required for the bands below 20 GHz The probabilistic nature of interference becomes of paramount importance in the above 20 GHz region because of the large and highly variable nature of the effects of the environment on propagation.

There is an urgent need for data to support design and modelling of telecommunications systems operating above 20 GHz.

Continued study by the CCIR and the various administrations are required to successfully manage the spectrum above 20 GHz.

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REPORT 1104*

LIMITATION OF RADIATION FROM INDUSTRIAL, SCIENTIFIC AND MEDICAL (ISM) EQUIPMENT

(Study Programme 4D/1)

(1990)

1. Introduction

This Report is written in response to Resolution No. 63 of WARC-79, CCIR Study Programme 4D/1 and CCIR Decision 54. The Report describes what Industrial, Scientific and Medical (ISM) equipment is, it then provides a summary of current and envisaged ISM applications and the frequencies involved. ISM propagation characteristics and interference effects on radio systems are also described. The CISPR ISM radiation limits, intended to replace CISPR Publication 11 dated 1975 are given with an explanation of their derivation.

2. ISM Applications

ISM Application is the operation of equipment or appliances designed to generate and use locally radio frequency energy for industrial, scientific, medical, domestic or similar purposes, excluding applications in the field of telecommunications. In dealing with the abatement of radio interference from this class of equipment CISPR has found it convenient and helpful to divide ISM equipment into two main categories, namely:

Group 1 ISM equipment: Group 1 contains all ISM equipment in which there is intentionally generated and/or used conductively coupled radio frequency energy which is necessary for the internal functioning of the equipment itself.

Group 2 ISM equipment: Group 2 contains all ISM equipment in which radio frequncy energy is intentionally generated and/or used in the form of electromagnetic radiation for the treatment of material; and spark erosion equipment.

Each group is sub-divided into two classes:

Class B equipment is equipment suitable for use in domestic establishments and for establishments directly connected to a low voltage power supply network which supplies buildings used for domestic purposes.

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* The Director of CCIR is requested to bring this document to the attention of: ICAO, IMO, INMARSAT, IALA, COSPAS/SARSAT

Class A equipment is equipment suitable for use in all establishments other than domestic and those directly connected to a low voltage power supply network which supplies buildings used for domestic purposes.

Note: Operation of equipment which does not meet the class A limits, but does not result in unacceptable degradation of radio sevices, may be sanctioned on a case by case basis by the competent national authority.

Note: Although the class A limits have been derived for industrial and commercial establishments administrations may allow with whatever additional measures are necessary the installation and use of class A equipment in a domestic establishment or in an establishment connected directly to domestic electricity power supplies.

This division into groups and classes was supported by a number of administrations. However, some administrations found it more suitable to divide ISM equipment into only three categories reflecting domestic, commercial and industrial application.

For the purpose of this Report Information Technology Equipment (ITE) is deemed not to be ISM equipment (ITE is dealt with by CISPR in their Publ 22).

The devices in group 1 in the opinion of some administrations are not those historically envsaged as ISM equipment. As such their treatment for interference control and abatement is likely to be quite different from that necessary for the devices in group 2. It is with the group 2 equipment, where for medical and industrial purposes the rf energy is used to generate heat for the treatment and processing of materials (including human tissue) and the manufacture of products that this report is mainly concerned. Heat can be generated within the material, on its surface or in close proximity as the process requires. It can readily be controlled in its level, distribution and period of application. The general advantages of electroheat processes include high efficiency, controlled quality, improved environment and great versatility.

Many countries are rapidly reducing their dependence on fossil fuels and this has proved an added incentive to the adoption of electroheat processes leading to higher industrial productivity and more competitive manufactured goods. One very important factor is the improvement in product quality which is obtained by the use of electroheat processes with the resultant reductions in wastage and consequent reduction in costs.

2.1 Current applications

The frequencies currently employed for industrial, scientific, medical and other non-communication applications cover a very wide spectrum and include also frequencies other than those designated by the Radio Regulations. A number of ISM equipment use frequencies of undefined tolerance and stability and some of them use frequencies allocated to the safety services and radionavigation services. Table I summarizes information on the frequencies used;

Table I shows the frequencies designated to ISM in the Radio Regulations.

TABLE I

ISM equipment in current use

FREQUENCY	TYPICAL EQUIPMENT	MAJOR APPLICATIONS
Below 75 Hz	60 kW-20 MW installations Rotary machines and solid state inverters (mainly at 50 Hz or 60 Hz)	Induction heating and melting of metals
75-1000 Hz Nominal frequency variation ±10 %	100 kW - 10 MW installations, some motor alternators, mainly solid state power sources, comprising semi-conductor rectifiers and inverters	Induction heating and melting of metals
1 - 9 kHz Nominal frequency variation ±10 %	40 kW - 10 MW installations, some motor alternators, mainly solid state power sources, comprising semi-conductor rectifiers and inverters	Induction heating and melting of metals
9 - 50 kHz Nominal frequency variation ±10 %	100 W - 5 kW installations, solid state power sources, comprising semi-conductor rectifiers and inverters; 10 kW - 300 kW installations, solid state power sources	Ultrasonic welding and cleaning
50 - 1000 kHz Nominal frequency variation ±5 %	50 kW - 1 MW installations with valve or solid state generators; 100 kW - 1 MW installations, solid state power sources comprising semi-conductor rectifiers and inverters, low power medical and dielectric generators	General induction heating and melting of metals; medical applications surgical diathermy; glue curing
1 - 5 MHz Nominal frequency variation ±5 %	1 kW - 1500 kW installations; valve induction generators, low power medical and dielectric generators	Sealing for packaging; production of semi-conductor materials; surgical diathermy; glue curing
5 - 20 MHz Nominal frequency variation ±2 1/2%	1 kW - 250 kW installations; valve dielectric generators	Wood glueing and drying; medical applications, medical diathermy
20 - 80 MHz Nominal frequency variation ±1 %	1 kW - 100 kW installations; valve dielectric generators	Plastic sheet welding; food processing; material manufacture; material drying and preheating; medical diathermy; low power pulsed short wave

TABLE I (continued)

FREQUENCY	TYPICAL EQUIPMENT	MAJOR APPLICATIONS
100 - 500 MHz Nominal frequency variation ±1 %	100 W - 5 kW; valve generators	Medical applications
500 - 3000 MHz	1 kW - 100 kW installations, microwave generators, (majority operating at 2450 MHz), rf excited plasma equipment; microwave oveus	Food processing; materials manufacture; chemical analysis medical applications

The frequencies employed by the ISM Industry cover a wide frequency spectrum and include frequencies other than those designated for ISM applications in the Radio Regulations.

50 Hz to 9 kHz

The frequency of 50 or 60 Hz is widely employed for induction heating as also are frequencies between 50 Hz and 9 kHz. Over the frequency range, individual equipment operate at power levels from 10 kW up to 10 MW. There are hundreds of thousands of installations operating throughout the world. No cases of interference are reported in spite of the relatively high levels of radiation emitted from the rectification systems and the terminal voltage levels of the harmonics.

9 kHz to 150 kHz

Until the early seventies, the only ISM equipment operating in this frequency band was ultrasonic cleaning and welding equipment which largely used a frequency band between 15 and 30 kHz. There are to-day tens of thousands of ISM installations operating between 9 and 150 kHz. Few cases of interference have been reported.

150 kHz to 1000 kHz

No frequency bands have been designated to ISM equipment in this range. As a result, the whole spectrum has been used over the past forty years and to-day there are hundreds of thousands of units in operation (some of them working near the distress frequency of 500 kHz).

Individual units operate at power levels of 1 kW to 500 kW and may radiate at levels much greater than those recommended by CISPR.

1 MHz to 100 MHz

The industry has tried to concentrate wherever possible on 27.12 MHz due to the greater frequency tolerance (as compared with other bands designated to ISM) but often this frequency is not practical for physical reasons and lower or higher frequencies have to be employed.

Two additional frequencies within this range are used extensively for ISM. The frequency 13.56 MHz is used for plasma etching of silicon wafers, wood glueing, and medical diathermy, while 40.68 MHz is used for dielectric heating.

100 - 1000 MHz

The two ISM designations in this frequency band are widely employed. Other frequencies are being used to a limited extent.

The band 902 MHz to 928 MHz is designated for Region 2 only.

Above 1000 MHz

The ISM radiation bands of 2450 MHz and 5800 MHz are used. The major use is the 2450 MHz and 2375 MHz (USSR only) bands where millions of domestic microwave ovens are in operation.

2.2 Future applications

The drive throughout the world for savings in energy and cleaner industrial environments is leading to an increasing use of electricity in industrial processes. In the change from the direct use of fossil fuels to electro-heat savings of 80 % in energy, improved control of quality and better working and environmental conditions have been recorded.

Four areas of future development directly affected by radiation limitation proposals stand out.

I) the development and use of high power plasma technology for air heating and material treatment;

II) the use of high power induction heaters for refining and treatment of materials for the semi-conductor industry. These processes are essential in the manufacture of high frequency semi-conductor devices for use in the communications industry;

III) the further development of high power and highly efficient induction power sources, and their more widespread use throughout the manufacturing industries. This is in part coupled to new high efficient transverse flux methods of heating flat materials.

IV) efficient point-to-point transport of energy by a highly collimated microwave beam.

V) high power dielectric drying.

2.2.1 Plasma torches at 13.56 MHz and 27.12 MHz

Plasma torches which produce large volumes of very hot gases may be used for heating in a conventional manner or may be employed to stimulate chemical reactions or plasma gas may be used to conduct electricity as a flexible electrode. Specific applications include:

a) use by the aerospace industry for simulating atmospheric re-entry conditions for ballistic missiles and other space hardware. A group of torches having a combined power of 20 MW is used for this purpose in France;

b) use to heat or superheat reducing gases in blast furnaces to introduce a saving of more than 80 % in consumption of foundry coke.

c) recovery of metals from ore or dust and the production of ferro-alloys, input powers of up to 6 MW are envisaged;

d) combination of induction heating and plasma torches to improve the efficiency of certain processes, input powers of 1 MW are envisaged.

2.2.2 Induction heating in the semi-conductor and information technology industries

The refining of very pure semi-conductor materials relies heavily on the use of induction generators. High power induction generators are being applied at the research level and will in the future be transferred to industrial applications. Work in other aspects of semi-conductor materials of a very confidential nature are in progress which will have a great impact on the information technology industries. The use of induction generators is central to this work.

2.2.3 High power induction equipment

Over the past 15 years high efficiency semi-conductor based induction generators have been applied for melting metals, particularly vacuum melting for the aerospace industry for heating for forging and steel rolling and heat treatment of engineering components for the machine tool and automotive industries. These generators have progressed from tens of kilowatts to tens of megawatts over this period. Application techniques such as transverse flux to improve still further the efficiency of such equipments will bring about the need for larger generators and their more widespread use. Already a pilot installation of 10 MW is in operation within the British Steel Corporation and another of 37 MW in Sweden. Their undoubted success both econimically and environmentally is acting as a stimulus to their very wide use throughout all industrially based countries.

2.2.4 Efficient point-to-point transport of energy by a highly collimated microwave beam

One of possible future non-communication applications of radio waves that have received substantial study is the free-space power transmission by microwave beam. It is defined as the efficient point-to-point transport of energy by a highly collimated microwave beam. "Free-space" is defined to exclude the use of any physically injected material such as waveguides between the transmitting and receiving points of the system but not to exclude the presence of gaseous, liquid, or congealed material that exists naturally in the earth's atmosphere or in space.

The following properties of microwave power transmission are of great importance:

1. No mass either in the form of wire conductors or ferrying vehicles is required between the source of energy and the point of consumption.

2. Energy can be transferred at the velocity of light.

3. The direction of energy transfer can be rapidly changed by repointing the transmitting antenna.

4. There is no loss of energy in the transfer of energy through the vacuum of space.

2.2.4.1 Energy from Satellite Solar Power Station

A conceptual design of a satellite solar power station (Corson, 1981) assumes a transmitted power level of 6500 MW at 2450 MHz. The projected overall efficiency of the microwave energy transmission system, including the conversion efficiencies at the satellite and earth stations is 68 %. The rest is transferred into heat or irradiated into space in the form of radio waves.

The peak power flux-density at the centre of the collecting area is 20 mW/cm^2. A guard area extending out to a 10 km radius will encompass power flux-densities in excess about 50 $\mu W/cm^2$. Outside the collecting area and guard band, the power flux-density will be attenuated to a level of about -13 $dB(W/m^2)$.

2.2.4.2 Microwave-powered vehicles

Experiments were undertaken with microwave-powered vehicles: helicopters and aeroplanes. Powers of the order of 600 kW at 2450 MHz were involved (Brown, 1974).

2.2.5 High power Dielectric Drying

Dielectric drying is not a fundamentally new application but has been employed in ovens at the end of drying because of the ability of dielectric heating to concentrate its energy in the areas of highest moisture content.

Looking into the future as energy costs increase, the potential for better thermal efficiency by use of dielectric heating will cause it to be more extensively used for the complete drying operation doing away with other processes. The power requirements can be well into the megawatts region where efficiencies can approach 65 % over all. Reduced pollution due to lack of combustion products, reduced floor space, opportunities for automation, will aid this product growth.

3. Radiation from ISM equipment

In order to assess the interference potential of ISM equipment in operational use, it is required that the results of radiation measurements made at factory installations be analysed and made available. Data is provided in section 3.1 and 3.2 on both attenuation and fieldstrength.

3.1 Industrial induction heating equipment operating in range 0.15 to 1 MHz

In the UK nineteen factory sites each using such an equipment were examined in detail. At each site rf field strength measurements were made over the range 10 to 100 meters along four radial lines having angular intervals of 90° above the equipment. The frequencies measured were the fundamental and up to three of its harmonics.

These measurements were analysed to show:

I) the attenuation of harmonics relative to the fundamental;

II) the attentuation of the rf field (F) with distance and hence the attenuation index "n" of the law

$$F \quad \left(d^{-n}\right)$$ were "d" is distance from equipment.

The results are given in Table XIII in annex II.

3.2 Equipment operating in 27 MHz band

In the UK three factory sites using a total of twenty equipments were examined in detail.

The results are given in Tables XIV and XV in annex II.

The measurements were analysed to show:

I) the attentuation of harmonics relative to the fundamental;

II) the attenuation index "n".

The results referred to in sections 3.1 and 3.2 reflect the actual attenuation in a real environment (building attenuation, reflections, refractions, near field effects etc.). Further measurements to both expand the data and improve confidence in these results would be beneficial.

4. Short range propagation

ISM radiation limits are usualy defined for fixed distances from an equipment with 10, 30, 100 metres being commonly used. At frequencies above 5 MHz all these distances can be considered as being in the far-field and, if only the direct ray ia considered an inverse law (1/d) could be used to interpolate the field strength from a measurement made at any one of the above fixed distances.

From theoretical considerations the near-field from a defined source can be calculated, however any object within the field would modify this result. Additionally, in most practical situations a field strength measurement is the resultant of the direct ray plus indirect ray(s) and a simple 1/d law can be seen not to apply.

During 1984 the CISPR sub-committee on ISM decided to see if there could be agreement as to a propagation law of the type:

$$\text{field strength} \quad d^{-n}$$

A questionnaire was circulated to members of the sub-committee requesting that they consult with their national experts then indicate the value of "n" which they would use if it were necessary to consider propagation paths of length 10 to 300 metres at various frequencies.

Fifteen administrations returned questionnaires and an analysis of these returns is given in Table II.

TABLE II - Mean and standard deviation values of exponent 'n' for short range propagation derived from returned questionnaires

FREQUENCY		DISTANCE FROM SOURCE 10 meters	30 meters	100 meters	300 meters
150/285 kHz	x	2.72	2.33	2.34	2.03
	σ	0.45	0.76	0.81	0.8
285/490 kHz	x	2.3	2.47	2.2	1.58
	σ	0.45	0.82	0.84	0.8
490/1605 kHz	x	2.67	2.2	1.67	1.31
	σ	0.52	0.77	0.71	0.44
1.605/30 MHz	x	2	1.7	1.32	1.32
	σ	0.82	0.7	0.47	0.47
30/68 MHz	x	1.31	1.4	1.4	1.35
	σ	0.46	0.49	0.52	0.47
68/108 MHz	x	1.22	1.3	1.36	1.3
	σ	0.44	0.48	0.5	0.48
108/156 MHz	x	1.22	1.26	1.33	1.35
	σ	0.44	0.51	0.54	0.51
156/174 MHz	x	1.25	1.29	1.41	1.44
	σ	0.46	0.54	0.58	0.62
174/216 MHz	x	1.28	1.38	1.48	1.44
	σ	0.44	0.53	0.57	0.62
216/400 MHz	x	1.17			
	σ	0.41			
400/470 MHz	x	1.17	1.22	1.25	1.36
	σ	0.41	0.53	0.61	0.63
470/585 MHz	x	1.24	1.38	1.36	1.25
	σ	0.42	0.54	0.63	0.61
585/614 MHz	x	1.29	1.33	1.36	
	σ	0.49	0.57	0.63	
614/854 MHz	x	1.27	1.4	1.36	1.25
	σ	0.46	0.56	0.63	0.61
854/960 MHz	x	1.17	1.22	1.25	
	σ	0.41	0.53	0.61	

x = mean value
σ = standard deviation

Note - It is necessary that complementary measurements be carried out to confirm these results.

5. Possible radiation from future ISM equipment

There is no substantially new physical mechanism expected for radio frequency radiation from future ISM applications. However, certain future applications require high power on the fundamental frequencies. The most significant unintended effects of free-space power transmission by microwave beam seem to remain:

1. out-of-band spurious radiation

2. spill-over radiation, i.e. the percentage of the total energy that is vasted in the form of stray radiation outside the energy collecting area.

6. Effects on receiving equipment

Effects of interfering signals depend upon the susceptibility of receiving equipment and upon the relationships among the power levels, types, modulations, and bandwidths of wanted and interfering signals at the receiver.

The following interference mechanisms may occur:

a - when the ISM operating frequency coincides with the legitimate receiver pass-band,

b - when the ISM operating frequency coincides with one or more receiver spurious responses,

c - when the ISM operating frequency or a ISM spurious frequency is strong enough to drive an active device in the receiver into a non-linear condition. This may occur in all stages of a receiver (front end, IF, audio amplifier, etc.)

d - when one or more ISM spurious frequencies coincide with the legitimate receiver pass-band,

e - when one or more ISM spurious frequencies coincide with one or more spurious responses of the receiver.

In the case of impulsive interference, much depends on the receiver time-response, which, in turn, is determined by the bandwidth and characteristics of the detector and automatic gain control circuits, if any. Receivers using narrow band filters with steep skirts have transient responses which persist for a long time under impulsive interference conditions, and are thus vulnerable to this type of disturbance.

Some relief from interference may be gained by improving the receiver immunity, that is by making it less suceptible. This may sometimes be done by the use of selective filters and/or a directive antenna at the receiver input, by suitable interference cancellation techniques and by RFI suppression filters and screening.

When directional receiving antenna is used, one can differentiate the following cases:

a) the source of interfering radiation lies on the direction of the main lobe of the receiving antenna, or

b) the source lies on the direction outside the main lobe of the receiving antenna.

In the second case the interfering signal is additionally attenuated due to directivity of receiving antenna. Similar classification can be introduced to include polarization effects.

Note - Relief from interference should first be provided by the filtering, screening or, if feasible, relocation of the interfering ISM equipment.

7. Radiation levels inside the bands designated for ISM use

Some measurements of the radiation levels generated by ISM equipment in the bands designated for their use have been carried out in different countries and at different locations. Table III gives a survey of the results. Further measurements at 30m distance from the outside wall of the buildings housing the ISM equipment are required.

TABLE III

Levels of field strength generated by ISM equipment in the frequency bands designated for their use

Centre Frequency (MHz)	Frequency Range (MHz)	No of appropriate footnote to the Table of Frequency allocations of the ITU Radio Regulations	Range of measured Field Strengths (dB(μV/m)) (1)
6.780	6.765-6.795	524	80 - 100
13.560	13.553-13.567	534	80 - 150
27.120	29.957-27.283	546	70 - 150
40.680	40.66-40.70	548	60 - 120
433.920	433.05-434.79	661(2)662 (Reg 1)	60 - 120
915.000	902-928 (2)	707 (Reg. 2 only)	60 - 130
2450	2400-2500	752	30 - 130
5800	5725-5825	806	No information
24125	24000-24250	881	No information
61250	61000-61500	911	No information
122500	122000-123000	916	No information
245000	244000-245000	922	No information

Note 1 - The field strength is that existing at a distance of 30 m from the boundary of the premises in which the ISM equipment is situated. Therefore the actual distance between the ISM equipment and the measuring point is not known. The CCIR has concluded that measurements should be referenced to the outside wall of the building housing the ISM equipment.

Note 2 - 896 MHz in the United Kingdom.

8. Frequencies designated for ISM use

8.1 By the ITU

These frequencies are for use in all ITU Regions unless otherwise stated.

TABLE IV

Frequencies designated for ISM use

Frequency band (MHz)	Appropriate footnote to the Table of Freq. Allocations of the ITU Radio Regs.
6.765 - 6.795	524*
13.553 - 13.567	534
26.957 - 27.283	546
40.660 - 40.700	548
433.050 - 434.790	REGION 1 ONLY 661* & 662
902.000 - 928.000	REGION 2 ONLY 707
2.400 - 2.500	752
5.725 - 5.875	806
Frequency band (GHz)	
24.00 - 24.25	861
61.00 - 61.50	911*
122.00 - 123.00	916*
244.00 - 246.00	922*

* Use of these frequency bands is subject to special authorisation by administrations concerned in agreement with other administrations whose radio communications services might be affected.

8.2 By Individual Countries

Permitted frequencies (MHz).

Countries which are not mentioned by name are assumed to use only ITU designated frequencies.

Australia : ITU designated and 918 - 926 MHz
Netherlands : ITU designated 3.370-3.410 MHz unlimited radiation
Poland : 13.56, 27.12, 40.68

USSR:	0.008 ± 7.5% (Ind. 1);	0.018 ± 7.5% (Ind. 1)
	0.022 ± 7.5% (Ind. 1, S1 M1);	0.044 ± 10% (Ind.1 S1 M1)
	0.066 ± 12% (Ind.1 S1 M1);	
	- 10%	
	0.880 ± 1.0% (Ind.2 S2 M2);	1760 ± 2.5% (Ind.2 S2 M2)
	2.640 ± 1.0% (M2);	5.280 ± 2.5% (Ind.2 S2 M2)
	13.560 ± 1.0% (Ind.2);	27.120 ± 1.0% (Ind.2)
	±0.05% (S1 M1);	±0.06% (S1 M1)
	40.68 ± 1.0% (Ind.2);	81.36 ± 1.0%
	± 2.0% (S2 M1)	
	152.50 ± 1.0% (Ind.2 S2 M2);	460.00 ± 1.0% (S1 M1)
	915.00 ± 2% (S1 M1);	2375 ± 2.0% (Ind.1 S1 M1 H1)
	2450 ± 2% (Ind.1 S1 M1 H1);	22125 ± 0.5% (S1)
	24125 ± 0.5% (Ind.1 M1);	42300 ± 2.5% (Ind.1 S1 M1)
	46200 ± 2.5% (Ind.1 S1 M1);	48400 ± 2.5% (Ind.1 S1 M1)

Notation:
- H Household apparatus
- Ind Industrial apparatus
- M Medical apparatus
- S Scientific apparatus
- 1 Use of this frequency is allowed without limitation of radio frequency level providing that no disturbing effect is produced for radio services which operate in this band.
- 2 Use of this frequency is allowed provided that the radio interference level does not exceed the established limits.

UK:	13.533 - 13.553 MHz	110 dB μV/m	100 m	
	13.553 - 13.567 MHz	unlimited		
	13.567 - 13.587 MHz	110 dB μV/m	100 m	All measured from the boundary of the users premises.
	26.957 - 27.283 MHz	unlimited		
	83.996 - 84.004 MHz	130 dB μV/m	30 m	
	167.992 - 168.008 MHz	130 dB μV/m	30 m	
	886.0 - 906.0 MHz	120 dB μV/m	30 m.	

At present above 1000 MHz there is no primary legislation of ISM equipment in the UK.

9. Radiation limits

9.1 Limits application

The determination of satisfactory limits for ISM equipment is greatly complicated by the difference in national goals and policies. Some Administrations use strict limits to prevent interference, while others use more relaxed limits considered more reasonable for manufacturers yet providing a level of protection as required by those Administrations. Some require manufacturers to meet limits, while other only apply these standards when interference is experienced. Some Administrations accept a level of liability for interference caused by equipment meeting international standards and therefore set stricter limits. Those accepting no such liability find relaxed limits more usefull. Some national policies are related to issues of geography and population density. Many of the differences in limit appliation are very difficult to reconcile. Therefore the process of reaching international agreement on ISM limits has been difficult. There follows a tabulation of CISPR limits proposed in 1989. (Annex III contains a description of how they were derived.) It was decided to adopt the limits as given below although some administrations have applied the limits specified by CISPR in 1975 and still had a preference for them.

9.2 Radiation limits

These radiation limits are the same as those proposed by CISPR in 1989. When reading the following Tables V, VI, VII and VIII, it should be noted that:

- CISPR has taken a statistical approach in calculating the likelihood of interference when deriving these limits. So even when equipment complies with these limits it may still cause radio interference under certain circumstances. In such cases additional measures may be required;

- some administrations have expressed their concern that measurements in the horizontal direction would not account for emissions which could propagate in the vertical direction and potentially interfere with aviation systems. However, CISPR (CISPR/B/WG1(Secretariat)44) concluded, based on an analysis of the problem, that there would be no significant difference between radiation in the horizontal or vertical direction; on the other hand, measurements carried out by the Ohio University for the United States Federal Aviation Administration indicate that radiation in the vertical direction from ISM equipment can exceed those in the horizontal direction by as much as 20 - 40 dB;

- Table VIII is giving special limits for the protection of specific safety services in particular areas. Regarding the protection of other safety services and the protection of certain frequency bands which might require additional protection (due to their assignment for public safety, search and rescue and emergency use), further studies are required. Information on this matter is given in Annex I.

TABLE V

Radiation limits for Class B equipment measured on a test site

Frequency Range (MHz)	Measured on a test site	
	Class B Group 1 dB (μV/m) Test distance 10 m	Class B Group 2 dB (μV/m) Test distance 10 m
0.15 - 30	under consideration	under consideration
30 - 80.872	30	30
80.872 - 81.848	30	50
81.848 - 134.786	30	30
134.786 - 136.414	30	50
136.414 - 230	30	30
230 - 1000	37	37

Note: Limits for frequencies below 150 kHz and above 1 GHz are still to be established

TABLE VI

Radiation disturbance limits for Group 1 Class A equipment

Frequency Range (MHz)	Group 1 Class A measured on a test sote 30 m test distance (dB(μV/m))	Group 1 Class A measured at 30 m distance outside building from exterior wall
0.15 - 30	under consideration	under consideration
30 - 230	30	30
230 - 1000	37	37

Note: Limits for frequencies below 150 kHz and above 1 GHz are still to be established

TABLE VII

Radiation disturbance limits for Group 2 Class A equipment

Frequency Range (MHz)	Group 2 Class A limits at 30 m	
	From exterior wall outside building housing the ISM equipment	On a test site (dB(μV/m))
0.15 - 0.49	75	85
0.49 - 1.705	65	75
1.705 - 2.194	70	80
2.194 - 3.95	65	75
3.95 - 20	50	60
20 - 30	40	50
30 - 47	48	58
47 - 68	30	40
68 - 80.872	43	53
80.872 - 81.848	58	68
81.848 - 87	43	53
87 - 134.786	40	50
134.786 - 136.414	50	60
136.414 - 156	40	50
156 - 174	54	64
174 - 188.7	30	40
188.7 - 190.979	40	50
190.979 - 230	30	40
230 - 400	40	50
400 - 470	43	53
470 - 1000	40	50

Note 1 - The lower limit shall apply at the transition frequencies.

Note 2 - Limits for frequencies below 150 kHz and above 1 GHz are still to be established

TABLE VIII

Protection of specific safety services in particular areas

Frequency Band (MHz)	Limit (dB μV/m)	Measuring Distance Outside Building from Exterior Wall (m)
0.2835 to 0.5265	65	30
74.6 to 75.4	30	10
108 to 137	30	10
242.95 to 243.05	37	10
328.6 to 335.4	37	10
960 to 1215	37	10

10. Information on complaint statistics

The different formats of and entries into RFI statistics in different countries make it difficult to compare data from various sources. However, it can be seen from the responses to a questionaire on RFI from ISM equipment which was sent to all participants in the work of CCIR Study Group 1 and CISPR members in 1980/81 that the number of verified complaints on interference from ISM equipment is small in all countries, not only in absolute terms but also in comparision with both the total number of verified complaints and the total amount of ISM installations (see Table IX).

However, these figures may not distinguish different radio services and do not represent the actual number of incidents of interference. The number of interference incidents is often much greater than the reported number of sources. Only a percentage of people suffering from interference do complain and, in cases of complaints, very often the inferference source cannot be identified. The real interference potential is therefore not known. Future investigations are necessary to obtain more information about the relation between the number of complaints and the real interference potential.

10.1 Services actually or potentially affected (Table X)

At present the Broadcasting Services, the Land and Aeronautical Mobile Services are mainly affected. As far as the Broadcasting Service is concerned, the LF, MF, and VHF bands are the most critical ones. In future, the Broadcasting Satellite Service in the 12 GHz band might be affected by the 5th harmonic of ISM equipment operated in the 2450 MHz band.

10.2 Main sources of ISM generated interference (Table X)

The main sources of interference are harmonics of ISM-designated frequencies, and ISM equipment which are operated outside the designated ISM bands.

10.3 Compliance with present CISPR limits

The Federal Republic of Germany, Japan and Sweden (for medical equipment) are the only countries where more than 90 % of the ISM equipment are assumed to comply with CISPR limits. In most countries, only a small percentage of ISM equipment installed comply with CISPR limits (Table X). However the number of interference complaints referring to ISM sources is also small in those countries. This indicates that relaxations in relation to the existing CISPR limits might be possible in some bands and in some areas.

TABLE IX - Complaints on interference from ISM equipment

No.	country/ organization	number of verfied complaints on interference from ISM equipment/year	total number of verified complaints from all man-made sources of interference per year	total amount of installed ISM equipment	figure in column 2 as percentage of column 3	figure in column 2 as percentage of column 4	year of interference investigation
	(1)	(2)	(3)	(4)	(5)	(6)	(7)
1	Denmark	= 40 *	= 2,000 *	4,000	2	1	1980
2	Finland	8	1,200		0,67		1986
3	France	= 100	27,824		= 0.36		1979
4	Germany	76	10,143	= 270,000	0.75	<0.03	1982
	Fed.Rep.of	113	11,175	= 310,000	1.0	>0.04	1983
5	Japan	543	31,000	11,920,000 including microwave ovens	1.75	<0.005	1980
6	Netherlands	72	5012	-	1.44	-	1983
		51	5018		1.02		1984
7	Sweden	48	(= 2,000) *	-	2-3	-	average over 1978-80
8	UK	170	22,987	123,000	0.74	0.14	average over 6 years 1978-83
9	USA	107	75.815	275,000	0.14	<0.04	average over 1978-80

* data not contained in the response to the questionaire, but calculated on basis of other data available

TABLE X - Services affected and sources of interference

No.	Country (1)	Services affected (2)	Main source of interference (3)	Compliance with CISPR radiation limits (4)	Causes of interference (5)
1	Denmark	LMOB,Coast radio stations, Radionavigation equipment, BC (Sound and TV)	1)	10 % of all ISM equipment	1)
2	Finland	LMOB TV	harmonic radiation of ISM equipment		1)
3	France	broadcasting (television and sound) aeronautical mobile aeronautical navigation maritime navigation land mobile networks (safety services)	ISM equipment operated outside the destignated ISM bands harmonics of ISM equipment wood glue drying RF - excited arc welders (LF - MF bands)	10 % of all ISM equipment	incorrect setting of the oscillator fundamental frequency, frequency pulling with work load variation, inadequate use
4	Germany Fed.Rep.of	BC, LMOB AERMOB	harmonics of ISM equipment, spark erosion equipment, ultra sound equipment, RF excited are welders	90% of all ISM equipment comply either at the specified distance from the equipment or at the specified distance from the industrial area	CISPR limits are not met
5	Japan	TV and Sound BC, BC-SAT	inductive heater di-electric heater	99% (including microwave ovens)	lack of immunity of sound and television broadcasting receivers, frequency related susceptability; audio rectification, caused by non-linear circuits in receivers and other electronic entertainment devices

TABLE XII (continued)

No.	Country (1)	Services affected (2)	Main source of interference (3)	Compliance with CISPR radiation limits (4)	Causes of interference (5)
6	Netherlands	BC, AERMOB LMOB	1)	1)	lack of immunity of interfered audio and video equipment
7	Sweden	all services	Wood glue drying, Plastic seam welders, Diethermy, RF-excited arc welders, Surface erosion of metals, Drying non-metals	medical equipment: 100% other ISM equipment: very few meet the limits	1)
8	United Kingdom	BC Services in the LF/MF/VHF and UHF bands, MOB in the VFH band, Aeronautical communication in the VHF band, BC-SAT, Amateur Radio Services, ILS radio-astronomy	RF excited are welders (LF, MF bands), plastic seam welders, (UHF bands) spark eroders	30% of all ISM equipment	incorrect setting of the oscillator fundamental frequency, frequency pulling with work load variation, removal of screening, inadequate maintenance or use
9	USA	TV-, AM-, FM-, BC, CB, Amateur, LMOB, AERMOB	4th and 5th harmonics of 27 MHz ISM equipment due to poor maintenance	majority of ISM equipment does not comply with CISPR limits	poor maintenance
10	Zimbabwe	TV	2nd harmonic of 27 MHz ISM equipment	1)	1)

1) No information available

ANNEX I

Additional protection required for specific frequency bands

In its studies on the protection requirements for specific Safety Services the CCIR concluded that Table VIII as given in section 9.2 of this report, did not cover all specific safety services, e.g. those in the Maritime Mobile Service. it was also recognized that a number of other frequency bands may require additional protection.

Table XI enumerates a number of frequency bands for Safety Services and suggests radiation limits. However, this table does not include all maritime safety services, likely to be used in coastal waters.

Table XII specifies frequency bands which might require additional protection in particular areas.

Both tables are intended to initiate further studies. Particularly CISPR is invited to continue its studies on this matter."

TABLE XI

Frequency bands for safety services and suggested limits

Frequency band (MHz)	Limit (dbμV/m)	Measuring distance outside building from exterior wall (m)
2.170 to 2.194	55	30
156 to 162.5	30	10
406 to 406.1	37	10
1544 to 1545	37	10
1645.5 to 1646.5	37	10
In Region 1		
1.625 to 1.635	55	30
1.800 to 1.810	55	30
2.160 to 2.170	55	30

TABLE XII

Frequency bands which might require additional protection in particular areas

Frequency Band	Type of Services	Center Frequency
70.0- 129.0 kHz	Radionavigation (Decca navigator, LORAN C)	
495.0- 505.0 kHz	International distress frequencies for radiotelegraphy	500 kHz
1820.0-1880.0 kHz	Radionavigation (LORAN A)	
2173.5-2190.5 kHz	International distress and calling for radiotelephony	2182 kHz
3021.5-3027.5 kHz	Aeronautical mobile (search and rescue operations)	3023 kHz
4123.5-4129.5 kHz	Aux. frequency for 2182 kHz	4125 kHz
4177.0-4178.0 kHz	International distress frequencies for narrow-band direct-printing telegraphy	4177.5 kHz
4207.0-4208.0 kHz	International distress frequencies for digital selective calling	4207.5 kHz
5678.5-5684.5 kHz	Same as 3023 kHz	5680 kHz
6213.5-6219.5 kHz	Aux. frequency for 2182 kHz	6215 kHz
6267.5-6268.5 kHz	Same as 4177.5 kHz	6268 kHz
6311.5-6312.5 kHz	Same as 4207.5 kHz	6312 kHz
8289.5-8295.5 kHz	Distress and safety communications using radiotelephony	8291 kHz
8362.5-8368.5 kHz	Search and rescue for manned space vehicles	8364 kHz
8376.0-8377.0 kHz	Same as 4177.5 kHz	8376.5 kHz
8414.0-8415.0 kHz	Same as 4207.5 kHz	8414.5 kHz
12288.5-12294.5 kHz	Same as 8291 kHz	12290 kHz
12519.5-12520.5 kHz	Same as 4177.5 kHz	12520 kHz
12576.5-12577.5 kHz	Same as 4207.5 kHz	12577 kHz
16418.5-16424.5 kHz	Same as 8291 kHz	16420 kHz
16694.5-16695.5 kHz	Same as 4177.5 kHz	16695 kHz
16804 -16805 kHz	Same as 4207.5 kHz	16804.5 kHz
74.8 - 75.2 MHz	Aeronautical Radionavigation (marker beacons)	75 MHz
108 - 118 MHz	Aeronautical Radionavigation (instrument landing system, VHF omnidirecyional range)	
118 - 137* MHz	Aeronautical mobile (VHF air/ground communications)	
121.45 - 121.55 MHz	Emergency frequency	121.5 MHz
156.509 - 156.541 MHz	Distress and safety communications using digital selective calling	156.525 MHz
156.7625- 156.8375 MHz	Distress communications used internationally and safety frequency frequency	156.8 MHz
242.95 - 243.05 MHz	Emergency frequency	243 MHz
328.6 - 335.4 MHz	Aeronautical Radionavigation (ILS)	
406 - 406.1 MHz	Emergency Position Indicating Radio Beacon (Earth-to-space)	
960 -1215 MHz	Aeronautical Radionavigation (DME, SSR)	
1300 -1350 MHz	Aeronautical Radionavigation (ground radar)	
1544 -1545 MHz	Distress and safety communications for moving satellite (Space-to-Earth)	
2700 -2900 MHz	Aeronautical Radionavigation (ground radar)	
4200 -4400 MHz	Aeronautical Radionavigation (radio altimeter)	
5000 -5250 MHz	Aeronautical Radionavigation (microwave landing system)	
9000 -9200 MHz	Aeronautical Radionavigation (ground radar)	
9300 -9500 MHz	Aeronautical Radionavigation (airborne weather radar, ground radar, radar beacons)	

* 144 MHz in Region 1 only

ANNEX II

Radiation from ISM equipment

This Annex contains the results of measurements obtained at a number of factory sites in the United Kingdom.

TABLE XIII

Summury of induction machine results

FREQ. (kHz)	RATED POWER (kW)	Fieldstrength of fundamental 30 m from equipment dB(μV/m)	Relative level of harmonic fields to fundamental (dB)			Attenuation of field with range (dB/decade of distance)	Attenuation index 'n'
			f_2	f_3	f_4		
430	25	78	-18	-25	--	80.47	4.02
365	45	102	-52	-52	-72	121.1	6.15
368	30	98	-68	-68	--	72.57	3.65
380	30	95	--	--	--	100.56	5
770	8	89	-10	-38	--	69.27	3.46
940	8	91	-35	-41	--	71.39	3.57
327	45	100	--	--	--	55.26	2.76
360	45	106	-40	-56	-66	81.02	4.05
314	50	91	-50	-65	--	80.47	4.02
316	30	92	-62	-62	-62	51.28	2.56
406	45	89	-47	-29	--	101.4	5.07
461	45	74	--	--	--	93.31	4.66
325	45	84	-54	-34	--	87.41	4.37
443	30	90	-24	-40	-60	66.47	3.32
510	30	88	-52	-58	--	83.28	4.16
425	30	71	-45	-45	--	93.26	4.66
435	25	67	-33	-41	--	75.6	3.78
365	25	76	--	--	--	59.18	2.96
169	50	84	-64	-64	-34	80.88	4.04
Mean value (x)			-44.3	-47.9	-58.8	80.82	4.007
Standard deviation (σ_{n-1})			17.5	13.8	14.6	17.17	0.87

TABLE XIV - Summary of 27 MHz machine results

SITE	MACHINE CATEGORY	RATED POWER (kW)	Fieldstrength of fundamental 30 m dB(μV/m)	Relative level of harmonic fiels to fundamental (dB)					Attenuation with distance (dB/decade)	Attenuation index 'n'
				f_2	f_3	f_4	f_5	f_9		
1	Food dryers (x5)	55 to 80	96	-23	-12	-17	-10	-23	38.59	1.93
2	Glue dryers (x3)	12	83	-30	-27	-11	-5	-26	33.7	1.69
3	Plastics Welders (x12)	6 to 17	104	-46	-51	-59	-54	-54	30.58	1.53
Mean value (x)				-33	-30	-29	-23	-34.3	34.29	1.71
Standard deviation (σ_{n-1})				11.8	19.7	26.2	27.0	17.0	4.04	0.20

TABLE XV - Relative level of harmonic fields at 100 m

SITE	Relative level of harmonic fields to fundamental (dB)				
	f_2	f_3	f_4	f_5	f_9
1	-26	-19	-21	-13	-22
2	-30	-30	-14	+2	-29
3	-45	-45	-46	-23	-38
Mean value (x)	-33.7	-31.3	-27	-14.7	-29.7
Standard deviation (σ_{n-1})	10	13	16.8	17.6	8

ANNEX III

Derivation of CISPR ISM limits

The CISPR Recommendations for methods of measurement and limits of interference embodied in CISPR Publication 11 have not found universal acceptance. CISPR has therefore taken the decision to review the provisions of CISPR Publication 11 to take into account comments which have been received from various sources and to reflect more accurately present day conditions.

Limits in Publication 11 were based on a model derised in the late 1950s to take into account the views and requirements of the users of the spectrum at that time. The first action of CISPR Sub Committee B was therefore to revise this model and to extend it by the introduction of an allowance for probability.

The CISPR model is based on a simple equation as follows:

$$L = F_s - P_r + d + p$$

Where L = Limit dB(μV/m) at a specified distance measured in a specified manner

F_s = minimum value of fieldstrength to be protected dB(μV/m)

P_r = Protection ratio dB

$$d = 20 \log_{10} \frac{d2^n}{d1^n}$$ where d_1 = measurement distance

d_2 = distance from the ISM equipment at which the minimum fieldstrength is to be protected

n = attenuation exponent.

p = an allowance in dB, to take account of a number of factors which modify the worst case limits to give a practical limit for the control of interference.

The factors include:

I) only a proportion of receivers are situated in areas of maximum wanted fieldstrength,

II) the radiation pattern of ISM apparatus is not normally isotropic and the receiving antenna may not be in the direction of maximum radiation,

III) there may be discrimination against the interference by virtue of antenna directivity and plane of polarisation,

IV) the ISM harmonic may not be co-frequency with the wanted signal or may not be in the most sensitive part of received passband,

V) the ISM and the receivers potentially affected may not be operating at the same time,

VI) the ISM may operate infrequently or for short periods only at full power.

In the control of interference it is desirable, from the point of view of the protection of the spectrum and for economic and other practical reasons, for equipment to be suppressed prior to sale or hire to the end user. For these reasons the CISPR model also includes provisions for the calculation of limits for measurement on a standard test site. The provisions include the insertion of an allowance for the attenuation of the buildings which house the ISM equipments and a factor to take into account the differences between measuring conditions on a standard test site and in the normal environments of commercial and industrial use of equipment.

Use of the CISPR model

Table XIII shows how the CISPR ISM model has been used for the calculation of the revised CISPR limits. In the calculations the values of minimum fieldstrength to be protected were those specified by the ITU, CCIR, ICAO or other expert body for the relevant service; protection ratios were determined by additionally taking into account the nature of the ISM emanations. Studies were being carried out in various countries.

The nature of CISPR limits is such that to achieve the desired goal, which is the protection of radio services by harmonised standards which are derived to avoid the introduction of technical or administrative barriers to trade, compromises must be made and single values have had to be adopted from data which in many instances have large statistical spreads. The determination of these single values has been a matter of some difficulty and the values eventually adopted were not always justifiable on a strictly engineering basis. Individual numerical values for the various components of the CISPR model used in the calculation of limits are therefore compromises arrived at by discussion.

It can also be seen by comparing the calculated limits in Table XIII with the limits proposed in CISPR/B(CO)19 (Table VII to Table X) that further compromises were introduced to meet specific practical difficulties, which, unless taken into account would have prevented the possibility of the adoption of the proposed limits by all Administrations and National Committees. A typical example of such a compromise was the inclusion of limits more stringent than calculated in the television bands generally in exchange for a more released limit at the 3rd, 5th and 7th harmonic of the 27.12$\pm$ 0.6 MHz ISM band.

In the evolution of the model the CISPR found it necessary to adopt a grouping and classification of equipment and limits to enable legal and administrative differences in the various countries to be accommodated and to ensure adequate protection of radio services without unnecessarily causing all ISM apparatus to meet stringent limits.

The classifications and groupings are explained in the main report.

Allowance for probability

The allowance for probability for services other than safety of life services forms a major portion of the limit. This may be seen from examination of the calculation for Band I Television in the CISPR calculation.

TABLE XVI - Calculation of Class A Limits for Group 2 ISM apparatus

Source: CISPR/B(Secretariat) 39

Frequency band (MHz)	Service to be protected	Signal to be protected dB(μV/m)	Protection ratio (dB)	Permissible interference field at receiving antenna dB(μV/m)	Distance from equipment at which signal is to be protected (m)	Attenuation law	Approximate equivalent interference field at 30 m from equipment dB(μV/m)	Building attenuation (dB)	Allowance for probability (dB)	Corresponding practical limit at 30 m from boundry dB(μV/m)	Corresponding limit at 30 m on a test site dB(μV/m)
1	2	3	4	5	6	7	8	9	10	11	12
0.150 to 0.285	LF BC	77	30	47	45	$d^{-2.7}$	57	10	30	87	97
	Aero Beacons	37	0	37	300	$d^{-2.8}$	93	10	NIL	93	103
0.285 to 0.490	Aero Beacons	37	0	37	300	$d^{-2.8}$	93	10	NIL	93	103
0.490 to 1.605	MF BC	56	30	26	45	$d^{-2.7}$	36	10	30	66	76
	Aero Beacon	37	0	37	300	$d^{-2.7}$	93	10	NIL	93	103
1.605 to 4.00	Fixed Links	20	0	20	1000	$d^{-2.0}$	81	10	10	91	130
	Aero Mobile	28	0	28	300	$d^{-2.0}$	68	10	NIL	68	107
4.00 to 15.00	Fixed Links	14	0	14	1000	$d^{-2.0}$	74	10	10	84	123
	Aero Mobile	21	0	21	300	$d^{-2.0}$	61	10	NIL	61	100
15.00 to 20.00	Fixed Links	6	0	6	1000	$d^{-2.0}$	67	10	10	77	116
	Aero Mobile	16	0	16	300	$d^{-2.0}$	56	10	NIL	56	95
20.00 to 30.00	Fixed Links	6	0	6	1000	$d^{-2.0}$	67	10	10	77	116
	Aero Mobile	0	0	0	300	$d^{-2.0}$	40	10	NIL	40	79
30.00 to 68.00	TV BC Land	48	50	-2	45	$d^{-2.0}$	5	10	30	35	74
	Mobile	5	-18	23	100	$d^{-2.0}$	44	10	NIL	44	83

TABLE XVI (Continued)

1	2	3	4	5	6	7	8	9	10	11	12
68.00 to 100.00	Aero Beacons	63	20	43	30	$d^{-1.0}$	43	10	NIL	43	53
	FM BC	60	36	24	45	$d^{-2.0}$	31	10	30	61	100
	Land Mobile	5	-18	23	100	$d^{-2.0}$	43	10	NIL	43	82
100.00 to 156.00	FM BC	60	36	24	45	$d^{-1.8}$	30	10	30	60	93
	ILS	46	6	40	30	$d^{-1.0}$	40	10	NIL	40	50
	Aero Mobile	39	0	39	300	$d^{-1.0}$	59	10	NIL	59	69
	Land Mobile	5	-18	23	100	$d^{-1.8}$	42	10	NIL	42	75
156.00 to 174.00	Land Mobile	5	-42	47	100	$d^{-1.8}$	66	10	NIL	66	99
174.00 to 216.00	TV BC	55	50	5	60	$d^{-1.8}$	16	10	30	47	79
	Land Mobile	6	-42	48	100	$d^{-1.8}$	67	10	NIL	67	100
216.00 to 400.00	ILS	46	6	40	30	$d^{-1.0}$	40	10	NIL	40	50
400.00 to 470.00	Fixed Links Land Mobile	26	0	26	1000	$d^{-1.8}$	81	10	10	91	124
		Figures not yet available									
470.00 to 585.00	TV BC	64	50	14	75	$d^{-1.3}$	24	10	30	54	72
585.00 to 614.00	Aero Nav	10	0	10	1000	$d^{-1.0}$	40	10	NIL	40	50
	TV BC	64	50	14	75	$d^{-1.3}$	24	10	30	54	72
614.00 to 854.00	TV BC	70	50	20	75	$d^{-1.3}$	30	10	30	60	78
854.00 to 960.00	Land Mobile	Figures not yet available									

* Relaxed by 10 dB at 81.36 and 135.6 MHz

NOTE: The method of calculation and the various values contained in this Annex should be considered provisional until later confirmation by additional CCIR studies.

Tabling: 48dB(uV/m) for F_s, 50dB for the protection ratio, 7dB allowance for the distance and 30dB allowance for probability leads to a limit of 35dB(uV/m) ($L=F_s-P_r+d+p$).

There is no simple method for the determination of a value to be allotted to probability in the CISPR model. However, first results of a study carried out in the USA using a computer simulation have shown that, in the simulation, which takes account of the factors listed as I), II), V) and VI) in the probability list, a change in limit of 30dB has little effect on the probability of interference.

The USA work does not take into account the effect of TV receiver antenna directivity and this factor was therefore taken into account in a study carried out in the UK to determine the distance d_1 at which protection is necessary. In the use of the model to calculate limits the CISPR proposed the use of the 30 dB probability allowance from the USA study and took into account the effect of receiver antenna directivity by adopting a distance d of 45 m for Band I television.

The study of probability is still actively proceeding.

REPORT 1101

PRELIMINARY STUDIES OF RADIO INTERFERENCE LIMITS FOR MAC/PACKET DECODERS

(Question 4/1)

(1990)

1. Introduction

1.1 This report presents the preliminary results of CCIR studies that have been carried out on the electromagnetic compatibility between television equipment employing MAC/packet decoders, and aeronautical/maritime emergency (distress) systems using Emergency Locater Transmitters (ELTs) and Emergency Position-Indicating Radio Beacons (EPIRBs).

1.2 MAC/packet decoders, in common with other digital television equipments, will use clock frequencies based on the internationally agreed digital television coding standard (Recommendation 601). This standard has been in use since 1981 and applies to equipment which may be used in television studios, for outside broadcast purposes and for medical systems incorporation picture processing, in addition to domestic TV receivers.

1.3 The CCIR world-wide standard for digitally coded video signals is based on a sampling clock frequency of 13.5 MHz. This particular frequency was carefully chosen to permit orthogonal sampling of both 625 line/50 Hz and 525 line/60 Hz systems (i.e. picture samples on adjacent television lines are vertically above one another in both systems) and is thus constrained to be a harmonic of both television line frequencies employed in the various TV systems currently in use around the world.

1.4 The 625 line/50 Hz MAC family of television systems (C, D and D2-MAC/packet) employs time compression and expansion, such that the highest clock frequency in the MAC decoder is 3/2 times that of the basic CCIR standard, i.e. 20.25 MHz. Other digital processing circuits for TV receivers may use a clock frequency related to 13.5 MHz, for example, 40.5 MHz.

1.5 The radio interference limits for electrical and electronic equipment are the subject of several existing and developing specifications [Annex A]. These limits are intended to ensure electromagnetic compatibility between such equipments and radiocommunication services.

1.6 In the case of digital TV equipment, using a sampling frequency based on 13.5 MHz, there are three factors which merit special attention in considering electromagnetic compatibility (EMC):

- Certain harmonics of the clock frequencies coincide with the centres of the emergency frequency channels at 121.5 MHz and 243 MHz in the aeronautical mobile frequency bands. In general the emissions from digital systems will consist of a discrete line spectrum related to the clock frequency and its harmonics together with a broadband noise-like spectrum produced by the data. It should be noted that any clock frequency below 34 MHz will have harmonics in the aeronautical mobile (R)/aeronautical radionavigation bands 108 - 137 MHz, MAC and MAC/packet radiation are such cases. Its harmonics will fall within the band 108 - 117.975 MHz of the aeronautical-radionavigation service, which is a safety service and within the band 117.975 - 137 MHz of the aeronautical-mobile (R) service, which contains the aeronautical emergency frequency 121.5 MHz and within 1 MHz of the satellite EPIRB frequency band 406 - 406.1 MHz. Analysis of potential interference to services in these bands (except for 121.5 and 243 MHz) are not included in this document [see Report 1169].

- The potential number of TV receivers employing digital processing in future use may be very large.

- All receivers tuned to the same channel signal will be locked to exactly the same clock frequency and all receivers using any MAC/packet channel will radiate within 20 Hz of each other at the fundamental.

Because of this combination of factors, and the vital importance of protecting the operation of emergency services from unwanted interference, studies have been conducted in order to investigate the necessary interference limits for MAC decoders and associated equipment.

1.7 It should also be noted that potential interference is not limited to MAC/packet systems and that other types of equipment specifically ISM may also generate interference (see 2.2 of Annex C).

1.8 These studies took into consideration a full deployment of the MAC/packet television system, which is likely to occur around 15-30 years in the the future. By then, the development of the COSPAS/SARSAT system may have reached its fourth generation. For the future it will also be necessary to take into account the great uncertainties concerning the utilization of 121.5 MHz and the manner in which it might be treated on-board the satellites.

2. Operation of aeronautical/maritime emergency, search and rescue services

2.1 SARSAT (Search and Rescue Satellite-Aided Tracking) is a part of the international COSPAS/SARSAT programme that is intended to facilitate prompt detection and location of aeronautical and maritime distress events. This capability is accomplished by placing a special instrument package aboard polar orbiting satellites. This instrument package contains equipment that receives and processes low-power signals emitted by ELTs and EPIRBs [COSPAS-SARSAT, 1981].

The aeronautical mobile band is divided into channels which are normally used for AM voice communication. One of these channels, centered on 121.5 MHz, is reserved for use as an emergency frequency for distress and calling. A second, harmonically related frequency, 243 MHz, in the fixed and mobile band is also reserved for the same purpose. The permitted use of these frequencies has been broadened to include use by EPIRBs.

2.2 Reception of ELTs may be at a ground station, on a ship, on an aircraft or on a satellite. Such reception may be used for alert and location, for example via one of the polar low-orbiting satellites of the SARSAT and COSPAS programmes [ICAO, 1986], or using search and rescue aircraft.

2.3 New ELT/EPIRB systems for satellite reception are now being introduced, operating at 121.5, 243 and 406 MHz frequencies; the 406 - 406.1 MHz band is being exclusively reserved for these emergency operations. The 406 MHz systems offer several performance advantages over existing systems, using the lower frequencies, which can suffer interference from shared use and from existing man-made sources of interference. ICAO is currently pursuing the adoption of improved characteristics for ELTs operating on 121.5 and 243 MHz [ICAO, 1988]. The characteristics being pursued are consistent with those adopted in the United States in 1985 [TSD, 1985]. Such characteristics have been adopted by the CCIR in the 1986-1990 study period.

3. Scope of studies

3.1 The frequencies 121.5 MHz and 243 MHz are the ninth and eighteenth harmonics of 13.5 MHz respectively. There is, therefore, a potential for interference.

3.2 In the case of 406 MHz, the nearest harmonic of 13.5 MHz is the thirtieth at 405 MHz.

Bearing in mind the high stability of the clock frequency (typically two parts in ten to the seventh and that the widest 1 dB bandwidth used in the 406 - 406.1 MHz band is 80 kHz, with corresponding 45 dB and 85 dB bandwidths of 120 kHz and 180 kHz), this offset of 1 MHz should ensure that the harmonic will not cause in-band interference in a 406 MHz receiver. However, receiver interference effects such as image and spurious responses (i.e. out-of-band effects) should be studied.

3.3 Studies have, therefore, been concentrated on interference at 121.5 MHz and 243 MHz. The necessary analysis consists of two main tasks as given below:

- determine the total potentially-interfering field strength which can be allowed without affecting the correct reception of the emergency-frequency signals for each particular receiving system (aircraft, satellite and ground station);

- calculate using a simulation mathematical model for several typical situations taking into account environmental factors and the number of sources deployed.

4. Presentation of analysis methods

4.1 Central to the analysis of interference potential is the calculation of the summation of the signals at the receiver from individual interfering sources. Annex B addresses this summation with respect to both the SARSAT* satellite and to terrestrial receivers (11 cases). Annex C addresses the summation of signals to the SARSAT satellite receiver only (3 cases).

4.2 Each annex approaches this summation problem from different directions. Annex B uses an analytical approach to calculate the summation of signals, while Annex C uses computer simulations. In addition there are differences in the assumptions made with regard to the following:

- manufacturers safety margin;
- satellite antenna pattern;
- interference criteria.

Other assumptions for which there is agreement are:

- number and distribution of interfering sources;
- activity factor;
- building attenuation;
- basic SARSAT parameters.

When both methods use comparable assumptions, their results agree within 1 or 2 dB thus providing confirmation of the analysis techniques.

4.3 Interference criteria

The Annex B analysis uses an interference threshold of -40 dBμV (equivalent to an increase in noise temperature of 2,880 K) which is based on the quoted minimum detectable signal. The Annex C analysis uses an interference threshold of -50.8 dBμV (equivalent to an increase in noise temperature of 240 K) which is based on noise temperatures measured by operating satellites while in view of the eastern United States.

4.4 Manufacturer's safety margin

A manufacturer's safety margin of 5.5 dB was taken into account in Annex B and not in Annex C.

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* A further analysis is needed to determine the radiation limits for the protection of the COSPAS satellite.

5. Conclusions

5.1 The analysis methods presented in Annexes B and C have used different methodologies and have reached different conclusions. However, studies [CCIR, 1986-90] have shown that if the same interference limits and parameters were used, almost identical results would be obtained.

5.2 In order to reduce the differences between the results given by the two anlaysis methods, it is necessary to carry out further studies on elements related to the interference limits and parameters. The following areas requiring further studies involving both COSPAS and SARSAT satellites, have been assembled in two groups. High priority items are:

a) analyse the distribution and density of decoders and O.B. units, by studying television ownership and activity factor on a world-wide basis, taking into account the type of system in use (cable, satellite ...) and the visibility area of 121.5 MHz receivers;

b) what is the adequate radiation limit required for the protection of SAR satellite receivers, by a comparison of signal-to-interference ratio against signal-to-noise ratio in terms of effects on sensitivity of satellite receivers, considering appropriate limits and the mechanism involved;

c) a study to identify noise temperature values for the satellite system around the globe, especially for Europe;

d) the overall probability of significant degradation to ELT Doppler trace due to MAC/packet interference taking into account: FFT algorithms, processing speed, threshold levels, effects of 121.5 MHz voice communication and other interference;

e) investigate "aircraft crash attenuation factor", which considers signal losses due to crash configuration and conditions;

*f) a study is required, which should quantify the electromagnetic emission in the hemisphere above the equipment in its normal mode of operation. The decrease of the field with distance and position should also be studied;

*g) a study of cabinet radiation parameters to develop a model describing how a cabinet radiates;

h) building attenuation values, with both inside and outside sources, should be studied in both horizontal and vertical planes, and for all types of construction;

i) study present manufacturer's safety margins,

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* The Director of the CCIR should ask CISPR "What is the most appropriate test method for determining the full three dimensional radiation pattern of operational domestic equipment?" in order to more accurately determine the source radiation pattern factor Ks [Annex B, § 3.4.1].

and the lesser priority items are:

j) study Search And Rescue (SAR) operations to identify the appropriate search aircraft height and visibility range to be used in propagation calculations;

k) as decoder population increases, carry-out flight tests over areas of various concentration to validate theoretical analysis of interference emissions - this study cannot be carried out at the present time;

l) studies of airfield arrangements around the world in order to determine the appropriate distance between the ground-based receivers and the O.B. units at their potential location;

m) potential interference to 406.05 MHz satellite receivers need to be studied, to verify that out-of-band effects do not occur;

n) the expected growth of the decoder population in relation to the expected reduction (or growth) of the 121.5 MHz beacon population should be determined - this study cannot be carried out at the present time;

o) a study to compare the use of the maximum antenna gain curve against a distribution approach, for example a Monte Carlo technique to obtain the gain in the direction of interest.

5.3 A revision of the analyses using the revised assumptions obtained above and a consideration of the overall resultant probabilities, should be made.

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ANNEX A

RADIATION LIMITS FOR TV EQUIPMENT

General limits for interference from broadcast receivers are treated in CISPR Publication No. 13 (1975). The measurement method conforms with that specified in CISPR Publication No. 22. The radiation limits currently applied to digital TV equipment used in the domestic environment are as follows:

CISPR Publication No. 13 (CENELEC EN55013)

Frequency (MHz)	Measuring distance (m)	Radiation Limit (dBμV/m)
121.5	3	40
243	3	47

In the case of digital studio equipment, the limits are as follows:

CISPR Publication No. 22 (class A) (CENELEC EN55022)

Frequency (MHz)	Measuring distance (m)	Radiation Limit (dBμV/m)
121.5	30	30
243	30	37

ANNEX B

ANALYTICAL APPROACH TO ANALYSIS OF POTENTIAL INTERFERENCE BY SPURIOUS EMISSION FROM MAC/PACKET DECODERS

1. Introduction

1.1 The analysis methodology considered in this annex was developed by the EBU [CCIR, 1986-90, a] and a more complete description of the theoretical background may be found in Report 1102. This analysis method was considered during the Interim and Final Meetings [1988 and 1989] and a number of modifications were proposed. The text presented below is a consolidation of the original EBU proposal and the modification already agreed.

1.2 The data collected and studied by the EBU for aircraft and ground stations were obtained from RTCA Documents 183 and 186 [RTCA, 1983] and from experimental simulations. In the case of the SARSAT system, data were based on ICAO data [ICAO, 1986].

Data were also collected on the characteristics of the spurious emissions from representative equipments, and on the effect of the environment (i.e. buildings, location of equipment, etc.). Statistical methods and computer models were developed to estimate the effects of a distributed array of sources.

2. Analysis

2.1 In general a distress receiver is evaluated at a height h, above a circular area of radius R containing N sources. The analysis may be split into three parts:

- establish what interfering signal voltage "V" can be allowed at the input of the distress receiver;
- compute the summation of all interfering sources within the defined area, taking into account building effect and other relevant factors, and assuming an emission limit based on CISPR Publication No. 22 (class B);
- determine whether any modification to the limit is desirable in order to ensure compatibility.

2.2 The following cases were assessed:

a) Aircraft/city: An aircraft flying over a city (population one million and radius 15 km) at a height of 10 km.

b) Aircraft/large city: London was chosen as representing a worst-case example of a). In this case, the population is eight million and the radius 28 km.

c) Aircraft/rural: A search aircraft at a lower altitude (780 m) above a rural area. This corresponds to a worst-case situation, an ELT being received at a range of 115 km.

d) Aircraft/studio: An aircraft flying over a TV studio at a height of 10 km. The studio was assumed to contain 1,000 sources of interference. This caters for a possibility of several studios in one town.

e) Aircraft/O.B. unit: An aircraft flying over an outside broadcast (O.B.) unit at a height of 10 km. The number of interfering sources was taken as ten.

f) Satellite/global: A polar orbiting SARSAT satellite was at an altitude of 850 km and assuming a uniform distribution of interfering sources appropriate to a population of 600 million.

g) Satellite/large city: The effect of a worst-case concentration of sources falling within the SARSAT receiver bandwidth. London (population eight million and radius 28 km) was taken for this case.

h) Satellite/studio: The effect of a TV studio (1,000 sources) on the SARSAT receiver.

i) Satellite/O.B. unit: The effect of an outside broadcast unit (10 sources) on the SARSAT receiver.

j) Ground station/city: A terrestrial receiver (e.g. air/ground communications receiver, VHF direction-finding receiver) located adjacent to a town.

k) Ground station/O.B. unit: An outside broadcast unit at a distance of 400 m from a terrestrial receiver.

These cases require additional analysis to improve the validity of the conclusions that have been drawn from them. In particular the assumptions for the aircraft height, the night-time noise level and the number of active sources and the cumulative effect of other numbers of these active sources should be investigated.

3. Allowable interfering signal voltage V

3.1 Aircraft receiver

3.1.1 Document [RTCA, 1983] shows that the interfering signal must be limited to -120 dBm at the receiver input terminals in order not to interfere with reception of distress signals. The table below indicates noise levels taken calculated from measurements of antenna noise temperature (including the effects of man-made noise).

Altitude	Population	Noise in a 25 kHz bandwidth (dBμV)	
		Around 121.5 MHz	243.0 MHz
10 km	1 million	-6	-2
10 km	8 million	+3	+7

3.1.2 For an ELT at the limit of range (115 km) and reduced power (5 mW), the above limit of interference would ensure a signal-to-interference (S/I) ratio of 13 dB [RTCA, 1983].

3.1.3 The power of 5 mW may not represent the worst-case situation. In an aircraft's final crash configuration, damage to the ELT antenna and terrain shielding effects may reduce the signal radiated from the crash site; studies [RTCA, 1983 and Taggart et al., 1971] indicate that a crash attenuation factor could be up to 20 dB, but the power radiated under these circumstances is not yet known.

3.2 Satellite receiver

3.2.1 For current production SARSAT receivers, the minimum detectable signal levels are -148.9 dBm for 121.5 MHz signals and -151.8 dBm for 243 MHz signals; -148.9 dBm corresponds to the power from a radiator with an e.i.r.p. of -16 dBm at a distance of 850 km, assuming isotropic antennas. In the study, a dipole satellite antenna was assumed which results in a value of -147 dBm (-40 dBμV) at the satellite receiver. Since a dipole was also used in the summation formula in § 3.4, the use of -40 dBμV will not affect the results obtained.

3.2.2 At 121.5 MHz, SARSAT antenna gains vary from -7 dBLi to +3 dBLi. At 243 MHz, SARSAT antenna gains vary from -7 dBLi to -1 dBLi.

3.2.3 An ELT at a worst-case distance of 2,450 km and worst-case power of 5 mW (7 dBm) will produce a receiver voltage of -26 dBμV. Provided the total interference is less than -40 dBμV, this ensures:

- no possibility of a false alarm;
- signal-to-interference ratio > 14 dB.

3.2.4 This calculation takes into account the power of the emission. In practice the signal processing in the SARSAT system might use only the carrier power. The S/I ratio mentioned above needs further study. The SARSAT operational characteristics also need to be taken into account.

3.3 Ground receiver

3.3.1 In this case, the interfering signal must be limited to -120 dBm (-13 dBμV) at the receiver input terminals (the same as for an aircraft receiver).

3.4 Summation of sources

3.4.1 The formula [Report 1102, Equation 15] for computing the summation of many interfering sources is as follows:

$$V_{rms} = C_A E_L K_S d_L \sqrt{K_D N}\ 10^{-\frac{m_B + m_m}{20}}\ e^{(\sigma^2_B + \sigma^2_m)\left(\frac{\ln 10}{20}\right)^2}\ \sqrt{E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right]}$$

where v_{rms} = root-mean square voltage at receiver input (μV)

C_A = receiving antenna factor

E_L = specified field-strength limit (μV/m) at a distance d_L (m) for a single interfering source (e.g. CISPR limit)

K_S = source radiation pattern factor

K_D = reduction factor due to Doppler effect

N = total number of active interfering sources

m_B = mean value of building effect (dB)

σ_B = standard deviation of building effect (dB)

m_m = mean value of manufacturer's safety margin (dB)

σ_m = standard deviation of manufacturer's safety margin (dB)

d_i = distance of source i to receive antenna

$a(d_i)$ = antenna gain in direction of source i at distance d_i.

$E\left[\left(\frac{a(d_i)}{d_i}\right)^2\right]$ = expectation value of the square of the antenna gain/distance factor $a(d_i)/d_i$

3.4.2 It should be noted that the summation of the signals gives a Rayleigh distribution which may be characterized by its r.m.s. value as calculated above. It was assumed that the Rayleigh distributed signal has the same interfering effect as already existing noise having the same root-mean square voltage.

3.4.3 The factor "building effect" was treated as a log-normal probability distribution.

3.4.4 A cosine law satellite antenna pattern was assumed, and this combined with the free space propagation equation to yield a signal attenuation factor which reduces received power level with the fourth power of d_i. Future satellites will most likely employ shaped beams which maximize the gain at or near the horizon, thus modifying the model-calculated received signal power. Calculations should be repeated employing the expected gain pattern of the next generation satellites.

3.5 Data used in the computation

3.5.1 Field strength produced by a single source E_L, d_L

3.5.1.1 In each case, the CISPR Publication No. 22 (class B) limit was used, i.e. E_L = 30 dB (μV/m) (37 dB (μV/m) at 243 MHz) measured at a distance d_L = 10 m.

3.5.2 Source radiation pattern factor K_S

3.5.2.1 Equipments will be required to meet the limit in § 3.5.1 in the direction of maximum field strength. However, the field strength in the direction of the distress receiver will in general be less due to the random orientation of receivers. A factor of 3 dB has been adopted to take account of this source radiation pattern effect. Thus K_S equals 0.7.

3.5.2.2 When sources are available, the radiation pattern should be measured around the hemisphere above the television receiver in its operational position. A representative value of K_S may then be derived for use in the equation in 3.4.1. It is also necessary to ensure that the values measured are not subject to near-field effects or cabinet radiation effects [CCIR, 1986-90, b] and measurements at greater distances than 3 metres as specified in CISPR Publication No. 13 will be needed for verification. The question was raised whether the field strength in the CISPR limit was to be considered as produced by a source of radiation equivalent to an antenna. The propagation law might differ from an inverse distance law and decrease more rapidly. Further studies in this area are therefore highly desirable.

3.5.3 Doppler effect K_D

3.5.3.1 The SARSAT polar orbiting satellite travels at a speed of 7 km/s and this results in a Doppler frequency shift of the interfering signal.

3.5.3.2 Satellite receivers have bandwidths as shown in the table below:

Frequency (MHz)	Bandwidths (kHz)		
	1 dB	20 dB	80 dB
121.5	25	35	92
243.0	46	64	180

3.5.3.3 Some Local User Terminals (LUTs), which are ground-based receivers used to process data transmitted from the SARSAT satellites, have signal processing bandwidths of 50 Hz at 121.5 MHz and 100 Hz at 243 MHz; other LUTs may have different processing bandwidths. Since each incremental signal processing bandwidth is extremely narrow (50 Hz), only a small fraction of all the interfering sources over a large area will be detected in any single processing filter.

3.5.3.4 It can be shown [CCIR, 1986-90, a] that, at 121.5 MHz, a bandwidth of 50 Hz is equivalent to a strip approximately 14 km wide beneath the satellite. Thus, although the satellite receiving antenna can see an area of radius 3,651 km, the received interference power in a processing band is reduced by as much as 20 dB. In the case of a large city (e.g. London), the attenuation is about 5 dB.

3.5.4 Number of active sources N

3.5.4.1 It was assumed that the number of TV receivers is equal to the population divided by three, based on statistics of existing TV receivers in Europe. In addition, an activity factor of 25% was assumed to allow for sets

which are switched off or tuned to a non-MAC terrestrial service. This results in a number of interfering sources N equal to the population divided by 12. This density represents a saturation level of DBS receivers and is unlikely to be approached for 15-30 years.

3.5.4.2 The factor 4 comes from the working assumption that one in four households would contain an active MAC/packet receiver at a given time. An activity factor of 25% seems to be reasonable in the light of the fact that some television receivers may be switched off or may not be of the MAC/packet type.

3.5.4.3 To avoid unnecessary pollution of the electromagnetic environment, it is highly desirable to ensure that MAC/packet decoders are not powered-on when a receiver is in a standby condition or switched to receive a non-MAC/packet transmission.

3.5.4.4 The factor 3 is obtained from the statistics of existing television receivers within the EBU Member countries; the statistics are yearly published in the EBU Review-Technical. These statistics show that the ratio between the number of inhabitants and the number of television receivers is not lower than three in most European countries. This factor does not change very much with the time since it represents the saturation situation and it has been therefore recognized to be a standard value in the planning of terrestrial broadcasting networks.

3.5.4.5 More studies are needed on TV ownership and activity factor, taking into account the type of receiving system in use, e.g. cable systems versus direct connect satellite TV.

3.5.5 Building effect m_B, σ_B

3.5.5.1 The effect of buildings is to produce both reflection and attenuation of the signal. Data has been taken from measurements reported by Mir and White [1974] at a frequency of 150 MHz. These give:

mean value	m_B = 22 dB
standard deviation	σ_B = 12.5 dB.

3.5.5.2 It should be noted that the standard deviation of this factor has a large influence on the value of v_{rms}. For example, a building effect with a mean value of 22 dB and a standard deviation of 12.5 dB will have the same effect as a mean value of 4 dB and standard deviation of 0 dB.

3.5.5.3 It was considered by some administrations that other building attenuation models might be more appropriate. Further study is required to compare certain models given in the references with the given model for building attenuation in order to conclude on the appropriateness of this model. In addition, the implications of the statistical distribution used to represent the building effect needs further study.

3.5.6 Manufacturers' safety margin m_m, σ_m

3.5.6.1 Equipments in large-scale production will be tested on a sampled basis. The procedure recommended by CISPR ensures that at least 80% of the items comply with the limit with at least 80% confidence. Europe has had experience in using the 80%-80% rule for over a period of more than ten years.

3.5.6.2 The manufacturer, in order to minimize his own testing, will apply a more stringent limit such that he has a higher confidence level in his product meeting the CISPR limit. The difference between these limits, i.e. the manufacturers' safety margin, is typically 6 dB with a standard deviation of 2 dB. Hence

mean value $m_m = 6$ dB

standard deviation $\sigma_B = 2$ dB.

3.5.6.3 It is understood that this safety margin may be used only where the limits above are obligatory for the manufacturer. It is also noted that CISPR measurements may not take into account the effects discussed in § 3.5.2.2.

3.6 Results of computation

3.6.1 Using the data as described above, the interfering signal has been computed for each case a) to k) using the method given in Report 1102 . The results are given in Table I.

TABLE I

Results of calculation of total interference margin at 121.5MHz and 243.0 MHz

Case	height[1] h (km)	radius R (km)	population	number of active sources N	calculated rms voltage at antenna terminals[3] (dBμV)	interfering limit[3] (dBμV)	margin[3] (dB)
a) Aircraft/city	10	15	10^6	$8.3.10^4$	-8 (-7)	-6(-2)	+2(+5)
b) Aircraft/London	10	28	8.10^6	$6.5.10^5$	-4 (-3)	+3(+7)	+7(+10)
c) Aircraft/rural	0.78	115	5.10^6	4.10^5	-18 (-17)	-13	5 (4)
d) Aircraft/studio	10	0.1	-	10^3	-22 (-21)	-6(-2)	+16(+19)
e) Aircraft/O.B. unit	10	0.01	-	10	-42 (-41)	-13	29 (28)
f) Satellite/global	850	3651	6.10^8	5.10^7	-47 (-46)	-40	7 (6)
g) Satellite/London	850	28	8.10^6	$6.5.10^5$	-38 (-37)	-40	-2 (-3)
h) Satellite/studio	850	0.1	-	10^3	-64 (-63)	-40	24 (23)
i) Satellite/O.B.	850	0.01	-	10	-81 (-80)	-40	41 (40)
j) Ground station/ city	0.01	15	$1.2.10^6$	$8.3.10^4$	-7 (-6)	-6(-2)	+1(+4)
k) Ground station/ O.B. unit[2]	0.4	0.01	-	10	-14 (-13)	-13	1 (0)

Note 1 - For cases j) and k), height = separation distance.

Note 2 - A study carried out by one administration indicated that 50 m was more appropriate [CCIR 1986-90, c].

Note 3 - Values in brackets relate to 243 MHz.

3.6.2 In Table I, the margin is defined as a difference between the specified interference limit and the calculated r.m.s. voltage at the antenna terminals of the aircraft, satellite or ground station. In the case of positive margin, no risk for interference exists. In the one case, where there is a small negative margin, interference may be just detectable. The largest negative margin is 3 dB and this is for satellite reception in the case of London. Since there will be very few cities in this category, and their locations are well-known, this would not appear to justify a reduction in the emission limit. Further studies are needed to note whether a Doppler trace is lost or significantly degraded or not.

3.7 Effect of dispersions of ELT frequencies

3.7.1 Although the short-term ELT frequency deviation is very small, the dispersion of the ELT frequencies covers several hundreds of Hz. On the other hand, the ninth harmonic of the MAC/packet clock is precisely on 121.5 MHz. The satellite signal processing bandwidth of 50 Hz is not centred on 121.5 MHz precisely, where the interference occurs, but has a dispersion of approximately ten times that bandwidth. In addition, interference due to the Doppler effect occurs only during a percentage of the time. Taking these effects into account, further study is required to evaluate the global interference probability.

4. Conclusions

4.1 The studies above have indicated that the radiation limits that should be applied to MAC/packet decoders and other digital TV equipment to be used in the domestic environments are as follows:

Frequency (MHz)	Measuring distance (m)	Radiation limit dB (μV/m)
121.5	10	30
243.0	10	37

This limit is identical with the existing CISPR Publication No. 22 (Class B) and CENELEC EN 55022 radio interference limit for information technology equipment at these two frequencies.

4.2 This study indicates that the radiation limits quoted above will ensure electromagnetic compatibility with the aeronautical service in possible distress situations.

4.3 In the case of digital studio equipment, the limit is given in CISPR Publication No. 22 (Class A) and CENELEC EN 55022 limits at the frequencies 121.5 and 243 MHz as follows:

Frequency (MHz)	Measuring distance (m)	Recommended radiation limits dB(μV/m)
121.5	30	30
243.0	30	37

This limit is effectively 10 dB higher than the CISPR (Class B) limit because it is measured at 30 m instead of 10 m. Some administrations are concerned about these limits.

4.4 In the case of an aircraft receiver, it has been demonstrated from the studies that have been carried out that the existing man-made noise degrades the sensitivity over densely populated areas.

4.5 These results are based on studies considering the situation taking into account a full deployment of the MAC/packet television system, which is likely to occur around 15-30 years in the future.

4.6 Based on the values found in Table I, there are no combinations of scenarios a) - k) of section 4 which will yield significantly lower margins.

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ANNEX C

SIMULATION APPROACH TO ANALYSIS OF POTENTIAL INTERFERENCE TO SARSAT BY SPURIOUS EMISSIONS BY MAC/PACKET DECODERS

1. Introduction

1.1 The analysis methodology considered in this annex was developed by the United States [CCIR 1986-90, b] and a more complete description of the theoretical background may be found in [Miller and Wallace, 1989].

2. Background

2.1 The SARSAT receiving and signal processing sub-systems have been designed to achieve the maximum thermal noise-limited sensitivity available today in order to reliably detect and locate distress events. Because the desired signals received by the SARSAT instrument package may be relatively weak, the SARSAT sub-systems are susceptible to interference.

2.2 Interference to the SARSAT system can be caused by the radiation of spurious signals in the bands of interest when the radiators are in view of the satellite. Such radiators include transmitters operating in adjacent or harmonically related bands, automobiles and machinery, electronic equipment (e.g., computers), consumer entertainment equipment, and other sources.

2.3 The International Special Committee on Radio Interference (CISPR) has established limits for spurious radiations from electrical and electronic equipment. For Class B equipment (which is typical of consumer equipment), the applicable radiation limits for information technology equipment are given in Table 1 [CISPR Publication 22, 1985]. These limits are designed to provide protection to the local environment from individual units, not the aggregate effect from many.

Table 1
CISPR RADIATION LIMITS
(Publication 22)

Frequency	Maximum Electric Field (Quasi-peak)
30 - 230 MHz	30 dB(μV/m) at 10 meters
230 - 1000 MHz	37 dB(μV/m) at 10 meters

2.4 This annex examines the effects of spurious emissions from MAC/packet decoders on the performance of the SARSAT system operating at 121.5 MHz and 243 MHz. Radiations are assumed to be at the CISPR limits of Table 1. As explained in section 5, the results documented herein indicate that the CISPR limits are inadequate [CISPR Publication 22, 1985].

3. SARSAT system characteristics

3.1 The SARSAT system consists of the space segment, the distress alerting transmitters (ELTs and satellite EPIRBs), and the signal processing sub-system.

3.2 Space segment

3.2.1 The space segment consists of a satellite in a circular, near-polar orbit at a nominal altitude of 850 km. It carries a SARSAT instrument package consisting of Earth coverage receiving antennas, receivers, signal processing sub-systems (406.05 MHz only), and a space-to-Earth transmitter, to receive, process, and relay signals transmitted by ELTs and satellite EPIRBs operating at 121.5, 243 and 406.05 MHz. Signal processing at the receiving earth station/Local User Terminal (LUT) is used to detect the occurrence of a distress event and to determine its location (see Report 761).

3.2.2 The spacecraft receiving antennas point toward nadir (straight down) and exhibit a circularly symmetric radiation pattern in order to receive linearly polarized emissions from ELTs and EPIRBs in any orientation. Figures 1 and 2 show the antenna gain patterns referenced to a linearly polarized signal (dBLi) measured at the receiver input, for 121.5 MHz and 243 MHz, respectively [COSPAS-SARSAT, 1988]. Equation (1) is used for the gain function for both 121.5 and 243 MHz antennas,

$$G(dBLi) = a_0 + a_1\Gamma + a_2\Gamma^2 \qquad (1)$$

where Γ (radians) is the angle off nadir and the values for a_0, a_1 and a_2 are given in Table 2. These approximations are least mean-square error, second-order curve fits (to within about 0.2 dB) to the gain functions shown in the figures.

Table 2: Coefficients for Antenna Gain Formula (1)

Frequency (MHz)	a_0	a_1	a_2
121.5	02.96	00.04	-03.51
243 (Γ<45 deg)	-03.01	04.06	-01.88
243 (Γ>45deg)	-28.95	66.92	-40.05

3.2.3 The receiving system noise temperature depends on a number of factors, including the operating frequency, time of day and type of area within view of the satellite. Post-launch measurements of the receiving system noise temperature were made on all NOAA satellites launched to date. The results of these measurements are shown in Table 3 [NASA: a, b, c, d, e]. The table shows the minimum and maximum noise temperatures measured in each series of tests. The tests were performed at the NASA Goddard Space Flight Center, United States, with the satellite viewing the eastern (industrial) United States. The noise temperature for NOAA-8 at 121.5 MHz was increased by electromagnetic interference from on-board instruments. As can be seen in Table 3, this problem does not exist on later satellites. On the basis of these data, minimum noise temperatures of 2,000 K for 121.5 MHz and 800 K for 243 MHz were assumed for analysis. Further study is needed to quantify noise temperature values in other parts of the world.

Table 3: Measured receiving system noise temperature.

SARSAT Satellite	Reference	Measured Noise Temperature (K)	
		121.5 MHz	243 MHz
NOAA-8	NASA, a 19	3630-18200	800
NOAA-9	NASA, b 20	1320-4680	1050-4680
NOAA-9	NASA, c 21	2009-2903	not measured
NOAA-10	NASA, d 22	2100-3200	500-800
NOAA-11	NASA, e 23	1182-3981	851-5495

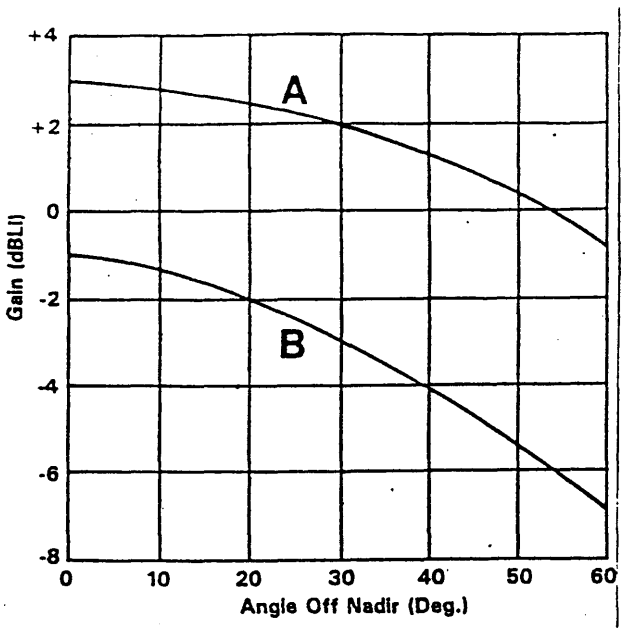


Figure 1. 121.5 MHz spacecraft antenna gain [COSPAS-SARSAT, 1988]
A: Maximum gain over 90% of region.
B: Minimum gain over 95% of region.

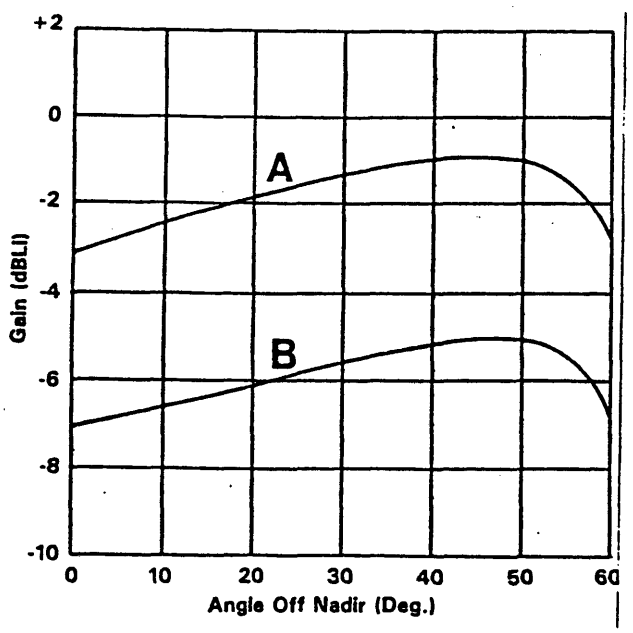


Figure 2. 243 MHz spacecraft antenna gain [COSPAS-SARSAT, 1988]
A: Maximum gain over 90% of region.
B: Minimum gain over 95% of region.

4. Analysis

4.1 An analysis was performed [Miller and Wallace, 1989] to estimate the level and nature of the signal received by a SARSAT receiver in three cases from two assumed distributions of emitters (MAC/packet decoders and studio equipments) within the satellite's field of view. The distributions assumed were 1) widespread, where emitters are randomly distributed over the entire field of view with a constant average number per square km; and 2) concentrated, where emitters are distributed uniformly throughout a circular area representing an isolated city, both on-nadir and off-nadir.

4.2 Satellite antenna pattern

The satellite antenna gain in all cases was assumed to be independent of azimuth angle and equal to the upper curves (maximum gain over 90% of the region) of Figure 1 or 2 as a function of the angle off-nadir. In the analysis, the gain was calculated using quadratic formulae that fit these curves to within ±0.2 dB. Further study is needed to compare this assumption with a statistical approach.

4.3 Uniformly distributed interferers

4.3.1 For the widespread distribution, a Monte Carlo computer analysis was used in which results from a large number of individual-emitter trials were combined. In each trial, an emitter location was selected at random, and the power and frequency of the signal received by the satellite were calculated. The histogram of the received power versus frequency over many trials then represented the power spectral density of the aggregate signal received from all emitters at once. Received power was calculated for each trial as e.i.r.p. (assumed constant at the radiation limit less 3 dB for random orientation) + building attenuation + propagation path loss + satellite receiver antenna gain. Signal frequency was determined by the relative emitter/satellite velocity and consequent Doppler shift.

4.3.2 Figure 3 shows the results of two preliminary simulations for an operating frequency of 121.5 MHz (using a simplification of Equation (1)) in order to determine the comparison between two building attenuation models. The power spectral density (PSD) is plotted as a function of Doppler shift for interferers uniformly distributed over the field of view of the satellite. Only half of the curve is shown due to symmetry about zero Doppler.

4.3.3 The solid curve in Figure 3 is the result of 2×10^6 trials. For each trial, a random value for building attenuation (log normal distribution, $\mu = 22$ dB, $\sigma = 12.5$ dB [CCIR, 1986-90, a]) was added to the path loss. The dotted curve in the figure is the result of 5×10^5 trials with the constant value of 4 dB used for the building attenuation. Because the two curves agree within ±2 dB, all subsequent analyses are based on the simpler approach of a constant building attenuation of 4 dB.

4.3.4 Figure 4 shows the results of 10^6 trials for 121.5 MHz, with two antenna patterns. The upper curve corresponds to the curve fit of Equation (1). The lower curve in Figure 4 assumes a dipole antenna gain [CCIR, 1986-90, a].

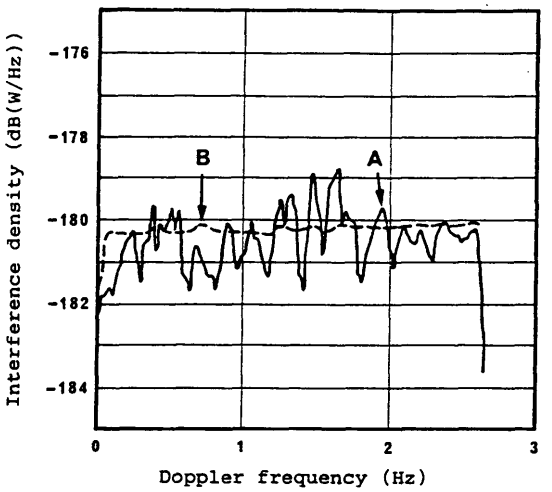


FIGURE 3 - PSD at the input to the 121.5 MHz receiver

(1 W total radiated power, antenna gain computed by equation (2) of [Miller and Wallace, 1989])

A: log-normal building attenuation
B: constant building attenuation, 4 dB

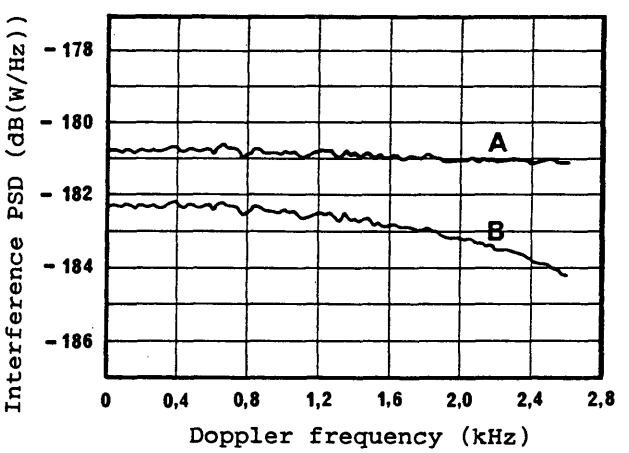


FIGURE 4 - PSD at the input to the 121.5 MHz receiver

(1 W total radiated power)

A: antenna gain = equation (1)
B: dipole antenna

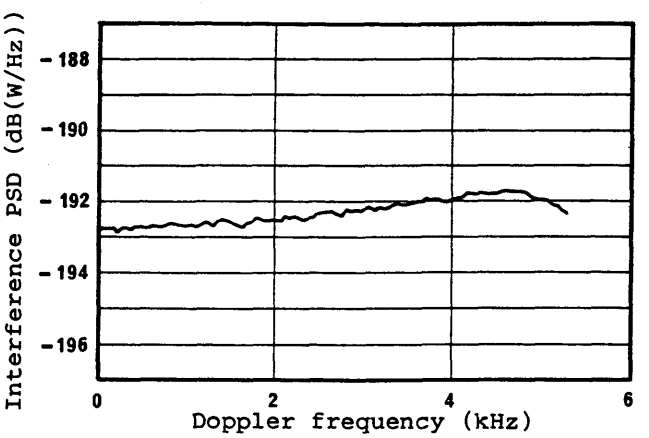


FIGURE 5 - PSD at the input to the 243 MHz receiver

(1 W total radiated power, antenna gain from equation (1))

4.3.5 Figure 5 shows the results of 10^6 trials for an operating frequency of 243 MHz, using the antenna gain given by Equation (1). The interferers are uniformly distributed over the field of view of the satellite. As seen, the PSD at 243 MHz is 11-12 dB less than the PSD at 121.5 MHz. This is consistent with the use of a satellite receiving antenna with 1-6 dB less gain at 243 MHz than at 121.5 MHz, a 6 dB reduction caused by the increased basic transmission loss, and a 3 dB reduction caused by the increased Doppler spread. The results reflect a constant 4 dB building penetration loss.

4.3.6 Referring to Figures 4 and 5, assume that there are 50×10^6 interferers uniformly distributed over the field of view and that each is radiating an e.i.r.p. of -87.8 dBW (equivalent to the CISPR limit for 121.5 MHz - 3 dB allowance for random orientation of the sources). The total radiated power from all sources added together is then $10\log(50 \times 10^6)$ - 87.8 dBW = -10.8 dBW. Figures 4 and 5 were calculated with a normalized total radiation level of 0 dBW. Consequently, the PSD determined by the simulation curve of Figure 4 would be decreased by 10.8 dB to give the total received power at the input to the receiver of (-180.8 - 10.8) = -191.6 dBW/Hz for emitters at the CISPR Publication No. 22 limit.

4.4 Interference from cities

For the concentrated distribution simulation, the received power level is calculated assuming that all emitters are colocated at the centre of the "worst-case city" (see Annex B, § 2.2). The shape of the PSD vs. frequency curve takes into account the distributed nature of the emitters within the city's circular boundary, with a radius of 28 km.

4.5 Increase in receiving system noise temperature

The effective increase in the receiving system operating noise temperature is an appropriate measure of the degree of degradation that this interference will have on the SARSAT system's performance. The increase in the effective noise temperature, delta T, due to interference is

$$\text{delta } T = PSD/k \qquad (2)$$

where PSD is the power spectral density (Watts/Hz) at the input to the receiver due to interference and k is Boltzmann's constant (1.380×10^{-23} J/K or -228.6 dB (W/Hz-K)).

5. Interference criteria

5.1 In section 3 it is shown that the radiations from a large number of MAC/packet units will be received aboard the satellite as a noise-like signal. The effect of this noise-like interference may be to measurably degrade the detection threshold thus reducing the probability of detecting and locating a distress event.

5.2 Analysis of real-world ELT signals received by the SARSAT satellites [King, 1986] has shown that a large percentage of the signals processed exhibit D/N_0 values that are near the processing threshold. Table 4 gives signal statistics observed in a seven-week period in 1983 for the NOAA-8 spacecraft. These data show that the SARSAT system operates near its threshold.

5.3 The data show the ability of the SARSAT system to detect and locate a distress event is noise-limited. For such noise-limited systems, the accepted criterion for specifying a permissible level of interference is in terms of the increase in the received noise power (or equivalently the receiver operating noise temperature) caused by the interference. This criterion is used by services seeking protection from harmful interference caused by spurious radiation from other systems.

Table 4. Received Signal C/N Statistics for NOAA-8 Spacecraft [King, 1986]

Frequency (MHz)	C/N_0 Threshold (dB-Hz)	C/N_0 Percentiles		
		25th	50th	75th
121.5	14.0	16.5	19.5	24
243	15.5	16.5	19.0	22

5.4 For example, in the radio astronomy service, passive measurements of thermal radiations of extraterrestrial origin are made using high-gain, low-noise receiving systems that incorporate sophisticated signal combining and signal processing techniques [see Report 224]. The level of harmful interference adopted by the radio astronomy service is the level which introduces an error of 10% or more in the measurement of either the power received from the extraterrestrial source or the noise temperature of the source.

5.5 The fixed-satellite service has allocated 20% of the noise budget for a telephony channel to interference from other FSS networks and 10% of the noise budget to interference from terrestrial systems operating in the band on a primary, coequal basis [see Recommendations 353, 356 and 466].

5.6 The maritime mobile-satellite service has allocated 12% of the overall noise performance objective for a telephony channel to in-band interference from other MMSS systems and from terrestrial systems operating on a primary, coequal basis [see Report 917].

5.7 Electromagnetic interference (EMI) from other equipment sharing the spacecraft with the SARSAT receiver could, like interference from the ground, increase the effective noise temperature and degrade the system's sensitivity. Stringent design goals for EMI levels within the receiving bands have been established to prevent such degradation [King, 1984]. They are based on a maximum degradation of 0.5 dB in the system threshold, which is equivalent to a 12% increase in the system noise temperature.

5.8 In the light of the above[1], MAC/packet radiations will be allowed to increase the receiving system's noise power by 0.5 dB or noise temperature by 12% before considering it to be harmful interference in this analysis. This limit may need further study with respect to this interference criterion.

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1 MAC/packet radiations are tolerated on a non-interference basis and have no allocation status in these frequency bands.

6. Interference analyses

This section contains the results of an analysis of interference to SARSAT operations at 121.5 MHz and 243 MHz that was performed using the approach outlined in section 3 and detailed in [Miller and Wallace, 1989]. Three cases were considered: 1) interference from spurious radiations from equipments uniformly distributed over the field of view; 2) interference from equipments located within a city with the satellite overhead; and 3) interference from equipments located within a city with the satellite not overhead. The assumptions employed in this analysis are given below and are consistent with those used in Annex B. The results are compared with the 0.5 dB change in noise temperature detection threshold.

6.1 Assumptions used in the analysis

6.1.1 The e.i.r.p. of a single source is -87.8 dBW at 121.5 MHz and -80.8 dBW at 243 MHz corresponding to the CISPR limits of 30 dB(μV/m) and 37 dB(μV/m), respectively, measured at 10 metres (with 3 dB reduction for random orientation).

6.1.2 Building penetration loss is 4 dB. This value may require further study.

6.1.3 The number of active TV decoders (activity factor) is equal to the population within the field of view divided by 12. This factor assumes that 1 in 3 of the population owns a decoder and that 1 in 4 of the decoders is in use at any one time. These statistics are based on existing TV receivers in Europe and representative of a saturation density that will not likely be reached for 15-30 years [CCIR, 1986-90, a].

6.2 Results

6.2.1 Three cases with differing geometry were considered for each of the two operating frequencies:

6.2.1.1 50 million active interferers that are uniformly distributed over the field of view of the satellite.

6.2.1.2 650,000 active interferers concentrated within a city whose dimensions approximate London. The satellite passes directly overhead.

6.2.1.3 650,000 active interferers concentrated within a city whose dimensions approximate London. The city is at an arbitrary position: relative latitude of 11° and a central angle of 21°.

6.2.2 The results of the analyses are summarized in Tables 5 and 6. The power spectral density values given for the uniform distribution cases were taken from Figures 4 and 5 and adjusted as described in § 3.3.6. The tables show that the increase in noise temperature due to spurious radiation at the CISPR Publication No. 22 limits exceeds the degradation threshold by from 10.3 to 19.5 dB at 121.5 MHz and from 10.3 to 15.5 dB at 243 MHz.

Table 5: Potential interference to SARSAT at 121.5 MHz.

Case	Uniform Dist.	28 km City (At nadir)	City Off nadir
Number of active sources	50×10^6	650×10^3	650×10^3
Received power (dBW)		-163.2	-177.2
Peak Doppler Spread (Hz)	2654	99.1	27.3
Power spectral density (dBW/Hz)	-191.6	-185.3	-194.5
Equivalent voltage (dBμV/50Hz)	-37.6	-31.3	-40.5
Noise temperature increase (K)	5012	21,380	2570
Permissible increase in noise temperature (K)	240	240	240
Excess increase in noise temperature (dB)	13.2	19.5	10.3

Table 6: Potential interference to SARSAT at 243.0 MHz.

Case	Uniform Dist.	28 km City (At nadir)	City Off nadir
Number of active sources	50×10^6	650×10^3	650×10^3
Received power (dBW)		-168.4	-179.2
Peak Doppler Spread (Hz)	5307	198.2	54.6
Power spectral density (dBW/Hz)	-195.6	-193.3	-198.5
Equivalent voltage (dBμV/50Hz)	-41.6	-39.3	-44.5
Noise temperature increase (K)	1995	3388	1023
Permissible increase in noise temperature (K)	96	96	96
Excess increase in noise temperature (dB)	13.2	15.5	10.3

Table 7. Maximum Radiation Limits from the above study to protect SARSAT

Frequency (MHz)	Measuring Distance (meters)	Radiation Limit (dB(μV/m))
121.5	10	10
243.0	10	21

6.3 Spurious radiation threshold levels

The emission levels given in Table 7 will ensure that interference from MAC/packet decoders and studio equipment does not cause the noise level in the SARSAT receivers to increase above the threshold level of 0.5 dB (12% increase in noise temperature). This results in radiation levels which are lower than the CISPR Publication No. 22 limits by 20 dB at 121.5 MHz and 16 dB at 243 MHz.

7. Conclusions

7.1 The analysis reported in this document concludes that MAC/packet TV decoders and television studio equipment radiating at the CISPR limits reproduced in Table 1 could, in the aggregate, cause harmful interference to the SARSAT system.

7.2 In the light of the different COSPAS orbit altitude, antenna gain and operating parameters, one would expect the interference limits to be different from those found for SARSAT. An analysis is needed to determine the radiation limits adequate for the protection of the COSPAS satellite.

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REPORT 1118

A STUDY ON REDUCTION IN RADIO NOISE FROM ELECTRIC POWER SUB-STATIONS

IMPORTANCE OF PROPER MAINTENANCE

(Question 46/1, Study Programme 46B/1)

(1990)

1. Introduction

Electric power sub-stations generate quite a high level of radio noise which is primarily due to gap discharge and corona breakdown. Loosely grounded connectors, faulty insulators, high resistance metal contacts and miscellaneous loose hardware may exist at any large or medium sub-station at any time which may give rise to gap discharge and corona discharge. It is therefore quite imperative that periodic maintenance of sub-stations should reduce the level of radio noise.

To study the behaviour of radio noise and effect of periodic maintenance, it is considered necessary to know:

i) the level of noise radiation from an electric power sub-station where several "hot spots" - loose contacts, bad insulators etc., may exist;

ii) the level of radio noise after elimination of such faults;

iii) the extent of reduction in radiated noise level after thorough maintenance of the sub-station.

2. Experimental study on a typical electric power substation

Experimental study was carried out on a typical 220 KV sub-station where three incoming and three outgoing lines were in existence. This sub-station was located in the suburban part of New Delhi, India. The vehicular traffic density on roads around the sub-station was very low and effects of ignition noise were thus, negligible. Moreover, there were no residential colonies, industries or any other source of man-made radio noise which could introduce errors in measurements. This power sub-station was commissioned in the year 1983. There were two bus-bars carrying 39 insulators in each bus-bar and a total of 106 joints.

3. Mode of measurement

Measurements were carried out as per CISPR publications. In the CISPR mode (i.e., Quasi-peak mode) of requirement, the microprocessor selects the 120 kHz IF bandwidth automatically. Two suitable sites were chosen after careful examination of the location of bus-bars, transformers, distribution network and gantry poles etc., in the sub-station. One site was selected between two transformers and distribution networks facing bus-bar I and bus-bar II. Another site was selected near gantry poles where several metal contacts existed. Measurements were undertaken at 20 metres distance from bus-bars at both the sites. Final readings were arrived at on the basis of raw data of more than 40 recordings at each spot.

4. Condition of sub-station

With the help of concerned organization responsible for regular maintenance and upkeep of the sub-station, a "hot spot" survey was carried out and four "hot spots" were detected apart from faults like faulty insulators. One set of measurements was carried out before maintenance of the sub-station. Subsequently all the "hot spots" were removed and faulty insulators replaced. Another set of measurements was undertaken after this maintenance exercise.

5. Weather conditions

Weather was fairly dry, sunny and clear during measurements. All the recordings were done during the daytime when the temperature was around 40-43°C and humidity was approximately 30-35%.

6. Findings

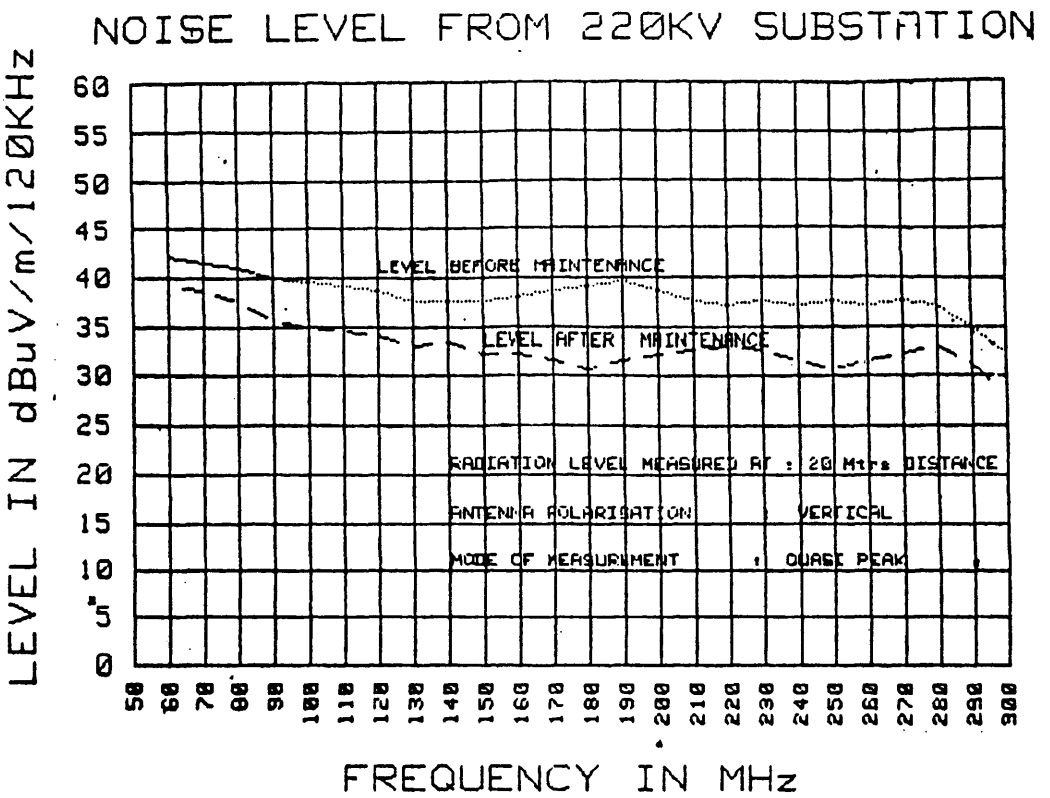
Radio noise spectra of 220 KV power sub-station is graphically presented in Figure 1 for both antenna polarizations. Radio noise radiations have been measured in the frequency band 60 MHz to 300 MHz in fair weather conditions.

Study of data collected during this exercise reveals a reduction in radio noise of the order of 10 dB after maintenance was carried out. Substantial reduction in radio noise thus can be achieved if power sub-stations are maintained properly and kept fault free.

The study indicates that regular preventive maintenance of sub-stations in order to eliminate the loosely grounded connectors, bad insulators, high resistance metal contacts and miscellaneous loose hardware, results in substantial reduction of radio noise generated in power sub-stations and thus reduces interference to radiocommunication services.

7. Conclusion

This study highlights the role of maintenance of electric power sub-stations. It must be borne in mind that electric power generation installations and distribution systems may be a source of severe interference to the radio spectrum (cf. CISPR Document 18) and that every effort has to be made to reduce and even eliminate the same.


NOISE LEVEL FROM 220KV SUBSTATION
LEVEL IN dBuV/m/120KHz
LEVEL BEFORE MAINTENANCE
LEVEL AFTER MAINTENANCE
RADIATION LEVEL MEASURED AT : 20 Mtrs DISTANCE
ANTENNA POLARISATION : VERTICAL
MODE OF MEASUREMENT : QUASI PEAK
FREQUENCY IN MHz

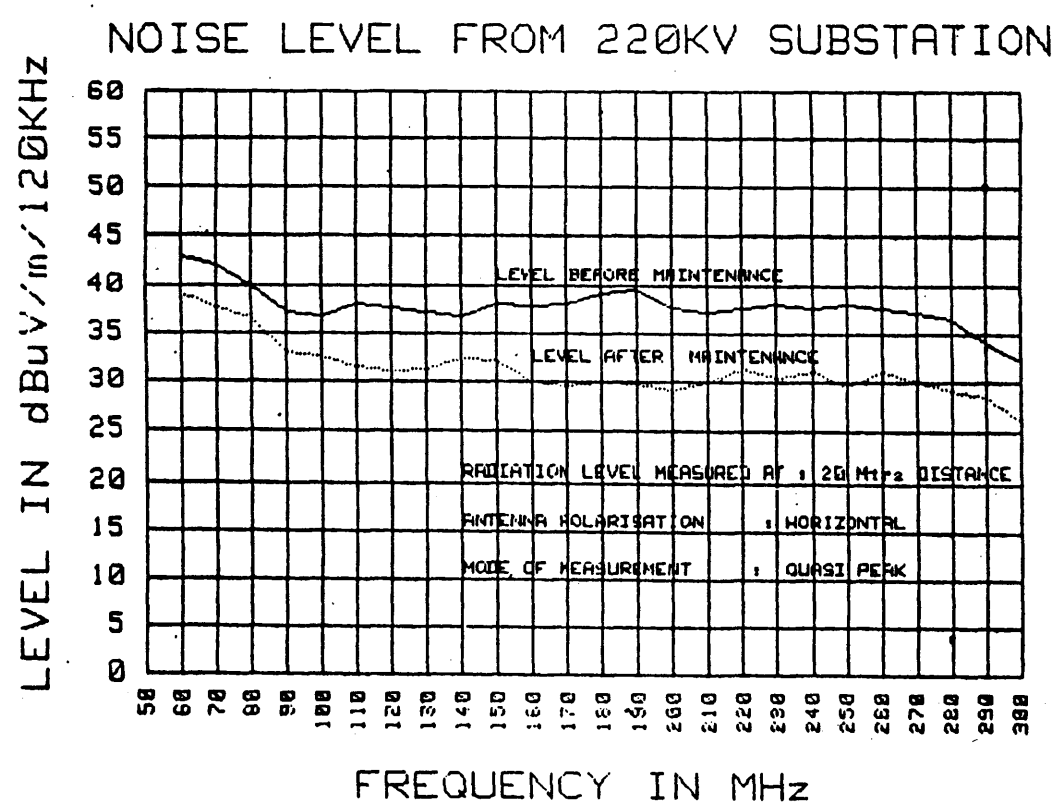

NOISE LEVEL FROM 220KV SUBSTATION
LEVEL IN dBuV/m/120KHz
LEVEL BEFORE MAINTENANCE
LEVEL AFTER MAINTENANCE
RADIATION LEVEL MEASURED AT : 20 Mtrs DISTANCE
ANTENNA POLARISATION : HORIZONTAL
MODE OF MEASUREMENT : QUASI PEAK
FREQUENCY IN MHz

FIGURE 1

REPORT 670-1

WORLDWIDE MINIMUM EXTERNAL NOISE LEVELS, 0.1 Hz TO 100 GHz

(Question 46/1, Study Programme 46A/1)

(1978-1990)

1. Introduction

This short Report is designed to specify the minimum external noise to be expected at terrestrial receiver sites from natural and man-made noise sources (excluding unwanted signals) in the frequency range 0.1 Hz to 100 GHz. The object is to be able to determine what minimum receiver noise figure, f_r, is appropriate so that a terrestrial receiving system will be close to being limited by external noise. Generally, it is inappropriate to use receiving systems with noise figures less than those specified by the minimum external noise.

2. Definitions

The overall operating noise factor, f, for a receiving system is composed of a number of noise sources at the receiving terminal of the system. Both internal and external noise must be considered. For receivers free from spurious responses, the system noise factor (see Report 413, Oslo, 1966) is given by

$$f = f_a + (l_c - 1)\left(\frac{t_c}{t_0}\right) + l_c\,(l_t - 1)\left(\frac{t_t}{t_0}\right) + l_c\,l_t\,(f_r - 1) \qquad (1)$$

where:

f_a: the external noise factor defined as

$$f_a = \frac{p_n}{k t_0\, b} \qquad (2)$$

F_a: the external noise figure defined as $F_a = 10 \log f_a$

p_n: the available noise power from a lossless antenna (the terminals of this lossless antenna do not exist physically, but this is the only appropriate point in the receiving system to reference the various noise factors).

k: Boltzmann's constant = 1.3802×10^{-23} J/K

t_0: the reference temperature in K taken as 288 K

b: the noise power bandwidth of the receiving system in Hz

l_c: the antenna circuit loss (available input power/available output power)

t_c: the actual temperature, in K, of the antenna and nearby ground

l_t: the transmission line loss (available input power/available output power)

t_t: the actual temperature, in K, of the transmission line

f_r: the noise factor of the receiver ($F_r = 10 \log f_r$: noise figure in dB)

If $t_c = t_t = t_0$, (1) becomes

$$f = f_a - 1 + f_c f_t f_r \tag{3}$$

where f_c is the noise factor associated with the antenna circuit losses,

$$f_c = 1 + (l_c - 1)\left(\frac{t_c}{t_0}\right) \tag{4}$$

and f_t is the noise factor associated with the transmission line losses,

$$f_t = 1 + (l_t - 1)\left(\frac{t_t}{t_0}\right) \tag{5}$$

Relation (2) can be written

$$P_n = F_a + B - 204 \qquad \text{dBW} \tag{6}$$

where $P_n = 10 \log p_n$ (p_n = available power in watts), $B = 10 \log b$, and $-204 = 10 \log kt_0$. For a short ($h \ll \lambda$) grounded vertical monopole, the vertical component of the r.m.s. field strength is given by:

$$E_n = F_a + 20 \log f_{\text{MHz}} + B - 95.5 \qquad \text{dB } (\mu\text{V/m}) \tag{7}$$

where E_n is the field strength in bandwidth b and f_{MHz} is the centre frequency in MHz. Similar expressions for E_n can be derived for other antennas [Lauber, 1977]. For example, for a half-wave dipole in free space,

$$E_n = F_a + 20 \log f_{\text{MHz}} + B - 99.0 \qquad \text{dB } (\mu\text{V/m}) \tag{8}$$

The external noise factor is also commonly expressed as a temperature, t_a, where, by definition of f_a,

$$f_a = \frac{t_a}{t_0} \tag{9}$$

and t_0 is the reference temperature in K and t_a is the antenna temperature due to external noise.

More detailed definitions and discussions (including the case with spurious responses) are contained in Report 413 (Oslo, 1966).

3. Results

The following three figures and the discussion for each specify the minimum expected value of F_a along with other noise levels of interest. Note that f_a is a dimensionless quantity, being the ratio of two powers. The quantity f_a, however, gives numerically the available power spectral density in terms of kt_0 and the available power in terms of kt_0b [Spaulding, 1976].

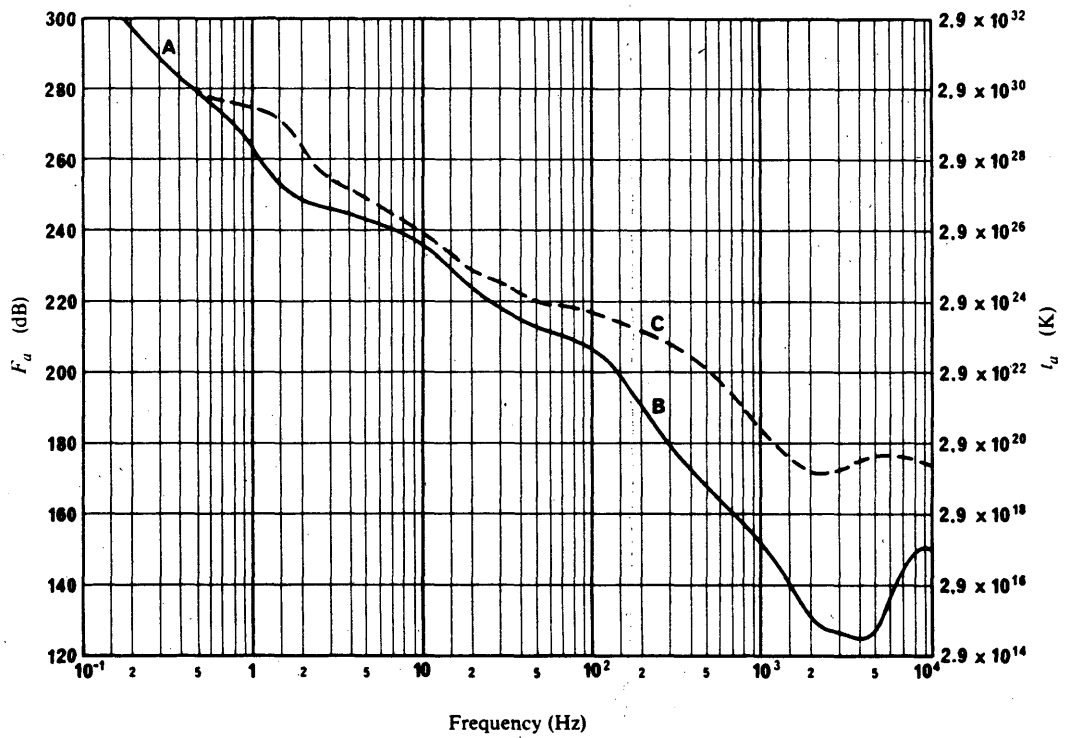


FIGURE 1 — *F_a, minimum and maximum, versus frequency (0.1 to 10^4 Hz)*

A: Micropulsations

B: Minimum value expected of atmospheric noise

C: Maximum value expected of atmospheric noise

Fig. 1 : This figure covers the frequency range 0.1 Hz to 10 kHz. The solid curve is the minimum expected values of F_a based on measurements (taking into account the entire Earth's surface, all seasons and times of day) and the dashed curve gives the maximum expected values. Note that in this frequency range there is very little seasonal, diurnal, or geographic variation. The larger variability in the 100 – 10 000 Hz range is due to the variability of the Earth-ionosphere wave-guide cutoff.

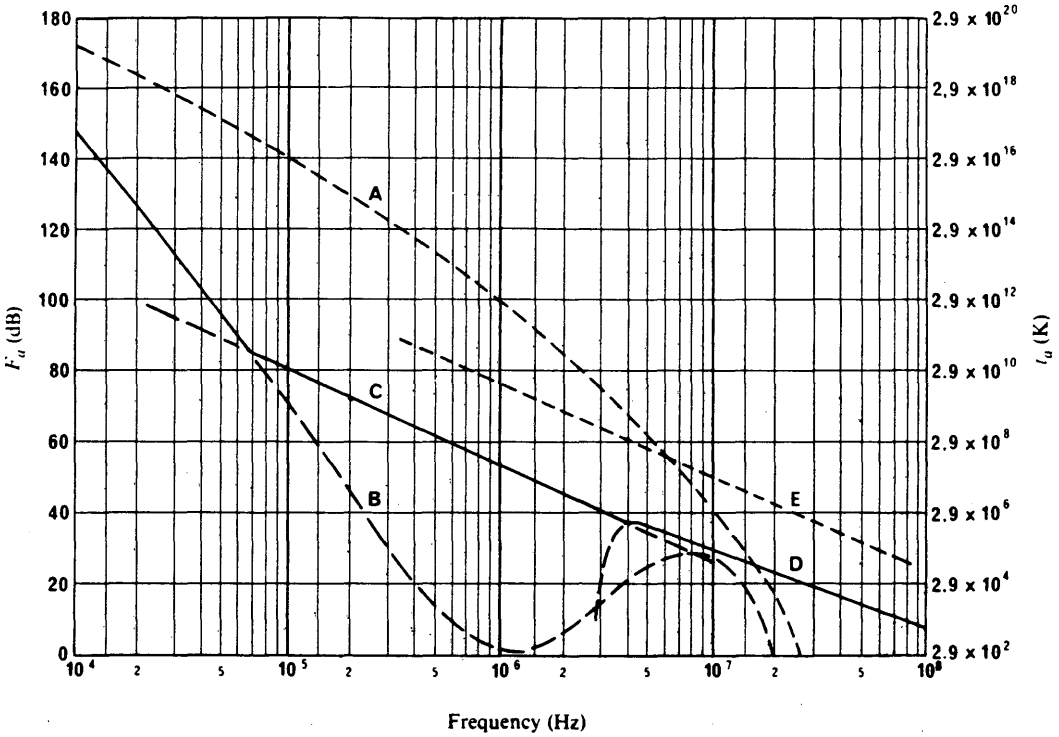


FIGURE 2 — *F_a versus frequency (10^4 to 10^8 Hz)*

A: Atmospheric noise, value exceeded 0.5% of time
B: Atmospheric noise, value exceeded 99.5% of time
C: Man-made noise, quiet receiving site
D: Galactic noise
E: Median business area man-made noise
———— : Minimum noise level expected

Fig. 2 : This figure covers the frequency range 10^4 to 10^8 Hz, i.e., 10 kHz to 100 MHz. The minimum expected noise is shown via the solid curves and other noises that could be of interest as dashed curves. For atmospheric noise, the minimum values expected are taken to be those values exceeded 99.5% of the time and the maximum values are those exceeded 0.5% of the time. For the atmospheric noise curves, all times of day, seasons, and the entire Earth's surface have been taken into account. More precise details (geographic and time variations) can be obtained from Report 322. The man-made noise (quiet receiving site) is that noise measured at carefully selected, quiet sites world-wide as given in Report 322. The atmospheric noise below this man-made noise level was, of course, not measured and the levels shown are based on theoretical considerations [see references]. Also shown is the median expected business area man-made noise. Further details concerning man-made noise and its variation can be obtained from Report 258 and [Spaulding and Disney 1974] and references therein.

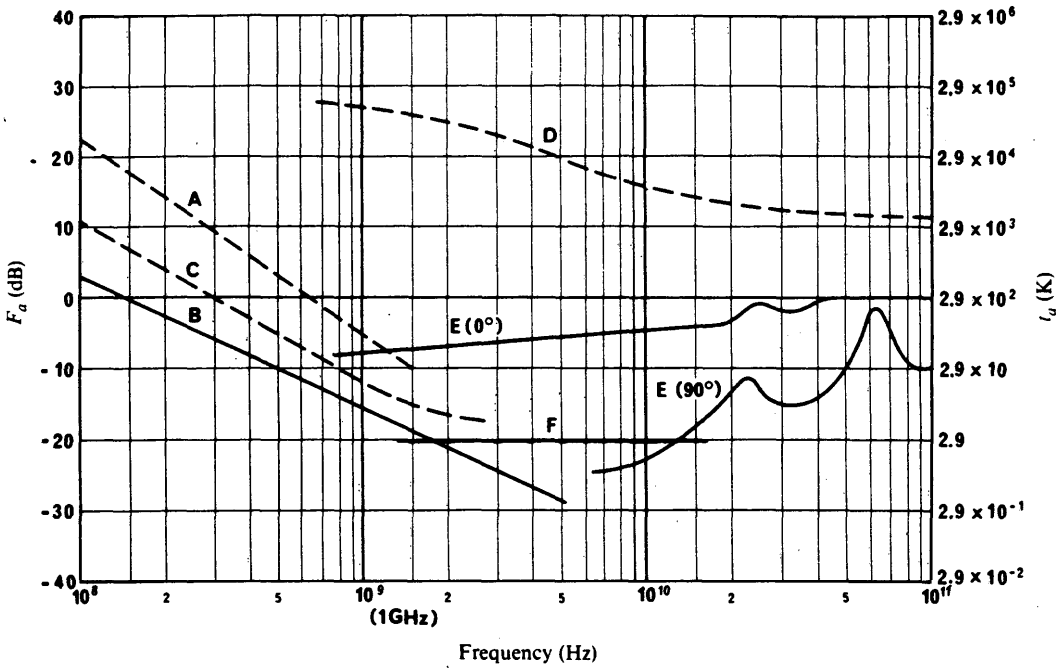


FIGURE 3 — *F_a versus frequency (10^8 to 10^{11} Hz)*

A: Estimated median business area man-made noise

B: Galactic noise

C: Galactic noise (toward galactic center with infinitely narrow beamwidth)

D: Quiet sun (½ degree beamwidth directed at sun)

E: Sky noise due to oxygen and water vapor (very narrow beam antenna); upper curve, 0° elevation angle; lower curve, 90° elevation angle

F: Black body (cosmic background), 2.7 K

———: Minimum noise level expected

Fig. 3 : On this figure, the frequency range 10^8 to 10^{11} Hz is covered, i.e., 100 MHz to 100 GHz. Again, the minimum noise is given by solid curves, while some other noises of interest are given by dashed curves.

The majority of the results shown on the three figures are for omni-directional antennas (except as noted on the figures). The average value of F_a for directional antennas will be the same if random direction is assumed. Studies have indicated that at HF (for example), for atmospheric noise from lightning for very narrow beam antennas, there can be as much as 10 dB variation (5 dB above to 5 dB below the average F_a value shown) depending on antenna pointing direction.

For galactic noise, the average value (over the entire sky) is given by the solid curve labled galactic noise (Figs. 2 and 3). Measurements indicate a ± 2 dB variation about this curve. The minimum galactic noise (narrow beam antenna towards galactic pole) is 3 dB below the solid galactic noise curve shown on Fig. 3. The maximum galactic noise for narrow beam antennas is shown via a dashed curve on Fig. 3.

We now want to consider a simple example to show how to determine the required noise figure. At 10 kHz, for example, the minimum external noise is F_a = 146 dB. If it is assumed $t_c = t_t = t_0$, and $l_c = l_t = 1$ (that is, no antenna or transmission line losses), then

$$f = f_a - 1 + f_r \tag{10}$$

We can take f_r to be that value which will increase f by only 1 dB. This gives a noise figure, f_r, of 140 dB, or an overall noise factor, f, of 147 dB. Any smaller noise figure, f_r, no matter how small, cannot decrease f below 146 dB. Consider now that $l_c = l_t = 100$, i.e., 20 dB antenna losses and 20 dB transmission losses. Then,

$$f = f_a - 1 + 10\,000\, f_r \tag{11}$$

In order to raise the f no more than 1 dB (to 147 dB) for the above situation, f_r can only be as large as 100 dB.

There are occasions where more than one type of noise needs to be considered because two or more are of comparable size. This can be true at any frequency, in general, but occurs most often at HF, where atmospheric, man-made, and Galactic noise can be of comparable size (Figure 2, 10 MHz, for example). The values of noise power, F_a, given on Figures 1-3 are median values (3 month-4 hour time blocks for atmospheric noise) and the f_a's have distributions about the median f_a. An appropriate method for obtaining the median value and distribution for the sum of two or more noise processes has been developed [Spaulding and Stewart, 1987].

Finally, it should be noted that many of the external noises are impulsive in nature (e.g. many forms of atmospheric and man-made noise). This means that the external noise can still limit performance of a communications system, even though the receiver noise is made as high as possible without increasing the overall operating noise factor f. That is, performance depends not only on the interfering noise power (signal-to-noise ratio), but also on the detailed statistical characteristics of the interference. Additional detailed information is contained in CCIR Reports 258, 322, 342, and 720.

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DECISIONS

DECISION 27-2

SPECTRUM MANAGEMENT AND COMPUTER-AIDED TECHNIQUES

(Questions 18/1, 44/1 and 45/1; Study Programmes 18B/1 and 18C/1)

(1978-1981-1985)

CCIR Study Group 1,

CONSIDERING

(a) that, due to the growing demands on the radio-frequency spectrum, there is a need to improve spectrum management techniques;

(b) that administrations are facing increasingly voluminous and complex tasks in spectrum management due to the increased use of existing telecommunications and new telecommunications technology;

(c) that the increase in the shared use of the spectrum among administrations requires the use of more complex analysis techniques;

(d) that the efficient solution of spectrum management problems requires data storage, data retrieval, and analysis capabilities, and consequently requires effective spectrum management and the application of computer methods;

(e) that, because of continuing technological developments, increasingly powerful micro- and minicomputers are available at reasonable cost;

(f) that frequency coordination between administrations and ntofication to the IFRB would be enhanced by the transfer of computer generated data;

(g) that Resolution 71 approved the publishing and maintenance of the Handbook on Spectrum Management and Computer-aided Techniques,

DECIDES

1. to maintain Interim Working Party 1/2, having the following terms of reference, and working mainly by correspondence according to Resolution 24:

to study the necessary principles and to develop techniques for efficient spectrum management, including specification of functional requirements, electromagnetic compatibility (EMC) models and computer-aided techniques for data management and frequency assignment. In particular the IWP shall be concerned with:

- maintaining and updating of the CCIR Handbook on Spectrum Management and Computer-aided Techniques, in particular to describe procedures on how to design and implement automated spectrum management systems;

- examination of computer programs for spectrum management and EMC analysis;
- development of recommended standards for computer generated national spectrum management data to be exchanged between administrations bearing in mind the advantages of compatibility with standards used for exchange of data between administrations and the IFRB;

2. that Interim Working Party 1/2 is presently composed as shown in Annex I to this Decision;

3. that any administration or organization desiring to participate in the work of Interim Working Party 1/2 should make known to the Director, CCIR, the name and address of its expert member;

4. that the Director, CCIR, shall communicate any changes to Annex I which may come to his attention, to the Chairman and Vice-Chairman of Study Group 1, as well as to the Chairman of IWP 1/2, without such changes being considered as modifications to the present Decision;

5. that IWP 1/2 shall prepare a report giving the results of its studies at the Final Study Group Meeting, 1989;

6. to review the present Decision at its next meeting.

ANNEX I

At the end of the XVIth Plenary Assembly, 1986, the following Administrations and International Organizations have indicated their participation in the work of Interim Working Party 1/2:

Administrations:

Germany (Federal Republic of)
Australia
Canada
China (People's Republic of)
Spain
United States of America
Finland
France
India
Italy
Japan
Jordan
Mexico
Norway
New Zealand
Papua New Guinea
Poland (People's Republic of)
United Kingdom
Sweden
Switzerland
USSR
Venezuela

International Organizations:

European Space Agency (ESA)
European Broadcasting Union (EBU)

International Telecommunication Union:

IFRB

Chairman of Interim Working Party 1/2:

Mr. R. J. Mayher
National Telecommunications and Information Administration
Department of Commerce
14th and Constitution Avenue N.W.
Room 4096
WASHINGTON DC 20230
United States of America

DECISION 53-2

UPDATING OF THE HANDBOOK FOR MONITORING STATIONS

(Question 32/1, Resolution 62)

(1981-1985-1989)

CCIR Study Group 1,

CONSIDERING

(a) the importance of monitoring to the general operation of expanding radio services;

(b) the difficulties encountered by administrations in collecting the information required for the establishment or modernization of monitoring stations;

(c) the importance of making available to the developing countries information which would enable them to set up their own stations and thus to take part in the international monitoring system;

(d) the usefulness of the Handbook for Monitoring Stations as a source of information on monitoring;

(e) the terms of Question 32/1 and § 6 of DECIDES in particular;

(f) the provisions of Resolution No. 39 (Mob-83) of the Radio Regulations;

(g) the provisions of Recommendation No. 30 of the Radio Regulations;

(h) the need to update the Handbook for Monitoring Stations in the light of the progress achieved in the techniques and equipment used by monitoring stations;

(j) the provisions of CCIR Resolution 62 relating to the need for visual and/or aural information,

(k) Recommendation No. COM6/B of the WARC-ORB(88) relating to the International monitoring of emissions originating from space stations,

DECIDES

1. that Interim Working Party 1/5 shall be maintained, pursuant to Resolution 24, with the following terms of reference:

- to update the Handbook as appropriate;
- to decide what new chapters might be included and to prepare them;
- to continue the studies concerning the space monitoring facilities;
- to study the possibilities that are available for monitoring the sprectrum by automatic means;
- to study various aspects of the inclusion of information contained in the relevant CCIR text in various Chapters of the Handbook as appropriate and report to Study Group 1 alternative proposals in this regard;
- to study the possibilities of using audio-visual materials to facilitate the training of staff at monitoring stations;
- to study specific monitoring questions submitted to it by the Director, CCIR, or the Chairman, Study Group 1;

2. that IWP 1/5 shall work according to Resolution 24, and by correspondence as far as possible;

3. that IWP 1/5 shall report on the progress made at the next Interim Meeting of Study Group 1;

4 that IWP 1/5 shall endeavour to complete its task before the next Final Meeting of Study Group 1;

5 that the Chairman and the composition of Interim Working Party 1/5 shall be as shown in Annex I.

6. that any administration and other organizations participating in the work of the CCIR, wishing to take part in the work of IWP 1/5 should communicate to the Director, CCIR, the name and address of its expert member;

7. that the Director, CCIR, shall notify the Chairman and Vice-Chairman of Study Group 1 as well as the Chairman of IWP 1/5 of any changes to Annex I that may be brought to his attention, without such changes being regarded as modifying this Decision.

ANNEX I

The following Administrations, International Organizations and Scientific or Industrial Organizations have indicated that they wished to participate in the work of Interim Working Party 1/5:

Administrations:

Germany (Federal Republic of)
Spain
United States of America
France
India
Iran (Islamic Republic of)
Italy
Japan
Portugal
United Kingdom
Canada

International Organizations and Scientific or Industrial Organizations:

EBU Rohde and Schwarz THOMSON-CSF

International Telecommunication Union:

IFRB

Chairman of Interim Working Party 1/5:

Mr. R. Lefort
CENTRE FRANCE TELECOM
BP9 78660 ABLIS
France

Telephone: 33 1 34842001
Telex: 698193 INTRAMB
Fax: + 33 1 34859208

DECISION 54-2

RADIATION FROM ISM EQUIPMENT

(Question 4/1 and Study Programme 4D/1)

(1981-1985-1989)

CCIR Study Group 1,

CONSIDERING

(a) Resolution No. 63 of the World Administrative Radio Conference, Geneva 1979;

(b) Opinion 2 of the CCIR;

(c) Recommendation 433-3 of the CCIR;

(d) that the Interim Working Party, known as IWP 1/4, has been constituted at the end of 1980;

(e) that although progress has been made in the work of IWP 1/4 final conclusions have not yet been reached (see Report 1104 which indicates the further work that is required),

DECIDES

1. to continue the work of IWP 1/4 mainly by correspondence according to Resolution 24 with the following terms of reference:

- to determine, in collaboration with the CISPR and IEC, limits for the radiation of ISM equipment in those parts of the spectrum which are not designated for the use of ISM equipment;
- to determine, in collaboration with the CISPR and IEC, limits for the radiation of ISM equipment in the frequency bands which are designated for their use, taking into account the needs of present and future ISM applications;

2. that the Chairman and the composition of IWP 1/4 shall be as shown in Annex I;

3. that the Director, CCIR, shall communicate this Decision to all administrations, and organizations participating in the work of the CCIR, with a view to requesting their contributions to the work of IWP 1/4;

4. that any administration or organization other than those listed in Annex I, desiring to participate in the work of IWP 1/4 should make known to the Director, CCIR, the name and address of its expert member;

5. that the Director, CCIR, shall communicate any changes to Annex I which may come to his attention, to the Chairman and Vice-Chairmen of Study Group 1, as well as to the Chairman of IWP 1/4, without such changes being considered as modifications to the present Decision;

6. that IWP 1/4 shall report to the next Interim Meeting of Study Group 1, and if possible, prepare an appropriate Recommendation for approval at the next CCIR Plenary Assembly.

ANNEX I

The following Administrations and International Organizations have indicated that they wished to participate in the work of Interim Working Party 1/4:

Administrations:

Germany (Federal Republic of)
Canada
China (People's Republic of)
Denmark
United States of America
France
India
Italy
Japan
Netherlands
Poland (People's Republic of)
United Kingdom
Sweden
USSR

International Organizations:

CISPR
EBU
IALA

Chairman of Interim Working Party 1/4:

Mr. G. Lehning
Fernmeldetechnisches Zentralamt
Referat S 24
Am Kavalleriesand 3
D-6100 DARMSTADT
Federal Republic of Germany

Telephone: +49 6151 832450
Telex: 419511 ftz d
Fax: + 49 6151 834791

DECISION 56*

TRANSFER OF TECHNOLOGY

(1980)

CCIR Study Group 1,

CONSIDERING

(a) Resolution No. 14 of the World Administrative Radio Conference, Geneva, 1979, relating to the transfer of technology;

(b) the importance of the application of science and technology for the purpose of telecommunications development throughout the world;

(c) the need to accelerate this application of science and technology,

DECIDES

1. it will undertake to assist, in whatever way is possible, the transfer of technology needed for telecommunications development throughout the world, and especially in developing countries;

2. it will encourage ITU member administrations to participate to the maximum extent practicable in the work of the CCIR, an important forum for the transfer of information related to application of telecommunication technology;

3. it will additionally undertake to publish in the form of manuals and handbooks, information of a permanent and directly applicable nature that will help provide for the needed technology transfer.

Examples:

(a) Reports of IWP 1/2 on Spectrum Management and Computer-aided Techniques (to assist modern frequency management);

(b) "Handbook for Monitoring Stations", CCIR (for which a special chapter is being prepared to address means of solving new problems encountered in establishing a new monitoring service);

4. it will provide technical aid to ITU-organized seminars and study cycles.

DECISION 100

PREPARATIONS FOR THE 1992 WORLD ADMINISTRATIVE RADIO CONFERENCE

CCIR Study Group 1, (1989)

CONSIDERING

(a) that the ITU Plenipotentiary Conference, Nice 1989, by Resolution No. PL-B/1, has decided:

- that the World Administrative Radio Conference for Dealing with Frequency Allocations in Certain Parts of the Spectrum, shall be held in the first quarter of 1992;
- that the agenda for this Conference shall be established by the Administrative Council, taking into account the Resolutions and Recommendations of WARC HFBC-87, WARC MOB-87 and WARC ORB-88 relating to frequency allocations;
- that, in addition, this Conference may consider defining certain new space services and consider allocations to these services in frequency bands above 20 GHz;

(b) that CCIR Study Group 1 has already developed some basic concepts and technical criteria for frequency sharing and compatibility assessment which may facilitate the work of the Conference;

(c) that a CCIR Study Group 1 report should be submitted to the Joint Interim Working Party preparing a consolidated report for WARC-92 which can be submitted to administrations in advance of the Conference, as indicated in Resolution No. 24;

(d) that Study Group 1 may be in a unique position with respect to providing guidance on basic concepts and technical criteria needed for frequency sharing within allocations and for compatibility between services,

DECIDES

1. to create an Interim Working Party 1/6 which shall prepare the Study Group 1 report to JIWP WARC-92 with the following terms of reference:

1.1 to identify appropriate assumptions and principles that might be applicable to both existing and anticipated frequency sharing situations;

1.2 to study basic concepts and technical criteria as appropriate for frequency sharing as called for in draft new Decision [JIWP WARC-92];

1.3 to report on any other specific sharing situations where Study Group 1 could provide guidance to the preparatory work for WARC-92;

1.4 to review existing Study Group 1 texts with a view to extracting those parts pertaining to WARC-92 and to forward those to the JIWP;

1.5 to study frequency tolerances and radio emissions which have impact sharing or compatibility criteria and which are applicable to those services and allocations included by the Administrative Council on the agenda for WARC-92;

1.6 to provide rapporteurs, as may be appropriate, to the IWPs and JIWPs established to prepare texts for JIWP WARC-92 with a view to coordinating further study that may help resolve frequency sharing issues pertaining to WARC-92,

2. that Mr Becker (New Zealand) shall serve as Chairman of the IWP and Mr. Skiffins (UK) and Mr. McCaughern (Canada) shall serve as Vice-Chairmen;

3. that IWP 1/6 shall establish its structure and working arrangements as required;

4. that interested administrations and other participants in the work of CCIR Study Group 1 should indicate not later than 10 January 1990 to the Director of the CCIR and to the Chairman of IWP 1/6 their intention to participate in the work of the IWP and should make known the name(s) and address(es) of their representative(s);

5. that the IWP shall meet as appropriate so that a report will be available for distribution in order to meet the time scale required by JIWP WARC-92.

ANNEX

IWP 1/6 participants will consist of representatives of:

Australia	DDR	EBU
Canada	Iran	
USA	Finland	
France	UK	
Italy	Japan	
Sweden	Netherlands	

