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XVIIth PLENARY ASSEMBLY
DÜSSELDORF, 1990



INTERNATIONAL TELECOMMUNICATION UNION

REPORTS OF THE CCIR, 1990

(ALSO DECISIONS)

ANNEX TO VOLUME X – PART 1

BROADCASTING SERVICE (SOUND)

CCIR INTERNATIONAL RADIO CONSULTATIVE COMMITTEE



Geneva, 1990



XVIIth PLENARY ASSEMBLY
DÜSSELDORF, 1990



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ANNEX TO VOLUME X-1

BROADCASTING SERVICE (SOUND)

(Study Group 10)

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SECTION 10A-1 : AMPLITUDE-MODULATION SOUND BROADCASTING IN BANDS 5 (LF), 6 (MF)
AND 7 (HF)

REPORT 1058

**MINIMUM AF AND RF SIGNAL-TO-NOISE RATIO REQUIRED FOR
BROADCASTING IN BAND 7 (HF)***

(Question 44/10, Study Programme 44B/10)

(1986)

1. Introduction

The RF signal-to-noise ratio at the receiver input depends, apart from other factors, on audio frequency S/N ratio for a defined grade of performance. Since the majority of the physiological and psychological factors ultimately influence only the AF signal-to-noise ratio, a series of subjective listening tests were carried out in India and the USSR to assess the minimum acceptable value of this ratio, from which the equivalent RF signal-to-noise ratio at the input of the receiver may also be derived.

* The First Session of the World Administrative Radio Conference for the Planning of HF Bands Allocated to the Broadcasting Service (WARC HFBC(1)) has already considered the content of the documents mentioned in the reference and adopted a value of 24 dB for AF signal/noise ratio for planning purposes.

2. Tests carried out in India [CCIR, 1982-86a]

2.1 Experimental procedure

2.1.1 A variety of pre-recorded programme samples (spoken word, instrumental music, vocal classical music and western music) played back from a tape deck were mixed with white noise obtained from a random noise generator. The bandwidth setting of the noise generator was kept at 20 kHz. These programme samples mixed with noise were recorded on a magnetic tape through a filter with a cut-off frequency of 3 kHz (3 dB attenuation) and an attenuation slope of 24 dB per octave. These filter characteristics represent the characteristics of an average HF receiver in India. Each sample was recorded for different combinations of signal-to-noise ratio ranging from 15 dB to 30 dB in steps of 2 dB. The average level of each programme sample was pre-determined by means of a level recorder. This level was maintained in the play-back system by using a standard vu-meter calibrated by a steady level tone. The r.m.s. AF noise was measured by a sound level meter conforming to the specification laid down in IEC 179-A publications. The signal-to-noise ratio measured with this set-up could be considered as representing the ratio of the audio frequency signal as measured on a standard vu-meter, to the r.m.s. noise, for a bandwidth of 3 kHz.

2.1.2 Each recorded sample was reproduced before an audience through a good quality reproducing system. The listeners were asked to assess whether a particular sample was acceptable to them, keeping in mind the inherent quality characteristics of HF broadcasting services.

2.2 Analysis and discussion

2.2.1 Minimum acceptable signal-to-noise ratios are shown in Table I and Fig. 1 for different types of programme. Audio frequency signal-to-noise ratio values of 16 dB, 17 dB and 19 dB were accepted by 50, 70 and 90% of the listeners respectively, for instrumental and western music programmes. For the spoken word and vocal classical music programmes, values of 17 dB, 19 dB and 21 dB were found acceptable by 50, 70 and 90% of the listeners, respectively. These values of AF signal-to-noise ratio apply to 3 kHz audio bandwidth under stable conditions.

TABLE I — Minimum audio frequency signal-to-noise ratio (dB)
accepted by various percentages of listeners

	50%	70%	90%
Instrumental (music Sarod)	16	17	19
Western music (pop song)	16	17	19
Spoken word	17	19	21
Vocal music (classical)	17	19	21

2.2.2 Thus an AF signal-to-noise ratio of 21 dB was determined as the minimum requirement under the worst conditions.

2.2.3 Further tests were carried out to determine the equivalent S/N ratio at the input of a receiver whose characteristics were representative of the average characteristics of HF receivers available in India [CCIR, 1978-82]. A steady RF signal modulated to 30% with 1 kHz tone was fed to the receiver. For an S/N ratio of 15 dB to 22 dB at the output, a corresponding S/N ratio of 24 dB to 31 dB was observed at the input. The differences between the output and the input S/N ratio was thus found to be 9 dB. It was therefore concluded that a value of 30 dB RF S/N ratio may be regarded as an acceptable value for broadcasting in band 7 (HF).

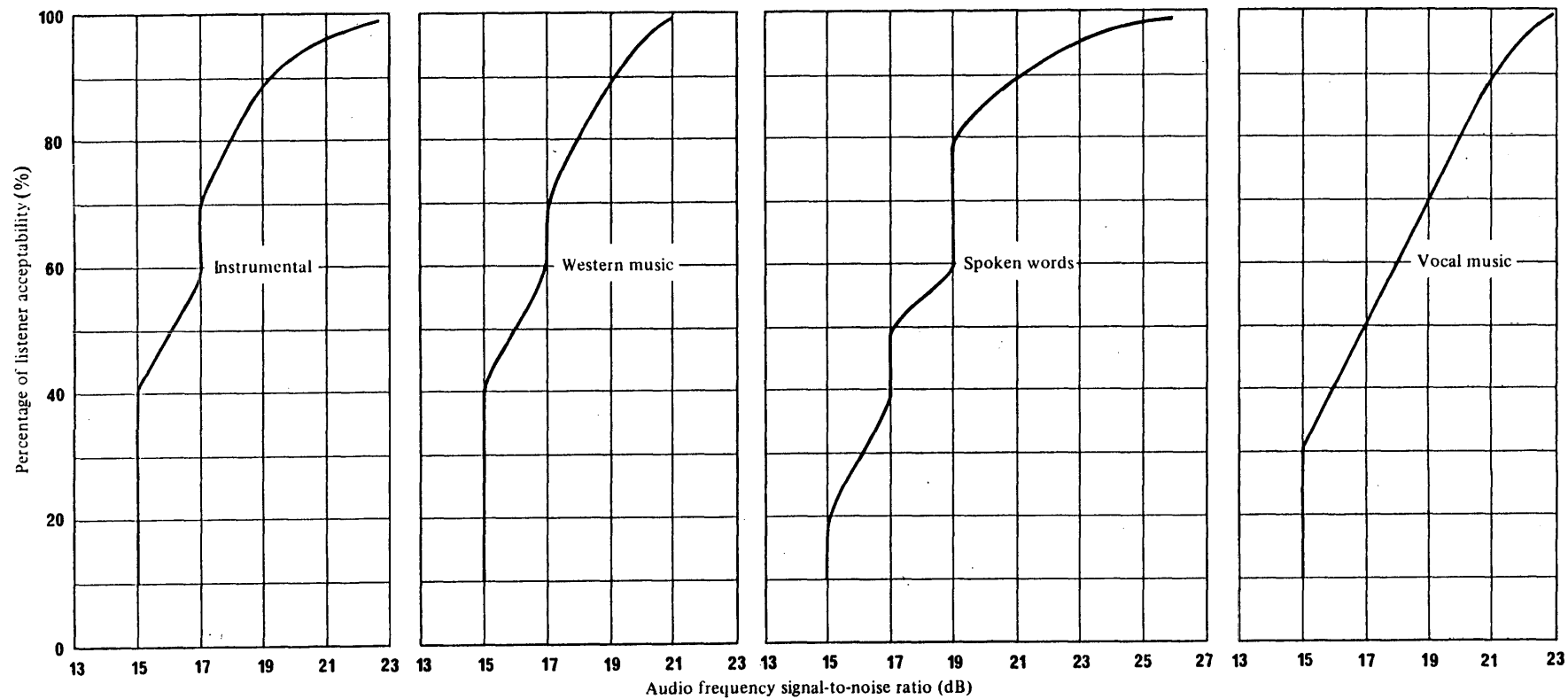


FIGURE 1 – *Signal-to-noise ratios acceptable for different types of programme*

3. Tests carried out in the USSR [CCIR, 1982-86b]

Several series of controlled listening tests in accordance with the method described in Recommendation 562 for programmes of various types and for various AF signal-to-noise ratios have been carried out in the USSR, the results of which are presented below.

3.1 Description of the experiments

Figure 2 contains a block diagram of the experimental arrangement used for the listening tests. A variety of programme samples (speech, music) were pre-recorded on magnetic tape and played back from a tape deck. The resulting signal was fed into the modulator of an HF generator. The same modulator also received white noise from the output of a noise generator. Before input into the modulator, the wanted audio modulating signal and the noise were mixed in a passive adding circuit. The HF signal at the HF generator output was fed via an artificial antenna to a standard broadcasting receiver with an RF bandwidth of 6 kHz and an AF bandwidth of 3 kHz. The HF signal at the receiver input was maintained at a sufficiently high level to enable the intrinsic receiver noise to be disregarded.

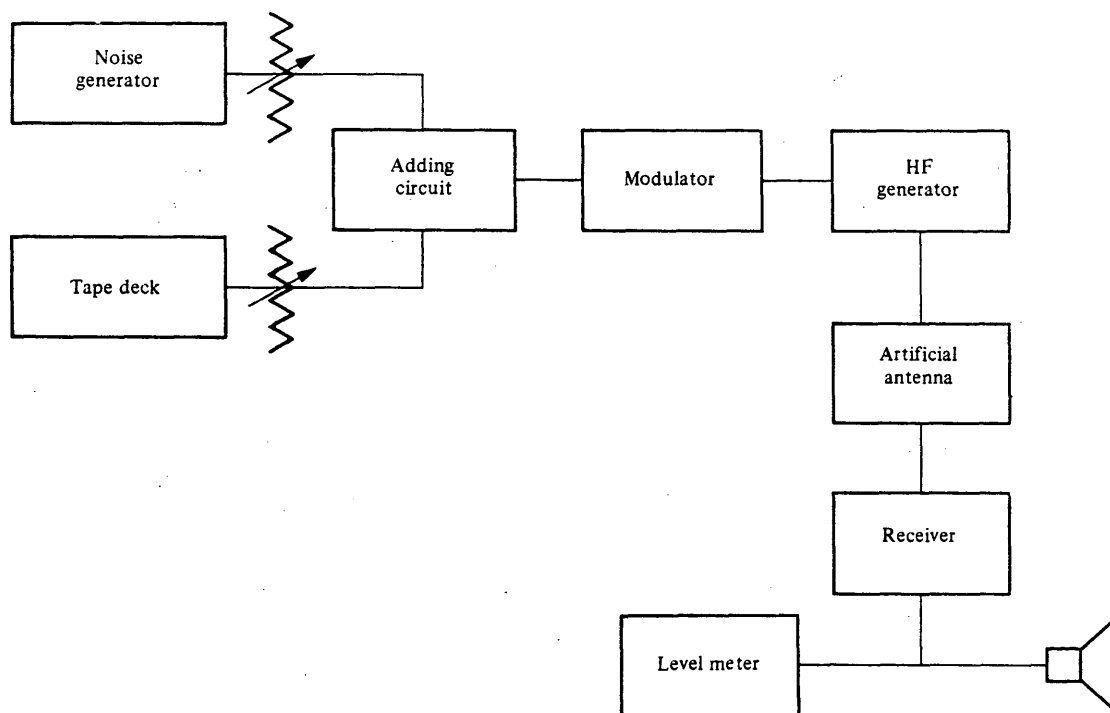


FIGURE 2 – Block diagram of the experimental arrangement used for the listening tests

The audio signal level at the receiver output was measured using a standard level meter with time integration of the order of 200 ms used to measure the average sound programme level. The r.m.s. noise level was also measured at this point. These two readings were then used to determine the signal-to-noise ratio.

The audio signals at the receiver output were reproduced before an audience through a good quality speaker.

The listeners judged the degree of noise impairment on the CCIR five-grade scale. Families of curves were then plotted for each specific value of the signal-to-noise ratio (in 3 dB increments) for the statistical evaluation of noise impairment according to the CCIR scale as a function of the number of listeners.

3.2 Findings and conclusions

Figure 3 portrays the relation between the grade of noise impairment judged on the CCIR scale and the AF signal-to-noise ratio. The graph is based on the selected opinions of 80% of listeners, for speech programmes. It is known that noise is more perceptible for speech programmes, and less obtrusive for music, particularly dance music.

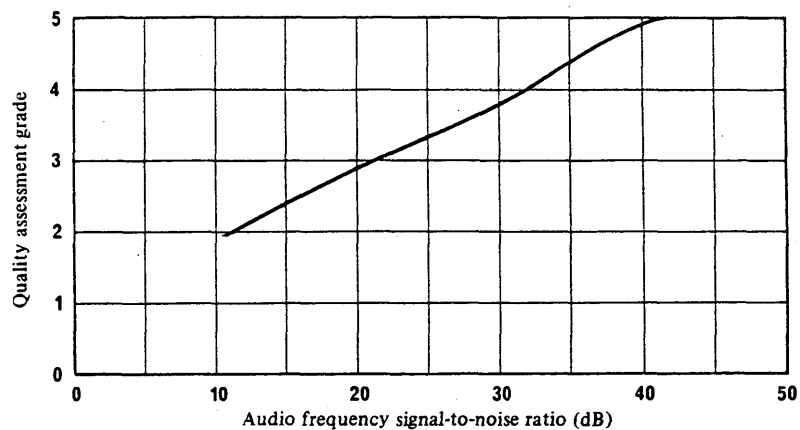


FIGURE 3 – Grade of noise impairment on the CCIR five-grade scale as a function of the AF signal-to-noise ratio

The graph in Fig. 3 shows that perceptible, but not annoying, noise, corresponding to grade 4 on the CCIR scale, occurs with an AF signal-to-noise ratio of approximately 31 dB. For a ratio of approximately 20-21 dB, 80% of the listeners evaluated the noise as slightly annoying, i.e. corresponding to grade 3 on the CCIR scale.

REFERENCES

CCIR Documents

[1978-82]: 10/204 (India).

[1982-86]: a. 10/67 (India); b. 10/227 (USSR).

REPORT 1201

NUMBER OF HF SOUND BROADCASTING TRANSMITTERS USING A SINGLE CHANNEL

(Study Programme 44F/10)

(1990)

Investigations were carried out in the USSR [CCIR 1986-90] with a view to determine the maximum number of transmitters using a single channel in HF sound broadcasting networks. These investigations were carried out using network models taking into account the specific design features of international and national sound broadcasting networks.

On the basis of the results obtained, it was possible to evaluate the capacity of the HF bands allocated to the sound broadcasting service as a function of the requirements governing the operation of the networks.

The maximum number of transmitters in a channel (K_f) depends on a multiplicity of geophysical factors and technical network parameters. Particular attention was paid to the influence of the following factors on K_f :

- type of network model;
- frequency band used;
- size of service area;
- protection ratios adopted;

Three types of model were used:

- a model corresponding to national and local broadcasting networks;
- a model representing the characteristic features of the organization of international and national broadcasting over large areas;
- a mixed model taking into account the characteristics of the first two models.

On the basis of a study of these models, the following conclusions were obtained:

- the maximum possible number of transmitters using a channel at the same time varies from 6-7 in the low frequency bands to 2-3 in the high frequency bands;

- the maximum number of transmitters is practically independent of the type of the model considered; a dependence on the type of the model is apparent in the low and high frequency bands;
- with large service areas (some CIRAF zones) which in practice are frequently notified by broadcasters, the maximum number of stations depends only slightly on the protection ratios; for smaller areas there is a more marked dependence.

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[1986-90]: 10/225 (USSR)

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veshchatelnykh peredatchikov (Determination by calculation of the number of
HF sound broadcasting transmitters operating on one frequency).
Elektrosviaz, 10, Moscow, USSR.

REPORT 516-4

FIELD STRENGTH RESULTING FROM SEVERAL ELECTROMAGNETIC FIELDS

(Question 44/10, Study Programme 44A/10)

(1970-1978-1982-1986-1990)

Studies have been carried out in Italy [CCIR, 1966-69] and in Hungary [CCIR, 1978-82] on the composition of several stable electromagnetic fields at the same point.

1. Field strength resulting from two stable electromagnetic fields

If two fields of different frequencies are considered at a point in space:

$$E_1 = A \cos \omega t \quad (1)$$

$$E_2 = B \cos [(\omega + \Delta\omega)t + \varphi] \quad (2)$$

where A , B , ω , $\Delta\omega$ and φ are constant in time, and if it is assumed that both fields are polarized in the same direction, the instantaneous amplitude of the vector representing the resultant field is:

$$E = \sqrt{A^2 + B^2 + 2AB \cos (\Delta\omega t + \varphi)} \quad (3)$$

The mean value of E in the period $T = 2\pi/\Delta\omega$ is:

$$E_R = \frac{1}{T} \int_{t_0}^{t_0 + T} E(t) dt = Af(A/B) \quad (4)$$

The $\Delta\omega$ frequency component of E for the same period is:

$$E_{\Delta\omega} = \left(\frac{2}{T}\right) \int_{t_0}^{t_0 + T} E(t) \cos (\Delta\omega t + \varphi) dt \quad (5)$$

If the values A , B , E_R and $E_{\Delta\omega}$ (dB(μ V/m)) are designated by E_1 , E_2 , $(E_1 + \Delta R)$ and $(E_2 + \Delta E)$ respectively and assuming that $E_1 \geq E_2$, a graph can be drawn of the values of ΔR , as indicated by a field measuring device, and of ΔE as a function of $(E_1 - E_2)$, and the curves shown in Fig. 1 (continuous line) and Fig. 2 are obtained.

The dotted line shown in Fig. 1 is obtained by calculating the r.m.s. of the amplitudes of the two fields:

$$E_R = \sqrt{A^2 + B^2} \quad (6)$$

The values of the first curve (continuous line) in Fig. 1 are always lower than those of the second curve, for each value of $(E_1 - E_2)$. The maximum difference is 0.9 dB for $E_1 = E_2$.

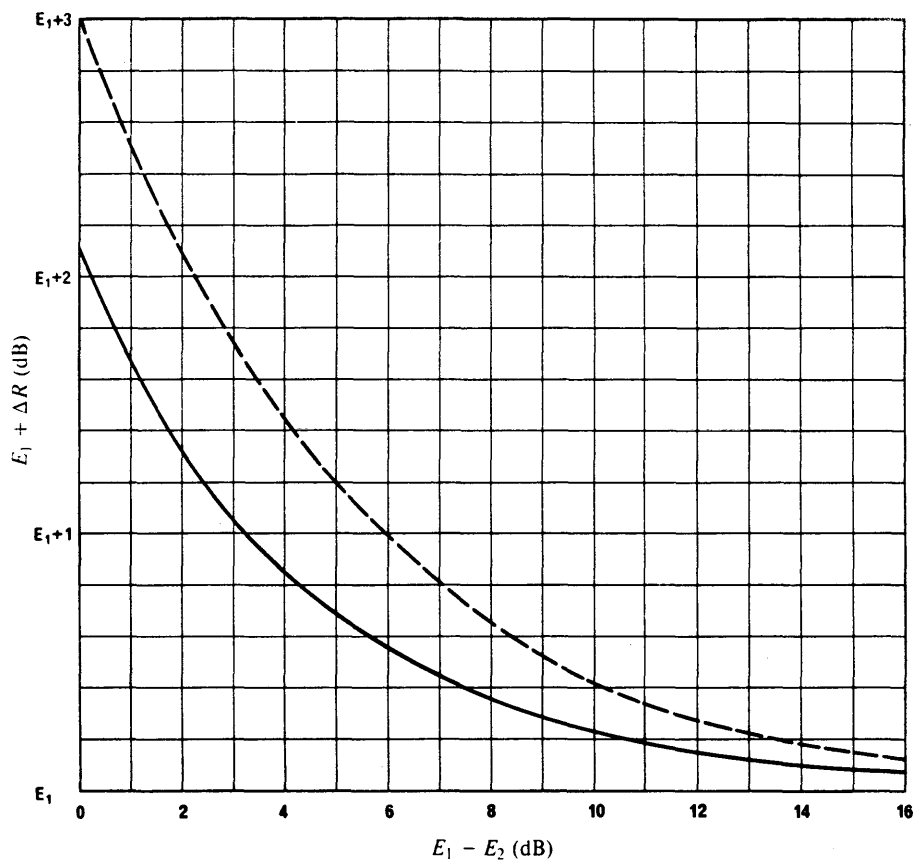


FIGURE 1 - Resultant ($E_1 + \Delta R$) of two stable electromagnetic fields ($E_1 > E_2$)

— : value indicated by a field measuring device
- - : value calculated by adding the powers of the two signals

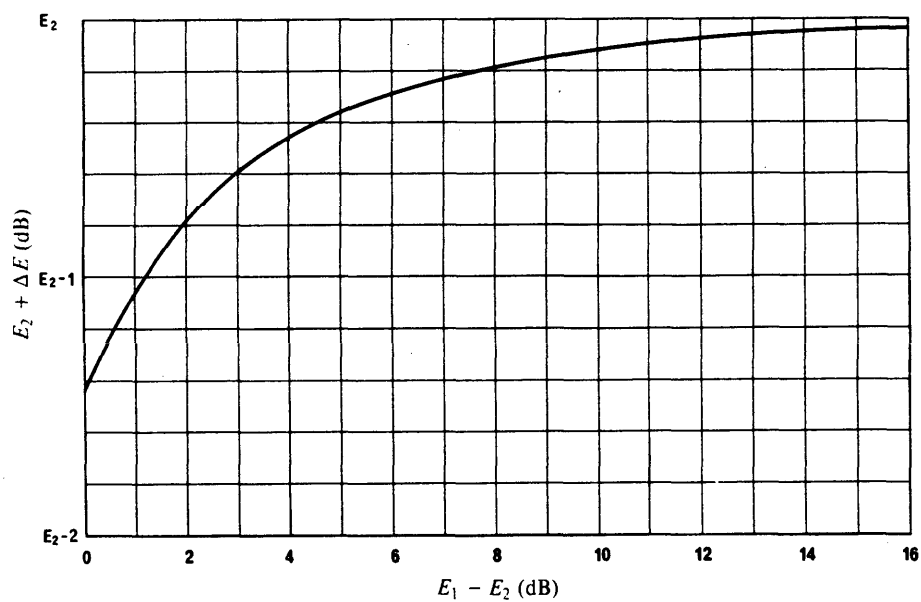


FIGURE 2 - ($E_2 + \Delta E$) component in the resultant of two stable electromagnetic fields ($E_1 > E_2$)

2. Field resulting from three stable fields

Let:

$$E_1 = A \cos \omega t$$

$$E_2 = B \cos [(\omega + \Delta\omega_1)t + \phi_1]$$

$$E_3 = C \cos [(\omega + \Delta\omega_2)t + \phi_2]$$

be three fields (with the same assumptions as before).

If the values A , B , C , the mean value and amplitude of the components of the vector representing the resultant (dB(μ V/m)) are designated by E_1 , E_2 , E_3 , $(E_1 + \Delta R)$, $(E_2 + \Delta E_1)$ and $(E_3 + \Delta E_2)$ respectively and assuming that $E_1 \geq E_2$ and $E_1 \geq E_3$, graphs can be drawn in which the curves give the values of ΔR and ΔE_1 , ΔE_2 as a function of $(E_1 - E_2)$ and $(E_1 - E_3)$.

The results are shown in Figs. 3 and 4.

It can be seen, for example, that when $E_1 = 63$ dB, $E_2 = 62$ dB and $E_3 = 60$ dB, then $\Delta R = 2.8$ dB, $\Delta E_1 = 3.7$ dB and $\Delta E_2 = 3$ dB.

For information concerning the measurement of ΔE , ΔE_1 and ΔE_2 , refer to Report 273.

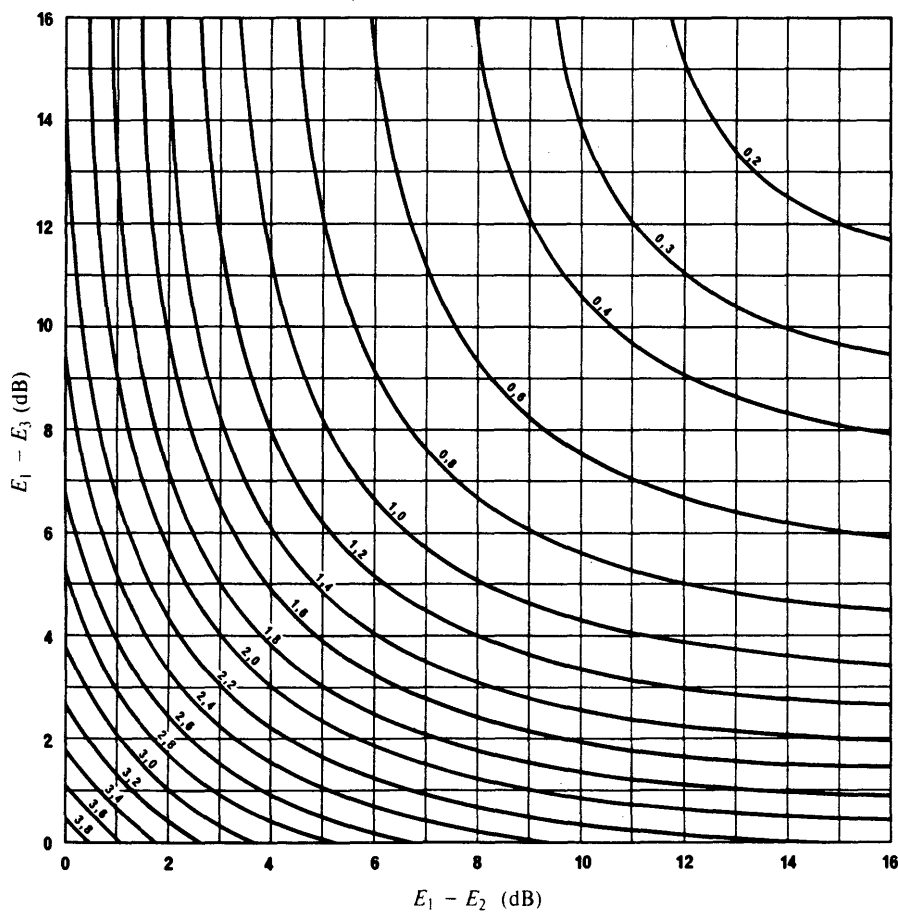


FIGURE 3 - Family of curves $(E_1 + \Delta R)$ resulting from three stable electromagnetic fields ($E_1 > E_2$, $E_1 > E_3$)

4. Experimental results

Measurements of field strength carried out at the RAI monitoring centre at Monza and at the Hungarian monitoring station at Tárnok have led to the following deductions:

- considering two or three amplitude-modulation emissions in band 6 (MF) on the same channel;
- assuming a difference of at least a few hertz between the carrier frequencies, and
- assuming that the signals at the reception point are stable,

it may be assumed that the measured value of the resultant field is, with an approximation of 0.2 dB, that which is deduced from the continuous lines in Figs. 1 to 4 respectively.

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CCIR Documents

[1966-69]: X/56 (Italy).

[1978-82]: 1/143, 10/198 (Hungary (People's Republic of)).

[1986-90]: 10/253 (USSR).

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REPORT 401-6

TRANSMITTING ANTENNAS IN LF AND MF BROADCASTING

(Question 44/10, Study Programme 44G/10)

(1966-1970-1974-1978-1982-1986-1990)

1. Antenna with reduced vertical radiation

A high-efficiency anti-fading antenna should be of sectionalized construction and have a total electrical height of $2\lambda/3$ to λ , to produce the necessary rapid rise of sky-wave field-strength near the point where it equals that of the ground-wave. The effect of the resistive component of the antenna current on the vertical radiation pattern of a sectionalized tower can be reduced or compensated by multiple feeding. It should be noted that the location and extent of the fading zone varies due to changes in the properties of the reflecting ionospheric layers.

In practice, the fading zone is somewhat larger than that calculated. This might be due, on the one hand, to variations of the E-layer reflection, and on the other hand, to F-layer reflections. The design of antennas should take care of these effects.

2. Influence of ground conductivity on the vertical radiation pattern (including ground system)

[CCIR, 1970-74a] gives the results of a theoretical study into the influence of ground conductivity on the vertical radiation patterns of typical vertical antennas. The study takes account of the diffraction of waves around the curvature of the Earth.

Ground loss may be defined as the reduction in field strength which occurs above a curved Earth when land replaces sea water. The corresponding increase which occurs when land is replaced by sea water is known as "sea gain".

[CCIR, 1966-69a] gives the results of field-strength measurements carried out to determine the effects of ground conductivity on low angle radiation over a relatively long transmission path of 1400 km. Transmissions from Rome at 845 kHz were measured simultaneously at coastal and inland sites along a radial extending 100 km inland from a coastal site in Southern England. Because of the principle of reciprocity it is immaterial whether the antenna transmits or receives. Figure 1 shows the results of measurements in terms of field strength at the inland sites. B to K. relative to the coast site A.

It is concluded that a transmitter operating in the MF band (band 6) will radiate sky-waves more efficiently at low angles of radiation if it is situated on a coastline facing the area to be served and that the ground loss does not reach its limiting value until the antenna is at least 50 km inland. To obtain maximum advantage open sea must extend from the coastline for a distance of at least 100 wavelengths in the direction of propagation.

Ground loss applies equally to transmitting and receiving antennas. For further information see Recommendation 435.

Measurements of the effect of sea gain have been reported by Italy [CCIR 1966-69b] and by the USSR [CCIR 1970-74b] in LF and MF. The results are in good agreement with the calculated theoretical values.

Studies carried out in Italy [CCIR, 1982-86a] have shown the combined influence of a soil of finite conductivity and of a ground system on the sky-wave vertical radiation pattern of MF transmitting antennas. By the use of a computer-based procedure, a correction factor has been derived which is applied to the radiation pattern of a single mast antenna over flat and perfectly conducting ground.

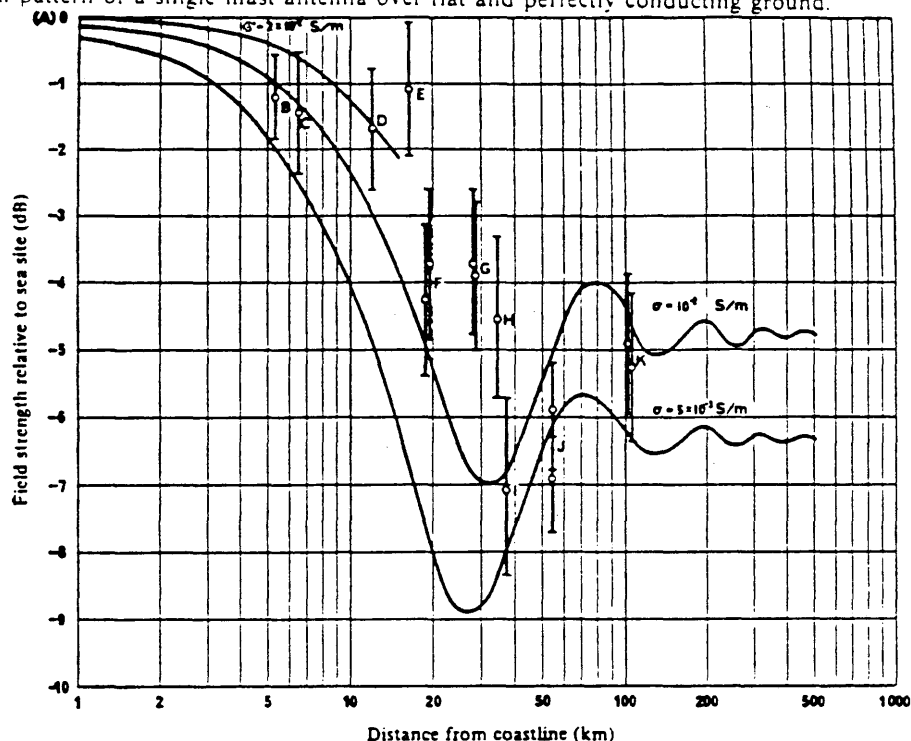


FIGURE 1 - Theoretical and measured ground loss

— : theoretical ground loss

○ : measured ground loss

(Vertical lines indicate confidence limits)

Figure 1 shows how ground loss varies with distance from the sea at the frequency 845 kHz for an angle of arrival of about 4° . Also shown in Fig. 1 are theoretical curves for ground conductivities of 5×10^{-3} S/m and 10^{-2} S/m which are believed to be the limits for most of the area. Part of the theoretical curve for 2×10^{-2} S/m is also included since the first 10 km inland was known to be of about this value. Figure 1 clearly demonstrates the large ground loss at sites well inland from the coast.

3. High-efficiency transmitting antennas

It should be emphasized that the capital and maintenance costs of a high-efficiency transmitting antenna should not be considered in isolation, but rather with regard to the cost and effectiveness of the broadcasting station as a whole. For example, for a broadcasting station with a transmitter output power of 100 kW or more, the cost of a simple antenna may be a relatively small part of the total expenditure.

3.1 LF anti-fading antennas

[CCIR 1966-69c] gives information on experience gained with the broadcasting ring antenna for use in band 5 at Motala (191 kHz). The antenna is a stationary field ring antenna consisting of one central element (a vertical radiator of height 250 m) and five vertical radiators with a height of 200 m, equally spaced on a radius of 630 m (0.4 λ). The measurements of field strengths at distances of up to 300 km from the antenna show a substantial extension of the fading-free zone.

3.2 MF anti-fading antennas

[CCIR, 1970-74c] describes a double half-wave cage-type antenna for use in band 6 (MF). This antenna consists of four cages of copper or aluminium wires. The cages are suspended on an earthed, non-divided mast, each cage ($\lambda/4$) being galvanically connected with the mast at its upper end and insulated from the mast at its lower end (see Fig. 2).

By appropriate selection of the ratio and phase of the currents in the two dipoles, the vertical radiation pattern can be shaped within a wide range. This enables selection of the optimum operating conditions for both day-time and night-time conditions.

[CCIR, 1970-74d] reports on the existence of a high-gain band-6 (MF) broadcasting antenna at Lille-Camphin on a frequency of 1376 kHz. This antenna is of the full-wave type and is designed to increase the horizontal gain by reducing radiation in directions other than that of the horizon. It is centre-fed, but its quarter-wave skirt structure avoids cutting the tower with insulators. There is also no insulator at the base of the tower. Feeder matching is effected without the classical antenna unit, inductance or capacitance. The radiation diagram was regulated with the help of measurements made using a helicopter. A detailed description of the antenna is available [Lacharnay, 1969].

[CCIR, 1970-74e] describes the design of an antenna, the radiation pattern of which can be regulated in the vertical plane. The upper part of the antenna is a normal vertical radiator whilst the lower part is a vertical radiator with a reactance in the form of a short circuit loop connected to the base. To reduce its characteristic impedance, inclined wires are suspended from the upper part of the antenna. A vertical wire screen is fixed to the lower part of the antenna, reaching from the ground to a height of $H_1 = 0.4 H$.

Regulation of the radiation pattern in the vertical plane is provided by adjusting the current distribution by means of distributed constant impedance lines. The antenna possesses anti-fading properties in a wide frequency band and has a gain of up to twice that of a base driven half-wave antenna.

3.3 MF Directional antennas

[CCIR, 1970-74f] describes a switchable transmitting antenna for optimized day and night service in band 6 (MF). Two antennas of this type were in use at two broadcasting stations (Ismaning, 1 602 kHz and Langenberg, 1 586 kHz). The antenna consists of three insulated sections and this permits very flexible adjustment of the current distribution and allows changes in the vertical radiation pattern over a wide range. A suitable combination of parameters results in an improvement in the co-channel radio-frequency wanted-to-interfering signal ratio of between 3 dB and 10 dB.

[CCIR, 1974-78a] describes a band 6 (MF) directional antenna with a controlled horizontal radiation pattern [Timonina, 1971].

The antenna consists of four radiators located at the corners of a square of side D . Two of the radiators are active whilst the other two act as passive reflectors tuned by means of a short-circuited stub. The radiation pattern in the horizontal plane is controlled over the sector 0° to 360° . The direction of maximum radiation may be changed in steps of 90° . Assuming $D = 0.25 \lambda$ to 0.28λ then the back radiation level is -18 dB to -20 dB relative to the main beam, whilst over an angular sector of 100° this level averages about -14 dB.

A four-mast antenna with an adjustable current distribution [Bielousov, 1974; Bielousov *et al.*, 1976] and height $H = \lambda$ with $D = 0.28 \lambda$ has a gain of 10 dB when tuned for maximum radiation in the forward direction.

[CCIR, 1974-78b] refers to a published article [Lacharnay, 1976] which describes various directional transmitting antennas employing vertical mast-radiators for LF and MF broadcasting, characterized by very low radiation over wide sectors, both in elevation and azimuth. Such antennas render it possible to reduce the interference in overcrowded frequency bands. Some practical installations are described and details are given of the attenuation that can be obtained (30 dB at present) as well as the stability of the antenna adjustments, verified over several years.

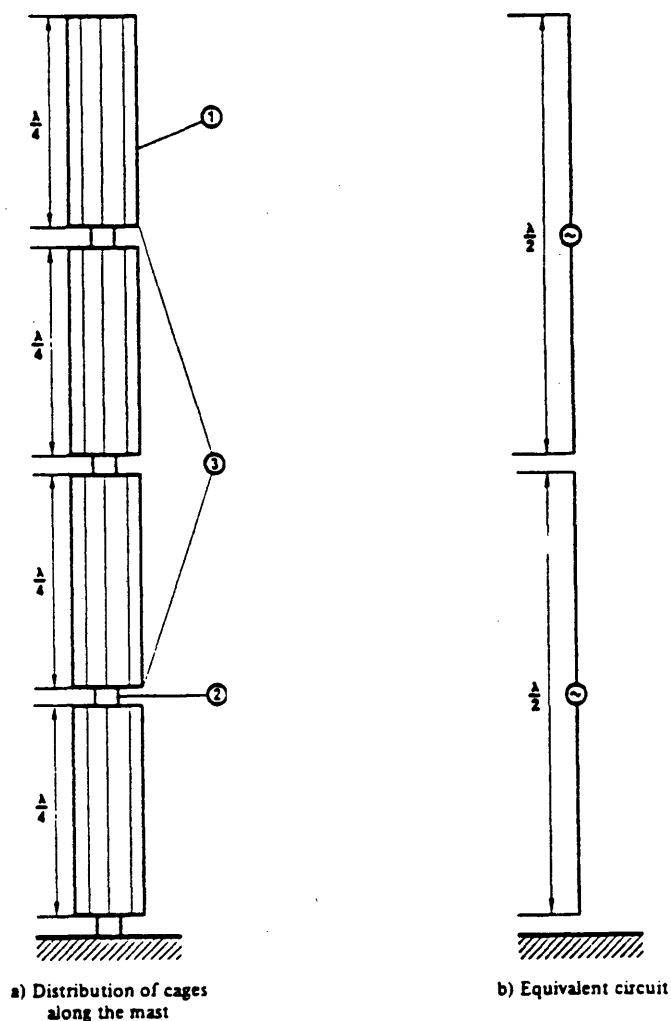


FIGURE 2 - Double half-wave cage-type antenna

4. Methods of prediction of the re-radiation effects

The problem of the re-radiation effects from structures, such as power transmission lines, metallic masts, buildings with metallic structure and buried metallic installations is a serious one since such structures may modify the desired radiation pattern. The elimination of the undesired effects is closely related to the knowledge of the current distribution along the antenna structure in the presence of such surrounding structures.

A number of the methods for determining the current distribution along the antenna structure has been reported. A simple method, the so-called "point matching method", a variant of "the moment method" [Harrington, 1968] has been developed in Yugoslavia [Popovic, 1970] and has been successfully applied in the treatment of the related applications [Surutka and Mitic, 1975].

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REPORT 1063

**PREDICTION AND CONTROL OF RE-RADIATION
IN MF BROADCASTING**

(Question 44/10, Study Programme 44G/10)

(1986)

1. Introduction**1.1 Re-radiation**

When the radiated field from a broadcast transmitting installation intercepts a metallic structure, currents are induced in that structure and it radiates a second field which adds vectorially to the original. This is known as re-radiation.

1.2 Re-radiating structures

Sound broadcasting at MF normally involves vertical radiators, usually insulated at the base, over an extensive conductive earth screen composed of many radial wires. Vertical metallic structures of appreciable height (about 15% or more of a wavelength) in the vicinity of such transmitting antennas may cause re-radiation problems. Typical structures include masts, towers, tall chimneys or heavy industry complexes. High-rise buildings may also be troublesome sources. However, the most common re-radiation hazard is high tension power transmission lines which use steel towers. It is normal on such lines to attach one or two "sky wires" to each tower, running parallel to and above the power conductors, so as to afford protection from lightning and to disperse any destructive currents present from either a lightning stroke or a fault involving the power conductors. The towers and sky wires, together with their images in the earth, form loops which tend to be one-wavelength resonant somewhere in the lower part of the MF band and two-wavelength resonant in the upper part. The power conductors themselves have little effect on the re-radiation.

1.3 Survey results

In Canada, about three-quarters of MF stations employ directional arrays.

A survey conducted in 1976 indicated that 28% of the respondents had suffered re-radiation problems and a further 25% were anticipating them. Steel-tower power lines were the chief offenders, with masts, tall buildings, smoke stacks and industrial complexes also mentioned. Costs of correction had varied from \$1000 to \$500 000 and in three instances the transmitter site had to be abandoned. One station anticipating problems for a new critical array required measures which involved the following: broadcast consultant — 5 days; station staff — 30 days; power utility engineers — 6 days; and riggers and linemen — 40 days [DOC, 1985].

2. Prediction methods

The extent of the re-radiation problem can be predicted by analytical methods, by scale-model tests or by full-scale measurements.

2.1 Analytical methods

An accurate method and one found to correlate reasonably well with scale model and full-scale tests, is the numerical electromagnetics code (NEC), developed at the Lawrence Livermore Laboratories in 1979 [Burke *et al.*, 1979].

It is a moment-method analysis, wherein all significant current conductors are broken into segments of 0.1 wavelength or less and represented by single wires of appropriate diameter or by plates of appropriate dimension. The program continues to adjust currents and phases between predetermined boundaries in each of these segments until the laws of electromagnetics are satisfied. NEC is a powerful tool, but has the disadvantage that it requires a major computer and is expensive in the time required for the run and for its analysis.

Another moment-method approach is provided by the Richmond program [Richmond, 1974a and b]. However, the NEC appears to be more versatile and has been used to analyse towers, buildings and power lines.

By use of transmission line theory, much less complex computer programs for analysis of power lines can be developed. One program by an electric power utility — Ontario Hydro — shows pattern distortions similar to those from the moment-methods but with re-radiation at resonant peaks lower by a factor of about 1.8. Another

program, simple enough to be run on a mini-computer and known as AMPL is recommended for initial analysis of power lines in the Canadian report [DOC, 1985]. The program is listed in that report. Agreement with the moment-methods is good up to 1100 kHz but less so, at the top of the MF band.

2.2 Scale model tests

Modelling of a transmitting antenna and of re-radiating objects usually at scales of 200 or 600 to one, has been done using radio ranges and an anechoic chamber in Canada. On the radio range, the models are placed on a conductive surface turntable and measurements are made at the edge of the conductive surface screen. In the anechoic chamber, no earth screen is used and models are made double their normal height so as to include their images in the earth.

2.3 Full-scale tests

Tests in the field have included numerous "before and after" tests on new power lines wherein the broadcast radiation pattern was measured with particular care ahead of and after construction of the power line and again after appropriate de-tuning procedures had been completed. Other tests involved the temporary installation of a low-power transmitter near isolated buildings or isolated test sections of power transmission lines. Details are included in the Canadian report [DOC, 1985].

3. Assessment procedures

Where construction of either a new broadcasting transmitter or a new re-radiation hazard is anticipated, a preliminary assessment of the potential effects is suggested, followed by the corrective action indicated in the circumstances.

3.1 Power transmission lines

Preliminary assessment using the program AMPL would give an indication of the seriousness of the problem. This could then be followed, if required, by an NEC analysis and appropriate procedures could be suggested. One Canadian station which had a fairly critical directional pattern, was faced with the construction of two nearby power lines. Computer projections by the moment-method indicated in which power line towers high re-radiating currents would flow and that re-radiation would indeed be excessive. The lines were built, and intolerable re-radiation was confirmed in field measurements. The most severely offending towers then had their sky wires isolated and the re-radiation dropped to acceptable levels.

3.2 Masts and towers

Using moment-method computation, the Communications Research Centre developed graphs of "scattering cross-section" (σ/λ^2) for a representative set of masts or towers which permit rapid assessment of the amplitude, though not the phase, of the re-radiation component in a field [Royer, 1985] (see Fig. 1).

At point, p , at distance, d :

$$|E| = |E_p| \pm |E_s|$$

where:

E_s : scattered field, and

E_p : normal field.

E_s can be determined, having obtained σ/λ^2 from the curve and knowing the incident field at the scatterer, E_{s0} , from:

$$\sigma = 4\pi d_p^2 \left| \frac{E_s}{E_{s0}} \right|$$

$|E|$ can be determined for any horizontal or vertical angle using moment-method analysis. Royer provides an extensive family of curves for this purpose. If the field value is of concern, appropriate action can be initiated.

3.3 High-rise buildings

Royer's work at the Communications Research Centre (see § 3.2) also included predictions for a representative set of buildings, modelled as wire grids. An example is given in Fig. 2. Techniques similar to those in § 3.2 can be employed.

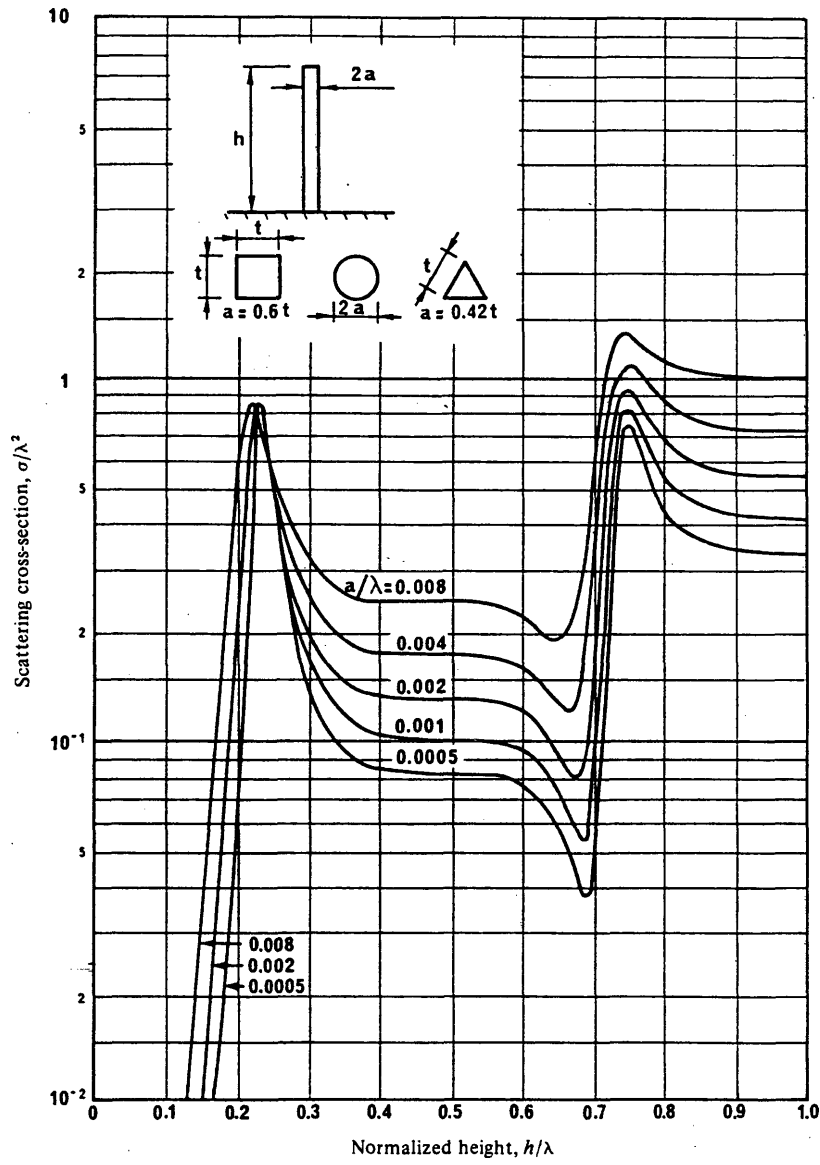


FIGURE 1 – Scattering cross-section for a representative set of masts and towers

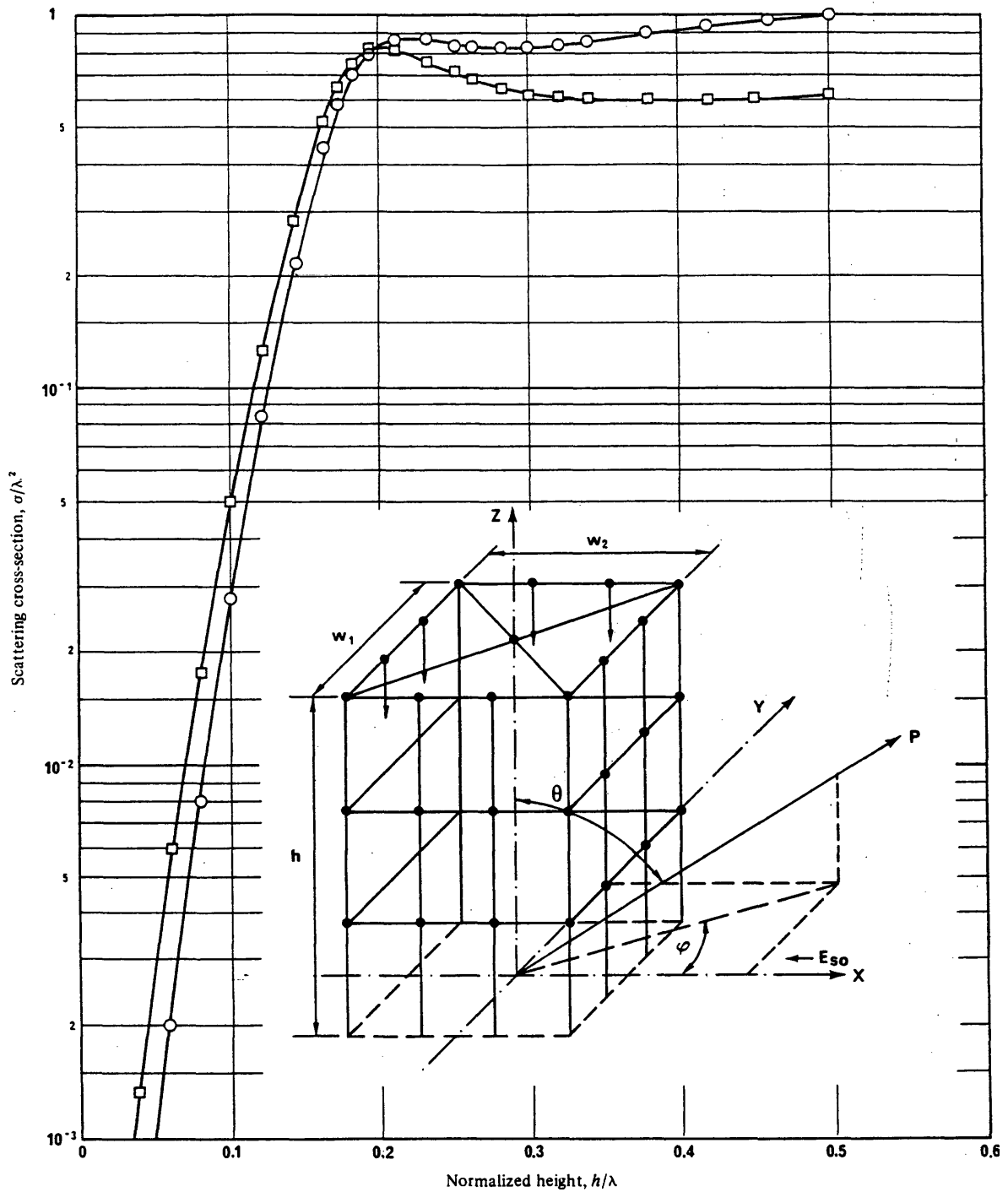


FIGURE 2 – Scattering cross-section for a representative set of buildings; modelled as wire grids

- $\square$: $\phi = 0^\circ$
 $\circ$: $\phi = 180^\circ$
 $w_1 / \lambda = w_2 / \lambda = 0.200$
 a_0 (Wire radius) / $\lambda = 0.0016$

4. De-tuning techniques

4.1 *Isolation of sky wires*

Where power lines create a problem, a common technique is to install insulators between the towers and the sky wire, which is normally earthed at these points. This minimizes current flow in the adjacent loops. This method must be used judiciously, since it creates additional hazards in the case of lightning or fault-induced current flow and also may cause resonance at another frequency, affecting another station in the area.

4.2 *Stubs*

When a structure or loop is resonant, standing waves of current and voltage occur. Introducing an insulator at the point of maximum current flow would be the most effective cure but in most instances, this is impractical. An open quarter-wave stub can introduce high impedance at the point and is normally used for this purpose. Unfortunately, on power lines, maximum current on the sky wire usually occurs some distance out from the tower. A stub or insulator inserted at that point is very effective but finds little favour with power authorities who want no extraneous installations mounted above the power conductors. Stubs are also difficult to install or adjust in this location. Normal practice therefore has been to connect stubs to the tower legs just below the power conductors, suspending them about 1 m out, parallel to the legs and terminating them on the tower leg some 5 m above ground via a tuned reactance. Since the stubs are invariably less than one-quarter wavelength long, a variable capacitor is used. It is tuned to minimize current at the base of the tower leg.

An alternative tower stub has been developed, known as the "elbow stub" [DOC, 1985]. This starts at the tower just below the power conductors but is guyed out from the tower so as to form a right angle at mid-length, then return to the tower near the earth where the tuning components can be adjusted. Elbow stubs for two frequencies are found to be very effective when mounted on orthogonally opposed corners of the tower. Elbow stubs have also been proved to have a wider bandwidth than stubs parallel to the tower legs.

4.3 *Buildings*

Scale model and analytical predictions of re-radiation from buildings show reasonable agreement and are amenable to de-tuning techniques. However, the uncertain impedance of footings and the lossy nature of material such as concrete used in their construction cause somewhat anomalous results in de-tuning procedures at full scale.

De-tuning can be provided by stubs such as used on towers but the stubs are usually not acceptable aesthetically. An alternate stub arrangement can be provided by a roof-top stub supported 3 or 4 m above the building periphery, and connected through a suitable reactance to the lightning protection system. An improved roof-top stub is afforded by mounting an umbrella of a few wires above the roof, connecting their common point in a manner similar to the above.

In full-scale measurements in Canada, it was found that de-tuning, computed to provide about 18 dB reduction, resulted in only about 4 or 5 dB improvement. As explained above, this is attributed to the lossy nature of the structure. Further investigation is warranted.

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REPORT 458-5

CHARACTERISTICS OF SYSTEMS IN LF, MF AND HF BROADCASTING

(Question 44/10)

(1970-1974-1978-1982-1986-1990)

1. Introduction

Question 44/10 relates to the possibility of standardizing one or more sound broadcasting systems on a world-wide basis. It is clear that the study of this complex question is not sufficiently advanced to achieve this aim. The present Report, therefore, is only a summary of the information available, intended to encourage administrations, broadcasting organizations and industry, to take an interest in these questions and to undertake the studies necessary to solve them.

At present, broadcasts in bands 5, 6 and 7, unlike sound and television broadcasting in band 8 (VHF), are operated throughout the world with an almost complete absence of internationally standardized transmission characteristics; with the exception of channel spacings and carrier frequencies for bands 5 and 6, but even these differ from Region to Region. The other transmission characteristics vary from country to country and in many cases even from transmitter to transmitter [EBU, 1971].

2. Systems available for standardization

The following list of possible systems cannot, at the present time, be considered complete. Several of these systems are compared in [Haviland, 1969]. This study also shows that interference between transmitters must be considered when defining a system and the importance of well defined channel spacings becomes apparent.

Each of the above systems may incorporate modulation processing devices (see Note 1). If such a device is necessary in the receiver to obtain full advantage of a similar device in the transmitter, the code should be completed by a suitable abbreviation. Thus, an amplitude-modulation double-sideband (DSB) system with envelope demodulation comprising a compressor in the transmitter and an expander in the receiver, is coded AM-DSB-ENV-COMPANDOR. An example is given in Annex I.

Descriptions of amplitude-modulation double-sideband systems with synchronous detection appear in [Netzbund, 1969] and [CCIR, 1966-69] and it should be noted that only these systems can serve as transition systems from amplitude-modulation double-sideband systems with envelope detection to single-sideband (SSB) systems with synchronous detection. Receivers based on synchronous detection would produce undistorted audio-frequency signals for both the above systems ("receiver compatibility").

The transmission of compatible single-sideband signals (see Note 2) in sound broadcasting should be considered to be inappropriate for regular programme services, and consequently be restricted to exceptional cases for the following reasons:

- it is difficult to reduce out-of-band emissions in the suppressed sideband of a compatible SSB system;
- strong out-of-band emissions will result from maladjustments of compatible SSB transmitters;
- imperfections of conventional DSB receivers which have no bearing on the reproduction quality of DSB transmissions may give rise to dynamic distortion, when the automatic gain control follows a portion of the spectrum with high energy density rather than the carrier, or to non-linear distortion due to the amplitude and phase response within the radio-frequency and intermediate frequency stages;

- additional non-linear distortions, similar to those which affect double-sideband signals, which are to be expected when receiving sky-wave signals, cannot be reduced by using a different demodulation technique than envelope detection.

Note 1. — By "modulation processing" is understood any process consisting of altering certain characteristics of the modulation, such as the dynamic range, audio-frequency bandwidth, etc.

Note 2. — A transmission is said to be "compatible" if it can be received on an existing conventional receiver, without any modification whatsoever to the receiver, and then gives a quality of reception at least as satisfactory as that obtained at present in a double-sideband system.

TABLE I

Modulation	Detection	Code
Amplitude-modulation, double sideband	Envelope detection	AM-DSB-ENV
Amplitude-modulation, double sideband	Synchronous detection	AM-DSB-SYNC
Single-sideband, amplitude-modulation	Synchronous detection	SSB-SYNC
Frequency modulation (narrow band)		FM

3. Characteristics to be specified

For the systems mentioned in § 2, the characteristics whose standardization will be required are given below. This list is not necessarily complete.

3.1 All systems

- channel spacing,
- carrier frequencies,
- intermediate frequency or frequencies,
- receiver oscillator frequency stability.

Note 1. — The relationship between these characteristics is shown in Annex II.

- audio-frequency bandwidth of the programme,
- necessary bandwidth of emission,
- overall bandwidth of the receiver.

Note 2. — For the relationship between these characteristics and the channel spacing, see Annex I to Recommendation 639. Deviations from standardized values may be tolerated, if they do not result in unacceptable interference.

- characteristics of modulation processing devices.

In addition, the following characteristics should be standardized:

3.2 Amplitude-modulation double-sideband systems

- maximum depth of modulation.

When laying down the characteristics of the transmission system and the reference receiver, in order to obtain the best adaptation of a frequency-assignment plan to receivers of reasonable performance, the following rules should be taken into account:

3.2.1 The transmission system and receiver characteristic should be suitably related, particularly in the amplitude/frequency response. Harmonic distortion should be reduced to acceptable values. In this respect, criteria for the determination of tolerances for the overall amplitude and phase characteristics and ways for the assessment of practical values are set out and discussed in [Makiedonski, 1974].

3.2.2 The audio-frequency bandwidth transmitted should be related to the carrier spacing (see Annex I to Recommendation 639). (The precise bandwidth cannot be given: some administrations are of the opinion that it should be one half of the spacing between carriers, and some of them believe that this value may even achieve the value of unity; but this depends on the absolute channel spacing.)

3.2.3 Uniform carrier spacings, with nominal carrier frequencies being an integral multiple of the carrier spacing, should be adopted, at least within the broadcasting bands 5 and 6. (There are technical advantages also in the adoption of uniform carrier spacing on a world-wide basis, including band 7.)

3.2.4 The intermediate frequency (or frequencies) of the receiver should be chosen so that interference created internally within the receiver is at a minimum. With stable local oscillators (see Annex II) this condition is best met when the intermediate frequency (or frequencies) of the receiver is an integral multiple of the carrier spacing.

3.3 *Single-sideband systems with synchronous detection*

- degree of reduction of the carrier wave (see Recommendation 326, § 1.5),
- degree of suppression of the unwanted sideband,
- maximum permissible values of intermodulation products,
- auxiliary signals to obtain receiver synchronization.

The considerations under § 3.2 would apply, and where possible the same characteristics should be adopted.

Note 3. — More information about SSB systems is contained in Report 1059.

3.4 *Frequency modulation*

- modulation index,
- maximum modulation frequency.

4. **Stereophonic systems in AM broadcasting**

[CCIR, 1974-78] describes four methods of stereophonic broadcasting for use in bands 5 and 6. All systems described in this Document use conventional amplitude modulation for the stereophonic sum signal and angle modulation of the same carrier frequency for the stereophonic difference signal. They differ in the way the modulations are generated and in the resulting distribution of the audio information in the various radio frequency sidebands.

4.1 *Amplitude modulated stereo system used in Australia and Brazil*

The Australian Administration has, from February 1985, allowed its MF stations to broadcast in stereophony. Compatible quadrature amplitude modulation (CQUAM) has been adopted as the stereophonic modulation standard. Field tests [CCIR, 1982-86] confirmed that the occupied bandwidth of stereophonic transmission is not significantly greater than monophonic transmission. The Brazilian Administration adopted the same system on an exclusive basis since January 1986.

5. **Preemphasis, deemphasis and 10 kHz audio bandwidth limitation for administrations employing 10 kHz channel spacing in band 6 (MF)**

With 10 kHz channel spacing broadcasting in band 6 (MF), a standard preemphasis, deemphasis and 10 kHz audio bandwidth limitation has been implemented [CCIR, 1986-90]. This produces a transmission/reception system with an overall audio frequency response that is essentially flat from 50 Hz to nearly 10 kHz (limited only by the receiver's choice of bandwidth), and reduces interference caused to stations operating +/- 20 kHz removed in frequency. Further, this entirely eliminates undesired high-order dynamic intermodulation products that contribute noise and interference in the MF band. This system is described in detail in Annex I.

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ANNEX I

THE USE OF VARIOUS MODULATION PROCESSING DEVICES

In the OIRT, studies have been made of the possibilities of increasing the modulation factor, i.e., the sideband power, by either trapezoidal modulation or dynamic-compression operation, with transmitters operating in bands 6 (MF) and 7 (HF) [CCIR, 1970-74a]. The results were that dynamic compression without clipping should be preferred, if the objective is to increase the sideband power with a minimum loss of quality (e.g. for music programmes). On the other hand, if a loss of quality is deemed unimportant (e.g. for speech programmes), trapezoidal modulation leads to a higher degree of sideband power.

Experiments carried out in the RFZ, Berlin, have confirmed the assumed increase of the sideband power as follows:

- *Compression of the dynamic range by 12 dB:*
average gain with a rise-time of 0.5 ms and a decay-time of 35 ms: ≈ 6 dB.
- *Compression of the dynamic range by 6 dB:*
average gain with a long decline period (1.5 s) for programmes with a wide dynamic range: ≈ 3 dB.
- *Trapezoidal modulation*
average gain with 5 dB increase in the level of the audio-frequency signal and clipping: ≈ 3 dB.

Studies have been carried out in Sweden concerning the improvement in radio-frequency wanted-to-interfering signal ratio obtained by using audio-frequency compression and expansion in connection with a double-sideband amplitude-modulation system and a frequency-modulation system with a maximum deviation of ± 5 kHz [CCIR, 1966-69].

The audio-frequency range was from 40 to 5000 Hz.

The compressor reduced the dynamic range of the audio-frequency signal, expressed in decibels, to half its value, the time-constants were 2 ms for the rise-time and 20 ms for the decay-time. The expander had characteristics reciprocal to those of the compressor.

The test results can be summarized as follows:

In the absence of interference, no change in quality was observed when using both compressor and expander in the system. The quality was also judged to be satisfactory by listening when only the compressor was used.

In the presence of co-channel interference, the radio-frequency protection ratios (dB) were found to be as follows:

	Type of modulation	
	Amplitude	Frequency
– without compressor and expander	40 to 50	40 to 45
– with compressor only	30 to 40	30 to 40
– with compressor and expander	20 to 25	25 to 30

It should be noted that these values were obtained when the unwanted transmitter was not equipped with a compressor.

In a more extensive study covering double-sideband amplitude-modulation only, the effect of compression and expansion on the radio-frequency protection ratio was investigated when the unwanted transmitter was also equipped with a compressor [CCIR, 1970-74b]. The study was made with different types of programmes transmitted by both the wanted and the unwanted transmitters.

When a speech programme was interfered with by another speech programme, the reduction of the radio-frequency protection ratio was about 15 dB when using compression and expansion. No deterioration in reproduction quality has been reported. When only compression was applied there was a smaller reduction in protection ratio of about 10 dB; in this case, the quality of the reproduced sound was considered lower than acceptable and significantly worse than when no compression was applied.

When a music programme was disturbed by either music or speech, the result depended to a great extent on the character of the wanted programme. The advantage obtained with compression plus expansion or with compression only, was always smaller than when the wanted programme was speech, and sometimes even negligible.

A new compression method has been developed in Japan based on digital techniques [CCIR, 1982-86a]. It is applicable to a wide variety of uses from high compression transmissions to high quality sound, with less degradation of signal quality.

The principle of operation is as follows:

- the signals are divided into segments of predetermined length that includes a certain number of zero crossing points;
- control is imposed individually on each segment;
- information for control of a segment is obtained from the segment itself;
- control is fixed within the same segment and changed only at the contours of segments.

For obtaining higher compression, this digital level compression method is able to solve most of the disadvantages inherent in conventional analogue methods. For various applications, the compression characteristics are easily selectable by reading the correct ROM addresses and the dynamic range of the original signal can be retrieved at the receiving end by the received control information.

With these features this compression method has a wide variety of applications.

Studies in the United States have resulted in the adoption of a standard preemphasis, deemphasis and 10 kHz bandwidth limitation that improves the overall audio frequency response and reduces interference [CCIR, 1986-90a]. Its characteristics are described below as they apply to U.S. broadcasting in the MF band.

MF Transmission Preemphasis

Preemphasis is the boosting of high audio frequencies prior to modulation and transmission. Today, most MF stations use preemphasis to varying extents. This preemphasis is employed in an attempt to compensate for the "narrow" response of most receivers. If preemphasis is not controlled, one station may interfere with MF receivers tuned to neighboring stations located on adjacent channels. Whether such interference is objectionable will depend on (1) the response characteristics of the receiver, (2) the amount and nature of transmission preemphasis, (3) the extent to which the station is employing compression/limiting techniques, and (4) whether the AM transmission system is bandlimited in the audio processor, transmitter or antenna.

Preemphasis is useful for improvement of the MF transmission-reception system audio response only to a limited extent for receivers using IF transformers. Many receivers using ceramic filters with narrow response characteristics cannot be improved by use of excessive preemphasis. These receivers do not recover the transmission of preemphasized high audio frequencies.

Each MF broadcast station should broadcast with audio preemphasis as close as possible (within the capabilities of the station's transmission system) to the recommended standard, without exceeding it. The curve applies for audio frequencies up to 10 kHz. See Figure 1. The curve describes the recommended transmission system static audio response of an AM station.

The recommended preemphasis curve is a single zero curve with a break frequency at 2122 Hz. It is similar to the 75 microsecond curve used for Region 2 FM broadcasting. To reduce the peak boost at high frequencies, a single pole with a break frequency of 8700 Hz is employed. Analysis has shown that the proposed curve is compatible with most existing AM receivers.

Measuring Performance.

The preemphasis curve for MF use is a static curve, and cannot be measured dynamically. Studies have shown that the dynamic and non-linear functions performed by most AM station audio processors will modify any given preemphasis curve. In addition, it is the audio response of the entire AM transmission system that indicates performance in accordance with the standard. For these reasons, measuring a station's preemphasis curve for the purpose of determining compliance with this standard should be performed in accord with the following specifications:

1. Compliance with the curve should be measured by sweeping the station's transmission system with audio tones. The dynamic functions of the MF station's processor, but not the frequency shaping circuits, must be disabled (i.e., in "proof" mode).
2. The net transmission system audio response is best measured by detecting the over-the-air signal. This will ensure that the AM transmitter and antenna combination are faithfully reproducing the preemphasized audio. Alternatively, if the transmitter and antenna combination is reasonably broadband, performance can be determined by static measurement of the audio signal prior to modulation.

MF Receiver Deemphasis

Receiver deemphasis results from the selectivity characteristics of a receiver's RF and IF stages and the response characteristics of the receiver AF section. A standard deemphasis curve permits MF broadcast stations to know, with certainty, the likely overall response characteristics of AM receivers.

AM receivers should complement the recommended transmission preemphasis characteristic by incorporating a net receiver system audio response described below. (The net system audio response of an AM receiver is the combined RF, IF, and AF audio response.) The deemphasis curve is characterized by a single pole at 2122 Hz and a single zero at 8700 Hz. It is the precise complement of the preemphasis standard described above. The preemphasis/deemphasis standards apply only for audio frequencies below 10 kHz; the implementation of preemphasis/deemphasis standards produces a transmission/reception system that is essentially flat to nearly 10 kHz and limited only by the AM receiver's choice of bandwidth.

Measuring Performance.

The deemphasis characteristic should be determined by measuring the overall frequency response in accordance with International Electrotechnical Commission ("IEC") Publication 315-3, Clause 11.2.

Although an optional enhancement to an AM receiver, using notch filters is recommended. If used, the notch filter should (1) have as high a "Q" as is practical, (2) adequately suppress the interfering carriers, and (3) not unduly degrade the desired bandwidth performance of the AM receiver.

10 kHz Bandwidth for MF Transmission.

Each MF broadcast station should modulate its transmitter with an audio bandwidth described by the specification in Figure 2. Appropriate and carefully designed audio low-pass filters as the final filtering prior to modulation can be used to implement this specification. The purpose of the bandwidth specification is to remove interference by controlling the occupied RF bandwidth of AM stations.

The audio envelope input spectrum to the MF transmitter should be -15 dB at 10 kHz, smoothly decreasing to -30 dB at 10.5 kHz, then remaining at -30 dB from 10.5 kHz until 11.0 kHz. At 11.0 kHz, the audio bandwidth should be -40 dB, smoothly decreasing to -50 dB at 15 kHz. Above 15 kHz, the audio bandwidth should remain at least -50 dB. The reference level is 1 dB above a 200 Hz sine wave at 90% negative modulation.

Measuring Performance.

An MF station is determined to be in compliance with this bandwidth characteristic by measurement of the station's audio bandwidth in accordance with the following parameters:

1. Audio bandwidth measurements should be obtained at the audio input terminals to the AM transmitter. For AM Stereo stations, audio bandwidth should be measured at the L+R audio input terminals to the RF modulator. Note that the bandwidth standard characterizes an audio bandwidth that represents station program material that has been modified by possibly non-linear circuits in the station's audio processor. For this reason, it is recommended that a test signal be used that adequately characterizes typical audio program material, rather than relying on static audio test tones. However, it may still be useful to measure bandwidth statically at the time that AM preemphasis is measured.

2. Audio bandwidth should be measured using a test signal consisting of USASI (United States of America Standards Institute) noise (see Figure 3) that is pulsed by a frequency of 2.5 Hz at a duty cycle of 12.5%. USASI noise is intended to simulate the long-term average spectra of typical audio program material. Pulsing of the noise is intended to simulate audio transients found in audio program material. USASI noise is a white noise source (*i.e.* noise with equal energy at all frequencies) that is filtered by (1) a 100 Hz, 6 dB per octave high-pass network and (2) a 320 Hz, 6 dB per octave low-pass network. See Figures 3 and 4. A pulsed USASI noise generator is shown in Figure 4. Using the attenuator pad, the ratio of peak-to-average amplitude should be 20 dB at the audio output of the pulser. The station's audio processor must be in normal operating mode.

3. A suitable swept-frequency or FFT (Fast Fourier Transform) spectrum analyzer should be used to measure compliance with the bandwidth specification. When a swept-frequency audio spectrum analyzer is used to measure compliance with the bandwidth specification, the analyzer's setup should consist of: (a) a 300 Hz resolution bandwidth, (b) 2 kHz/horizontal division, (c) 10 dB/vertical division, (d) a reference of 1 dB above 200 Hz (sine wave) 90% negative modulation, and (e) a display set to maximum peak hold (or equivalent function). The analyzer's operating span and sensitivity are adjusted as necessary to determine compliance. When a FFT analyzer is used to measure compliance with the bandwidth specification, the analyzer's setup should consist of (a) a reference set to 1 dB above 200 Hz (sine wave) 90% negative modulation, (b) Hanning window, (c) a horizontal span of 20 kHz, (d) a dynamic range of 80 dB (or available range), (e) a display set to maximum peak hold (or equivalent function).

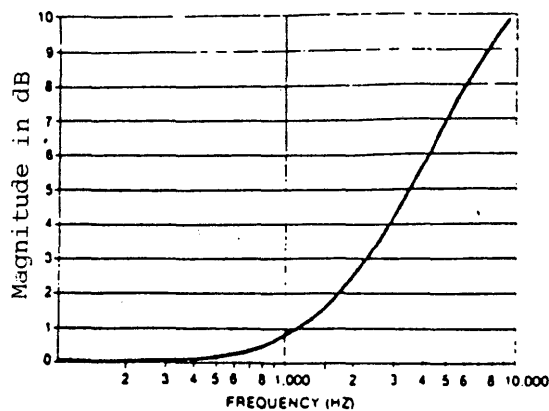


FIGURE 1 - Modified 75 μ s AM Standard Preemphasis Curve

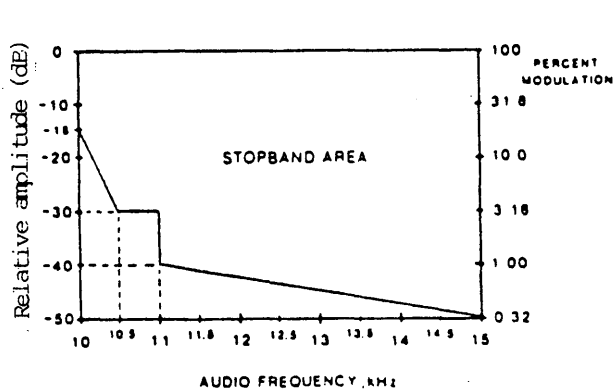


FIGURE 2 - NRSC Stopband Specification
(Audio Envelope Input Spectrum to AM Transmitter)

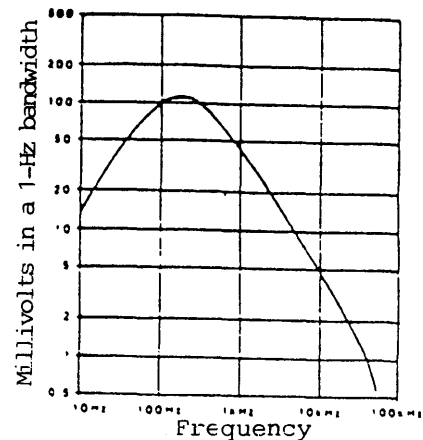


FIGURE 3 - Spectra of USASI Noise

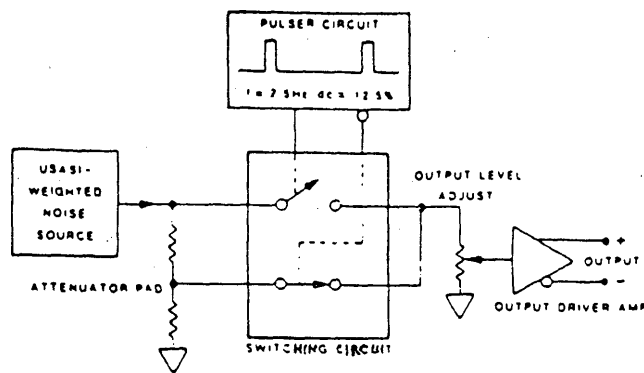


FIGURE 4 - Pulsed USASI Noise Generator

4. In the USSR [CCIR, 1986-90b], a method of measuring the mean modulation depth of double-sideband AM broadcasting signals has been developed.

The mean modulation depth of the AM signal is determined by the formula:

$$\bar{m} = \frac{\pi}{2} \cdot \frac{\int_0^T |U_A(t) - U_0| dt}{U_0 T}$$

where:

U_0 : signal carrier voltage amplitude,

$U_A(t)$: instantaneous value of AM signal amplitude,

T : averaging time.

The mean modulation depth is equal to the product of the maximum value of the modulation depth and the parameter of the AF modulating signal, designated the "relative mean voltage".

$$U_{rmv} = \frac{\pi}{2\sqrt{2}} \cdot \frac{\int_0^T |U(t)| dt}{U_H T}$$

where:

U_{rmv} : relative mean voltage

$U(t)$ - instantaneous value of the modulating signal,

U_H - voltage corresponding to the nominal level at the point of measurement.

The parameter U_{rmv} may be used for the numerical calculation of the variation in modulation depth obtained as a result of the dynamic compression of the modulation signal.

A digital broadcasting signal parameter meter has been developed which uses an algorithm to calculate the mean modulation depth and the RMV in accordance with the formulae given above. The instrument has two measuring channels, which enables simultaneous measurements to be made at the modulator input and the transmitter output, or at the input and output of the processor. It is used to monitor double-sideband AM transmitters and broadcasting signal preliminary processing devices.

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ANNEX II

CHANNEL SPACING, PROTECTION RATIO AND INTERMEDIATE FREQUENCY

When choosing the carrier frequencies, channel spacing and also the intermediate frequencies to be used in receivers, it is important that they should be chosen so as to minimize interference from:

- the local oscillators of the receivers in use, or of nearby receivers, either by a fundamental or a harmonic frequency;
- harmonics of a transmitted frequency, or other possible intermodulation products [CCIR, 1963-66; SCART, 1966; CCIR, 1970-74].

If both the carrier frequencies and the intermediate frequency are an integral multiple of the carrier spacing, then all interfering products will also be integral multiples of the carrier spacing. Theoretically, therefore, maximum protection could then be obtained because the frequency difference between any interfering signal of this kind and the wanted carrier frequency, would be zero or a multiple of the channel spacing.

If these requirements are to be met with in a particular broadcasting band, it would be essential for the channel spacing to be uniform throughout the band. It would be more advantageous, moreover, if this condition could be met with in both bands 5 and 6, or better still, throughout bands 5, 6 and 7. On the other hand, this condition should be satisfied on a world-wide scale or at least in those areas, where a single frequency assignment plan exists or will be established [Eden, 1967].

However, it must be noted that the disturbance caused by an interfering signal increases rapidly as its frequency difference from the wanted signal increases from zero.

Under present-day conditions in band 7 the frequency differences might have any possible value and this may require an additional protection ratio of up to 17.5 dB. With the adoption of the proposed arrangement, the maximum frequency difference would depend on the accuracy with which the local oscillator frequency and the centre frequency of the intermediate-frequency passband can be controlled. To achieve an improvement close to the maximum possible it would be necessary to achieve stabilities of the order of 100 Hz. As far as the intermediate-frequency stability is concerned this could be achieved by using ceramic or mechanical filters rather than using conventional intermediate-frequency coils. The control of the initial tuning operation and the frequency drift of the local oscillator may require special techniques in which automatic frequency control may be required. The adoption of the proposal would, therefore, give little improvement in the short term, with existing receivers, but would offer the chance of substantial improvement in the future, without any disadvantages under present-day conditions.

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REPORT 1059-1

CHARACTERISTICS OF SINGLE-SIDEBAND SYSTEMS IN HF BROADCASTING

(Question 44/10)

(1986-1990)

1. Introduction

This Report has been developed in recognition that consideration is being given to the progressive introduction of single-sideband broadcasting systems into the bands now allocated to the Broadcasting Service at HF. Introduction of this form of broadcasting can be accomplished with greater technical harmony if there is sufficient guidance concerning the technical parameters involved. The following considerations will be concentrated on single-sideband amplitude modulation with synchronous demodulation. With respect to a necessary transition period from DSB to SSB, some consideration must also be given to the reception of SSB signals with reduced carrier by receivers with envelope detection. At the end of the transition, all of the advantages of SSB transmissions could then be realized, as follows:

- a more efficient utilization of the frequency spectrum and a reduction of interference;
- the capability of improving the required protection ratio between adjacent channels in the case of a sufficient carrier reduction;
- the capability of improving the quality of reception, in particular under poor propagation conditions (selective fading), with SSB receivers.

The modulation technique considered most suitable for the achievement of bandwidth saving is some form of SSB system. WARC-79 Recommendation 501 and No. 302 of the Radio Regulations suggest the use of SSB emissions to the maximum possible extent in AM systems. There are two types of such systems, namely single-sideband (SSB) and compatible single-sideband (CSSB).

The CSSB system is not suitable for use in amplitude-modulated sound broadcasting, principally because of its increased distortion; furthermore, a greater radiofrequency bandwidth is needed and adequate suppression of out-of-band emissions is likely to be difficult at HF.

In the event of the introduction of single-sideband amplitude modulation broadcasting, it would seem desirable to use the definitions existing in Recommendation 326.

According to this Recommendation, the carrier component is defined in relation to the peak envelope power P_p of a radio transmitter, by the acceptable intermodulation level D_n .

For single-sideband broadcasting transmitters, the acceptable intermodulation level D_n determines the non-linear distortion (quality), and the out-of-band radiation (adjacent channel interference).

In single-sideband reduced carrier systems, the precision of the locally re-inserted carrier is important for the reception quality.

The system parameters for a future SSB-system for sound broadcasting in band 7 (HF) must be chosen in such a way that the different requirements of the transition period (reception of SSB-signals with receivers using envelope detection) as well as those of the period thereafter, when only receivers with synchronous demodulation will be used, are taken into account [CCIR, 1978-82a].

The system specification of SSB is presented in Recommendation 640.

2. Fundamental technical characteristics of single-sideband broadcasting systems

2.1 Nature of single-sideband (SSB) modulation

References to single-sideband (SSB) frequently are made considering the transmitted signal to be amplitude modulated (AM). However, in fact, the transmitted SSB-signal is a composite of both amplitude and phase modulation. Both components of modulation contain the same information, but differ in phase by 90° . Both the amplitude and the phase information can be recovered independently using a synchronous detector that is phase-locked to the incoming carrier. If the phase of the reinserted carrier is the same as that of the incoming carrier, then the amplitude modulation is recovered. On the other hand, if the phase of the reinserted carrier is shifted by 90° relative to the incoming carrier then the phase information is recovered.

For reasons mentioned above, it can be concluded that:

- additional services (data, etc.) using additional phase modulation are difficult to implement in an SSB-signal;
- an SSB-transmission is less affected by selective fading, compared to a DSB transmission. This improved performance results from the constant level of the regenerated carrier, but also from the fact that the synchronous detector's performance is not degraded with shifts in phase of the incoming carrier (during periods of selective fading) because either or both phase and amplitude components of the SSB signal are used in detection.

2.2 Peak envelope power (P_p)

The nominal power of an SSB transmitter is designated by its peak envelope power P_p . This P_p is specified as the mean power produced during peak conditions employing full modulation. Therefore it is difficult to measure utilizing a thermic power meter connected with a dummy load unless the input signal is sinusoidal.

2.3 *Carrier and sideband power relationships in DSB and SSB systems**

The sideband power P_s of a transmitter, depends on the peak envelope power P_p and the chosen carrier reduction, a .

For a single-sideband broadcasting transmitter, the most appropriate value of a will depend mainly on the carrier recovery requirements in a low cost single-sideband receiver. To be able to produce the reference carrier for the synchronous demodulator at an acceptable cost, carrier reduction must be restricted to between about 6 and 12 dB, at the transmitter.

* According to the Radio Regulations (see also Recommendation 326) the following power terms should be used:

- peak envelope power (PX);
- mean power (PY)
- carrier power (PZ)

To enable uniform presentation in this Report, which includes a large number of other power and voltage terms the following designations differing from those recommended by the Radio Regulations and Recommendation 326 are used:

$$\begin{aligned} PX &= P_p \\ PY &= P_m \\ PZ &= P_c \end{aligned}$$

Figure 1 shows the relation between the sideband power P_s and the carrier reduction, a , for a given peak envelope power of the transmitter. The indication (1) denotes the values for modulation by a sinusoidal signal and (2) denotes the values for noise or programme modulation. The values of the voltages U and the powers P are given as percentages related to their peak envelope values. With programme modulation and a carrier suppression of more than 30 dB, the sideband power P_s (2) will be about 10% of P_p . A transmitter operating with a carrier reduction of 6 dB can, when modulated with a programme signal, only radiate a sideband power P_s (2) of about 2.5% of its nominal peak envelope power.

For the calculation of P_s (2) it has been assumed that the ratio of the mean power to the peak envelope power is 0.1 (see Recommendation 326, Table I).

2.4 Equivalent sideband power

To replace a DSB transmitter (carrier power P_c) for sound broadcasting in band 7 (HF) with an equivalent SSB transmitter, the sideband power of the SSB transmitter must be double the value of the sum of both sideband powers of a DSB transmitter. This is justified at least during the transition period, when SSB emissions will also have to be received by DSB receivers having a bandwidth equal to $2B_v$, twice as large as necessary for reception of the SSB emission, and also because the coverage of the target area must be the same.

After the transition period, when only receivers with ~~synchronous demodulation~~ and half the bandwidth of a DSB receiver will be used, theoretically the sideband-power of all SSB transmitters could be reduced to one half of the original value. The signal-to-noise ratio at the output of the receiver, however, would be unchanged only if the spectrum causing the co-channel and adjacent-channel interference within the receiver passband had a uniform power density. Because of the presence of a carrier and the unequal distribution of power in the spectrum of sound-broadcasting emissions, this condition is not met in practice [Gröschel, 1978].

Thus, for the equivalent sideband power P_s of the future SSB transmitters in HF sound broadcasting the following expression might be more suitable:

$$P_s(\text{SSB}) \approx (1 \text{ to } 2) \times P_s(\text{DSB})$$

A factor nearer to 2 might be realistic.

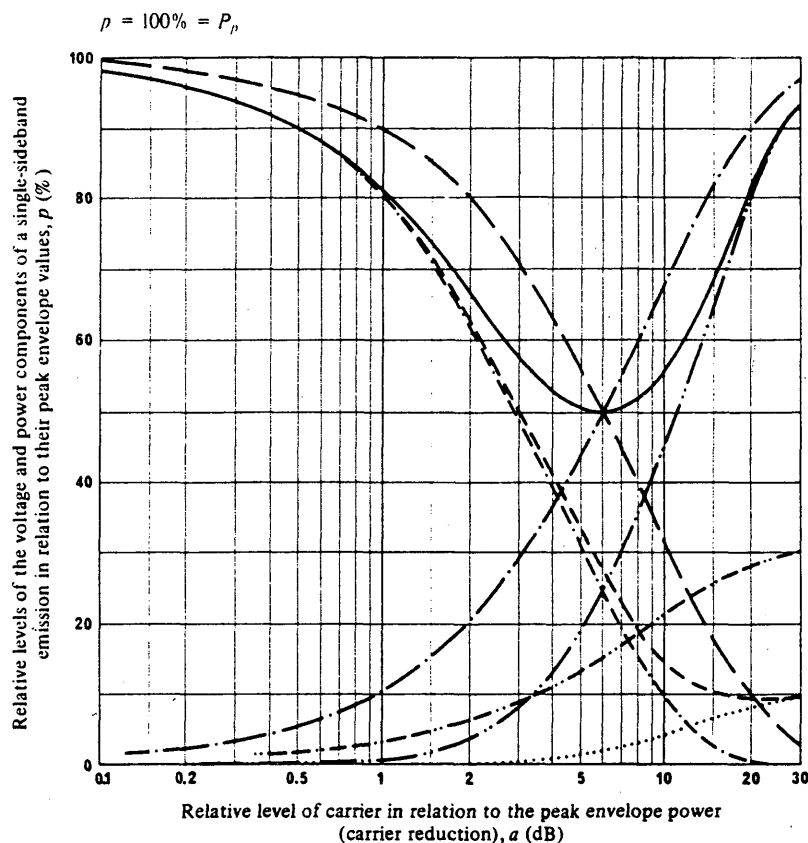


FIGURE 1

- P_p : peak envelope power of the transmitter (defined by means of the acceptable intermodulation level D_n)
- U_c : carrier voltage (r.m.s. value)
- - - $U_s(1)$: sideband signal voltage for modulation with a sinusoidal signal (r.m.s. value)
- · - · $U_s(2)$: sideband signal voltage for programme or noise modulation (r.m.s. value)
- · · P_c : carrier power
- - - - $P_s(1)$: sideband power for modulation by a sinusoidal signal
- · · · $P_s(2)$: sideband power for programme or noise modulation
- $P_m(1)$: transmitter mean power for modulation by a sinusoidal signal
- - - $P_m(2)$: transmitter mean power for programme or noise modulation

Some more details on the necessary peak-envelope power of future SSB transmitters replacing present DSB transmitters in HF sound broadcasting are given in Fig. 2. During the transition period the special receiving conditions will necessitate a peak-envelope power of the transmitter about 4 times higher than will be needed after the transition period [CCIR, 1978-82a].

The WARC HFBC-1987 defined equivalent sideband power in the following manner:

When the carrier reduction relative to peak envelope power is 6 dB, an equivalent SSB emission is one giving the same audio-frequency signal-to-noise ratio at the receiver output as the corresponding DSB emission, when it is received by a DSB receiver with envelope detection. This is achieved when the sideband power of the SSB emission is 3 dB larger than the total sideband power of the DSB emission. (The peak envelope power of the equivalent SSB emission and the carrier power are the same as that of the DSB emission.)

2.5 Comparison of power consumption of equivalent DSB and SSB transmitters

One of the aims of the introduction of SSB modulation in sound broadcasting in the HF bands, apart from a better utilization of the spectrum, is a considerable reduction of transmitter power consumption. Optimum saving of power can eventually be achieved when SSB receivers (having only half the bandwidth of DSB receivers) are in common use, thus permitting a change in carrier level from -6 dB to -12 dB.

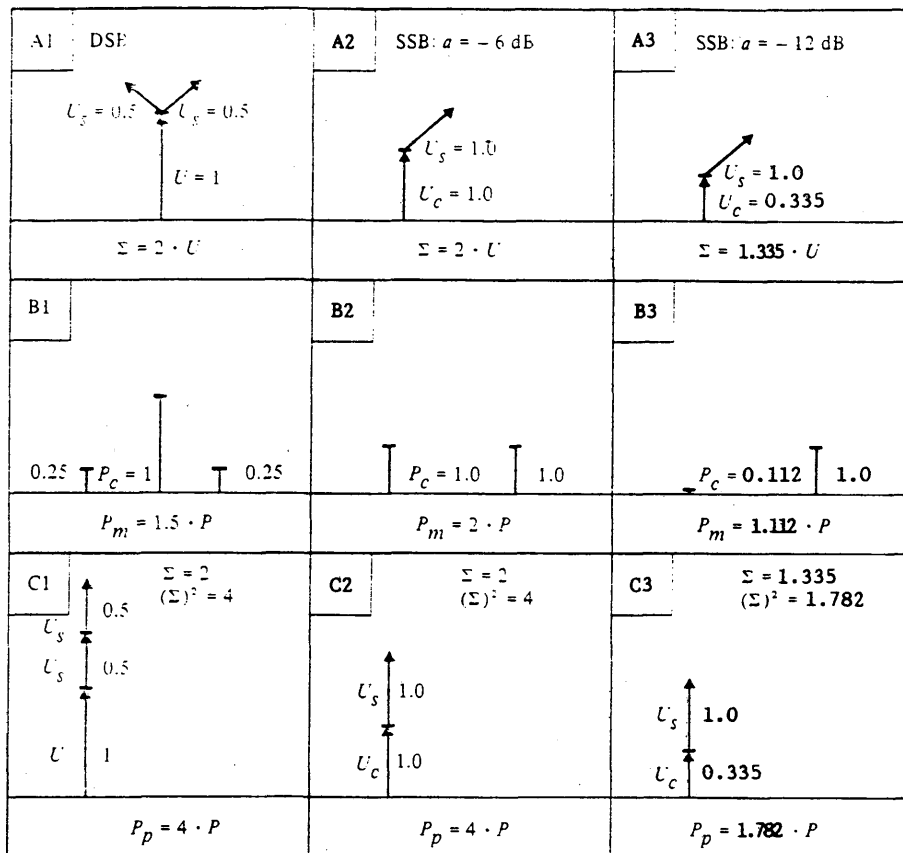


FIGURE 2 – Relation between carrier- and sideband-levels and powers with DSB and SSB transmission depending on carrier reduction

(For comparison, carrier power level of the DSB transmitter was assumed to be equal to 1. Full modulation with a single sinusoidal tone.) $P_S(\text{SSB}) = 2 \times P_S(\text{DSB})$

A1...A3: level of spectral-components (at the output of the transmitter)

B1...B3: average output power of the transmitter

C1...C3: peak-envelope power at the output of the transmitter

Index 1: DSB signal

Index 2: SSB signal, carrier reduction: -6 dB (class of emission H3E)

Index 3: SSB signal, carrier reduction: -12 dB (class of emission R3E)

U : carrier level of the DSB transmitter

U_s : level of sidebands

U_c : carrier level of the SSB transmitter

P : carrier power of the DSB transmitter

P_c : carrier power of the SSB transmitter

P_m : average power

P_p : peak envelope power

On the assumption of programme modulation with 40% effective modulation factor, a value which is reached if the modulation signal is substantially compressed, representative values for the power consumption of transmitters for DSB and SSB modulation with equivalent sideband power have been calculated and are summarised in Tables I and II.

This comparison is based on an overall efficiency factor of 60% for DSB transmitters and of 38% or 60% for SSB transmitters. These values have been derived from existing SSB transmitters. In this case an SSB transmitter with e.g. 12 dB carrier reduction and with an equivalent sideband power (i.e. equal to the DSB sideband power) will have respectively about 40% or 25% of the total power consumption of a DSB transmitter (Figure 2 conditions apply).

During the transition period when reception with DSB receivers must be assumed, a carrier reduction of 6 dB is recommended. In this case an SSB transmitter with double sideband power has respectively about 70% or 7% more power consumption than a DSB transmitter (see Table II).

TABLE I - *Output power ratios*

Carrier reduction	Sinusoidal modulation $m_{\text{eff}} = 100\%$	Programme modulation $m_{\text{eff}} = 40\%$
6 dB	2.0	1.16
12 dB	1.11	0.27

Average output power of an SSB transmitter relative to DSB carrier power. ($m = 100\%$, conditions as in Figure 2.)

TABLE II - *Power consumption ratio*

Carrier reduction	Sinusoidal modulation $m_{\text{eff}} = 100\%$		Programme modulation $m_{\text{eff}} = 40\%$	
	SSB efficiency 38%	60%	SSB efficiency 38%	60%
6 dB	2.11	1.33	1.70	1.07
12 dB	1.17	0.74	0.40	0.25

Power consumption of an SSB transmitter relative to that of a DSB transmitter. DSB efficiency 60%. ($m = 100\%$, conditions as in Figure 2.)

2.6 Carrier reduction

The amount of carrier reduction of an SSB signal determines the additional non-linear distortion that occurs when this signal is demodulated by envelope detection in a conventional DSB receiver. This additional distortion depends both on the carrier reduction α , and on the modulation depth. Because of this additional distortion to be expected during the transition period, the carrier reduction should then not be more than -6 dB relative to peak envelope power (class of emission H3E). This conclusion was confirmed by a compatibility reception field test carried out at a distance of 2000 km from a transmitting site in Japan. Subjective assessment of the impairment by distortion of the received SSB programmes was made with reference to the received DSB programmes using the five-grade impairment scale of Recommendation 562. The results are shown in Fig. 3.

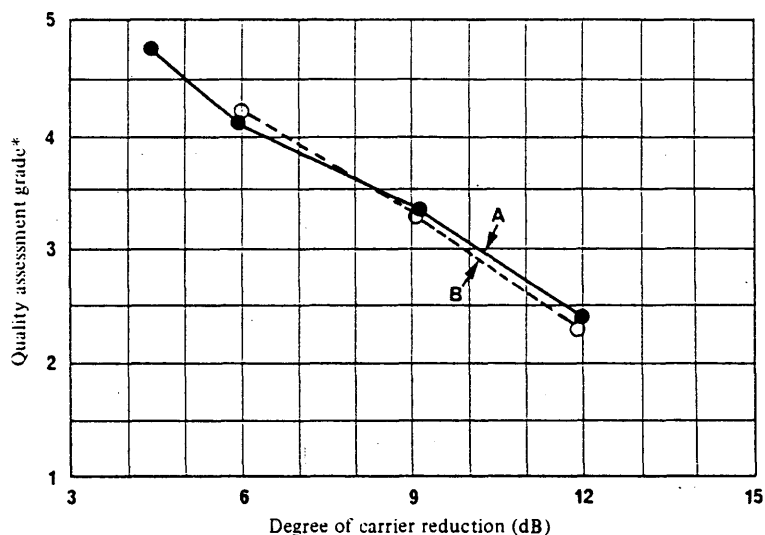


FIGURE 3 – Assessment in the reception test
using conventional DSB receivers for an SSB system

A: speech
B: light music

* See Table I of Recommendation 562.

At a carrier reduction of 6 dB, the average impairment of reproduced SSB signals with reference to the DSB signals was slightly higher than grade 4 [CCIR, 1982-86a].

In this case, intermodulation distortion is the prevailing cause. The amount of distortion reached with full modulation is quite large. In practice, with average modulation factors 35 to 45%, this distortion will be smaller.

2.7 Acceptable intermodulation level in SSB transmitters

The value for the acceptable intermodulation attenuation of D_N will largely depend on the required quality of the transmission with respect to non-linear distortion. Theoretical considerations based on a mathematical model for the calculation of adjacent-channel interference in SSB systems [Gröschel, 1978] show that with respect to adjacent-channel interference the intermodulation attenuation D_N of SSB broadcasting transmitters need not be better than 20 to 25 dB. In the future SSB system the acceptable intermodulation level therefore must be determined according to the wanted quality of the received signal. A range of $D_N = 26$ dB ($\approx 6.5\%$ harmonic distortion) to $D_N = 35$ dB ($\approx 2.3\%$ harmonic distortion) seems to be suitable for future SSB sound-broadcasting transmitters depending on the type of the prevailing programme (speech, music).

2.8 Demodulation

AM-based signals can be detected by two different methods; with conventional envelope detection or by mixing the input signal with an internally generated injection carrier. Envelope detection uses peak rectification which accurately follows the envelope of the modulated waveform.

In the mixing method the frequency of the internally inserted carrier must be equal to that of the carrier of the incoming SSB or DSB signal in which case the RF sideband information will be transferred into the audio range. Demodulation by mixing is normally referred to as product demodulation in a case where the frequency of the inserted carrier is nearly the same as that of the incoming carrier. In the case of these carriers being phase locked to each other, demodulation is referred to as synchronous or coherent demodulation.

DSB (A3E) or DSB with changing carrier level can be detected by envelope or synchronous demodulation.

Product demodulation is possible only with a SSB receiver, where the other sideband and carrier are cancelled or suppressed by filtering before demodulation.

SSB (H3E, R3E) signals can be demodulated by product or synchronous demodulation. SSB H3E (-6 dB) allows for envelope detection. The audio quality achieved is, however, marginal and therefore this method should be considered only during the transition period with old receivers.

Synchronous demodulation has the following advantages compared with product demodulation:

- locking function leads to less stringent requirements of the frequency stability in receiver oscillators;
- proper tuning is more easily accomplished by unskilled operators;
- allows both A3E and SSB detection;
- less sensitive to marginally adjusted transmitters (AM-PM conversion).

Because of the above advantages, synchronous demodulation should be recommended for new receiver designs as a standard demodulator and the system parameters should be based on this concept.

2.9 The influence of selective fading on demodulation systems

Broadcasting in the HF bands is based almost exclusively on sky-wave coverage. Because of selective fading, the reception of sky-wave signals with double sideband (DSB) modulation and envelope (linear) detection at the receiver, suffers severely from non-linear distortion. This degradation may largely be avoided when synchronous or product demodulation (see also 2.1 and 2.8) is used.

Test transmissions in band 6 (MF) using sinusoidal modulation (800 Hz) have been analyzed to determine the non-linear distortion occurring with envelope detection (A3E) and with synchronous detection (R3E).

The results are shown in Fig. 4. They show clearly that the non-linear distortion caused by effects of ionospheric propagation, such as selective fading, will be greatly reduced by the use of product demodulation in the receiver. The same improvement would be possible, even with class A3E emissions, if synchronous demodulation were used. No improvement will be achieved in compatible single-sideband (CSSB) emissions [CCIR, 1978-82b].

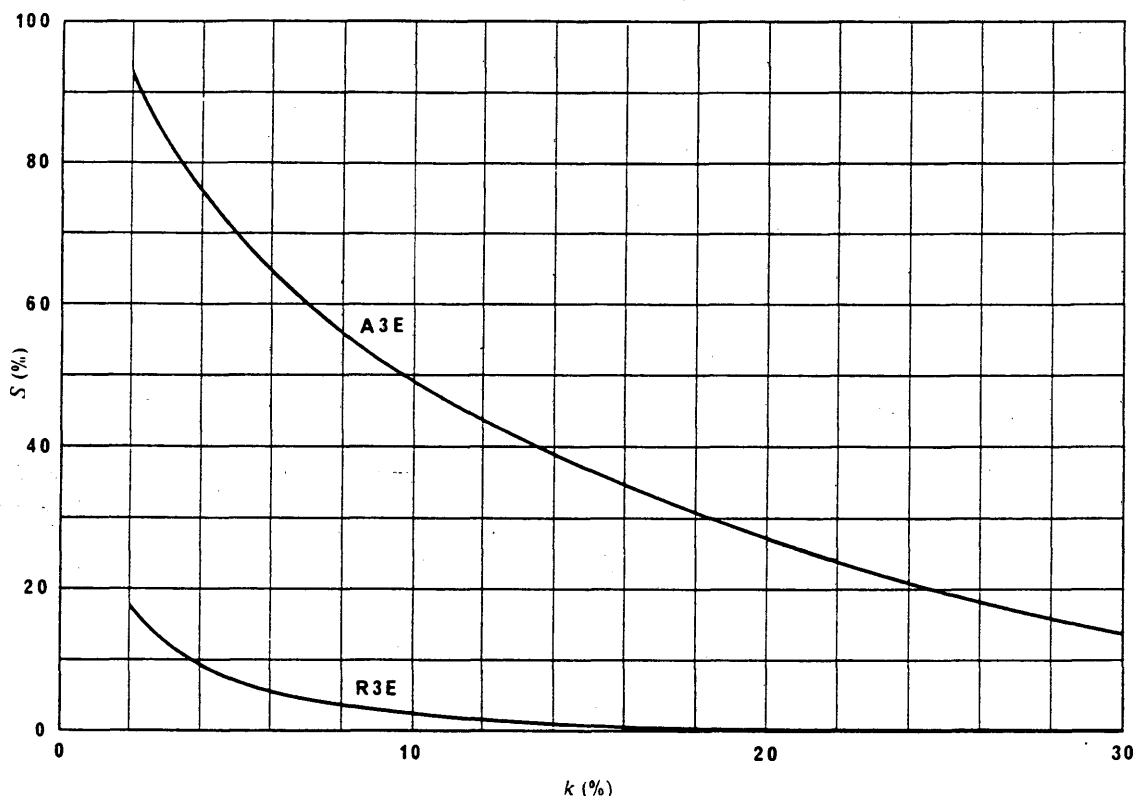


FIGURE 4 - Cumulative probability (S) of the non-linear distortion (k) averaged over the whole measuring period. Comparison between A3E (envelope) and R3E (synchronous) detection

2.10 The precision of the locally re-inserted carrier in SSB receiver

In single-sideband reduced carrier systems, the precision of the locally re-inserted carrier is important for the reception quality. Comprehensive subjective listening tests have shown [Thiessen, 1973] that the effects of non-linear distortion and inaccuracy of the re-inserted carrier superimpose and can be described by:

$$\Delta Q = (k_2/10\%)^2 + (k_3/6\%)^2 + (\Delta f/12 \text{ Hz})^2 \quad (1)$$

where

ΔQ : impairment of quality on the basis of a 6-grade scale,

k_2, k_3 : distortion factors of the 2nd and 3rd harmonics,

Δf : frequency error, (Hz), of the re-inserted carrier.

A maximum value $\Delta Q_{max} = 0.25$ appeared to be just tolerable. Values of $k_2 = 2.9\%$, $k_3 = 1.7\%$ and $\Delta f = 3.5$ Hz would be jointly admissible if the disturbing effects are assumed to be evenly distributed.

2.11 Further factors to be taken into account

2.11.1 The suppression of the same sideband (upper or lower) would have to be adopted within each broadcasting band. The present state of research indicates that, in the technology of intermediate-frequency and audio-frequency filters, suppression of the lower sideband is preferable. It is therefore suggested that for broadcasting the upper sideband should contain the full audio-frequency modulation.

2.11.2 The degree of carrier reduction should not exceed 12 dB. This was confirmed from the performance point of view of an SSB receiver by a reception field test carried out in Japan over a distance of 2000 km from the transmitting site, using experimental SSB receivers developed by incorporating synchronous detectors in conventional commercially available HF receivers.

Figure 5 shows the subjective assessment of the quality of the reproduced sound with the experimental SSB receivers using the seven-grade impairment scale of Recommendation 562, with reference to reception by DSB receivers with envelope detection.

The reception quality of SSB signals was worse than that of the DSB signal in cases where carrier reduction exceeded 12 dB because much distortion was caused within the SSB receiver due to difficulty in proper carrier extraction.

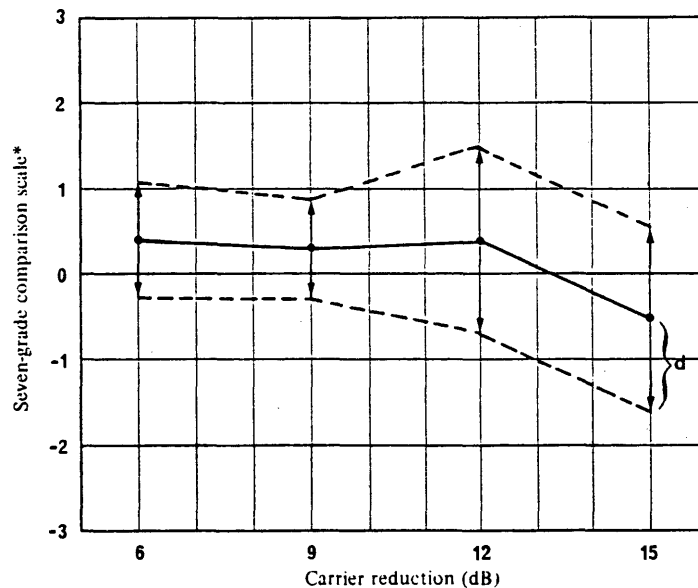


FIGURE 5 – Assessment in the reception test using the experimental SSB broadcast receiver
d: standard deviation

* See Table II of Recommendation 562.

2.11.3 The audio-bandwidth transmitted should be related to the carrier spacing. (The precise bandwidth cannot be given, the fraction of the channel spacing might well approach unity.)

3. Transition period

1987	31.12.1990	2000	31.12.2015
Start of transition period	New transmitters should be capable of SSB	Latest review of the final date for introducing SSB	End of transition, SSB with 12 dB carrier reduction
SSB: 6 dB carrier reduction	Low cost receiver should be equipped with synchronous demodulation		
	Review of the final date of cessation of DSB		

FIGURE 6 – Organization of the transition period from DSB to SSB

During the transition period, which is shown in Figure 6, SSB transmissions will be mainly received by conventional DSB receivers using envelope detection. In order to avoid undue deterioration of reception quality, it will be necessary to operate SSB transmitters with a carrier reduction not exceeding 6 dB. This, on the other hand, will have some consequences on the necessary output power of SSB transmitters if the same loudness level is required. To obtain with a conventional DSB receiver using envelope detection the same loudness level with both SSB and DSB, the sideband power of the SSB emission has to be 3 dB larger than the total sideband power of the DSB emission. Alternatively, if the sideband power of the SSB emission cannot be increased, one has to accept some reduction of the coverage area. One way to reduce this difficulty is to group together the SSB transmissions.

The possibility of conversion of DSB transmitters to the SSB mode is described in § 5.3.

3.1 Evaluation of compatibility aspects of a proposed SSB system

One of the major aspects to the implementation of SSB broadcasting in the HF bands is the fact that suitable SSB receivers will not yet be at the disposal of most listeners. Therefore, during the transition period, SSB transmissions should be receivable with adequate quality on conventional receivers using envelope detection. Among other parameters, e.g. audio-frequency bandwidth and compression, the dominant parameters controlling the compatibility are the amount of carrier reduction and the sideband powers. In addition to the transmission characteristics, the reception quality depends on the type of receiver and its tuning, as well as on propagation conditions such as, for example, type and intensity of fading.

3.2 SSB receivers

The WARC HFBC-87 decided that the immediate introduction of SSB is to be encouraged and also decided the procedure of transition from DSB to SSB (Resolution No. 517 HFBC-87). This transition has the following consequences with respect to receiver development.

- Conventional receivers employing only envelope detection should no longer be produced.
- New receivers to be used during the transition period, and afterwards, should be equipped with a synchronous demodulator, and use for carrier acquisition, a method whereby the carrier is regenerated by means of a suitable control loop which phase locks the receiver to the incoming carrier. Such receivers are expected to work equally well with conventional DSB transmissions and with SSB transmissions having a carrier reduced to 6 or 12 dB relative to peak envelope power.

There are essentially three reasons for recommending such synchronous demodulators:

- They permit the introduction of spectrum-saving SSB techniques without impairment of tuning ease.

- They permit the reception of both DSB transmissions and SSB transmissions having carrier reductions of 6 dB and 12 dB, respectively.
- They generate less distortion in the case of selective fading, which occurs frequently in present-day receivers with envelope detection.

In order to attain the best reception quality achievable with the new SSB system, the new SSB receivers should have an AF bandwidth of about 4 kHz and a much improved selectivity, with a slope of attenuation of the order of 35 dB/kHz. They can be readily manufactured using modern ceramic IF filters. It is also possible to have somewhat less selectivity, but then only at the expense of audio bandwidth.

3.3 *RF protection ratios in the transition period from DSB to SSB*

In the following discussion, it is assumed that the transmitter passband has a width of 4.5 kHz (see Annex I to Recommendation 639), that its passband is limited with a slope of attenuation of 40 dB/kHz, that its modulation signal is subject to a high degree of compression (see Recommendation 560) and that in the case of an SSB emission the carrier reduction is 6 dB, and the upper sideband is used.

The high audio compression used would result in a short-term r.m.s. modulation factor of about 50%.

3.3.1 *RF co-channel protection ratios in the transition period*

Due to the need to increase the radiated sideband power by 3 dB in the case of equivalent SSB emissions, there is a consequent need to make an allowance of the same 3 dB in the co-channel protection ratio for the case of a wanted DSB signal interfered with by an SSB signal, if the same quality of reception is to be maintained. Practical tests carried out in Japan have confirmed that an increase is necessary. One way to reduce this difficulty is to group together the SSB transmissions.

3.3.2 *Relative RF protection ratios in the transition period*

- It is first assumed that a wanted DSB signal is received by a conventional DSB receiver with envelope detection which is interfered with by an SSB emission with equivalent sideband power.

The resultant RF protection ratio of the wanted DSB signal in the lower adjacent channel at for example $\Delta f = -5$ kHz would be impaired by about 1 dB, while under the same conditions reception of the wanted DSB signal in the upper adjacent channel at $\Delta f = +5$ kHz would be impaired by about 4 dB in comparison to the present RF protection ratios, as specified in Recommendation 560.

If the tuning of the receiver is offset slightly, however, equal impairments of about 2.5 dB on both sides can be obtained.

- For the case of a wanted SSB signal interfered with by a DSB signal, the results obtained practically coincide with the values specified in Recommendation 560 (curve D). In this case reception of the wanted SSB signal will not be degraded during the transition period.
- In the case of a wanted SSB signal interfered with by an SSB signal, the results are the same as for a wanted DSB signal interfered with by an SSB signal, because in both cases the receiver is the same and the sideband powers are equivalent. In this case, therefore, the same impairment will occur as with a DSB signal being interfered with by an SSB signal.

The results described above show that during any transition period the protection ratios at $\Delta f = \pm 5$ kHz are impaired by about 2.5 dB in the case where SSB interferes with either SSB or DSB. When SSB is interfered with by DSB, it will need the same protection ratios as are valid now for DSB interfered with by DSB.

It can be expected that new SSB receivers which are suitable also for the reception of DSB signals will become available during the transition period, and that reception of DSB signals will be improved by their advantages (such as a reduction in harmonic distortion due to selective fading and better selectivity with enlarged AF bandwidth).

3.4 *RF protection ratios after the transition period for SSB operation with reduced carrier*

According to the proposed technical parameters for a future SSB system, a carrier reduction of 12 dB is assumed while all other assumptions made in § 3.3 for the transmitters are maintained. Both wanted and interfering signals are SSB signals.

Figure 7 shows a set of RF protection ratios calculated with the above assumptions for different SSB receivers which have different bandwidths and slopes of attenuation.

In Fig. 7, the relative RF protection ratios, A_{rel} , are given with respect to the frequency difference Δf between the wanted carrier f_w and the interfering carrier f_i :

$$\Delta f = f_w - f_i$$

Thus, negative Δf describes interference from the upper adjacent channel.

Curve ① is valid for an SSB receiver whose frequency response corresponds to that of the EBU reference receiver ($B_R = 4$ kHz) and the now obligatory tuning offset leads to an effective receiver AF passband of 4 kHz. Reception in the lower adjacent channel at $\Delta f = -5$ kHz is impaired by about 7 dB while the improvement in the upper adjacent channel would be almost 17 dB.

If a receiver with the skirt selectivity of the EBU reference receiver is to be used after the transition period, this receiver should not lead to any reduction in the relative RF protection ratio in the lower adjacent channel beyond the value now reached, according to CCIR Recommendation 560, and its AF bandwidth could be increased to 3 kHz (Fig. 7, curve ②), which would mean 50% more AF bandwidth than at present (2 kHz).

It would, however, be much more desirable to take advantage of the necessary introduction of SSB receivers and improve their selectivity. A considerable improvement of the relative protection ratio could be reached by using a ceramic IF filter with a skirt selectivity of about 35 dB/kHz. The corresponding protection ratios for receiver passbands of 4.4 kHz, 3.7 kHz and 3.0 kHz are given in Fig. 7 by the curves ③, ④ and ⑤, respectively. With 4.4 kHz bandwidth, protection would be improved by about 7 dB in the lower adjacent channel and by more than 40 dB in the upper adjacent channel. As can be seen, even a receiver bandwidth of 3.8 kHz would be achievable with at least 27 dB protection ratios in the lower and the upper adjacent channels. This example clearly shows the great advantage of an improvement of the skirt selectivity of the receiver.

3.5 Conclusions

Table III summarizes the effects of the introduction of SSB emissions with 6 dB carrier reduction during the transition period and with 12 dB carrier reduction in the final situation. These effects relate to the use of receivers with the EBU reference receiver frequency response during the transition period and an SSB receiver equipped with a ceramic or mechanical IF filter having a bandwidth of about 4 kHz and a slope of attenuation of 35 dB/kHz.

If an existing DSB transmitter has been converted to SSB operation, the achievable sideband power will be lower than the equivalent sideband power. This will result in an AF signal-to-interference ratio about 3 dB lower in comparison to DSB.

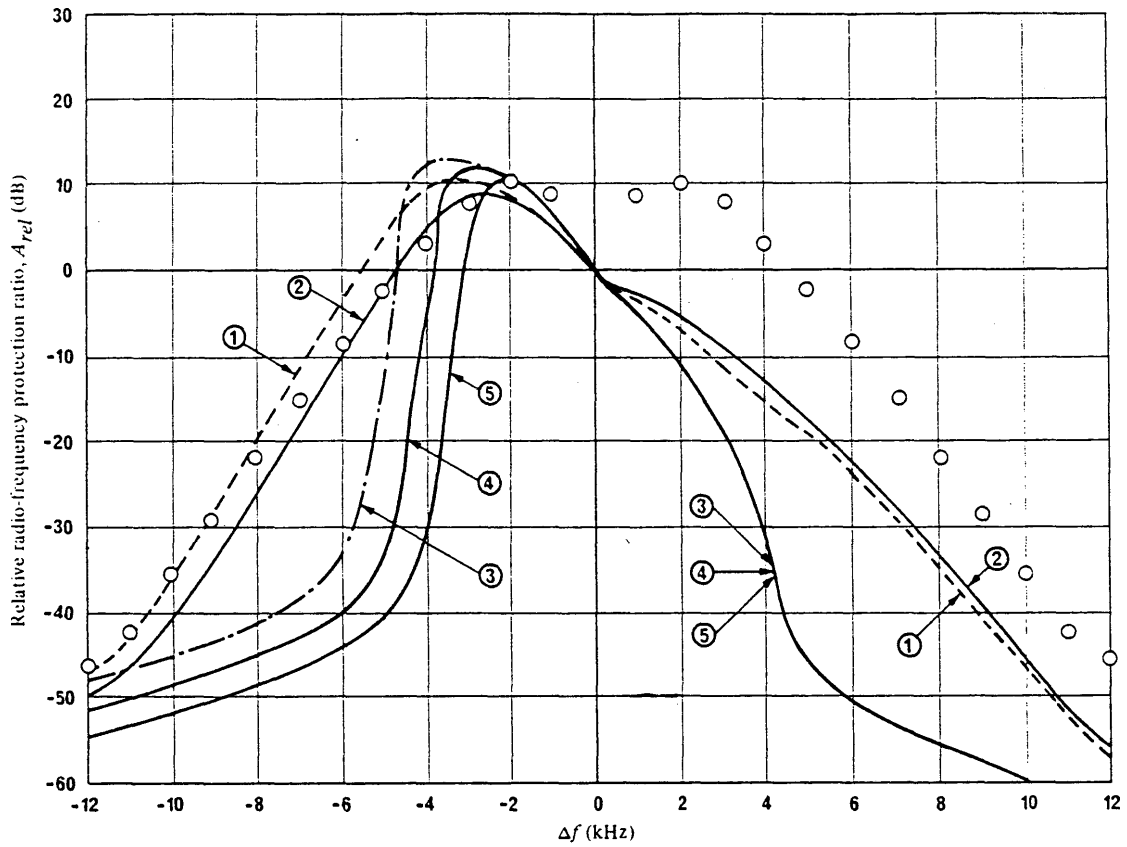


FIGURE 7 – RF protection ratios after the transition period

Wanted signal: SSB; interfering signal: SSB

Receiver:

- ① EBU $B_R = 4$ kHz; slope 8 dB/kHz
 - ② EBU* $B_R = 3$ kHz; slope 8 dB/kHz
 - ③ ceramic filter $B_R = 4.4$ kHz; slope 35 dB/kHz
 - ④ ceramic filter $B_R = 3.7$ kHz; slope 35 dB/kHz
 - ⑤ ceramic filter $B_R = 3.0$ kHz; slope 35 dB/kHz
 - ○ ○ values of CCIR Recommendation 560, curve D
- $\Delta f = f_w - f_i$ B_R : receiver RF bandwidth
 f_i always coincides with $\Delta f = 0$

* Offset by 1 kHz.

TABLE III — *Summary of the effects of the introduction of SSB transmissions with 6 dB carrier reduction during the transition period, and with 12 dB carrier reduction afterwards*

	Transition system		Final system
Transmitter carrier reduction	6 dB		12 dB
Total transmitter power	With constant carrier: more power With floating carrier: less power (comparable to final system)		Less power
Transmitter capital cost	More than for conventional DSB		Less than DSB
Transmitter operating cost	More than for conventional DSB		Less than DSB
Transmitted bandwidth	DSB: ± 4.5 kHz/SSB: +4.5 kHz		SSB: +4.5 kHz
Receiver selectivity characteristics	DSB: EBU	SSB: ceramic filter	SSB: ceramic filter
Receiver AF bandwidth	2 kHz	3.7-4 kHz	3.7-4 kHz
Receiver RF bandwidth	4 kHz	3.7-4 kHz	3.7-4 kHz
Slope of IF filter	8 dB/kHz	35 dB/kHz	35 dB/kHz
Receiver costs	No change	Initially more expensive	Less expensive than during transition
Channel spacing	10 kHz		5 kHz
Spectrum usage	No saving, slightly less efficient		Up to twice the number of emissions
Relative adjacent-channel protection ratio	Worse by 2.5 dB	Better	Better approx. 25 dB
Quality	Slightly worse	Better	Better

4. Relative values of RF protection ratio

The WARC HFBC-87 considered using the following relative values of RF protection ratio.

TABLE IV

Relative RF protection ratio values with reference to the co-channel RF protection ratio for DSB wanted and unwanted signals (dB)¹ for use in the HF bands allocated exclusively to the broadcasting service

	Wanted signal	Unwanted signal	Carrier frequency separation $f_{\text{unwanted}} - f_{\text{wanted}}, \Delta f(\text{kHz})$								
			-20	-15	-10	-5	0	+5	+10	+15	+20
1	DSB	SSB (6 dB carrier reduction relative to P_p)	-51	-46	-32	+1	3	-2	-32	-46	-51
2	SSB (6 dB carrier reduction relative to P_p)	DSB	-54	-49	-35	-3	0	-3	-35	-49	-54
3	SSB (6 dB carrier reduction relative to P_p)	SSB (6 dB carrier reduction relative to P_p)	-51	-46	-32	+1	0	-2	-32	-46	-51
4	SSB (12 dB carrier reduction relative to P_p)	SSB (12 dB carrier reduction relative to P_p)	-57	-57	-57	-45	0	-20	-47	-52	-57

¹ Frequency separations Δf less than -20 kHz, as well as Δf greater than 20 kHz, need not be considered.

The values of relative RF protection ratio given in Table IV should be used whenever SSB emissions in conformity with the specification in Appendix 45 to the Radio Regulations are involved in the use of the HF bands allocated exclusively to the broadcasting service.

The values given refer to the case of co-channel DSB wanted and unwanted signals for the same reception quality.

For the reception of DSB and SSB (6 dB carrier reduction relative to peak envelope power) wanted signals, a conventional DSB receiver with envelope detection designed for a channel spacing of 10 kHz is assumed.

For the reception of a SSB wanted signal (12 dB carrier reduction relative to peak envelope power), the reference receiver as specified in Appendix 45, Part B, section 3, to the Radio Regulations is assumed.

SSB signals with 6 dB carrier reduction relative to peak envelope power assume equivalent sideband power as specified in Appendix 45, Part B, section 1.2, to the Radio Regulations.

The figures for case 2 in Table IV relate to a situation where the centre frequency of the intermediate frequency pass-band of the DSB receiver is tuned to the carrier frequency of the wanted SSB signal. If this is not the case, the value for a difference of +5 kHz may increase to -1 dB."

5. Transmitter technology

5.1 DSB transmitters

Modern high power double-sideband transmitters are still using a valve in the final amplifier stage for generating the high radio-frequency power. Amplitude modulation is generated using a digital modulator which modulates the plate voltage by means of pulse width of pulse step modulation. This DC-coupled method enables the use of a dynamic carrier level operation or generating SSB-signals using a class C radio-frequency power amplifier.

Typical transmitter efficiencies for modern DSB transmitters in the range of 100 - 500 kW are between 65 and 75%, with a modulation depth of 100%. Decreasing modulation depth decreases these values by only a few per cent.

5.2 SSB transmitters

In the modern transmitters SSB can be generated by either of the two following methods:

- a) low level SSB generation followed by linear class AB or B power amplification;
- b) low level SSB generation followed by separation of the amplitude information and the phase information. The amplitude information is amplitude modulated by an anode modulator as described in § 5.1. The frequency source for the transmitter is phase modulated by the derived phase information. The composite signal, available at the transmitter output, resembles the SSB signal. This method is known as Kahn method.

The first method yields a total transmitter efficiency around 40% with full P_p , i.e. 100% modulation, by using additional energy saving circuitry. With 100% modulation, the second method reaches 65-75% efficiency which is similar to the DSB value. With a 40% average modulation level and -12 dB carrier reduction the total efficiency for the Kahn method decreases by 10%, resulting in a 55-65% efficiency. This reduction in efficiency results from the fact that the mean RF output power is reduced while the input power requirements of the transmitter (cooling, heaters, etc.) remain the same.

The maximum P_p , achievable for SSB is less than the maximum P_p of DSB for transmitters operatable for both DSB and SSB modes. For the method of linear amplification the P_p reaches 0.25 times the DSB value. The P_p generated using the Kahn method reaches 0.5 - 0.75 times the DSB value.

Keeping P_p constant, the change from -6 dB to -12 dB in carrier level can be utilized for the increase of wanted sideband power up to 3.5 dB.

5.3 Possibilities for conversion of existing DSB transmitters to SSB

The majority of existing DSB transmitters apply anode modulation using a modulation transformer and a modulation choke. The anode modulator is AC-coupled to the final radio-frequency stage and therefore only the linear amplification type of SSB generation is possible. The accompanying modifications are significant. Resulting SSB P_p cannot exceed original DSB carrier power.

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CCIR Documents

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- [1982-86]: a. 10/7 (Japan).

REPORT 1061-1*

**TRANSMISSION OF SUPPLEMENTARY INFORMATION
IN AMPLITUDE MODULATION SOUND BROADCASTING**

(Study Programme 44J/10)

(1986-1990)

By multiplexing supplementary information with amplitude modulation sound broadcasting, new broadcasting services can be realized. Some examples are radio data broadcasting, emergency alert broadcasting and programme identification code broadcasting.

1. AM data broadcasting system (AMDS)**1.1 AMDS in the United Kingdom**

Trials have been carried out in the United Kingdom of a system for the transmission of supplementary information in amplitude modulation sound broadcasting [CCIR, 1978-82a; 1982-86a; Wright, 1984].

A data signal has been applied to the British Broadcasting Corporation Radio 4 (United Kingdom) transmitters operating on 200 kHz (from 1 February 1988: 198 kHz). Phase-modulation of the main carrier is used to carry the data signals and the peak phase deviation is 22.5° either side of the normal unmodulated carrier phase. The information rate is 25 data bit/s and each data bit is biphase-encoded so that the minimum time between transitions in the biphase-encoded waveform is 20 ms. The input to the phase-modulator is low-pass filtered and this, combined with similar filtering in the receiver, provides a full cosine roll-off characteristic.

The British Broadcasting Corporation 200 kHz (from 1 February 1988: 198 kHz) network consists of a three-transmitter synchronized group with mush areas, i.e. zones of mutual interference, in the regions of near-equal field strength between the transmitters. During the development of the experimental system, very careful attention was paid to preserve compatibility (i.e. no interference by the data signal) with reception of the programme signal [Lyner, 1982] and, in the special case of mush areas, the data signal timings were adjusted so that identical data signals from the different transmitters arrive in the mush area within ± 1 ms of each other.

The data are formatted into blocks of 50 bits. Each block contains 32 useful message bits and a 13-bit cyclic redundancy check word as well as addressing and synchronization for up to 16 independent transparent data channels.

Currently two of the data channels are assigned for use by the United Kingdom Electricity Supply Industry for load management purposes by means of radio-controlled time switches. Experimental work is in progress using two further channels for a coded weather information service for mariners around the British Isles. Each application will occupy a maximum of about one sixth of the channel capacity, i.e. 5 frames/min, which is equivalent to about 3 bit/s.

The data signals have been radiated continuously from all three transmitters since 1 October 1981 and there have been no listener complaints. The data signals themselves have been successfully received at ranges well in excess of 400 km, including all parts of the United Kingdom and some sites in France, using simple experimental receiver/decoders. These receivers operate on field-strengths down to 1 mV/m. They use an internal ferrite-rod antenna and have often been mounted in positions where the radio signal was subject to considerable local attenuation or electrical interference.

* This Report should be brought to the attention of the IEC.



The system became a fully operational service on 1 April 1985 for data messages originated and received by the United Kingdom Electrical Supply Industry. The data source equipment is able to accept messages from other potential users but these remaining channels have yet to be exploited.

It should be noted that although the specification relates to long-wave transmissions, the system is suitable also for general use in amplitude-modulation medium-wave sound broadcasting.

1.2 AMDS in the German Democratic Republic

[CCIR, 1982-86b] first presented the results of tests carried out at LF only on the AM data broadcasting system (AMDS) developed in the German Democratic Republic and shown in Figure 1. [CCIR, 1986-90a] gives summary results of additional laboratory tests and field trials carried out in the German Democratic Republic on this system at LF and MF bands.

After ten years theoretical and experimental research, a 200 bit/s AMDS system concept is sufficiently proven for conventional monophonic AM broadcasting [CCIR, 1986-90b].

More recently further tests under severe LF and MF propagation conditions (influence of sky-to-ground wave interference, effects of interferences near the boundary of the service area etc.) have been carried out. The results of field tests carried out on this system as well as on other systems in the Federal Republic of Germany at LF and MF are reported in § 1.3 [Augustin and Kühne, 1986 and 1988], [Raufmann and Schramm, 1988], [Schramm, 1989] and [Augustin *et al.*, 1989].

The results of these long-term tests are as follows:

- Data system operation is satisfactory within the whole AM sound broadcasting coverage area under normally defined conditions and beyond the boundaries of this area under the following conditions:
 - 1) no interference should be caused by co- or adjacent-channel transmitters;
 - 2) in the case of synchronized operation particularly in the mush-area time-delay correction for both the data and sound signals should be taken into account [Augustin and Schulze, 1973].
- With a phase deviation of $\pm 15^\circ$ and a simple second order RC filter with a cut-off frequency of about 100 Hz to form the pulse shapes of the digital signal, optimum spectrum shaping can be achieved.
- Data transmission at lower bit rates allows the use of higher values of the phase deviation (see ——— Recommendation 706).
- For full compatibility between the sound and data signals the following conditions are to be met:
 - optimal transmitter tuning to minimize sound signal unwanted synchronous phase modulation;
 - amplitude modulation factor for the sound signal should be limited to $m \leq 0.98$ to avoid at any rate overmodulation and to provide a sufficient residual carrier for the digital signal recovery in the decoder;

- receivers should be stable to avoid unwanted phase modulation caused by mechanical shocks or vibration or by electronic noise or peaks on the supply voltages for the tuning elements;
 - "allowed" detuning of the receiver should be less than the 3 dB passband of the IF band filters.
- The GDR transmission system AMDS at 200 bit/s (so-called "AMDS-200") fulfils all DSB requirements proposed in _____ Recommendation 706 for common AM broadcasting sound transmission systems.
- This system can also be used for transmitting time information as dealt with within Study Programme 4B/7 "Dissemination of time signals by addition of phase modulation on amplitude-modulated sound broadcasting transmitters", Question 4/7 or 7/7, Report 577 _____ and Report 578 _____ concerning the problem "time codes" (see also Recommendation 583).
- "AMDS-200" can be regarded as a system for the distribution of all kinds of digital information (RZ, NRZ and bi-phase codes) for example:
- facilities for automatic or assisted tuning functions related to the receiver user;
 - traffic information (like TMC) for the motorists, inaudible to the normal broadcasting listeners;
 - radiopaging;
 - etc.
- If overall reception quality of the sound programme is rated 3 or better on SINPO code a very good trade-off is established between acceptable quality, stability and reliability of data and sound broadcasting signals. The most important causes of bit errors are:
- 1) deep fading effects (disrupting reception of sound and data signal) caused by long periodic ionospheric absorption fading effects, by long periodic co-channel interference and by ground and skywave interference where the RF protection ratios will no longer be maintained;
 - 2) interference effects caused by strong adjacent or co-channel operated transmitters whose sidebands are partially or fully overlapping the useful bandwidth of the data channel.

However, in many cases, even when the sound signal becomes useless due to strong adjacent channel transmitters, the data signal may still be received in good quality.

- A bit error rate (BER) $< 10^{-3}$ (or $< 10^{-4}$ in particular conditions) can be achieved within the whole service area (and even outside) for 50% of the time or a BER $< 10^{-2}$ (or $< 10^{-3}$ in particular conditions) for 90% of the time without particular error protection strategies, provided that the wanted to unwanted RF-signal ratio is:
 - ≥ 8 dB (protection ratio value used for planning) for correlated sound and data information signals of the co-channel station, for example: in synchronized broadcasting networks or when ground or skywave reception exists, but
 - ≥ 30 dB (protection ratio value used for planning) for uncorrelated sound and/or data transmission signals of independent radio stations.

On the basis of the successful results of the laboratory and field tests, "AMDS-200" is suitable for a modified, low bit-rate radio data system (RDS) as applicable to AM sound broadcasting.

The system has been under trial for several years using high power broadcasting transmitters of the Deutsche Post in the German Democratic Republic and has met fully the expected performance with real data signals.

The system operates in a stable manner and is relatively uncritical with regard to tuning conditions and phase modulation parameters, and no interference to the received broadcast programme has been experienced.

For transmitters which are used as high stability frequency and phase references including phase controlling systems, bi-phase encoding should be used.

For digital signal recovery, suitable fixed and/or mobile AM sound broadcasting receivers or commercial measuring or multiband receivers were used.

1.3 Field tests with AMDS in the Federal Republic of Germany

[CCIR, 1986-90c] Extensive field tests have been carried out in the Federal Republic of Germany with supplementary data-transmission systems including, amongst others, the system referred to in § 1.2. These field tests were performed with a total measuring time of more than 600 hours and in various transmitting and propagation conditions including synchronized and DAM-transmitter operation as well as multipath propagation. On the basis of their results it appears possible to implement supplementary data-transmission systems in AM broadcasting networks with data rates of about 100-200 bit/s.

Comparative measurements with circuit developments of different institutions have demonstrated that a high technically compatible standard of the coder systems has been achieved. Special measurements have shown that the excess of a certain maximum modulation factor adjusted at the transmitter increases the number of errors in the data transmission considerably. This effect can be kept small by periodical and accurate adjustments of the transmitter.

Although the system reliability seems to be sufficient with the receiving equipment already available, it has become obvious that there is still room for substantial improvements especially in the field of receiver design in which up to now supplementary data transmission with carrier phase modulation has not had to be taken into account.

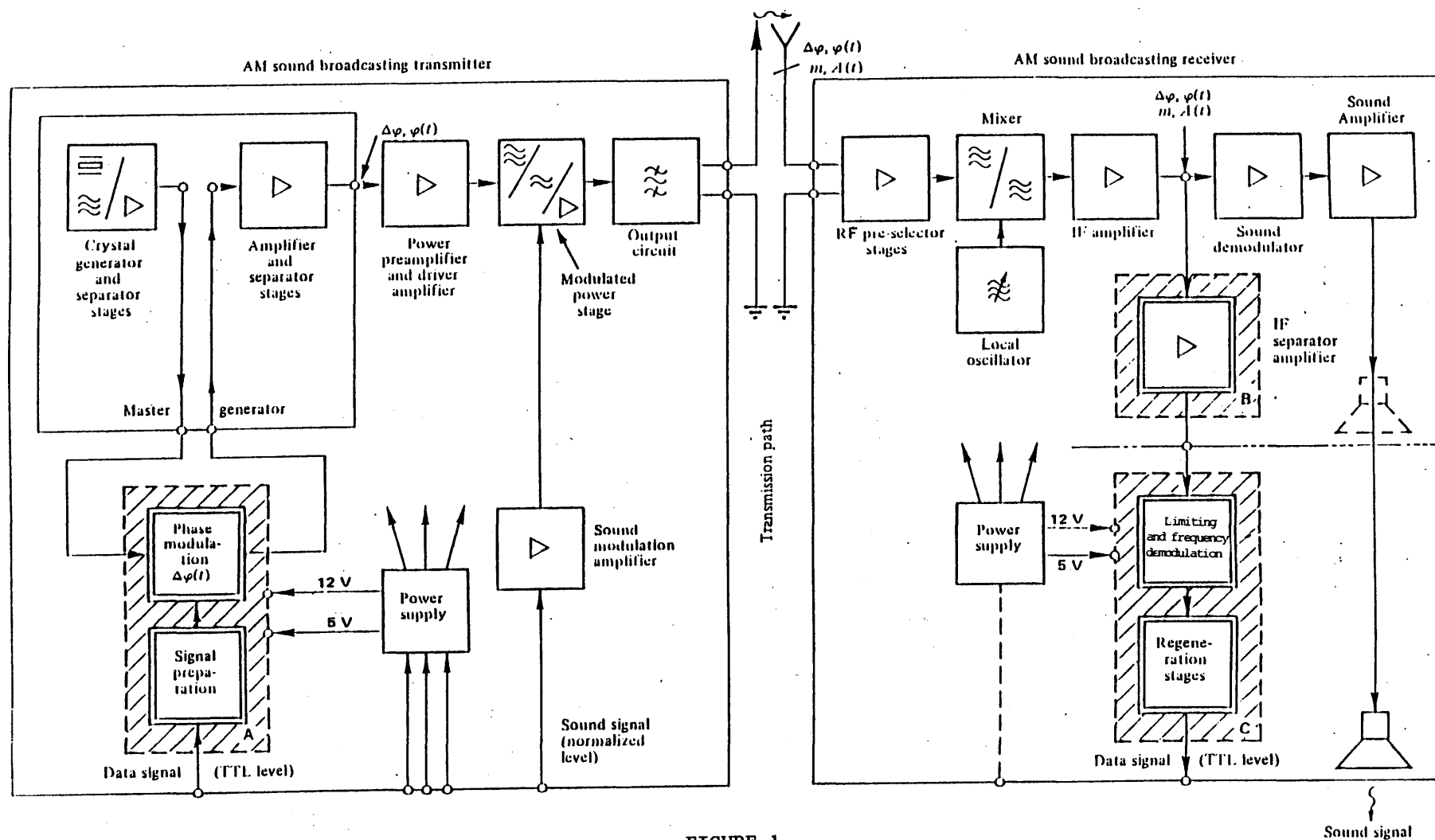


FIGURE 1

Basic circuit diagram of a monophonic AM sound broadcasting system (AMDS)
for supplementary data transmission

- A: Additional module (phase modulator)
 B: Additional module (output decoupling unit)
 C: Additional module (frequency demodulator and data signal regenerator)

1.4 AMDS in France

[CCIR, 1986-90d, e] In France, a system intended to broadcast subsidiary information associated with amplitude modulation sound broadcasting has been in experimental operation since 1986 from a transmitter located in Allouis, broadcasting the Radio-France "France Inter" program on a frequency of 162 kHz.

The system is multiservice; it incorporates two levels of data multiplexing:

- the first level multiplexes "time data" with "other data",
- the second level provides the temporal multiplexing of various data signals coming from independent sources as the set of "other data"; this multiplexing is done at the level of "network" layer.

The binary rate (time data + 16 useful bits per second) and the modulation parameters in use make the system directly compatible with the broadcasted sound and with the special requirements in terms of phase and frequency stability imposed on this transmitter, whose carrier is used as a frequency reference.

The modulation in use is frequency keying with continuous phase with a binary rate of 40 bit/s. It is a partial response modulation that can be considered as the juxtaposition of 100 ms symbols, transmitted every 25 ms. The phase of the carrier during each time slot $[kt, (k+1)T]$ depends on bit b_k and the 3 preceding bits b_{k-1} , b_{k-2} , b_{k-3} .

The information contents of the two services described above is organized into 1 minute frames each divided into 60 time slots.

The 60 one-second time slots making the "minute" field are quoted S_0 , $S_1, \dots, S_{59}$ (S_0 indicates the first second of the minute).

- Each second from S_0 to S_{57} contains time information and data;
- second S_{58} contains time information, a field synchronization word and a transmitter identifier word, with the same temporal organization as for seconds S_0 to S_{57} ;
- second S_{59} provides "minute" synchronization.

In each second, the data signal has a duration of 700 ms, the time signal a duration of 200 ms and there is a 100 ms interval between the data signal and the time signal, during which the carrier is not phase modulated. This interval is used for phase locking the transmitter.

The capacity that can be used for the "other data" during one minute is $58 \times 25 = 1450$ bits normally organized in 29 packets of 50 bits.

A packet is made of 5 address bits, 32 data bits, 13 check bits.

- The first five bits define the packet address code, (the most significant bit comes first); address 00000 is reserved to broadcast a stuffing packet when there is no useful data to transmit.
- The 32 following bits bear the useful information.
- The 13 last bits are reserved for error protection.

It can be seen that the system offers a transmitter identification service and allows multiplexing of up to 63 low rate data channels. This system has been developed for use on amplitude modulation sound transmitters in the kilometric band. It can also be used in the hectometric and decametric bands.

The system has been adopted as a French national standard [France NFC90-002].

2. Emergency alert broadcasting system

In Japan, the control signal for an emergency alert broadcasting system, that uses a continuous-phase frequency-shift keying (CP FSK) coded signal with signal frequencies in the middle of the audio frequency range, is recommended by the Radio Technical Council of Japan [CCIR, 1982-86c; Radio Technical Council, 1983].

For such an emergency alert broadcasting system, in which the content of the alert broadcast is very important and where the emergency message interrupts other programmes, a middle frequency audio signal is most suitable as the control signal because it has the best transmission characteristics and reliability. It can also be used as an alarm.

The signal frequencies of the CP FSK signal are 640 Hz and 1024 Hz and the bit rate is 64 bit/s. The time necessary to actuate the alert receiver is usually 2 s.

The types of signals are the "start" signal, the "end" signal and the "test" signal. The signal is composed of fixed type codes that identify the control signal, and variable type codes. The length of the fixed type code and variable type code are both 16 bits. As it is necessary for at least four fixed type codes (a total of 64 bits) to be received before the alert receiver is actuated, the probability of mis-triggering by misinterpreting noise or programme sound as the control signal, is negligible.

The system is equally applicable to other broadcasting media, such as FM sound broadcasting or television and the transmitting equipment can easily be added to existing equipment.

The signal can be used as an alarm that will effectively attract people's attention.

Field trials of the signal system were performed using all sound broadcasting and television media. These tests confirmed that in those areas where broadcast sound is received satisfactorily, the receiver would operate consistently. However, in those areas where deep, frequency-selective fading takes place at night-time because of interference between sky-wave and ground-wave signals, it was observed that some code words were missed and there was some delay in activation of certain types of receiver. When the signal was transmitted for about 15 s, the system operated satisfactorily even in these areas. Thus the signal system is highly reliable under various reception conditions.

3. Programme identification code broadcasting

When considering the use of a low frequency audio signal for the identification of daily programmes that can be transmitted several times per day, it is necessary that the signal does not impair the content of the programme.

Considering the narrow bandwidth available when a low frequency audio signal is employed, a method must be developed that has a high data transmission rate in order to control receivers without a long delay at the start of the programme.

NHK have been studying a system in which the code signal is a continuous-phase frequency-shift keying (CP FSK) signal, which has signal frequencies of 24 Hz and 32 Hz and a data transmission rate of 16 bit/s [CCIR, 1982-86c; Namba and Saito, 1985].

The signal in the experimental system is composed of four 12-bit code words and the receiver is activated when two of them are received, i.e. 1.5 s after the signal is sent. About 70 programmes can be identified from the pattern of the code word.

The signal is transmitted by either multiplexing in the lower frequency portion of the programme sound or by frequency modulating the main carrier. The former case is also applicable to FM sound broadcasting or television. The latter method of frequency or phase modulation of the main carrier may be found to be preferable for AM transmitters [Wright, 1984].

It has been demonstrated through laboratory experiments that the signal is inaudible under normal listening conditions, because the receiver can be activated by a modulation level of less than 5% or by a maximum frequency deviation of about 10 Hz of the main carrier. It is thus possible to obtain satisfactory identification of daily programmes.

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REPORT 1060-1

**ENERGY SAVING METHODS IN AMPLITUDE MODULATION BROADCASTING
AND THEIR INFLUENCE ON RECEPTION QUALITY**

(Question 44/10, Study Programmes 44A/10 and 44F/10)

(1986-1990)

1. Introduction

For high-power amplitude modulation transmitters used for LF, MF and HF broadcasting, reduction of power consumption has always been of major importance. Power consumption savings can be achieved by improving the efficiency of transmitters and/or by using other amplitude modulation techniques.

2. Improvement of transmitter efficiency

In the past, several transmitter systems have been developed with more efficient power amplification, e.g. the Doherty system, the Taylor principle (addition of third harmonics), etc. In modern transmitters, modulation amplifiers in switched mode have been introduced, e.g. pulse duration modulators (PDM) or pulse step modulators (PSM). Using such modulators, the overall efficiency of the transmitter has been improved by 10% or more compared with transmitters using conventional modulation techniques. However, such transmitters may radiate spurious signals caused by the switching frequencies and these spurious signals can cause interference in adjacent channels. Another disadvantage is that even small levels of over-modulation can lead to unacceptable amounts of distortion. However, both of these disadvantages can be reduced by appropriate control of the audio signal input and by the choice of a switching frequency that is a multiple of the channel spacing. Further measurements should be carried out to determine the full effects of radiated spurious signals from PDM or PSM transmitters on the adjacent-channel protection ratio, since this type of high-power transmitter will become increasingly available in the future.

3. Energy saving using other amplitude modulation techniques

It is already well established that the future use of SSB transmissions will achieve a considerable energy saving – in Table II of Report 1059 reductions in power consumption of up to 80% compared to equivalent DSB transmissions are indicated.

Two alternative amplitude modulation transmission systems employing carrier control proportional to the mean level of modulation have been proposed. Both can be introduced easily into modern transmitters. These are:

3.1 Dynamic amplitude modulation (DAM)

DAM was proposed in early 1939, and has been in operation successfully for several years in a number of European countries at MF and more recently in the Netherlands at HF. An overall reduction in power consumption of 50% for average programme material using moderate audio compression has been achieved. The level of energy saving depends upon the reduction of carrier power for zero modulation (**Fig. 1**). Additional dynamic compression occurs in the receiver due to the action of the automatic gain control (AGC).

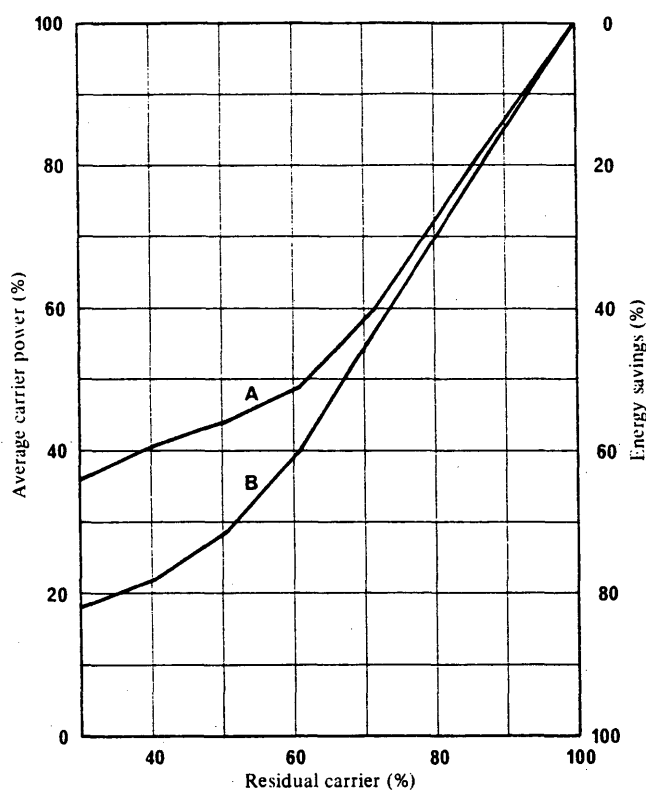


FIGURE 1 – Relationship between carrier power, energy savings and residual carrier (knee-shaped characteristic with offset)

Curves A: 9 dB compression gain

B: 0 dB compression gain

Rapid changes in the level of programme modulation can cause distortion. Moreover, programme pauses, during which the carrier level is reduced, result in an increase of noise and interference. However, these disadvantages can be reduced by the choice of an appropriate level of carrier control plus a limited reduction of the carrier during zero modulation (Fig. 2). Also the increase in noise and interference during programme pauses will be partially compensated by the increase in the mean programme level of the received DAM signal (i.e., depth of modulation). Based on theoretical investigations, a reduction of the wanted-to-interfering signal ratio of the order of 1 dB can be expected [CCIR, 1982-86]. The carrier regulation characteristic of some transmitters can be easily modified to take account of specific conditions of programme content and possible interference levels.

Further studies would be useful to determine the consequences on reception quality and effective coverage area for transmissions using DAM, and to determine any change in protection ratio for conventional amplitude modulation transmissions experiencing interference from a DAM transmission.

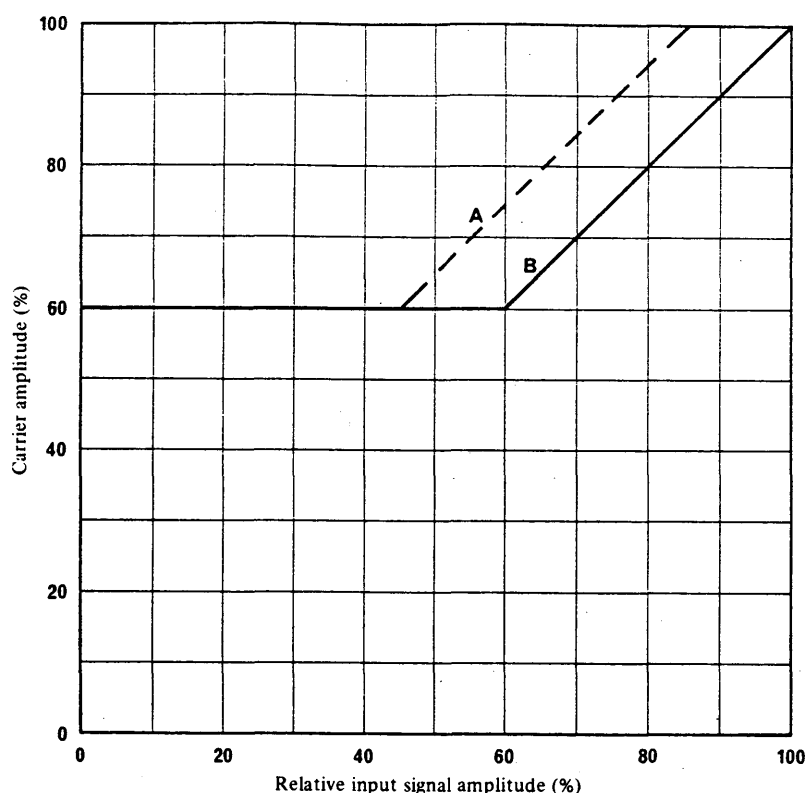


FIGURE 2 – Carrier control characteristics for a transmitter using DAM technique

A: knee-shaped characteristic with offset

B: knee-shaped characteristic

3.2 AM companding system*

This method has been developed in the UK, [Lawrence *et al.*, 1986 ; Williams, 1986] and differs from DAM in that the compression is increased approximately linearly from zero in the unmodulated condition to a maximum value at 100% modulation, whilst maintaining the appropriate depth of modulation. This is seen as having the following advantages:

- receiver AGC operation compensates for the applied compression ;
- the reduction of peak voltages and powers allow further savings due to redesign of the transmitter output; reduced peak demand may allow lower electricity charge rates ;
- maintenance of normal power levels during periods when the programme content is most susceptible to interference, i.e. under conditions of low modulation.

* The United States of America points out that the term "companding" is used extensively in North America in a different system.

The results of the various laboratory tests and field trials conducted in the UK have indicated that the application of Amplitude Modulation Companding (AMC) with a maximum compression of 3 dB causes negligible loss in reception quality, even in fringe areas. Tests involving interfering signals were conducted with the interfering programme the same as the wanted programme and the interfering transmitter synchronised with the wanted transmitter, and with the interfering programme different from the wanted programme. Co-channel and adjacent-channel protection ratios were not significantly affected by changing from conventional AM to AMC.

A number of UK MF transmitters have been using AMC since November 1987.

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REPORT 943-1*

**PROTECTION OF SOUND-BROADCASTING STATIONS
AGAINST ATMOSPHERIC ELECTRICITY**

(Question 48/10)

(1982-1986)

1. Introduction

The WARC-79 identified the requirement for the service Study Groups of the CCIR to give consideration to the needs of their services to effect protection of installations against atmospheric electricity, particularly lightning discharges. This Report, relating to sound broadcasting, complements information for the fixed and radio-relay services contained in Reports 861 and 932 respectively.

CCITT Study Group V is currently considering the protection of telecommunication lines and equipment against lightning discharges and the overall subject of lightning protection is under review by IEC Committee TC 81.

2. The lightning process

Lightning discharges occur when the electrostatic potential created by convective charge generation in thunderclouds exceeds the breakdown potential. Discharges may occur from cloud-to-cloud, cloud-to-air or cloud-to-ground. With the cloud-to-ground type, both the discharge itself and the electrostatic stress before and during a storm which can be as great as 5-10 kV/m [Surutka and Veličković, 1973], may cause damage.

As a charged cloud approaches a mast, it induces charges of opposite sign on the sections of the stays. When the cloud is discharged by a lightning stroke to the ground, the field collapses in a time interval of the order of a microsecond. Charges on the stays produce high voltage across the insulators. Voltages as great as 400 kV arise, leading to flashovers and breakdowns. It has been suggested [CCIR, 1982-86a] that arcs caused by static atmospheric electricity represent a greater risk of damage than direct or near lightning strokes.

* The Director, CCIR, is requested to bring this Report to the attention of the IEC.

3. Frequency of incidents

There are marked seasonal, diurnal and geographical variations in the incidence of thunderstorms. Storms tend to occur in the afternoon hours and are least frequent in the post-dawn period. They are most prevalent in the local summer at mid- and high-latitudes, but have little seasonal variation in the tropics. Their incidence is greatest over tropical land masses, but the proportions of cloud-to-cloud and cloud-to-air discharges are also higher in the tropics than at other latitudes.

Current evidence on reported incidents causing damage is scanty, but there is a marked dependence on geographical position and the precise locations of installations. In France, they are appreciably more frequent in the southern part of the country and in high mountainous areas, and in Japan, in the northern part of the country which is subject to heavy snowfalls.

4. Power dependence

Although atmospheric electricity can give rise to problems to domestic receivers, these are minor compared with those occurring at transmitting installations which generally use high masts. High-power installations (with transmitter powers in excess of 10 kW) tend to be more prone than lower-power installations, because the transmitter power is sufficient to maintain an arc discharge across components, once it is initiated, either by a lightning stroke or by wind-borne static. This situation becomes more serious as the power of the transmitters is raised.

5. Locations of flashover points and damage caused

Antennas for LF and MF use include base-fed or loop-fed steel masts which are either self-supporting or guyed with steel ropes divided by insulators into sections shorter than the resonant length. Lightning strikes can cause flashovers between the mast and the ground or on the guy insulators. Moreover, the surges thus created may result in equipment being damaged. Solid-state transmitter circuitry, antenna feed lines and antenna matching-unit capacitors, are particularly vulnerable.

Base insulators are generally ceramic pillars and provided these are shunted by a discharge gap (see § 6.1) damage is rare. For low transmitter powers, stay insulators are invariably of the simple slotted glazed-ceramic egg-type set between interlocking wire loops. Although flashovers occur during lightning storms, no obviously observable mechanical damage or electrical deterioration to antenna performance is noticed in the long term. However, with the sustained discharge arising with high-power transmitters, cracked stay insulators and arc-erosion of stay wires adjacent to insulators may occur.

6. Protection methods

6.1 *Traditional spark gaps*

Base-fed radiators may have discharge gaps permanently fitted to the structure. Generally, the earthed side is connected to the radial ground system through copper conductors. Discharge gaps have a variety of forms, ranging from spheres and cylinders with gaps of less than half the diameter, to rods and loop structures. Gap settings are generous, typically 8-16 mm, to prevent rain droplets from coalescing.

In Japan, judiciously placed spark gaps made of cylindrical hard-carbon electrodes are used. Gap shapes and sizes are optimized as a function of the dominant lightning waveform encountered [CCIR, 1978-82a; NHK, 1982].

6.2 *Non-linear resistances at the guying insulators*

Silicon carbide non-linear resistances of about 1 M Ω at the working RF voltage can be employed at the guy-insulator terminals. These release the static charges on the insulated parts of the gaps, but are often destroyed if lightning strikes the mast directly.

6.3 *Transmitter shut-off*

A lightning detector placed on one of the mast stays may be installed which, when it registers a lightning discharge, momentarily (for about 1 s) causes the transmitter drive to be interrupted. In this way any flashover on a stay insulator is not followed by a power arc; the flashover extinguishes immediately and no damage is caused. The device can be made sensitive to about 5 A which is much less than the current required to cause an arc.

Alternatively, the transmitter may be shut off by measuring the standing-wave ratio at the antenna terminals arising from the arc mismatch.

6.4 *Power reduction during storms*

Substantial reduction of transmitter power during storms for high power carriers can be a useful procedure to avoid self-sustained arcs.

6.5 *Fast EMP gas discharge tubes*

These can be fitted across the output terminals of solid-state transmitters to provide protection, but have only limited life and may themselves become damaged.

6.6 *Earthing arrangements*

It is important to pay particular attention to earthing arrangements.

In some station earthing systems the radial-wire ground systems in a directional array are bonded together and connected to the earth at the transmitter building via a copper tape of cross-section not less than 75 mm² in a trench alongside the buried coaxial feeders from the antenna. The tape is connected to transmitter earth so as to by-pass any joints or connections in the coaxial feeders and thus reduce lightning surges in their outer and inner conductors. This practice has resulted in no known instance of lightning damage to inaccessible buried feeders.

Stay ropes are earthed alongside their concrete anchor blocks using either copper tape, aluminium tape or multi-strand stainless-steel wire. A straight path down to earth is adopted in order to prevent lightning from smashing the concrete block. Local earthing plates or spikes are used, which are not always connected to the main radial earthwire systems.

6.7 *Protection of VHF antennas*

It is common practice with VHF antennas for the radiating elements to form a d.c. short circuit (e.g. folded dipoles) and to be bonded to the mast so that lightning surges are suppressed between the inner and outer sections of their coaxial feeder cables. Decoupling of VHF feeders from MF masts using inductor-wound methods [CCIR, 1982-86b] has hitherto prevented lightning damage.

6.8 *Use of plastic ropes*

Plastic fibre "Parafil" stay ropes for low-power stations can be used to overcome the problems encountered with conventional guys.

In Yugoslavia, "Philistran" stay wires have been used successfully with transmitter powers of up to 600 kW.

7. *Specific antenna designs*

7.1 *France [CCIR, 1978-82b]*

7.1.1 *The Allouis LF antenna (164 kHz)*

An LF antenna specially designed for protection against lightning has been developed in France. It consists of masts less than $\lambda/4$ high and the ring of guys at the top of each mast has no insulators; this means that lightning strikes are released directly into the ground and the guys form a protective cage for the other parts of the antenna. The radiation pattern is the same as for a conventional mast. The efficiency of this type of antenna has been proved during many years of operation.

7.2 *The Netherlands [CCIR, 1982-86c]*

7.2.1 *The Flevoland MF antenna (simultaneous 400 kW on 747 kHz and 400 kW on 1008 kHz)*

The $5\lambda/8$ antenna mast is divided in half and centre-fed. The lower part is grounded and the upper section is electrostatically grounded via a $\lambda/4$ stub in the mast (the insulated ladder). The guys are not divided into sections but are insulated at both ends. The lower insulators are bridged by a network which keeps the RF current in the guys as low as necessary in order not to deform the radiation pattern. Each guy has such a network so that the total antenna system is protected against atmospheric electricity and no other special protection devices are needed.

7.2.2 *The Lopik MF antenna (120 kW on 675 kHz)*

In this base-fed $5\lambda/8$ antenna, the insulators in the guys are also bridged by a copper wire, except for those two at the ends. Coils parallel to the lower insulators minimize the currents in the guys. With this arrangement, the transmitter no longer keeps tripping in order to protect the guy insulators.

7.3 Yugoslavia [CCIR, 1982-86a]

7.3.1 The Belgrade MF antenna (2000 kW on 684 kHz)

An efficient and simple solution for eliminating static electricity has been applied to the 235 m high Belgrade MF antenna. It is based on the use of static-drain coils, which are connected in parallel to the guy insulators, forming anti-resonant circuits with them at the operating frequency [Surutka, 1978, 1979; Surutka and Veličković, 1984].

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SECTION 10A-2: SOUND BROADCASTING IN THE TROPICAL ZONE

REPORT 303-3*

**DETERMINATION OF THE EFFECTS OF ATMOSPHERIC NOISE
ON THE GRADE OF RECEPTION IN THE TROPICAL ZONE**

(Question 45/10, Study Programme 45D/10)

(1978-1986)

1. Introduction

One of the primary requisites for establishing the limits of power for sound broadcasting is the determination of the value of the signal-to-noise ratio required for satisfactory listening in the absence of any other interference. Availability of reliable noise data is, therefore, necessary. In most of the countries in the Tropical Zone, which are either underdeveloped or developing and where man-made noise is comparatively low, atmospheric radio-noise is the principal source of interference in band 7 (HF).

2. Results of studies

2.1 The characteristics of atmospheric radio noise such as amplitude, duration and repetition rate of individual impulses are important in determining the noise response of the receiver. Atmospheric radio noise can be classified as either "fluctuation noise" or "impulsive noise". A theoretical analysis of the problem, i.e. dependence of the response of the receiver on fluctuation and impulsive noise has been made by [Thomas and Burgess, 1947]. It has been concluded there that:

2.1.1 the r.m.s. and peak values of the output response produced in a receiver by fluctuation noise are both proportional to the square root of the receiver bandwidth;

2.1.2 the r.m.s. and peak values of the output response produced in a receiver by a unit impulse are respectively proportional to the bandwidth and the square root of the bandwidth;

2.1.3 the output waveform is independent of the waveshape of the impulse if its duration is less than the order of the reciprocal of the receiver bandwidth, and is dependent only on its time integral;

2.1.4 the minimum value of tolerable signal-to-noise ratio for a high quality sound-broadcasting service is 40 dB.

[Landon, 1936; Jansky, 1939] have also found, on the basis of some experiments, the same bandwidth relationship.

2.2 Studies in India [CCIR, 1956a and CCIR, 1959] and South Africa [CCIR, 1956b] have led to the conclusion that a signal-to-noise ratio of approximately 40 dB is necessary to establish a satisfactory broadcasting service.

* In view of the opinion expressed by Study Group 6, Study Group 10 will consider the possibility of transferring this text to Study Group 6 at the appropriate time.

3. Conclusion

Results of the studies, relating bandwidth and noise response, are mainly based on theoretical considerations and laboratory experimentation. It is, therefore, necessary that a detailed study under actual conditions, correlating the receiver response and noise encountered in the Tropical Zone, be made to find an answer to § 1 of Question 45/10.

On the basis of the studies described in this Report, it is also reasonable to recommend provisionally a signal-to-noise ratio of 40 dB to ensure a satisfactory reception for at least 90% of the time. Further investigation by different administrations in the Tropical Zone is required to confirm this.

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REPORT 302-1

INTERFERENCE TO SOUND BROADCASTING IN THE SHARED BANDS IN THE TROPICAL ZONE

(Question 45/10 and Study Programme 45E/10)

(1956-1959-1963-1978)

1. Introduction

This Report is a summary of the information available on the question of interference in the bands shared with sound broadcasting in the Tropical Zone for the determination of the radio-frequency protection required for a broadcast signal in the presence of interference. The radio-frequency (RF) protection ratio is the value of the radio-frequency wanted to-interfering signal ratio that enables, under specified conditions, the necessary audio-frequency protection ratio to be obtained at the output of a receiver (Recommendation 638). These specified conditions include such diverse parameters as spacing Δf of the wanted and interfering carrier, emission characteristics (type of modulation, modulation depth, frequency deviation, etc.), receiver input and output levels as well as the receiver characteristics (selectivity and susceptibility to cross-modulation, etc.). The audio-frequency (AF) protection ratio is the agreed minimum value of the audio-frequency signal-to-interference ratio considered necessary to achieve a subjectively defined reception quality. This ratio may have different values according to the type of service desired (Recommendation 638). The wanted signal is from a broadcasting station operating in the shared bands in the Tropical Zone and the interfering signal may be either:

- from A1A/A1B and A2A/A2B telegraphy,
- A3E telephony and broadcasting or combinations of more than one type of emission.

Agreed values of protection ratios are necessary for the solution of frequency assignment problems in amplitude modulation sound-broadcasting systems and may also serve as basic reference data for the evaluation of the relative effectiveness to be expected of various amplitude modulation transmission systems.

The minimum wanted field to which this protection ratio is to be maintained is also to be determined taking into account the noise levels prevalent in different parts of the Tropical Zone.

In this Report, Part I deals with the considerations for the determination of protection ratios and Part II is concerned with the determination of the minimum value of the wanted field to which the recommended protection ratio is to be maintained.

PART I (PROTECTION RATIO)

2. Studies and measurements

Various administrations have carried out studies to determine, by subjective tests, the protection ratios for different percentages of listener satisfaction. In general, the measurements have been carried out at the output of the receiver, fitted with simple filters having an audio-frequency cut-off of 5 kHz. Some measurements have been made for cut-off frequencies of 6, 7, 8 and 9 kHz also. These measurements have been carried out for various carrier frequency separations up to 10 kHz. The permissible frequency tolerances of both wanted and unwanted signals, as specified in Appendix 7 of the Radio Regulations, have been taken into account while determining the protection ratios.

3. Results

3.1 [CCIR, 1956a], gives the results of an extensive series of subjective tests carried out by generally simulating domestic broadcast listening conditions in the absence of fading. A broadcast receiver, with a substantially flat response up to about 4 kHz, but with a filter giving an attenuation of about 8 dB at 5 kHz and a sharp cut-off above this frequency, was used.

For unwanted signals of classes of emission A2A/A2B and A3E and for various frequency separations between the carriers of the wanted and unwanted signals, listeners were presented with various ratios of wanted-to-unwanted signal in random order and asked to state whether they considered the reception satisfactory or unsatisfactory. Table I gives information for the wanted-to-unwanted signal ratios required to provide 90%, 70% and 50% listener satisfaction for unwanted signals of classes of emission A2A/A2B telegraphy, A3E telephony and broadcasting and for:

- frequency separation of 0 kHz and 5 kHz exactly; and
- nominal frequency separations of 0 kHz and 5 kHz under the most unfavourable conditions that could arise within the maximum permissible frequency tolerances of both wanted and unwanted signals, as were specified in the Radio Regulations, Atlantic City, 1947.

Utilizing the information given in the above document, curves giving relative protection ratios for frequency separation between 0 and 5 kHz have been derived and are shown in Fig. 1. The actual values for any frequency separation up to 5 kHz can, however, be obtained from these curves and from the value of co-channel protection ratio given in Table I for different percentages of listener satisfaction.

3.2 [CCIR, 1956b], gives details of the results of subjective tests made to determine the ratio of wanted-to-unwanted signal as a function of the frequency separation of the carriers of the two signals. Two typical broadcast receivers were used, having a fairly uniform response up to about 4 kHz falling to about –8 dB to –10 dB at 5 kHz. The unwanted signal was modulated by speech with a frequency range limited to 3 kHz. The ratio necessary to satisfy nearly all listeners varied from about 54 dB at 1 kHz separation, to a maximum of 56 dB between 2 and 3 kHz separation, falling to 52 dB at 5 kHz separation. The corresponding ratios when nearly all the listeners found the conditions unsatisfactory, were about 15 dB lower. Subsequent tests to determine the ratio at which interference was “perceptible” gave intermediate values.

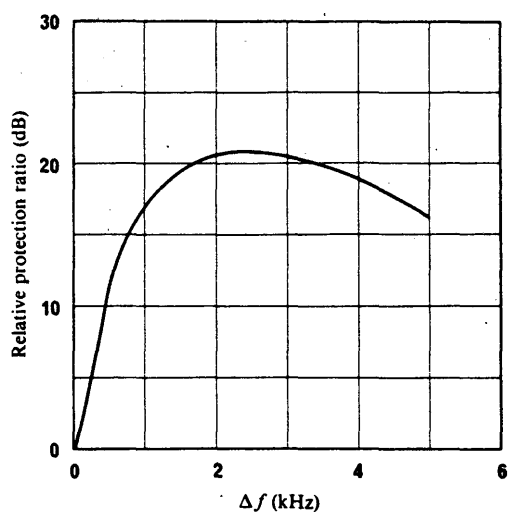
3.3 [CCIR, 1956c], gives the results of similar tests made with two types of receiver, one a narrow-band receiver with considerable attenuation above 3 kHz, and the other a wider-band receiver with an attenuation of about 8 dB at 5 kHz. For the wide-band receiver, as commonly used for broadcast listening, the wanted-to-unwanted signal ratio for various frequency separations follows the same general curve as before and, at a frequency separation of 5 kHz, is 43 dB for 90% listener satisfaction.

3.4 Comparison of the results arrived at in the three documents shows that there is a considerable degree of agreement. The values are within ± 5 dB and those in the United Kingdom and Federal German Republic documents are almost of the same order as those given by India. There is, therefore, sufficient justification to assume that the values of the wanted-to-unwanted signal ratio which provide the various degrees of listener satisfaction given in Table I are reliable.

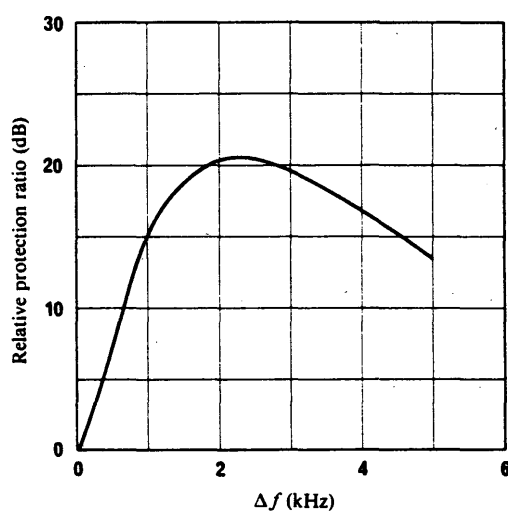
From an examination of Table I, it will be seen that, when the unwanted signal is a mobile A3E emission, there is a considerable increase in the required wanted-to-unwanted signal ratio when allowance is made for maximum frequency tolerances. The possibility of interference to broadcasting services from mobile services would be appreciably reduced, particularly where the two services have the same nominal frequency, if the mobile services operated within closer frequency tolerances, if possible with the same tolerance as the fixed and broadcasting services.

TABLE I

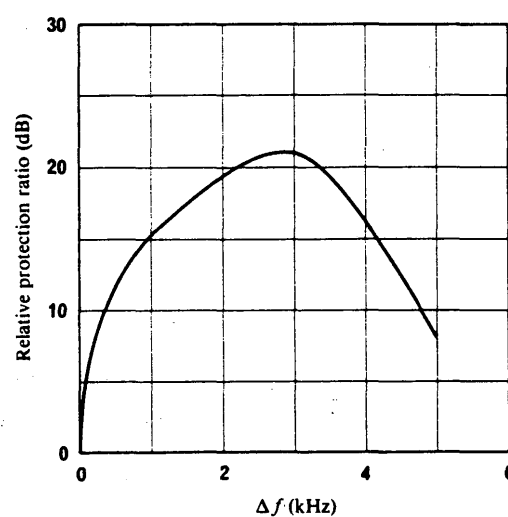
Interfering emission	Maximum frequency tolerance (App. 7 to Radio Regulations) (Hz)	Frequency separation (kHz)	Signal-to-interference ratios for 90%, 70% and 50% listener satisfaction (dB)					
			Ignoring frequency tolerances			Allowing for maximum frequency tolerances		
			90%	70%	50%	90%	70%	50%
A2B – fixed (525 Hz tone)	150	0	35	31	28	42	38	34
A2B – mobile (525 Hz tone)	1000	0	35	31	28	49	45	42
A3E – fixed (3 kHz maximum modulation)	150	0	33	30	28	40	36	33
A3E – mobile (3 kHz maximum modulation)	1000	0	33	30	28	50	47	44
A3E – broadcasting	150	0	33	30	28	44	40	36
A2B – fixed (525 Hz tone)	150	5	39	37	36	43	40	38
A2B – mobile (525 Hz tone)	1000	5	39	37	36	49	46	43
A3E – fixed (3 kHz maximum modulation)	150	5	48	44	40	50	46	42
A3E – mobile (3 kHz maximum modulation)	1000	5	48	44	40	52	48	45
A3E – broadcasting	150	5	48	46	44	49	46	44



a) Wanted: broadcasting
Unwanted: broadcasting



b) Wanted: broadcasting
Unwanted: telephony (A3E)



c) Wanted: broadcasting
Unwanted: telegraphy
(A3B with 525 Hz tone modulation)

FIGURE 1 - Relative protection ratio as a function of frequency separation, Δf

Although the sidebands of the unwanted signal contributed to some extent to the interference, the heterodyne beat note between the carriers of the wanted and unwanted signals was always predominant. This was the case for a frequency separation of 5 kHz between the two signals and, although the receivers used provided an attenuation of some 8 to 10 dB to the beat note, the use of a filter to provide further attenuation would have reduced the required wanted-to-unwanted signal ratio. Further studies are needed to ascertain what additional attenuation at 5 kHz could usefully be provided and what would then be the required wanted-to-unwanted signal ratio. For this purpose, consideration should also be given to the possibility of providing suitable filters in new and existing receivers.

It is agreed that since the figures shown in Table I are derived from measurements made under steady-state conditions, appropriate allowance should be made for fading, when using these figures to derive the protection ratios to be used in practice. The value of fading allowance to be used for broadcasting in the Tropical Zone requires further study.

India has given the opinion that protection ratios should be based on those figures in Table I that provide 90% listener satisfaction and that the figures for 70% and 50% are for information only and should not be regarded as the lower limit of acceptability. Australia, the Republic of South Africa and France are of the opinion that a listener satisfaction higher than 50% should be provided. The Republic of South Africa and Australia consider that it would be impracticable to achieve 90% listener satisfaction from the point-of-view of signal-to-noise ratio, particularly under heavy static conditions, and therefore, to aim at achieving about 80% listener satisfaction for signal-to-interference would be more realistic. The United Kingdom is of the opinion that the wanted-to-unwanted signal ratios given in Table I are based on too critical an assessment and that lower figures would be generally acceptable. The United Kingdom is also of the opinion that, because of practical considerations, protection ratios will have to be based on figures providing about 50 to 70% listener satisfaction.

3.5 [CCIR, 1959a], summarizes the data presented at various times regarding the protection ratio of the wanted-to-unwanted signal that is required for just tolerable interference at various values of separation of carrier frequencies. The curves obtained in the document are given in Fig. 2. A summary of the conditions of the Post Office tests is also given in the document and is reproduced in § 3 of Annex I.

3.6 [CCIR, 1959b], describes work carried out in connection with Question 1/XII. Protection ratios required against A1A and A1B emissions, both for speech and music programmes, A2A/A2B and A3E emissions for music programmes have been assessed. The results are in § 4 of Annex I.

[CCIR, 1959b] takes into consideration the standards for frequency tolerances laid down in the Radio Regulations, Atlantic City, 1947. The summary is confined to two limiting cases, namely frequency separations of 0 and 5 kHz respectively and indicates the protection ratios required for various types of emission. The document also states that the results refer to steady-state conditions and that an appropriate allowance should be made for fading.

An analysis of selectivity characteristics of receivers in use in India is also given in the document. Extensive tests were carried out to investigate, from the point of view of listener satisfaction, the effect of reducing the bandwidth of broadcast transmissions on overall quality.

The conclusions in the Indian document is that it is necessary to maintain the normal bandwidth of modulating frequencies to well beyond 5 kHz. Any modifications to the design of broadcast receivers tending to attenuate frequencies at 5 kHz and lower will, therefore, result in serious deterioration in the quality of reception.

3.7 [CCIR, 1962 and 1963a], give results of further listening tests for frequency separation of 5 to 10 kHz. The experimental set-up was the same as before, except that, as recommended in Study Programme 1C/XII (1966), filters with cut-off frequencies at 5, 6, 7, 8 and 9 kHz were incorporated at the output of the receiver (ordinary domestic type with average selectivity and fidelity characteristics). The attenuation at 1 kHz below the nominal cut-off frequency of the filter is 6 dB; it is more than 30 dB at the cut-off frequency and very steep thereafter. The wanted signal was modulated with a speech programme, the interfering signals were modulated with music, speech, A1A/A1B and A2A/A2B telegraphy (Morse — 525 Hz tone modulation). The documents show that, on the basis of these experiments, which relate to values required for 90% listener satisfaction, the protection ratio required in each case gradually decreases as the carrier separation increases beyond 5 kHz and also as the cut-off frequency of the filter decreases. Table II gives the values of the protection ratios required for various types of interference when no filter is used in the output circuit of the receiver.

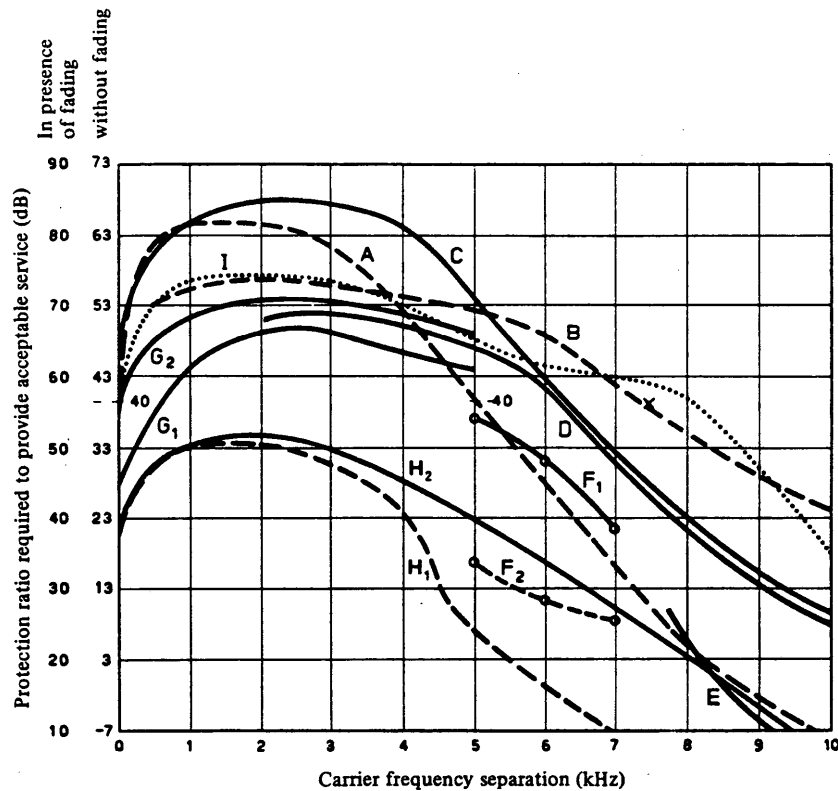


FIGURE 2 - Protection ratios required to provide acceptable service

DESIGNATION OF CURVES

A*	VAN DER POL (1933)	G ₁ **	Indian tests (50% satisfaction)
B*	BRAILLARD (CCIR, Bucharest, 1937)	G ₂ **	Indian tests (90% satisfaction)
C*	B.P.O. tests, 1948	H ₁ **	Curve used by the IFRB, 1956, for HF broadcast plans
D**	B.P.O. tests, 1950	I***	French tests, 1962
E**	B.P.O. tests, 1951	H ₂ **	Curve drawn from values of "Receiver Discrimination" given in IFRB Technical Standards Series "A" 1968.
F ₁ **	B.P.O. tests, 1956 (no filter)		
F ₂ **	B.P.O. tests, 1956 (whistle filter)		

- * Criterion of test: Just perceptible interference
- ** Criterion of test: Just tolerable interference
- *** Corresponding to a "tolerable" interference for five different types of receivers.

TABLE II

Wanted signal	Interfering signal	Frequency separation (kHz)	Desired protection ratio (dB)
Speech	Music	5	46
Speech	Music	10	22
Speech	Speech	5	44
Speech	Speech	10	16
Speech	A3B telegraphy	5	38
Speech	A3B telegraphy	10	8
Speech	A1B emission	5	38
Speech	A1B emission	10	8
Music	Music	5	38
Music	Music	10	12
Music	Speech	5	38
Music	Speech	10	6

Note. — The protection ratios quoted refer in all cases to the ratios at the input to the receiver, no account having been taken of the effect of using directional receiving antenna or of the advantage that can be obtained by using different polarization for transmission of the wanted and unwanted signals.

With the incorporation of filters at the output of the receiver, the required protection ratios become less; the degree of reduction depends upon the frequency separation, the cut-off frequency of the filter and to some extent on the nature of the interference. The details are shown in [CCIR, 1962], (Table II and Figs. 1 to 5) and [CCIR, 1963a] (Table III and Figs. 1 to 3).

The documents conclude that, in assessing the required protection, the allowable frequency tolerance limits of various emissions must be taken into account. Considering the frequency tolerance standards, as laid down in the Radio Regulations, the increase in the required protection has been estimated. For a frequency of operation of 5 MHz, in the limiting case of interference from a broadcast station, the increase in required protection would be of the order of 2 dB for 5 kHz frequency separation and 1 dB for 10 kHz frequency separation. For frequencies of operation higher than 5 MHz, such protection ratios would be still higher. In the other limiting cases of interference from mobile stations, the increase in protection ratio at 5 MHz would be of the order of 4 dB for 5 kHz frequency separation and 3 dB for 10 kHz frequency separation. The incorporation of filters with lower cut-off frequencies would result in the reduction of the required protection ratios, only at the cost of the quality of received music programmes.

Since the results presented have been derived under steady-state conditions, appropriate allowance should be made for fading under actual operating conditions.

3.8 [CCIR, 1963b], describes the results of measurements of protection ratio made in the medium-frequency band under steady-state conditions on various types of receivers of recent manufacture.

The measurements were made by the usual method with a wanted signal and an unwanted signal, the level of the latter being set to give a level of interference considered tolerable by the listeners. Various types of programme were used for the tests but programme material particularly susceptible to interference (e.g. music of high quality with a wide dynamic range) and programmes only slightly susceptible to interference (e.g. modern dance music with a restricted dynamic range) were excluded.

The results of the measurements are shown by curve I of Fig. 2.

There are two main conclusions to be drawn from [CCIR, 1963b], namely that, for medium frequency reception at least, and with the great majority of receivers in use, the protection ratio necessary with no frequency separation (± 20 Hz) is 40 dB, and with 5 kHz spacing, 46 to 50 dB.

The incorporation of special cut-off filters would reduce the required protection ratios, but only at the expense of the quality of received music programmes.

4. PART II (MINIMUM FIELD TO BE PROTECTED)

4.1 [CCIR, 1956d] indicates that listening tests have shown that for most of the time a satisfactory signal-to-noise ratio (estimated at 40 dB) can be obtained with a broadcasting system using arrays having vertical incidence radiation and a transmitter power of 5 kW, within a local service area.

For a given power and frequency the day-time field within a given service area is less than the night-time value. On the other hand the day-time noise levels are considerably less than the night-time values.

With the limited amount of information available on atmospheric noise levels in the tropical areas, the day-time and night-time median signal strengths of 100 $\mu\text{V/m}$ and 1000 $\mu\text{V/m}$ respectively would be necessary. These would provide a satisfactory broadcasting service in the Tropical Zone using bands shared with broadcasting having a median signal to upper decile noise ratio of 30 dB (median signal to median noise ratio of 36 dB).

In the light of the above and the fact that the fixed and mobile services must set up a satisfactory signal-to-noise ratio in their own service areas besides giving the necessary protection to the broadcast services, it has been suggested by the United Kingdom that the minimum value of the field intensity of a broadcast signal to be protected in the bands 2300 kHz to 5060 kHz within the local service area, should be 150 $\mu\text{V/m}$ by day and 250 $\mu\text{V/m}$ by night.

4.2 [CCIR, 1959c], gives the results of studies carried out for determination of noise level in tropical regions and the minimum signal required for satisfactory listening. Based on these studies, the minimum signal required for various frequency bands for different times of the day and seasons is given in Fig. 3. These figures indicate that for satisfactory reception higher values of signal are required during night-time as compared with day-time.

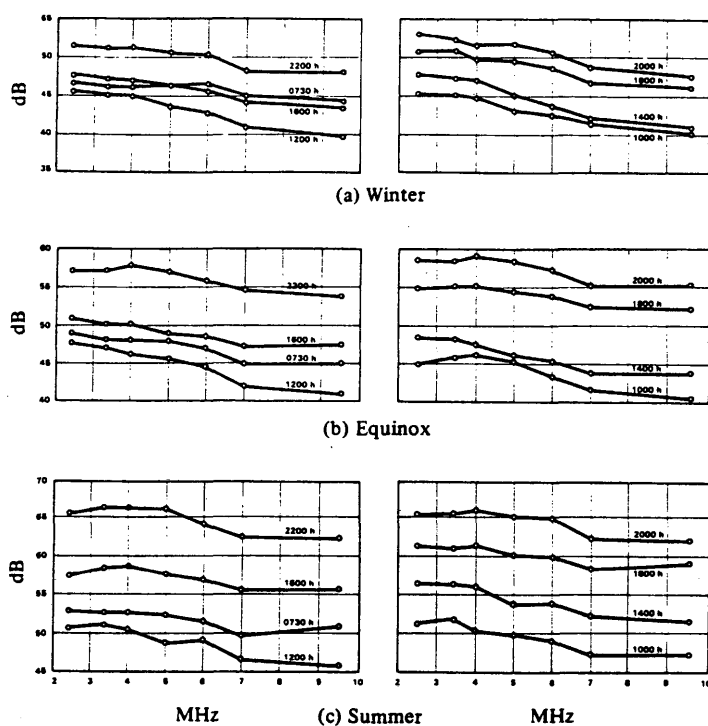
4.3 The conclusions of the two documents referred to in §§ 4.1 and 4.2, that noise level is considerably higher at night-time as compared with day-time, are also indicated in Report 322.

5. Conclusions

The protection ratio values given in Tables I, II and IV and those in Fig. 2 indicate that there is some divergence in the results that have been obtained. This may be due to the variety of experimental set-ups, types of receiver and different percentage of listener satisfaction employed by various administrations for the determination of protection ratios. Nevertheless, the values given in Tables I, II and IV and those in Fig. 2 for different frequency spacings and percentages of listener satisfaction give a consistent trend and the results could be applied to the solution of frequency assignment problems. However, the curves 'H1' and 'H2' give protection ratios much lower than any of the measured data (Fig. 2).

With the present state of unsatisfactory interference position in high frequency bands, it would be difficult to realise in practice the values of protection ratios given by the curves A to G and I of Fig. 2. Comprehensive data on radio-frequency protection ratios may perhaps be collected now by other than the subjective methods hitherto used, employing objective methods as described in CCIR Recommendation 559. Based on the data so collected, further studies may be necessary to determine the lower limit of protection ratios which would be acceptable in practical considerations.

Regarding the minimum field to be protected, it may be mentioned that due to the prevalence of much higher noise levels during night-time as compared with day-time in the Tropical Zone the night-time field should be higher than the day-time field. This aspect has to be taken into consideration while stipulating the minimum signal to be protected for satisfactory listening.

FIGURE 3 – Minimum signal required for satisfactory listening (dB ($\mu V/m$))

ANNEX I

1. Conditions of test for curve A

The tolerable signal-to-interference ratio (with the receiver tuned to the wanted signal), was chosen as the criterion of receiver sensitivity.

The receiver sensitivity was adjusted to apply 150 mW low-frequency power to the loudspeaker, the wanted signal being modulated by a 400 Hz tone to a depth of 30%. The quasi-maximum of the modulation of the interfering signal corresponded to a modulation index of 90%.

The amplitude of the interfering signal was then increased, up to the point when its interfering effect on an unmodulated wanted signal was just perceptible to the ear at a distance of about 50 cm from the loudspeaker.

Furthermore, if the wanted signal was also modulated, the above ratio may be multiplied by a factor of between 3 and 5.* (Documents of the European Radiocommunication Conference, Lucerne, 1933, 280-282, and Documents of the Fourth Reunion of the CCIR, Bucharest, 1937, Vol. I, 109-112.)

2. Conditions of test for curve B

Similar conditions to those of curve A, but relating to very high-quality reception (Documents of the CCIR, Bucharest, 1937, Vol. I, 241).

* Note by the Director, CCIR. – The original van der Pol curve was plotted as the ratio of the "tolerable interfering signal to the wanted signal". Accordingly, with the curve A of Fig. 2, which is plotted as the ratio of the wanted signal to the interfering signal, this factor should be 1/3 to 1/5 (–9.5 to –14 dB).

3. Conditions of tests for curves C, D, E and F (British Post Office tests)

The conditions under which tests F were conducted require some detailed comment.

A "standard" condition of co-channel interference was set up, this in the first place providing an interfering broadcast signal 23 dB below the wanted carrier level. As finally set up, however, short-term Rayleigh-type fading was introduced and, on the basis of some practical evidence, this was taken to require the co-channel protection ratio to be increased to 33 dB. An allowance of 7 dB for long-term fading would thus give the figure of 40 dB for planning purposes, as used at the Mexico City Broadcasting Conference, 1948. The tests F, however, were carried out only with artificially produced short-term fading and all adjacent channel figures, therefore, relate to a co-channel protection ratio of 33 dB. In the final presentation, these figures have therefore been reduced by 10 dB to equate the results to the non-fading conditions used for all other tests. The ordinate has been so labelled that protection ratios can be read off either for non-fading conditions, or for full fading conditions incorporating a total allowance of $10 + 7 = 17$ dB for short-term and long-term fading. For some of these measurements under test F, a simple whistle filter was placed in the loudspeaker input leads, so that the improvement in protection ratio, that might readily be gained by reducing the audible heterodyne whistle at 5, 6 and 7 kHz, could be assessed.

TABLE III

Test	Date of test	Wanted signal		Unwanted signal	
		Type	Modulation index	Type	Modulation index
C	1948	Music (0 – 8 kHz)	30% average, peaking to 100% occasionally	Speech (0 – 8 kHz)	30% average, peaking to 100% occasionally
D	1950	Broadcast speech	idem	Telephony (0 – 3 kHz)	70%
E	1951	Broadcast speech	idem	Speech (0 – 6 kHz)	30% average, peaking to 100% occasionally
F	1956	Broadcast speech	idem	Music (6 dB down at 4.6 kHz)	idem

4. Conditions of test for curves G1 and G2 (Indian tests with 50% and 90% listener satisfaction respectively for broadcasting)

See Tables I and IV.

5. Conditions of test curves H1 and H2

5.1 Curve H1 was the protection ratio used by the IFRB in the preparation of the tentative 1956 HF broadcasting plans only.

5.2 Curve H2 is the modified H1 curve based on the data of "receiver discrimination" given in IFRB Technical Standards, Series A, 1968 (Ref. CCIR Recommendation 449-2 (Geneva, 1974)), where both wanted and the interfering emissions are broadcasting emissions.

6. Conditions of test for curve I

See § 3.8 of the Report.

TABLE IV

Wanted signal	Interfering emission	Frequency separation (kHz)	Protection ignoring frequency tolerance (dB)	Maximum frequency tolerance in the shared bands (Atlantic City) (Hz)	Protection taking into account the tolerance in column 5 (dB)
Speech	A1B-fixed (40 w.p.m.)	0	26.5	150	33.5
	A1B-mobile (40 w.p.m.)	0	26.5	1000	44.5
	A3B-fixed (mod. at 525 Hz)	0	35	150	42
	A3B-mobile (mod. at 525 Hz)	0	35	1000	49
	A3E-fixed (mod. at 3 kHz max.)	0	33	150	40
	A3E-mobile (mod. at 3 kHz max.)	0	33	1000	50
	A3E-broadcasting	0	33	150	44
	A1B-fixed (40 w.p.m.)	5	41.5	150	43
	A1B-mobile (40 w.p.m.)	5	41.5	1000	47
	A3B-fixed (mod. at 525 Hz)	5	39	150	43
	A3B-mobile (mod. at 525 Hz)	5	39	1000	49
	A3E-fixed (mod. at 3 kHz max.)	5	48	150	50
	A3E-mobile (mod. at 3 kHz max.)	5	48	1000	52
	A3E-broadcasting	5	48	150	49
	A3E-broadcasting	0	33	645	51.5
	A3E-broadcasting	5	48	645	50.5
Music (vocal)	A1B-fixed (40 w.p.m.)	0	27.5	150	36
	A1B-mobile (40 w.p.m.)	0	27.5	1000	42.5
Music (instrumental)	A3B-fixed (mod. at 525 Hz)	0	24	150	28.5
	A3B-mobile (mod. at 525 Hz)	0	24	1000	36
Music (vocal)	A3E-fixed (mod. at 3 kHz max.)	0	26	150	34
	A3E-mobile (mod. at 3 kHz max.)	0	26	1000	41.5
	A1B-fixed (40 w.p.m.)	5	37	150	39
	A1B-mobile (40 w.p.m.)	5	37	1000	43
Music (instrumental)	A3B-fixed (mod. at 525 Hz)	5	39	150	40
	A3B-mobile (mod. at 525 Hz)	5	39	1000	43
Music (vocal)	A3E-fixed (mod. at 3 kHz max.)	5	42.5	150	44
	A3E-mobile (mod. at 3 kHz max.)	5	42.5	1000	46.5

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REPORT 304-3*

FADING CHARACTERISTICS FOR SOUND BROADCASTING IN THE TROPICAL ZONE

(Question 45/10, Study Programme 45B/10)

(1956-1959-1963-1978-1986-1990)

1. Introduction

The fluctuations of the received signal in time, space and frequency which occur as random short-period variations, irregular long-period variations and more or less regular variations, are collectively described as "fading". To plan sound broadcasting services, it is necessary to have precise information on the types and characteristics of fading that may be expected to be encountered in the received signal.

2. Types of fading

Fading may be caused by various effects of ionospheric phenomena. Depending on the cause, the different types of fading, in general, may fall into the following principal classes.

2.1 Interference fading

Interference fading results from interference between two or more waves which travel by different paths to arrive at the receiving point. This type of fading may be caused by:

- interference between sky wave and ground wave;
- interference between multiple reflected sky waves;
- interference between the two oppositely polarized magneto-ionic components - the ordinary and the extraordinary waves;
- interference between various wavelets diffractively scattered from different scattering centres, caused by the steady horizontal drift of the irregularities present in the ionosphere or by their random movement or both.

Interference fading may last for a period of a fraction of a second to a few seconds, during which time the resultant field intensity may vary over wide limits.

2.2 Polarization fading

Polarization fading occurs as a result of changes in the direction of polarization of the downcoming wave, relative to the orientation of the receiving antenna, due to random fluctuations in the electron density along the path of propagation.

Like interference fading, polarization fading also lasts for a period of a fraction of a second to a few seconds.

* The attention of Study Group 6 is drawn to this Report.

2.3 *Absorption fading*

Absorption fading is caused by variation in the absorption of radiowave propagated through the ionosphere due to changes in the densities of ionization and it usually lasts longer and sometimes may last even longer than one hour.

2.4 *Skip fading*

Skip fading may be observed at receiving locations near the skip distance at about sunrise and sunset, when, because of the instability of ionization density of ionosphere, MUF for a transmission path may oscillate around the actual frequency. The amplitude of the received signal may fall abruptly when the skip distance crosses over the receiving point and may suddenly increase with the decrease in the skip distance, thus causing fading.

2.5 *Selective fading*

In general, fading is faster at high frequencies than at low frequencies, because for given movement in the ionosphere, there is a greater phase-shift on the shorter wavelengths either on reflection or on scattering from irregularities. Sometimes on a modulated carrier, the frequency components in the sidebands fade differently and independently, giving rise to distortion of the modulation envelope. This is called "selective fading".

In general, all the above types of fading are encountered in the Tropical Zone. However, some special types of fading observed only in the Tropical Zone and at low magnetic latitudes are discussed in § 4.

3. *Character of fading*

The terms "severity" and "rapidity" of fading refer to the characteristics of the variation of the amplitude of the received signal.

3.1 *Severity of fading*

Severity or depth of fading is measured by the amplitude distribution or probability distribution function.

3.1.1 *Analysis of severity of fading*

In a fading curve, i.e. the variation in the received signal versus time, if R is the amplitude of the downcoming wave at any instant t , and dR represents a slight change in R in the time interval $(t + dt)$, the probability of finding the amplitude between the limits R and $(R + dR)$ is $P(R) dR$, $P(R)$ being the probability density function of R .

Probability density functions, computed analytically to describe the envelope of a fading signal, differ in form and depend upon the mechanism of fading of the received signal. Depending upon the amplitude and phase of the signals which form the resultant descending wave, the amplitude distributions normally observed conform in general to the following three standard statistical curves:

- Rayleigh distribution,
- normal or Gaussian distribution,
- log-normal distribution.

3.1.1.1 *Rayleigh distribution*

If the downcoming wave consists of a large number of waves scattered from the irregularities and moving in a random manner, the resultant amplitude of the signal at any instant will be the vector sum of the scattered waves each having a random phase. The probability distribution of the resultant amplitude R of the envelope will then be represented by a Rayleigh distribution.

$$P(R) = (R/\psi) \exp(-R^2/2\psi) \quad (1)$$

where,

R : amplitude at any instant,

ψ : density in the power spectrum and a parameter related to the average value $\bar{R}$, the r.m.s. value R_0 and the most probable value R_m as follows:

$$\bar{R} = \sqrt{\pi\psi/2}, R_0 = \sqrt{2\psi}, R_m = \sqrt{\psi}$$

It is sometimes convenient to evaluate ψ from the plot of $\log [P(R)/R]$ as a function of R^2 . The corresponding log plot is a straight line, the slope of which is $\psi/2$ and the intercept $\log (1/\psi)$.

Fig. 1(a) shows the Rayleigh type of probability density function of amplitude R , and Fig. 1(b), shows the corresponding plot of $\log [P(R)/R]$ against R^2 .

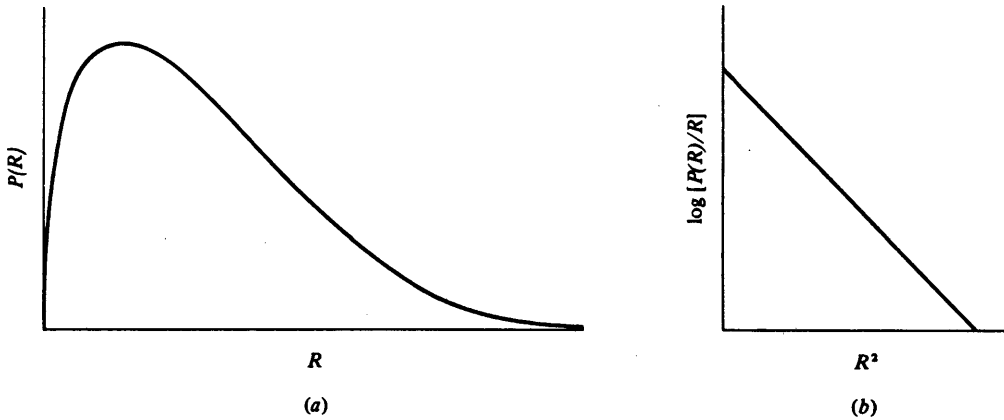


FIGURE 1

3.1.1.2 Normal or Gaussian distribution

The normal distribution is represented by:

$$P(R) = \left(1/\sigma \sqrt{2\pi}\right) \exp \left[-\frac{1}{2} \left(\frac{R - R_M}{\sigma} \right)^2 \right] \quad (2)$$

where σ is the standard deviation in Gaussian distribution and R_M is the median value of R .

The normal distribution is usually observed when the downcoming wave consists of a random component and a steady component because of specular reflection in the ionosphere. The contribution due to the specular component is of a considerable amount in comparison with the random one.

Fig. 2 shows the probability density function following a normal or Gaussian distribution:

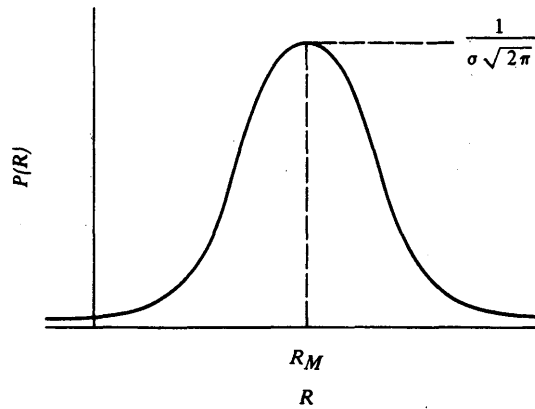


FIGURE 2

3.1.1.3 Log-normal distribution

The amplitude density function of fading records representing a log-normal distribution is given by:

$$P(x) = (1/\sigma_x \sqrt{2\pi}) \exp(-x^2/2\sigma_x^2) \quad (3)$$

where,

$$x = 20 \log(R/R_M)$$

R_M is the median value of the amplitude R . In this case the field strength measured logarithmically is distributed normally. Fig. 3 gives a typical log-normal distribution curve.

3.1.1.4 A general formula for intensity distribution

For a general form of fading, the signal at the receiving point may be assumed to be composed of a steady sinusoidal component and a random fluctuating component. This leads to a probability density function described by several workers [Nakagami, 1943; McNicol, 1949; Nakagami, 1960].

In this case, following work of Rice [1944-1945], the probability density of the resultant amplitude Q of the envelope may be expressed as:

$$P(Q) = (Q/\psi) \exp[(-Q^2 + B^2)/2\psi] I_0(QB/\psi) \quad (4)$$

where,

I_0 : is the Bessel function of zero order and imaginary argument;

B : is the amplitude of the steady signal;

ψ : is related to the amplitude of the random fluctuating component in the following manner:

$$\psi = R_0^2/2 \text{ or } R_m^2$$

where,

R_0 : is the r.m.s. value;

R_m : is the most probable value of the random component.

In general $\overline{Q^2} = B^2 + 2\psi$.

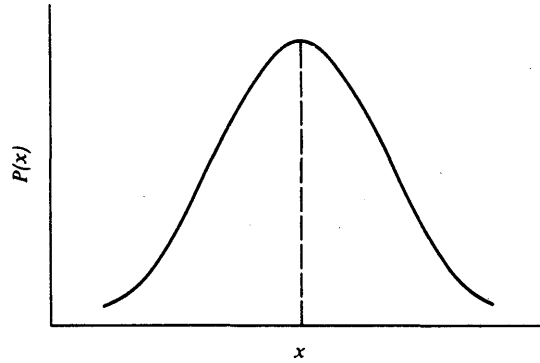


FIGURE 3

Fig. 4 describes the general behaviour of $P(Q)$ in terms of b , where $b = B/\sqrt{\psi}$, a ratio of the amplitude of the steady signal and the most probable value of the random signal.

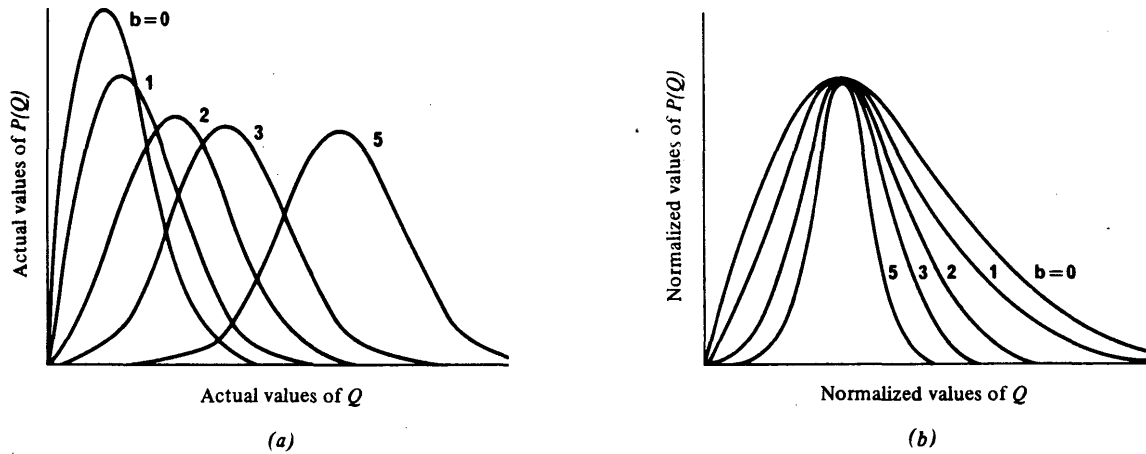


FIGURE 4 - Probability densities of the amplitude Q of a steady signal plus a random signal

(a) Actual curves

(b) Normalized curves

If $b < 1$, then equation (4) reduces to:

$$P(Q) \approx (2Q/R_0^2) \exp(-Q^2/R_0^2) \quad (5)$$

which approximates to a Rayleigh distribution.

If $b > 3$,

$$P(Q) \approx (1/\sqrt{2\pi\psi}) \times \sqrt{Q/B} \times \exp [-(Q-B)^2/2\psi] \quad (6)$$

The distribution is Gaussian, with a standard deviation of $\sqrt{\psi}$.

McNicol [1949] has shown that (6) can be rewritten in the form:

$$P(Q) = (1/\sqrt{2\pi\psi}) \times \exp [-(Q-Q_m)^2/2\psi] \quad (7)$$

within an accuracy of 1%;

where,

$$Q_m = \sqrt{\psi(b^2 + 1)}$$

A plot of $\log P(Q)$ as a function of $(Q - Q_m)^2$ gives a straight line, the slope of which is $1/2\psi$ as shown in Fig. 5. Knowing ψ and Q_m (from the distribution curve) one can evaluate b , which represents the relative proportion between the steady and the random signals in the received wave.

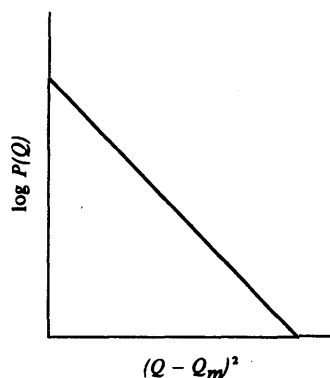


FIGURE 5

In addition to the above probability density functions, another distribution, " m -type" may also occasionally be encountered.

The functional form of the m -distribution by Nakagami [1960] as:

$$P(R) = \frac{2m^m R^{2m-1}}{\Gamma(m) \Omega^m} \exp -(m/\Omega) R^2 \quad (8)$$

where, $\Omega = \overline{R^2}$ (time average of R^2)

and

$$m = (\overline{R^2})^2 / (R^2 - \overline{R^2})^2 \geq 0.5$$

always, that is the inverse of the normalized variance of R^2 .

Fig. 6 shows an m -type amplitude distribution.

Quantitative estimates of the severity of fading which can be obtained from the statistical distributions of the received signal are dealt with in § 5.

3.1.2 Experimentally observed amplitude distribution

Measurements on fading on transmissions in several areas in the Tropical Zone at different times of the day and at various distances from the transmitters have been reported in CCIR documents.

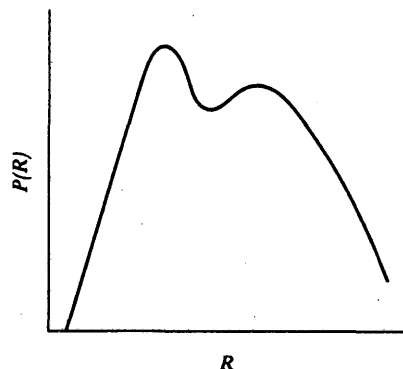


FIGURE 6

The fading records obtained at Barbados, Trinidad, Ghana, Singapore and Johannesburg by the Administration of the United Kingdom [CCIR 1959b; 1962b] agree fairly closely with the Rayleigh type of distribution for ranges between 0 and 350 km. However, for ranges exceeding 350 km, considerable departure from a Rayleigh type of distribution has been observed.

Measurements on the fading of CW transmissions at oblique incidence on 4.7, 9 and 15 MHz bands and on pulsed transmissions on the equivalent vertical incidence frequency were carried out by All India Radio. The distributions observed from the analysis of the fading curves were found to be Rayleigh, normal or log-normal [CCIR, 1959a].

Apart from the usually observed distribution of fading signal i.e. Rayleigh, normal or log-normal, *m*-type probability distribution was also reported in India [Das Gupta and Vij, 1960; Khastagir *et al.*, 1968; Srivastava *et al.*, 1972].

3.2 Rapidity of fading

The rapidity or speed of fading can be characterized in different ways [Fürth and MacDonald, 1947; McNicol, 1949; Booker *et al.*, 1950; Rice, 1958]. The various forms, characterizing the rapidity of fading, are discussed below.

3.2.1 Analysis of rapidity of fading

An important statistical property of the instant-to-instant variation of amplitude is given by the autocorrelation function. The rapidity or the speed of fading can be described in terms of the time auto-correlation function of the amplitude or in terms of the power spectrum of fading, which is the Fourier transform of the auto-correlation function.

If $A(t)$ represents the amplitude as a function of time, such that the average value is constant over the interval t to $t + \tau$, then the auto-correlation function $\rho(\tau)$ is defined by:

$$\rho(\tau) = \frac{1}{\tau} \int_t^{t+\tau} A(t) \cdot A(t + \tau) \cdot dt \quad (9)$$

If f represents the fading frequency and $F(f)$ is the Fourier transform of $A(t)$, i.e.

$$F(f) = \frac{1}{2\pi} \int_{-\infty}^{\infty} A(t) \cdot \exp(-i\omega t) dt$$

then the power spectrum $W(f)$ of the fading signal is given by $W(f) = |F(f)|^2$ which is proportional to $\int \rho(\tau) \exp(-i\omega\tau) d\tau$, the Fourier transform of $\rho(\tau)$.

Hence the frequency spectrum of the fading signal may be obtained with the aid of the auto-correlation function. The width of the fading power spectrum is related to the speed of fading.

Assuming that fading is produced by diffractive scattering from a large number of ionospheric irregularities and the scattered wavelets, having random amplitudes and phases, combined at the ground to give a resultant signal at any instant; [Booker *et al.*, 1950] have shown that the auto-correlation function $\rho R(\tau)$ of the amplitude R of the received fluctuating signal is proportional to:

$$\exp [(-16\pi^2 v_0^2 \tau^2)/\lambda^2] \quad (10)$$

where v_0 is the r.m.s. velocity of the irregularities in the line of sight.

The auto-correlation function falls to e^{-1} after a time $\tau = \lambda/4\pi v_0$ which can conveniently be called "fading time", since it will, on the average, represent the time which must elapse before the received signal has altered appreciably.

For oblique incidence,

$$\rho R(\tau) = \text{const.} \exp [(-16\pi^2 v_0^2 \tau^2 \cos^2 \phi)/\lambda^2] \quad (11)$$

where ϕ is the angle of incidence of the radiowave to the ionosphere.

Whenever fading is random, the amplitude distribution is Rayleigh type and the auto-correlogram is a smooth curve having an exponential decrease. If however, the fading is not absolutely random, but has a tendency towards periodicity, the auto-correlogram will show a number of maxima and minima. The corresponding amplitude distribution is often a Gaussian curve [CCIR, 1959a].

Figs. 7(a) and 7(b) show the auto-correlograms for two types of fading curves.

Other forms used for characterizing the rapidity of fading do not represent as complete a description of fading as auto-correlation functions, but nevertheless they also provide useful information concerning fading rates, as discussed below.

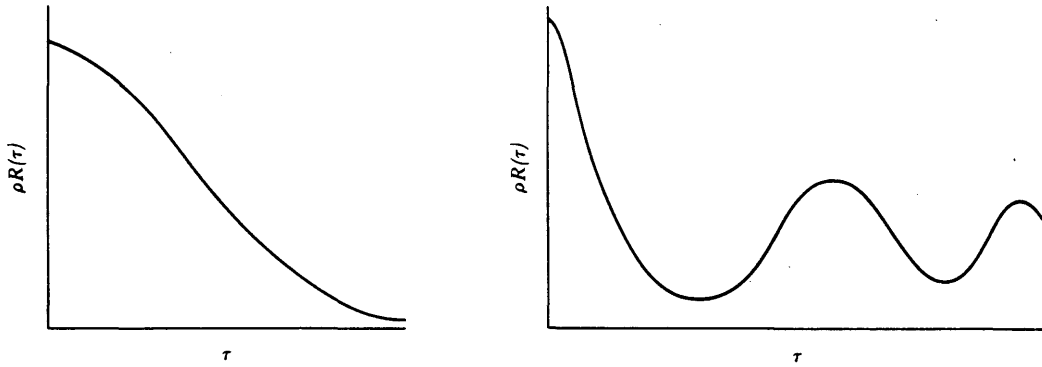


FIGURE 7

(a) Auto-correlogram for a Rayleigh type of fading

(b) Auto-correlogram for Gaussian type of fading

Following the work of [Fürth and MacDonald, 1947] on noise, speed of fading may be obtained by analysing successive differences of amplitudes of signals in the fading curve. They have shown from an extension of Rice's theory on random noise that, in the case of Rayleigh amplitude distribution, the probability distribution of v_τ , the successive difference between two amplitudes ($R_2 - R_1$) within a small interval of time τ , is given by:

$$\rho v_\tau dv_\tau = (1/\pi) \exp(-x^2) dx \quad (12)$$

where,

$$x = v_\tau / (2\pi\sigma\tau/\sqrt{2\psi})$$

and the power spectrum is Gaussian with standard deviation σ .

The power spectrum $W(f)$ is given by:

$$W(f) = (\psi/\sigma\sqrt{2\pi}) \exp[-(f - f_0)^2/2\sigma^2] \quad (13)$$

where $\psi = \int_0^\infty W(f) df$ is the total energy in the power spectrum.

A similar expression may be derived in the case of a randomly fading signal. In that case, f_0 is the incident wave frequency and f is the Doppler shifted frequency of the reflected wave from the moving irregularities.

The average value of $|v_\tau|$ is defined by:

$$|\bar{v}_\tau| = 2 \int_{-\infty}^{\infty} v_\tau \cdot \rho v_\tau \cdot dv_\tau$$

From (12) and (13) above, we get:

$$|\bar{v}_\tau| = 2 \sqrt{2\pi\psi} \cdot \sigma\tau$$

$$|\bar{v}_\tau/\tau| = \sqrt{2\pi\psi} \cdot 2\sigma$$

$|\bar{v}_\tau/\tau|$ is called the "speed of fading".

A simple form of defining fading rate is the number of positive crossings per unit time through any specified level. The number of maxima, N , of the amplitude of the received signal envelope per unit time may be used as a measure of the rapidity of fading. Rice has shown that if σ is the standard deviation of the power spectrum of the received signal, then $N = 2.52\sigma$.

The rapidity of fading may also be calculated from the derivative of the amplitude of the received signal and is given by $|d\bar{R}/dt|/\bar{R}$, where R is the instantaneous amplitude of the received signal.

3.2.2 Some measured values of fading rates

A few typical values of the fading rate measured in the Tropical Zone are given below:

In [CCIR, 1959a], India reported a fading rate of 11.9 fades per minute on an average, while the fading rate in general is observed to vary between 4 and 16 per minute.

Average fading frequency taken on 2.3 MHz in the E-region during daytime and night-time was found to be 10.96 cycles/min (0.183 Hz) and 14.002 cycles/min (0.233 Hz). In the case of F2 region, average fading frequency taken on 5.6 MHz during daytime and night-time was observed to be 11.22 cycles/min (0.187 Hz) and 8.37 cycles/min (0.14 Hz) respectively [Rao and Rao, 1967].

Increased rate of fading is observed almost throughout the Tropical Zone after sunset. The Administration of the United Kingdom reported this phenomenon through [CCIR, 1959b]. Typical value of the rapidity of fading is 10 Hz at 15 MHz [Davies and Barghausen, 1967]. Rapid fading rates up to 15 Hz were noticed by [McNamara, 1971].

The fading rate of signals reflected from the Es layer is found to be more than that from the normal E or F region. Rates of 0.2 to 5 Hz are typical [Rastogi *et al.*, 1966; Skinner and Wright, 1962].

[Skinner and Wright, 1962] found that, for the equatorial q-type Es, the fading rates are of the order of 4 Hz and the fading rate increases almost linearly with increasing wave frequency. For the blanketing type of Es, fading rates are much smaller than those of the q-type Es and are of the order of 0.8 Hz.

Observations in India at Thumba on the magnetic equator have shown that the speed of fading of the radiowaves reflected from the Es layer, has a strong diurnal variation, being very rapid in the early morning and evening hours and very slow at midday [Rastogi *et al.*, 1966].

4. Features of fading specially encountered in the Tropical Zone

Fading of signals in the tropical regions at low geomagnetic latitudes has some special characteristics due to:

- a particular type of sporadic E layer associated with the equatorial electrojet, which is a regular daytime occurrence;
- irregularities present in the night-time F layer, often mentioned as spread F.

4.1 Fading due to sporadic E

Fading observed during daytime in the equatorial zone is very often attributed to sporadic E. An analysis carried out by All India Radio [CCIR, 1962a] shows that sporadic E is one of the causes of the day-to-day fluctuations of signals.

4.1.1 The characteristics of sporadic E in the Tropical Zone

In the Tropical Zone, depending on the latitude, two types of sporadic E are usually prevalent:

- temperate zone sporadic E;
- equatorial zone sporadic E.

Temperate zone sporadic E observed at middle latitudes is predominantly a summer-time phenomenon occurring both by day and night and more intensely by day. In a narrow zone near the magnetic equator ($\pm 6^\circ$ magnetic dip), a special type of highly transparent sporadic E called equatorial sporadic E or Es-q appears regularly during daytime [Matsushita, 1962; Knecht and McDuffie, 1962]. It is observed that there is a narrower zone right at the dip equator, where the blanketing type of sporadic E is usually absent [Knecht and McDuffie, 1962; Cohen *et al.*, 1962].

But [Oyinloye, 1971] suggests that blanketing Es would sometimes occur over the dip and geomagnetic equator.

4.1.2 *Observed effects on fading due to sporadic E*

The fading characteristics of radiowaves reflected from Es layers studied by different workers indicate different results.

Some of the characteristics of echoes from the two principal daytime types of Es at Ibadan (Geomag. Lat. 10.6° N), Nigeria, have been investigated by means of vertical incidence pulsed transmissions in the frequency range 4 to 8 MHz. It has been observed that for the higher frequencies in the frequency range 4.3 to 8.5 MHz, the signal reflected from q-type Es is almost randomly phased with Rayleigh amplitude distribution, whereas at the lower frequencies near 4.5 MHz, amplitude distribution follows Rice function. Rayleigh type amplitude distribution is observed from the fading records at 5.7 and 7.5 MHz from the blanketing type of Es observed at midday at Ibadan [Skinner and Wright, 1962].

It is observed at Tirupati (Geomag. Lat. $3^{\circ}45'$ N) in India, that at times well before ground sunrise, the amplitude distribution of the echoes from sporadic E fitted most closely with the Rayleigh function, whereas just before ground sunrise and during the day, the amplitude distribution followed Gaussian or Nakagami-Rice functions [Venkateswarlu and Ramu, 1964]. However, observation at Thumba, situated at the magnetic equator, has shown that the probability distribution of the amplitude of signal strength does not fit with the theoretical Rayleigh or Gaussian distribution [Rastogi *et al.*, 1966]. Analysis of fading records at Calcutta (Geomag. Lat. 12.3° N) shows that the statistical distributions of amplitude are both Rayleigh and Nakagami-Rice types at noon, whereas during morning, evening and early night, other types of amplitude distribution – very often double-peaked or m-type have been observed [Ganguly and Samanta, 1967].

§ 3.2.2 may be referred to for the fading rates on high frequency signals reflected from the Es region.

4.2 *Fading due to F-region irregularities (flutter fading)*

A peculiar type of very rapid fading has been observed in the equatorial region after sunset. This is called flutter fading, the cause of which is attributed to F-region irregularities commonly known as spread F.

4.2.1 *Characteristics of F-region irregularities of spread F in the Tropical Zone*

In the equatorial zone after sunset, irregularities develop in the F-region ionization inside a belt extending approximately between geomagnetic latitudes of 30° N and 30° S. Its occurrence is high within $\pm 20^{\circ}$ geomagnetic latitude. It is seen that just after local sunset the F-region over the equatorial belt exhibits a marked increase in height and seems to break up progressively into patchy regions or irregularities in patches, generally known as spread F [Lyon *et al.*, 1961]. The characteristics of equatorial spread F have been studied by many workers in the past few years [Osborne, 1952; Bhargava, 1958; Shimazaki, 1959; Lyon *et al.*, 1960; Rangaswamy and Kapasi, 1963; Krishnamurthy, 1966; Rao, B.C.N., 1966; King, 1970; Davies, 1972; Bowman, 1975; Röttger, 1976]. The main features are as follows:

The presence of the irregularities in the F region is manifested by echoes whose range has been spread in the ionograms and they are classified as either range spread or frequency spread echoes. The range spread echo which is due to equatorial type spread F observed after sunset is seen to influence HF propagation. The irregularities are known to be magnetically field-aligned [Cohen and Bowles, 1961]. Recent observations have shown that both large-scale, non-field-aligned irregularities and small-scale field-aligned irregularities coexist in the F layer, when spread F is observed in the ionograms at equatorial latitudes [Bates, 1971; Cole and McNamara, 1975; Röttger, 1976].

The diurnal variation of range spread F is controlled by the sunset and the range spread F irregularities are seen normally after sunset and before midnight.

The seasonal variations show maxima of range spread F at the equinoxes. A longitudinal anomaly is seen in this case. Significant differences exist between the Far Eastern and American Equatorial region in the seasonal behaviour of spread F [Singleton, 1960].

The range spread F occurrence is positively correlated with sunspot number and indicates a negative correlation with magnetic activity [Huang, 1970].

Various workers have observed the same characteristics of range spread F as for equatorial flutter fading. Their observations are discussed in the next section.

4.2.2 *Experimental observations on flutter fading due to spread F in the Tropical Zone*

Flutter fading has been observed in India on All India Radio shortwave regional broadcast transmissions, ever since they were introduced in 1938. It was found that while flutter fading was very intense and annoying on shortwave transmissions from AIR Madras (Geomag. Lat. $3^{\circ}05' \text{ N}$) and Radio Ceylon, (Geomag. lat. $3^{\circ}38' \text{ S}$), it was considerably less on SW transmissions from AIR Bombay (Geomag. lat. $10^{\circ}00' \text{ N}$) and almost absent on transmissions from AIR Delhi (Geomag. lat. $19^{\circ}11' \text{ N}$). This indicates that flutter occurs nearer the geomagnetic equator. [Subba Rao and Somayajulu, 1949] called attention to the occurrence of equatorial flutter fading while studying field-strength variations of broadcasting stations of AIR in the 41 and 60 metre bands at Waltair in India. [Osborne, 1952] mentioned intolerable fading of ionospheric reflections at nearly vertical incidence associated with HF broadcast transmissions near the magnetic equator. [Yeh and Villard, 1958] have also described a high frequency fluctuation (up to 40 Hz) of signals observed at Stanford on BBC transmissions in the 9 and 21 MHz bands from Singapore, propagated on paths crossing the equator. They termed this phenomenon as "Doppler Fading". During the month of August 1957 they noticed this fading almost daily in the case of Singapore, but only sporadically from the transmissions in the same frequency range from Australia to Peking and to Buenos Aires. [Humby, 1959] reported similar behaviour of transequatorial Admiralty circuits from England to Colombo and England to Singapore during the period 1947 to 1958. [Chaman Lal, 1960] also reported rapid type of fading which is seen to be particularly pronounced in equatorial Southern India at Madras, Tiruchirapalli and Trivandrum. A rather similar observation was published by [Bennington, 1960] on the reception of United Kingdom transmissions in Johannesburg (South Africa) and Singapore during the period 1954 to 1958. Sunset fading effect on flutter fading has been investigated in Ghana since 1958. [Koster, 1963] gave a detailed analysis of the flutter fading observed on BBC (15.07 MHz) broadcasts to West Africa and Accra and on a broadcast signal locally originating from Accra on 5 MHz at Kumasi (270 km north-west of Accra). [Davies and Barghausen, 1967] recorded the effects of equatorial spread F on 5 paths in Africa near the magnetic equator, and observed flutter fading on the equatorial path. [McNamara, 1971] observed the evening type transequatorial propagation with deep flutter fading of the order of 5 to 15 Hz on the circuits between Okinawa and Yamagawa (Japan) and Townsville (Australia). The results of a study of night-time records of the field strength of radio signals on 11.8 MHz transmitted from Colombo and received at Ahmedabad indicate the presence of flutter fading [Chippa and Patel, 1973]. [Cole and McNamara, 1975] enumerated the effect of spread F on equatorial propagation. [Röttger, 1976] also observed flutter fading on transequatorial circuits between Lindau (Federal Republic of Germany) and Tsumeb (Republic of South Africa).

4.2.3 *Characteristics of flutter fading*

Based on the above observations, the characteristics of flutter fading are summarized below.

4.2.3.1 *Geographical extent*

Equatorial flutter fading appears to be most serious within a geographical area between $\pm 20^{\circ}$ geomagnetic latitude. Both the north-south circuits and east-west circuits are affected by this flutter fading.

4.2.3.2 *The diurnal variation of flutter fading*

Flutter fading is normally observed only between evening and midnight, starting almost always around sunset and continuing for three to four hours thereafter. It is found to be most intense within two hours of local sunset at the point where the propagation path crosses the magnetic equator. The time of start is quite well-defined and sudden, but the time of disappearance is rather gradual. The fades are very deep and the signal is almost completely drowned into noise, even though the mean signal strength remains high [Koster, 1963].

4.2.3.3 *Seasonal variation of flutter fading*

Seasonally, the phenomenon of post sunset flutter fading has been observed to be more pronounced during equinoctial months as compared to the other seasons of the year. It has been noted that definite maximum occurs during the two equinoctial periods well after the true equinox. The autumnal equinox shows significantly more flutter fading than the vernal equinox. The comparatively rapid falling-off of the phenomenon as the sunspot cycle approaches minimum is also observed [Koster, 1963].

4.2.3.4 *Solar-cycle-dependence of flutter fading*

Observations of [Osborne, 1951, 1952; Humby, 1959; Koster, 1963] tend to confirm that there is a positive correlation between mean sunspot number and amount of flutter fading. Equatorial flutter fading is most serious during sunspot maximum and becomes almost negligible during periods of sunspot minimum.

4.2.3.5 *Fading rates at various frequencies*

The observed fading rates are different for different wave frequencies. [Koster, 1963] analyzed fading on signals received in Ghana from London at frequencies approximately 7, 12, 15 and 21 MHz and concluded that the flutter fading rate was directly proportional to the wave-frequency. [Chaman Lal, 1960] also remarked that flutter increased with increase in operational frequency. Observations by [Cohen and Bowles, 1961] had shown that the effects of spread F on signals over transequatorial paths varied with frequency. [Davies and Barghausen, 1967] observed that the fading rate varied over a wide range, but a value of 10 fades per second was typical at a frequency of 15 MHz. A comparison of fading rates of signals at 10.1 and 20.2 MHz, observed over the same path from Monrovia to Accra, showed that the fading rates at 10 MHz were five to seven times faster than those at 20 MHz. This is, however, contrary to the observations of Koster and Chaman Lal, as mentioned above. [McNamara, 1971] observed the evening type transequatorial propagation (TEP) which had a high correlation with spread F with fading rates up to 15 Hz. Fading rates of transequatorial HF signals have been evaluated from observations on the paths Lindau to Rome and Tsumeb to Lindau which cross the equatorial zone of Africa [Carman *et al.*, 1974] and have been found to vary between 50 and 180 maxima per minute.

4.2.3.6 *Doppler fading*

Doppler fading of signals propagated on paths crossing the Equator noticed by [Yeh and Villard, 1958], has been observed by others also. [Calvert *et al.*, 1962], while measuring Doppler shifts of signals propagated from Tripoli to Accra, found that in the evening hours the spread in Doppler spectrum was of the order of 10 to 15 Hz. The systematic movement of a group of irregularities results in Doppler shift of the transmitted frequency up to 30 Hz at 20 MHz, while random motion of individual irregularities leads to Doppler spreading up to about 20 Hz [Bradley *et al.*, 1972]. [McNamara, 1975] also observed the evening type TEP with Doppler shift that sometimes exceed 40 Hz.

4.2.3.7 *Flutter fading and magnetic activity*

Flutter fading correlates negatively with magnetic activity [Wright *et al.*, 1956; Shimazaki, 1959; Briggs, 1965; Davies and Barghausen, 1966]. On quiet days flutter fading is noticed appreciably, whereas on magnetically disturbed days, flutter fading is absent [Chippa and Patel, 1973].

4.2.3.8 *Flutter fading and radio star scintillation*

In the equatorial region, observations of scintillation have been made at Accra in Ghana, at Ibadan in Nigeria and at Kodaikanal in South India. The temporal, seasonal, solar cycle and geomagnetic activity variations of scintillation have been found to be similar to spread F and flutter fading.

Wright *et al.* [1956] pointed out the high correlation between the occurrence of flutter fading and radiostar scintillation. This was further confirmed by [Koster, 1963]. Radiostar scintillation also correlates negatively with magnetic activity.

[Kent, 1961] who conducted scintillation observations utilizing satellite signals, obtained similar results on scintillation correlation with spread F. [Bhargava, 1964] at Kodaikanal also found the same diurnal and solar cycle variation of scintillation but the maximum rate of scintillation was found to occur several hours before the maximum of spread F.

4.2.4 *Effect of flutter fading on broadcasting*

From the standpoint of degradation in communication circuits, flutter fading is one of the most important factors. It affects the overall merit of broadcast programmes to a considerable extent. Bennington [1960] indicates that, in the audio spectrum, voice is still somewhat intelligible in presence of flutter, whereas music becomes very disagreeable and jarring to listeners.

4.2.5 Other types of fading observed in the Tropical Zone

Another peculiar type of fading has been observed in India on the shortwave broadcast transmission of All India Radio. This type of fading, which is characterized by sudden and violent fluctuations of the received signal intensity, has been designated as "surge fading". As compared with flutter fading, "surge fading" is slower but a deeper form of fading accompanied by severe distortion. This peculiar type of fading gives the impression of the signal being received in powerful surges.

4.2.5.1 Characteristics of surge fading

Surge fading is experienced at its worst after sunset. Seasonally, surge fading has been noted to be more pronounced during winter and the equinoctial months preceding winter. With surge fading, the signal varies over a wide range of amplitude with a recurrence rate of a few surges per minute. Surge fading may be due to phase interference or polarization changes of the received signal. The cause for this type of occurrence is not clearly known.

5. Fading allowances to be provided for planning broadcasting services in the Tropical Zone

Quantitative estimate of the fading range may be obtained from the statistical distribution of the signal, which can be characterized by indicating the levels exceeded for a number of specified time-percentages. The following Table I gives the levels for two time percentages derived theoretically from the distributions already discussed [Norton *et al.*, 1955].

TABLE I

Distribution	Level (dB) relative to median, exceeded for the following percentages of time	
	10%	90%
Rayleigh	5.21	- 8.18
Normal and log-normal (where σ is the standard deviation)	1.282 σ	- 1.282 σ

[CCIR, 1962a] serves as a preliminary basis for suggesting the values of fading allowances to be made for planning shortwave broadcasting services in the Tropical Zone. This document deals with two basic parameters associated with fading: the intensity fluctuation factor (IFF) and fading safety factor (FFS).

The intensity fluctuation factor, defined as the ratio (in dB) between the monthly median and the monthly lower decile values of the hourly-median field intensity, is considered to make allowance for day-to-day variations for 90% of the days. It is found that the values of IFF fluctuate widely between 1.0 dB and 15.4 dB, the overall average being 7.6 dB. The fading safety factor, defined as the ratio (in dB) between the median intensity of a received signal and the intensity exceeded 90% of the time, is estimated to make an allowance for instant-to-instant variation of the signal for 90% of time. The fading safety factor in summer varies from 2.4 to 27.0 dB, the most probable value being 11.2 dB and the corresponding figures for winter months are 1.6 to 14.4 and 7.0 dB respectively. The overall FSF is found to vary from 1.6 dB to 27.0 dB, the most probable value being 9.0 dB. The fading safety factor, calculated theoretically from the Rayleigh type of distribution which is predominant on short-term fading in the Tropical Zone, is 8.18 dB.

The overall fading allowance for planning may be obtained from the short-term and long-term fading allowance. Intensity fluctuation factor and fading safety factor, when referred to the received signal, take into account the variations of the wanted signal only. However, to arrive at the appropriate value of the fading

allowance for planning any sound broadcasting service for a given service probability, it is important to take into account the character and variability of the interfering signal also. Considering this aspect, Recommendation 411 proposed certain values of fading allowance under three conditions of reception, as follows:

- radio frequency signal-to-interference ratio: 16 dB;
- wanted signal-to-atmospheric noise ratio: 17 dB;
- wanted signal-to-industrial noise ratio: 12 dB.

The above-recommended overall variability allowance must be made to ensure that the steady-state ratio is attained for 90% of the time.

There has not been any further documentation on this subject since the CCIR Recommendation 411 was adopted. However, it would appear that the due verification of this Recommendation for the Tropical Zone can be made only after further studies, especially, because it has been suggested that fading, referred to as instant-to-instant variations of field intensity, is more severe in the Tropical Zone than in the countries of the temperate zone [CCIR, 1962a].

Investigations in the USSR [CCIR, 1986-90] have shown that selective fading caused by multipath propagation of HF signals may give rise to distortions in the amplitude-frequency characteristic of the transmitted frequency spectrum. Additionally, with an envelope detector the occurrence of non-linear signal distortions is increased, and a deterioration in intelligibility occurs. The frequency at which the distortions occur and their duration are determined by the fading characteristics of the signal and their intensity is proportional to the signal modulation coefficient. Non-linear distortion may cause the level of the second harmonic of the signal to reach -12 dB with respect to the level of the sideband frequency.

Similar signal distortions may be observed in the case of the synchronized operation of two transmitters serving the same reception area even where no multipath propagation occurs.

6. Conclusion

In this Report, the nature, type and characteristics of fading encountered in the Tropical Zone have been discussed. Some of the special features of fading in the Tropical Zone for example flutter fading, surge fading and fading due to equatorial sporadic E, have also been described.

The fading records analysed by various workers in different parts of the Tropical Zone show different types of distribution such as, Rayleigh, normal, log-normal and *m*-type. An idea of the allowance for short-term and long-term fading may be obtained from the distribution curves.

Propagation in the equatorial belt is highly influenced by q-type sporadic E during daytime. The fading rate of the radiowave reflected from the Es layer is observed to be more than in case of normal E or F region.

Flutter fading is a typical evening time phenomenon and depends upon the geographical area, local time, season, sunspot number and magnetic activity. Equatorial flutter causes circuit degradation and affects the music transmission appreciably. Presence of irregularities in the lower part of F region appears to be the cause of flutter fading. Hence a shift to higher frequencies is recommended wherever possible in the HF band; at least to alleviate the deterioration of circuit performance in the presence of spread F. [Davies, 1972] provided a scheme of frequency usage, designed to minimize as far as possible, the adverse propagation effects of spread F.

"Surge" fading occurring in this zone reduces the satisfaction value of the programme considerably. This type of fading is slower but deeper in form accompanied by severe distortion. The causes that lead to surge fading are not yet firmly established. Since the information available at present is too meagre, further studies on surge fading appear to be necessary.

The fluctuations in the field strength which occur primarily in the HF broadcasting range due to interference and variations of absorption and polarization impair reception quality. These fading effects may be minimized by use of diversity systems. Depending upon the mechanism of propagation various diversity systems such as space diversity, frequency diversity and polarization diversity are in use. It may be mentioned that the use of space diversity or polarization diversity does not help much in reducing the degrading effects in reception when the fluctuation is very rapid (e.g. in "flutter" and "surge" fading). However, single-sideband reception may be used in minimizing the degrading effects of selective fading.

Sufficient data relating to the quantitative estimates of short-term and long-term fading in the Tropical Zone is not available. Further studies will have to be continued to arrive at specific values of fading allowances to be provided for planning sound broadcasting service. It would also be worthwhile to have detailed analysis of the fading characteristics on the types of fading especially those encountered in the Tropical Zone, and to make a study with a view to correlating the fading character and the fading allowance which should be provided to minimize annoyance to satisfactory listening.

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REPORT 472-2

**SINGLE-SIDEBAND RECEPTION FOR RE-BROADCASTING APPLICATIONS
WITHIN THE TROPICAL ZONE****Reduction of fading effects**

(Question 45/10, Study Programme 45C/10)

(1970-1978-1990)

1. Introduction

One of the most economical methods of re-broadcasting is the use of sky-wave signal from one broadcasting transmitter by another. In view of the simplicity and economy, it is of special importance to developing countries. The efficacy of this method, however, depends to a large extent on the quality of reception of the sky-wave signal at the re-transmission site. The quality is often impaired due to selective fading and "surge" and "flutter" fading peculiar to the Tropical Zone. In order to improve the quality of sky-wave reception by minimizing the adverse effects of the above types of fading, use of single-sideband reception may be considered as one of the effective means for re-broadcasting application.

2. Extent to which the use of single-sideband reception can improve the quality of HF broadcasting

2.1 Single-sideband (SSB) reception of a conventional double-sideband (DSB) amplitude-modulation emission has certain advantages, the principal ones being:

2.1.1 it is possible to select either of the sidebands of the received broadcasting signal, thereby rejecting the sideband most affected by adjacent channel interference;

2.1.2 the demodulation process is more linear at high levels of modulation and does not produce the objectionable audio distortion of an envelope detector, associated with selective fading of the carrier;

2.1.3 this system of reception has been reported to minimize other propagation impairments (e.g. flutter and surge fading) as discussed below. Preliminary studies reported in the CCIR texts have indicated the possibility of reducing the effects of selective, surge and flutter fading by the use of single-sideband reception technique.

2.2 The deterioration of the quality of sky-wave reception due to selective fading results from the non-linear harmonic distortion when envelope detector is adopted in DSB (AM) reception. However, in SSB system of reception, selective fading leads to only linear distortion which is less severe than non-linear distortion as in the DSB (AM) reception.

Studies in the Federal Republic of Germany [CCIR, 1966-69a] have led to the following broad conclusions:

For double-sideband AM emissions, selective fading near the carrier frequency causes an apparent increase in the depth of modulation. This gives rise to distortion as soon as the carrier voltage falls below the sum of sideband voltages. Besides, the amplitudes of the low frequency components of the modulating signal are attenuated — thereby affecting the dynamic range and the tonal quality of the programme. If selective fading occurs in one of the sidebands, an apparent phase modulation of the carrier occurs. This causes a frequency shift of the modulation. However, the change in the depth of modulation is not as pronounced as with selective fading of the carrier because of the retention of the second sideband.

For single-sideband emissions, the interfering effect of selective fading of the carrier is completely eliminated. Compared with the amplitude of the sideband, the amplitude of the carrier re-inserted in the receiver is very high and causes a reduction in the phase modulation of the vector sum of the carrier and sideband to a very low level. Consequently, the signal is not distorted by carrier fading.

Selective fading in the sideband changes the tonal quality of the signal. The effect is particularly noticeable when the low frequencies of a musical programme are affected by fading. This could be remedied by frequency diversity reception of the two sidebands. But this method fails if one sideband suffers interference from another transmitter. In this case, diversity reception is only possible when space or polarization diversity is applied to the undisturbed sideband but it requires more complex equipment. It may, however, be mentioned that the assessment of comparative merits of DSB (AM) and SSB reception by the Federal Republic of Germany was subjective. An experimental method for subjective comparisons developed in the Federal Republic of Germany is described in [CCIR, 1966-69a].

An HF-double-sideband emission (6 MHz band) was received on a receiver with 3 outputs:

- double-sideband;
- single-sideband;
- carrier signal strength measurement.

The signals from the first two outputs were recorded simultaneously on the two tracks of a tape-recorder while the signal from the third output was recorded on a paper chart-recorder. The counter on the tape-recorder was used to mark the paper-recording simultaneously. Switching from track 1 to track 2 and vice-versa of the tape-recorder allowed immediate comparison between the effects of fading on double-sideband and single-sideband reception. Different types of fading were recorded and could be compared with one another. Such comparative reproduction of characteristic passages indicated the improvement that could be obtained with SSB reception.

At the interim meeting of Study Group XII, Palma de Mallorca, 1968, the Federal Republic of Germany gave a demonstration using this experimental method. On the basis of this demonstration Study Group XII agreed unanimously that SSB reception led to a significant improvement in the quality of reception when selective fading occurred.

2.3 Some preliminary listening tests carried out in India [CCIR, 1966-69b] on spoken-word programmes have indicated that in the presence of surge and flutter fading, reception is worst while using DSB reception with automatic gain control (a.g.c.), but perceptible improvement is observed without a.g.c. When surge and flutter fading are severe, a noticeable improvement is obtained by the use of SSB reception even with a.g.c. The study concludes that reception on a single-sideband receiver with a.g.c. can apparently give acceptable quality of the spoken-word programmes when surge and flutter fading are present. In the experiment a commercially-built single-sideband adaptor was used, together with a standard HF communication receiver.

The single-sideband adaptor had the following main characteristics:

- sharply tuned bandpass filters to select either the upper or lower sideband of the emissions;
- carrier rejection filter (30 dB);
- stable, tunable local oscillator for carrier re-insertion;
- automatic frequency control of the local carrier re-insertion oscillator;
- oscilloscope indicator for correct tuning.

3. Preferred Characteristics of SSB receiving system.

The characteristics of the single side-band receiver would be important in determining the degree of improvement in quality of reception. An SSB receiver should have the following characteristics:

- The selectivity characteristics of the receiver must be sharp with an overall bandwidth (-3 dB) of 4 kHz, and a slope of attenuation of 35 dB/kHz,
- The receiver must be equipped with a synchronous demodulator and for carrier acquisition must use a method whereby a carrier is regenerated by means of a suitable control loop which phase locks the regenerated frequency to the incoming carrier.
- A built-in synthesizer for automatic tuning is essential in this type of receiver.
- The receiver must work equally well with both conventional DSB as well as SSB transmissions regardless of whether the carrier reduction is 6 dB or 12 dB relative to peak-envelope power.

4. Conclusion

Studies on single-sideband reception have indicated that there is a considerable improvement in the quality of reception in the presence of selective fading.

Preliminary studies [CCIR, 1966-69b] have also shown the possibility of reducing the effect of surge and flutter by single-sideband reception. But there is a need for further detailed and extensive tests, both subjective and objective, to assess the extent of improvement that could be achieved and the determination of the preferred characteristics of such a system. It may be worthwhile to find out, whether the improvement in reception is due to the reduction of the effects of selective fading which may be associated with "surge" and "flutter".

The introduction of a suitable assessment scale is necessary when observed effects are subjective and are not open to exact measurement.

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[1966-69]: a. XII/7 (Federal Republic of Germany); b. XII/5 (India).



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SECTION 10B: FREQUENCY-MODULATION SOUND BROADCASTING IN BANDS 8 (VHF) and 9 (UHF)

REPORT 1066-1

CONTROL OF MODULATION LEVEL IN FM SOUND BROADCASTING

(Study Programme 46B/10)

(1986-1990)

Study Programme 46B/10 draws attention to the possible over-deviation of FM sound transmitters caused by:

- occasional peaks in the sound-programme signals exceeding the permitted maximum level for the transmitters;
- applying pre-emphasis to the modulating signal when this contains high-audio-frequency components at high amplitude.

Attention is also drawn to the peak-programme meter (as specified in Report 292) for monitoring programme level.

Interim Working Party 10/6 studied the control of level for the international exchange of sound programmes, and during the course of this work obtained information on the control of modulation level for FM sound transmitters, and the following points of interest have been noted:

1. Several different types of programme level meter are in use (see Report 292) and although the peak-programme meter is generally recognized as the most suitable for monitoring level during programme transmission, there are many situations where such meters are not available.

Incremental-display peak-programme meters [IEC, 1976] are attracting interest.

2. Signal levels should be set in the distribution network and at the inputs to transmitters using tones, and all of the meters mentioned in Report 292 are suitable for measurements on test tones.

It should be noted that the prolonged application of tones at high level to the distribution network or transmitters may cause interference to other services, and should therefore be avoided. Signals for alignment and testing of sound-programme circuits have been considered by IWP 10/6, and some may be suitable for use with FM transmitters.

3. Some organizations use limiters to control modulation level and, in situations where pre-emphasis is used, a variable-emphasis limiter offers advantages over a conventional limiter. This type of limiter controls the peaks either by a conventional limiting action or by reducing the degree of emphasis, as appropriate. When set to control the occasional peaks which might otherwise over-deviate the transmitter, the variable-emphasis limiter provides the means for making good use of the transmitter dynamic range without introducing unacceptable impairment of the sound signal [CCIR, 1982-86a].

4. If the level of the sound-programme signal at the input to any type of limiter is set too high, the resultant compression of the signal can cause unacceptable impairment of the sound programme itself and an increase in the level of interference to other services sharing the radio spectrum, even though the limiter may be effective in preventing over-deviation of the transmitter.

5. Limiters may be used either at the inputs to the transmitters or at the input of a stable distribution network. In the latter case, there are the advantages that both the transmitters and the distribution network are protected against excessive signal levels, and that in general fewer limiters are required (with a consequent saving in cost).

It is generally undesirable to subject sound signals to more than one limiting process, as unacceptable impairment may result. The connection of limiters at a number of different points in a sound-programme circuit is therefore inadvisable.

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REPORT 947*

**RADIO-FREQUENCY PROTECTION RATIO REQUIRED BY FM SOUND
BROADCASTING IN THE BAND BETWEEN 87.5 MHz AND 108 MHz AGAINST
INTERFERENCE FROM D/SECAM TELEVISION TRANSMISSIONS**

(Question 46/10, Study Programme 46J/10)

(1982)

1. Introduction

In some areas of Europe the part below 100 MHz of the VHF band II is used for television broadcasting. For planning of sound broadcasting in that band it is, therefore, necessary to know to what extent the reception of sound broadcasting signals is interfered with by television transmissions.

2. Measurements

To study the radio-frequency protection ratios required by sound broadcasting reception against interference from D/SECAM television transmissions, measurements have been made in the Federal Republic of Germany and Finland.

2.1 In Finland five representative domestic sound broadcasting receivers were measured making use of a D/SECAM television signal generator as the unwanted transmitter. The measurements were performed in accordance with Recommendations 468 and 641.

The D/SECAM television signal generator used was modulated by a 13×17 grid electronic test pattern signal with 90% modulation depth. The sound carrier was at a level of -10 dB with respect to the vision carrier and was modulated by a 1000 Hz signal with ± 50 kHz deviation. The wanted reference signal was 1000 Hz with ± 75 kHz deviation. The input level was 66 dB(1 μ V) into 75 Ω . The signal-to-noise ratio required was 50 dB measured according to Recommendation 468.

In the range of wanted signal frequencies from 0.5 MHz to 4 MHz above the vision carrier, the required protection ratio values were found to be greatly dependent on picture content. The worst interference with varying degrees of frequency dependence was produced by test patterns. In this range, therefore, an envelope curve has been taken to represent the interference potential of commonly-used test pictures.

The results obtained on the basis of these measurements are given in Table I and Figure 1. The effect of offset frequency operation should be taken into account separately especially near the carriers.

The results of the laboratory measurements are for steady interference. According to Recommendation 412 these values may be reduced by 8 dB for tropospheric interference (protection 99% of the time).

2.2 In the Federal Republic of Germany the RF wanted-to-interfering signal ratios were measured using one re-broadcast receiver and one domestic sound broadcast receiver of medium quality which meets the RF protection ratio curves of Recommendation 412 fairly well.

The measurements were made using a PAL transmitter but with a vision-sound-carrier spacing of 6.5 MHz. The vision-to-sound-carrier ratio was chosen as 10 dB. Taking account of the greater video bandwidth of the D/SECAM standard the interference in the frequency range between 5 MHz and 6 MHz was extrapolated from the results at lower frequencies.

The frequency of the chrominance sub-carrier was at about 4.433 MHz according to PAL standard but when evaluating the results, the special conditions of the D/SECAM standard using two sub-carriers at frequencies 4.25 MHz and 4.4 MHz were taken into account.

The measurements were made according to Recommendation 641 using the D/SECAM television signals as the interfering signal. However, the interfering sound carrier was only modulated with a quasi-peak deviation of ± 24 kHz, thereby taking account of the smaller maximum deviation of ± 50 kHz for the sound in television systems. For the video signal, different test patterns were used.

* This Report should be brought to the attention of Study Group 11.

TABLE I – *Radio-frequency protection ratio required by frequency modulation sound broadcasting at frequencies between 87.5 MHz and 108 MHz against interference from D/SECAM television transmissions*
(Steady interference)

Wanted signal frequency relative to vision carrier (MHz)	RF-protection ratio (dB)	
	Mono	Stereo
-2.0	-30	-12
-1.0	-2	18
-0.5	0	20
-0.15	19	25
-0.1	24	35
-0.05	30	50
0.0	35	45
0.05	30	50
0.1	24	35
0.15	19	31
0.25	10	25
0.5	0	20
1.0	-1	20
2.0	-3	18
3.0	-4	17
4.0	-5	15
4.18	8	25
4.25	10	26
4.41	10	26
4.48	8	25
4.7	-5	15
5.0	-15	0
6.0	-25	-5
6.25	-13	-6
6.3	-5	5
6.4	6	26
6.45	15	40
6.475	25	43
6.5	28	35
6.525	25	43
6.55	15	40
6.6	6	26
6.7	-3	0
7.0	-30	-13

Note 1. – For tropospheric interference (protection 99% of the time) these values may be reduced by 8 dB.

Note 2. – Values for frequencies from 0.5 to 4 MHz are greatly affected by picture content. The figures given are for a test pattern and are representative of the on-the-air test picture transmissions.

3. Comparison of the results

The results obtained in the measurements effected in the Federal Republic of Germany generally confirm those given in Table I and Fig. 1. As a result of the different measuring methods and equipment used, the radio-frequency protection ratios differ below -1 MHz relative to the vision carrier and around the colour sub-carrier frequencies above 4 MHz. The results obtained in the Federal Republic of Germany require lower protection ratios below -1 MHz for stereophonic reception and above 4 MHz for monophonic reception.

4. Conclusions

The radio-frequency protection ratio values given are based on a preliminary series of measurements and slight modifications are possible when further measurements are carried out. The numerical values of Table I and Figure I could, however, be used provisionally until results based on other measurements are available.

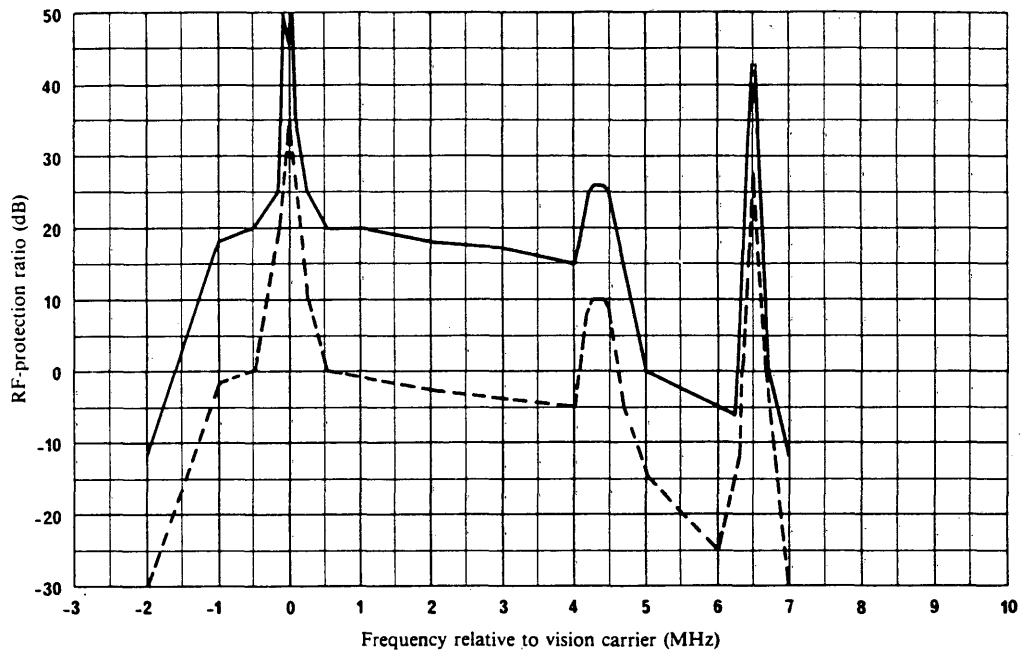


FIGURE 1 – Radio-frequency protection ratio required by frequency modulation sound-broadcasting at frequencies between 87.5 MHz and 108 MHz against interference from D/SECAM television transmissions

(Steady interference)

— : stereo
 - - - : mono

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[1978-82]: 10/236 (Finland); 10/235 (Poland (People's Republic of)); 10/239 (Germany (Federal Republic of)).

REPORT 1064-1

PARTICULAR INTERFERENCE CASES IN FM BROADCASTING

(Question 46/10, Study Programme 46N/10)

(1986-1990)

1. Interference caused by an overmodulated transmitter

[CCIR, 1982-86a] describes the results of laboratory measurements made to evaluate the sensitivity of several receivers to interference in the case where the interfering transmitter is over-modulated.

Interference was measured as described in Annex I to Recommendation 641, for a stereophonic signal and a wanted RF receiver input level of -50 dBm (0.7 mV/ $50\ \Omega$).

The -3 dB and -40 dB bandwidths of the RF filter added at the output of the interfering transmitter were respectively 500 kHz and 2600 kHz.

Two overmodulation values were used: $+3$ dB and $+6$ dB. It was found that, for interfering signals within the receiver passband, the increase of protection ratios was not related to the type of receiver; thus, for a 100 kHz carrier frequency spacing, the increases in protection ratio were 11 dB and 15 dB for increases in modulation depth of 3 dB and 6 dB respectively.

On the other hand, it was found that in the case of interference with a (non standard) 150 kHz carrier frequency spacing, the change in protection ratios could be as high as 6 dB for a 1 dB change in modulation depth of the interfering transmitter.

2. Interference for large carrier frequency differences

Document [CCIR 1982-86a] also describes tests to evaluate the effect of interference from transmissions having large frequency differences, which were carried out under similar conditions to those given in § 1 above.

In this case, measurements were made with normal modulation of the interfering transmitter and for carrier frequency spacing up to 1 MHz. The measurements showed that, beyond 400 kHz, there was no relation whatsoever between protection ratios and whether or not the unwanted transmitter was modulated.

With a professional receiver, the protection ratios decrease when a narrow-band RF filter (-40 dB bandwidth equal to 1200 kHz) is inserted at the output of the interfering transmitter. This shows that reception is disturbed only by the residual noise sidebands of the unwanted carrier.

On the other hand for the domestic receivers used, the protection ratios are almost constant from 400 kHz onwards, and have a value of around -40 dB which is practically independent of the type of filtering used on the interfering carrier. In this case, it is only the presence of the interfering carrier which impairs reception, with many possible causes of disturbance, such as desensitization of the input stage, local oscillator drive, etc.

3. Interference when the protection ratio is not respected

Document [CCIR, 1986-90a] describes the results of tests carried out on three receivers (professional, semi-professional, commercial) when the protection ratio is not respected.

Interference tests on all three receivers were carried out in monophony and stereophony, at a wanted receiver input RF level of -50 dBm (0.7 mV/50 ohms) and for positive frequency offsets. The test conditions described in Recommendation 641 were followed except as regards the wanted/unwanted AF signal ratios, which were taken as 50 dB (Recommendation 641 value), 40 dB and 30 dB.

Document [CCIR, 1986-90b] describes the results of similar measurements for 31 domestic receivers of different price categories (lower, medium and higher), however, for audio-frequency signal-to-interference ratios of 47 dB, 50 dB, 53 dB, 56 dB and 59 dB.

It was found that for a frequency difference up to (and including) 50 kHz in monophony and 100 kHz in stereophony, an increase of the interfering signal level leads to a similar reduction of the audio-frequency signal-to-noise ratio at the output of the receiver.

On the other hand, for a frequency difference larger than these values but smaller than 250 kHz approximately, a very small increase in RF interference can cause a considerable deterioration in reception quality, more pronounced in monophony than in stereophony. In such cases of offset, it is essential at the planning stage to allow a substantial margin for the uncertainties of propagation, multipath interference, obstacles, etc. Based on the results obtained, a margin of around 10 dB would not appear excessive. Considering the small number and the types of receivers tested additional studies should be carried out.

4. RF protection ratios for different wanted signal levels

Document [CCIR, 1986-90c] describes the results of measurements made to evaluate the influence of the wanted signal level on RF protection ratios. The RF protection ratios for 31 domestic and 16 car receivers of different price categories were measured with different wanted signal levels.

The measurements were performed according to CCIR Recommendation 641. Input levels for the wanted signal of 30 dB(pW), 40 dB(pW) and 50 dB(pW) were applied.

The mean value curves of measured RF protection ratios are shown in Figures 1 and 2. Each figure shows curves for stereophonic and monophonic reception. For comparison purposes the RF protection ratio curves for steady interference according to Recommendation 412-4 are also shown. Figure 1 presents curves for domestic receivers, and Figure 2 shows the comparable results for car receivers.

The figures show that the influence of the wanted signal level on measured RF protection ratios is not as large as expected, at least if only mean values and not single receivers are considered. The increase of the measured RF protection ratio is ≤ 5 dB for stereophonic reception with domestic receivers, if the wanted signal level is increased from 40 dB(pW) to 50 dB(pW). For car receivers this value is slightly above 5 dB. For monophonic reception the increase of the measured RF protection ratios above 300 kHz carrier frequency separation is somewhat higher than 5 dB (up to 9 dB). In this case, however, the wanted-to-interfering signal levels are already considerably below the RF protection ratios.

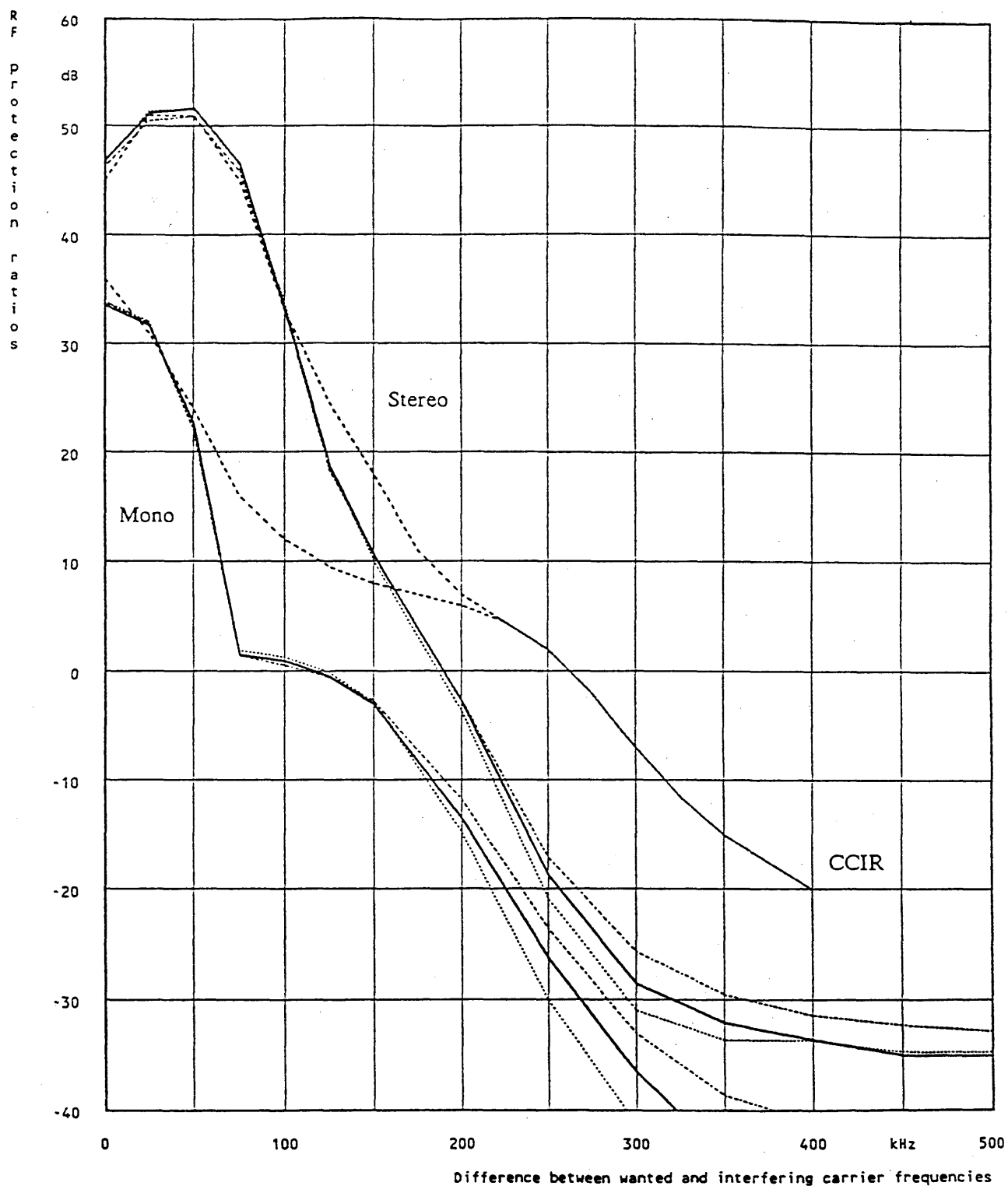
5. Interference caused by intermodulation of strong RF signals

Document [CCIR, 1986-90d] describes results of an investigation of domestic and car FM radio receivers on their tendency to intermodulate in the presence of strong signals. This receiver performance in the presence of strong RF signals is measured with three RF-signals and expressed as a protection ratio.

Thirty-one domestic and 16 car receivers of different price categories were measured. Two interfering signals of equal levels were positioned above or below the frequency of the wanted signal at equal differences of frequencies, i.e.

$$\begin{aligned}\Delta f &= f_w - f_{i2} = f_{i2} - f_{i1} \quad \text{or} \\ \Delta f &= f_{i2} - f_w = f_{i1} - f_{i2}\end{aligned}$$

The interfering signal f_{i2} was unmodulated, and the interfering signal f_{i1} was modulated with coloured noise according to Recommendation 641. The RF protection ratios were measured according to Recommendation 641, the only difference being that two interfering signals were applied as mentioned above. The mean values of the so-called strong signal protection ratios for stereophonic and monophonic reception with domestic and car receivers are presented in Figures 3 to 6. The standard deviation of the measured receivers ranged from 5 to 7 dB.



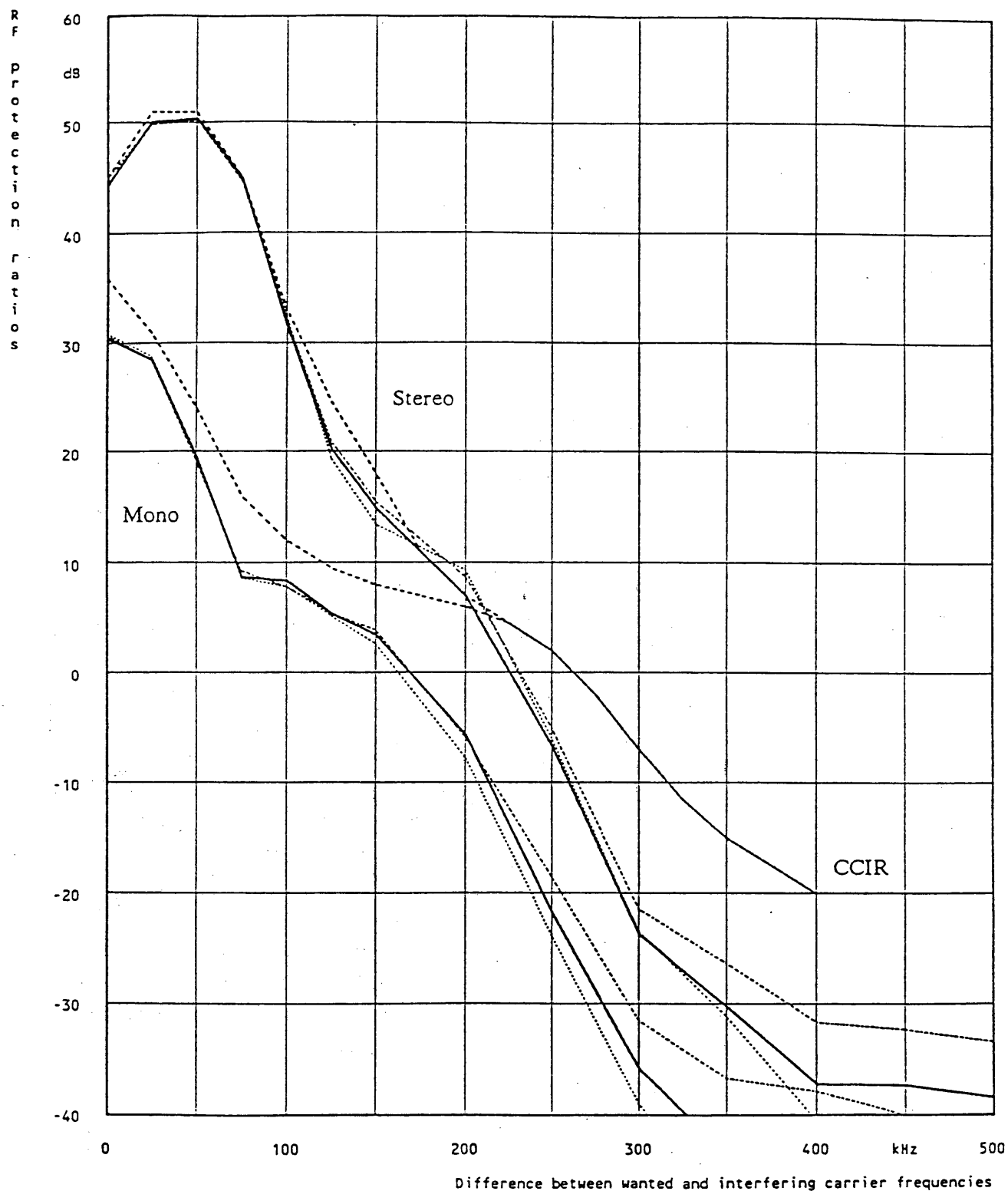
Difference between wanted and interfering carrier frequencies

curve	input power	number of receivers
.....	50 dB(pW)	31 stereo / 31 mono
—	40 dB(pW)	31 stereo / 31 mono
.....	30 dB(pW)	26 stereo / 31 mono

FIGURE 1

Radio-frequency protection ratios for different input powers

Domestic receivers



curve	input power	number of receivers
.....	50 dB(pW)	10 stereo / 16 mono
————	40 dB(pW)	10 stereo / 16 mono
-----	30 dB(pW)	8 stereo / 16 mono

FIGURE 2

Radio-frequency protection ratios for different input powers
Car receivers

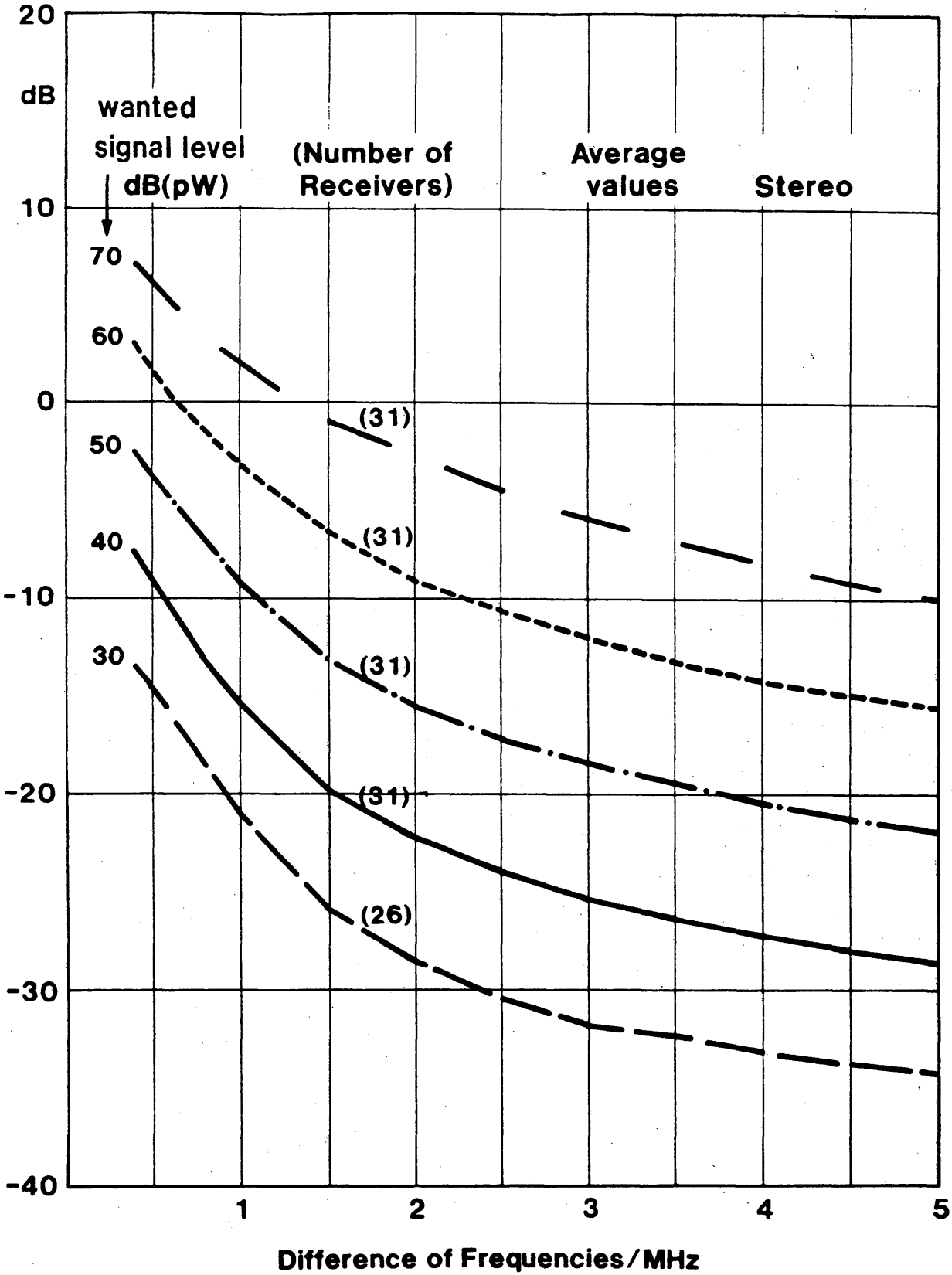


FIGURE 3

Strong signal protection ratios of domestic receivers
for different wanted signal levels - stereo

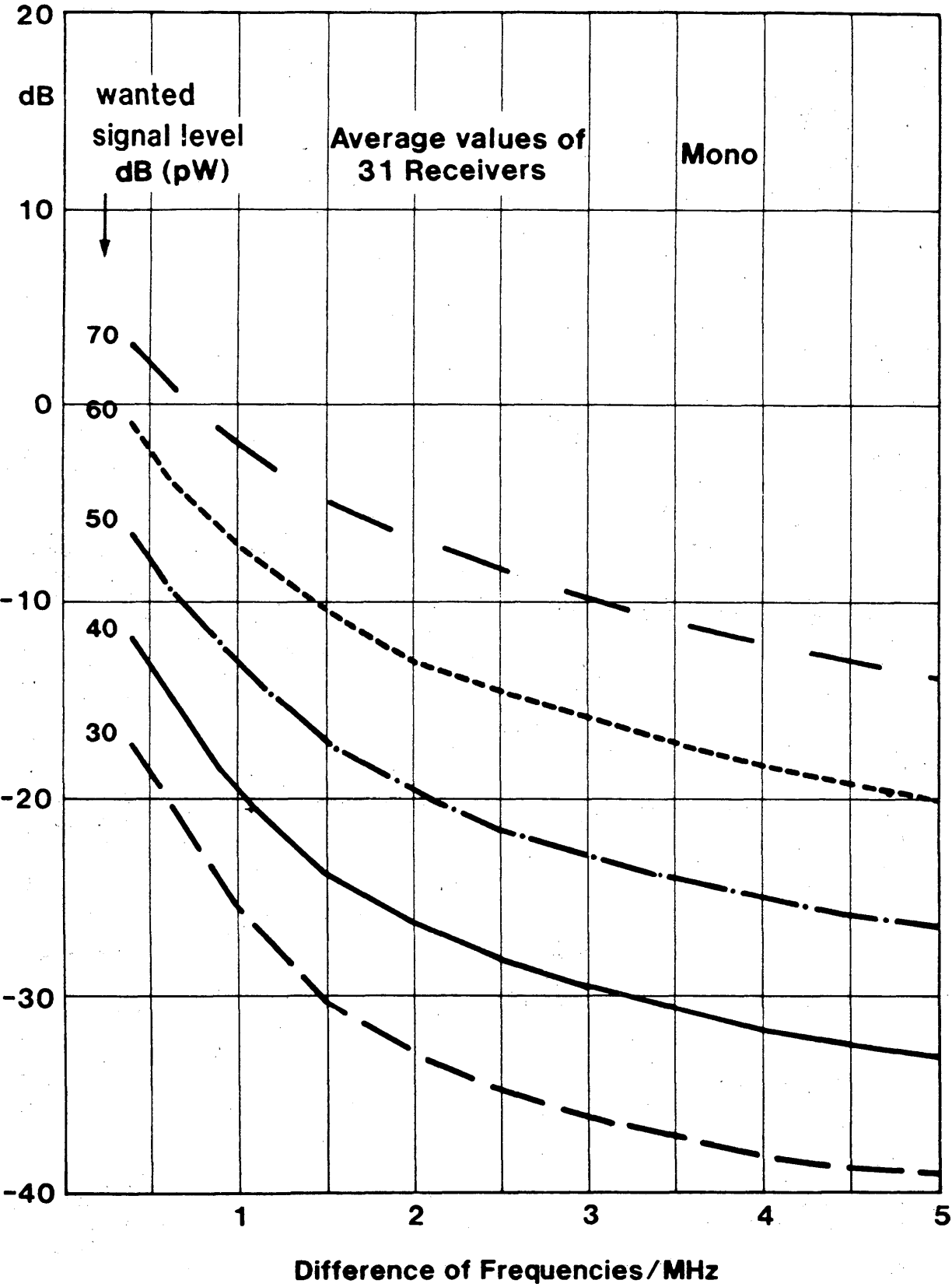


FIGURE 4

Strong signal protection ratios of domestic receivers
for different wanted signal levels - mono

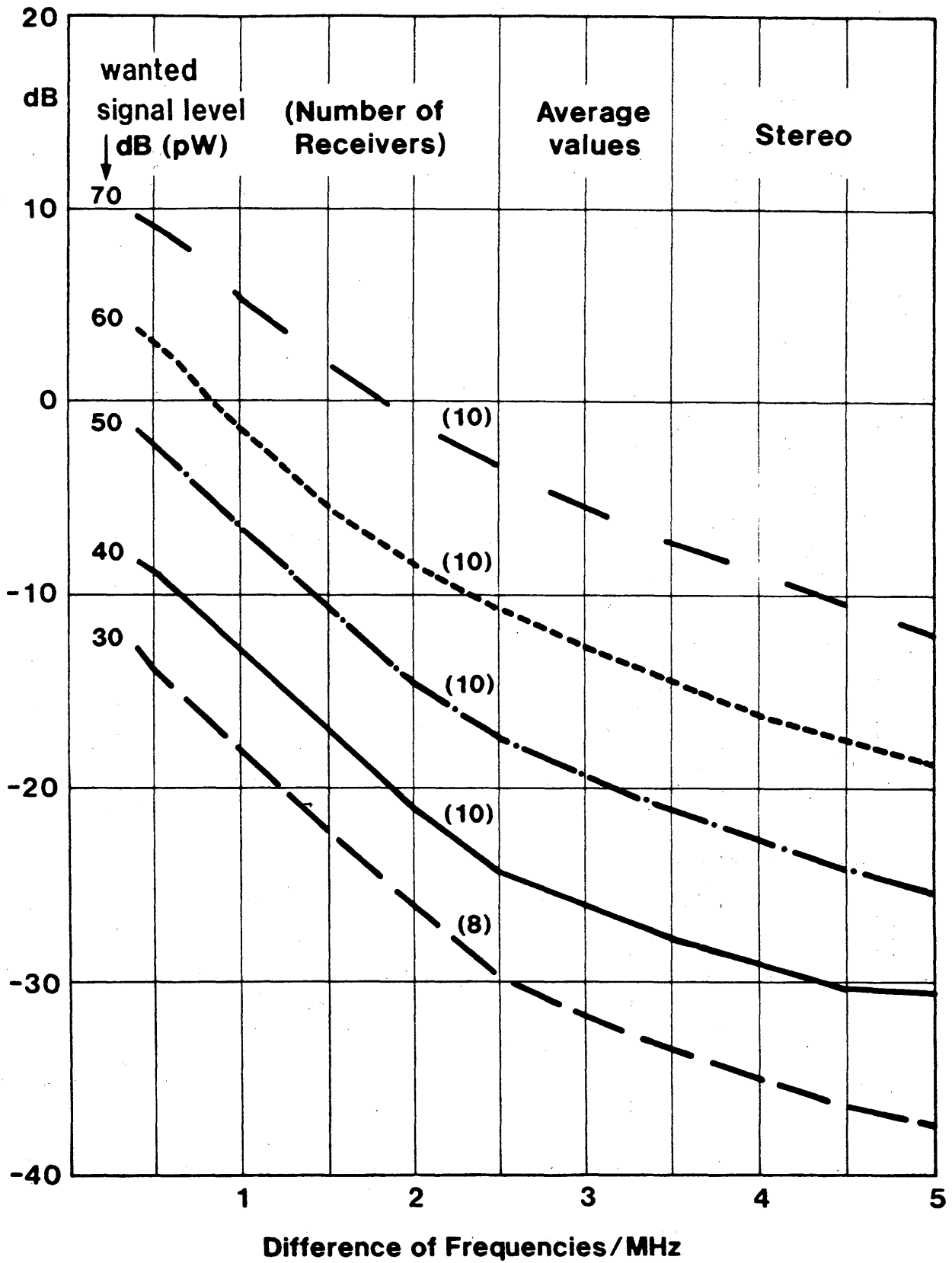


FIGURE 5

Strong signal protection ratios of car receivers
for different wanted signal levels - stereo

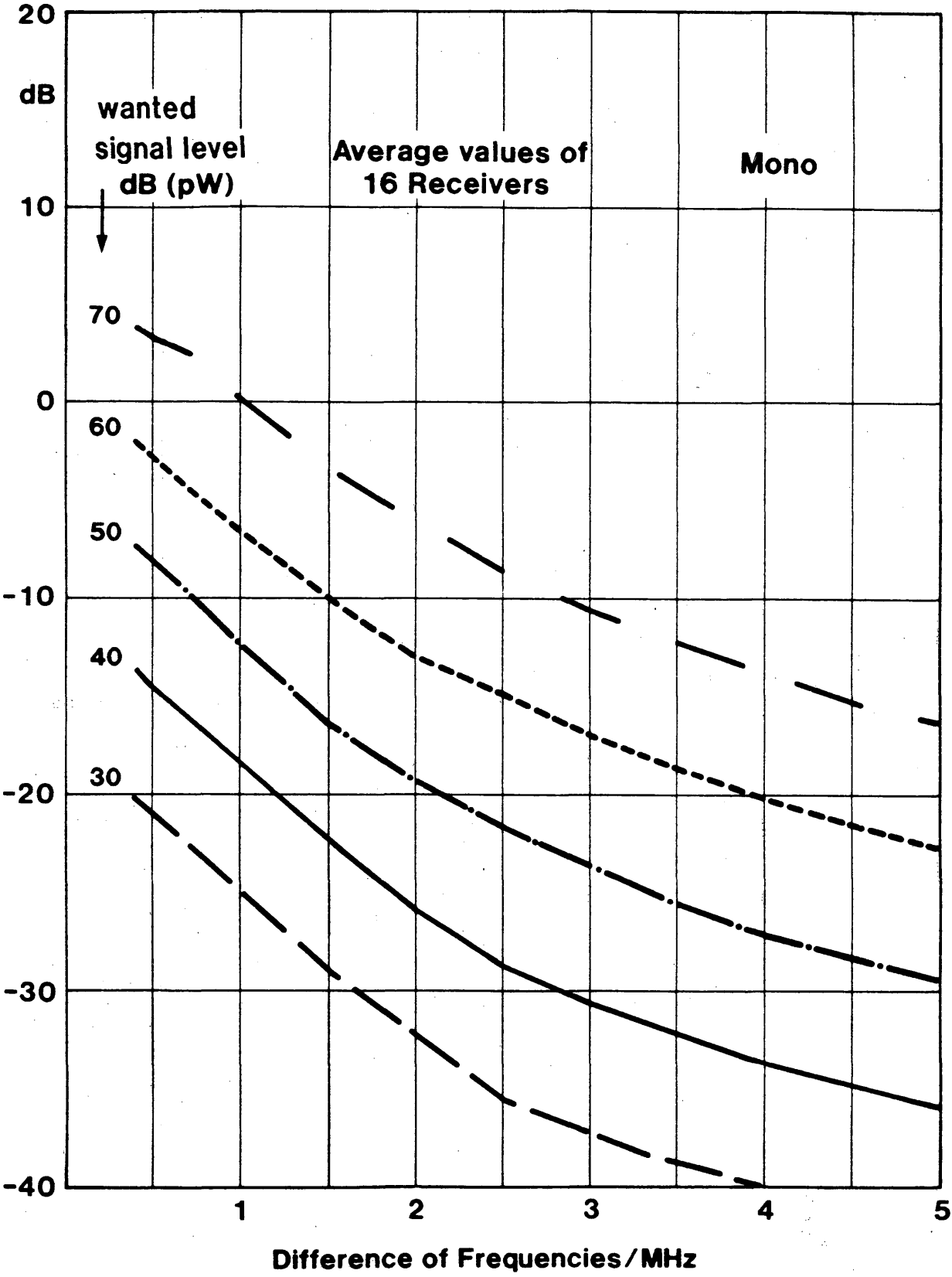


FIGURE 6

Strong signal protection ratios of car receivers
for different wanted signal levels - mono

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[1986-90]: a. 10/5 (France); b. 10/265 (Germany (Federal Republic of)); c. 10/264 (Germany (Federal Republic of)); d. 10/308 (Germany (Federal Republic of)).

REPORT 1202PROTECTION RATIO FOR FM SOUND BROADCASTING IN THE
CASE OF THE SAME PROGRAMME AND SYNCHRONIZED SIGNALS

(Study Programme 46L/10)

(1990)

1. Introduction

In the case of single frequency operation in which the interfering transmitter broadcasts the same programme as the wanted transmitter, the applicable protection ratio values can be expected to be lower than those given in Recommendation 412 for the general case. If in addition the wanted and interfering signals are identical in programme, frequency and modulation depth, apart from a slight difference of level and time delay, the interfering signal appears as an echo of the wanted signal, thus rendering the situation even better.

The reception quality is also influenced by the phase between the wanted and interfering signals.

Protection ratios were evaluated [CCIR, 1986-90a, b] in France and in Italy having regard to this particular case, in which synchronization of the interfering transmitter with the wanted transmitter guarantees that the instantaneous phases of both signals are identical.

2. Measuring conditions2.1 Arrangement

The tests were arranged in such a way as to simulate the reception of a signal formed by the combined field strengths of two synchronous transmitters. The parameters for consideration are:

- the level difference between the signals received from both transmitters;
- the time delay between the wanted and interfering signals;
- monophonic and stereophonic transmitter modes (the presence of an RDS sub-carrier was not considered);
- the phase shift between the signals received.

The measurements carried out in France and in Italy followed in principle the same philosophy. An RF sound signal modulated by a high quality audio frequency source (compact disk) was introduced in two separated channels: one of these channels was equipped in such a way as to permit the attenuation and the time delay of the signal to be varied, step by step, before recombining the two signals.

For the French tests a professional FM stereo receiver was used, and for the Italian tests a commercial good quality FM stereo receiver was used.

2.2 Evaluation

A subjective evaluation procedure was adopted because the interference revealed both noise and distortion; the CCIR 5-grade impairment scale was adopted.

2.2.1 For the measurements carried out in France, preliminary tests showed that speech was more critical than music for test purposes. Consequently, a "speech sample" was used as the test signal; the duration of each sample was 20 seconds, making it possible to evaluate the impairment of the worst phase configuration.

To permit practical exploration of all the phase configurations, it was decided to offset the wanted and interfering transmission frequencies by 0.1 Hz (phase shift of 360° in ten seconds).

The task of the five listeners participating in the tests was to give scores for the impairment in relation to the reference, of 30 samples corresponding to the 30 selected configurations resulting from the combination of operating modes (mono/stereo), time delay values (2, 5, 10 μ s) and the five level ratio values (chosen as a function of the other parameters).

The protection ratio values corresponding to both grade 3 and 4 of the impairment scale were determined from the obtained scores.

2.2.2 For the measurements made in Italy three types of programme modulation were employed: piano solo, violin solo and modern music; piano solo was the most critical, whilst the modern music was the most tolerant.

Four delay conditions have been examined; 5 μ s, 10 μ s, 20 μ s, 40 μ s. For each delay and for each type of modulation, the protection ratio corresponding to grade 4 on the impairment scale, was determined by four groups of expert listeners each consisting of 12 people.

In carrying out the subjective tests, the phase shifter was adjusted in each test to obtain the condition of maximum distortion.

It was also noted that the degradation results are proportional to the modulation depth, so that particular care was taken not to exceed the correct modulation limit.

3. Results and considerations

The tests carried out in France (mono and stereophonic mode) and in Italy (monophonic mode) have given results perfectly in agreement comparing those corresponding to the more critical types of modulation employed ("speech" and "piano solo" respectively). The results obtained during the two series of tests are summarized in the following table for both modes of operation (mono and stereo) and for the values of time delay considered:

TIME DELAY (μ s)	PROTECTION RATIOS (dB)			
	MONOPHONIC MODE		STEREOPHONIC MODE	
	I = 3	I = 4	I = 3	I = 4
2	< 1	1	4	6
5	1	2	10	12
10	1	3	14	16
20	NE	11	NE	NE
40	NE	20	NE	NE
NE = not evaluated I = impairment grade				

The figures in the table indicate the worst values found during the tests.

The results corresponding to monophonic mode can be considered reliable, whilst those corresponding to stereophonic mode are purely indicative, being based on a limited number of subjective evaluations; it may be noted that:

- in the most favourable case (monophonic mode, 2 μ s time delay) the protection ratio is very close to 0 dB using synchronized transmitters;
- the time delay between wanted and interfering signals creates transmission channel selectivity which worsens as the time delay increases and, of course, stereophonic signals are more sensitive to this;
- up to 5 μ s time delays, the protection ratio is independent of the modulation type; for increasing delay, the protection ratio becomes a function of the modulation type.

4. Application

In application of such a system in Italy [Angeli and Gunetti, 1989] along a mountainous section of one of the main motorways (section Bologna-Florence of about 85 km) a new FM monophonic synchronized broadcast service addressed to car-drivers was implemented.

The main purpose was to ensure good in-car reception without retuning along the complete section of the motorway including tunnels, in which a radiating RF cable was employed. It is foreseen that such a service will extend to most parts of the main Italian motorways.

5. Conclusion

On the basis of the collected data and of the results obtained, it is possible to plan a synchronized monophonic network for special applications with protection ratios of only 2 dB, provided that the relative time delay between the modulating signals is maintained within 5 μ s over the whole area to be served and that the maximum deviation does not exceed ± 75 kHz.

Thus in the case of co-channel interference, the protection ratio evaluations made for synchronized wanted and interfering transmitters broadcasting the same programme, give values very much below those indicated in Recommendation 412.

Further evaluations should be pursued for a larger number of configurations including transmissions with multiplexed data signals.

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REPORT 1065*

THE RF SPECTRUM OF FREQUENCY-MODULATION
SOUND-BROADCASTING TRANSMITTERS

(Question 46/10, Study Programme 46L/10)

(1986)

1. Introduction

The RF spectrum of frequency-modulation sound-broadcasting transmitters is infinite in theory. In practice, the need may arise to have a precise knowledge of either the necessary (No. 146 of the Radio Regulations) or the occupied bandwidth (No. 147 of the Radio Regulations). For example, only when the occupied bandwidth is known, is it possible to plan for acceptable service in adjacent bands.

Rules for the calculation of the necessary bandwidth of emission for numerous services are laid down in the Radio Regulations and examples are given in Appendix 6 to these Regulations. According to these rules the necessary bandwidth of frequency-modulation emissions is calculated by using Carson's rule:

$$B_n = 2f_{max} + 2DK$$

where:

B_n : necessary bandwidth,

f_{max} : highest modulation frequency,

D : maximum deviation of the RF carrier,

K : factor which is largely undefined and in the examples set to be equal to 1.

In frequency-modulation sound broadcasting the maximum deviation is $D = 75$ kHz, and the highest modulation frequency $f_{max} = 15$ kHz or $f_{max} = 53$ kHz for monophonic or stereophonic emissions, respectively. When supplementary programme or information signals are transmitted, the corresponding value may be as high as 76 kHz.

With these values and $K = 1$ the necessary bandwidths resulting from the application of Carson's rule would be:

$$B_n = 180 \text{ kHz}, B_n = 256 \text{ kHz}, \text{ or } B_n = 302 \text{ kHz, respectively.}$$

* This Report should be brought to the attention of Study Group 1.

Measurement results, particularly when obtained for stereophonic emissions, differ greatly from the calculated values. In the light of Report 418 it is rather doubtful whether Carson's rule may be used in bandwidth calculations for stereophonic emissions, since the pre-supposition that $D \gg f_{max}$ is not met, and since consequently, a value $K \neq 1$ may have to be used.

Difficulties are also experienced when calculating the occupied bandwidth and also the out-of-band emissions. For example, no reference to out-of-band emissions in FM sound broadcasting is contained in Recommendation 328 "Spectra and bandwidths of emissions", although out-of-band emissions are defined there. Even Reports 275 and 324, which are quoted in the Recommendation and which deal with measurements of bandwidths of emissions, do not include any useful information with respect to this question. Measurements of the frequency spectrum of transmitters which are frequency-modulated with white noise are described in Report 977. However, the results of these measurements do not permit conclusions to be made with regard to the radio-frequency spectrum of frequency modulation sound-broadcasting transmitters. It was with this in mind that the measurements described below were performed [CCIR, 1982-86a].

2. Simulated programme signal

With respect to the time variations, there exists a close interrelation between the RF spectrum of an FM transmitter and the modulating programme signal. Therefore, only statistical information can be given either on the average power-density distribution or on the percentage of time for which certain levels are exceeded.

Because of the close interrelation of the RF spectrum and the modulating AF signal, the first aim was to determine a representative programme signal. For various measurements, the coloured noise described in Recommendation 559 is used to simulate a programme signal. The spectral power density distribution of this coloured noise is in rather good agreement with that of a representative non-compressed broadcasting programme. Because of its low dynamic range it is preferred for certain measurements. In the case of spectrum measurements however, an AF modulating signal should be used which has dynamic characteristics similar to those of an average broadcasting programme. For this purpose, the amplitude distribution of a non-compressed stereophonic broadcasting programme was determined over a period of one week. The continuous line in Fig. 1 shows the cumulative time distribution of programme levels for this AF signal. Moreover, the measurements showed that the day-to-day variations were less than 2 dB.

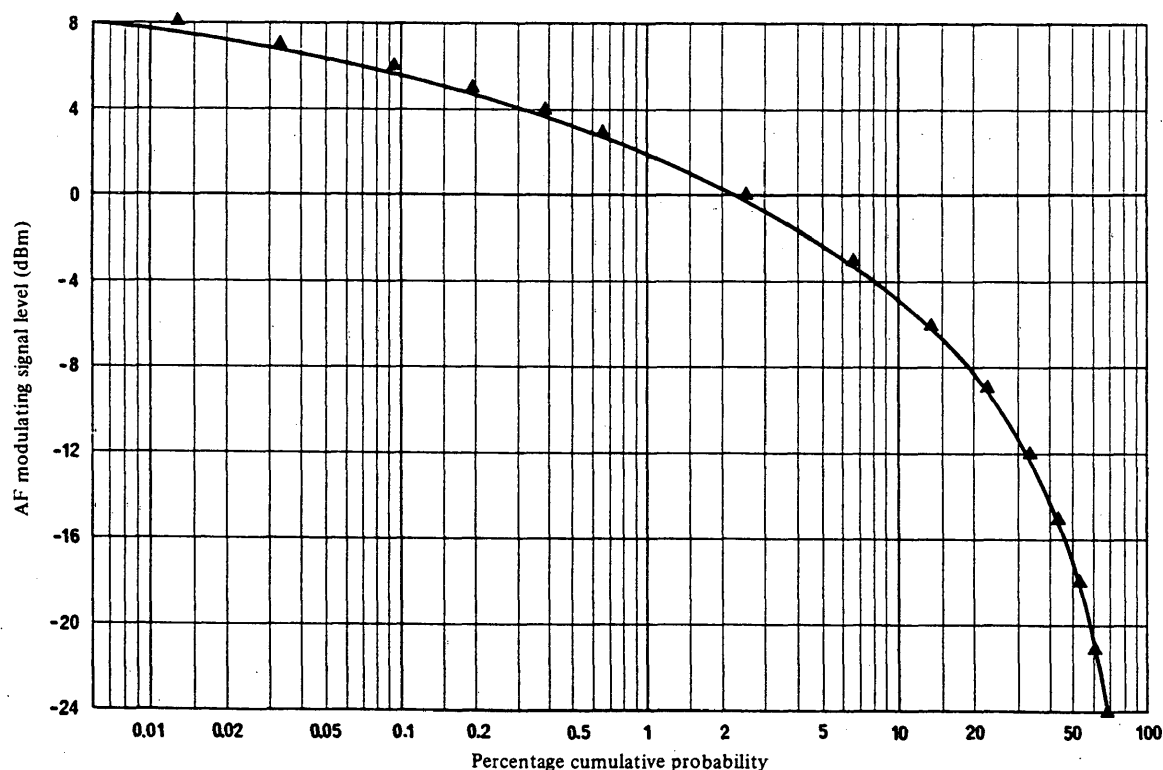


FIGURE 1 – Amplitude distribution of FM sound-broadcasting emissions

- ▲ : Distribution of the test-tape sequence of 60 programme samples (representative of a non-compressed high quality programme)
- : Distribution of a week's programme

To reduce the time required for the measurements, samples of 1 min duration were taken at equidistant time intervals. In this way a sequence of 60 programme samples was obtained which is representative of a 24 hours stereophonic programme. This can be seen from Fig. 1, where the data points indicate the cumulative distribution of the sequence of samples. The use of this sequence of samples as a representative substitution for a real programme is justified by the proximity of the data points to the curve in Fig. 1.

3. Measurements of spectra

Under certain assumptions the power-density spectrum of a time function can be measured with a spectrum analyser. As it was intended to determine not only the power-density distribution, but also spectra for different percentages of time, the spectrum analyser was only used as a filter having a noise bandwidth of 1.2 kHz and as a logarithmic demodulator. The analyser was tuned to various frequencies having different frequency separations from the RF carrier. By means of a stochastic counter connected to the output of the analyser, the percentage of time was determined during which various levels were exceeded. For this purpose, an RF signal generator whose output was directly connected to the analyser input, was frequency-modulated via a pre-emphasis network (50 μ s) and a stereo coder by either coloured noise or by the AF signal representing a typical stereophonic programme. With coloured noise the frequency deviation was ± 32 kHz as in the method described in Annex I to Recommendation 641, while for AF signals the frequency deviation corresponded to normal operating conditions, i.e. ± 40 kHz when the AF signal is a 500 Hz sinusoidal tone at studio level (+6 dBm). This operating condition (CCITT maximum amplitude indicator (CCIR Report 292)) is obtained with a transmitter adjusted so that the maximum deviation in the stereo mode does not exceed ± 75 kHz.

Stereophonic measurements with coloured noise lead to a further problem. Until now FM transmitters were modulated with coloured noise in the monophonic mode only. If for the A and B channel the same noise source is used, both signals are correlated and thus no difference signal, S , is produced as long as the levels of A and B are equal. As two different noise sources with uncorrelated signals are not representative of broadcasting programmes, it is possible to find a solution based on the amplitude distribution of FM emissions. To be specific, the ratio of the sum signal M and the difference signal S was measured. Several measurements which were in good agreement with one another showed that on average:

$$M \approx S + 10 \text{ dB}$$

In the case of measurements with coloured noise in the stereophonic mode the appropriate levels were obtained by supplying different levels to the A and the B inputs to the stereo coder. The ratio chosen was:

$$A = B - 6 \text{ dB}$$

That means the signal level in the A channel was half that in the B channel. Thus one obtains on a linear scale:

$$\text{Sum signal:} \quad M = (A + B)/2 = 1.5 B/2$$

$$\text{Difference signal:} \quad S = (A - B)/2 = 0.5 B/2$$

and, consequently:

$$M = 3 S$$

i.e. the M signal is roughly 10 dB greater than the S signal.

4. Results of measurements

Initially, the cumulative amplitude distribution with time was determined in various spectral bands, each 1.2 kHz wide, within the FM spectrum of a monophonic or stereophonic transmission. The unmodulated RF carrier served as a reference level (0 dB). Because of the limited dynamic range of the measuring arrangement (typically 85 dB) only in one case could measurements be made at frequency displacements up to 150 kHz. As the RF carrier in the monophonic mode remains unmodulated during programme pauses, the levels tend towards 0 dB at 0 kHz frequency difference and small time percentages. This result is in contrast to that obtained with coloured noise.

From the amplitude distributions measured, the power contained in the corresponding frequency bands can be calculated. The resulting power-density distribution within the frequency spectrum of an FM transmitter depending on the modulation mode and the simulated programme signal is depicted in Figs. 2 and 3. It should be noted that the power density is not related to 1 Hz, but the curve identified by "eff" is the level difference between the unmodulated RF carrier and the power measured within 1.2 kHz noise bandwidth.

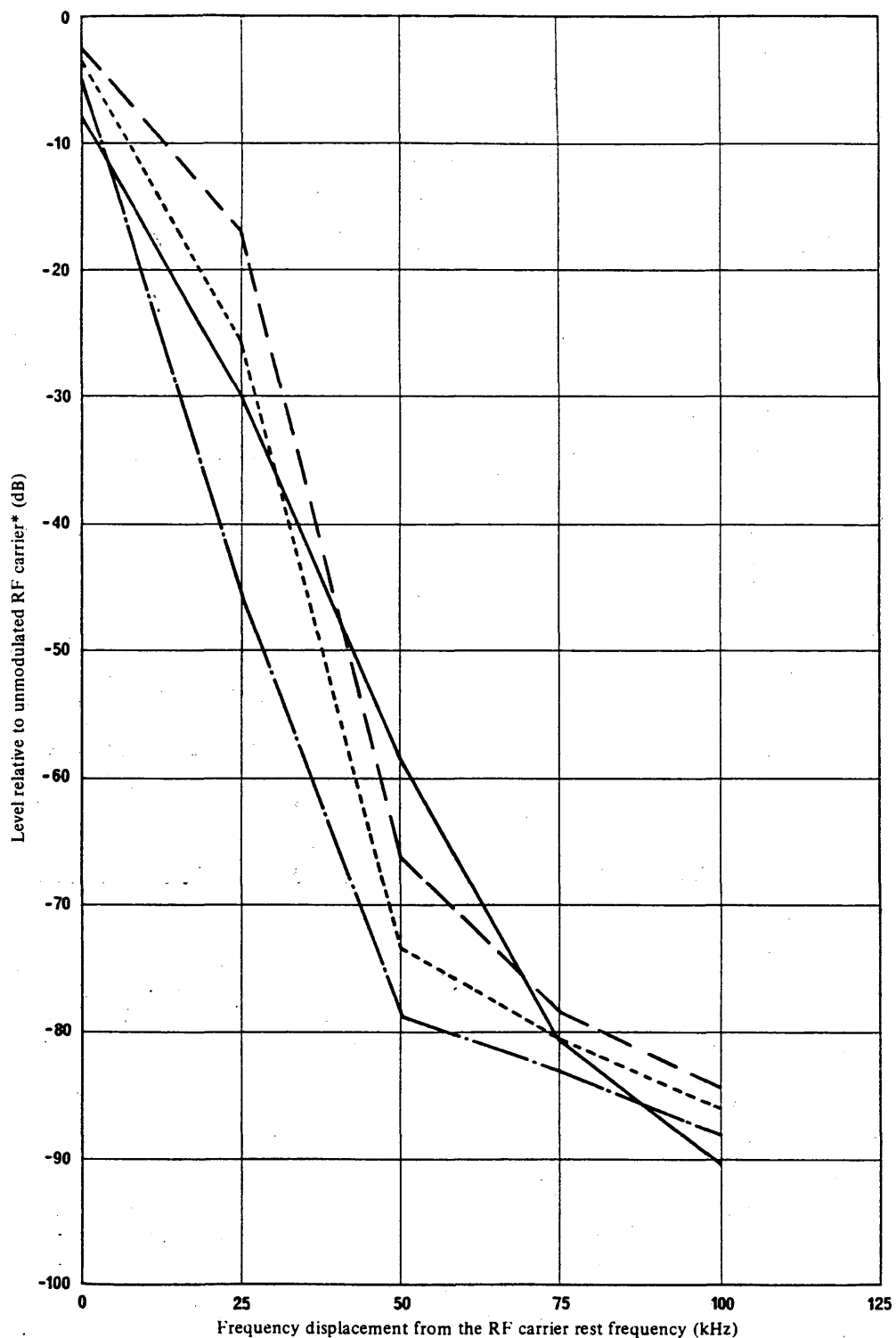


FIGURE 2 - FM spectrum with programme (mono)

- 1% } time for which a level is exceeded
- - - 3% }
- . - 10% }
- eff: r.m.s. level

* Measured within 1.2 kHz noise bandwidth.

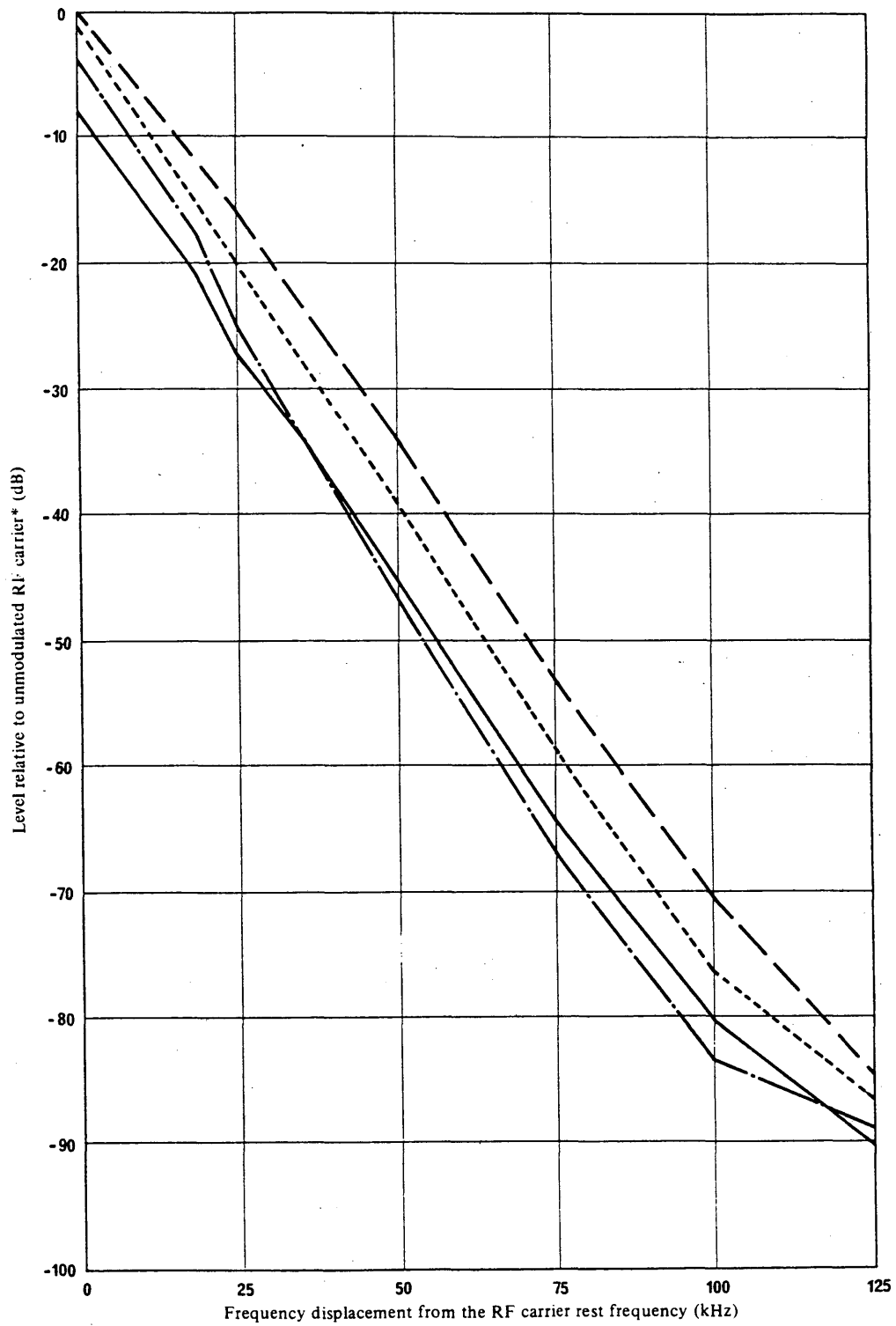


FIGURE 3 – FM spectrum with programme (stereo)

---	1%	} time for which a level is exceeded
...	3%	
-.-.-	10%	
—	eff:	r.m.s. level

* Measured within 1.2 kHz noise bandwidth.

Frequently, not only the power-density distribution is of interest but also those levels which are exceeded or reached for small percentages of time. Therefore, the values corresponding to 1%, 3% and 10% of the time have been evaluated and depicted as curves in Figs. 2 and 3.

A comparison of those figures clearly shows that for stereophonic modulation the spectrum is wider than for monophonic modulation. It is worth mentioning that the frequency spectrum generated by the representative broadcasting programme is considerably smaller than that generated by coloured noise.

For the purpose of this study it was assumed that 99%* (99.9%) of the power of the FM signal is contained within the occupied bandwidth. Calculating the occupied bandwidth from the spectral power-density distributions, one obtains the following values:

- coloured noise, mono: $B_n = 80 \text{ kHz}$ (100 kHz)
- coloured noise, stereo: $B_n = 74 \text{ kHz}$ (112 kHz)
- programme, mono: $B_n = 44 \text{ kHz}$ (62 kHz)
- programme, stereo: $B_n = 50 \text{ kHz}$ (80 kHz)

In general, the spectra for stereophonic modulation are somewhat wider than for monophonic modulation. However, with coloured noise the occupied bandwidth calculated for 99% of the power is 6 kHz wider for monophony than for stereophony. This can be explained by the way the energy is distributed in the vicinity of the carrier.

However, when considering the necessary bandwidth, i.e. the bandwidth necessary to obtain a defined quality standard (e.g. small signal distortions) other values are obtained. If one accepts for instance, that outside the necessary bandwidth levels may exist for only 1% of the time which exceed -60 dB relative to the unmodulated RF carrier, a condition which corresponds to non-linear distortions $\leq 0.1\%$, one obtains the following bandwidths:

- coloured noise, mono: $B_n = 147 \text{ kHz}$
- coloured noise, stereo: $B_n = 211 \text{ kHz}$
- programme, mono: $B_n = 94 \text{ kHz}$
- programme, stereo: $B_n = 170 \text{ kHz}$

The values of occupied and necessary bandwidth both seem to be relatively small and the values for the occupied bandwidth differ from the values given elsewhere. The necessary bandwidth, however, depends on the quality requirements. A requirement for lower distortion or smaller percentages of time would lead to an increase in necessary bandwidth.

The results obtained may contain small errors due to the limited measuring accuracy. However, it can clearly be seen that modulation with coloured noise will lead to larger occupied and necessary bandwidths than a representative programme signal, for monophony as well as for stereophony.

5. Conclusions

On the basis of the measurement results presented, the RF bandwidths of FM sound-broadcast emissions are obviously smaller than one would expect from calculations using Carson's rule. The results obtained so far have not, however, suggested a more reliable formula to replace Carson's rule. Supplementary measurements and theoretical studies are needed to develop a reliable model describing the relationship between typical modulating signals and the associated values of the occupied or the necessary bandwidth in FM sound broadcasting.

It should be stressed that the measurement results presented are still insufficient to permit conclusions to be drawn with respect to the specification of passband characteristics of FM receivers.

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* Unless otherwise specified by the CCIR (which is not the case as yet) the value of 99% is in agreement with No. 147 of the Radio Regulations.

REPORT 464-5

POLARIZATION OF EMISSIONS IN FREQUENCY-MODULATION
BROADCASTING IN BAND 8 (VHF)

(Question 46/10, Study Programme 46C/10)

(1970-1974-1978-1982-1986-1990)

1. Introduction

From the outset of broadcasting in band 8 (VHF), the provision of high quality was a key feature and was given further impetus by the introduction of stereo transmissions. Thus, during the planning of these new services, consideration was given primarily to properly installed receiving installations employing roof-mounted antennas where studies had shown some advantage in the use of horizontal polarization.

The advent of the transistor has led to the mass production of cheap portable VHF/FM radio sets of high sensitivity and capable of satisfactory operation from built-in rod antennas. Also, the use of VHF/FM with car radios is increasing. Generally, these newer categories of receivers enable a mass audience to be reached and, although they are not able to derive the fullest benefit from high-quality transmissions, they are used in such numbers that they cannot be ignored by broadcasters. It must also be mentioned that there is generally a continuing increase in demand for receivers giving high-quality reception, including stereo.

Thus fixed home receivers may employ fixed antennas with the advantage of greater gain and directivity than is the case for a car or portable radio-antenna which has to be nominally omnidirectional in the horizontal plane.

This Report is intended to provide broadcasters with additional data from which they may more easily choose the most appropriate polarization to use for any new service, according to the individual circumstances.

2. Forms of polarization for broadcast services

The polarization employed for broadcast services may take several forms, and these are defined as follows:

Horizontal polarization: with the electric vector in the horizontal plane.

Vertical polarization: with the electric vector in the vertical plane.

Slant polarization: with the electric vector inclined at 45° to the horizontal. This may be considered as the result of equal-amplitude vertically- and horizontally-polarized components combined in the same phase.

The polarization is said to be right or left slant when the electric vector of the propagating wave has been rotated clockwise, or anti-clockwise, respectively, 45° from the vertical position, as seen from the transmitting point.

Elliptical polarization: with the electric vector rotating in a circular motion. This may be considered as the result of vertically- and horizontally-polarized components differing in phase and/or amplitude.

The polarization is said to be right-hand or left-hand elliptical when the electric vector of the propagating wave rotates clockwise, or anti-clockwise, respectively, as seen from the transmitting point.

Circular polarization: with the electric vector rotating in a circular motion. This may be considered as a special case of elliptical polarization where equal-amplitude vertically- and horizontally-polarized components are combined in phase quadrature.

The polarization is said to be right-hand or left-hand circular when the electric vector of the propagating wave rotates clockwise, or anti-clockwise, respectively, as seen from the transmitting point.

Dual polarization: when substantially equal-amplitude vertically- and horizontally-polarized components are radiated without particular control of the phase relation between them. Typically, the vertically- and horizontally-polarized sources may be displaced one from the other so that the resultant polarization varies between circular and slant, according to azimuth angle.

Mixed polarization: the collective term applied when both vertical and horizontal components are radiated, embracing slant, elliptical (including circular) and dual polarization.

In practice, because of the types of antenna employed for domestic reception, one type of mixed polarization is unlikely to have any significant advantage over another.

When the broadcaster wishes to use mixed rather than horizontal or vertical polarization, the choice may be made with regard to the most convenient arrangement for realisation of the transmitting antenna. In such cases, it is necessary to increase the transmitting power by 3 dB when it is intended to provide the same input voltage to receivers employing horizontally- or vertically-polarized receiving antennas. It is to be noted that, where directional antenna patterns are employed, the specified form of polarization is likely to be obtained only within the main beam.

3. Use of horizontal, vertical and mixed polarization

The evidence available suggests that for the type of antenna used with portable and car receivers, there is, in some circumstances, a signal-strength advantage to be gained through the transmission of a vertically polarized component. The use of a vertically polarized component transmitted in addition to the existing horizontal component has become popular in North America and new services frequently use circular polarization. This has been found to offer advantages for car reception in filling in the standing-wave pattern as well as providing generally higher signal strength in open areas. Tests described in [CCIR, 1982-86a] support the above but for flat terrain environment only. In mountainous terrain, coverage is limited by the topography; long-range multipath distortion degrades the service and there is little difference in the average signal level for home reception between the horizontal and circular polarized emissions. In Ireland, **most band-8 transmissions are vertically polarized but mixed polarization is used** by stations covering major urban areas while in the United Kingdom most stations (including 13 BBC stations each having three or four high power transmissions in the range 130 - 250 kW) now radiate mixed (circular or slant) polarization and the remaining horizontally polarized stations are being converted to mixed polarization.

Tests carried out in Sweden [CCIR 1978-82a] also show that car-radio reception is improved in built-up areas close to the transmitter by using circular or vertical polarization rather than horizontal. However, work in the Federal Republic of Germany and in Italy has shown that horizontally-polarized transmissions are less susceptible to multipath distortions in certain types of terrain and that this is especially important for stereo or other multiplexed signals.

Investigations to evaluate the difference in propagation of vertical and horizontal waves were carried out in 1988 within the Socialist Federal Republic of Yugoslavia by RTV Ljubljana and the Institut für Rundfunktechnik (Federal Republic of Germany) [CCIR 1986-90]. The station under test was a high power VHF FM sound broadcasting transmitter within a service area including rolling as well as rugged terrain. The results show that vertical waves have a much stronger diffraction loss at most locations in the measuring area and were found to produce much more diffuse reflections. This caused a perceptible increase of distortions in home reception and a significant multiplication of noisy drop-outs in car reception.

Experience in the United Kingdom [CCIR 1978-82b], particularly as a result of tests carried out in the London area showed that, compared with horizontal polarization, mixed polarization (see § 2) provided an improved service to listeners using portables or car radios without causing a significant degradation to stereo listeners using properly installed, horizontally polarized, outside antennas. Using an e.r.p. of 400 W the measurements were made within a range up to 25 km from the transmitter site and it was found that the addition of a vertical component (of equal power) to an existing horizontally polarized transmission led to an effective increase in signal for portable and car receivers of between 5 and 7 dB. Canadian tests confirm the above for a flat terrain environment and up to approximately 160 km. A circularly polarized antenna was used and station powers varied from 21 kW to 360 kW e.r.p. The advantage of using circular polarization appears to occur mainly for the mobile receivers, in the fringe areas, but the investment and operational costs needed are higher than for horizontally polarized emissions.

4. Use of orthogonal polarization

Report 122-3 (Geneva, 1982) describes advantages which may be obtained by transmitting orthogonally polarized signals (vertical and horizontal) from different stations and using polarization discrimination in reception. In principle, this technique is also applicable in the frequency bands used for FM sound broadcasting. However, full advantage can only be obtained when the polarization of receiving antennas conforms to that of the wanted signal. In cases where the polarization of the receiving antennas is random, e.g. when in VHF/FM sound broadcasting, antennas for domestic and car reception have different polarization, no general advantage can be expected from making systematic use of orthogonal polarization and this may be detrimental to reception under these conditions.

Furthermore, where mixed polarization is used, the opportunity is generally lost for the provision of mutual protection between services through the use of orthogonal polarization (vertical and horizontal). In the case of mixed polarization, no advantage in polarization discrimination can be gained over horizontal or vertical polarization by using, at the receiving site, the same type of antenna as at the site of the wanted transmitter. This would also apply for right and left circular polarization, because the direction of rotation of a signal coming from the back of an antenna is opposite to that of a signal coming from the front.

5. Definition of effective radiated power for interference calculations in the case of mixed polarization [CCIR, 1978-82c]

Any kind of mixed polarization can be represented by a horizontally- and a vertically-polarized component which, depending on the type of receiving antenna used, may provide different input voltages to the receiver. Because of the general loss of mutual protection between services through the use of orthogonal polarization — as mentioned above — it is not important to take the phase relationship between the two components into account when carrying out interference calculations. However, the effective radiated power (e.r.p.) has to be clearly defined. It is therefore appropriate to specify separately the powers associated with the two orthogonal components of the radiated signal (i.e. the horizontally and vertically polarized components), and not merely to indicate the value of the total e.r.p. only.

6. Propagation factors

6.1 *Direct or diffracted component*

Broadcast signals arriving at a receiving point are normally made up of several components, the strongest of which is either a direct wave or a diffracted wave. The magnitude of this direct or diffracted wave is substantially independent of polarization, except where diffraction occurs over smooth rounded hills (without vegetation or obstructions higher than about one metre), in which case a vertically-polarized wave will be attenuated less than a horizontally polarized wave. Another special case is where propagation is over hills wooded with coniferous trees, as the greater tendency of trees to scatter the vertically-polarized component contributes to a greater loss for this polarization. Close to the ground, the boundary conditions favour a vertically-polarized component, as any horizontally-polarized component is attenuated. Thus, at low antenna heights, vertical polarization will give the stronger signal.

6.2 *Reflected component*

In addition to the direct or diffracted component there will generally be one or more reflected signals, and these may arise from a number of causes which may be categorised into the following three groups:

6.2.1 *Grazing reflections from the ground in front of the receiving point*

The path difference between direct and reflected signals is normally very small, and the echo, in consequence, does not contribute to distortion but only to variation in field strength. In most cases where the grazing angle is not more than a degree or so, the effect is substantially independent of polarization; however, if transmission is between high transmitting and receiving points, e.g. across a valley or bay, the echo delay time may become significant, and horizontal polarization will normally give a rather larger echo.

6.2.2 *Reflections from the side of the path, where the reflecting object may be a man-made structure, trees, hills or mountains*

When the reflections are from vertical or near-vertical surfaces, their magnitude will generally be significantly greater for vertically-polarized transmissions as compared with horizontally-polarized ones. It is this factor which is likely to make vertically-polarized transmissions rather more liable to multipath effects when compared with horizontally-polarized transmissions. In practice, the effect is more pronounced when the direct path is obscured and there are strongly illuminated structures or topographical features to the side of the path. The effect can become important where the path difference between direct and reflected signals exceeds about 3 km (10 μ s). With this path delay, perceptible distortion may occur at an echo amplitude of about 15% for stereo or 40% for mono, although the actual value will depend upon the amount of AM suppression in the receiver.

6.2.3 *Reflections from behind the receiving point and occurring from vertical or near-vertical surfaces such as buildings, trees or mountains*

Where dimensions of the reflecting object are large in both the vertical and horizontal plane, then polarization is often found to be unimportant, although some evidence has been found in the Federal Republic of Germany that buildings tend to reflect vertically-polarized band 8 signals more than those which are horizontally polarized. Where the reflecting object is a tree, particularly of the coniferous type, vertically-polarized signals will be reflected more strongly. As in §6.2.2 the effects are important only when the path difference exceeds about 3 km.

6.3 *Summary of propagation factors*

It is evident, therefore, that transmissions of a vertical component can result in an increased signal strength where receiving antennas are essentially vertically polarized and used at low heights. Offset against this advantage is the possibility of an increase in multipath distortion in certain types of terrain. This distortion is the more serious when multiplexed signals are transmitted.

Present evidence shows that, in flat or rolling country, multipath effects are minimal, while the increase in received signal strength obtainable by transmitting a vertical component is the greatest. At the other extreme, in rugged mountainous country, the increase in received signal strength is small, whereas multipath effects may be severe, especially when reflections are from mountains with coniferous trees.

Thus, in these extreme cases, it is possible to make a decision as to the most suitable polarization to employ for transmission. For the more general intermediate cases, the decision may be less clear-cut and be dependent upon the type of service the broadcaster is attempting to provide.

7. **Man-made interference**

The main source of interference is from vehicle-ignition systems, and experience seems to show that the magnitude of the interference differs from one country to another. The explanation for this is not known with certainty, but it may be due to the different standards adopted for vehicle electrical system suppression.

Interference radiated from vehicles has a semi-random polarization but with a preponderant vertical component. Thus, where receiving antennas are well above ground level and of the same polarization as the transmissions, it is advantageous to use horizontal polarization. Where the receiving antenna is used near ground level and responds primarily to vertically polarized signals, as is the case at present for portable and car receivers, the signal-to-interference ratio can be improved only by the radiation of a vertical component.

8. **Existing services**

It is important to know whether services already exist in the area to be served. If so, it will presumably be necessary to employ a polarization of transmission such that existing receiving systems can, in most cases, provide satisfactory results. Generally, this is likely to mean that where existing services use horizontal or vertical polarization, the new service should use either the same or mixed polarization.

9. Discussion

The various factors affecting the choice of polarization for new services in band 8 (VHF) have been reviewed. It appears that certain advantages exist for the use of horizontal, vertical or mixed polarization, and that these advantages take on a different weight according to the type of terrain, the type of receiving installation and the type of service being provided. No one type of polarization is best for all circumstances.

Table I is intended, with this in mind, to assist the broadcaster in making a choice according to circumstances.

TABLE I - Choice of polarization appropriate for new services

Type of service intended	Polarization of existing services	Type of terrain	Preferred polarization for new services
Primarily for high-quality receiving installations, probably with stereo, with no improvement to reception conditions for portable or car sets envisaged.	nil	any	horizontal
	horizontal	any	horizontal
	vertical	flat or rolling	vertical
	vertical(1)	rugged	mixed(2)
Primarily to reach the largest audience, especially those using portable or car sets. Account to be taken of those installations already equipped to receive any existing transmissions.	nil or horizontal	flat or rolling	mixed
		rugged	horizontal
	vertical	flat or rolling	vertical
	vertical(3)	rugged	mixed

(1) It would be preferable to change any existing services to horizontal polarization.

(2) Horizontal, if existing services can be changed to horizontal polarization.

(3) It would be preferable to change any existing services to mixed polarization.

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REPORT 944

THEORETICAL NETWORK PLANNING

(Question 46/10, Study Programme 46L/10)

(1982)

1. Introduction

Broadcasting-transmitter networks should be planned in such a way that the required coverage of the area is provided using the minimum number of frequencies. From the purely technical standpoint, the coverage area of each transmitter depends upon a number of factors, for example: transmitter power, minimum usable field strength, radio-frequency protection ratio, the distance between transmitters sharing the same or adjacent channels, channel spacing, bandwidth of emission and other factors influencing wave propagation. It also depends on the channel distribution scheme.

When a large number of channels are to be planned or replanned for a particular AM or FM sound or television service, it has been found that effecting an efficient use of the spectrum can prove difficult when employing empirical methods only. For this reason, a theory of uniform transmitter networks was developed during the late 1950s and early 1960s [EBU, 1960]. This method can be applied with success when some uniformity of standards exists for the service to be planned. Furthermore, the frequency band to be planned should be constrained as little as possible, i.e. there should ideally be complete freedom in assigning any frequency to any transmitter.

The theory is not only useful in designing new or remodelling actual transmitter networks, but also provides a powerful tool for determining optimal technical parameters such as channel spacing, transmitter characteristics, etc., and identifying the best attainable coverage possible.

The method described below has already been used during the VHF/UHF European Broadcasting conference, Stockholm 1961, during the African VHF/UHF Broadcasting Conference, 1963, and helped in preliminary studies for the Geneva 1975 LF/MF Broadcasting Conference.

2. The theory of regular networks

It should be noted that the networks which will be studied in the present section are purely theoretical in the following sense:

- all the transmitters are identical: their power and antenna height are the same;
- they are equipped with non-directional antennas;
- propagation is isotropic and independent of frequency, at least within the band to be planned.
- for the purpose of calculating distances, the planning area is assumed to be flat and the population evenly spread over its surface; there are neither political nor natural boundaries.

In those conditions, and provided interference is negligible, the coverage area, i.e., the region where a good reception is achievable with a normal domestic receiver, is limited by a contour within which the electromagnetic field strength is greater than or equal to the value necessary to obtain a given signal-to-noise ratio. In the ideal situation described above, this contour is a circle whose radius depends on the type of service and the propagation laws valid for the frequency range considered. If it is desired to cover the whole of the planning area using such circular coverage areas, it is quite obvious that the number of transmitters per unit surface will be minimized if they are situated at the vertices of a lattice of equilateral triangles (Fig. 1). It should be noted that there is not one transmitter per triangle, but rather one transmitter for each pair of triangles forming a rhombus (shaded area of Fig. 1). This remark is of importance when computing the network efficiency.

With this configuration, every location on the planning area is served by at least one transmitter and overlapping of service areas is minimized. In what follows, we will take as a unit of length the side of the elementary equilateral triangle. The transmitter density is then one per elementary rhombus, the area of which is $\sqrt{3}/2$. The network efficiency can be expressed as the ratio of the total area where the service is acceptable (the area of the elementary rhombus) to the sum of the (overlapping) coverage areas. A regular lattice of equilateral triangles gives the optimum value, i.e., 0.83 (each transmitter has a coverage area of $\pi/3$ with a radius of $\sqrt{3}/3$). The ideal value would obviously be 1, but it is not attainable as it is impossible to cover a surface with circles without overlapping areas. A regular lattice of squares (or of isosceles right triangles) only achieves a ratio of 0.64 (each transmitter has a coverage area of $\pi/2$ with a radius of $\sqrt{2}/2$).

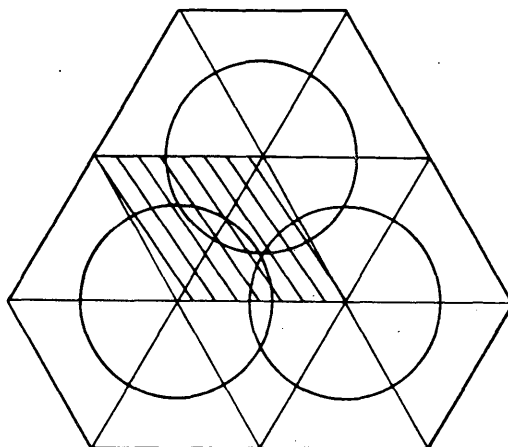


FIGURE 1 – Location and coverage areas of identical broadcasting transmitters in a regular network

It should be noted that actual transmitter networks are neither geometrically regular, nor do their technical characteristics correspond to those of the theoretical network. Deviations from geometrical regularity and from the power and antenna height of the theoretical network towards lower values will inevitably result in lower network efficiency. Nevertheless it is possible to derive, from the results of studies concerning theoretical networks, a fairly clear picture of the relationship between these factors and network efficiency.

3. An example of theoretical network

It is assumed first of all that the available frequency band has been divided, in an appropriate manner, into channels sufficiently wide for a sound or television programme to be located within its limits and adequately spaced (9 kHz for AM sound broadcasting; 100 or 200 kHz for FM sound broadcasting; 6 or 8 MHz for television (200 kHz and 6 MHz correspond to the American channels)) to make co-channel interference predominant vis-à-vis adjacent-channel interference.

It is then easy to derive, from this channel spacing, the number of channels that can be used when planning. As this number is obviously finite, it is inevitable that, at some more or less large distance from the starting point, the first channel assigned has to be re-used. The smaller the distance D , between two transmitters operating on the same channel, the smaller the number of channels needed to cover the whole planning area. But it is necessary that D be large enough for the two co-channel transmitters not to interfere unduly with each other.

It is not difficult to evaluate D . With the help of suitable propagation curves, field strengths of the wanted transmitter and of the interferer are computed, for given percentages of time (see Recommendations 368 and 370), on the circle defined in § 2, limiting the coverage area of the wanted transmitter. D is chosen sufficiently large so that the difference, expressed in dB, between the two field strengths is larger than the protection ratio which is to be found in the relevant Recommendations 560, 412, 418 and Report 306. The case of multiple interference will be dealt with later.

With N channels, it is possible to cover N elementary rhombi, that is an area equal to $N\sqrt{3}/2$, taking as unit area the surface of a square whose side is equal to the side of the elementary triangle. So another kind of network arises using rhombi consisting of N elementary rhombi, and it is logical to stick to the conclusion found in § 2: a given channel will be repeated at the nodes of a lattice of rhombi whose shorter diagonal is equal to the side, the area of which is $N\sqrt{3}/2$. This rhombus will be called the "co-channel rhombus" and the length D of its side is $\sqrt{N}$ ($\sqrt{N}$ times the side of the elementary triangle). The characteristics of rhombic lattices will be shown in a simple example, where $N = 13$ (see Fig. 2).

- **Period:** all the co-channel rhombi are identical. A two-dimensional periodicity appears. It may remind the reader of the theory of elliptical functions.
- **Channel distribution:** within a co-channel rhombus, channels might be distributed at random, but in every rhombus their locations should be identical. However, a random distribution is not suitable if it is necessary to take into account interference other than co-channel interference, which is usually the case. Assuming that a homogenous network leads to the most efficient utilization of the spectrum, it is necessary for the interfering signals on any channel to have the same strength, wherever the transmitter is.

Hence if, taking account of the protection ratio between adjacent channel transmitters (referred to here as 1-channel transmitters; likewise 2-channel transmitters are spaced by 2 channels, etc.), channel 1 is placed at a point which is sufficiently far from channel 0, this defines for channel 1 a co-channel rhombic which is offset from the co-channel rhombic for channel 0, by the vector $\vec{01}$. To meet the requirement referred to above, for homogeneity, channel 2 will be placed such that $\vec{12} = \vec{01}$, and so on for the other channels. Channel N falls on channel 0. If we consider the 13 channel network in Fig. 2, it is seen that someone moving along the straight lines of the elementary triangle lattice will find the channel numbers are in an arithmetic progression, modulo 13. Starting from channel 0 and heading west, he will encounter channels: 2, 4, 6, 8, 10, 12, 1, 3, 5, 7, 9, 11, 0, etc. It is therefore very easy to define the entire network on the basis of the progression step in two distinct directions: + 2 towards the left and + 5 towards the top righthand corner.

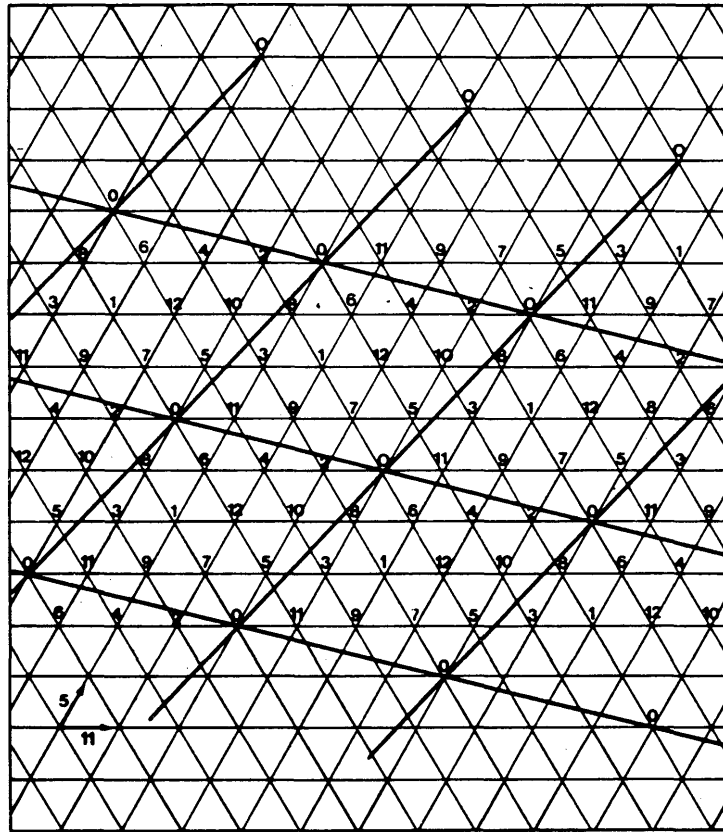


FIGURE 2 - Example of regular lattice for 13 channels

4. Ideal and non-ideal networks

To make the calculations easier, it is convenient to refer the plan to a system of two coordinate axes, bearing equal scales but making an angle of 60° rather than the conventional 90° (see Fig. 3).

We already know that the side of the co-channel rhombus has a length equal to $\sqrt{N}$, but it is necessary for the rhombus to coincide with the intersections of the elementary lattice. In the tilted coordinate system, the distance of any point (x, y) from the origin is:

$$(x^2 + xy + y^2)^{1/2}*$$

If two integers a and b can be found such that $a^2 + ab + b^2 = N$ (number of channels), then there is a co-channel rhombus whose vertices coincide with the intersections of the elementary lattice. $N = 13$ corresponds to the case where $a = 3$ and $b = 1$. In the following discussion, such numbers will be called "rhombic numbers". Some of them are given on Fig. 4.

* All the distances in this section are calculated on the base of this formula and expressed in the multiples of units equal to the length of a side of an elementary triangle.

There are other conditions for the choice of a and b : they shall not have any common divisor, either between themselves or with N , and neither of them can be equal to zero. For instance, if they are both even numbers, only the even-numbered channels will appear. The other restriction will be given later. The theory still works if N is not exactly a rhombic number. The co-channel rhombus becomes a co-channel parallelogram, with an area equal to N . In this case, the sides of the parallelogram are not equal; one of them is shorter than $\sqrt{N}$ and as a consequence this co-channel distance is slightly smaller than for the rhombus.

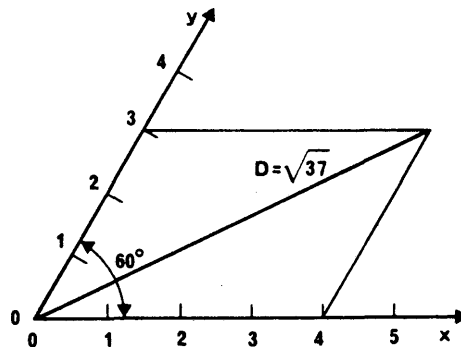


FIGURE 3 - Reference coordinates of regular lattices

$\begin{smallmatrix} a \\ b \end{smallmatrix}$	1	2	3	4	5	6	7	8	9	10	11	12
1	3	7	13	21	31	43	57	73	91	111	133	157
2			19		39		67		103		147	
3				37	49		79	97		139		
4					61		93		133			
5						91	109	129	151			
6							127					

FIGURE 4 - Table of rhombic numbers

Figure 5 shows a network with 26 channels, 26 not being a rhombic number. However, in this particular case, it is seen that the shortest co-channel distance is the shorter side of the parallelogram whose length equals 4.358. This is not much less than the distance $\sqrt{N} = 5.099$ which would have been obtained if 26 were a rhombic number. The lengths of other co-channel distances, e.g. the longer side of the parallelogram or the shorter diagonal of it, are 5.291 and 6.082 respectively. The next rhombic number is 31 an example of which is given in Fig. 6.

Figure 7 shows a network with 120 channels, 120 not being a rhombic number. However, in this particular case, it is seen that the shortest co-channel distance is the diagonal whose length equals 10.58. The lengths of the sides are 11.53 and 10.82, respectively. None of these numbers is much less than the distance $\sqrt{N} = 10.95$, which would have been obtained if 120 were a rhombic number. This lattice may not be optimum for VHF planning.

In the case of N not being a rhombic number, there are three possibilities:

- not to take note of this value N and consider the nearest rhombic number which could be used instead (e.g. 111; see Fig. 4);
- to consider the resulting regular lattice despite its consisting of co-channel parallelograms;
- to distort the lattice so as to transform co-channel parallelograms to rhombi. In these circumstances the elementary rhombi would become parallelograms, and network efficiency would decrease. This decrease will only be slight when N is fairly large.

The best choice among the three options depends on the circumstances to be taken into account.

It is not always easy to find the progression steps such that the channel assigned to the origin will reappear at the vertices of the co-channel rhombus or parallelogram. Referring Fig. 2 to the slant coordinate system of Fig. 3, the steps p along the x -axis and q along the y -axis must satisfy the following equations:

$$1 \cdot p + 3 \cdot q = kN \quad \text{and} \quad 4 \cdot p - 1 \cdot q = k'N$$

k and k' are integers.

In Fig. 2, $p = 11$, $q = 5$, $k = 2$, and $k' = 3$, for $N = 13$. These steps are indicated in Fig. 2. In the case of a non-rhombic number, there is more than one parallelogram and many attempts are necessary to find the best one [Arnaud, 1962]. Even in the case of rhombic numbers it may happen that two different co-channel rhombi exist, as it appears on Fig. 4 for 91 (9 and 1 or 6 and 5).

The case where a or $b = 0$ is a trivial one. If, for example, $b = 0$, the co-channel rhombus sides coincide with the x and y axes, and $a = \sqrt{N}$. The sum of steps along the axes from one vertex to the next must be N ; consequently their value is $\sqrt{N}$ and it is then only possible to write on the lattice channels that are multiples of $\sqrt{N}$.

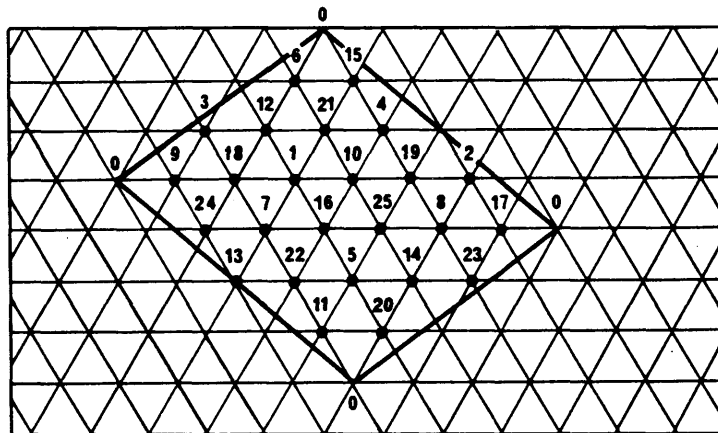


FIGURE 5 - Network with 26 channels

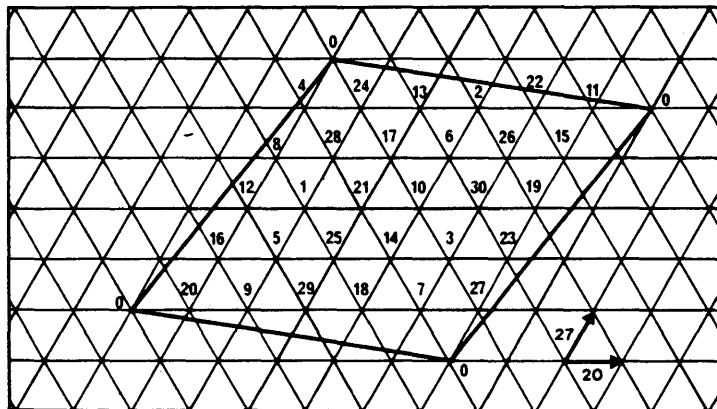


FIGURE 6 - Example of an optimum regular lattice for 31 channels

5. Evaluation of interference

5.1 Co-channel interference

It has been explained in § 3 how the minimum necessary number of channels derives from the consideration of propagation curves and protection ratios, but only one interfering transmitter has been taken into account. On Fig. 8 are drawn a few co-channel rhombi for a co-channel distance D , the wanted transmitter being at the centre and the eighteen others surrounding it are the interferers. In the case of VHF or UHF broadcasting, the coverage area is evaluated on the basis of the 50% or 1% of the time propagation curves of Recommendation 370 (for the wanted or interfering transmitter, respectively), taking due account of the characteristics of the

transmitter (effective radiated power, antenna height, frequency) and the minimum usable field strength. (See Recommendation 412). In § 2, the coverage area radius has been found to be $\sqrt{3}/3$ times the side of the elementary rhombus. So the scale of the lattice is known. The first six interfering transmitters are at a distance $D = \sqrt{N}$, and as a first approximation it will be assumed here that D is large enough and that the interfering field strength is almost the same at the wanted transmitter location and on the fringe of its coverage area. In a more accurate study, the ratio of wanted/interfering signals should be calculated at the limit of the coverage area, at least at the six locations shown by roman numerals I to VI on Fig. 8.

D has to be long enough for the difference (in decibels) between the wanted field strength exceeded for 50% of the time and the combined interfering field strengths exceeded for 1% of the time, to be greater or equal to the co-channel protection ratio. Then it should be checked whether the twelve transmitters located on the second hexagon (see Fig. 8) have any influence, but in general this is not the case unless D is very small.

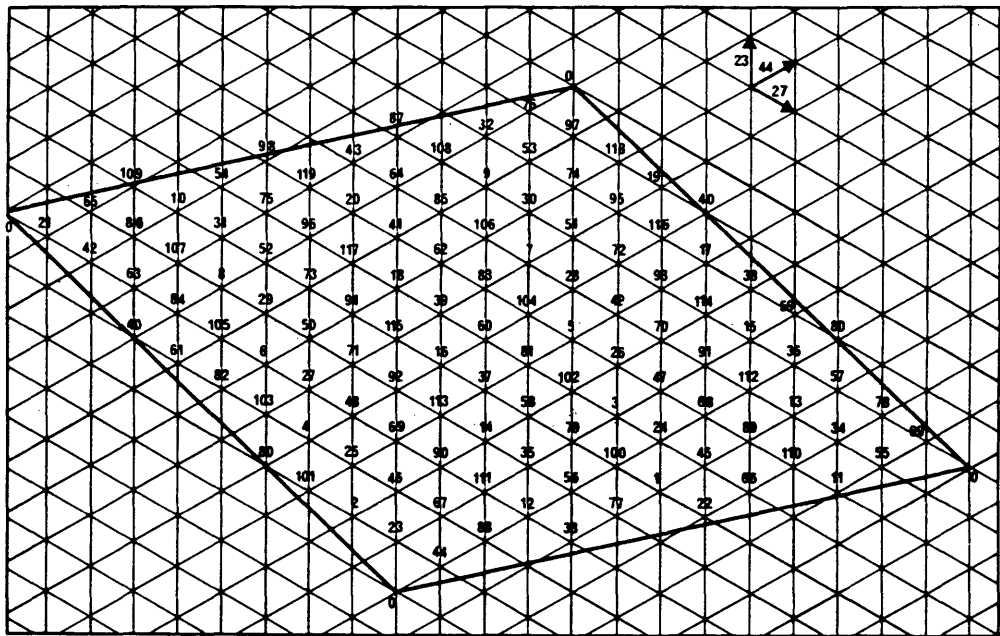


FIGURE 7 - Network with 120 channels

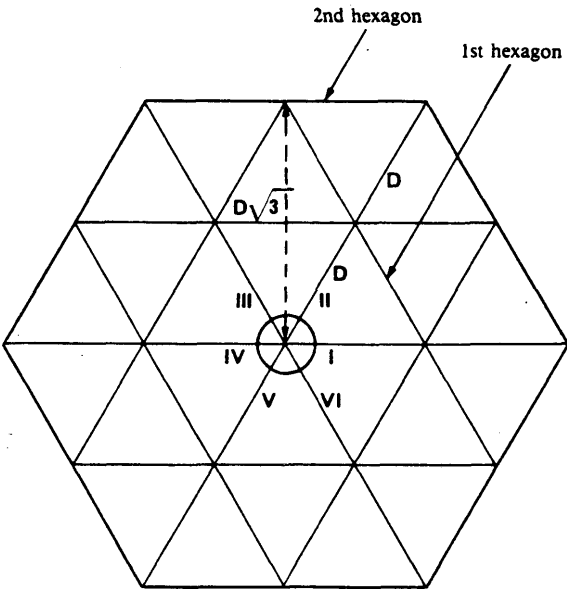


FIGURE 8 - Location of co-channel transmitters

5.2 Adjacent channel interference

The network of Fig. 2 will serve as a practical example. D is assumed to have been found to be equal to $\sqrt{13}$. By definition, the network is regular and any given channel can be studied. Take for instance channel 7. The adjacent channel distances are $\sqrt{3}$ and 2. There are 4 adjacent channel interfering transmitters. As in the case of co-channel interference, their total field strength will be evaluated at the transmitter location first, and then, if the wanted/interfering signal ratio is only slightly different from the protection ratio, at the most critical points at the limit of the coverage area. Other adjacent channel transmitters are much more remote. Their distances are $\sqrt{7}$, $\sqrt{12}$, $\sqrt{13}$, etc.

For a wanted transmitter on channel C , there are in each co-channel rhombus at least two adjacent channel interfering transmitters on channels $C + 1$ and $C - 1$. The further the transmitters $C + 1$ and $C - 1$ are from the corners of the co-channel rhombus of C , the better will be the network; the ideal positions will be the centres of gravity of the two equilateral triangles making up the rhombus. Hence if transmitter $C + 1$ is exactly at one of these centres of gravity, $C + 2$ will be exactly at the other and $C + 3$ will coincide with C which is absurd. This occurs with the network with $N = 21$ channels. If the coincidence of a point in the network with the centres of gravity is only approximate, channel $C + 2$ will be twice as far from the centre of gravity of the triangle, however channel $C + 3$ will be still fairly close to channel C which is unfavourable in FM sound broadcasting as this corresponds to a carrier spacing of 300 kHz (in parts of Europe) for which the protection ratio is only slightly negative (-7 dB).

5.3 Multiple interference

Both in theory and in practice more than one interfering transmitter has to be respected. At the site of the wanted transmitter in the network of Fig. 2 there are co-channel transmitters interfering with near identical signal strength and in addition there are four interfering adjacent-channel transmitters. It should not be overlooked that at least 2-channel and 3-channel transmitters will also need to be taken into account. In practice it has proved useful to compute the usable field strength on the basis of at least those interferers which cause the strongest interference potential. It cannot however be a matter for this description of a planning method to advocate one or the other method of computing multiple interference effects.

6. Planning constraints

When planning a particular type of service, some constraints have to be kept in mind and most of them result from the receiver design. It is useful to distinguish between internal and external constraints. Such constraints and their generating mechanisms are discussed in Report 946.

7. Use of theoretical networks

Theoretical networks may be of some help in trying to find the optimum characteristics of transmitters [Sauvet-Goichon, 1980] or the most suitable type of modulation parameters or channel spacing. There is a good probability that a solution found to be optimum in a regular network would be the best one for an actual network as well and in any event it will be far easier to study the former with the help of computerized methods.

7.1 Simple network

The allocated band, 16 MHz wide, is divided into 159 channels. Below 159 the largest rhombic number is 157 (12 and 1, see Fig. 4). Two channels are then left empty. They may replace channels whose use is precluded by some local constraints (see § 6) or fill gaps in mountainous areas or near borders. As a general rule, more than one programme is broadcast from a given location and some geographical neighbouring assignments will be moved to the same point. Figures 9 and 10 show two examples of channel clustering. Assignments made at the vertices of the shaded areas are concentrated to the centre of gravity of each area. It should be checked that this does not involve an unacceptable increase in interference.

Figure 11 gives the steps to build a 157 channel network. Their derivation is given in Annex I. The 1961 Stockholm Plan for UHF television has been based on a channel clustering C , $C + 3$, $C + 6$.

7.2 Multiple networks

When there is a large number of channels (more than about 50), and if, as is usual, each centre consists of several transmitters, it is possible to split the band into sub-bands, each of them including the same number of channels and possibly reserved for a separate programme. It should be noted that the African Plan has been designed on a channelling of 86 kHz, i.e. 185 channels in 16 MHz, for 6 networks of 31 channels or 5 networks of 37 channels (37 is a rhombic number).

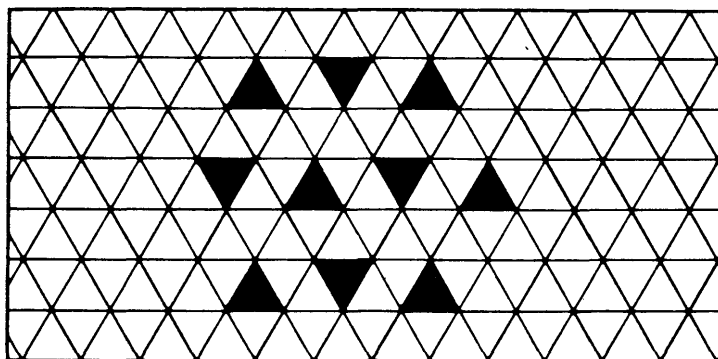


FIGURE 9 - Clustering of neighbouring channels

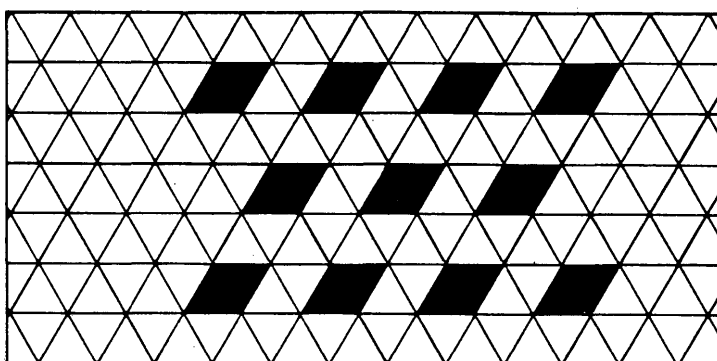


FIGURE 10 - Clustering of neighbouring channels

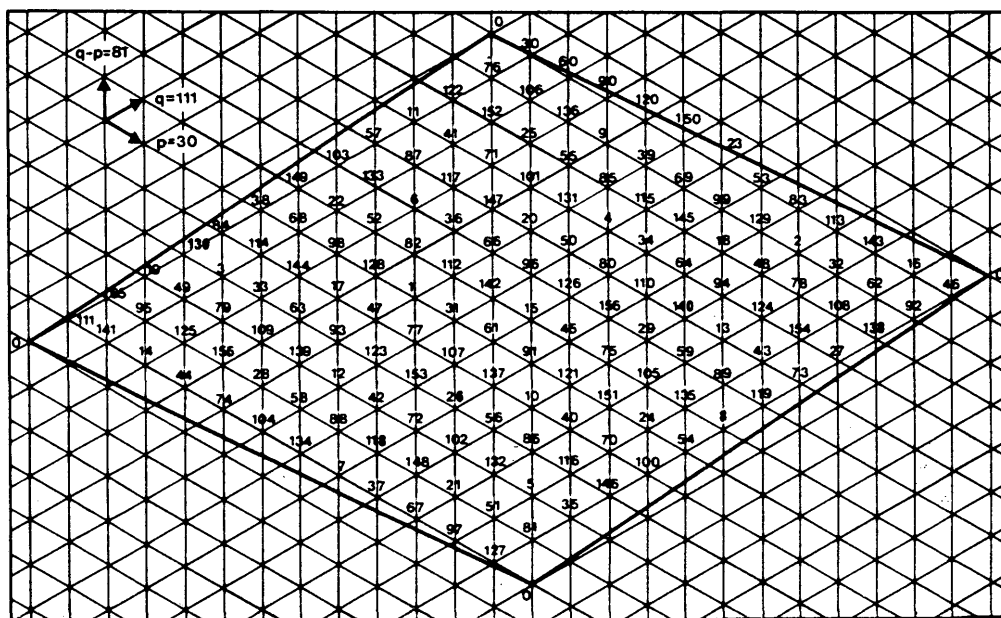


FIGURE 11 - Example of an optimum regular lattice for 157 channels

8. Number of channels needed for full area coverage with one programme in VHF sound broadcasting

8.1 General considerations

Based on the previously discussed method the number of channels needed for the dissemination of one programme in VHF sound broadcasting to provide full coverage was evaluated.

The investigations are made for different average distances between transmitters, two antenna heights, (300 m and 600 m), two coverage factors (90% and 100%) and an effective radiated power of 100 kW. Monophonic and stereophonic services were taken into account. The calculation of the interferences was carried out using the simplified multiplication method (see Report 945).

This method had previously been used during:

- the European Broadcasting Conference, Stockholm, 1961,
- the African Broadcasting Conference, Geneva, 1963.

8.2 Performance of the calculations

8.2.1 Basic assumptions

The investigations are based on idealized networks with a regular lattice of transmitters, linear distribution of channels and equilateral co-channel triangles. The so-called elementary triangles which are formed by three neighbouring transmitters are in this case not equilateral in general, but only for certain numbers of channels [EBU, 1960].

The calculations are based on the following conditions:

- field strength $E(50,50)$ (Recommendation 370),
- field strength $E(50,1)$ (Recommendation 370),
- protection ratios (Recommendation 412),
- local variation of field strength: 8.3 dB,
- receiving antenna,
 - monophony : omnidirectional,
 - stereophony : directional with 12 dB front-to-back ratio,
- minimum usable field strength (Recommendation 412)
 - monophony : 48 dB(μ V/m),
 - stereophony : 54 dB(μ V/m),
- channel spacing : 100 kHz.

8.2.2 Calculation of the usable field strength

8.2.2.1 Method of calculation

The calculation of these interferences was carried out using the simplified multiplication method.

Experience in some countries has shown that the coverage situation found in actual practice is, on average, in acceptable agreement with computation results obtained with the simplified multiplication method. In the subsequent sections only this method has been used for the determination of the required number of channels.

8.2.2.2 Number of interfering transmitters

For the calculation of interferences in the theoretical lattice, the 18 strongest interfering co-channel transmitters and the 40 strongest interfering adjacent channel transmitters having frequency separations up to 400 kHz were taken into account. (The relatively large number of interfering adjacent channel transmitters is only relevant with close channel spacings.)

8.2.3 Variation of parameters

The average distance of transmitters was varied between 40 and 120 km in steps of 10 km. For the effective height of the transmitter antenna, 300 and 600 m were chosen, because the average height values in many European countries fall into that range. The effective radiated power was set to 100 kW to ensure that interference rather than noise will limit the coverage area; this being a condition for efficient spectrum utilization (see Report 414 (Kyoto, 1978)). For small distances between transmitters however, considerably less power would provide equal percentage of coverage with the same number of channels.

8.2.4 Results

The number of channels needed in a theoretical network under different conditions can be found in Figs. 12 to 15. For a monophonic service with an omnidirectional receiving antenna and for a stereophonic service with a directional receiving antenna (12 dB front-to-back ratio), the number of channels required is almost equal.

In an idealized theoretical network based on the above assumptions with an effective transmitter antenna height of 300 m, an effective radiated power of 100 kW and an average distance between transmitters greater than 70 km:

- for 90% coverage about 25 channels are necessary (Fig. 12),
- for 100% coverage about 31 channels are necessary (Fig. 13).

Examples of possible channel distribution for corresponding numbers of channels are given in Figs. 5 and 6.

The number of channels needed for one programme in a theoretical network can only give an approximation to the number needed for a real network. The less the uniformity of the real network considered the more extra channels will be needed. As experience with existing networks in the range 87.5-100 MHz in continental Europe has shown, up to 4 programmes are possible in this range.

Based on theoretical study and practical experience, it may be assumed that for the entire frequency band 87.5-108 MHz should be practicable:

- for 90% coverage about 7 programmes,
- for 100% coverage about 6 programmes,

In practical networks, the power and the effective antenna height of any single transmitter should be set to the lowest possible value to obtain the intended coverage.

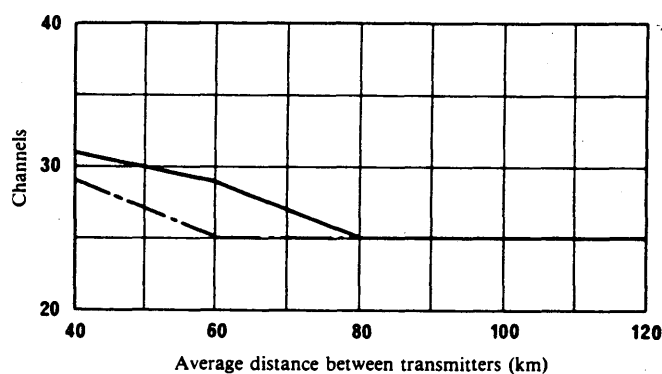


FIGURE 12 - Number of channels needed in a theoretical network for coverage 90 % and antenna height 300 m

--- : monophony (0 dB - antenna gain)
 — : stereophony (12 dB - front-to-back ratio of the antenna)

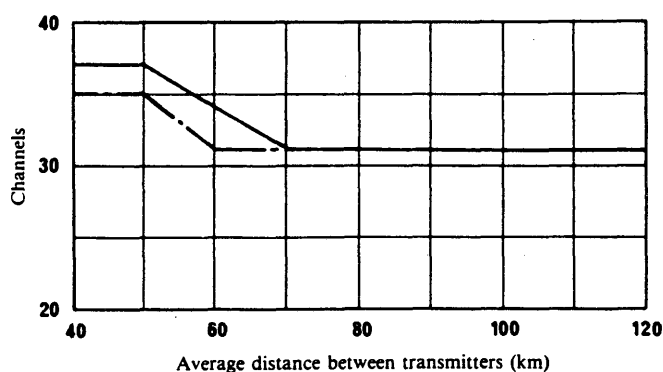


FIGURE 13 - Number of channels needed in a theoretical network for coverage 100 % and antenna height 300 m

--- : monophony (0 dB - antenna gain)
 — : stereophony (12 dB - front-to-back ratio of the antenna)

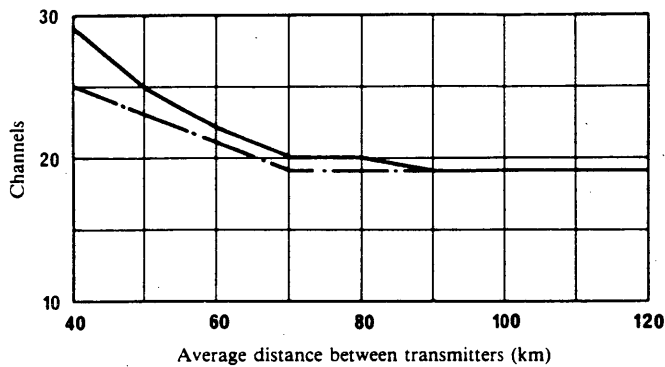


FIGURE 14 - Number of channels needed in a theoretical network for coverage 90 % and antenna height 600 m

--- : monophony (0 dB - antenna gain)
 — : stereophony (12 dB - front-to-back ratio of the antenna)

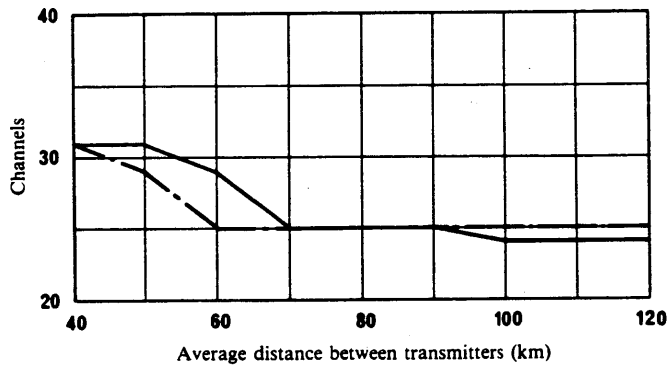


FIGURE 15 - Number of channels needed in a theoretical network for coverage 100 % and antenna height 600 m

--- : monophony (0 dB - antenna gain)
 — : stereophony (12 dB - front-to-back ratio of the antenna)

9. Conclusions

When the number of channels is high, it seems difficult to achieve by purely empirical means the best use of the frequency spectrum. The theory of regular networks appears as an efficient tool to reach an optimum. Moreover, in an area where the surface of the earth is divided amongst many countries, it is a good way to assure them they have obtained a fair share of the spectrum. Consideration could also usefully be given to means of planning irregular networks in special cases where these may be more appropriate.

ANNEX I

CHOICE OF THE PROGRESSION STEPS IN AN OPTIMUM
157 CHANNEL LATTICE

A systematic study has been made of all the rhombic numbers less than 160, involving calculations of distance for 1-, 2- and 3- channel distances for all sets of values of p and q .

Those results that appear the most useful are shown below (Table I) (to simplify the presentation, the square of the distance is shown). The Table extends only to $N = 100$ to limit its size.

TABLE I

	p	q	Adjacent	2-channel	3-channel
$N = 19 (a = 3, b = 2)$	6	10	4	3	3
$N = 31 (a = 5, b = 1)$	7	27	7	7	3
$N = 37 (a = 4, b = 3)$	6	29	9	3	7
	9	25	9	7	3
$N = 39 (a = 5, b = 2)$	4	29	7	7	9
$N = 43 (a = 6, b = 1)$	5	13	12	7	3
$N = 49 (a = 5, b = 3)$	12	29	13	9	3
$N = 57 (a = 7, b = 1)$	11	37	13	7	9
$N = 61 (a = 5, b = 4)$	6	23	13	7	12
	15	27	16	13	3
$N = 67 (a = 7, b = 2)$	4	53	13	13	12
	8	39	19	13	3
	21	27	12	7	13
$N = 73 (a = 8, b = 1)$	5	33	19	7	12
	6	25	13	9	16
	7	17	19	19	3
	11	58	12	13	19
$N = 79 (a = 7, b = 3)$	4	17	13	19	21
	6	65	16	7	19
	11	27	19	9	12
	22	54	21	19	3
$N = 91 (a = 6, b = 5)$	4	68	16	19	21
	9	62	21	13	12
	15	73	25	21	3
$N = 91 (a = 9, b = 1)$	4	55	19	21	12
	6	37	19	9	21
	8	19	27	19	3
$N = 93 (a = 7, b = 4)$	5	61	25	13	9
	17	40	16	13	21
$N = 97 (a = 8, b = 3)$	6	81	19	13	21
	13	30	21	19	12
	26	60	28	21	3

As an example we give here the details of the calculations used to find the best progression steps for $N = 157$ channels.

As $157 = 12^2 + 12 \times 1 + 1^2$, the step p and q should be such that

$$12p + q = 0 \text{ modulo } 157$$

p and q must be greater than 3 so that transmitter coverages on channels C and $C + 1$, $C + 2$ or $C + 3$ do not overlap.

The first values to be tested are $p = 4$ and $q = 109$. Then $q - p = 105$ (see Fig. 11 for definition of p , q , $p - q$). The question now is to find out where channel 1 (or 156) falls, as here the co-channel rhombus of channel 0 is considered. As the lattice is homogeneous and regular, we can begin with any channel. The ideal location for channel 1 is at the centre of gravity of any of the two triangles which join up to form the co-channel rhombus. The adjacent channel distance would then be

$$\sqrt{157} \times \sqrt{3}/3 = 7.23$$

For $p = 4$ and $q = 109$, channel 1 will be found at three ($q - p$) steps: $q - p = 105$; $3 \times 105 = 315 = 1 \text{ modulo } 157$. Adjacent channel distance is 3, which is far shorter than the maximum of 7.23.

For $p = 5$ and $q = 97$ ($q - p = 92$) channel 156 will be found at five ($q - p$) steps plus two p steps: $5 \times 92 + 2 \times 5 = 470 = 156 \text{ modulo } 157$. Adjacent channel distance is $\sqrt{19}$. Hopefully a better solution exists.

For $p = 6$ and $q = 85$ ($q - p = 79$) channel 1 will be found at two steps, as $2(q - p) = 158 = 1 \text{ modulo } 157$. This is far too close.

For $p = 7$ and $q = 73$ ($q - p = 66$) channel 156 will be found at three p steps plus four q steps: $3 \times 7 + 4 \times 73 = 313 = 156 \text{ modulo } 157$. The 1-channel distance is $\sqrt{37} = 6.08$. Channels 156 and 1 are near the centres of gravity of the triangles as are channels 155 and 2; channels 154 and 3 at a distance of $\sqrt{12}$ from channel 0. A thorough search amongst all the pairs, p and q , shows that there are only a few sets of steps resulting in 1-channel distance greater than or equal to 6. They are given in the following Table II, together with the 2-channel and 3-channel distances.

TABLE II

p	q	Adjacent	2-channel	3-channel
7	73	$\sqrt{37}$	$\sqrt{39}$	$\sqrt{12}$
11	25	$\sqrt{48}$	$\sqrt{37}$	$\sqrt{3}$
20	74	$\sqrt{37}$	$\sqrt{21}$	$\sqrt{21}$
24	26	6	1 ⁽¹⁾	
30	111	$\sqrt{39}$	$\sqrt{19}$	$\sqrt{21}$
33	75	$\sqrt{39}$	$\sqrt{7}$	$\sqrt{48}$

(1) This eliminates the combination.

The final choice depends on the protection ratios for 2 and 3 channel interferences. The pair 24 and 26 will not be suitable.

If it were planned to cluster channels in groups of 3 or 4 (see Figs. 9 and 10) it would be more appropriate to have all steps greater than, say, 20 to avoid multiplexing problems (see [Arnaud, 1962]). Such a network is drawn on Fig. 11 ($p = 30$, $q = 111$). If no clustering is foreseen, or if a channel spacing of 20 is acceptable for colocated transmitters, then the network $p = 20$, $q = 74$ is slightly better.

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REPORT 945-2*

METHODS FOR THE ASSESSMENT OF MULTIPLE INTERFERENCE

(Question 46/10, Study Programme 46L/10)

(1982-1986-1990)

1. Introduction

The usable field strength, E_u , is defined in Recommendation 638 and may be used as a criterion for the interference situation in a given channel and in a given area. It takes account, in principle, of natural and man-made noise as well as of the combined effect of the entirety of interfering transmitters. If the usable field strength is large, the influence of the interfering transmitters is also large, whereas this influence is small if the usable field strength is small.

The usable field strength is independent of the characteristics of the wanted transmitter and does not normally exhibit large variations with location. It can be determined for any location of interest. For international planning purposes, it may be convenient to calculate the usable field strength at the site of the wanted transmitter. As a first approximation, this value may be considered to be representative of the situation in the whole coverage area.

For the calculation of these interferences basically two categories of methods are in use:

- statistical methods;
- non-statistical methods.

Normally use is made either of the 'simplified multiplication method' as an example of a statistical method or of the non-statistical 'power-sum method'. Experience has also been gained in the successful use of the statistical 'log-normal method' [Kubrakov *et al.*, 1985].

These three methods are described and indications are given for their efficient use.

The simplified multiplication method and the log-normal method incorporate the effects of location variability on the assumption that they follow a Gaussian law and are intended for use at VHF and higher frequencies. The power-sum method applies for point-to-point reception and must be applied successively with different receiver locations in coverage studies.

Details for the use of the simplified multiplication method and its physical bases are to be found in Annex I and in [O'Leary and Rutkowski, 1982], while similar information on the log-normal method is contained in Annex II and in [Bobkova and Pavliouk, 1987].

2. Power-sum method

Use has been made of the power-sum method for the assessment of multiple interference at the LF/MF Broadcasting Conference for Regions 1 and 3, Geneva, 1975, and at the Administrative Radio Conferences for the Broadcasting-Satellite Service, Geneva, 1977 and 1983. A similar method called the "RSS method" has been used at the Regional Administrative MF Broadcasting Conference (Region 2), Rio de Janeiro, 1981.

* This Report should be brought to the attention of Study Groups 1 and 11.

3. Statistical methods

Four statistical computation procedures which can be used to calculate the effects of multiple interference have been developed to date [*Ad hoc* Committee 1949, 1950]. These are:

- the integration method,
- the log-normal method,
- the multiplication method,
- the simplified multiplication method.

These procedures attempt to make use of the statistical (in locations) nature of propagation curves such as those given in Recommendation 370. Although, essentially based on a single theoretical approach, these procedures differ from one another as a result of different physical assumptions. These are made in each procedure in order to simplify the mathematical calculations. The use of any of these procedures results in distribution functions which describe the location probability of reception used in conjunction with Reports 228 and 485 to calculate coverage.

The simplified multiplication method and the log-normal method are the least complex. The former was used for the assessment of multiple interference at the VHF/UHF European and African Broadcasting Conferences in 1961 and 1963, and at the Regional Administrative Conference for FM Sound Broadcasting in the VHF band (Geneva, 1984).

3.1 The simplified multiplication method

This method is based on the following assumptions:

- no correlation exists between the fields of interest;
- the time dispersion of the field strength of the desired transmitter can be neglected, compared to that of the unwanted transmitter;
- one interfering field dominates at the reception location;
- the influence of noise can be neglected, to compensate for the errors introduced by the other approximations.

The usable field strength E_u , is determined for a specified coverage probability (with respect to time and location) and depends on the values of the nuisance fields:

$$E_{si} = P_i + E_{ni(50, T)} + A_i + B_i \quad (3)$$

where:

- E_{si} : nuisance field corresponding to the i -th unwanted transmitter.
- P_i : the e.r.p., in dB(kW), of the i -th unwanted transmitter.
- $E_{ni(50, T)}$: the field strength, in dB(μ V/m), normalized to an e.r.p. of 1 kW, of the i -th unwanted transmitter. This field strength is exceeded at 50% of the locations during at least $T\%$ of the time.
- A_i : the radio-frequency protection ratio associated with the i -th unwanted transmitter, expressed in dB.
- B_i : the receiving antenna discrimination, expressed in dB.

Assuming a normal distribution of the field strength, expressed in dB(μ V/m), appropriate account of the effect of multiple interference can be taken by the use of the simplified multiplication method. With this method the usable field strength, E_u , can be calculated by iteration from:

$$p_c = \prod_{i=1}^n L(x_i) \quad (4)$$

with:

$$x_i = \frac{E_u - E_{si}}{\sigma_n \sqrt{2}}$$

where:

- p_c : coverage probability (e.g. 50% of locations (100 - T)% of time), in the presence of n nuisance fields;
- $L(x)$: coverage probability in the presence of a single nuisance field, which equals the probability integral for a normal distribution (see Annex I);
- σ_n : standard deviation with location of the wanted and interfering field strengths (dB(μ V/m)) (see Annex I).

For further details see Annex I and [O'Leary and Rutkowski, 1982].

3.2 The log-normal method

This method is based on the following assumptions:

- no correlation exists between the fields of interest;
- the location variability in the field strength of the wanted and all interfering transmitters is taken to be identical;
- the effect of certain interfering field strengths, each of which obeys a log-normal law, is replaced at the point of reception by the effect of one resultant interfering field strength subject to the log-normal law with the following parameters [Fenton, 1960; Bobkova and Pavliouk, 1987]:

$$E_r = 0.1152 \sigma^2 + 10 \log \left(\sum_i M_{si} \right) - 5 \log U$$

$$\sigma_r = 6.58 \sqrt{\log U}$$

$$U = \frac{(k - 1) \sum_i M_{si}^2}{\left(\sum_i M_{si} \right)^2} + 1$$

$$k = \exp \left[\left(\sigma / 4,34 \right)^2 \right]$$

$$M_{si} = 10^{E_{si}/10}$$

- where: E_{si} - nuisance field corresponding to the i th unwanted transmitter, given by equation (3);
- E_r - median value of resultant nuisance fields, expressed in decibels;
- $\sigma; \sigma_r$ - standard deviations with location of interfering field strengths and resultant interfering field strength respectively, expressed in decibels;

- the usable field strength (for purposes of sound and TV broadcasting planning) is defined for L% of locations, at which the following conditions are simultaneously fulfilled:

a) the wanted signal exceeds the noise level by the required amount, and

b) the necessary protection ratio, i.e.:

$$E_u \geq E_{\min} \text{ (50\% of locations and 50\% of time)}$$

$$E_u \geq E_{si} \text{ (50\% of locations and T\% of time),}$$

Condition b) is fulfilled when the usable field strength E_{u1} is equal to E_r . The probability of the fulfilment of both conditions a) and b) is equal to the product of their probabilities:

$$P_c = P_1 \cdot P_2 = L \left(\frac{E_{u1} - E_r}{\sqrt{\sigma_u^2 + \sigma_r^2}} \right) \cdot L \left(\frac{E_{u1} - E_{\min}}{\sigma} \right)$$

where: P_1, P_2 - probabilities of fulfilment of conditions a) and b) respectively; their values are determined by a normalized distribution curve for the normal law.

and, $L(x)$ is the probability integral, see equation (8) of Annex I.

If the value found for p_c satisfies the predetermined value of the coverage probability, p_{cp} (generally $p_{cp} = 0.5$, i.e., E_u is defined for 50% of the points of reception), then $E_u = E_{u1}$ and the calculation is completed. However, if $p_c \neq 0.5$, the further calculation of the usable field strength is performed by iteration using the formula:

$$E_u = E_{u1} + \frac{0.5 - p_c}{0.05}$$

The detailed calculation is shown in Annex II and in [Bobkova and Pavliouk, 1987].

The simplified log-normal method

If it is assumed, in the same way as for the power-sum method, that the effect of man-made interference and other noise may be covered by taking into account the minimum usable field-strength as a source of interference, the fluctuation in the corresponding field-strength also being subject to the log-normal law, this method may be simplified to give a non-iterative procedure for the determination of the usable field-strength for 50% of the receiving locations [Bobkova, 1988].

On this assumption, $E_{\min}$ may be introduced into the formula for determining the resultant field-strength as an additional source of interference:

$$E_{rs} = 0.1152 \sigma^2 + 10 \lg \left(\sum_i M_i + M_{\min} \right) - 5 \lg U_s,$$

$$\text{where } U_s = \frac{(k-1) \left(\sum_i M_i^2 + M_{\min}^2 \right)}{\left(\sum_i M_i + M_{\min} \right)^2} + 1,$$

$M_{\min}$ - median value of minimum usable field-strength, expressed in relative power units.

The value obtained for E_{rs} directly determines the median usable field-strength (i.e., for 50% of receiving locations). The usable field-strength value for any other percentage of locations required may be obtained by the same iterative procedure as that used in the standard log-normal method, by means of the formula:

$$P_{cp} = L \left(\frac{E_u - E_{rs}}{\sqrt{\sigma^2 + \sigma_{rs}^2}} \right)$$

where P_{cp} - predetermined value of the coverage probability

σ_{rs} - r.m.s. deviation of the resultant interference field, determined by the above formula, substituting into it the value U_s .

4. Comparison of the results obtained by the various methods

Sections 4.1 and 4.2 relate in particular to the simplified multiplication method and the power-sum method whilst section 4.3 includes considerations of the log-normal method.

4.1 *General considerations*

The comparison of calculations, using the two methods under equal conditions (no receiving antenna discrimination) is given below. This comparison is intended to enable the reader to estimate the difference to be expected, when applying one method or the other, rather than to advocate a specific method. It should be noted in this respect that either method can be expected to yield only approximate values of the usable field strength.

Differences between the results of both methods may be considerable, and the utmost care should be exercised when deriving further conclusions from these results, e.g. with respect to the number of channels required for satisfactory coverage by one programme.

It should, however, also be recognized that it would be possible to use, with comparable degree of precision, either method as a basis for comparison of different variants of a plan.

4.2 *Results*

Figures 1 and 2 exhibit the differences ΔE in usable field strengths when computed according to the simplified multiplication method (E_M) for a percentage of receiving locations of 45% and 50% respectively, or to the power sum method (E_P) for an undefined percentage of receiving locations. The differences ΔE depend on the relationship between the interference potentials of the various individual transmitters, i.e. their technical characteristics, their distance from the wanted transmitter and the frequency spacing. Normally there is one predominant source of interference producing nuisance field E_1 , following by a second and less harmful source of interference creating nuisance field E_2 and the remaining interfering transmitters causing gradually decreasing nuisance fields $E_3 \dots E_n$.



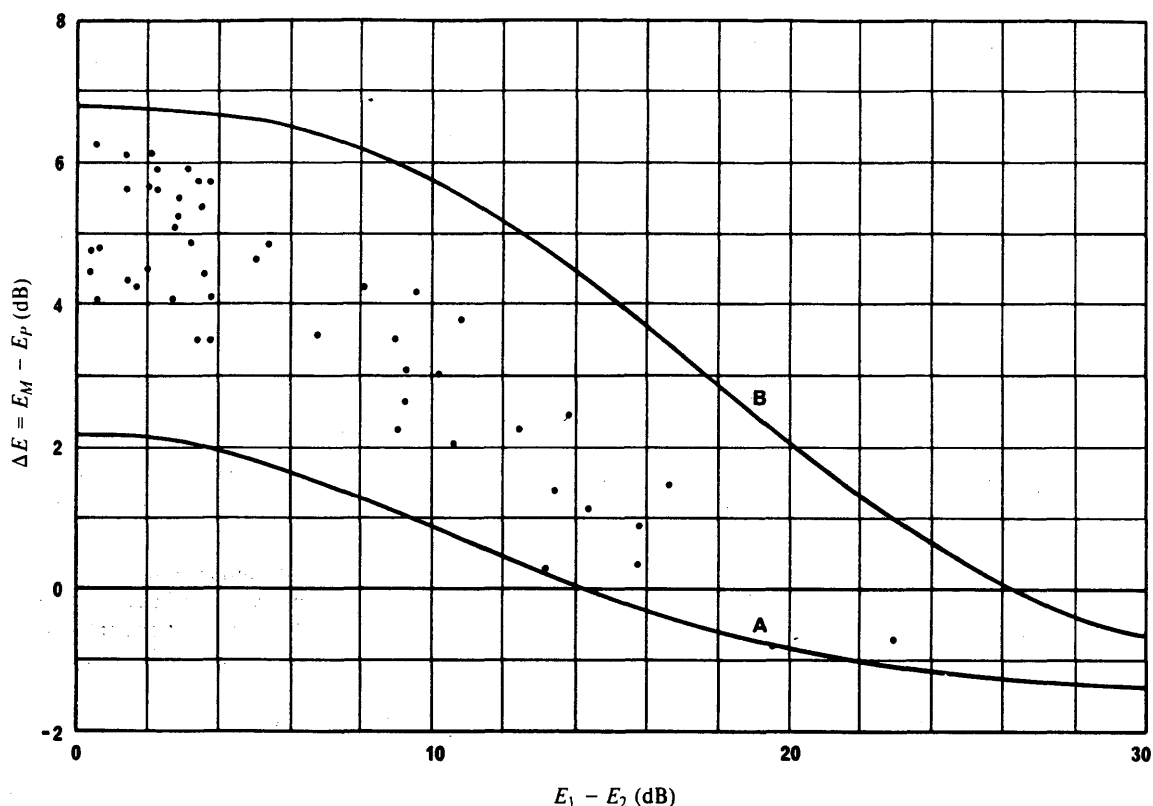


FIGURE 1 - Difference in usable field strength, ΔE , as a function of the difference between nuisance fields, E_1 and E_2 . Percentage of locations for the simplified multiplication method: 45%.

- : 50 transmitters
- $E_1 - E_2$: difference between nuisance fields of the two strongest interfering transmitters
- $\Delta E = E_M - E_P$: difference between the values of usable field strength obtained with either the simplified multiplication method (E_M), or the power sum method (E_P).
- Curves A : limiting curve obtained for the case of 2 nuisance fields, E_1 and E_2
- B : limiting curve obtained for the case of 10 nuisance fields, E_1 and 9 times the value of E_2

In Figs. 1 and 2 values of

$$\Delta E = E_M - E_P = f(E_1 - E_2)$$

are plotted for 50 ΔE values obtained for the first (in alphabetical order) 50 out of 345 transmitters operating at present in the Federal Republic of Germany. These 50 values are thought to be representative of these 345 VHF/FM transmitters. Moreover, two limiting curves, A and B, have been included between which all the 50 values obtained are situated. These limiting curves take account of the n interfering sources $E_1 \dots E_n$ in different ways: for the lower limiting curve (A) only nuisance fields E_1 and E_2 are taken into account, whereas for the upper limiting curve (B) there are $(n - 2) = 8$ additional sources of interference equal in severity with E_2 :

$$E_2 = E_3 = E_4 = \dots = E_{10}$$

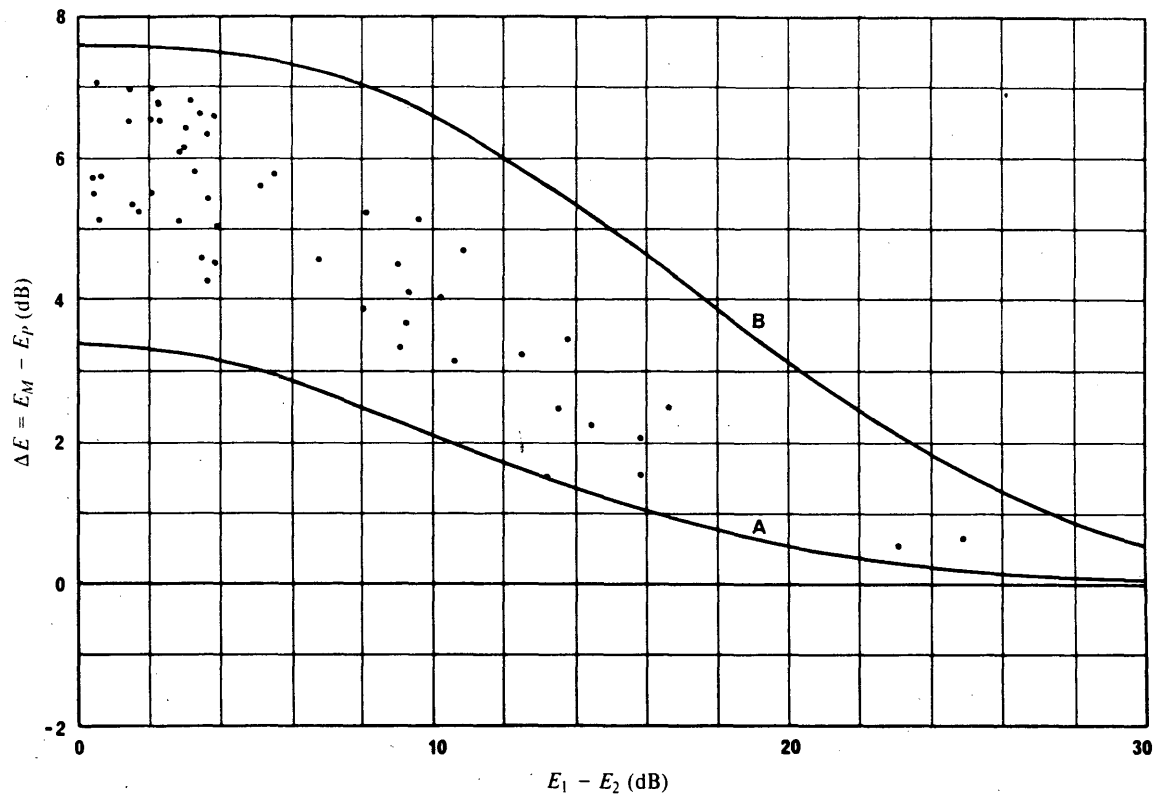


FIGURE 2 - Difference in usable field strength, ΔE , as a function of the difference between nuisance fields, E_1 and E_2 . Percentage of locations for the simplified multiplication method: 50%.

• : 50 transmitters

$E_1 - E_2$: difference between nuisance fields of the two strongest interfering transmitters

$\Delta E = E_M - E_P$: difference between the values of usable field strength obtained with either the simplified multiplication method (E_M), or the power sum method (E_P)

Curves A : limiting curve obtained for the case of 2 nuisance fields, E_1 and E_2

B : limiting curve obtained for the case of 10 nuisance fields, E_1 and 9 times the value of E_2

It can be seen from Figs. 1 and 2 that in the majority of cases the simplified multiplication method yields values which are up to about 7 dB higher than those obtained with the power sum method, depending on the network configuration.

Almost identical results were obtained from an analysis of usable field-strengths of 1177 UK assignments obtained in the Region 1 VHF/FM Plan, Geneva, 1984 as indicated below:

Percentage of assignments	Ratio (dB) of usable field-strengths exceeded Simplified multiplication: power sum
4	7
50	5
85	3

Note. - Calculations took account of the first 20 interfering sources.

- Results for simplified multiplication method based on 50% location probability.

It should be noted that all the above comparisons relate to the VHF bands. Differences between the two methods will be greater at UHF because of the larger values of standard deviation with location.

4.3 Consideration of location correlation between the fields

If there is no location correlation between the wanted and interfering fields, the overall standard deviation of variation with location is derived from:

$$\sigma = \sqrt{\sigma_n^2 + \sigma_i^2} = \sigma_n \sqrt{2} \quad (5)$$

where σ_n and σ_i , the standard deviations with location of wanted and interfering signals respectively, are considered identical and equal to:

- 8.3 dB for bands I to III,
- $9.5 + 0.405 g$ dB in bands IV and V (g being a function of Δh , see Annex I, § 2).

However, equation (5) is the particular case, for location correlation coefficient $\rho = 0$, of the general expression:

$$\sigma = \sqrt{\sigma_n^2 - 2\rho\sigma_n\sigma_i + \sigma_i^2}$$

Positive values of ρ will thus result in reduced values of σ and hence also of resultant usable field-strength when using the simplified multiplication method.

Figure 3 shows the relationship between the usable field strengths obtained by the simplified multiplication method (s.m.m), the power-sum method and the log-normal method. For simplicity, the calculations are based on the case in which all interfering field strengths are of equal magnitude. The data given in Figure 3 and also in [Bobkova and Pavliouk, 1987] show that the log-normal method consistently provides intermediate values between the results calculated by the simplified multiplication method (limit case when $\rho = 0$) and the power-sum method. The curve C (log-normal method) is closest to the curve B. (s.m.m. for $\rho = 0.25$).

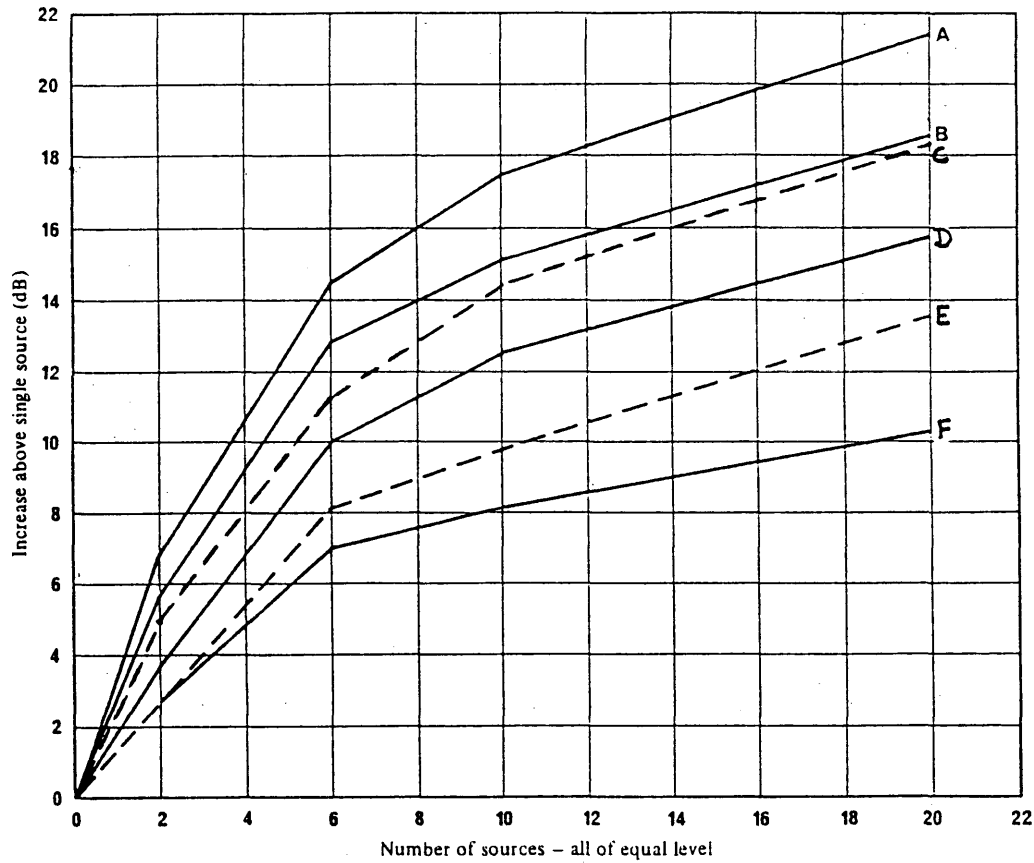


FIGURE 3

- A: s.m.m. for $\rho = 0$
- B: s.m.m. for $\rho = 0.25$
- C: log-normal method
- D: s.m.m. for $\rho = 0.5$
- E: power-sum method
- F: s.m.m. for $\rho = 0.75$

A series of tests has been carried out in the United Kingdom, at both VHF and UHF, to establish the typical values of ρ occurring in practice. Most of these gave results within the range 0.25 to 0.75 with a tendency for the higher values to occur when signals arrived from the same direction; also, as might be expected, for values lower than the above range (but still positive), to occur when one of the transmitters lies within the area under investigation. "

Figure 4 shows the difference in the values of usable field-strength (ΔE_u) obtained by the log-normal method (E_r) and by the simplified (E_{rs}) log-normal method $\Delta E_u = E_r - E_{rs}$ for the case in which there are six equivalent interfering fields at the receiving location, as occurs in the case of the regular network (Report 944). The abscissa shows the difference in the values of (ΔE_s) interfering field-strength E_{si} and the minimum usable field-strength: $\Delta E_s = E_{si} - E_{min}$.

Figure 4 shows that for the case $E_{si} \geq E_{min} + 2$ dB, both methods produce identical results: when $E_{si} = E_{min}$, the difference is only 0.1 dB. The maximum difference, 1.7 dB, is obtained when $E_{si} \leq E_{min} - 10$ dB, i.e. when the active interference level is very low and the size of the wanted transmitter's service area depends entirely and exclusively on the selected value of minimum usable field-strength.

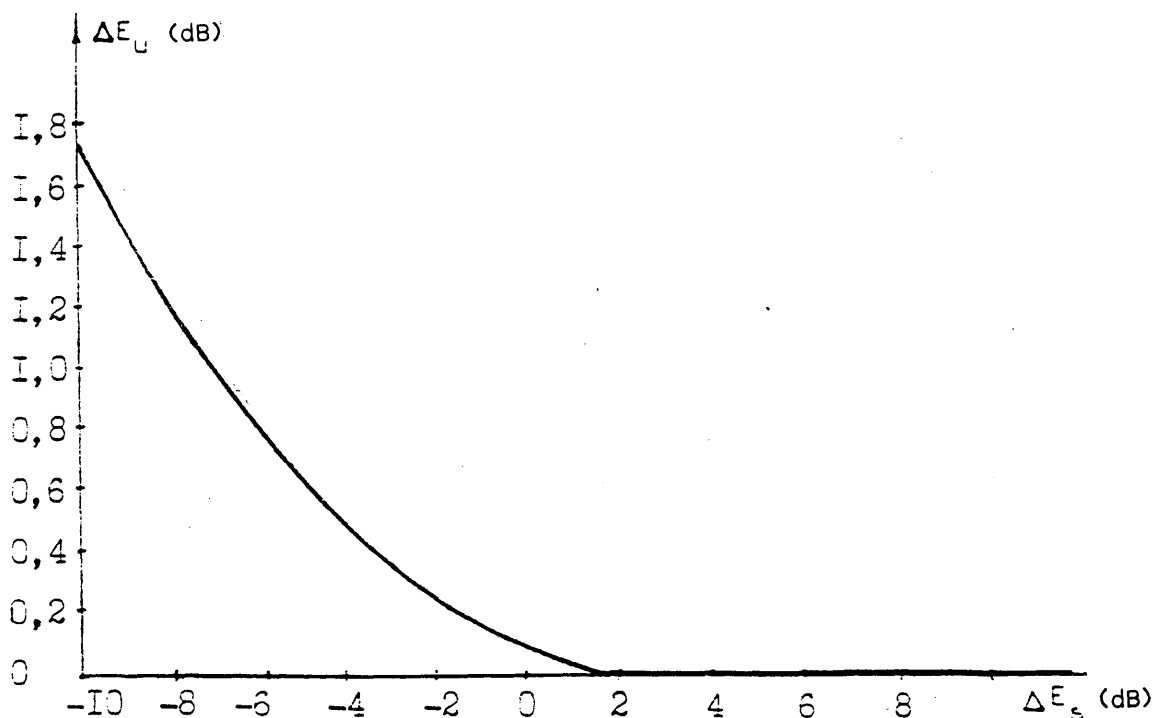


FIGURE 4

Difference between the usable field-strength values obtained by the standard log-normal method and the simplified log-normal method

5. Conclusions

The simplified multiplication method, the log-normal method and the simplified log-normal method are statistical in nature and can be used for interference assessments for any desired percentage of locations. In calculations for 50% of the receiving locations and 50% of the time, the simplified log-normal method is the most economical of these statistical methods in terms of calculation time. The power-sum method is likewise fairly simple, but its use leads to results which tend to be more optimistic than those obtained with the simplified multiplication method.

The inclusion of realistic values of correlation coefficient in a modified simplified multiplication method would substantially reduce the differences between this and the power-sum method, especially in those cases where the correlation coefficient lies between 0.5 and 0.75. However, further studies are needed to provide values of the location correlation coefficient, ρ , which can be applied in different types of terrain.

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ANNEX I

HOW TO USE THE SIMPLIFIED MULTIPLICATION METHOD FOR CALCULATING USABLE FIELD STRENGTHS (IN FM SOUND BROADCASTING)

1. Introduction

It has been proposed, on an international level [CCIR, 1961], to determine the influence of interfering transmitters (co-channel, adjacent channel and image channel) by means of the simplified multiplication method, which was developed by the [Ad hoc Committee 1949, 1950] and is described in detail in [Grosskopf, 1952]. In the following, a step-by-step explanation of the method is given for the practical user without deeper theoretical justification.

2. The concept of the usable field strength

The usable field strength E_u , is a quantity characterizing the coverage situation. To calculate the usable field strength, it is necessary to determine all those transmitters:

- which lie within a definite range of the wanted transmitter (according to experience: up to 800 km), according to the value of Δf);
- which might cause interference in relation to the required protection ratio (A_i).

For the n interfering transmitters, so determined, the nuisance field, E_{si} , is:

$$E_{si} = P_i + E_{ni(50, T)} + A_i + B_i \quad (6)$$

where:

- $E_{ni(50, T)}$: field-strength (dB(μ V/m)) of the unwanted signal normalized to 1 kW effective radiated power (e.r.p.) at 50% locations for $T\%$ time (from field-strength curves of Recommendation 370);
- P_i : e.r.p. (dB(kW)) of the interfering transmitter;
- A_i : protection ratio (dB);
- B_i : receiving antenna discrimination (dB).

The usable field strength, E_u , is a function of the n nuisance fields, E_{si} , and is calculated according to the formula:

$$p_c = \prod_{i=1}^n L(x_i) \quad (7)$$

with:

$$x_i = \frac{E_u - E_{si}}{\sigma_n \sqrt{2}}$$

where:

p_c : the coverage probability. To initiate the iterative process of calculating E_u a predetermined value, p_{cp} , of the coverage probability is given, e.g. $p_{cp} = 0.5$. With the value of E_u obtained at the end of the iterative process the coverage probability is $p_c = p_{cp} = 0.5$, i.e. 50% of locations.

Note. — p_c can be set to any other value of coverage probability (e.g. 45% $\rightarrow p_c = 0.45$).

L : the probability integral for a normal distribution:

$$L(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^x [\exp(-t^2/2)] dt \quad (8)$$

In this function x is the difference between the levels of the usable field strength, E_u , and the nuisance field, E_{si} , related to σ , the standard deviation (with location) of the resulting difference in level.

Identical values are assumed for the standard deviations (with location) of the wanted and interfering field-strength levels: $\sigma_n = \sigma_s$. Thus, the standard deviation of the resulting level difference is:

$$\sigma = \sqrt{\sigma_n^2 + \sigma_s^2} = \sigma_n \sqrt{2}$$

The value $\sigma_n = 8.3$ dB is assumed for the frequency bands I to III. For band IV/V this value is dependent on the terrain attenuation, g , and σ is then calculated according to the formula $\sigma_n = 9.5 + 0.405 g$. The attenuation correction factor g (dB) can be derived from Δh (see Recommendation 370).

3. Calculation of the probability integral

3.1 Tabular evaluation

The probability integral in the form:

$$\varphi(x) = \frac{2}{\sqrt{2\pi}} \int_0^x [\exp(-t^2/2)] dt \quad (9)$$

can be found evaluated in Table I.

Since

$$\frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} [\exp(-t^2/2)] dt = 1$$

and

$$\frac{1}{\sqrt{2\pi}} \int_{-\infty}^0 [\exp(-t^2/2)] dt = 1/2$$

it follows that:

$$L(x) = \frac{\varphi(x)}{2} + 1/2$$

3.2 Evaluation using Hastings approximation

If the calculations are to be done with a computer (or programmable pocket or table calculator) the following rational approximation is very useful:

$$\begin{aligned} \text{for } x \geq 0: \quad L(x) &= 1 - \frac{1}{(2\pi)^{1/2}} e^{-x^2/2} H(y) \\ \text{for } x < 0: \quad L(x) &= 1 - L(-x) \end{aligned} \quad (10)$$

where:

$$H(y) = C_5 y^5 + C_4 y^4 + C_3 y^3 + C_2 y^2 + C_1 y^1$$

$$\text{and: } y = [1 + 0.2316419 |x|]^{-1}$$

$$C_5 = 1.330274429$$

$$C_4 = -1.821255978$$

$$C_3 = 1.781477937$$

$$C_2 = -0.356563782$$

$$C_1 = 0.319381530$$

By means of equation (10) the integration in equation (8) and also the use of tables can be avoided when evaluating the probability integral. The error involved by using this approximation is less than 10^{-7} .

4. Practical calculation procedures to determine the usable field strength

Since it is impossible to solve equation (7) explicitly for E_u for a predetermined value p_{cp} (e.g. $p_{cp} = 0.5$) it must be solved iteratively. We begin with an initial value for E_u , which, according to experience, should be some 6 dB larger than the largest of the E_{si} , and determine, successively, for each E_{si} :

$$z_i = E_u - E_{si} = \Delta_i$$

$$x_i = \frac{\Delta_i}{\sigma_n \sqrt{2}} \quad (\text{in bands I to III: } x_i = \Delta_i / 11.738)$$

$\varphi(x_i)$ from Table I

$$L(x_i) = \frac{\varphi(x_i)}{2} + \frac{1}{2}$$

As for the standard deviation, a value $\sigma_n = 8.3$ dB is assumed to apply for bands I to III, it seems appropriate to introduce Table II where $L(x_i)$ is presented as a function of Δ_i for $\sigma_n = 8.3$ dB. In bands IV and V, where $\sigma_n = 9.5 + 0.405 g$, Table II may also be used once the Δ_i values have been corrected according to:

$$\Delta'_i = \Delta_i \cdot \frac{8.3}{9.5 + 0.405 g}$$

p_c is then determined by means of equation (7). If p_c is different from p_{cp} (e.g. $p_{cp} = 0.5$), the value so obtained is used as a basis to correct, as a part of the iterative process, the initial E_u value. From experience, the correction may be assumed to correspond approximately to:

$$\Delta E_u \approx \frac{p_{cp} - p_c}{0.05} \text{ dB}$$

Then the determination of E_u has to be continued by repeating, with the corrected E_u , the determination of new Δ_i and $L(x_i)$ for each E_{si} and of a new p_c . This procedure has to be carried out until the correction ΔE_u falls below the accuracy limit. Table III gives an example for the iterative determination of E_u in the presence of 5 nuisance fields ($\sigma_n = 8.3$ dB). The values of $L(x_i)$ are taken from Table II.

TABLE I - Probability integral

$$\varphi(x) = \frac{2}{\sqrt{2\pi}} \int_0^x [\exp(-t^2/2)] dt$$

x	$\varphi(x)$	x	$\varphi(x)$	x	$\varphi(x)$	x	$\varphi(x)$
0.00	0.0000	0.60	0.4515	1.20	0.7699	1.80	0.9281
01	0.0080	61	0.4581	21	0.7737	81	0.9297
02	0.0160	62	0.4647	22	0.7775	82	0.9312
03	0.0239	63	0.4713	23	0.7813	83	0.9328
04	0.0319	64	0.4778	24	0.7850	84	0.9342
0.05	0.0399	0.65	0.4843	1.25	0.7887	1.85	0.9357
06	0.0478	66	0.4907	26	0.7923	86	0.9371
07	0.0558	67	0.4971	27	0.7959	87	0.9385
08	0.0638	68	0.5035	28	0.7995	88	0.9399
09	0.0717	69	0.5098	29	0.8029	89	0.9412
0.10	0.0797	0.70	0.5161	1.30	0.8064	1.90	0.9426
11	0.0876	71	0.5223	31	0.8098	91	0.9439
12	0.0955	72	0.5285	32	0.8132	92	0.9451
13	0.1034	73	0.5346	33	0.8165	93	0.9464
14	0.1113	74	0.5407	34	0.8198	94	0.9476
0.15	0.1192	0.75	0.5467	1.35	0.8230	1.95	0.9488
16	0.1271	76	0.5527	36	0.8262	96	0.9500
17	0.1350	77	0.5587	37	0.8293	97	0.9512
18	0.1428	78	0.5646	38	0.8324	98	0.9523
19	0.1507	79	0.5705	39	0.8355	99	0.9534
0.20	0.1585	0.80	0.5763	1.40	0.8385	2.00	0.9545
21	0.1663	81	0.5821	41	0.8415	05	0.9596
22	0.1741	82	0.5878	42	0.8444	10	0.9643
23	0.1819	83	0.5935	43	0.8473	15	0.9684
24	0.1897	84	0.5991	44	0.8501	20	0.9722
0.25	0.1974	0.85	0.6047	1.45	0.8529	2.25	0.9756
26	0.2041	86	0.6102	46	0.8557	30	0.9786
27	0.2128	87	0.6157	47	0.8584	35	0.9812
28	0.2205	88	0.6211	48	0.8611	40	0.9836
29	0.2282	89	0.6265	49	0.8638	45	0.9857
0.30	0.2358	0.90	0.6319	1.50	0.8664	2.50	0.9876
31	0.2434	91	0.6372	51	0.8690	55	0.9892
32	0.2510	92	0.6424	52	0.8715	60	0.9907
33	0.2586	93	0.6476	53	0.8740	65	0.9920
34	0.2661	94	0.6528	54	0.8764	70	0.9931
0.35	0.2737	0.95	0.6579	1.55	0.8789	2.75	0.9940
36	0.2812	96	0.6629	56	0.8812	80	0.9949
37	0.2886	97	0.6680	57	0.8836	85	0.9956
38	0.2961	98	0.6729	58	0.8859	90	0.9963
39	0.3035	99	0.6778	59	0.8882	95	0.9968
0.40	0.3108	1.00	0.6827	1.60	0.8904	3.00	0.99730
41	0.3182	01	0.6875	61	0.8926	10	0.99806
42	0.3255	02	0.6923	62	0.8948	20	0.99863
43	0.3328	03	0.6970	63	0.8969	30	0.99903
44	0.3401	04	0.7017	64	0.8990	40	0.99933
0.45	0.3473	1.05	0.7063	1.65	0.9011	3.50	0.99953
46	0.3545	06	0.7109	66	0.9031	60	0.99968
47	0.3616	07	0.7154	67	0.9051	70	0.99978
48	0.3688	08	0.7199	68	0.9070	80	0.99986
49	0.3759	09	0.7243	69	0.9090	90	0.99990
0.50	0.3829	1.10	0.7287	1.70	0.9109	4.00	0.99994
51	0.3899	11	0.7330	71	0.9127	4.417	1 - 10 ⁻⁵
52	0.3969	12	0.7373	72	0.9146		
53	0.4039	13	0.7415	73	0.9164		
54	0.4108	14	0.7457	74	0.9181		
0.55	0.4177	1.15	0.7499	1.75	0.9199	4.892	1 - 10 ⁻⁶
56	0.4245	16	0.7540	76	0.9216		
57	0.4313	17	0.7580	77	0.9233		
58	0.4381	18	0.7620	78	0.9249		
59	0.4448	19	0.7660	79	0.9265	5.327	1 - 10 ⁻⁷
0.60	0.4515	1.20	0.7699	1.80	0.9281		

TABLE II

Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$
.0	.50000	7.000	5.0	.66493	4.121	10.0	.80288	2.217	15.0	.89936	1.071	20.0	.95580	.457
.1	.50340	6.932	5.1	.66803	4.074	10.1	.80523	2.188	15.1	.90085	1.054	20.1	.95659	.448
.2	.50680	6.864	5.2	.67112	4.028	10.2	.80757	2.158	15.2	.90233	1.038	20.2	.95737	.440
.3	.51020	6.796	5.3	.67419	3.981	10.3	.80989	2.129	15.3	.90379	1.022	20.3	.95813	.432
.4	.51359	6.729	5.4	.67726	3.936	10.4	.81219	2.101	15.4	.90524	1.005	20.4	.95889	.424
.5	.51699	6.663	5.5	.68031	3.890	10.5	.81448	2.072	15.5	.90667	.989	20.5	.95964	.416
.6	.52038	6.596	5.6	.68335	3.845	10.6	.81675	2.044	15.6	.90808	.974	20.6	.96037	.408
.7	.52378	6.531	5.7	.68638	3.801	10.7	.81900	2.016	15.7	.90948	.958	20.7	.96109	.401
.8	.52717	6.466	5.8	.68939	3.756	10.8	.82124	1.989	15.8	.91086	.943	20.8	.96180	.393
.9	.53056	6.401	5.9	.69239	3.712	10.9	.82345	1.962	15.9	.91222	.928	20.9	.96251	.386
1.0	.53395	6.337	6.0	.69538	3.669	11.0	.82565	1.935	16.0	.91357	.913	21.0	.96320	.379
1.1	.53733	6.273	6.1	.69836	3.626	11.1	.82784	1.908	16.1	.91491	.898	21.1	.96388	.372
1.2	.54071	6.209	6.2	.70132	3.583	11.2	.83000	1.882	16.2	.91623	.884	21.2	.96455	.365
1.3	.54409	6.147	6.3	.70427	3.541	11.3	.83215	1.856	16.3	.91753	.869	21.3	.96521	.358
1.4	.54747	6.084	6.4	.70721	3.499	11.4	.83428	1.830	16.4	.91882	.855	21.4	.96586	.351
1.5	.55084	6.022	6.5	.71013	3.457	11.5	.83639	1.804	16.5	.92009	.841	21.5	.96650	.344
1.6	.55421	5.960	6.6	.71304	3.416	11.6	.83848	1.779	16.6	.92135	.827	21.6	.96713	.338
1.7	.55758	5.899	6.7	.71593	3.375	11.7	.84056	1.754	16.7	.92259	.814	21.7	.96775	.331
1.8	.56094	5.839	6.8	.71881	3.334	11.8	.84262	1.729	16.8	.92382	.800	21.8	.96836	.325
1.9	.56430	5.778	6.9	.72168	3.294	11.9	.84466	1.705	16.9	.92503	.787	21.9	.96896	.318
2.0	.56765	5.719	7.0	.72453	3.254	12.0	.84669	1.681	17.0	.92623	.774	22.0	.96955	.312
2.1	.57099	5.659	7.1	.72737	3.215	12.1	.84869	1.657	17.1	.92742	.761	22.1	.97013	.306
2.2	.57434	5.600	7.2	.73019	3.176	12.2	.85068	1.633	17.2	.92858	.748	22.2	.97071	.300
2.3	.57767	5.542	7.3	.73300	3.137	12.3	.85265	1.610	17.3	.92974	.736	22.3	.97127	.294
2.4	.58100	5.484	7.4	.73579	3.098	12.4	.85461	1.587	17.4	.93088	.723	22.4	.97183	.289
2.5	.58433	5.426	7.5	.73857	3.060	12.5	.85654	1.564	17.5	.93200	.711	22.5	.97237	.283
2.6	.58765	5.369	7.6	.74134	3.023	12.6	.85846	1.541	17.6	.93312	.699	22.6	.97291	.277
2.7	.59096	5.312	7.7	.74408	2.985	12.7	.86036	1.519	17.7	.93421	.687	22.7	.97344	.272
2.8	.59427	5.256	7.8	.74682	2.948	12.8	.86225	1.497	17.8	.93530	.676	22.8	.97396	.266
2.9	.59757	5.200	7.9	.74954	2.912	12.9	.86412	1.475	17.9	.93637	.664	22.9	.97447	.261
3.0	.60086	5.144	8.0	.75224	2.875	13.0	.86596	1.453	18.0	.93742	.653	23.0	.97497	.256
3.1	.60415	5.089	8.1	.75492	2.839	13.1	.86780	1.432	18.1	.93846	.641	23.1	.97546	.251
3.2	.60743	5.035	8.2	.75760	2.804	13.2	.86961	1.411	18.2	.93949	.630	23.2	.97595	.246
3.3	.61070	4.980	8.3	.76025	2.768	13.3	.87141	1.390	18.3	.94051	.619	23.3	.97643	.241
3.4	.61396	4.926	8.4	.76289	2.733	13.4	.87319	1.369	18.4	.94151	.609	23.4	.97690	.236
3.5	.61722	4.873	8.5	.76551	2.699	13.5	.87495	1.349	18.5	.94250	.598	23.5	.97736	.231
3.6	.62046	4.820	8.6	.76812	2.664	13.6	.87670	1.329	18.6	.94347	.588	23.6	.97781	.227
3.7	.62370	4.768	8.7	.77071	2.630	13.7	.87843	1.309	18.7	.94443	.577	23.7	.97826	.222
3.8	.62693	4.715	8.8	.77328	2.597	13.8	.88014	1.289	18.8	.94538	.567	23.8	.97870	.217
3.9	.63015	4.664	8.9	.77584	2.563	13.9	.88183	1.270	18.9	.94632	.557	23.9	.97913	.213
4.0	.63336	4.612	9.0	.77838	2.530	14.0	.88351	1.251	19.0	.94724	.547	24.0	.97956	.209
4.1	.63657	4.561	9.1	.78091	2.497	14.1	.88517	1.232	19.1	.94815	.538	24.1	.97997	.204
4.2	.63976	4.511	9.2	.78342	2.465	14.2	.88681	1.213	19.2	.94905	.528	24.2	.98038	.200
4.3	.64294	4.461	9.3	.78591	2.433	14.3	.88844	1.195	19.3	.94994	.519	24.3	.98078	.196
4.4	.64611	4.411	9.4	.78838	2.401	14.4	.89005	1.176	19.4	.95081	.509	24.4	.98118	.192
4.5	.64928	4.362	9.5	.79084	2.370	14.5	.89164	1.158	19.5	.95167	.500	24.5	.98157	.188
4.6	.65243	4.313	9.6	.79328	2.339	14.6	.89322	1.140	19.6	.95252	.491	24.6	.98195	.184
4.7	.65557	4.264	9.7	.79571	2.308	14.7	.89478	1.123	19.7	.95336	.482	24.7	.98232	.180
4.8	.65870	4.216	9.8	.79811	2.277	14.8	.89632	1.105	19.8	.95418	.474	24.8	.98269	.176
4.9	.66182	4.168	9.9	.80050	2.247	14.9	.89785	1.088	19.9	.95500	.465	24.9	.98305	.173

TABLE II (continued)

Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$	Δ	$L(x)$	$-\log L(x)$
25.0	.98341	.169	30.0	.99470	.054	35.0	.99857	.014	40.0	.99967	.003	45.0	.99994	.001
25.1	.98376	.165	30.1	.99483	.052	35.1	.99861	.014	40.1	.99968	.003	45.1	.99994	.001
25.2	.98410	.162	30.2	.99496	.051	35.2	.99864	.014	40.2	.99969	.003	45.2	.99994	.001
25.3	.98443	.158	30.3	.99508	.050	35.3	.99868	.013	40.3	.99970	.003	45.3	.99994	.001
25.4	.98476	.155	30.4	.99520	.049	35.4	.99872	.013	40.4	.99971	.003	45.4	.99995	.001
25.5	.98509	.152	30.5	.99532	.047	35.5	.99875	.013	40.5	.99972	.003	45.5	.99995	.001
25.6	.98541	.148	30.6	.99543	.046	35.6	.99879	.012	40.6	.99973	.003	45.6	.99995	.001
25.7	.98572	.145	30.7	.99554	.045	35.7	.99882	.012	40.7	.99974	.003	45.7	.99995	.000
25.8	.98603	.142	30.8	.99565	.044	35.8	.99886	.012	40.8	.99975	.003	45.8	.99995	.000
25.9	.98633	.139	30.9	.99576	.043	35.9	.99889	.011	40.9	.99975	.002	45.9	.99995	.000
26.0	.98662	.136	31.0	.99587	.042	36.0	.99892	.011	41.0	.99976	.002	46.0	.99996	.000
26.1	.98691	.133	31.1	.99597	.041	36.1	.99895	.011	41.1	.99977	.002	46.1	.99996	.000
26.2	.98719	.130	31.2	.99607	.040	36.2	.99898	.010	41.2	.99978	.002	46.2	.99996	.000
26.3	.98747	.127	31.3	.99617	.039	36.3	.99901	.010	41.3	.99978	.002	46.3	.99996	.000
26.4	.98775	.125	31.4	.99626	.038	36.4	.99904	.010	41.4	.99979	.002	46.4	.99996	.000
26.5	.98802	.122	31.5	.99636	.037	36.5	.99906	.009	41.5	.99980	.002	46.5	.99996	.000
26.6	.98828	.119	31.6	.99645	.036	36.6	.99909	.009	41.6	.99980	.002	46.6	.99996	.000
26.7	.98854	.116	31.7	.99654	.035	36.7	.99912	.009	41.7	.99981	.002	46.7	.99997	.000
26.8	.98879	.114	31.8	.99663	.034	36.8	.99914	.009	41.8	.99982	.002	46.8	.99997	.000
26.9	.98904	.111	31.9	.99671	.033	36.9	.99917	.008	41.9	.99982	.002	46.9	.99997	.000
27.0	.98928	.109	32.0	.99680	.032	37.0	.99919	.008	42.0	.99983	.002	47.0	.99997	.000
27.1	.98952	.106	32.1	.99688	.032	37.1	.99921	.008	42.1	.99983	.002	47.1	.99997	.000
27.2	.98976	.104	32.2	.99696	.031	37.2	.99924	.008	42.2	.99984	.002	47.2	.99997	.000
27.3	.98999	.102	32.3	.99704	.030	37.3	.99926	.007	42.3	.99984	.002	47.3	.99997	.000
27.4	.99021	.099	32.4	.99711	.029	37.4	.99928	.007	42.4	.99985	.002	47.4	.99997	.000
27.5	.99043	.097	32.5	.99719	.028	37.5	.99930	.007	42.5	.99985	.001	47.5	.99997	.000
27.6	.99065	.095	32.6	.99726	.028	37.6	.99932	.007	42.6	.99986	.001	47.6	.99997	.000
27.7	.99086	.093	32.7	.99733	.027	37.7	.99934	.007	42.7	.99986	.001	47.7	.99998	.000
27.8	.99107	.091	32.8	.99740	.026	37.8	.99936	.006	42.8	.99987	.001	47.8	.99998	.000
27.9	.99127	.089	32.9	.99747	.026	37.9	.99938	.006	42.9	.99987	.001	47.9	.99998	.000
28.0	.99147	.087	33.0	.99753	.025	38.0	.99940	.006	43.0	.99988	.001	48.0	.99998	.000
28.1	.99167	.085	33.1	.99760	.024	38.1	.99941	.006	43.1	.99988	.001	48.1	.99998	.000
28.2	.99186	.083	33.2	.99766	.024	38.2	.99943	.006	43.2	.99988	.001	48.2	.99998	.000
28.3	.99205	.081	33.3	.99772	.023	38.3	.99945	.006	43.3	.99989	.001	48.3	.99998	.000
28.4	.99223	.079	33.4	.99778	.022	38.4	.99946	.005	43.4	.99989	.001	48.4	.99998	.000
28.5	.99241	.077	33.5	.99784	.022	38.5	.99948	.005	43.5	.99989	.001	48.5	.99998	.000
28.6	.99259	.075	33.6	.99790	.021	38.6	.99950	.005	43.6	.99990	.001	48.6	.99998	.000
28.7	.99276	.073	33.7	.99795	.021	38.7	.99951	.005	43.7	.99990	.001	48.7	.99998	.000
28.8	.99293	.072	33.8	.99801	.020	38.8	.99953	.005	43.8	.99990	.001	48.8	.99998	.000
28.9	.99309	.070	33.9	.99806	.020	38.9	.99954	.005	43.9	.99991	.001	48.9	.99998	.000
29.0	.99326	.068	34.0	.99811	.019	39.0	.99955	.005	44.0	.99991	.001	49.0	.99999	.000
29.1	.99341	.067	34.1	.99816	.019	39.1	.99957	.004	44.1	.99991	.001	49.1	.99999	.000
29.2	.99357	.065	34.2	.99821	.018	39.2	.99958	.004	44.2	.99992	.001	49.2	.99999	.000
29.3	.99372	.064	34.3	.99826	.018	39.3	.99959	.004	44.3	.99992	.001	49.3	.99999	.000
29.4	.99387	.062	34.4	.99831	.017	39.4	.99961	.004	44.4	.99992	.001	49.4	.99999	.000
29.5	.99402	.061	34.5	.99835	.017	39.5	.99962	.004	44.5	.99992	.001	49.5	.99999	.000
29.6	.99416	.059	34.6	.99840	.016	39.6	.99963	.004	44.6	.99993	.001	49.6	.99999	.000
29.7	.99430	.058	34.7	.99844	.016	39.7	.99964	.004	44.7	.99993	.001	49.7	.99999	.000
29.8	.99444	.056	34.8	.99849	.015	39.8	.99965	.004	44.8	.99993	.001	49.8	.99999	.000
29.9	.99457	.055	34.9	.99853	.015	39.9	.99966	.003	44.9	.99993	.001	49.9	.99999	.000

TABLE III

Approximation		1		2		3	
<i>i</i>	E_{si} (dB)	$E_u = 78$ dB		$E_u = 76.6$ dB		$E_u = 76.44$ dB	
		z_i (dB)	$L(x_i)$	z_i (dB)	$L(x_i)$	z_i (dB)	$L(x_i)$
1	64	14	0.8835	12.6	0.8585	12.44	0.8554
2	72	6	0.6954	4.6	0.6524	4.44	0.6474
3	60	18	0.9374	16.6	0.9214	16.44	0.9193
4	50	28	0.9915	26.6	0.9883	26.44	0.9878
5	45	33	0.9975	31.6	0.9964	31.44	0.9963
p_c		0.5696		0.5082		0.5010	
ΔE_u (dB)		≈ -1.4		≈ -0.16		≈ -0.02	

The result of the iterative computation is $E_u = 76.42$ dB.

The necessity to carry out numerous multiplications using at least four-digit numbers suggests a further simplification of the method consisting in substituting the $L(x_i)$ by the logarithms of their reciprocal value. This would reduce the computation work to a summation of the $-\log L(x_i)$ values. To further facilitate the computation of ΔE_u , it is appropriate to select a basis for these logarithms in such a way that ΔE_u immediately results from a comparison of the sum with $-\log p_{cp}$ (logarithm to the same basis), e.g. $-\log 0.5$ (50%).

For convenience, the logarithms of $-L(x_i)$ are included in Table II. As an example these logarithms are used in Table IV. The underlying interference problem is identical in Tables III and IV and so are the results.

TABLE IV

Approximation		1		2		3	
<i>i</i>	E_{si} (dB)	$E_u = 78$ dB		$E_u = 76.7$ dB		$E_u = 76.45$ dB	
		z_i (dB)	$-\log L(x_i)$	z_i (dB)	$-\log L(x_i)$	z_i (dB)	$-\log L(x_i)$
1	64	14	1.251	12.7	1.519	12.45	1.575
2	72	6	3.669	4.7	4.264	4.45	4.386
3	60	18	0.653	16.7	0.814	16.45	0.848
4	50	28	0.087	26.7	0.116	26.45	0.123
5	45	33	0.025	31.7	0.035	31.45	0.037
—	$-\log p_c$	5.685		6.748		6.969	
	$-\log 0.5$ ⁽¹⁾	-7.000		-7.000		-7.000	
ΔE_u (dB)		≈ -1.3		≈ -0.25		≈ -0.03	

⁽¹⁾ For $p_{cp} = 0.5$;

for other values of p_{cp} : $-\log p_{cp} = (-7 \log p_{cp})/\log 2$;

e.g. for $p_{cp} = 0.45$: $-\log p_{cp} = 8.064$.

The result of the iterative computation is $E_u = 76.42$ dB.

In addition to the procedure described above a number of other approaches to making use of the simplified multiplication method exist and are contained in a more complete description of the method [EBU, 1984]. Which of the procedures will be preferred may depend on the computation facilities available to the user.

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ANNEX II

USE OF THE LOG-NORMAL METHOD FOR USABLE FIELD STRENGTH CALCULATION
(IN FM SOUND AND TELEVISION BROADCASTING)

1. Order of manual calculation of E_{u1} by the log-normal method. The symbols used are defined in section 3.2.

The calculation is performed in the following order:

- a) find a median value of the resultant of n interferers at the point under consideration and its standard deviation:

$$E_r = 0.1152 \sigma^2 + 10 \log \left(\sum_{i=1}^n M_{si} \right) - 5 \log U \quad (\text{dB})$$

$$\sigma_r = 6.58 \sqrt{\log U} \quad (\text{dB})$$

The value $\sigma = 8.3$ dB is assumed for the frequency bands I to III. For band IV/V this value is dependent on the terrain attenuation, g , and σ is then calculated according to the formula $\sigma = 9.5 + 0.405 g$. The attenuation correction factor g (dB) can be derived from Δh (see Recommendation 370).

- b) take $E_{u1} = E_r$ (see note)

- c) determine the probability that $E_{u1} \geq E_r$:

$$p_i = L(\Delta E_r) = 0.5 + \frac{\varphi(\Delta E_r)}{2}$$

where: $\Delta E_r = \frac{E_{u1} - E_r}{\sigma_m}; \quad \sigma_m = \sqrt{\sigma^2 + \sigma_r^2}$

The probability $\varphi(\Delta E_r)$ can be determined from section 3 of Annex I, so that $x = \Delta E_r$

d) determine the probability that $E_{u1} \geq E_{min}$:

$$P_2 = L(\Delta E_{min}) = 0.5 + \frac{\varphi(\Delta E_{min})}{2}$$

where:
$$\Delta E_{min} = \frac{E_{u1} - E_{min}}{\sigma}$$

The probability $\varphi(\Delta E_{min})$ can be determined from section 3 of Annex I, so that $x = \Delta E_{min}$.

e) determine the probability of the simultaneous fulfilment of both inequalities:

$$P_c = P_1 \cdot P_2.$$

f) If the value obtained for p_c satisfies the given $p_{cp} = 0.5 \pm 0.01$, then the calculation is completed and $E_u = E_{u1}$. Otherwise, we find the value

$$E_{u2} = E_{u1} + \frac{0.5 - p_c}{0.05}$$

and the calculation is repeated from the second point with the new value of E_{u2} , and so on until the required precision is obtained.

Note - It should be noted that in cases where $E_{si \max} - E_{min} \geq 16.5$ dB, the unknown usable field strength value is equal to the value obtained for E_r , and no further calculation is required.

2. Examples of manual calculation of usable field strength.

An example of the calculation of the usable field strength is given for the same values of E_{si} as in Table III of Annex I. This is shown in Table V with respect to two different values of E_{min} : 50 and 57 dB.

Table V shows that, after the calculation of the resultant interference E_r , equivalent to n interfering fields E_{si} , the calculation of the usable field strength E_u will require a minimum number of steps depending on the correlation of the value obtained for E_r with the value of E_{min} .

TABLE V

1. Calculation of median value of resultant interference and its standard deviation			
i	E_{si} , dB	$E_r = 0.1152 \sigma^2 + 10 \lg \sum_i M_{si} - 5 \lg U$ (dB)	$\sigma_r = 6.58 (\lg U)^{1/2}$ (dB)
1	64	73.71	7.85
2	72		
3	60		
4	50		
5	45		

2. Calculation of the value of E_u when $E_{min} = 50$ dB				
1) Approximation $E_{u1} = E_r = 73.71$ dB				
$\Delta E_r = \frac{E_{u1} - E_r}{\sigma_m}$	$L(\Delta E_r)$	$\Delta E_{min} = \frac{E_{u1} - E_{min}}{\sigma}$	$L(\Delta E_{min})$	$p = L(\Delta E_r) L(\Delta E_{min})$
0	0.5	2.86	0.9978	0.4989
$E_u = E_{u1} + \frac{0.5-p_c}{0.05} = 73.71 + 0.02 = 73.73$ dB				

3. Calculation of the value of E_u when $E_{min} = 57$ dB				
1) Approximation $E_{u1} = E_r = 73.71$ dB				
0	0.5	2.01	0.9772	0.488
$E_{u2} = E_{u1} + \frac{0.5-p_c}{0.05} = 73.71 + 0.23 = 73.94$ dB				
2) Approximation $E_{u2} = 73.94$ dB				
0.02	0.508	2.04	0.9798	0.498
$E_u = E_{u2} + \frac{0.5-p_c}{0.05} = 73.94 + 0.04 = 73.98$ dB				

Table VI shows an example of calculation of usable field-strength by the simplified log-normal method (the values taken for E_{si} , E_{min} and σ are the same as those in Table V).

TABLE VI
Calculation of median resultant field-strength

i	E_{si} , dB	$U_s = \frac{(k-1)(\sum_i M_i^2 + M_{min}^2)}{(\sum_i M_i + M_{min})^2} + 1$	$E_{rs} = 0,1152 \sigma^2 + 10 \lg(\sum_i M_i + M_{min}) - 5 \lg U_s$
I	64	I. $E_{min} = 50$ dB	
2	72	$U_s = 26.2378$	$E_{rs} = 73.73$ dB
3	60	2. $E_{min} = 57$ dB	
4	50	$U_s = 25.2356$	$E_{rs} = 73.95$ dB
5	45		

A comparison between Tables V and VI shows that the simplified log-normal method using a non-iterative procedure produces the same usable field-strength values as those obtained by the standard log-normal method.

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REPORT 946-1*

FREQUENCY-PLANNING CONSTRAINTS ON FM SOUND BROADCASTING IN BAND 8 (VHF)

(Question 46/10, Study Programme 46L/10)

(1982-1990)

1. Introduction

In order to ensure effective planning of sound-broadcasting services in the frequency range 87.5-108 MHz, it may be necessary to take into account certain constraints on the use of frequencies in order to avoid interference to other sound-broadcast transmissions within this band and to ensure compatibility with television broadcasts in the frequency ranges 47-68 MHz and 174-230 MHz.

This Report identifies the various possible interference mechanisms resulting from the technical limitations of receiver design, and also those due to constraints resulting from the transmission of several broadcast programmes from the same site.

A listing of the main interference mechanisms generated in the radio receiver is provided in § 2, and § 3 specifies those which are of importance in planning. No account is taken of interference resulting from radiation of harmonics and intermodulation products at transmitter sites, on the assumption that the broadcaster can take the necessary precautions to reduce such spurious radiation to acceptable levels.

2. Interference mechanisms arising from the characteristics of broadcasting receivers

2.1 Local oscillator radiation

In the following, consideration is given only to the case where the local oscillator frequency is 10.7 MHz above that of the wanted transmission, this representing the condition for the great majority of current receivers. The following interference mechanisms may occur:

2.1.1 Radiation at the fundamental frequency

Radiation from VHF-FM receivers may affect VHF-FM reception at frequency separations of 10.7 ± 0.2 MHz. This occurs if adjacent receivers are in use, tuned to transmissions separated by this frequency difference.

Radiation from television receivers in the range 47-68 MHz may affect VHF-FM reception. This may occur when the television oscillator frequency, which is equal to the intermediate frequency (IF) plus the frequency of the vision carrier, lies near the carrier frequency of a VHF-FM transmission. VHF-FM frequencies which may suffer possible interference are given in Table I below.

TABLE I

System	Channel	Channel limits (MHz)	Vision carrier (MHz)	IF video channel (MHz)	VHF-FM frequencies which may be interfered with (MHz)
B	3	54-61	55.25	38.9	94.1; 94.2
B	4	61-68	62.25	38.9	101.1; 101.2
D	R1	48.5-56.5	49.75	38	87.7; 87.8
D	R2	58-66	59.25	38	97.2; 97.3
I	IB	52.5-60.5	53.75	39.5	93.2; 93.3
I	IC	60.5-68.5	61.75	39.5	101.2; 101.3

* Sections 2.5 and 3.2.5 of this Report should be brought to the attention of Study Group 8.

2.1.2 Harmonic radiation

Second harmonics of local oscillators of VHF-FM receivers may interfere with reception of television on all channels in the range 196-230 MHz. The interfering VHF-FM frequency can be calculated using the formula

$$2(f_{FM} + 10.7) = f_{TV} \quad \text{MHz} \quad (1)$$

2.2 Harmonic conversion

In a VHF-FM receiver the second harmonic of the local oscillator may combine with television sound and vision carriers in the range 174-230 MHz to produce a frequency difference of 10.7 MHz, thereby causing interference to radio reception. The interfered VHF-FM frequency can be calculated using the formula

$$2(f_{FM} + 10.7) = f_{TV} \pm 10.7 \quad \text{MHz} \quad (2)$$

2.3 Harmonics and intermodulation products generated under overload conditions in the receivers

High input signal levels can lead to non-linearities in the input stage of receivers, giving rise to the generation of harmonics and intermodulation products. The following cases may occur:

- second harmonics of television transmissions in the range 47-54 MHz fall in the range 94-108 MHz;
- second harmonics of VHF-FM transmissions in the range 87.5-108 MHz fall in the range 175-216 MHz;
- third-order intermodulation products due to reception of more than two VHF-FM transmissions separated by equal frequency spacings occur on the frequencies of these transmissions. For example, with three frequencies f_1, f_2, f_3 such that:

$$f_2 = f_1 + \Delta f; f_3 = f_1 + 2\Delta f \quad (3)$$

where Δf is the frequency spacing,

$$\text{then } 2f_2 - f_1 = 2(f_1 + \Delta f) - f_1 = f_1 + 2\Delta f = f_3 \quad (4)$$

There exist additional possibilities of double or triple beats.

- The second harmonic of an interfering VHF-FH transmission separated by 5.3 or 5.4 MHz from the wanted transmission can combine with the second harmonic of the local oscillator to produce a frequency difference of 10.6 or 10.8 MHz, thereby causing interference to radio reception.
- Two transmissions with a frequency separation of 10.7 ± 0.2 MHz may give rise to interference as a result of intermodulation processes in a receiver.

2.4 Harmonics of the intermediate frequency

Non-linearities in the output intermediate frequency stage can produce harmonics of 10.7 MHz. The ninth and tenth harmonics (96.3 MHz and 107.0 MHz respectively) fall within the VHF-FM broadcasting band 8, and interference may result from feedback to the input stages of receivers tuned to these frequencies.

2.5 Image channel interference

The use of a receiver intermediate frequency of 10.7 MHz effectively eliminates the possibility of image channel interference arising from other emissions within the FM sound broadcasting band (87.5 - 108 MHz). However, emissions in neighbouring frequency bands, in particular the aeronautical communications band (118 - 137 MHz), may cause such interference. At critical locations, the field strength at ground level arising from aeronautical emissions may exceed 100 dB($\mu\text{V/m}$)[CCIR, 1986-90a].

Since the image channel is some 21.4 MHz above the frequency of the wanted signal, broadcasting reception above about 96.6 MHz is potentially at risk if the approach to an airfield lies over a populated area that is close to the limit of service of a broadcasting station. Whether interference is experienced in practice depends on there being the critical frequency relationship between the broadcasting and aeronautical frequencies and on the operational use of the aeronautical frequency. As the broadcasting and aeronautical frequency plans are not normally coordinated to avoid this critical relationship, it appears desirable to provide a minimum value of some 40 dB of image channel rejection in the design of broadcasting receivers. This would be the most appropriate long-term solution.

3. **Relevance of the various interference mechanisms to frequency planning**

3.1 *Main constraints*

3.1.1 The use of VHF-FM transmission frequencies separated by 10.7 ± 0.2 MHz in the same area should be avoided to the extent possible (see §§ 2.1.1 and 2.3).

3.1.2 It is common practice to radiate several VHF-FM transmissions from a single site using either a common antenna or separate antennas. In the former case, the minimum possible frequency separation is determined by the design limitations of combiner units.

Today there are combining units which can operate with frequency separations as low as 800 kHz[CCIR, 1986-90b], for unit transmitter powers of 10 KW, while maintaining signal quality and minimizing intermodulation products.

However, for planning purposes in general, the minimum frequency spacings should not be less than 2 MHz, with a reduction down to 1.5 MHz accepted in special cases.

With separate antennas for each transmission, frequency spacings of less than 800 kHz may be feasible in some cases.

3.2 *Constraints of less importance*

3.2.1 It is suggested that those interference mechanisms which depend on the generation of the second harmonic of the unwanted transmission in the VHF-FM receiver need not in general be taken into account in planning. In general, such interference is only likely to occur in areas close to the unwanted transmitter, and the problem may be overcome either with improved receiver selectivity or by providing appropriate filters in the antenna lead. In the case of portable VHF-FM receivers it will often be possible to reduce the probability of overload by retracting the receiving antenna should wanted signal levels allow this.

3.2.2 Although some earlier experience has indicated that significant interference to stereo reception of transmission on 96.3 MHz may result from the mechanism described in § 2.4, a number of European countries now operate such services on this frequency with little complaint. This may also apply to the frequency 107.0 MHz. In the People's Republic of China, these frequencies are used by FM broadcasting service and no interference has been noticed [CCIR, 1986-90c].

3.2.3 The use of VHF-FM transmission frequencies for which the second harmonic of the local oscillator falls within the band occupied by a television transmission operating in the range 196-230 MHz within the same area should be avoided. In view of the severe limitations it would impose on the use of frequencies for the VHF-FM service, it is probably impractical to take this into account in planning (see § 2.1.2). Tests carried out in the People's Republic of China show that this kind of interference is fairly weak and need not be taken into account in frequency planning [CCIR, 1986-90c].

3.2.4 Although the use of equal frequency spacings can in principle cause problems as described in § 2.3 (third inset), one European country provides three stereo transmissions at equal spacings from nearly all sites providing national network services at VHF-FM. It could be desirable to avoid using equal spacings for transmitters listed close to areas of high population density. Tests carried out in a large, industrialized city with a high population density (Shanghai), using a quintuplexer and an equal carrier-frequency spacing of 1 MHz show that the transmission of five monophonic and stereophonic broadcasting programmes with this mode of operation is possible when the radiation power of the five transmitters is not greatly different [CCIR, 1986-90c].

3.2.5 In situations where image channel interference from aeronautical communications is found to be particularly troublesome, consideration may be given to:

- a) a change in the frequency of either the broadcasting or the aeronautical assignment which together bear the critical relationship;
- b) varying, if at all possible, the operational use of the aeronautical frequency so that emissions at critical locations are largely avoided.

3.3 *Conclusions for the planning procedure*

In the planning procedure the main constraints (§ 3.1) should be taken into account generally. Other constraints may be taken into account only in individual cases.

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REPORT 300-7

STEREOPHONIC OR MULTI-DIMENSIONAL SOUND IN FREQUENCY-MODULATION SOUND BROADCASTING

(Question 46/10, Study Programme 46D/10)

(1963-1966-1970-1974-1978-1982-1986-1990)

1. Stereophonic sound in frequency-modulation sound broadcasting

Since the XIth Plenary Assembly of the CCIR, Oslo, 1966, certain administrations and broadcasting organizations have conducted theoretical and experimental work relating to stereophonic broadcasting.

As a result of this work, stereophonic transmissions intended for the public, using single frequency-modulation transmitters, were introduced in a large number of countries.

Specifications for stereophonic broadcasting systems already in use, i.e. the pilot-tone system and the polar-modulation system, are given in Recommendations 450 and 412.

A variant of the pilot-tone system has been investigated in the Netherlands using a single-sideband modulation of the sub-carrier, having a vector amplitude twice that of the sidebands generated in the case of normal sub-carrier modulation, making it compatible for stereophonic receivers having a synchronous type of decoder. The variant could offer interesting applications in certain circumstances, while maintaining the normal receiving and decoding principles.

2. Multi-dimensional sound-broadcasting systems

2.1 *Desirable basic characteristics*

The principal desirable characteristics of any system providing 3 or more output sound signals from a single radio-frequency channel, are as follows:

- the system should be compatible: that is to say, it should be possible to obtain both stereophonic and monophonic reception of a multi-dimensional sound transmission without reduction of quality in comparison with the reception of normal stereophonic or monophonic transmissions;
- the system should provide high-quality multi-dimensional sound reproduction;
- it should be possible to construct multi-dimensional sound receivers at reasonably economic prices;
- the introduction of multi-dimensional sound transmissions at an existing stereophonic broadcasting station, should not significantly reduce the coverage area of the station for stereophonic or monophonic reception;
- the coverage area of the broadcasting station for multi-dimensional sound reception should be substantially equal to that required for stereophonic reception;
- the protection against interference required for multi-dimensional sound reception should not be significantly greater than that required for stereophonic reception;
- the introduction of multi-dimensional sound transmissions should not necessitate extensive changes to existing frequency assignment plans;
- any system adopted should, as far as possible, acknowledge the existence of non-ideal receivers in the market and be so designed as to minimize the effect of any resulting cross-modulation on their monophonic and stereophonic performance.

2.2 Results of tests on multi-dimensional sound systems

[CCIR, 1970-74a] describes a variant of the pilot-tone system in which quadrature modulation is used for an additional sound channel *C* combined with the two stereophonic channels *A* and *B*. By so doing, the system is capable of transmitting three separate sound signals which can be demodulated in the same way as the pilot-tone system. With a baseband width the same as that required for stereophony and by a correct choice of signal levels, the radio-frequency bandwidth remains unchanged and hence, also, the radio-frequency protection ratios. This system is in accordance with Recommendation 450.

Subjective tests have shown that with three separate channels and matrixing signals *A*, *B* and *C* corresponding with the positions of the studio microphones feeding four loudspeakers, no significant subjective differences occur when compared with reproduction in four separate channels.

A third variant of the pilot-tone system is described in [CCIR, 1970-74b], in which four separate channels are transmitted by one transmitter. The system uses quadrature modulation of the normal suppressed sub-carrier of 38 kHz but additionally modulates a fourth audio-frequency signal onto a second suppressed sub-carrier of 76 kHz. Two possibilities of dividing the four signals into the different sub-carrier channels are given and with a sub-carrier frequency of 76 kHz, either double-sideband or single (lower) sideband modulation may be used.

[CCIR, 1974-78] outlines the work that has been carried out involving the study of some of the fundamental properties of hearing involved in the subjective aspects of multi-dimensional sound systems. It includes a proposal for a quadraphonic broadcasting system known as Matrix H, which encodes the incoming audio signals into two audio channels for distribution and transmission. At the receiver these signals can be reproduced directly as stereophony (or monophony) and can be decoded to give quadraphony. A considerable number of tests and calculations carried out in the United Kingdom have shown that Matrix H meets the basic characteristics as listed in § 2.1.

[CCIR, 1978-82a] describes surround-sound systems used experimentally in the United Kingdom for broadcasting. It is acknowledged that while all systems require an artistic compromise to be made between surround-sound and stereophonic balances, a broad distinction can be made between hierarchical and non-hierarchical systems. In hierarchical systems surround-sound decoding is possible with two or two and a half channels, and it achieves a similar coverage area to the stereophonic service at the expense of reducing the sharpness of sound images because of phase differences occurring in the audio band. In non-hierarchical three channel systems the stereophonic sound image can be excellent, but with the penalty of a reduced surround-sound coverage area.

[CCIR, 1978-82b] describes the stereophonic and monophonic compatibility of various multi-dimensional sound systems which utilize an N-2-N matrix format. This describes five such systems: SQ (CBS), HJ (BBC/NRDC), H (an earlier BBC system), BMX (Cooper/Nippon Columbia) and QS (Sansui). As described in this document, these characteristics differ significantly under conditions of ordinary stereophonic and monophonic reception.

3. Compandor applications

3.1 The FM-FM system using a compressor/expander in the S-channel

In the Documents mentioned in the Note to this Report, results are given of tests on a number of systems using a compressor for the *S*-channel in the transmitter and a corresponding expander in the receiver. These tests were carried out with an FM-AM system, a pilot-tone system and an FM-FM system. The results obtained have shown that noise in the *S*-channel is considerably suppressed by the compressor/expander. Recent tests have shown furthermore that, when transmitting two independent monophonic programmes, only the FM-FM system can give sufficiently low cross-talk from the *S*-channel, into the *M*-channel. The FM-FM system is defined by the following specifications:

- a compatible signal *M* produces a deviation of the main carrier of not more than 80% of the maximum frequency deviation for monophonic transmissions; in the case of two-programme transmission, the deviation is equal to that of the first programme signal. In the case of stereophonic transmissions, it is equal to one half the sum of the left-hand signal *A* and the right-hand signal *B*;
- a signal *S* produces frequency-modulation of a sub-carrier; in the case of two-programme transmissions equal to the second programme signal and, in the case of stereophonic transmissions, equal to one half the difference between the left-hand signal *A* and the right-hand signal *B*;
- the frequency of the sub-carrier is 33.3 kHz $\pm$ 100 Hz;
- the maximum frequency deviation of the sub-carrier is $\pm$ 10 kHz;
- the sub-carrier produces a deviation of the main carrier of between 18% and 20% of the maximum frequency deviation for monophonic transmissions;

- the pre-emphasis of the *S*-signal is identical to that of the compatible *M*-signal;
- a compressor with transfer ratio 2/1 (in dB) is inserted in the *S*-channel of the transmitter before the pre-emphasis network. This compressor has time-constants respectively equal to 2 ms for the rise-time and 20 ms for the decay-time;
- an expander with characteristics reciprocal to those of the compressor, is inserted in the *S*-channel of the receiver after the de-emphasis network;
- in stereophonic transmissions, an *A*-signal produces a frequency deviation in the same direction for the sub-carrier and the main carrier.

3.2 *Compatible companding in the pilot-tone system*

Broadcast tests [CCIR, 1986-90] in the United States of America have demonstrated the feasibility of providing improved stereo reproduction in fixed and mobile reception. A new technique uses an additional stereophonic sub-carrier in quadrature at 38 kHz for transmission of a compressed audio difference (L-R) signal. In new receivers, an adaptive expander utilizing the uncompressed difference signal as a decoding reference, provides accurate decoding over a range of compression characteristics. With this system, the size of the original stereophonic coverage area has been increased while maintaining compatibility with properly aligned receivers. The system has the following specifications:

- all existing characteristics of the pilot-tone composite signal are maintained;
- the *S'* signal, which is a compressed version of the regular *S* signal, amplitude modulates a second stereophonic sub-carrier;
- the frequency of the second stereophonic sub-carrier is also 38 kHz. It is transmitted in phase quadrature with respect to the first (regular) stereophonic sub-carrier;
- the polarity of the *S'* signal is reversed with respect to the *S* signal;
- complimentary transmitter and receiver low-frequency equalization with time constants of 200 and 1000 μ s is employed in the *S'* signal;
- an identification signal at approximately 10 Hz, for automatic actuation of receivers, is included in the quadrature channel, at a level which produces 1% deviation of the main carrier. The identification signal is frequency-locked to the pilot sub-carrier divided by 1920 (approximately 10 Hz);
- the compression function for the *S* signal is characterized at low input levels by a 14 dB gain. For input levels greater than -22dB with respect to the level providing maximum modulation of the sub-carrier, the output is gradually attenuated in order to reduce possible over-modulation of the main carrier. With the exception of the 14 dB fixed gain, this compression characteristic is only approximate, and, for the reason stated below, may be modified somewhat by a broadcaster without concern about rendering obsolete receivers already using the new service;
- in receivers, a unique expander processing the sum of the *S* and *S'* signals and using the *S* signal alone as a decoding reference produces a new signal equivalent to the *S* signal, but with an improved signal-to-noise ratio at low programme levels.

Note. - A complete list of CCIR documents on stereophonic broadcasting from the periods between 1963 and 1970 is to be found in Volume X of the XIIIth Plenary Assembly, Geneva 1974, pages 124 and 125.

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REPORT 463-5

TRANSMISSION OF SEVERAL SOUND PROGRAMMES OR OTHER SIGNALS
WITH A SINGLE TRANSMITTER IN FREQUENCY-MODULATION SOUND BROADCASTING

(Question 46/10, Study Programme 46G/10 and 46H/10)

(1970-1974-1978-1982-1986-1990)

1. Introduction

The growing need for more programmes in almost all parts of the world makes it desirable to examine the present broadcasting standards to determine what changes are possible to provide a better utilization of the available frequency bands. The frequency-modulation service uses a large bandwidth to provide a programme service of exceptional quality. However, it has been found that some additional broadcasting services can be added without significant deterioration of the programme quality. This is an important consideration when the need for additional broadcasting services might make a reconsideration of the standards a necessity.

The transmission of data information as an alternative to the main sound programme, is described in Reports 802, 1207 and 1208.

2. Requirements for monitoring and test instruments

The basic purpose of the supplementary sub-carrier monitoring instruments is:

- to provide equipment for use in the routine operation and adjustment of the broadcasting station in its day-to-day operations;
- to provide the instrumentation necessary to test and prove the integrity of the transmitting system.

In keeping with the above requirements, the following parameters become subjects of interest:

- the degree of modulation of the main carrier by the sub-carrier;
- centre frequency of the sub-carrier;
- spectrum of the modulated sub-carrier;
- cross-talk and noise;
- distortion, frequency response, or error rate, as applicable.

3. Systems using one or more sub-carriers for supplementary sound programmes

The analysis of supplementary sub-carrier operation in [CCIR, 1966-69a], shows that the use of a frequency-modulated sub-carrier provides the best signal-to-noise ratio for a given sub-carrier amplitude. Although the theoretical analysis of cross-talk is omitted, qualitative analysis, borne out by experience with sub-carrier operation, indicates that the system of modulation with the best signal-to-noise ratio also has the least cross-talk when other system parameters are the same.

Although the choice of the lowest possible sub-carrier frequency will provide the best signal-to-noise ratio in the supplementary sub-carrier channel, the sub-carrier frequency should be high enough to prevent audible beats. The use of a still higher sub-carrier frequency reduces the probability of cross-talk but unnecessarily reduces the signal-to-noise ratio.

3.1 An FM-FM system using a compressor/expander in the S channel is described in Report 300.

Note - The system did not conform to Recommendation 450.

3.2 An FM-FM system is described in [CCIR, 1966-69b]. In this instance, the frequency of the sub-carrier was 67 kHz with a maximum deviation of the main carrier by the sub-carrier of ± 7.5 kHz during stereophonic broadcasting and a maximum deviation of the main carrier by the sub-carrier of ± 11.25 kHz during monophonic broadcasting. The maximum audio frequency of the supplementary channel is restricted to 6 kHz.

3.3 [CCIR, 1966-69c] gives the results of tests showing that it is possible to transmit simultaneously a main channel programme and a second programme using a supplementary sub-carrier. The frequency response is as good and the distortion as low, for the supplementary sub-channel as for the main channel. A sub-carrier frequency of 58 kHz was used. The deviation of the main carrier by the sub-carrier was ± 15 kHz and the deviation of the sub-carrier ± 12 kHz. The cross-talk figures were -70 dB or better, for the main channel and about -50 dB for the sub-channel.

3.4 A system for the transmission of four supplementary sound programmes in frequency-modulation broadcasting is described in [CCIR, 1966-69d]. In this instance four sub-carriers with frequencies of 70.75 kHz, 55.25 kHz, 39.75 kHz and 24.25 kHz are used. The maximum deviation of the main carrier by all four sub-carriers simultaneously, is 22.5 kHz. The maximum audio frequency of each sub-carrier is 3 kHz.

3.5 A similar FM-FM system providing two supplementary programmes is described in [CCIR, 1966-69e]. The first sub-carrier has a frequency of 33.3 kHz and carries the second programme as described in Report 300. A third programme is carried on a second sub-carrier having a frequency of 66.6 kHz. This second sub-carrier, which has a deviation of ± 8 kHz, determines the deviation of the main carrier which is ± 3 kHz. The audio-frequency range is 40 to 5000 Hz. The system includes a compressor/expander with the same characteristics as mentioned in § 3.1. The cross-talk figures are better than -70 dB for the first and second channels and better than -60 dB for the third channel (when modulated at 100% in the third channel, and at 30% in the others, the cross-talk figure was better than -45 dB).

Note - This system does not conform to Recommendation 450.

3.6 [CCIR, 1970-74a] describes a two sound programme system, the supplementary channel intended to be received with high-grade receivers for feeding wired distribution systems. The maximum deviation of the main carrier caused by the first channel programme is ± 40 kHz (80% of the maximum system deviation). The sub-carrier frequency is 62.5 kHz and is frequency modulated with a maximum deviation of ± 30 kHz. The sub-carrier level is varied, deviating the main carrier ± 15 kHz during periods of no modulation and is reduced to a minimum value of ± 6 kHz (8 dB) at 100% modulation. The cross-talk figure for the first channel from the second channel is better than 55 dB and better than 50 dB for the second channel from the first channel. The difference in immunity to interference between the two channels is calculated to be 18 dB. This value can be reduced to 5 dB if the audio bandwidth of the second channel is reduced from 15 kHz to 6 kHz and highly directional antennas are used. More details are given in the bibliography.

3.7 [CCIR, 1970-74b] describes a system providing for a maximum of eight supplementary low-quality channels, in each of which suppressed carrier single-sideband modulation is employed. The corresponding eight frequencies are chosen within the band 64 kHz to 100 kHz, so as to avoid interference with the stereophonic signal within the frequency band up to 53 kHz.

3.8 Tests in the UK [Whyte, 1977] have shown that a subsidiary programme using a 41 kHz sub-carrier at any worthwhile level of injection caused unacceptable impairment to reception of a monophonic main programme. When a 67 kHz sub-carrier was used, at 15% or less injection, the subsidiary programme was found to be acceptably compatible with reception of a monophonic main programme. The quality of reproduction of the subsidiary programme was found to be markedly inferior to that of the main programme.

3.9 [CCIR, 1982-86a] shows that a supplementary monophonic broadcasting programme can be effectively transmitted to relay-distribution networks in FM stereophonic broadcasting of pilot-tone systems by employing a 67 kHz (± 4 kHz deviation) supplementary channel with a compandor. By using antennas with moderate directivity, the coverage area of the supplementary programme can approximate that of stereophonic broadcasting with conventional receiving antennas.

3.10 [CCIR, 1982-86b] shows that a subsidiary music programme using a 67 kHz (± 6 kHz deviation) sub-carrier transmitted with a data channel using a 92 kHz (± 4 kHz deviation) sub-carrier, exhibited some impairment in multipath reception areas. Reception of the stereophonic main programme was unaffected except in the area of the most extreme multipath conditions. Generally, in modern receivers fitted with sub-carrier trap filters and/or phase locked loop demodulators, the impairment of the main stereophonic programme was minimal or negligible. Where impairment of the subsidiary programme was evident (due to multipath) the reception could always be improved by relocating the receiver or by using a directional antenna.

Note. — The frequency of the 92 kHz sub-carrier did not conform to Recommendation 450.

3.11 An FM multiplex broadcasting transmission system that has a 76-kHz subcarrier, ± 1.875 -kHz deviation and 48-kbps transmission capacity has been developed and a digital sound broadcasting system using 32-kbps ADPCM has been put into service for fixed receivers [CCIR, 1986-1990a].

4. Systems using one or more sub-carriers for supplementary information

A number of applications are foreseen for the transmission of various types of information in addition to the main programme and which may, or may not, be related to that programme. Such applications include the control of receivers, operation of display devices built into the receiver or operation of a separate display device such as a television receiver. Additionally, the broadcaster may make use of the system for monitoring purposes.

Recommendation 643 (Annex I) provides details of a Radio Data System (RDS) for applications of this type.

Earlier work has shown that the transmission of sound is best accomplished through the use of frequency-modulation of a sub-carrier. Whilst this method may also be used for other types of information, more recent work [CCIR, 1974-78a] has shown that if the information can be digitally coded, other forms of modulation may be preferred. Further studies should be undertaken in order to examine any possibility for the use of a common transmission standard for different applications.

4.1 [CCIR, 1970-74c] describes the ARI traffic broadcast identification system, which is used in several countries in broadcasting networks. Results of field trials are given. The traffic information is broadcast in the normal way. Those listeners wishing to operate their receiver automatically, have three different control signals available. A sub-carrier at 57 kHz (see § 5.3) is used as an identification signal for networks transmitting traffic information. This sub-carrier is amplitude modulated. One of six possible modulation frequencies in the range 23 Hz to 54 Hz indicates the traffic area. A modulation frequency of 125 Hz is used for receiver muting and is present only during traffic announcements. The frequency deviation of the main carrier due to the unmodulated 57 kHz sub-carrier should not exceed ± 3.8 kHz [CCIR, 1982-86c]. Operational practice has shown that even under difficult receiving conditions a high degree of reliability was obtained, of both the control functions and the sound reception. No mutual interference was observed between the identification signals and either monophonic or stereophonic channels.

4.2 [CCIR, 1970-74d] describes the use of a supplementary sub-carrier for the simultaneous transmission of voice and a graphic display. The graphic display portion of the system is designed for the transmission of handwritten messages, diagrams, maps, and sketches at the time when written. The co-ordinates are frequency determined in ranges from 1310 to 1490 Hz and 2060 to 2340 Hz for vertical and horizontal travel respectively. Actual writing causes one of the co-ordinate signals to become modulated in amplitude with double the power line frequency. Frequency modulation on a carrier of 5100 Hz makes it possible to achieve frequency division multiplex of an audio-frequency channel from 200 to 2000 Hz from the writing channel. The sum signal is used for modulating a frequency-modulation sub-carrier as mentioned in Recommendation 450. The voice transmission can be related to the graphic display or not, as the user wishes. The system is especially useful for educational purposes. The entire information can be stored on audio quality magnetic tape.

4.3 [CCIR, 1970-74e] describes the transmission of television picture signals using a frequency-modulation sub-carrier with sampling scan conversion techniques to reduce the video bandwidth to the audio-frequency range. The transmission system consists of a source of composite video from a Vidicon camera, a scan converter, a sub-carrier generator, and a frequency-modulated transmitter. The receiver consists of a sub-carrier demodulator, a scan converter and a video monitor. The bandwidth of the output signal will correspond to one half of the line rate of the wide-band input signal. A line rate of 15 750 Hz will give a bandwidth of approximately 8 kHz. The picture period for a bandwidth of 8 kHz is 7 s. If the picture period is increased to 14 s, a bandwidth of only 4 kHz will be required.

4.4 Since an ideographic character has more information than an alphabetic character, an error-correction technique is essential for its reliable transmission. In [CCIR, 1982-86d], a comparison of error-correcting codes for FM data broadcasting under noise limiting conditions is made.

The error-correcting codes compared were:

- (a) the (8 : 4) extended Hamming code;
- (b) the (26 : 16) shortened cyclic code;
- (c) the (24 : 12) extended Golay code; and
- (d) the (272 : 190) code.

Packet structures which included 190 or 192 information bits were assumed.

The results were that in comparison with (a) at a bit error ratio of 10^{-3} , (b) has 10^2 , (c) has 10^4 and (d) has 10^{12} times improvement of error-correction capability. It was also shown that in the case of error detection, error patterns of 1 out of 2^{82} were undetected and extensive error detection capability is obtained with (d) in comparison with (a) or (b) or (c). By adopting a powerful error correcting and detecting code such as the (272 : 190) code it would be possible to decrease the deviation of the main carrier by the data sub-carrier in order to increase the cross-talk margin to the main programme channel.

4.5 [CCIR, 1982-86b] shows that data transmissions at 1200 Bd were possible using a 92 kHz (± 4 kHz deviation) sub-carrier together with a subsidiary programme using a 67 kHz sub-carrier (± 6 kHz deviation). In multipath reception conditions, impairment to the data did occur but this was eliminated with the use of a linearly polarized directional antenna.

Note. — The frequency of the 92 kHz sub-carrier did not conform to Recommendation 450.

4.6 [CCIR, 1986-1990 (b)] describes the feasibility for the supplementary transmission of three sub-channels simultaneously with the main stereo program: the RDS (Radio Data System) or a Paging service, an audio service and a data service. The RDS or Paging service sub-channel, both having their sub-carrier set at 57 kHz, were tested separately during this test program. The injection level was set at 3% of 75 kHz for the RDS system and between 3% and 7% for the Paging service. Amplitude modulation double sideband suppressed carrier was used.

For the audio sub-channel, the carrier frequency was set at 72 kHz for a 10 kHz audio frequency bandwidth (between 61 and 73 kHz). The injection level of the main carrier was set at 5% of 75 kHz and amplitude modulation, single sideband (lower sideband) suppressed carrier was used.

For the data service sub-channel, the carrier frequency was set at 92 kHz for an occupied bandwidth between 80.5 to 90 kHz. The injection level was set at 6% of 75 kHz for a 9.6 kbps data rate. Amplitude modulation, vestigial sideband suppressed carrier was used.

Field and laboratory tests were used for the feasibility study.

5. Cross-talk into the stereophonic channels caused by supplementary sound programmes or other signals

5.1 [CCIR, 1974-78b] gives typical test results of measurements of supplementary sub-carrier cross-talk into the M-channel and the S-channel in the transmission chain. It suggests that the deviation of the 67 kHz sub-carrier should be limited to ± 4 kHz when used with stereophonic transmissions.

5.2 [CCIR, 1970-74f] underlines the difficulty in avoiding cross-talk caused by sub-carrier channels similar to the specifications given in Recommendation 450. Receivers with insufficient filtering between the stereophonic multiplex-signal and the supplementary signals will incur cross-talk in the stereophonic programme. Those administrations not at present using supplementary broadcasts should be alert to this interference possibility before commencing such services.

5.3 [CCIR, 1970-74f] indicates that a supplementary carrier could be tolerated having a frequency of three times the pilot tone frequency and being phase-locked to it. In this case, only the deviation of the main carrier by the sub-carrier may be quite high (± 7.5 kHz). This sub-carrier, for example, can be used as a means of identification for a particular transmitter programme (e.g. a road traffic information programme).

5.4 [CCIR, 1974-78c] and [CCIR, 1974-78a] show results from tests with various sub-carrier frequencies and types of modulation. The Documents indicate that a supplementary sub-carrier having a frequency of three times the pilot-tone frequency and being phase-locked to it, causes minimum interference in existing stereophonic receivers. The Documents also indicate that with that choice of sub-carrier frequency there can be interference in receivers utilizing the phase-lock loop technique for the 38 kHz carrier regeneration. The phase of this carrier is altered by the modulated supplementary sub-carrier unless the spectrum of the modulating signal lies above the loop-bandwidth of the filter in the phase-lock loop. Such phase changes cause a change in amplitude of the demodulated S-signal resulting in an increase in A to B cross-talk and an intermodulation between the demodulated S-signal and the modulating signal of the supplementary sub-carrier.

5.5 Tests in the U.K. [Whyte, 1977] have shown that a subsidiary programme using a 67 kHz sub-carrier at any worthwhile level of injection caused unacceptable impairment to reception of a stereophonic main programme on at least 50% of existing stereo receivers. The use of a highly compressed main programme signal could, to some extent, conceal the interference from the subsidiary programme. The quality of reproduction of the subsidiary programme was found to be markedly inferior to that of the main programme.

5.6 [CCIR, 1982-86a] shows that the cross-talk from the supplementary channel with a compandor into the stereophonic channel can be attenuated by more than 60 dB by a correct choice of the supplementary carrier frequency (67 kHz), its maximum frequency deviation (± 4 kHz) and maximum audio frequency (6 kHz).

5.7 [CCIR, 1982-86b] shows that with a subsidiary programme using a 67 kHz sub-carrier (± 6 kHz deviation) and a data channel using a 92 kHz sub-carrier (± 4 kHz deviation), there was no noticeable cross-talk from either of the two sub-carriers into the main stereophonic channel (± 75 kHz deviation) in multipath reception areas. However, cross-talk from the main stereophonic channel into the sub-channels was noted. The reduction of this cross-talk was always possible with the use of a linearly polarized directional antenna or by relocating the receivers. The deviation of the main carrier by each sub-carrier was ± 7.5 kHz.

5.8 [CCIR, 1986-90a] shows results from tests with a 76 kHz sub-carrier frequency. In both laboratory and field tests, using fixed receivers, a 48 kbps digital signal transmitting system which can transmit a narrow-band sound signal and data signals simultaneously, had very little influence on the level of interference to stereo reception at 2.5% (± 1.875 kHz) multiplexing level. The two sub-carrier system which uses a 76 kHz frequency-modulated sub-carrier for the sound signal and a 57 kHz sub-carrier for the data signal, affected about twice the number of receivers compared with that of the digital system.

5.9 Tests in Australia [CCIR, 1986-90d] have shown that for a 67 kHz sub-carrier injected at a level not exceeding 10% (± 7.5 kHz deviation), beat frequency interference in the main programme will be subjectively insignificant if sub-carrier deviation and percentage injection level conform to the following relationships,

(a) For direct FSK of the sub-carrier:

$$I = [20/(F+2)] + \sqrt[3]{0.83F}$$

where F is the sub-carrier deviation in kHz, and
I is the sub-carrier injection level in percentage of
 ± 75 kHz.

- (b) For analogue modulation of the sub-carrier, a sub-carrier deviation greater than 8 kHz is not permissible as this would produce crosstalk into the 57 kHz portion of the baseband used for other services.
 - (i) for music and speech, and a sub-carrier deviation up to 4 kHz, the maximum permissible modulating frequency is 8 kHz. For a deviation in excess of 4 kHz, the maximum modulating frequency must be progressively reduced until it does not exceed 3 kHz at a sub-carrier deviation of 8 kHz.
 - (ii) for audio frequency shift keying, the permissible sub-carrier deviation is related to the audio modulating frequency. For a modulating frequency up to 2.6 kHz, the sub-carrier deviation should not exceed 6 kHz. High modulating frequencies in the range of 6 to 8 kHz require the sub-carrier deviation to be reduced to between 4 and 2 kHz respectively.

It has been found that interference due to the sidebands associated with the modulated 67 kHz sub-carrier overlapping the S signal was inaudible provided that the sideband products below 53 kHz averaged in a 3 kHz bandwidth, were less than 60 dB below the level of the unmodulated sub-carrier.

5.10 Tests in India [CCIR, 1986-90e] show that a data channel using a sub-carrier at 57 kHz with a deviation of ± 2.4 kHz, and a subsidiary speech quality programme using a 67 kHz sub-carrier with ± 5 kHz deviation intended for auxiliary services, along with a main stereo programme, exhibited no impairment in reception to any of the services. The data channel used Group 5A (CCIR Recommendation 643) for text transmission for educational and public utility services.

5.11 [CCIR, 1986-90b] shows the impact of transmitting, simultaneously with the main stereo programme, three sub-channels: the RDS or Paging service, an audio service and a data service. Interference into the main channel was measured as an increase in background noise (weighted) and evaluated subjectively by using the CCIR 5-Grade impairment scale. Based on results from 13 receivers (field and laboratory results), crosstalk into the main programme by the different combinations of sub-channel arrangements was detectable on four of the receivers and only when the Paging service was transmitted at the 7% injection level.

Interference into the RDS, the Paging and data service sub-channels was evaluated by observing errors in messages received while interference into the audio sub-channel was measured as increase in background noise (weighted). Errors in the messages were easily corrected by bit error correcting circuits. The increase in noise background depended on the tone used as the modulating signal for the main programme. Improvements to the prototype equipment can reduce this increase in noise.

6. Radio frequency problems

When additional signals are multiplexed onto the main monophonic or stereophonic programme, account must be taken of certain radio-frequency effects. These effects include a possible change in the required protection ratio and a possible increase in susceptibility to the deleterious effects of multipath. Administrations proposing systems requiring the use of additional sub-carriers are, therefore, urged to include this information.

Investigations into possible protection ratio changes should explore the case where the wanted transmission is stereophonic and the unwanted transmission includes additional sub-carriers, also the case where both transmissions include additional sub-carriers. Results are needed for co-channel and for adjacent channel protection ratios, noting that in different parts of the world channel spacings of 100 kHz and 200 kHz are used.

Note - A complete list of CCIR documents on supplementary sub-carriers in frequency modulation sound broadcasting from the period between 1966 and 1986 can be found in Volume X, Part I, of the XVth Plenary Assembly, Dubrovnik, 1986, pages 243-248.

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REPORT 1067

IMPROVEMENT OF THE RECEPTION QUALITY IN AUTOMOBILES FOR
FREQUENCY MODULATION SOUND BROADCASTS IN BAND 8 (VHF)

(Question 49/10)

(1986)

1. Introduction

There are major difficulties in achieving high-quality reception of FM sound-broadcasting signals in a moving automobile. Severe signal drop-outs may be experienced by the receiver as the vehicle encounters a continuously changing multipath receiving environment. At high vehicle speeds, these drop-outs appear as short staccato-like bursts of noise. At lower speeds, losses of programme are replaced by noise created by two completely different effects. In cases of non-selective fading caused by short-distance multipath propagation, the receiver attempts to compensate for a diminished signal level. In cases of selective fading, caused by long distance multipath propagation, the distortions appear similar to a strong modulation-dependent noise. The rate of drop-out occurrences is usually random, with the greatest incidence occurring in urban areas where numerous buildings may shadow or reflect the transmitted radio signals. The incidence of the selective fade-out is additionally dependent upon topography.

2. Non-selective fading**2.1 Diversity reception**

Theoretical studies and field measurements were carried out in the United States of America in order to determine the multipath characteristics for band 8 (VHF) sound-broadcast transmissions. As a result of these studies two major conclusions were reached. Firstly, it was determined that the probability of a signal drop-out occurring during the itinerary of a vehicle may be represented approximately by the well-known Rayleigh probability distribution. Secondly, it may be possible to achieve a considerable improvement in quality of reception by mounting two antennas on the vehicle to provide space diversity reception in a selection combining mode.

A classical technique for dealing with the problem of audio reception of fading signals is the use of diversity reception. Diversity is typically achieved by the combined (or alternate) use of two or more receiving elements. Any of several diversity modes may be employed: frequency (more than one channel), polarity (antenna with different polarization), or space. The last mode is especially attractive for automobile reception since it promises great benefit with no change of broadcasting standards.

Typically, space diversity systems have utilized antennas placed many wavelengths apart, usually ten wavelengths or more. Such spacing is not possible on an automobile where the overall length of the vehicle may be typically less than 1.5 wavelengths at band 8 broadcast frequencies. However, for correcting a drop-out, it has been observed that the movement of a vehicle over only a short distance (approximately 1 m) usually results in a completely adequate improvement in reception. This observation provided the rationale for investigating how effectively space diversity reception might be employed in an automobile.

The theoretical study based on actual field measurements and computer processing of the data has confirmed that significant improvement of FM radio reception can be achieved; Fig. 1 illustrates this. The upper chart shows a representative plot of relative field strength along a line in space from 0-14 m. The downward fluctuations are of varying severity and may or may not result in audible drop-outs, depending on station power, distance of the vehicle from the transmitter, and receiver sensitivity. The lower chart illustrates the degree of signal uniformity which would be achieved if a 2.8 m space diversity antenna system were used. (The vehicle shown is drawn to the distance scale of the chart.) It is easy to perceive how a radio receiver, automatically selecting the better output from one of the two antennas, can maintain an optimum signal level as both antennas are moved in tandem along the distance scale.

2.2 Measurements and analysis

A number of measurements were made on various streets and highways in New York City and surrounding suburban areas. A test vehicle with a single antenna connected to a data recorder was used to monitor the changing RF signal envelope as a function of vehicle speed. Computer sampling and processing of the data was later accomplished to predict what degree of improvement might be achieved by the use of two receiving antennas instead of one. Figure 2 represents the average system performance of the diversity reception for an antenna separation of 2.6 m for a constant value of probability, arbitrarily chosen to be 1.0%. Included in this chart is the reference curve for Rayleigh fading which appears as a diagonal line for the case of a single antenna ($m = 1$).

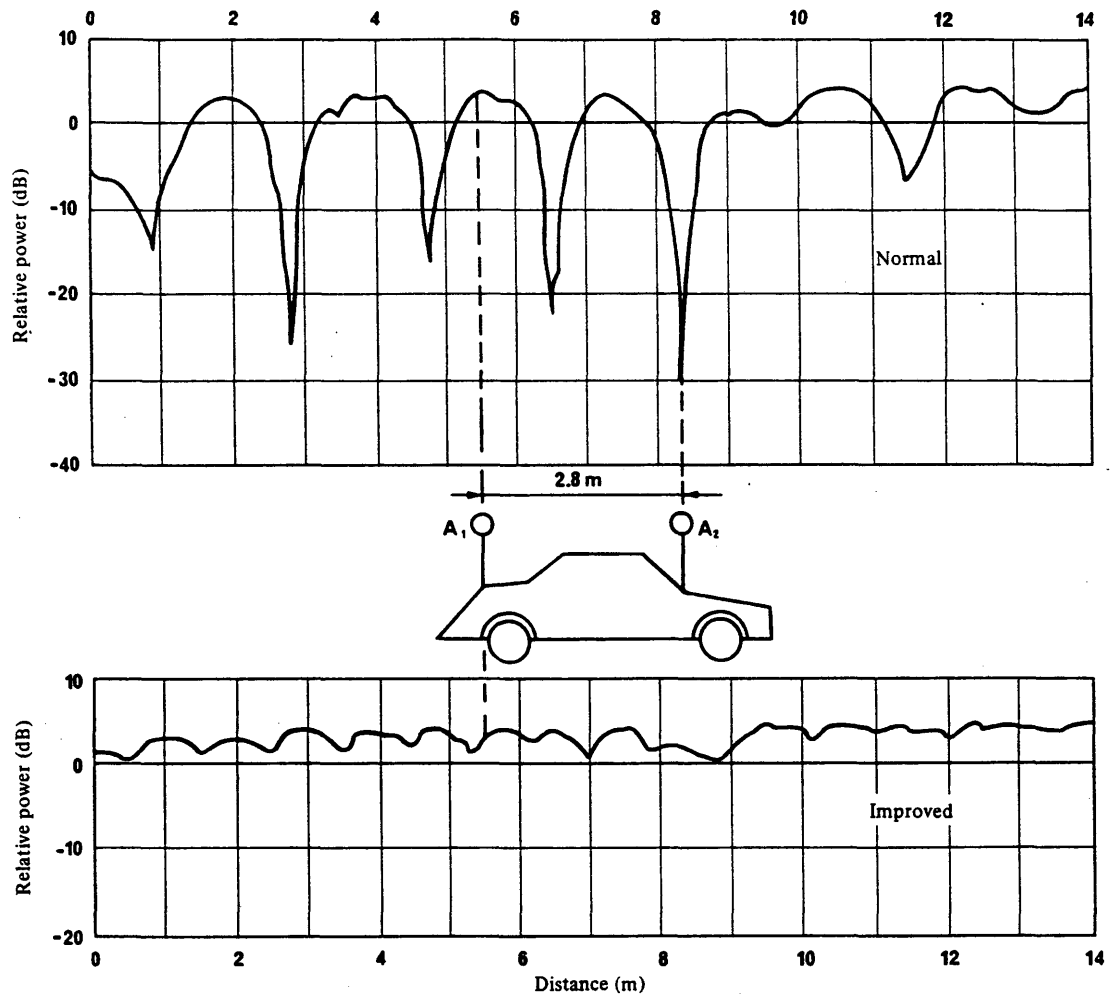


FIGURE 1 – Fading of FM carrier as a function of distance (above) and resultant FM carrier using space diversity and selection combining

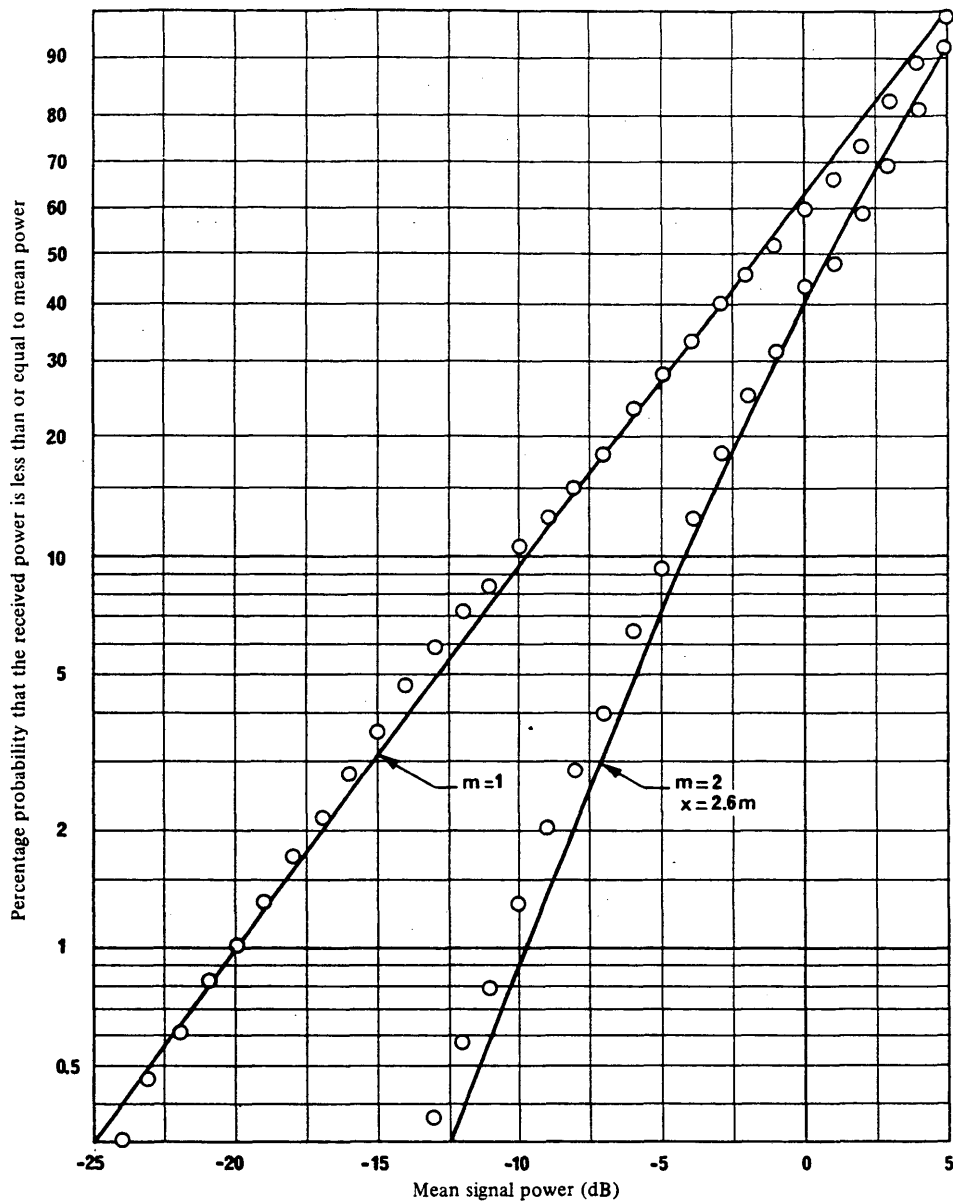


FIGURE 2 – Cumulative probability distribution of m -branch selection diversity combiner

m : number of car receiving antennas involved

x : separation between antennas

Figure 3 illustrates the effect of antenna separation on the performance of a selection combining system utilizing two antennas. The figure suggests that the diversity receiving system would achieve nearly theoretically predicted performance with antenna separation greater than 2.8 m. It also suggests that a single separation distance would be suitable for good reception across the entire FM broadcast band. Since the entire band extends only about $\pm 10\%$ from the 101.1 MHz mid-frequency employed in the tests; it appears that any antenna separation greater than 2.8 m will be acceptable. Fortunately, this dimension can be realized on ordinary passenger automobiles.

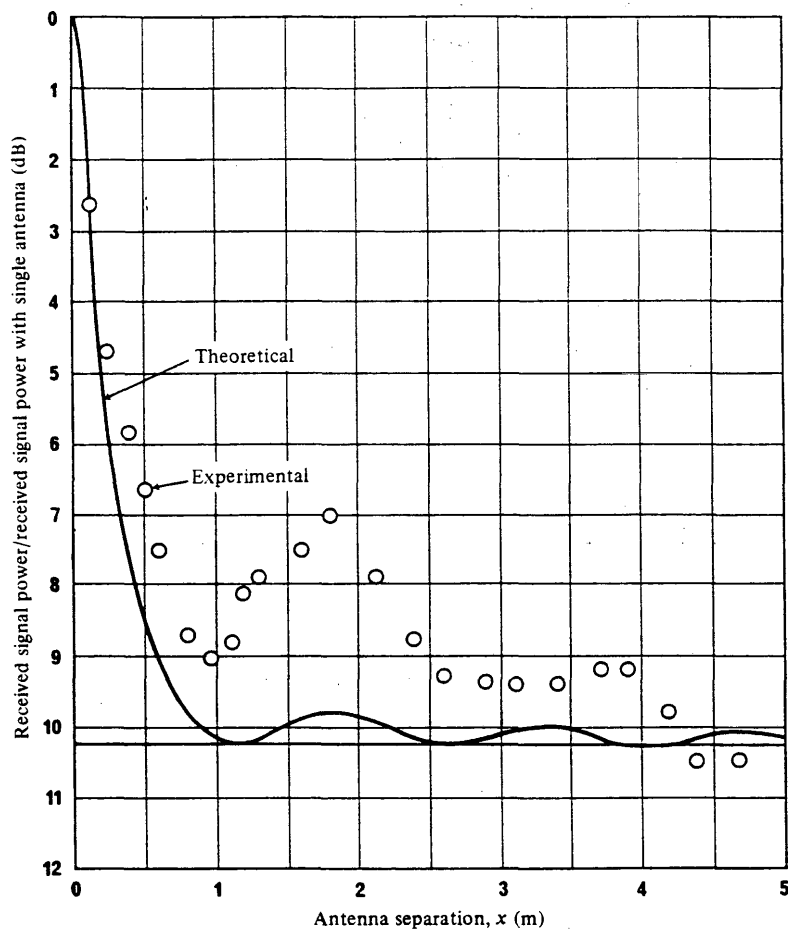


FIGURE 3 – Effects of antenna separation on performance of two-branch selection diversity combiner over that for single antenna

3. Selective fading

3.1 Effects of long-distance multipath propagation

Theoretical studies and field measurements were carried out in the Federal Republic of Germany in order to determine the effects of long-distance multipath propagation. From the result of these studies, two major conclusions can be drawn:

- sometimes the probability of noise bursts is much higher than that of level drop-outs and a relation cannot always be found between the instantaneous RF level and reception quality. Therefore, no level-sensitive diversity can completely solve the problems of selective fading;
- it may be possible to achieve a considerable improvement in reception quality by using new receiver techniques.

3.2 Adaptive de-emphasis

Selective fading causes incidental phase and amplitude modulation. When there is no level drop-out coincident with the selective fade-out, the unwanted amplitude modulation will be cancelled by the limiter. But there still remains the incidental phase modulation.

As shown in Figs. 4 and 5, there exists a relation between the envelope of the RF signal and the instantaneous value of interference in the baseband. The sharp extremes of the disturbance occur simultaneously with the zero crossing of the RF envelope. Therefore, a substantial part of the distortions may be eliminated by adaptive filtering in the baseband (Fig. 6). With an amplitude modulated-driven adaptive de-emphasis as shown in Fig. 7, the strong noise bursts caused by selective fade-out may be eliminated.

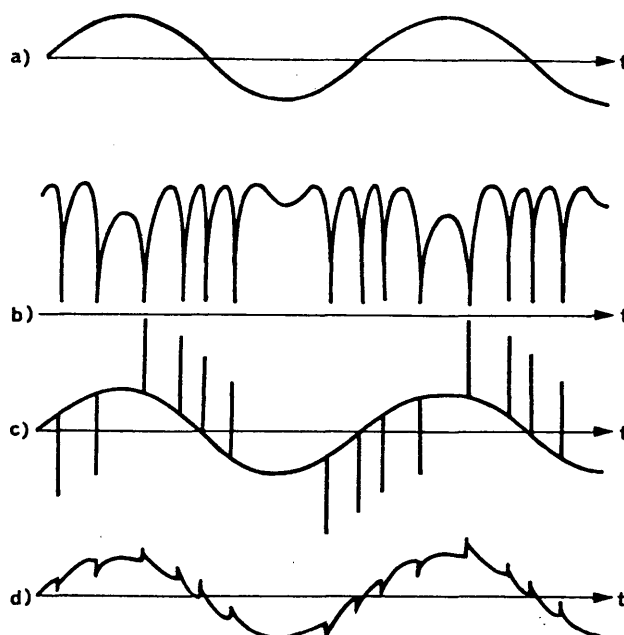


FIGURE 4 – Disturbances to the audio signal caused by selective fading when the antenna signal level is high

- a) modulating signal
- b) RF envelope
- c) audio signal before de-emphasis
- d) audio signal after de-emphasis

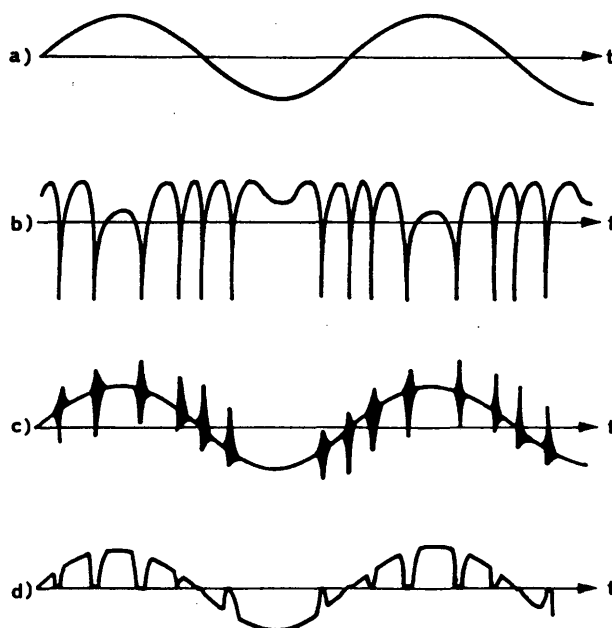


FIGURE 5 – Disturbances to the audio signal caused by selective fading when the antenna signal level is low

- a) modulating signal
- b) RF envelope
- c) audio signal before de-emphasis with low limiting threshold
- d) audio signal before de-emphasis with high limiting threshold

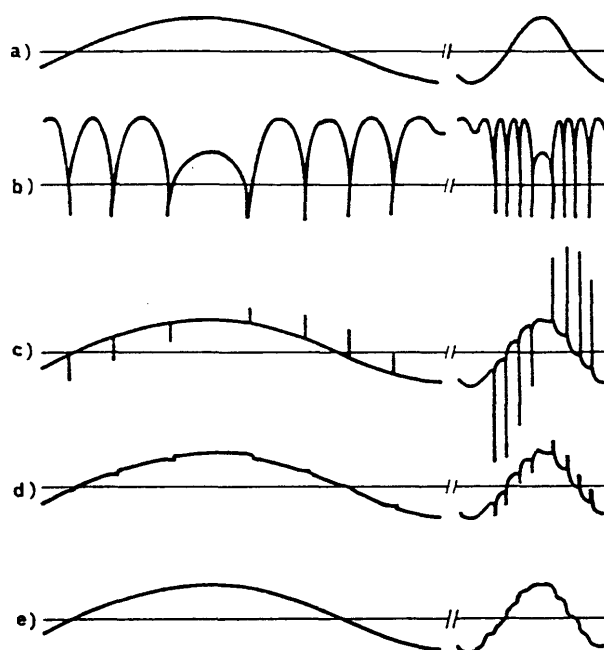


FIGURE 6 – Differences in disturbances depending on the modulating frequency

- a) modulating signal
- b) RF envelope
- c) audio signal disturbances without de-emphasis
- d) audio signal after sample and hold process
- e) audio signal after adaptive low-pass filtering

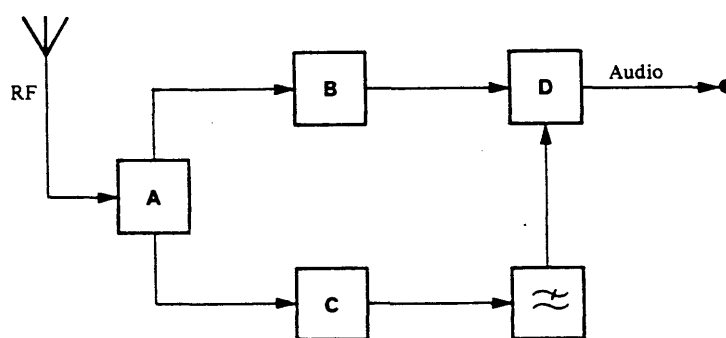


FIGURE 7 – Control of the de-emphasis to eliminate disturbances caused by selective fading

- A: RF-IF stages
- B: FM demodulator
- C: AM demodulator
- D: de-emphasis

4. Conclusion

The multipath effect produces severe signal drop-outs of FM broadcasts in a variety of typical motor vehicle operating environments, and thereby represents an impediment to the acceptance of this listening medium by the public. Space diversity receivers appear to offer significant performance advantages over conventional single-antenna arrangements for the reduction of non-selective fading effects and the improvement of FM broadcast reception in moving vehicles.

Adaptive filtering in the baseband offers significant performance advantages over the known receiver techniques when selective fading is present. Analysis shows that both diversity and adaptive de-emphasis can be realized separately or in conjunction for reception in moving vehicles. Thus, there exists the potential of changing intolerable listening conditions to good ones, for a wide range of problem situations.

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REPORT 1203

DIGITAL SOUND BROADCASTING TO MOBILE, PORTABLE AND
FIXED RECEIVERS USING TERRESTRIAL TRANSMITTERS1. INTRODUCTION

(1990)

Digital techniques have been used in sound programme production and transmission by some broadcasters for many years now, and more recently have become cheap enough to be introduced into the domestic consumer market, leading to wider public appreciation of high-quality sound, albeit via non-broadcast digital media. At the same time, there is rapidly growing congestion in the VHF/FM radio bands in many countries; thus the FM broadcasting services, which can deliver unimpaired sound quality into the home, are under threat of erosion of the quality deliverable. Overcrowding will inevitably increase the levels of interference which must be tolerated, particularly for mobile and portable receivers, which do not benefit from elevated, directional, receiving antennas, usually assumed in planning service coverage.

Although VHF/FM services can still provide excellent service to properly-installed fixed receivers, the solution for the future development of sound broadcasting is to provide an entirely new digital sound broadcasting service, designed at the outset to meet all the reception requirements of the diverse listening audiences. Also, a complete digital sound programme chain can be established from studio to domestic receiver.

This Report describes studies undertaken in Federal Republic of Germany, France, Netherlands, United Kingdom* and by EBU, using new types of digital coding and modulation specifically designed to provide a sound broadcasting service for mobile as well as static reception**. Additionally, new frequency allocation and planning techniques may be used to achieve significantly greater spectrum efficiency in network planning than has been possible previously using analogue FM systems.

2. PROPAGATION PATH AND CHANNEL CHARACTERISTICS

The diverse reception requirements for mobile and portable receivers create problems of shadowing by the local environment and use of frequencies in the VHF or UHF bands is desirable. This also leads to the need to cope with multipath propagation and man-made (in particular impulsive) noise; these can be particularly severe in cities. Conventional digital modulation systems are particularly sensitive to multipath signals since these create severe inter-symbol interference related to the differences in path delays.

* Under the European Eureka "Digital Audio Broadcasting (DAB)" joint research and development project.

** The Eureka DAB project is also studying modulation and channel coding schemes other than COFDM.

Early tests conducted in the UK <Eaton & Edwardson, 1978> demonstrated that mobile reception of experimental digital sound broadcasts at VHF was fairly disastrous, whereas fixed reception generally worked well. A more recent test at UHF using the digital sound signal associated with television broadcasting (see CCIR Report 795) has shown similar, if not so serious, failures in a mobile vehicle, but otherwise achieved a most satisfying high level of reception quality.

In relation to the digital channel bandwidth necessary, the multipath propagation may be frequency selective and also time varying. For a conventional digital modulation scheme, the achievable error performance is then strongly limited by the frequency selectivity and the fast field-strength variations with mobile reception. Studies of the statistical distribution of the field strength <Okumura *et al.*, 1968> have shown that it follows a log-normal distribution over large areas, coupled with a Rayleigh (no line-of-sight path) or Rice (direct path and Rayleigh) distribution within small areas (which are of dimensions of the order of a few hundred wavelengths).

In most respects the Rayleigh channel within dense urban areas is the least favourable and thus any new system must be designed to operate in this propagation environment. A certain amount of wideband propagation data is available <Cox & Leck, 1975, Bell & Williams, 1989>. Typically, at UHF, the 90% correlation bandwidth is of the order of 30 kHz with independence (i.e. <10% correlation) at frequency separations of the order of 3 MHz. Also the delay spread in urban areas is found to be of the order of 1-2 μ s, but exceeds 3 μ s for about 1% of locations. Some information is given in Report 955 (Annex II) and studies should be continued to provide a comprehensive wideband propagation model for the VHF and UHF bands and the different topographies encountered assuming terrestrial emissions*.

Nevertheless, a generalized broadcast model has been established <Pommier & Ratliff, 1988> which assumes a mobile receiver in a selective Rayleigh-fading environment. This is described in Annex I and is adequate for basic system design by considering where the channel is invariant on the one hand and statistically independent on the other. This leads to consideration of a broadband digital channel coding scheme.

3. CHANNEL CODING

3.1 General approach

A well-known solution to the selective fading channel is to employ spread-spectrum techniques which permit the temporal separation of the signals corresponding to the different transmission paths <Turin, 1980>; this permits their coherent addition through a transversal filter matched to the channel impulse response. The main weakness is the low spectrum utilisation attainable, usually below 0.25 bit/s/Hz, and this is unacceptable for broadcasting where severe restrictions on available spectrum apply.

* This requirement should be brought to the attention of Study Group 5.

Accordingly a new scheme, known as Coded Orthogonal Frequency Division Multiplex (COFDM) <Pommier & Wu, 1986>, has been devised which is well suited to a selective Rayleigh channel and yet despite being broadband, provides for efficient frequency planning.

Following from the broadcast channel model of Annex I, the first element of channel coding involves splitting the information to be transmitted into a large number of elementary sub-channels. Each sub-channel carries a low bit rate such that it remains locally invariant.

The second element involves using convolutional coding in conjunction with a Viterbi maximum-likelihood decoding algorithm, to provide a coding gain in excess of 20 dB at a bit error ratio of 10^{-3} , when the free distance of the code is larger than 10. Such a coding gain can only be achieved if the conditions of independent Rayleigh fading are met through successive symbols at the Viterbi decoder input. This independence is realised by appropriate frequency and time interleaving (in such a way that successive symbols are transmitted from large square to large square of the channel model given in Annex I).

Figure 1 shows the comparison of system performance in a Rayleigh fading channel with the case of the Gaussian additive channel, in relation to the free distance of the convolutional code used.

Furthermore, a convolutional code is well suited to realize a remarkable flexibility in bit-error protection performance <Weck, 1989>. If a low code rate is used as a basic convolutional code, each high code rate convolutional code can be derived by puncturing (i.e. not transmitting) certain code bits of the basic code. The desired code rate and therefore the desired bit-error protection performance can be realized with a single encoder and a single decoder using different puncturing. Such a system allows optimized bit-error protection with respect to the sensitivity to bit errors of the transmitted data. This adapted bit-error protection can provide a "graceful degradation" characteristic.

3.2 Orthogonal Frequency Division Multiplex (OFDM) technique

The OFDM technique splits the data stream of one broadcast channel into a large number, N , of modulated elementary carriers <Alard & Lassalle, 1987>. These modulated carriers (the OFDM signal) are simply derived from the orthogonal base of the Discrete Fourier Transform by using a vector of complex coefficients defining the I and Q components of each element of the base (each sub-channel carrier).

As shown in Figure 2, the main difference between conventional Frequency Division Multiplexing and this technique is that the spectra of the different carriers mutually overlap, giving an optimum spectrum efficiency (2 bit/s/Hz for a 4-PSK modulation of each carrier). Also, the sum of these N carriers exhibits a flat rectangular power spectrum. Because the signal satisfies orthogonality conditions, it is possible to extract the information from each modulated carrier separately by application of the Discrete Fourier Transform. Moreover, the modulation and the demodulation processes can be undertaken by means of Fast Fourier Transform algorithms.

Regarding the multiple paths of the broadcast channel, the condition of perfect orthogonality between carriers is no longer maintained at the receiver input, due to intersymbol interference. The solution to this problem comprises the addition of a guard interval before each symbol to absorb the intersymbol interference of duration shorter than this guard interval. Thus orthogonality between all carriers is maintained even in a multipath transmission channel under the condition that the channel memory is shorter than the guard interval.

3.3 Multiplexing technique

The multiplexing technique is chosen to permit reduced receiver complexity when each given service occupies only a part of the total channel capacity. Indeed, as the receiver has then to deal only with a part of the transmitted data multiplex, the coding and modulation is arranged such that one part of the data multiplex can be processed separately from the others. This is achieved by either a frequency or a time division multiplex, provided that the sub-channels which form the multiplex are encoded separately in the transmitter.

For instance, Figure 2 shows a frequency division multiplex of P programme channels each having N/P interleaved elementary carriers; P is effectively the ratio of the uncorrelated to highly-correlated bandwidth in the Rayleigh fading channel. Hence the Fast Fourier Transform of the receiver has to deal only with a sub-set of all N received carriers. It is not necessary to process all the Fast Fourier Transforms followed by programme selection, but, by combining them so as to use the programme selection as a procedure for decimating the Fast Fourier Transform (Alard & Lassalle, 1987), significant reduction in circuit complexity is achieved.

Although the receiver processing circuitry required is complex, all the processing functions are being developed on VLSI chips so that receiver cost should not be prohibitive if a large consumer market is stimulated.

3.4 Performance in a selective Rayleigh channel

A wide choice of modulation and channel coding parameters is under study allowing a trade-off between bit error ratio (BER) performance and spectrum efficiency. In particular, a 2, 4 or 8 PSK modulation scheme can be applied to each OFDM carrier, whilst either coherent or differential demodulation can be used. Additionally, a remarkable BER performance may be achieved by using an outer code of the Reed-Solomon type, concatenated with the inner convolutional code, at the price of a slight reduction of spectrum efficiency.

Figure 3 shows the COFDM performance in a Rayleigh fading channel using 2 and 4 PSK modulation for both coherent and differential demodulation. Figure 4 shows the performance with concatenated codes. The inner convolutional code has a constraint length of 7 and a rate of $1/2$, and the outer code is a Reed-Solomon type (336, 288) code, able to correct up to 24 erroneous words of 12 bits within a block of 336 transmitted words (Alard & Lassalle, 1987). It can be seen that the performance of 4 PSK modulation is close to that of 2 PSK, but it offers double the spectrum efficiency, i.e. 0.68 bit/s/Hz for this particular arrangement.

4. SOURCE CODING

Currently broadcasting systems incorporating digital sound transmission methods (e.g. see Reports 795, 1073) require between about 400 and 900 kbit/s capacity to convey a stereophonic programme. However, in order to make most efficient use of the scarce radio frequency spectrum for digital sound broadcasting, maximum bit-rate reduction of the digital sound source signal is also necessary, whilst conserving the full fidelity of the original studio signal.

Recent progress in sound coding compression techniques permits a high-quality studio signal, linearly coded to 16 bit quantisation accuracy at a 48 kHz sampling rate, to be reduced to around 2 bits/sample, yet retaining excellent sound quality (see Report 1199). Thus source transmission bit rates of only around 100 kbit/s per monophonic channel may be required.

Two different source coding techniques are under study, both of which attempt to match the permissible quantizing noise closely to the perception characteristics of the human ear. They both use audio spectrum analysis techniques and frequency-dependent bit allocation.

1) Transform coding

This coding method involves the conversion of a block of consecutive samples into the frequency domain (e.g. Fourier transform or the cosine transform) <Krahé, 1986, Brandenburg, 1988, Johnston, 1988>. This is equivalent to the time efficient computation of a filter bank with a large number (e.g. 512) of sub-bands. This strategy enables a reduction in the redundancy of the audio signal and also a matching of the quantization to the thresholds of perception of quantizing errors. Because of the large number of spectral coefficients (i.e. sub-bands), the masking threshold can be calculated very accurately. Only those values that are relevant with respect to the masking effects of the human ear are quantized.

ii) Sub-band coding

This coding method divides the broadband signal into a number of sub-band signals with a suitable filter bank and into digital frames of about 4 to 10 ms long <Theile et al., 1988, Dehery, 1988, Vörös, 1988>. Once in each frame, the maximum level attained by each sub-band signal, known as the scale-factor, is quantized and transmitted. Each sub-band is quantized with the bit-allocation based on the masking thresholds. This method avoids problems with time domain windowing and optimizes the distribution of the quantizing noise across the spectrum with regard to perceptibility, whatever the dynamic range of the input signal.

Sub-band coding is characterised by an inherent low sensitivity to transmission errors because the spectrum of the noise energy is confined to a single sub-band and is limited by the scale-factor provided that this is satisfactorily protected.

5. DEMONSTRATION OF SYSTEM APPROACH

An experimental system has been successfully demonstrated under the auspices of the EBU at the WARC-ORB(88) <Dosch et al., 1988> using a terrestrial UHF transmitter and a mobile receiver in the streets of Geneva. The system used COFDM* channel coding and MASCAM** sub-band sound source coding to provide 16 stereophonic sound programmes in a total bandwidth allocation of about 7 MHz. Excellent sound performance was demonstrated under a variety of severe multipath conditions. Further details are given in Report 955 (Annex IV).

6. FREQUENCY PLANNING

At a time when there is growing public interest in many countries in increasing the quality of sound broadcasting services available, it is of great concern that the existing frequency bands are becoming congested and there is, as yet, no further frequency allocation in the Radio Regulations for accommodating new sound broadcasting systems and services, even in the medium to long-term future. Although an allocation within the existing VHF or UHF broadcasting bands for a new digital service might in some cases be possible, one of the major problems in considering use of these bands is the difficulty of obtaining common channels universally in a given geographical area and the constraints placed by the existing systems in use. Careful assessment also needs to be made of the various possible uses of the available spectrum such as for new enhanced picture quality television services or new data broadcasting services. Consideration of extensions to the existing VHF and UHF broadcasting bands may therefore be required.

A minimum block of bandwidth is required to provide the necessary independence of multipath fading but typically up to 16 high-quality stereophonic programmes may be accommodated in a bandwidth of between 4 and 8 MHz. Nevertheless a single band of only 4 or 8 MHz would give very limited options for planning where different services are desired in adjacent geographical areas, e.g. in the border regions between neighbouring countries. In order to cover the large majority of the population, conventional area coverage terrestrial network planning requires some 10-20 times the amount of the basic spectrum necessary for a single transmitter.

The use of a single frequency network of terrestrial transmitters for a new digital service may be possible, at least in the VHF frequency range. Because the receiver is able to cope with multipath signals, it does not, in principle, know from which transmitter the signals have originated. It is only necessary to choose a sufficiently long guard interval to accommodate the difference in arrival times of the signals from the different synchronized transmitters, consistent with the overall symbol period being short enough to ensure that time coherence is maintained for mobile reception. Furthermore,

* COFDM stands for "Coded Orthogonal Frequency Division Multiplex"
** MASCAM stands for "Masking-pattern Adapted Sub-band Coding and Multiplexing"

any deficiencies in the coverage area may be filled in by providing a local, low-power, low-cost, 'repeater' transmitter on the same frequencies as those already used in the network. This novel feature may also find application in providing re-broadcast service within buildings and other heavily shielded areas (e.g. tunnels).

In practice, provided that signals from the different transmitters arrive with time differences of the order of the guard interval, T_G , or less they will combine constructively in the receiver. With a time difference greater than approximately $T_G + T_s/\sqrt{2}$ signals will combine destructively, (T_s : useful symbol period).

Consider as an example, an idealized VHF transmitter plan based on a regular triangular lattice with an adjacent transmitter separation corresponding to a propagation delay equal to the guard interval. It can be demonstrated that the effective minimum network-generated "carrier-to-interference" ratio provides an adequate margin for noise power in respect of system performance and long-range interference contributions with the system examples considered. However, the additional effect of sporadic-E interference which may be significant for the lower VHF broadcasting bands, has not been taken into account and requires further study.

7. SHARING CONSIDERATIONS

A further important factor is the significant reduction in interference protection required by this digital system, in relation to analogue broadcast systems, and its reduced potential to interfere with other radiocommunication services because of its uniform spectral characteristics. This offers better prospects of accommodating this proposed new broadcasting system into an already congested radio frequency spectrum. Further studies to establish protection ratios and inter-service sharing criteria are required.

In general, however, most efficient use of the spectrum for area coverage will be obtained with an exclusive VHF, or possibly UHF, band allocated to this system and using single frequency network techniques.

8. CONCLUSIONS

Studies of the concept and basic principles underlying an entirely new type of digital sound broadcasting system, which can be designed at the outset to serve mobile, portable and fixed home receivers effectively, have shown it to be a viable and attractive development for sound broadcasting.

The ability of a new channel coding scheme to overcome selective multipath propagation creates the opportunity to provide a reliable service to mobile and portable receivers with omni-directional antennas, even in dense urban and wooded rural areas. Additionally, new, more efficient, sound coding methods offer a significant reduction of spectrum requirements.

Further study is required to determine the optimum parameters for a digital sound broadcasting system. However, the necessary technological development is already in progress to provide for low-cost mass-production receivers within the next few years. There is an urgent need to determine protection ratios and sharing criteria for this type of system, but its basic broadband noise spectral characteristic makes it potentially much more attractive than analogue broadcasting systems.

It was demonstrated using an experimental system that a frequency block of 7 MHz (8 MHz channel) is required to provide up to 16 high quality stereophonic programmes in a particular service area, and that the ability to re-use such a single band of frequencies to provide area coverage from a number of geographically-separated co-channel transmitters offers the prospect of greatly improved utilisation of the radio spectrum. Complementary work recently undertaken, concerning the multiplex structure, planning methods and a further reduction of the digital audio bit-rates, suggests that smaller frequency blocks may be possible while preserving these advantages. For example, by broadcasting 12 to 16 stereophonic programmes within a 4 MHz block.

Further study is required to maximise the applicability of this new planning concept.

Whilst in principle, frequencies in Bands 8 and 9 are technically suitable, no specific band is reserved for such a service and in many countries the existing broadcasting bands are becoming full to capacity. Much of the spectrum efficiency of this new system may be lost if it has to share with analogue broadcasting services in the existing bands and thus most effective spectrum management will be effected if a contiguous block of spectrum can be allocated, at least in a given geographical region.

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ANNEX I

BROADCAST CHANNEL MODEL

For the critical case of a mobile receiver, the broadcast channel can be represented by a two-dimensional function of frequency and time. Considering the frequency domain, at a given time t , the channel response is the Fourier Transform of the channel impulse response $h(\tau, t)$ where τ denotes the time delay of the channel at instant t .

Now consider the time domain. For a given frequency f , or a narrow band around f , the field strength follows a random function defined by:

- the Rayleigh distribution

$$S(r) = \frac{r}{P} \exp\left(-\frac{r^2}{2P}\right) \text{ if } r \geq 0$$

where P is the mean power of the received field over a small area,

- the Doppler spectrum of which the Fourier transform gives the autocorrelation of the field strength variation

$$\gamma(\nu) = \frac{P}{\pi \sqrt{\left(\frac{\nu}{c} f\right)^2 - \nu^2}} \text{ if } -\frac{\nu}{c} f < \nu < \frac{\nu}{c} f$$

where $\frac{\nu}{c}$ denotes the ratio between the speed of the mobile and the speed of light

and $\frac{\nu}{c} f$ denotes the maximum Doppler frequency.

Figure 5 combines the channel frequency response and the time variation to give a two-dimensional channel representation, above which is a pattern of squares of different size:

- small squares indicate the frequency-time area where the channel may be considered as locally invariant,
- large squares show the minimum separation area for which two small squares are statistically independent.

Thus the size of the small square depends on the channel delay spread for the frequency axis and on the maximum Doppler frequency for the time axis. On the frequency axis, its length is less than one hundredth of the inverse of channel time spread, since, on the time axis, the length is less than one hundredth of the inverse of the maximum Doppler frequency.

The large square size is defined from the cross correlation of two frequencies spaced by ΔF <Jakes, 1974>

$$\rho(\Delta F, \Delta T) = \frac{J_0^2\left(2\pi \frac{v}{c} f \Delta T\right)}{1 + (2\pi \Delta F T_0)^2}$$

where $J_0(x)$ is the Bessel function of order zero,

ΔT is the time shift between frequency f and $f + \Delta F$ respectively and

T_0 is the mean delay spread.

To obtain independence on the frequency axis, whatever the speed of the vehicle, the condition

$$\frac{1}{1 + (2\pi \Delta F T_0)^2} \ll 1$$

must be satisfied. Thus for $\rho \ll 0.1$ the length of a large square on the frequency axis is about $\frac{1}{2 T_0}$

Independence with regard for the time axis depends only on the maximum Doppler frequency. The length of large square along the time axis is given by the first zero of $J_0(x)$, i.e.

$$\Delta T = \frac{2.5}{2\pi} \cdot \frac{c}{v} \frac{1}{f}$$



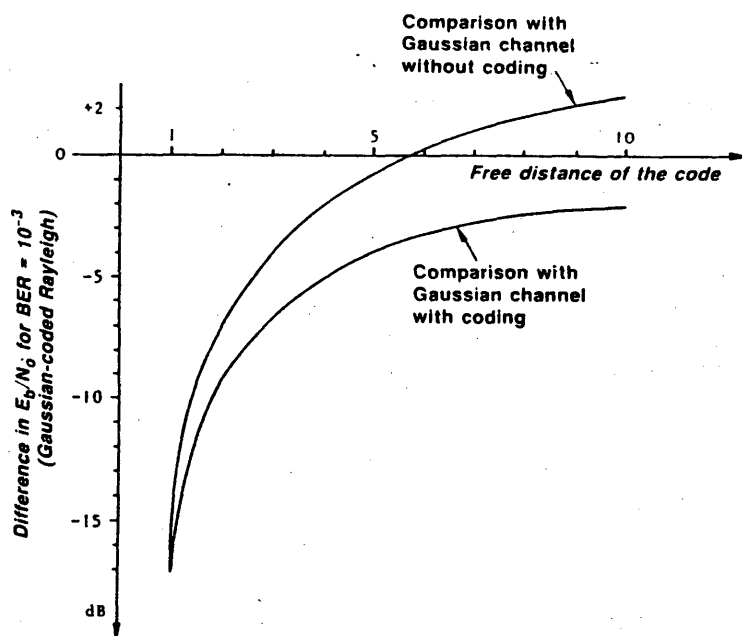


Figure 1 Comparison of convolutional coded system performance in a Rayleigh channel with the case of Gaussian additive channel

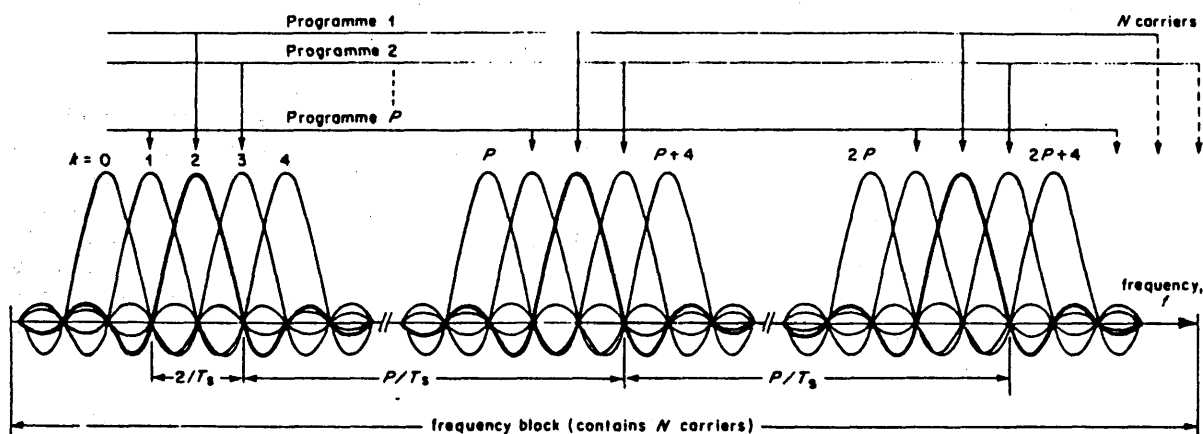


Figure 2 Spectrum arrangement for interleaved multi-carrier sound signals using OFDM technique

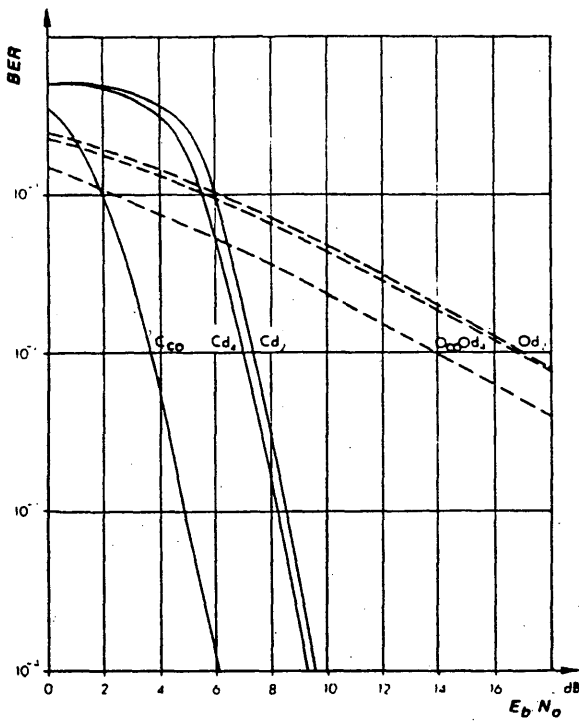


Figure 3 Performance comparisons for the convolutional (rate 1/2, $k = 7$) COFDM systems in a Rayleigh channel

- C convolutional code
- O no code
- co coherent demodulation
- d2 differential demodulation (2-PSK)
- d4 differential demodulation (4-PSK)

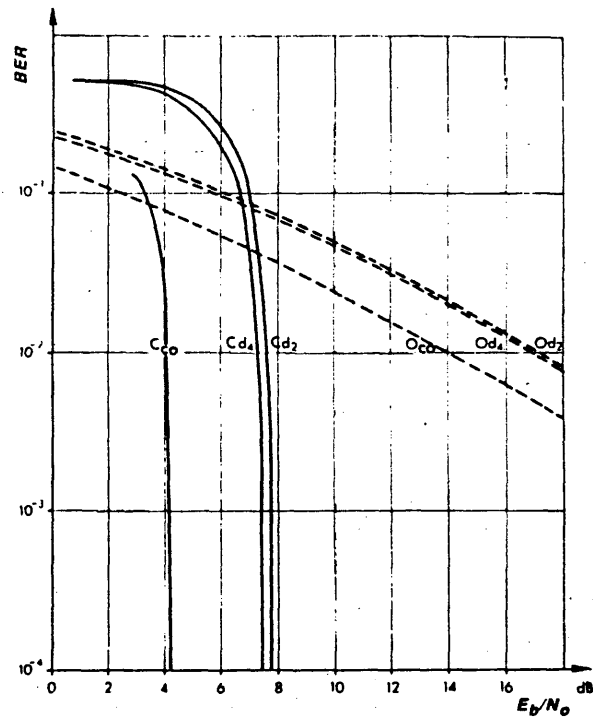


Figure 4 Performance comparisons for the concatenated convolutional (rate 1/2, $k = 7$) and R-S type (336, 288) block COFDM system in a Rayleigh channel

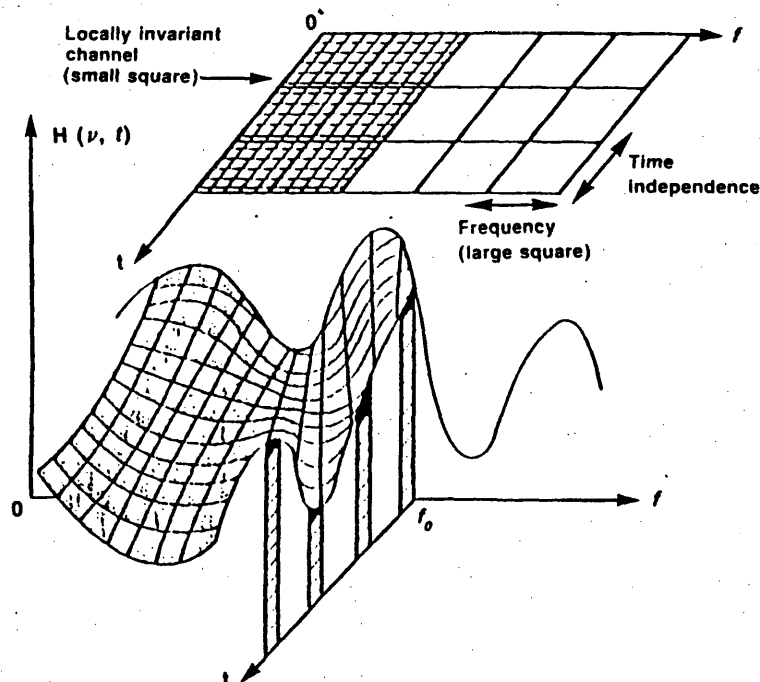


Figure 5 Channel frequency-time response for mobile reception

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REPORT 1198

COMPATIBILITY BETWEEN THE BROADCASTING SERVICE IN
THE BAND 87.5 - 108 MHz AND THE AERONAUTICAL
SERVICES IN THE BAND 108 - 137 MHz

(Study Programme 46J/10)

(1990)

This Report is published separately.

REPORT 795-3

TRANSMISSION OF TWO OR MORE SOUND PROGRAMMES OR
INFORMATION CHANNELS IN TELEVISION

(Question 47/10, Study Programme 47A/10)

(1978-1982-1986-1990)

1. Introduction

High quality multi-channel sound has been shown to be an asset to television programme transmissions for stereophonic sound reproduction and for separate sound channels. Studies conducted since 1959 have led to the development of several different methods of adding additional sound channels and data transmissions to television systems in use throughout the world. Some of these systems have been implemented with two being recommended, see Recommendation 707, and two being proposed in Appendices I and II to Annex I to this report.

A system for transmission of stereophonic or two separate sound signals using two sound carriers, "the two-sound carrier FM system", is in regular service as described in section 3. This service began in the Federal Republic of Germany in 1981 and has also been introduced in other countries, e.g. Australia, Italy, Netherlands and in the Republic of Korea.

The NICAM-728 digital system, where multi-channel sound and data are carried on a digitally modulated carrier additional to the analogue sound carrier has been studied in the United Kingdom and a number of other countries, and by the EBU. This system was put into regular service in Denmark, Finland and Sweden in 1988 and in New Zealand and the United Kingdom in 1989. The system is described in section 4. Studies are proceeding in France on the adoption of the digital system to television system L [CCIR, 1986-90a].

In 1978, an FM-FM system of multiplexing two sound channels on the single sound carrier of television system M was put into service in Japan. Stereophonic sound or two separate sound channels may be provided by this system as described in section 5. Work is under way to implement data services, for example facsimile broadcasting to provide additional text services.

In North America, the United States and Canada have implemented the BTSC* FM-AM multiplex system of multi-channel sound with television broadcasting beginning in 1984. The BTSC system permits the simultaneous transmission of

* Broadcast Television Systems Committee of the Electronic Industries Association, which adopted this system after five years of study.

stereophonic sound, an additional monophonic sound programme (second audio programme or "SAP") service, and a narrowband or data channel for station operation purposes. At this time, the BTSC system as described in section 6 is preferred for use in the United States but improved systems are possible.

2. System requirements

In addition to the capacity requirements of the additional sound or information channels, the following requirements should be fulfilled:

- the system should be compatible with existing receivers and networks;
- the system should not involve any increase in the bandwidth of a television channel;
- the additional sound or information channel should have at least the same coverage area as that of the picture channel;
- the system for an additional sound channel should provide high-quality reproduction of both channels and stereophony;
- for receiving the additional sound channel, an ordinary receiver and a relatively simple adapter can be used.

3. Two-sound carrier FM system

The system is recommended for analogue multisound in PAL television systems B, G and H. Its principal specifications are given in the proposal for a new Recommendation shown in Annex I.

3.1 Principles of the system

The two-carrier FM system provides an additional sound carrier for the second sound signal. The additional sound carrier is located at a higher frequency than the normal sound carrier and is set at a level lower than that of the first sound carrier.

3.2 Transmitter and receiver aspects

The system is implemented using simple well established techniques. Receiver cost is low. The circuits for the two sound carriers are identical for transmitters as well as receivers.

3.3 Stereophonic coding

The first sound channel is modulated by signal M, equal to one half of the sum of the left-hand signal A and the right-hand signal B and the second channel by signal B only. This coding is chosen in order to improve the audio S/N ratio in receivers which use intercarrier demodulation.

3.4 Coverage area

The service area of both sound channels is larger than the service area of the vision signal. Even when propagation conditions are unfavourable (long-distance and multipath propagation), reception is still satisfactory for both sound channels.

3.5 Quality of the two sound channels

The following figures are typical for modern television receivers using inter-carrier demodulation [CCIR, 1986-90b]. Further improvements are possible, e.g. with split-sound receiver types, synchronous detection or audio-frequency companding:

- weighted signal-to-noise ratio (CCIR Recommendation 468 peak-to-peak): 50 dB for normal pictures, 45-50 dB for those pictures causing the most interference (the possibilities for the use of an additional compatible companding system to improve the S/N ratio are being studied);
- cross-talk attenuation: dual-sound mode >62 dB;
- cross-talk attenuation: stereophonic mode >36 dB;
- audio-frequency response: ± 0.7 dB between 40 Hz to 15 kHz;
- harmonic distortion: <1%.

3.6 Compatibility

More than 10 million dual sound receivers are in use in 1989 in the Federal Republic of Germany. The system is operating without compatibility problems in the whole receiver population (i.e. including receivers not designed for the dual sound service) [CCIR, 1986-90b]. A similar situation exists in other countries which have introduced the system.

3.7 Data transmission

Laboratory tests carried out in Italy [CCIR, 1986-90c] have shown that during monophonic television programmes the second sound channel is available to transmit data at 4.8 kbit/s without affecting the picture or sound. In this particular service the data signal is sent via the peritelevision connector to a home computer [CCIR 1986-90d, e], [d'Amato et al., 1987].

4. NICAM 728 digital carrier system

This system is recommended for digital multisound in PAL television systems B, G, H and I, see Recommendation 707. A summary specification is given in Annex II to that Recommendation.

4.1 Principal characteristics

The NICAM 728 system uses an additional sound carrier located at a higher frequency than the normal sound carrier and set at a level lower than that of the first carrier. This second carrier is modulated using a process of quadrature phase shift keying by a digital signal with a bit rate of 728 kbit/s. This signal can carry either two high quality sound signals, which may be a stereo pair or independent mono, or one such sound signal with data, or data alone; associated signalling bits indicate which option is in use. The use of these signalling bits is not defined to allow for the possibility of other sound/data combinations in the future. There is also a provision for 11 kbit/s of ancillary data. High quality sound is linearly encoded at 32 kHz sampling rate and 14 bits per sample, the latter being reduced to 10 bits per sample by a process of near-instantaneous companding as defined in Report 647.

The system has the following features:

- it provides two high-quality sound channels in addition to the existing analogue sound channel;
- it is sufficiently rugged to ensure that reception of the vision signal fails before reception of the digital sound signal under difficult reception conditions;
- it gives satisfactory compatibility with existing services and receivers in over-the-air transmission and in distribution on cable systems;
- the sound coding is identical with that of one option available in the MAC/packet family of systems [CCIR, 1988].

4.2 Coverage area

Field trials with systems B, G and I show that the digital signal can be transmitted from a main station and travel acceptably through up to seven television transposers. In addition, the digital signal was found to be rugged against impaired reception due to weak signal conditions, multipath propagation and ignition interference [CCIR, 1986-1990f, g, h]. The service area for satisfactory reception is therefore equal to or greater than the area for satisfactory colour television reception.

4.3 Quality of the digital channels

The signal-to-weighted-noise ratio of a digital sound system is limited only by the inherent quantization accuracy of the digital sound coding system used. The main characteristics of the 14/10 near-instantaneous companding system are given in section 4.1.

The cross-talk attenuation between the two additional digitally coded sound channels is found to exceed 70 dB. Cross-talk attenuation from the normal mono sound channel into the digital sound channel also exceeds 70 dB. Such cross-talk is due only to implementation imperfections in the analogue parts of the transmission chain.

In the digital system, harmonic distortion is not inherent. Any harmonic distortion that is produced will depend upon the performance of the associated analogue circuits, including the A/D- and D/A- converters.

4.4 Compatibility with existing receivers

The introduction of the digital sound carrier could in principle cause interference into either the analogue sound or the picture. Over-air tests with system I and systems B and G have been carried out [CCIR, 1986-90i, j]. No significant problems with patterning on the picture were reported. Test results are similar to those obtained with the analogue two-carrier system.

Regarding interference from the digital signal into the analogue sound channel, results from over-air tests and supplementary laboratory measurements show that no serious problems are envisaged [CCIR, 1986-90f].

Results from operation of the digital system show no significant problems with existing receivers [CCIR, 1986-90k, l].

4.5 Compatibility with existing networks

Relating to interference from the NICAM 728 signal on a lower adjacent channel into the picture signal of a wanted channel, laboratory tests, tests in cable networks, off-air tests as well as regular operation with system I, B and G show no significant problems [CCIR, 1986-90f, k, m, n, o, h, p, q].

For the case of interference from the picture signal of an upper adjacent channel interfering into the NICAM channel of the wanted signal, tests in the United Kingdom as well as regular operation in Denmark demonstrated no significant problems with upper adjacent channel interference [CCIR, 1986-90k, p, q]. Any such interference could be limited by a slight modification of the vestigial sideband filter for the upper adjacent channel [CCIR, 1986-90r].

Calculations and laboratory measurements in Finland relating to interference from system B/PAL with NICAM 728 into system D/SECAM has demonstrated that the protection ratio to be applied in planning decreases 2 dB since the level of the FM sound carrier is lowered from -10 to -13 dB relative to the peak vision carrier [CCIR, 1986-90s].

Calculations and laboratory tests [CCIR, 1986-90f, t] have shown that the use of the digital sound system in system I will comply with the protection ratios given in CCIR Recommendation 655 for continuous and tropospheric interference for this system. Protection ratios applicable to co-channel interference from system I with NICAM 728 into system L are under study in the United Kingdom, in Study Group 11 and by the EBU.

Report 1214 gives information on the protection ratios applicable to situations with a digital signal involved.

4.6 Receiver population

Within one year of the introduction of the digital system in 1988 in Denmark, Finland and Sweden, more than 200,000 receivers for this system were in use in these countries.

4.7 Transmitters

The system can be easily implemented in both new and older transmitters. Experience in Denmark shows that in order to allow for margins in large cable networks, the power spectrum of the NICAM 728 signal at the output of the terrestrial transmitter should be within ± 1 dB relative to the ideal response for frequencies in the range $+5.85$ MHz ± 150 kHz relative to the vision carrier [CCIR, 1986-90k].

5. FM-FM system for NTSC television system M

5.1 Principles of the system

Specifications of the first sound channel and the vision channel in the system are the same as that of television system M.

The FM-FM system makes use of a sub-carrier in the normal sound channel. The frequency of the sub-carrier, which is frequency-modulated by the second sound signal, is chosen to be the second harmonic of the picture line frequency, because this reduces interference from the picture into the second sound channel.

The second sound channel has an audio-frequency response of up to 14 kHz. The system can be used for high quality transmission of either two independent programmes or one stereophonic programme.

The system specifications are given in Appendix I to Annex I to this report.

5.2 System performance

5.2.1 Receivers

Laboratory and broadcasting tests made in Japan showed that, with suitable design of the sound intermediate-frequency circuits, inter-carrier receivers were suitable for reception of a second sound channel having an audio-frequency response of up to 14 kHz.

5.2.2 Stereophony

In the FM-FM system, compatibility with existing standards necessitates modulation of the first and the second channels by the signals M and S respectively, as defined in Recommendation 450.

5.2.3 Coverage area

In the FM-FM system, the first sound channel has the same coverage area as the ordinary monophonic transmission and the second sound channel has the same coverage area as that of the picture channel.

Even when propagation conditions are unfavourable (long-distance multipath propagation), the system is still satisfactory for stereophony.

The FM-FM system was examined with respect to the effects on the audio-frequency signal-to-noise ratio under the conditions of line and precision-offset [CCIR, 1978-82].

5.2.4 Quality of the second channel

Signal-to-weighted noise ratio, including buzz:

- 51 dB (inter-carrier receiver).

Cross-talk from the main sound channel into the second sound channel:

- Under normal operation, the cross-talk attenuation from the first sound channel into the second sound channel was found to exceed 55 dB.

Harmonic distortion:

- 44 dB (inter-carrier receiver).

5.2.5 Stereophonic quality

The cross-talk attenuation between the stereophonic signals A and B is better than 30 dB (0.4 to 7 kHz; inter-carrier receiver).

5.2.6 Compatibility (vision and sound)

5.2.6.1 Compatibility with existing receivers

In the operational service in Japan, no complaint concerning interference into the picture channel has been reported and very few complaints of interference into the first sound channel were received. Since it is estimated that more than thirty million viewers watch the programmes every day, the compatibility of the FM-FM system is said to be good.

In the case of the digital broadcast facsimile system described in § 5.3.1, there is no interference to the picture concerned as well as to both sound channels of existing receivers of FM-FM multi-channel sound systems, if the maximum frequency deviation of the main sound carrier by the facsimile sub-carrier is limited to ± 2 kHz.

In the case of the analogue broadcast facsimile system described in § 5.3.2, the maximum frequency deviation of the main sound carrier by the facsimile sub-carrier has to be limited to ± 10 kHz. With this deviation interference occurred in very few receivers, but if a trap for the sixth line frequency multiple is inserted into the receiver, this problem would be eliminated.

5.2.6.2 Compatibility with existing networks

In Japan, with rebroadcasting by 1355 stations operated using the FM-FM system, no interference to the picture and the first sound channel was detected even when five rebroadcasting stations were operated in cascade. Interference from the picture signal into the second sound channel did not increase significantly with inter-carrier reception.

In broadcasting the FM-FM system operating in Japan no increase in interference to the picture and the first sound channel from adjacent channels and co-channels was observed.

5.3 Broadcast facsimile systems

Broadcast facsimile signals can be transmitted by a DQPSK- or FM-modulated sub-carrier to the main sound carrier simultaneously with the sound signal carried by the sound sub-carrier.

5.3.1 Digital system

A digital broadcast facsimile system developed in Japan [CCIR, 1986-90u] emphasizes compatibility with G3 (conforming to CCITT Recommendation T.4) apparatus and provides for facsimile transmission of ISO A4 page size text at 9600 bit/s net data signalling rates.

As to the signal format, the system employs a packet type signal which incorporates a majority logic decodable (272, 190) shortened difference set cyclic code for error correction.

The transmission characteristics are shown in Appendix I to Annex I of this report.

5.3.2 Analogue system

An analogue broadcast facsimile system has been developed in Japan [CCIR, 1986-90u]. It has a 2.5 minutes constant transmission time per page of ISO A4 size. Transmission of colour documents takes the same time as transmission of monochrome documents. In the colour transmission mode, chrominance components compressed to a 1/2 time axis are transmitted every two lines by substituting for the even numbered luminance lines. By this substitution a resolution corresponding to the normal mode on G3 apparatus is obtained. In the monochrome mode, a resolution corresponding to the fine mode on G3 apparatus is obtained.

The transmission characteristics are shown in Appendix I to Annex I of this report.

6. BTSC FM-AM system

6.1 Principles of the system

The BTSC system for multichannel television sound consists of an M signal derived from the sum of the left and right sound signals. This signal is compatible with monophonic receivers in that it has the same preemphasis and peak sound carrier deviation as a monophonic signal. An S signal is derived from the difference of the left and right sound signals and is compressed to make an S' signal that modulates a suppressed-carrier amplitude modulated sub-carrier. The frequency of the sub-carrier is 31 468 Hz, the second harmonic of a pilot-tone of line frequency. The pilot-tone is used to synchronize the encoding of the S sub-channel.

The compressor forms the encode portion of a compressor/expander (comparator) used to reduce the noise level on the S sub-channel to an acceptable level. This sub-channel noise level is higher in television stereophonic operation than in FM sound broadcasting because virtually all NTSC system M television receivers employ intercarrier sound detectors to recover the sound signal.

To maintain maximum stereophonic separation, the S sub-channel encoder (compressor) must be matched in amplitude exactly by the corresponding decoder (expander) in the receiver. In order to maintain the required amplitude match, each portion of this comparator system (both the encoder and decoder) must have a common amplitude reference. The absolute deviation of the transmitter is this reference.

The system specifications are given in Appendix II of Annex I to this report.

6.2 Additional signals

Provision is also made in the BTSC for the transmission of additional signals on the sound carrier.

One signal is a monophonic sound sub-channel (separate audio programme: SAP) at five times the horizontal line frequency. It uses noise reduction encoding similar to what is used on the S channel, and provides a 50 Hz to 10 kHz audio frequency response. It is intended for reception by a wide audience on stereophonic receivers.

Another signal is a non-programme sound and data channel at 6.5 times the horizontal line frequency. It is a voice bandwidth sub-channel intended for operational (non-public) use by the broadcast authority.

6.3 System performance

The BTSC system was chosen by the United States as its preferred system following comprehensive evaluation [EIA/BTSC, 1983].

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ANNEX I

PROPOSAL FOR A DRAFT NEW RECOMMENDATION

TRANSMISSION OF MULTISOUND IN TERRESTRIAL TELEVISION SYSTEM M

(Question 47/10, Study Programme 47A/10)

The CCIR,

CONSIDERING

- (a) the increasing requirement world-wide for suitable means of broadcasting stereophonic and/or multi-channel sound and/or data from terrestrial television transmitters;
- (b) the technological development in this area and in particular the relative merits of various possible analogue and digital methods, and described in Report 795;
- (c) the fact that the United States and Canada have used the BTSC FM-AM system as their preferred system beginning in March 1984 and that a public service of bilingual and stereophonic television programmes using the FM-FM system has been operation in Japan since 1978;
- (d) the advantage of low-cost analogue circuitry for multisound television receivers for system M;
- (e) the desirability of establishing standards for the introduction of stereophonic and/or multi-channel should for the terrestrial television broadcast services using system M,

RECOMMENDS

1. that if analogue multisound is to be introduced in terrestrial broadcasting in countries using NTSC television system M, it should be implemented using the FM-FM or BTSC system, and be implemented as defined in Appendices I and II to this draft new Recommendation.

Note 1 - The transmission systems described, in some cases, can be used for data services. Where applicable, reference to these data services will be found in the annexes containing the system specifications.

Note 2 - A modified two-sound carrier FM system similar to the system described in Table I of Recommendation 707, Annex I, has been introduced in the Republic of Korea for system M.

Appendix I to Annex I

System specifications for the FM-FM system

TABLE I

Transmission characteristics of the FM-FM system
(Television system M)

1. The first channel (the same as the monophonic channel)	
Maximum frequency deviation of the main carrier (kHz)	± 25
Audio-frequency range (Hz)	50 to 15 000
Pre-emphasis (μ s)	75
2. The second channel	
Sub-carrier frequency	second harmonic of the line frequency
Maximum frequency deviation of the sub-carrier (kHz)	± 10
Frequency deviation of the main carrier by the sub-carrier (sub-carrier level) (kHz)	± 15
Audio-frequency range (Hz)	50 to 14 000
Pre-emphasis (μ s)	75
3. Stereophony	
(A + B) channel	the same as the first channel
(A - B) channel	the same as the second channel except the sub-carrier level
Frequency deviation of the main carrier by the sub-carrier (sub-carrier level) (kHz)	± 20
The left-hand signal produces a deviation in the same direction in both the sub-carrier and the main carrier	
Compensation for receiver delay time (μ s)	20
4. Control signal	
Sub-carrier frequency	3.5 times the line frequency
Modulation frequency:	two sound programmes: 922.5 Hz stereophonic programme: 982.5 Hz
Modulation	60% (AM)
Maximum frequency deviation of the main carrier by the control signal sub-carrier (kHz)	± 2

TABLE II

Transmission characteristics of digital facsimile broadcast
multiplexed with the FM-FM system

Facsimile channel	
Sub-carrier frequency	70.804 kHz
Bandwidth	12.8 kHz
Modulation method	DQPSK
Data signalling rates	16 kbit/s
Spectrum shaping	60% square root cosine shaping (equally shared between transmitter and receiver)
Data scrambling	11th M-sequence, pseudo-random noise signal $g(X) = X^{11} + X^2 + 1$
Maximum frequency deviation of the main carrier	± 2 kHz

TABLE III

Transmission characteristics of analogue facsimile broadcast
multiplexed with the FM-FM system

Facsimile channel	
Sub-carrier frequency	94.4 kHz
Bandwidth	63 kHz
Modulation method	FM
Frequency deviation	+16 kHz, -8 kHz
Pre-emphasis	50 μ s
Maximum frequency deviation of the main carrier	± 10 kHz

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APPENDIX II TO ANNEX I

SPECIFICATIONS FOR THE BTSC FM-AM SYSTEM

1.1 A main (sum) channel containing audio frequencies from 50 to 15 000 Hz, with peak modulation corresponding to +/- 25 kHz deviation and with 75 microsecond pre-emphasis.

1.2 A pilot sub-carrier at the horizontal line frequency 15 734 Hz +/- 2 Hz, peak deviation of +/- 5 kHz. This pilot sub-carrier is to be frequency locked to the horizontal line frequency of the transmitted video signal. Stations not transmitting multi-channel sound according to this specification must limit modulation of the main sound carrier at 15 734 Hz +/- 20 Hz, to no more than 0.125 kHz deviation to prevent false activation of stereo mode reception in receivers.

1.3 A stereophonic sub-carrier at the second harmonic of the pilot sub-carrier, double sideband amplitude modulated, suppressed carrier, with the stereophonic difference signals from 50 to 15 000 Hz. This sub-carrier must cross the time axis with a positive slope simultaneously with each crossing of the time axis by the pilot carrier, and within 3 degrees (approximately +/- 0.53 μ s) of the zero crossings of the pilot sub-carrier. During the time period in which a left-only signal applies, the main channel modulation must cause an upward deviation of the sound carrier. The unmodulated sub-carrier must be suppressed to a level less than 0.25 kHz deviation of the sound carrier.

1.4 A frequency-modulated second audio programme sub-carrier, frequency locked to the fifth harmonic of the horizontal line frequency (pilot sub-carrier). When modulated, the centre frequency should nominally be that of the the fifth harmonic of horizontal line frequency with a tolerance of +/- 500 Hz. The frequency response of this channel is 50 to 10 000 Hz and modulates the sub-carrier with a peak deviation of +/- 10 kHz. This sub-carrier modulates the sound carrier with a peak deviation of +/- 15 kHz.

1.5 Audio amplitude and spectral compression (for complementary expansion in receivers) providing approximately 30 dB of background noise improvement in both the stereophonic sub-channel and the second audio programme channel. The compression has the following characteristics, where f is expressed in kHz (see also Figure 1).

1.5.1 Fixed pre-emphasis ($F(f)$) whose transfer functions is as follows:

$$F(f) = \frac{(jf/0.408)+1}{(jf/5.230)+1} \cdot \frac{(jf/2.19)+1}{(jf/62.5)+1}$$

1.5.2 Wideband amplitude compression wherein:

- (a) The decibel gain (or loss) applied to the audio signal during encoding is equal to minus one times the decibel ERMS value of the encoded signal (the result of the encoding process), weighted by a transfer function ($P(f)$) as follows:

$$P(f) = \frac{(jf/0.3254)}{((jf/0.0354)+1) \cdot ((jf/2.09)+1)}$$

where ERMS (Exponentially time-weighted root mean square value equals

$$\text{ERMS value} = \sqrt{\frac{1}{T} \int_{-\infty}^t S^2(u) e^{-(t-u)/T} du}$$

and where $S(u)$ is the waveform in question, a function of time, T is the exponential time-weighting period, and t is the time at which the ERMS value is computed.

- (b) The exponential time weighting period T_1 of the ERMS detector referred to above in (a) is 34.7 ms.
- (c) The zero decibel reference ERMS value for the encoded signal referred to above in (a) is 8.99% modulation of the sub-carrier at 0.300 kHz.

1.5.3 Spectral compression wherein:

- (a) The transfer function $S(f,b)$ applied to the audio signal during encoding is:

$$S(f,b) = \frac{1 + (jf/F) \cdot (b+51)/(b+1)}{1 + (jf/F) \cdot (1+51b)/(b+1)}$$

where $b=10^{D/20}$

$F=20.1$ kHz

D = decibel r.m.s. value and b is the decibel ERMS value of the encoded signal (the result of the encoding process) weighted according to a frequency transfer function $Q(f)$ as follows:

$$Q(f) = \frac{(jf/5.86)^3}{((jf/7.66)^2 + (jf/7.31) + 1) \cdot ((jf/26.9) + 1) (jf/3.92) + 1}$$

- (b) The exponential time weighted period T_2 of the ERMS detector referred to above in 1.5.3(a) is 11.4 ms.
- (c) The ERMS zero decibel reference for the encoded signal referred to above in 1.5.3(a) is 5.16% modulation of the sub-carrier at 8 kHz.

1.5.4 Overmodulation protection functionally follows the functions of 1.5.1, 1.5.2, and 1.5.3 above.

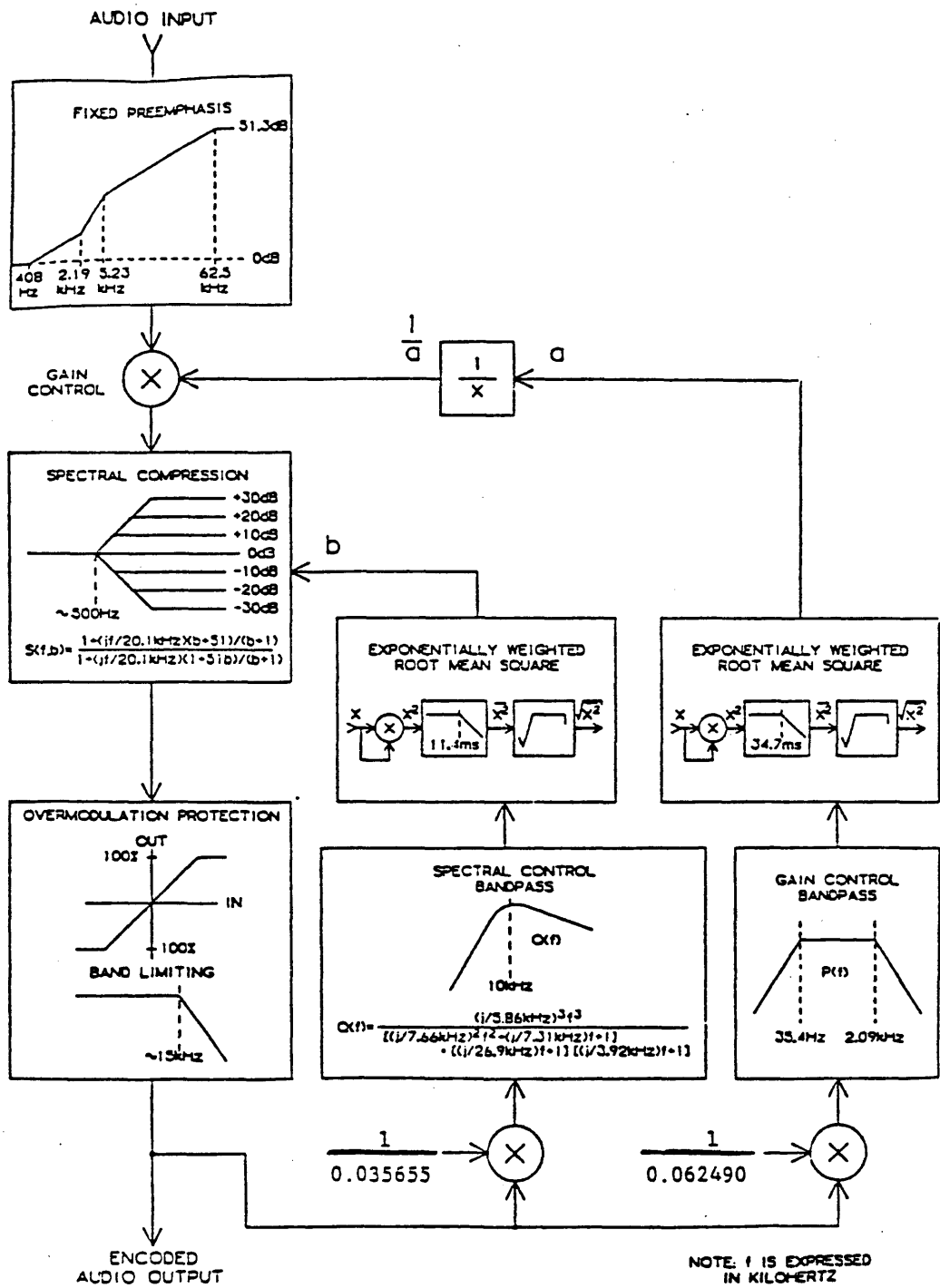
1.5.5 Band limiting to appropriately restrict bandwidth functionally follows the functions of 1.5.1, 1.5.2, and 1.5.3 above.

1.6 A non-programme related sound sub-carrier at 6.5 times the horizontal line frequency (pilot sub-carrier frequency) with a tolerance of ± 500 Hz. The peak deviation of this sub-carrier is limited to ± 3 kHz. This sub-carrier may be used to transmit audio or telemetry and control data signals. When the second audio programme is not transmitted a sub-carrier frequency between 47 and 120 kHz may be used for which the deviations are not specified.

1.7 A limit of 73 kHz instantaneous deviation of the aural carrier by the sum of all multi-channel sound signals.

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Note: Certain values in this figure are approximate. Exact values are found in the accompanying text.

FIGURE 1
STEREOPHONIC DIFFERENCE AND
SECOND PROGRAM AUDIO SIGNAL
ENCODING

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SECTION 10C: AUDIO-FREQUENCY CHARACTERISTICS OF SOUND-BROADCASTING SIGNALS

REPORT 292-6*

MEASUREMENT OF PROGRAMME LEVEL IN SOUND BROADCASTING

(Question 50/10. Study Programme 50E/10)

(1963-1966-1970-1974-1978-1982-1986)

The CCIR and CCITT have considered for many years: "by what methods and by means of what equipment should the programme level be controlled in connection with recording, reproduction and transmission over lines or radio links?" (Former Question 151.)

As far as transmission over lines or radio links is concerned, CCITT Recommendations J.14, J.15 and N.13 apply and should be respected (see Annex I).

It has not been possible to reach agreement on the use of a single type of meter. It is not possible, therefore, to do more than to state the present practices. Table I shows the fundamental characteristics of meters recommended at present by the CCITT, EBU, OIRT and IEC.

It is hoped that further contributions will be submitted which will make it possible to recommend uniform fundamental characteristics of the meters.

* This Report should be brought to the attention of the CMV.

TABLE I – Principal characteristics of the various instruments used for monitoring the volume or peaks during programme transmissions

Type of instrument	Rectifier characteristics (Note 3)	Time to reach 99% of final reading (ms)	Integration time (ms) (Note 4)	Time to return to zero (value and definition)
(1) CCITT – “Speech voltmeter” British type 3 (S.V.3) identical to the speech power meter of the ARAEN	2	230	100 (approx.)	equal to the integration time
(2) CCITT – V.U. meter (United States of America) C 16.5 – 1954 ASA (Note 1)	1.0 – 1.4	300	165 (approx.)	equal to the integration time
(3) CCITT – Speech power meter of the “SFERT” volume indicator	2	around 400 – 650	200	equal to the integration time
(4) CCITT – Peak indicator for programme transmission used by the BBC (B.B.C. Peak Programme Meter) (Note 2). Type IIa of IEC Publ. 268-10A	1		10	3 s for the pointer to fall 26 dB
(5) CCITT – Maximum amplitude indicator used by the Federal Republic of Germany (U 21), Type I of IEC Publ. 268-10	1	around 80	5 (approx.)	1.7 s from 100% to 10% of the reading in the steady state
(6a) OIRT Programme level meter (OIRT Recommendation No. 59)	1	less than 300	5	1.75 ± 0.25 s for the pointer to fall 20 dB
(6b) Level meter for international programme exchange over lines or links between OIRT and EBU (OIRT Recommendation No. 69)	1	less than 300	5 and 1000	2.5 to 4 s for the pointer to fall 20 dB
(7) EBU Standard peak programme meter (Note 5) EBU Tech. 3205 and Type IIb of IEC Publ. 268-10A	1	–	10	2.8 s for the pointer to fall 24 dB

Note 1. – In France, a meter has been standardized which is similar to that defined in item (2) of Table I.

Note 2. – In the Netherlands, a meter has been standardized (Type NOS-SN-411) which is similar to that defined in item (4) of Table I.

Note 3. – The number given in this column of the Table is the exponential n in the formula $V_{out} = (V_{in})^n$ applicable for each half-cycle.

Note 4. – The “integration time” was defined by the CCIF as the “minimum period during which a sinusoidal voltage should be applied to the instrument for the pointer to reach to within 0.2 N, or nearly 2 dB of the deflection which would be obtained if the voltage applied indefinitely”. A logarithmic ratio of 2 dB corresponds to 79.5% and a ratio of 0.2 N to 82%.

Note 5. – This meter is intended specifically for use in monitoring sound signals transmitted internationally, and therefore incorporates a scale conforming with CCITT Recommendation N.15, calibrated in dB from –12 to +12 relative to a level marked “TEST” corresponding to 0 dBm at a zero relative level point. In addition to the normal mode of operation having the characteristics shown above, the meter may be operated temporarily in a “slow” mode facilitating the comparison of observations made at widely separate points. The peak values indicated in this mode have no absolute significance, and may only be used for such comparisons.

ANNEX I

dBu0s AND dBrs TERMINOLOGY*

1. This Report specifies that transmission over lines or radio links should be in accordance with CCITT Recommendation J.14, considering the definitions given in Recommendation 645.

2. CCITT Recommendation J.14 uses terminology that is not always familiar in broadcasting practice. As a result transmission over lines and radio links has not always been satisfactory.

Report 820 details conditions to be observed by method A (vu meters) and method B (peak programme meters) users to ensure optimum performance in programme exchange.

Terms used in CCITT practice are introduced in Report 820.

3. This Annex describes the terms in broadcasters' terms.

These terms are:

3.1 dBu0s

3.1.1 The term dBu0s is used to describe test signal levels in sound programme circuits and sound systems. The test signals used to align sound programme circuits represent the sound signal levels, defined in Recommendation 645.

For broadcasting systems, the alignment signal should be defined in terms of a sine wave having a frequency of 1 kHz.

3.1.2 A level of 0 dBu0s corresponds to the nominal alignment level of the system concerned. It will be 9 dB below the permitted maximum level allowable at points of interconnection between broadcasting organizations.

In broadcasting practice, the alignment level generally corresponds to 35% of the maximum allowable modulation of a broadcast transmitter if no compression and/or pre-emphasis is used.

3.1.3 The measurement signal level of -12 dBu0s is used for long-term measurements and measurement at all frequencies.

3.2 dBrs

3.2.1 The term dBrs describes the designed voltage level at a point in a transmission path relative to that at a reference point of zero relative level (0 dBrs).

3.2.2 A 0 dBrs point is a point where the alignment signal of 0 dBu0s level has a voltage level of 0 dBu.

3.2.3 The dBrs value of a point is the gain or loss at that point with respect to a point of zero relative level (0 dBrs point). Zero gain or loss exists between the different 0 dBrs points throughout the transmission path.

Note 1. — The term dBu0s is a normalized test signal level. The dBrs value of the point at which a measurement is to be made is the normalizing factor which allows the absolute voltage level of the signal at any point in the transmission path to be determined.

Note 2. — From the descriptions of the dBu0s and dBrs above, it will be evident that:

- dBu (voltage level) = dBu0s (normalized test signal level) + dBrs (normalizing factor);
- the dBrs value is determined by circuit gain or loss;
- the addition sign must be taken to indicate algebraic addition.

Note 3. — It is necessary to distinguish clearly between the normalized test signal level as expressed in dBu0s, and the voltage level of the test signal at the measuring point in dBu. Thus, a meter will measure the voltage level of a signal in dBu (i.e. as a voltage ratio expressed in dB with respect to the reference voltage $U_0 = 0.775 \text{ V} = \sqrt{R \cdot P_0}$ with $P_0 = 1 \text{ mW}$ and $R = 600 \Omega$), but this information alone tells nothing about the relationship between the measured level and the level which one would expect to have at that point when a test signal (e.g. at alignment level) had been applied to some other point.

* CCITT Study Group XV has proposed that the suffix "s" be added to the terms dBu0 and dBrs when these refer to sound programme transmissions and not to telephone circuits. This text has adopted the proposed practice.

On the other hand, one requires a clear, unambiguous way of describing the level of a given test signal. The voltage level is not adequate, because for a given test signal the voltage level will vary with the amplification and losses which occur throughout a transmission system. The term dBu0s is such a means of describing a normalized test signal.

Thus, if one is applying the alignment signal level at a reference point, then this is 0 dBu0s signal level regardless of the voltage level in dBu. Furthermore, regardless of the variations of voltage level resulting from normal circuit gains or losses, this remains a 0 dBu0s test signal level even though its voltage level varies from point to point.

If one applied a measurement signal which is 12 dB below the alignment signal at that point, then this would be a -12 dBu0s signal regardless of its actual voltage level as read on a meter (in dBu).

Note 4. — The frequency of the alignment signal should be 1 kHz to avoid difficulties with equipment containing pre-emphasis, etc. 1 kHz is recommended in Recommendation 645 because this is used on most source test material (standard discs, tapes and film). For frequency response tests, a measurement signal 12 dB below the alignment signal should be used.

Note 5. — Definitions of "voltage level" and "transmission reference point" are given in § 1.2.2 and 2.4 of Appendix I to Recommendation 574.

Examples of the special notations used to indicate absolute voltage level in various cases are contained in § 8 of Annex I to the same Recommendation.

REPORT 1072-1*

**SUITABLE SOUND SYSTEMS TO ACCOMPANY HIGH-DEFINITION
AND ENHANCED TELEVISION SYSTEMS**

(Study Programme 47C/10)

(1986-1990)

1. The applications of a sound system to accompany high-definition television (HDTV) will range from the most complex such as multi-channel cinematographic productions (where it is desirable to strive for a higher degree of realism than is available in conventional cinematographic practices) to much simpler domestic uses where additional realism over conventional television may be desired but where furniture and space considerations will severely limit the number and placement of loudspeakers. Other intermediate levels of performance may also be desirable, subject to possible limitations on the number of audio channels available in reproduction by such factors as transmission system bandwidths, recorder capabilities and an interest in multilingual performance in some regions.

Recent studies [CCIR, 1982-86a, b, c, d; 1986-90a, b, c, d, e, f, g, h, i, j, k] have indicated that two-channel stereo is not sufficient for HDTV, but there is no agreement yet on the maximum number of audio channels which should be provided. However, a sound system suitable for HDTV applications should satisfy the following requirements:

- downward compatibility to more simple reproduction for less demanding situations;
- system compatibility for programme material originally prepared for single or multi-channel sound presentation;
- good localization of sound sources and correspondence with the visual image as far as possible for viewers and listeners in optimal and extreme seating positions;
- imaging capability of auditory ambience of the scene (e.g. reverberation, environment, and off-picture sound sources);
- reproduced sound ideally corresponding to the subjective quality given by a 16-bit linear digital encoding with a sampling frequency of 48 kHz;
- maximum economy in all respects, including transmission efficiency*.

* Bit-rate reduction techniques are described in Report 1199.

2. The CCIR has considered a number of documents on this subject.

2.1 In [CCIR, 1982-986a, 1986-90a] the EBU has pointed out that the introduction of HDTV certainly calls for the simultaneous introduction of a high-quality sound system, and that a two-channel system may not be adequate.

In [CCIR, 1986-90h] the EBU concludes that more than four sound reproduction channels are useful. Screen loudspeakers are necessary to stabilize the sound localization irrespective of the listening position; surround loudspeakers contribute to the enhancement of the sound realism. In an optimum configuration of sound system, the maximum number of reproduction channels is likely to be 8 (basic stereo: 2; screen: 2; surround: 4). A reduced number of channels configurations may nevertheless be possible, depending on the desired performance and programme requirements.

2.2 [CCIR, 1982-86b] describes a CBS triphonic sound system incorporating an audio matrix which provides for a third discrete audio channel, preferably for the transmission and reproduction of a centre-front dialogue programme. With a centre channel loudspeaker located in proximity with the video monitor, a stable source of centre sound is achieved in contrast to the highly unstable phantom centre sounds achieved by widely-spaced left and right loudspeakers. The triphonic system, which is applicable to both conventional and high-definition television services, provides the additional benefit of permitting non-critical placement of the two conventional stereo loudspeakers in a quasi-surround sound mode.

2.3 [CCIR, 1982-86c] reports on investigations carried out in Germany (Federal Republic of) on sound localization under the special circumstances of HDTV reproduction with large screens.

If more than one person watches the programme, they will sit side by side. This does not allow satisfactory perception of stable phantom sources generally used in stereophony. To convey picture-related localization and stability of perceived direction within a defined listening area, more than two screen related loudspeakers are necessary.

In the case of HDTV reproduction with large screens, the use of five screen-related loudspeakers is proposed. Such a system would make use of five separate sound channels which should be available for recording, transmission and reproduction. Compatibility with existing monophonic or stereophonic systems can be achieved by simple matrixing networks.

2.4 In [CCIR, 1986-90c and i] also from Germany (Federal Republic of) the required characteristics of an HDTV digital sound system - as summarized in § 1 - and results of some investigations are considered, leading to some conclusions regarding suitable loudspeaker arrangements, allocation of audio signals to individual loudspeakers, and number of transmission channels. With respect to the multi-language situation in Europe, the requirement of providing alternative languages is particularly considered.

2.4.1 In order to meet all requirements, and for reasons of economy, an HDTV sound system is proposed in [CCIR, 1986-90i] to ensure a flexible use of sound transmission channels.

Therefore, a studio-controlled allocation of audio signals to the reproducing channels is described, providing a final mix of dialogue, music and effects at the receiving end. The final mix is proposed to depend both on sound engineers' choice of sound format and intention as well as on listeners' choice of loudspeaker arrangement and language. A so-called "sound allocation control matrix" at the receiving end provides studio-controlled sound allocation to individual loudspeakers, depending on:

- an allocation control signal, which accompanies the multi-channel sound signal packet being transmitted;
- a loudspeaker arrangement status signal, which is generated at the receiving end;
- a language select signal, which is generated by the listener.

The proposed flexible use of transmission channels, in connection with the standardization of loudspeaker arrangements of different levels of performance, leads to particular advantages:

- the limited number of transmission channels can be used either for providing alternative languages and a reduced number of reproducing channels per sound presentation, or for providing a high performance multi-channel sound presentation without alternative languages;
- the sound engineers can use different recording/mixing techniques, depending on the production, type of programme, and number of languages to be provided, without technical compatibility problems. For example, emphasis can be put either on enhanced realism/effects or on stability of localization. Aesthetical aspects with respect to downward compatibility can be considered in the studio;
- technical developments of recording/mixing techniques, the result of which cannot be foreseen for the time being, will be able to be introduced without having to compromise or change the standard;
- if there is any restriction of transmitting channel capacity, the level of performance can be adjusted automatically in the continuity studio, by using the sound allocation control matrix.

Further studies are required for the development of an HDTV sound system, in particular the determination of:

- hierarchy of loudspeaker arrangements with respect to different applications and to handling aspects;
- the number of channels in the studio and for transmission;
- the conversion process for reducing the bit-rate in the case of transmission.

2.4.2 In [CCIR, 1986-90i] a hierarchy of loudspeaker arrangements is proposed. Both screen speakers and surround speakers complete the basic stereo speakers in order to stabilize localization or to enhance realism, depending on individually desired performance. Screen and surround category should be realizable with different levels of performance. Each level of performance should be specified by the number and location of the basic stereo loudspeakers, as follows:

basic stereo	1)	2 speakers, $\pm 30^\circ$
screen	1)	2 speakers, $\pm 10^\circ$
or	2)	1 speaker 0°
or	3)	0 speaker
surround	1)	4 speakers, $\pm 90^\circ$, $\pm 120^\circ$
or	2)	2 speakers, $\pm 90^\circ$
or	3)	1 speaker 180°
or	4)	0 speaker.

Free choice of any combination would ensure high adaptability to different types of programmes and individual desires. The maximum number of reproduction channels is 8 (basic stereo: 2; screen: 2; surround: 4).

The location of surround speakers is not critical because they are not used for stereophonic imaging rather than for the reproduction of reflections and diffuse sound or effect sound which is uncorrelated with the frontal sound.

As regards the number of transmission channels, there is no restriction due to the bit rate: assuming a high performance bit rate reduction system for each channel, the necessary bit rate for 10 channels will be about 1 Mbit/s, that is less than 1% of the bit rate consumed by the HDTV picture. Accordingly, 10 high-quality sound transmission channels should be provided in the HDTV system, in order to allow a high performance level of sound reproduction as well as multi-language services.

2.5 Several kinds of psychological experiments have been made in Japan in order to evaluate the quality of the stereophonic sound field and in order to study interaction effects between picture and sound images [CCIR, 1982-86d; CCIR, 1986-90b and d]. The results obtained from the experiments, along with some considerations described below, suggest the following desirable choices for the studio standard for the HDTV sound system.

2.5.1 It is desirable to make use of four sound signals in order to reproduce a better sensation of reality than is possible with stereo.

Reproduction of a sound field using two or three sound signals and with two or three loudspeakers located in front of the viewer gives an inferior sensation of reality. Judging from past experience, the situation cannot be improved with other three-loudspeaker arrangements, for example, loudspeaker arrangements in which a viewer faces the left and right loudspeakers with a wider listening angle e.g. 90° to 120° between the loudspeaker axes.

Although the use of more than four sound channels would provide a superior sense of reality, it would increase the cost and technical difficulties of the system. Consequently, a four-channel sound system would appear to be the most desirable choice for HDTV from the point of view of obtaining a better sensation of reality.

2.5.2 It is desirable to adopt loudspeaker arrangements which include three loudspeakers set in front of viewers, one under the centre of the screen and others to the left and right, in order to maintain a better coincidence between the direction of the sound image and that of the visual image on the screen for viewers on either side of the central perpendicular axis to the screen. At least two loudspeakers placed symmetrically at the rear using a pseudo-stereo system should be used to produce a greater sensation of realism.

The main trend of sound reproduction systems for movie theatres is a four sound-signal system with three signals fed to the front three loudspeakers and the remaining signal fed to several rear and side loudspeakers. HDTV sound systems should be compatible with movie sound reproduction systems.

2.5.3 As for the digital coding of sound signals for HDTV, in the studio, a sampling frequency of 48 kHz and linear coding with 16 bits/sample would be desirable for the international exchange of recorded programmes, taking into account Recommendation 646.

2.5.4 The sound system should be capable of being used as a one sound-signal system, a two sound-signal system, a three sound-signal system or a four sound-signal system. The number of sound signals would be chosen by considering necessary sound and picture effects and costs for production. The loudspeakers used in each case will be as follows:

one sound-signal system	centre loudspeaker
two sound-signal system	left and right loudspeakers
three sound-signal system	centre, left and right loudspeakers
four sound-signal system	centre, left, right and rear loudspeakers.

2.5.5 A sound transmission system using a baseband multiplex method, with its subjective evaluation results, is reported as a suitable sound transmission system for HDTV [CCIR, 1986-901]. On this sound transmission system, either two sound signals with the quality equivalent to that of a linear 16 bit - 20 kHz bandwidth signal or four sound signals with the quality slightly better than that of linear 14 bit - 15 kHz bandwidth signal, are transmitted within 1.35 Mbit.

2.5.6 [CCIR, 1986-90d] showed that the 3 front channels - 1 rear channel system allows broader range of viewing area and the sound quality in that range is better than that of the conventional stereophonic system at the centre viewing position and that the sound-visual difference of 15° was permissible for 75% of the non-expert viewers. Through the experiment on microphone arrangements, no particular difficulties were met in the recording of orchestra sound for the 3-1 system. Judging from these results, the Japanese Administration believes that the 3-1 system provides a desirable sound system for HDTV studio standard.

2.6 Experience in the United Kingdom with stereophonic sound for television [CCIR, 1986-90e] has lead to the conclusion that satisfactory stereo sound reproduction may be obtained using three loudspeakers in front of the viewer to reproduce the sound accompanying high definition television.

If more than one person watches the programme, the viewers may sit side-by-side. In any case, it is most unlikely that they will all sit on the axis of symmetry of the loudspeaker layout. To achieve reasonably acceptable sound image placement under these conditions using 2 loudspeakers, situated to the left and right of the television display, requires that each loudspeaker axis should point to the opposite end of the front boundary of the desired listening area. The loudspeakers should also have controlled and optimised directional characteristics [Kates, 1980] which it may not always be convenient to provide. To allow more flexibility in loudspeaker positioning, and to provide improved stereophonic sound source placement for viewers situated away from the axis of the loudspeaker layout, it is desirable to provide a third, centrally-placed loudspeaker.

The provision of 3 loudspeakers should be quite practicable. Ideally, all loudspeakers should reproduce the whole range of useful audio frequencies. However, it would be feasible, as a receiver option, to have a relatively large central loudspeaker enclosure incorporating one or more bass drivers in a stand or console for the television display. The left and right loudspeakers, which would normally be 2-3 metres apart, and separated from the television display, could be relatively small, reproducing frequencies from about 150 Hz upwards. Where such an arrangement is used, it is likely that compensating bass lift would need to be applied to the central signal.

The balancing of reproduced sound levels from the 3 loudspeakers should not present too severe a problem for the viewer. However, HDTV systems requiring the use of more loudspeakers could make balancing more of a problem in the viewing environment. If the placing of some loudspeakers behind the viewer is necessary, there could, additionally, be problems setting up the correct signal timing at the viewer's location.

2.7 [CCIR, 1986-90i] from Italy also discusses the number of independent sound signals to be provided for the production, delivery and presentation of HDTV programmes. Building on a parallel with the well established and fully satisfactory experience in the cinema, the view is expressed that:

- the number of independent sound signals to be produced is at most equal to four;
- those four independent sound signals can probably be multiplexed on just two sound transmission channels;
- consumer HDTV displays will be offered on the market with a number of loudspeaker positions that will vary depending on the display model and the consumer taste.

2.8 In [CCIR, 1986-90j] Australia has pointed out that a satisfactory balance between dialogue and accompanying music and effects is normally achieved in radio and television sound broadcasts for people with normal hearing. Of necessity, the hearing-aids used by the hearing-impaired incorporate compression to reduce the dynamic range to permissible limits. The compression can degrade the balance between the dialogue and the music leading to a serious reduction in intelligibility for the hearing-impaired. A possible solution to this problem is a special sound track for the hearing-impaired to provide a suitable relationship between the dialogue and music.

Furthermore, Australia has pointed out in [CCIR, 1986-90k] that countries using 50 Hz television systems replay film programmes that have been recorded at 24 frames per second, at a rate of 25 frames per second. The resultant 4.2% change of pitch causes a significant change in the emotional quality in spoken words and is an unacceptable artifact for HDTV sound where the highest quality is required. It is proposed that administrations be invited to study techniques for correcting or avoiding pitch change during film to television transfers.

3. One of the multi-channel sound systems under consideration in the United States [CCIR, 1986-90h] proposes to digitally transmit two discrete channels which contain four channels of directional information in matrix form (4-2-4). The four channels are designated as Left, Centre, Right, and Surround. Prior to matrix encoding, the Surround channel is filtered to a bandwidth of 100 Hz to 7 kHz and dynamically preemphasized with a modified Dolby B encoding circuit. This circuit provides exactly one-half of the normal Dolby B encoding boost. The Left, Centre, and Right channels are filtered with all-pass networks so that they remain in proper phase relationship with the Surround channel. The four channels are encoded into a pair of channels denoted Lt and Rt. The encoding equations are:

$$Lt = \text{Left} + 0.707 \text{ Centre} - 0.707 j \text{ Surround}$$

$$Rt = \text{Right} + 0.707 \text{ Centre} + 0.707 j \text{ Surround.}$$

The encoded signal is compatible with the following reproduction systems:

Monophonic

The Surround channel is not reproduced. The L, C and R signals are reproduced as in the case of a conventional stereo signal summed to mono.

Stereo

The L, C and R signals are reproduced as in normal stereo. The Surround signal is reproduced out of phase and is not well localized. The Centre channel is a phantom image.

Stereo with Surround

The L, C and R signals are reproduced as in normal stereo, with the C signal being a phantom image. The Surround signal is reproduced out of the rear channel speakers.

Stereo with Centre

The L, C and R channels are reproduced out of separate speakers. Dynamic decoding circuitry in the decoder can provide a hard Centre channel for precise dialogue localization, while maintaining good left-right separation for music and effects. The Surround information is reproduced by the Left and Right speakers out-of-phase.

Stereo with Centre and Surround

All four channels, L, C, R and S, are reproduced from their own loudspeaker channels. Dynamic decoding circuitry can improve localization for all channels. An audio delay line in the Surround decoder additionally prevents Centre channel dialogue from appearing to emanate from the rear Surround channel. This mode is identical to that used for the reproduction of stereo-optical 35 mm film soundtracks.

It has been further proposed that the 4-2-4 matrix signals be carried by a two-channel digital system using ADM sound coding with a total data rate of 512 kbit/s. The data would be transmitted by adding an additional QPSK-modulated sound carrier 4.85 MHz above the video carrier. The results of subject assessments of this ADM coding are in Annex VII of Report 953.

4. Taking into account these different contributions together with other results of new work, IWP 10/12 is expected to prepare a draft Recommendation before the next interim meeting.

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REPORT 798-2

SIMULATED PROGRAMME SIGNALS

(Question 44/10, Study Programme 44A/10)

(1978-1982-1990)

1. Present situation

Noise signals simulating programme signals are specified in Recommendation 559, Report 497, and Recommendation 571. It is also known that this subject has been studied by the IEC and a standard published [IEC, 1985].

It has been proposed that there should be only one such signal but before standardization can be achieved it is necessary to study the differences between the signals as specified at present, and the reasons for these differences.

2. Characteristics of the signals

The signals proposed at present differ significantly in the high-frequency regions of their spectra. The signal in the IEC standard has the least high-frequency energy, while the CMTT signal has the greatest amount.

Table I shows the main differences between the noise signals in terms of their purpose and their relationship with real programme signals.

TABLE I

Characteristic	CCIR		IEC
	Recommendations 559 & 641	Recommendation 571 Report 497	IEC 268-1, Second Edition
Type of signal which is simulated by the noise spectrum	Modern dance music (1967)	Average of all types (1976?)	Average of all types (1975-77)
Derivation of spectrum from measurements on actual programme signals	In each 1/3 octave band, the level which was exceeded for between 20% and 40% of the time was measured	The spectrum is typical of signals whose levels lie between those exceeded for 1% and 10% of the time, over a period of 1 min.	The spectrum is typical of signals whose levels are equal to that exceeded for 50% of the time
Purpose	Measurement of selectivity of AM and FM receivers for frequency planning and determination of protection ratios	Measurement of cross-talk and cross-intermodulation in circuits and multiplex systems, excluding measurements of non-linearity within a channel	Specifying the performance of amplifiers and loud-speakers. Measurement of non-linearity within a channel

3. Conclusions

It should be noted [CCIR, 1986-90] in particular that the noise signals specified in Recommendations 559 and 641 are used to determine radio-frequency protection ratios. The effect of a change in the characteristics of the noise signals used is discussed in [CCIR, 1978-82] and in Report 491.

It is not possible at present, considering the differences shown in Table I, to recommend the adoption of any one of the signals for *all* purposes. It is necessary for further studies to be carried out (see Opinion 59).

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REPORT 1068

DIGITAL AUDIO CODING STANDARDS

(Question 51/10, Study Programme 51B/10)

(1986)

The coding of sound signals has been defined in various Recommendations e.g. for broadcasting studios (Recommendation 646), for recording (Recommendation 648) and for emissions in satellite broadcasting (Recommendation 651).

These Recommendations make use of the following three methods of coding:

- Method 1: uniform quantization (linear mode) with at least 16 bits/sample;
- Method 2: uniform quantization (linear mode) 14 bits/sample;
- Method 3: near-instantaneous companding 14/10 bits/sample.

While in a studio environment a resolution of "at least 16 bits/sample" is demanded, methods 2 and 3 have been proposed for satellite services. Their common feature is to start with samples in two's complement representation uniformly quantized with a resolution of 14 bits. Until recently, it was generally agreed that the remaining quantization noise was acceptably low and caused no impairment of transmission quality in a normal listening environment. An additional increase of resolution would result in higher costs which would not be acceptable in a domestic environment.

Recently, the situation has changed substantially with the introduction of the compact disc; it was possible to demonstrate that a resolution of 16 bits represents neither a technical nor economic problem for consumer applications. This higher resolution is demanded for many reasons, among which are the following:

- broadcasters will have to take into account the growing preference for improved quality by listeners who will compare the quality of transmission with that achieved by the compact disc. A resolution of 16 bits has been justified practically by the BBC, among others, as a resolution lower than 16 bits will cause audible quantization noise when listening to critical signals in an ideal listening environment;
- as previously mentioned digital audio signals coming from television and radio studios will be available in 16 bits resolution, and it can be asked why this high quality cannot be transmitted to the listeners;
- an increase of resolution is also of interest to the broadcasters themselves as the extended dynamic range provides a higher overload reserve. This avoids the necessity of using limiters, which are necessary for a coding technique based on a 14 bits resolution.

Very often the transmission capacity does not allow the transmission of code words with a length of 16 bits, in which case a bit rate reduction by means of coding methods 2 or 3 is necessary. Nevertheless, the basic resolution of these two coding methods can easily be increased to 16 bits by the subsequent use of a scale factor like that already used in the MAC/packet systems. The scale factor shows how many of the MSBs following the sign bit have the same values as the sign bit itself (Fig. 1). The value of the MSBs is already determined by the scale factor and the value of the sign bit. Therefore, these MSBs do not have to be transmitted. Consequently, at the transmitter they are suppressed and the remaining bits (hatched region in Fig. 1) are shifted directly behind the sign bit. At the receiving end the samples are reconstructed perfectly by shifting back the transmitted bits, and the MSBs (which have not been transmitted) are inserted with the value of the sign bit according to the instruction of the scale factor. This procedure, which has been termed "floating point technique", closely resembles the 14/10 bits near-instantaneous companding procedure. In fact, the only significant difference is that 14/10 bits near-instantaneous companding shifts only at scale factors 0 to 4. By shifting at every available scale factor (0 to 7) it is possible to transmit at the appropriate low signal levels, also the 15th and 14th bit of a sample (Fig. 1). It is suggested that this coding procedure could be used not only for the transmission of 10 bits (16/10 bit companding) but also for 14 bits (16/14 companding) per sample.

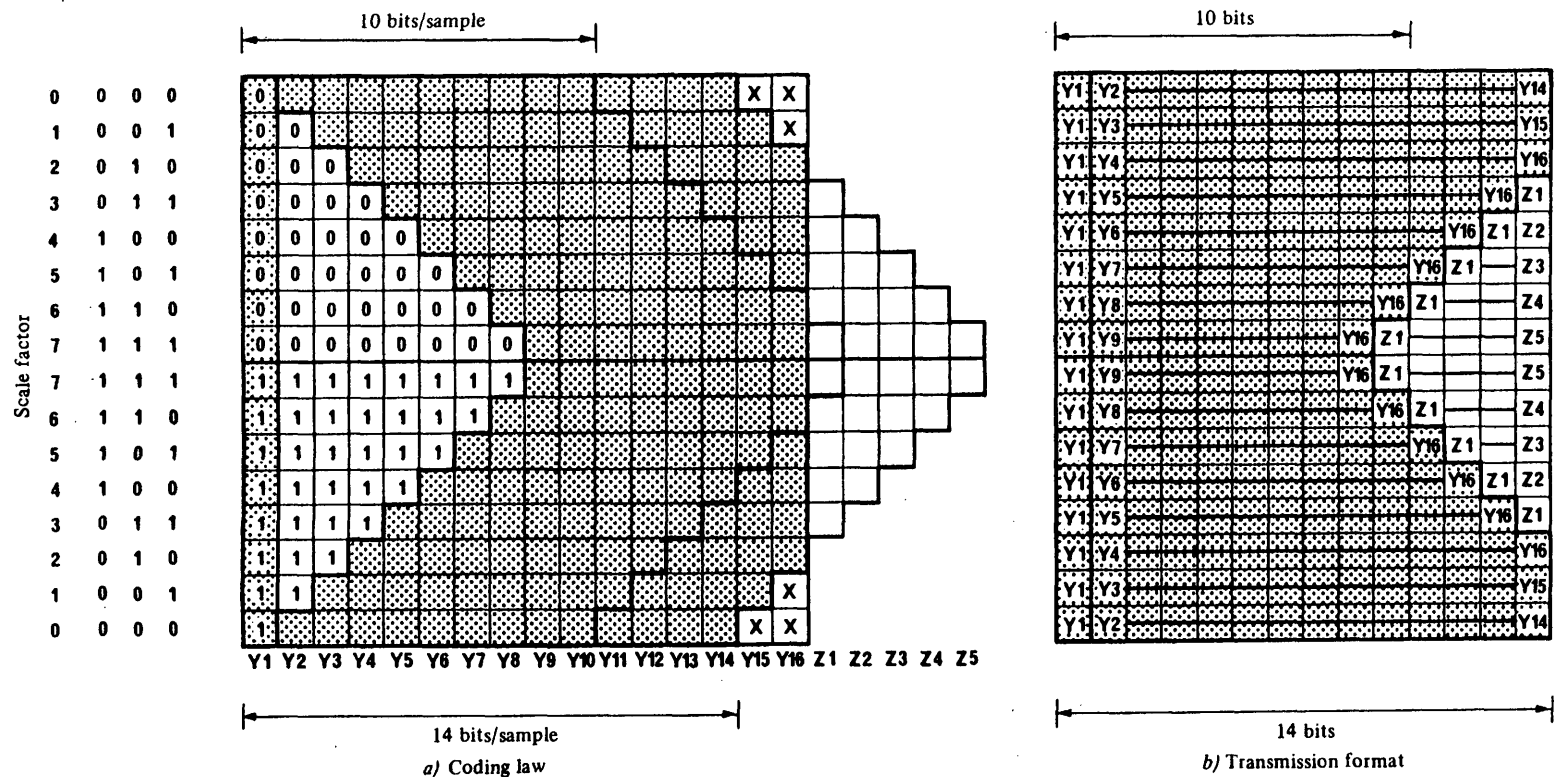


FIGURE 1 – Floating point technique

- Non-redundant range of the 16 bits source signal words
- X X Non-transmissible bits of the 16 bits source signal words

This is an important extension of the near-instantaneous companding technique which permits a significant improvement in quality to be made for the listener, and enables the broadcaster to transmit the signal with the same bit rate as for 14/10 companded signals, without the need for audio limiters.

[CCIR, 1982-86a] expressed the opinion that a new family of coding standards based upon the following 16 bits/sample initial coding, could be of benefit to multi-standard receivers:

- uniform quantization (linear mode) 16 bits/sample (see Note 1);
- near-instantaneous companding mode 16/14 bits/sample (see Notes 1 and 2);
- near-instantaneous companding mode 16/10 bits/sample (with pre-emphasis).

Note 1. – The use of pre-emphasis is still under consideration.

Note 2. – During the Final Meeting of the CCIR, 1985, this mode was included in Recommendation 651.

In evaluating this proposal, it should be noted that all the features of the linear mode and of the 14/10 bits companded mode will be available when the 16 bits companded modes are used. Broadcasters will not be compelled to use this improvement as there remains the possibility of transmitting sound signals with 14 bits with linear quantization or with 14 bits resolution in a five-scale near-instantaneous coding format.

It is important to note that this proposal:

- does not require additional data capacity; i.e. no change in multiplexing and organization of data packets is necessary;
- does not result in additional costs of the receiver hardware.

Therefore, the improvement obtained by changing to 16 bits/sample coding will not lead to additional expense. If initially this improvement is not used, and 14 bits/sample coding is adopted, the possibility of improving the sound by changing subsequently to 16 bits/sample will still exist.

Because of the format chosen, a later change would not jeopardize compatibility with already-existing receivers.

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REPORT 1199

LOW BIT-RATE DIGITAL AUDIO CODING SYSTEMS

(Study Programme 51B/10)

(1990)

1. Introduction

Different techniques are under investigation for reducing the bit rate of digital audio signals to values less than 192 kbit/s per channel without causing perceptible impairment. Beside the requirements of a low audio bit rate, which is necessary for efficient utilization of the radio spectrum, the use of low bit rate audio coding has certain implications, for example the introduction of delay, the reduction of processing margin, which can be a problem for the broadcasters and for production studios. In the text that follows, a list of implications is shown, which are of importance especially for broadcasting, but nevertheless for most applications:

- a) encoding of the full 20 kHz audio band will be necessary for some applications;
- b) in applications requiring more than one audio channel, the coding of unrelated audio signals must be possible;
- c) resistance to bit errors is an important aspect since robust coding schemes will require less redundancy for protection in transmission channels;
- d) it must be possible to access small blocks of data in the encoded signal for the purpose of applying error-concealment and to provide adequate time resolution for random-access and editing of the signal in encoded form;
- e) some capacity must be available for ancillary data;
- f) excessive delay can be a problem in some areas of application and must be kept to a minimum for such applications;
- g) the decoder should be as simple as possible to lend itself to an easy VLSI implementation;
- h) it could be advantageous for the decoder to work with different bit rates as selected by the encoder;

* This Report should be brought to the attention of Study Group 11 and the CMTT.

- i) for studio applications and contribution links, sufficient coding margin must be provided in the audio coding to allow for post-processing and subsequent coding and decoding.

The relative importance of each of these requirements in relation to different possible applications are assessed in reference [CCIR, 1986-90a].

2. Low bit rate coding methods

The principle coding methods studied to date involve:

- predictive coding;
- sub-band coding;
- transform coding.

2.1 Predictive coding

This type of coding harnesses short-term linear prediction techniques, whereby the difference between the real signal and an estimated value is transmitted. By using efficient predictors, the differences and hence the bit rate can be reduced. The bit rates obtained are between 192 kbit/s and 224 kbit/s, i.e. a reduction by a ratio of 2 with respect to coding in accordance with Recommendation 660.

Examples of predictive coding systems are given in the references [CCIR, 1986-90b, c].

2.2 Sub-band coding

With this type of coding, the initial band is split up into a number of sub-bands by means of a filtering network such as, for example, a tree network with quadrature mirror filters (QMF) or a polyphase network with pseudo-QMF. The signal in each sub-band is coded after sub-sampling by means of quantizers and/or adaptive predictors specific to each spectral zone. For high-quality sound channels (20 kHz bandwidth, 48 kHz sampling rate), configurations of 4 and up to 32 sub-bands have been used to obtain bit rates of the order of 128 kbit/s or lower, i.e. a reduction by a ratio of nearly 6.

The references [CCIR, 1986-90d, e, f, g], [Stoll, et al., 1988; Theile, et al., 1988; Dehery, et al., 1988 and 1989], contain details of sub-band coding systems. A more detailed description of these techniques is given in Annex I to this Report. Systems which combine sub-band coding with predictive coding are described in references [CCIR, 1986-90h, i].

2.3 Transform coding

The principle of this type of coding consists in converting a block of consecutive samples (e.g. 512) to the spectral domain (e.g. Fourier transform or cosine transform). The spectrum components are then quantized in order of size, allowing for the properties of frequency masking and aural sensitivity. A dynamic allocation of the bits required for the coding of each component makes it possible to optimize the bit rate at values of the order of 128 kbit/s or lower for high quality sound, representing a reduction by a ratio of nearly 6.

The references [Krahé, 1986; Brandenburg, 1988; Johnston, 1988] contain detailed descriptions of transform coding systems.

3. Quality assessment

Low bit rate coding systems do not always preserve the waveform or the spectrum of the original signal and may have characteristics which are not readily measured by conventional objective test methods. They may also produce artifacts which are not easily identified by the normal range of listening material. Critical test signals that would allow the identification of the following known artifacts should be sought:

- high slew signals to identify slew rate distortion;
- percussive sounds to identify pre-echo phenomena;
- sustained sounds to identify blocking effects and unwanted amplitude modulation;
- broadband sound signals to identify bandwidth limitation and colouration;
- stereophonic signals to identify changes in the stereophonic image.

It is therefore necessary to devise new measurement methods, and to find suitable listening material for use in subjective tests. To this effect, CCIR Recommendation 562 contains references to some suitable programme material for subjective assessment.

4. Situations to be tested

- single low bit rate codecs;
- cascaded low bit rate codecs;
- low bit rate codecs used in conjunction with other forms of processing (e.g. digital filtering, mixing, gain adjustment, transcoding to or from other systems, editing);
- low bit rate codecs operating under conditions where errors may be experienced.

5. Quality in different areas of application

Applications of low bit rate coding which can be envisaged are the conveying of high-quality audio and medium-quality audio via media providing limited bit rates or in which economies may be made by reducing the bit rate (e.g. point-to-point links and recording). The terms "high-quality" and "medium quality" are not yet precisely defined. "High-quality" is normally taken to mean audio with bandwidth restricted to not less than the frequency range 40 Hz - 15 kHz, and "medium-quality" normally refers to audio restricted to not less than 50 Hz - 7 kHz. High-quality audio is required in the studio, for programme contributions, for distribution links as well as for satellite and terrestrial digital broadcasting. Medium-quality may be adequate for lower-quality services (such as LF, MF and HF transmitters) and for voice-quality coordination signals.

The audio quality depends upon the system characteristics, the transmitted bit rate and delay. It will be important in future to define more precisely what is meant by "high-quality" and "medium-quality".

6. Delay

Low bit rate coding systems can introduce significant delay. The more efficient coding systems with the lowest bit rates tend to introduce the greater delays, and may, in consequence, cause operational difficulties in some circumstances (in television applications as well as radio). Report 1200 gives further information on some of the consequences of delay.

7. Connections via digital networks

There will be the need for signals coded by broadcasters to be exchanged, in digital form, via PTT or common carrier connections. Because of the operational consequences of using low bit rate coding (the effect upon quality and the introduction of delay), broadcasters should therefore be responsible for selecting the coding appropriate for their purposes.

8. Further studies

Administrations are invited to contribute to the following areas under Study Programme 51B/10:

- description of coding algorithms;
- definition of quality levels;
- subjective assessment methods;
- objective measurement techniques;
- results of evaluation tests.

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ANNEX I

DESCRIPTION OF A HIGH QUALITY SUB-BAND CODING TECHNIQUE USING PSYCHOACOUSTICAL MASKING

1. Introduction

A sub-band coding system family has been developed in the Federal Republic of Germany and France [CCIR 1986-90a, b] for different applications and requirements in the field of broadcasting [CCIR 1986-90c]. Prototype equipment (COFDM modem/MASCAM codec) was demonstrated at WARC-ORB(88) in Geneva in September 1988 (see [CCIR, 1986-1990d]). Recent development has succeeded in further reduction of bit rate, complexity and delay, and in the specification of a family of encoders by using one universal decoder which can easily be implemented into VLSI. This new coding family is called "MUSICAM" (Masking-pattern Universal Sub-band Integrated Coding And Multiplexing).

2. Basic principles of the sub-band coding system

2.1 Encoder

A polyphase filter network is used for dividing the broadband audio signal into 32 sub-band signals with a constant bandwidth of 750 Hz with 48 kHz sampling frequency. These sub-band signals are divided into digital frames of 12 successive audio samples with a duration of 8 ms. Once in each frame, a scale-factor corresponding to the maximum level attained by each sub-band signal is derived, and those values of the scale-factors are quantized. The bit rate of the scale-factors is reduced in particular by transmitting only one scale-factor for more than one adjacent sub-band in the higher frequency region.

In parallel to the filtering of the audio signal, a Fast Fourier Analysis is done in the frequency region up to 8 kHz. However, only the samples of the sub-bands are coded and transmitted. This analysis is necessary only in the encoder for the calculation of the dynamic bit allocation to the 32 sub-bands. By using such a transform, a high frequency resolution is provided. In combination with the scale-factors, which allow a high time resolution, the

human ear's masking thresholds can be estimated, both, in time and frequency, with a very high accuracy.

For each of the 32 sub-bands, the minimum of the masking threshold, which allows the maximum level of inaudible quantizing noise, is calculated. The necessary resolution of quantizing the sub-band samples in an individual sub-band is derived directly from the difference between the maximum level and the minimum masking threshold. No transmitting capacity is necessary for those sub-bands, which are carrying irrelevant parts of the signal (i.e. sub-bands signals which are completely masked).

For transmitting the coded signal, the scale-factors and the bit allocation information are coded and integrated into the multiplex signal as side-information beside the sub-band samples. This information is necessary in the decoder for decoding the audio signal.

2.2 Decoder

First of all, the side-information, i.e. the scale-factors and the bit allocation information, and 12 successive samples of each sub-band signal are separated from the multiplex signal. The reconstruction process is characterized by filling up the data format of the sub-band samples to 16 bit linear regarding the scale-factor and allocation control data for each sub-band and frame. By a filter network, which is completely inverse to the network used in the encoder, the complete audio signal with a bandwidth of up to 24 kHz can be reconstructed.

3. Technical features

Sampling frequency

For direct transcoding of different digital audio standard signals (EBU/AES-interface, consumer/professional), three different sampling frequencies, 32, 44.1 or 48 kHz can be chosen in the encoder.

Bit rate and quality

The bit rate per monophonic channel can be chosen in the encoder to be:

64 kbit/s: The subjective quality comes close to that of the CD for most of the signals. This low bit rate is very useful for high quality reporting links.

96 or 128 kbit/s: The subjective quality is identical to that of the CD for any critical signal. These bit rates are important for most applications and for broadcasting especially for radio sound-programme emission, television sound emission, distribution and consumer storage.

192 kbit/s: This bit rate is intended for any critical signal, allowing some "headroom" for post-processing such as equalizing, editing, cutting and cascading (multiple coding and decoding). This bit rate may be suitable for contribution links, production and professional storage.

Overall delay

The minimum delay using a polyphase filter network without additional buffering is about 24 ms for encoding and 6 ms for decoding, which will be obtained by future VLSI. The resulting short overall delay is important for some broadcasting applications, in particular for monitoring during the production process, synchronizing with video and bidirectional commentary links.

An increased overall delay, e.g. up to 80 ms, allows an additional buffering of the coded sub-band samples for reduction of the peak bit rate values. This is particularly useful when encoding at 64 kbit/s.

Error ruggedness

The encoded signal includes a protection of a certain number of important bits, such as side-information and most-significant bits in sub-bands at low frequencies. This protection, in addition to the inherent insensitivity of the audio samples to bit errors, provides a high degree of error ruggedness. At bit-error ratios of 10^{-3} (gaussian error distribution), the subjective quality is negligibly impaired.

Complexity

The decoding process needs significantly less computation power than the encoding process. Using either, a Motorola DSP 56001 24-bit fixed point, or an AT&T DSP 32C 32-bit floating point signal processor, the decoding process of a stereo decoder can be implemented into only one processor. As the computation power required by the decoder is mainly determined by the inverse filter network, the algorithm can easily be implemented in a custom VLSI circuit.

The complexity of the prototype stereo encoder can be reduced to one signal processor, either one DSP 56001 or one AT&T DSP 32C by a simplified calculation of the masking threshold without using the FFT. In this case, a higher bit rate is necessary for the same quality level but with the advantage of a shorter overall delay.

Coding family compatibility

For reasons of flexibility and low cost, one universal decoder integrated circuit can be used for all different applications. This means that the filter network, which is the same in the encoder and in the decoder, is optimized with respect to all required technical features.

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REPORT 1069*

**DIGITAL AUDIO INTERFACE FOR THE INTERCONNECTION OF
DIGITAL BROADCASTING STUDIOS WITH DIGITAL TRANSMISSION NETWORKS**

(Study Programme 51B/10)

(1986)

1. When sound programmes originating in digital broadcasting studios are to be transmitted to an earth station of a direct broadcasting satellite or to a remote studio, it will in most cases be advantageous to use a purely long-distance digital transmission circuit.

A specific interface different from that described in Recommendation 647 which will provide the error correction scheme most suited to the transmission link, may be needed to facilitate the interconnection between the broadcasting and long-distance equipment. This will also enable some necessary operations, such as auxiliary data management to be made by the broadcaster.

2. [CCIR, 1982-86a and b] describe an interface format which provides for the transmission of a digital stereophonic audio programme with a bit rate of 1024 kbit/s via the digital telephony network as defined in Recommendation G.732 of the CCITT (2048 kbit/s digital signals).

In accordance with Recommendation 651 on digital coding for the emission of high quality sound signals in satellite broadcasting (15 kHz nominal bandwidth) a near-instantaneous 16/14 bits floating point technique and a sampling frequency of 32 kHz are used for audio signal coding; in this case there is no pre-emphasis.

An additional bit rate of 32 kbit/s is provided for the transmission of data accompanying the programme signal.

Fully synchronous operation is envisaged as defined in Recommendation G.811 of the CCITT. Synchronizing signals may be derived from the first multiplexer of the digital network, which combines either two 1024 kbit/s signals or one 1024 kbit/s signal together with digital 64 kbit/s signals to form one 2048 kbit/s bit stream for transmission.

3. [CCIR, 1982-86c] describes a flexible multiplexing method that has been developed and is proposed for a new digital programme network that is now under study in Sweden.

A prototype system that can be interfaced to any transmission channel with bit rates up to 4000 kbit/s has been built and successfully tested. Interfaces to 3000 kbit/s and 2048 kbit/s have also been implemented. The latter is compatible with CCITT Recommendation G.732, and transmissions have been carried out over a 1500 km circuit including 140 Mbit/s links.

* This Report should be brought to the attention of the CMTT.

Any type of codec can be connected to the system, provided that the tributary bit-stream is an integer multiple of 100 bit/s. The present equipment includes linear 16 or 14 bits PCM codecs operating at 48, 32 or 16 kHz sampling frequency, as well as synchronous data interfaces operating at user-selectable, standard data speeds ranging from 110 bit/s to around 200 kbit/s. It is prepared for companded 14/10 bits codecs and asynchronous data interfaces.

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CCIR Documents

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REPORT 1070*

QUALITY MEASUREMENT METHODS FOR DIGITAL SOUND

(Question 51/10, Study Programme 51B/10)

(1986)

This Report is intended to collect information on measurement methods suitable for digital audio equipment.

1. Methods for measuring distortion caused by bit errors

A bit error in a data-stream causes an interference pulse, and for linear (uniform) coding the magnitude depends only on where it occurs in the code group. However, when companding is used, the audio signal itself has an effect on the magnitude, so in this case a method of measuring distortion in the presence of programme material is needed.

A method of this kind is applicable to both companded and linear systems; the major benefit in this latter case is that it allows measurements during the course of normal programme operations.

One method which is described in [CCIR, 1978-82a] suggests measuring the energy in a window above an audio band limit. Here the energy would be almost solely due to the interference pulses. For more complete details, see CMTT Report 968.

* This Report should be brought to the attention of the CMTT.

2. Measuring method of distortion noise

To evaluate the quality of digital audio equipment and transmission systems for broadcast use, it may be necessary not only to measure quantizing noise but also other kinds of distortion components [CCIR, 1982-86a]. For this purpose the following methods have been developed:

- a test signal, in which a certain frequency band is eliminated, is applied to the equipment of systems under test and the noise in the eliminated band of the output signal is measured [Nikaido, 1972; CCIR, 1982-86a];
- a method which is intended specifically for the detection and measurement of discontinuities in the transfer characteristic of digital audio equipment. In this method, a low-frequency sinusoidal tone is applied to the equipment under test, and the quantizing noise and non-linear distortion products in the band above the test signal are converted into digital form by an analogue-to-digital converter (ADC) and measured by a computer. At the same time, the test signal itself is taken from the output of the equipment, re-converted into digital form and applied to a second input of the computer. The computer is programmed to act as a sensitive peak noise meter whilst at the same time recording the instantaneous voltage level of the test signal. It is thus able to measure the normal quantizing noise with the brief bursts of noise resulting from discontinuities, and also indicate, approximately, the positions of discontinuities in the transfer characteristic [CCIR, 1982-86b].

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REPORT 799-2*

SUBJECTIVE ASSESSMENT OF QUALITY OF SOUND IN BROADCASTING USING DIGITAL TECHNIQUES

(Question 51/10, Study Programme 51A/10)

(1978-1982-1986)

1. Decision 18 co-ordinates the studies undertaken by the CCIR in the field of digital systems. It has allocated responsibility for the following studies to Study Group 10:

- study of encoding methods (including reduction of redundancy), transmission over various transmission media (including methods of monitoring), measurement of operational quality and test signals; these studies should apply in the field of sound studio equipment (including recording) and sound broadcasting from terrestrial transmitters and satellites;

* This Report should be brought to the attention of Interim Working Party CMTT/1.

- study of methods for protection against impairments associated with digital transmission in the field of sound studio techniques;
- study of the subjective quality of encoded and digitally transmitted signals for sound broadcasting;
- study of the problems of switching, mixing and special effects in sound studios.

2. During the Final Meeting, Geneva, 1977, the following documents were considered: [CCIR, 1974-78a, b, c, d, e and f].

These documents were studied by the CMTT but they did not provide a sufficient basis to formulate solutions to the problems given to Study Group 10 and listed above and also in Study Programme 51A/10. They have been used for information only.

Study Programme 51A/10, the aim of which is subjective assessment of the quality in digital sound broadcasting systems, should, in particular, provide an answer to the question raised by the CMTT, with regard to whether a subjective impairment of grade 4.5 is appropriate for a reference circuit including digital sections (Decision 18, Annex I; Report 647; [CCIR, 1974-78e]). The lack of detailed studies concerning the addition of the specific impairments associated with the digital systems does not, at the present time, make it possible to define a precise limit for the subjective impairment acceptable to the reference circuit.

Further studies are therefore necessary. They would allow definition of quality characteristics for intermediate or short distance transmission circuits (such as between production centres).

3. [CCIR, 1978-82a and b] report listening tests which have been carried out to determine the upper cut-off frequency of the audio signal. To reduce the bit rate, this frequency should be as low as possible without deteriorating the quality of the audio signal.

The result of these tests shows that, for the listener, a bandwidth of 15 kHz is sufficient and an extension of the frequency range up to 20 kHz gives an insignificant improvement in quality even taking into account future developments in sound systems because the result is related to the hearing ability of the listener.

4. For digital audio signals transmitted via satellite, the EBU proposes to use either a 14/10 bits near-instantaneous compandor or a 14 bits system using linear quantizing. In connection with the compandor, the use of emphasis is believed to be essential and the CCITT Recommendation J.17 pre-emphasis is proposed. [CCIR, 1982-86a] reports on listening tests which have been carried out to examine the value of emphasis for linear coding, where slightly higher quality criteria apply.

Two different emphasis circuits were studied:

- CCITT Recommendation J.17 pre-emphasis (additional attenuation of 5 dB);
- pre-emphasis characterized by the two time constants 50 μ s and 15 μ s.

The results of these subjective listening tests show that the 50/15 μ s de-emphasis leads to a noise reduction of 1 to 2 dB. The CCITT de-emphasis leads to an increase of impairment by noise of about 1 to 1.5 dB.

Because of the need to avoid exceeding the clipping point of a digital audio system, it is necessary to expand the headroom when using pre-emphasis. This expansion of the headroom can only be obtained by attenuation of the maximum programme level. This attenuation is equivalent to a decrease of S/N ratio.

Compared with this loss, the gain in S/N ratio by the use of de-emphasis is very small (50/15 μ s de-emphasis) or even can be an additional loss (CCITT de-emphasis).

Considering carefully the balance of evidence, it seems to the EBU that there is no advantage in introducing emphasis in a digital 14 bits (or more) system using linear quantizing for the purpose of reducing the audibility of quantizing noise. Nevertheless, the use of pre-emphasis in linear coding may be advantageous for other purposes (e.g. to reduce the impairment caused by errors or processes involving requantization).

5. A method to assess the perceptibility of group-delay distortion in low-pass filters is described in [CCIR, 1982-86b].

The group-delay distortion produced by a variable number of 11th order Chebyshev filters are reproduced by means of a digital processor running a fast Fourier transform program. According to these studies, the distortions caused by at least three 15 kHz low-pass filters in tandem, without group-delay correction, are inaudible with a specially selected test signal.



6. A number of contributions [CCIR, 1982-86c, d, e, f, g] have been received on the relative merits of the CCITT Recommendation J.17 pre-emphasis and of the 50 and 15 μ s pre-emphasis. Particular consideration is given in these contributions to the modification of peak programme levels resulting from their introduction as well as to their effect on the perceptibility of the modulation noise associated with companded systems.

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 [1978-82]: a. 10/13 (EBU); b. 10/49 (Germany (Federal Republic of)).
 [1982-86]: a. 10/21 (EBU); b. 10/212 (Japan); c. 10/269 (Germany (Federal Republic of)); d. 10/270 (France); e. 10-11S/139 (Japan); f. 10-11S/201 (EBU); g. 10-11S/205 (Japan).

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REPORT 1200*

THE EFFECT OF DELAY IN SOUND-PROGRAMME OPERATIONS

(Study Programme 51B/10)

(1990)

1. Introduction

With the increasing use of digital sound transmission equipment, delay effects are becoming significant on the circuits linking sound programme originating points and studio - centres as well as on circuits linking studio - centres and sound broadcast transmitters; the use of satellite links introduces even more serious problems. It is clear that the subject requires particular study and the following text presents a few examples of how delay effects can arise and the problems that ensue.

2. The problem of time differences between commentary and cue-programme signals in sound broadcasting.

Where a person presenting a sound broadcast programme has to receive cue programme via headphones, it is often unavoidable that the presenter's own speech signal is present in the cue-programme feed. Where the circuits carrying the broadcast and/or cue programme introduce significant delay, the resulting time-difference between the presenter's own speech and what is heard in the headphones can cause serious psychological effects. The following examples of the phenomenon are drawn from current operational practice:

* This Report should be brought to the attention of the CMTT.

a) Local Radio Operations

The presenter (who is often the only person running the radio station) listens to the programme output via an "off-air" receiver since he is responsible for checking that the signal is being radiated by the transmitter. This is shown in Fig 1. Problems can occur when the presenter is making an announcement whilst hearing his own voice via the transmitter, if more than a few milliseconds delay occurs between the studio output and the received signal fed to the headphones. In extreme cases, presenters have been so affected that they have disconnected the off-air cue-feed from the receiver, in consequence failing to detect shut-down of the transmitter

b) Outside-broadcast Operations

A contributor at an outside broadcast (OB) site will often be listening to the off-air signal, which is used as cue programme. An example of this is a commentator at a sports event. The off-air signal will be delayed by digital codecs in the distribution network, and there could also be significant delay in the feed from the commentator into the studio centre (for example, if there are digital codecs in the connections made via the telecommunication network). Figure 2 shows such a situation with a number of commentators at different events.

The commentator (who is alone at the event, without technical assistance) uses the cue signal to determine the starting point for his contribution (i.e. he is "cued" from the presenter in the studio) and he needs to listen to the cue signal whilst making his contribution. This is because he may be interrupted by questions from the presenter or by the need to terminate his contribution because of an event taking place elsewhere.

For some outside broadcasts, separate return feeds of cue programme can be provided. The return connection may be implemented with less delay than the off-air signal path, or (if the delay cannot be reduced to an acceptable level) it may carry a "mix minus" signal. In this case, the source receives back a mix of the full programme minus its own contribution. For sports events, however, it is often impractical (and invariably expensive) and off-air cueing tends to be used.

c) Laboratory tests

Preliminary tests have been conducted by the BBC in which both trained and untrained speakers have listened to their own speech, delayed, through headphones. These tests also included the effect of mixing some undelayed microphone output into the speaker's headphones. This technique has been found to produce a significant reduction in the disturbance caused by the delay.

Experienced broadcasters can normally cope with a delay of about 13 ms although it has been reported that some local-radio presenters find this unacceptable for continuous periods of duty. An inexperienced speaker, unaccustomed to hearing his own speech in headphones, may be affected by shorter values of delay, in the range 5 to 7 ms. This can cause problems, for example, when a member of the public is interviewed.

Initial results from the laboratory tests have, however, suggested that greater attention to accurate simulation of the operational conditions is necessary if anomalous results are to be avoided. Further work should therefore include study of the effect of a noisy environment, the loudness and frequency spectrum of the signal heard in the headphones and the degree of concentration required of the subject with regard to the subject matter of the test-speech material.

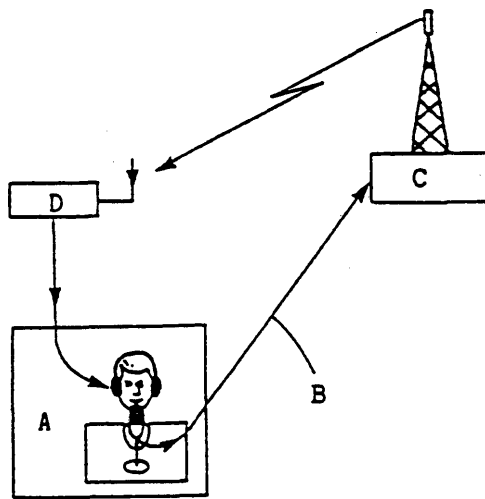
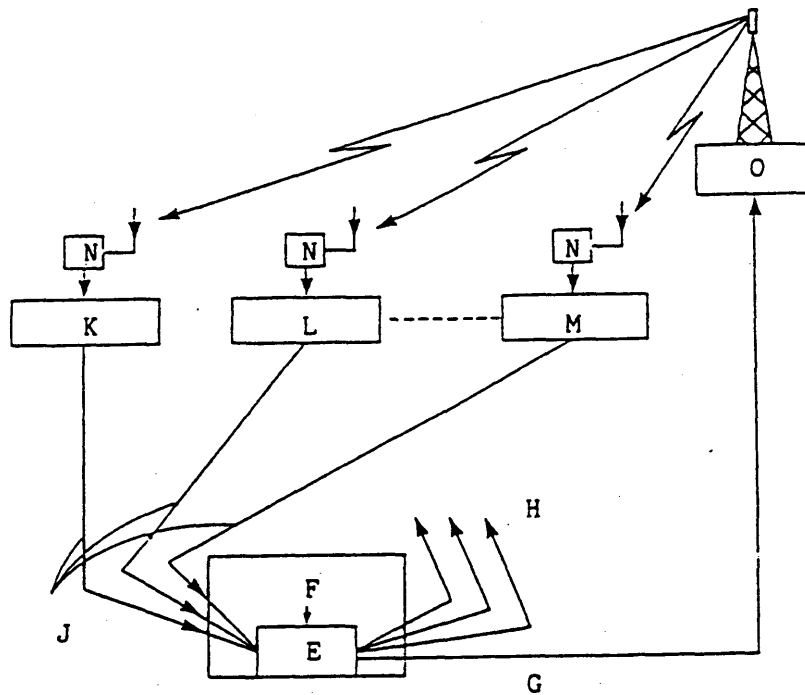


FIGURE 1 - Local radio situation

- A: studio
- B: distribution link
- C: transmitter
- D: receiver

FIGURE 2 - Outside broadcast situation

E: desk
 F: from presenter's microphone
 G: distribution link
 H: return feeds of cue programme, if available
 J: contribution links
 K: commentator at event 1
 L: commentator at event 2
 M: commentator at event n
 N: receiver
 O: transmitter

Note - Both contribution and distribution links may introduce delays.

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[1986-90]: a. 10/20 (United Kingdom).

REPORT 1204

AUTOMATIC SYNCHRONIZATION OF VIDEO AND AUDIO
AFTER TRANSMISSION

(Question 47/10)

(1990)

1. Introduction

Whenever the video and audio components of a television signal are transmitted over separate paths, there is the possibility that their transmission times will be different. When the difference is such that the viewer perceives a separation between vision and sound, enjoyment of the programme can be impaired. The subject has been studied in detail in Australia and is reported in [CCIR, 1986-90a].

A guide to the amount of allowable time difference, may be taken from the standards for film, where the picture can lead the sound by no more than two frames (83 ms) or lag behind it by no more than one frame (42 ms). The reason for the asymmetry in these allowable time differences is that sound arriving after vision is a familiar experience while sound arriving before vision is not.

Frame synchronizers are used increasingly in long distance transmissions because they confer so much operational flexibility. However, large delay differences can easily be introduced unless steps are taken to have the sound delayed equally with the vision within the synchronizer.

Delay differences are often compensated manually by an operator trying to judge the correction before other viewers can perceive it, but this is a difficult process. A method is under study by which the relative delays between vision and sound can be corrected automatically.

An audio signal in digital form is easily delayed by feeding it into a shift register and the delay can be varied by changing the address from which the output is taken. The scheme shown in Figure 1, only provides an increased delay of the audio component.

A similar delay of video is more difficult to achieve. However there is little need for such a video delay because sound following vision is more easily tolerated, and is less likely to occur. The only foreseeable cause of increased delay in the audio path is the insertion of a digital FIR low-pass filter with a sharp cut-off. Such a filter might be needed, for example, in converting the audio sampling rate from 48 kHz in the studio to 32 kHz for transmission, but this would only introduce a delay of 5.3 ms.

2. Generation of time codes

No correction method is proposed for sound in analogue form as a method has been proposed already for this purpose [Cooper, 1988]. However, it is anticipated that sound and vision will be transmitted almost entirely in the future by digital means. So long as sound and vision are transmitted in the same bit stream, no problems of synchronization need arise. Once the video signal is processed, though, for frame synchronization or whatever other purpose, a re-synchronization method will be needed.

In the digital audio interface for broadcasting studios given in Recommendation 647, § 3.5 the channel status data carries, in bytes 18-21, a 32 bit binary code which conveys the number of the first sample of the current block. Even at a 48 kHz sample rate, the 2^{32} states of this code can convey unique numbers of samples over a period of more than 24 hours. Thus, in the channel status data of the audio interface, a time indication is present, every block of 192 samples, i.e. every 6 ms with a 32 kHz sampling frequency or 4 ms with 48 kHz, to a precision of 31.25 μ s.

However, the digital audio interface is primarily designed to carry monophonic or stereophonic programmes in a studio environment and is unlikely to be used for transmission between studio centres. Thus any time-code information would have to be transcoded to take its place in the transmitted bit stream.

It is proposed therefore that a separate Audio Vision Synchronization (AVS) code be generated and transmitted with each signal, AVS(A) code in the audio bit stream and AVS(V) code in the vertical interval of the video. It is estimated that this would require 11 bits per block, i.e. an additional 1.83kbit/s with 32kHz sampling frequency or 2.75 kbit/s with 48kHz sampling.

These 11 bits would describe the timing of the first sample of the current block to the nearest millisecond. This would be done by taking the binary sample address for each block already present in bytes 18-21 of the channel status data of the Digital Audio Interface, as described earlier, and dividing it by the sampling rate in kilohertz, e.g. by 32, 44.1, or 48 as appropriate.

The resulting binary quotient would time the start of each block of digital audio in precise units of 1ms with 32kHz or 48kHz sampling and to within a maximum error of ± 1 ms with 44.1kHz sampling. With 11 bits, a total of 2048 separate states can be specified, i.e. the time code would have a complete cycle time of 2.048 seconds, from a count of zero to a count of 2047 and back to zero again.

The EBU time code specified for video indicates a count of pictures only, i.e. one every 40 ms in PAL. For NTSC, the count would be one every 33.37ms. Thus there seems no way of relating the audio time code to the conventional video time code. It is proposed therefore that an additional 11 bit signal, similar to that inserted into the audio signal, be inserted into every vertical interval of the video, indicating the time, in units of 1ms, at which it was inserted. It should be noted that this audio video Synchronizing time code is entirely independent of the conventional video time code.

It is not necessary to carry information on the sampling frequency within the AVS code as this has been taken care of already in the division of the binary address number of the Digital Audio Interface by the sampling frequency in kHz. However information about the sampling frequency can be recovered from the AVS code by counting the difference between the AVS time code numbers of successive blocks.

If the difference is taken between the numbers two blocks apart, the time will be 12.0 ms for 32 kHz sampling frequency, so that the number difference will be 11, 12 or 13. For 48 kHz sampling, the period will be 8.0 ms, producing a difference of 7, 8 or 9. Thus logic recognizing one or other group of difference numbers will identify the sampling frequency. A similar procedure applies for digital audio sampled at other rates.

The proposed procedure depends on high clock stability in the time codes. The stability of television timing signals is specified in Report 624 as ± 5 Hz in 4.433 MHz for PAL and that of digital audio in Recommendation 646 as ± 1 part in 10^5 . This comparatively wide tolerance was allowed so as to provide for audio signals originating from remote portable equipment.

However, Recommends 3 of that document states - "when an item of digital audio equipment is operating in a free-running mode, the maximum tolerance for the internal sampling frequency should be ± 1 part in 10^5 . When items of digital audio equipment are interconnected, in sound broadcasting or television applications, provision must exist for locking the internal sampling frequency (e.g.: television synchronizing signals, broadcasting house master clock, high accuracy clock from a telecommunication network); - "

The proposed scheme should work satisfactorily within these tolerances.

3. Correction of Delays.

On reception, each of the two AVS time codes, which arrive only at intervals, of a block of audio or a field of video, is used to synchronize its own continuously-running time code, as in Fig.1. These two derived time codes are then compared and their difference used as the correction signal to vary the audio delay until that difference is zero.

It might seem that, with a cycle time of 2048 states, the system could correct delay differences up to 2.048 seconds. However, as the output difference, $N(D)$, is taken as $N(V) - N(A)$. So long as the video lags behind the audio, $N(D)$ is positive, but if the sound ever lags behind the video, $N(D)$ will be negative and be indicated by a large number, $N(D) + 2048$. Thus if the sound lags by 100 ms, then the indication will be 1948.

It is therefore proposed that if the difference between the two derived time codes lies between decimal numbers 1536 to 2047, it should be interpreted as lying between -512 and -1, and therefore used to set the audio delay to zero, along with an indication "Video leads audio". Thus the system can correct errors of audio leading video by a maximum of 1.535 seconds, sufficient for the largest errors that can be foreseen.

Because the video syncs operate asynchronously from the audio-derived time code, the indicated times of the video can be in error by a maximum of ± 1 ms, so the indicated time difference $N(D)$ between audio and video will also vary by this amount. It is therefore proposed that hysteresis be built into the system so that the delay correction number $N(C)$ does not change until $N(D)$ differs from it by more than ± 2 ms, and that once change does occur, it then proceeds until the correction $N(C)$ is equal to $N(D)$. This feature is not shown in Fig.1.

It is also proposed that initial errors greater than 3 ms be corrected immediately, but that errors of ± 2 ms be corrected at the rate of 1 in 500, i.e. over a period of one second.

The variable delay can be implemented using already known techniques.

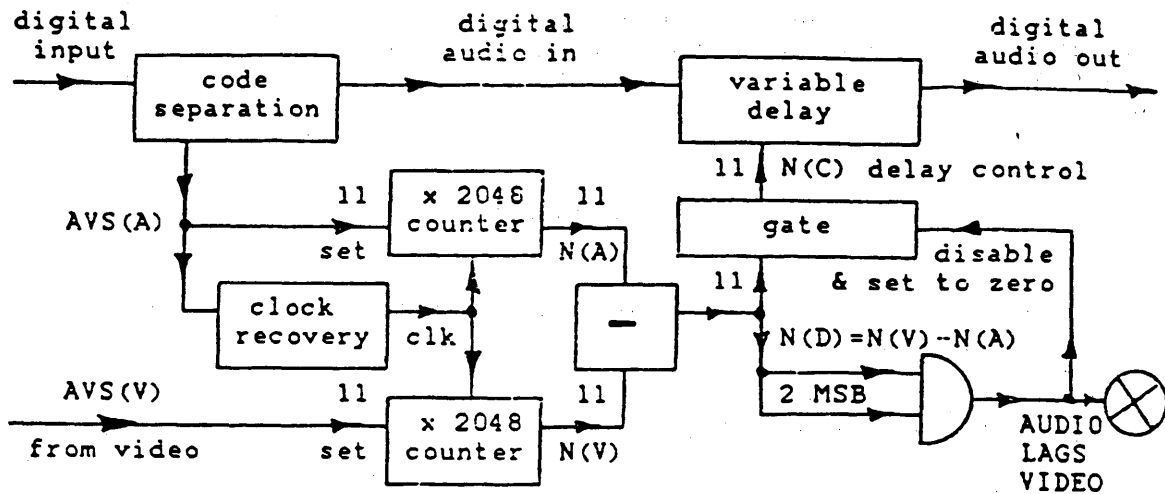


Figure 1

Schematic of automatic audio/video synchronizer

4. Transmission of A/V synchronization code

A possible method for transmitting the AVS code within each of the component signals is given in the following sections.

4.1 Audio

[CCIR 1986-90b] proposes a method by which high quality sound programme signals comprising a stereo pair sampled at 48 kHz can be carried in the ISDN, either at 1920 kbit/s (for H12 access) or 1536 kbit/s (for H11 access).

In the method proposed in the document, the channel status data of the digital audio interface, which include the 32 bit binary time of day clock, are transmitted with 48:1 time expansion. In other words, the data are transmitted transparently, but one complete channel status block is transmitted every 192 ms, instead of every 4 ms as in the case of the studio interface. The 11 bit A/V synchronizing code can be derived from these data as described earlier, the one disadvantage being that any change in transmission delay is detected only after a delay of up to 192ms. However this would not seem an important consideration.

Once the timing difference between the recovered time code and the initial time code, which might be up to 192ms, has been established, provision must be made to make any necessary fixed adjustment to it before the AVS code is derived. This aspect is not addressed in any detail in [CCIR 1986-90b] beyond the observation in section 10, Channel Status, that "Because only one data block out of 48 is transmitted, the two counters (local sample and time of day address code) must be incremented in the decoder by the appropriate amount".

4.2 Video. Before an 11 bit code can be inserted into the vertical interval, agreement would be needed on where it is to be put. The best place would be somewhere such as in the Programme Delivery Code proposed in [EBU, 1989].

5. Conclusion.

It is believed that the method proposed can synchronise the audio and video components transmitted over different paths, without occupying more than a minimum of "real estate" in either signal.

REFERENCES

COOPER, J.C. [1988] - Video-to-Audio Synchrony Monitoring and Correction - JSMPTE, September, pp 695 - 698.

EBU [1989] - Specification of the domestic video programme delivery control (PDC) EBU SPB 459 rev., March 1989.

CCIR Documents

[1986-90]: a. 10/315 (Australia), IWP CMTT/4-8 (IWP CMTT/4).

REPORT 1071

SAMPLING FREQUENCY CONVERSION AND SYNCHRONIZATION
OF DIGITAL SOUND SIGNALS

(Question 51/10, Study Programme 51B/10)

(1986)

When two or more digital sound signals are brought together in digital processing equipment, the sampling frequencies of the signals must be synchronous for the necessary arithmetic operations to take place.

In [CCIR. 1982-86a] a variable-delay synchronizer for digital sound signals having the same nominal sampling rate, is described.

The synchronizer comprises a buffer store into which sound samples are written, to be read out at the required output sampling rate. The buffer store operates effectively as a variable delay to compress or expand the input signal in time very slightly in order to achieve synchronization.

From time to time samples need to be repeated or discarded in order to prevent the store becoming completely full or empty (and thus being able to continue operating as a synchronizing element). Provided that the repetition or discarding of samples takes place during naturally-occurring "pauses" (or silences) in the programme, no audible impairment occurs.

An experimental sampling-rate synchronizer working on a variable-delay principle has shown that it can satisfactorily synchronize digital audio signals which have sampling rates which differ by up to $\pm 1 \times 10^{-5}$ from their nominal rate. However, it is recommended that sampling rates be stable to within $\pm 1 \times 10^{-6}$ of nominal rate where this type of synchronizer is used.

The experimental results verified that no perceptible impairment results from the use of this type of synchronizer. Moreover, the technique does not cause any increase in the level of quantization noise.

REFERENCES

CCIR Documents

[1982-86]: a. 10/19 (EBU).

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GILCHRIST, N. H. C. [October, 1980] Sampling-rate synchronization of digital sound signals by variable delay. *EBU Rev. Tech.*, 183, 222-226.

REPORT 953-2

DIGITAL CODING FOR THE EMISSION OF HIGH-QUALITY SOUND SIGNALS IN
SATELLITE BROADCASTING (15 kHz NOMINAL BANDWIDTH)

(Question 51/10, Study Programmes 51C/10, 2F/10 and 11)

(1982-1986-1990)

This text may be found in the Annex to Part 2 of Volumes X and XI.

REPORT 1237

SATELLITE NEWS GATHERING

(1990)

This text may be found in the Annex to Volume XII.

SECTION 10D: RECORDING OF SOUND PROGRAMMES

The texts relating to this Section may be found in the Annex to Part 3 of Volumes X and XI.

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SECTION 10E: BROADCASTING SERVICE (SOUND) USING SATELLITES

The texts relating to this Section may be found in the Annex to Part 2 of Volumes X and XI.

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DECISIONS

DECISION 18-6*

**DIGITAL SYSTEMS FOR THE TRANSMISSION OF
SOUND-PROGRAMME AND TELEVISION SIGNALS**

(1974-1976-1978-1981-1983-1989)

The CMTT,

CONSIDERING

- (a) that CCITT Study Group XVIII is studying the problems associated with the establishment of an integrated services digital network (ISDN) for the transmission of all kinds of signals, including those for sound-programme and television;
- (b) that CCITT Study Group XV is studying digital multiplex and line systems, including equipment for the digital transmission of sound-programme and television signals, and coding for interactive audio-visual services;
- (c) that CCITT Study Group I is studying the end-to-end service quality of all interactive audio-visual services supported on the ISDN;
- (d) that there is an urgent need to ensure that the standards adopted for the integrated services digital network shall be compatible with the requirements necessary for the transmission of sound-programme and television signals;
- (e) that CCIR Study Groups 10 and 11 are studying the use of digital techniques in programme production and emission;
- (f) that CCIR Study Groups 10 and 11 have the primary responsibility for specifying the overall performance, apportioning and methods of assessment of systems in the broadcasting service;
- (g) that the CMTT is studying the performance requirements applicable to _____ digital transmission circuits for sound-programmes and television, either for contribution or distribution circuits;
- (h) that the CMTT has to consider the transition period during which technological changes will take place;
- (j) that there is a need to co-ordinate the studies undertaken by Study Groups 10 and 11 and the CMTT, together with those in progress in CCITT Study Groups XV and XVIII,

DECIDES

1. that a Joint Interim Working Party CMTT-10-11/2 be established with the following terms of reference:
 - 1.1 to examine the results of studies on digital techniques effected in Study Groups 10 and 11 and the CMTT and to make any proposals necessary to ensure maximum compatibility between the recommendations of the Study Groups;
 - 1.2 to co-operate with CCITT Study Group XV and XVIII to ensure maximum compatibility between the Recommendations of the CCIR and the CCITT related to the digital transmission of sound-programme and television signals;

* This Decision should be brought to the attention of CCITT Study Groups I, IV, XV and XVIII.

- 1.3 to cooperate with CCITT Study Group I to ensure appropriate compatibility between interactive audio-visual services and the broadcasting services;
2. that the Chairman of JIWP CMTT-10-11/2 shall be a representative of the Administration of the United Kingdom;
3. that JIWP CMTT-10-11/2 shall be composed of the Chairmen and Vice-Chairmen of the CCIR Study Groups 10, 11 and the CMTT and representatives of the CCITT Study Groups concerned, as well as of representatives nominated by the following Administrations: Germany (Federal Republic of), Australia, Canada, Denmark, United States of America, Finland, France, India, Italy, Japan, Norway, the Netherlands, United Kingdom, Sweden, Switzerland, the USSR and Yugoslavia;
4. that, as far as possible, the work of JIWP CMTT-10-11/2 shall be carried out by correspondence;
5. that coordination of digital studies in the CCIR shall be carried out in accordance with Annex I;
6. that the specification of performance for digital transmission circuits in general shall be in accordance with the principle that the first objective for systems employing digital or digital and analogue methods should be that the overall performance is not worse than the equivalent analogue system. Later in the development of digital systems it may be possible to improve standards of performance.

ANNEX I

COORDINATION OF DIGITAL STUDIES IN STUDY GROUPS 10, 11 AND CMTT

1. Organization of work

To provide effective coordination of studies of Study Groups 10, 11 and the CMTT on digital techniques, the results of the work of these Study Groups will be examined by the Joint Interim Working Party before transmission to the CCITT Study Groups I, XV and XVIII.

—If necessary, it will make proposals to ensure compatibility between the results of the Study Groups. Between Study Group meetings this work will continue by correspondence. The results of the study by correspondence will be transmitted to the Study Groups prior to their meeting. The timing of the work of the Joint Interim Working Party must take account of meetings of the CCIR Study Groups concerned.

The JIWP CMTT-10-11/2 considers that the work of coordination would be expedited if the meetings of Study Groups 10, 11 and the CMTT could be held simultaneously and separate from meetings of the other CCIR Study Groups.

2. Coordination of studies

The JIWP CMTT-10-11/2 proposes that particular parts of studies concerning digital techniques should receive emphasis by Study Groups in the manner given in Table I.

TABLE I

*Distribution of the emphasis on studies concerning
digital techniques among certain Study Groups of the CCITT and the CCIR.
the CMTT, and the Joint Interim Working Party CMTT-10-11/2*

1. Specification of all ISDN network aspects (CCITT Study Group XVIII); specification of transmission equipment and systems (CCITT Study Group XV); and specifications relating to telecommunication services (CCITT Study Group I).
- 2.* Coding methods, including redundancy reduction and standards to be adopted for foreseen application of:
 - 2.1 Sound-broadcasting and television (see Note 1)
 - 2.1.1 studio installations including recording

sound (SG 10) television (SG 11);	
 - 2.1.2 terrestrial radio-relay links
 - 2.1.3 cable transmission links
 - 2.1.4 satellite transmission links
 - 2.1.5 broadcasting from terrestrial transmitters and from satellites

sound (SG 10) television (SG 11).	
 - 2.1.6 B-ISDN (distribution) (CMTT)
 - 2.2 Interactive audio-visual services (Study Group XV)
- 3.* Compatibility of the relevant coding methods (see § 2) for:
 - 3.1 studio installations, including recording;**
 - 3.2 terrestrial radio-relay links;
 - 3.3 cable transmission links;
 - 3.4 satellite transmission links;
 - 3.5 broadcasting from terrestrial transmitters and from satellites between themselves and with the other services envisaged under § 1
 - 3.6 Interactive audio-visual services
- 4.* Study of ways of protecting against digital transmission impairments (e.g. errors, jitter, slip) for:

	<i>Studio</i>	<i>Transmission link</i>
4.1 sound programmes	(SG 10)	(CMTT).
4.2 television	(SG 11)	
- 4.3 Interactive audio-visual services (Study Group XV)

* Close cooperation between the Joint Interim Working Party CMTT-10-11/2 and Study Groups XV and XVIII is essential for these studies.

** Aspects of compatibility relating to the exchange of recordings are matters for Study Groups 10 and 11.

5. Study of transmission methods over various transmission media _____ for:
 - 5.1 studio installation, including recording | sound (SG 10)
television (SG 11);
 - 5.2 terrestrial radio-relay links (SG 9);
 - 5.3 optical and metallic cable transmission links (SG XV);
 - 5.4 satellite transmission links (SG 4);
 - 5.5 broadcasting from terrestrial transmitters and from satellites | sound (SG 10)
television (SG 11);
 - 5.6 multiplex equipment and digital line systems on optical fibre cables in local networks, including ISDN access (Study Group XV).
6. Equipment for digital coding and multiplexing for sound and television programmes, including insertion of programme signals into telephony digital links (Study Group XV).
7. Methods of code conversion of programme signals in digital form, for example, conversion of signals encoded in a form suitable for studio application into a form suitable for long-distance transmission (see Note 1):
 - 7.1 sound programmes | (CMTT).
 - 7.2 television
8. Definitions of hypothetical reference circuits and chains for sound programme and television circuits (CMTT).
9. Transmission requirements to be satisfied by the hypothetical reference circuits and chains defined in § 8 (CMTT).
10. Definitions of hypothetical reference circuits and reference connections for the integrated services digital network (Study Group XVIII).
11. Measuring methods for performance and test signals for sound programmes and television:
 - 11.1 within each part of the system for:
 - 11.1.1 studio installations, including recording | sound (SG 10)
television (SG 11);
 - 11.1.2 terrestrial radio-relay links (SG 9);
 - 11.1.3 cable transmission links (Study Groups IV and XV);
 - 11.1.4 satellite transmission links (SG 4);
 - 11.1.5 broadcasting from terrestrial transmitters and from satellites | sound (SG 10)
television (SG 11);
 - 11.2 between the input ports and output ports of the transmission chain of § 11.1.2, 11.1.3 and 11.1.4 (CMTT).
12. Study of subjective quality of signals encoded and transmitted in digital form, or which have been processed (e.g. chroma-key/colour-matte or other special effects) using signals transmitted in digital form for (see Note 1):
 - 12.1 sound programmes (SG 10);
 - 12.2 television (SG 11).

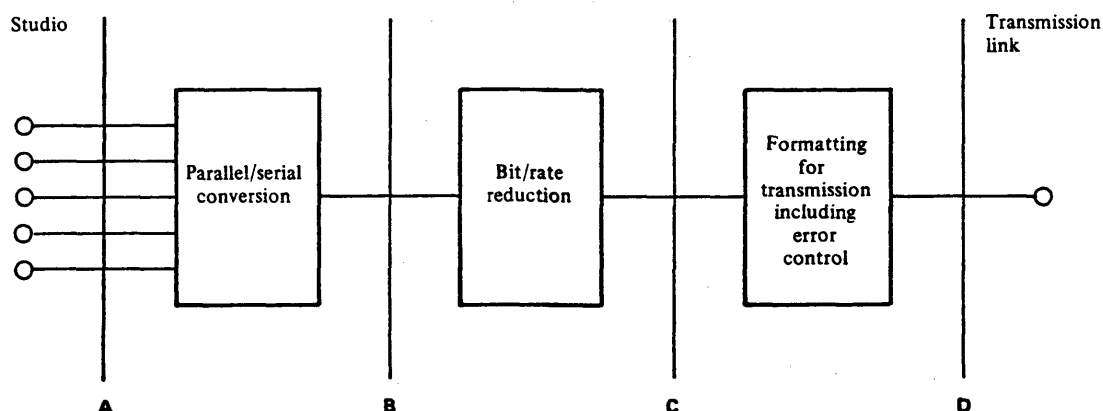
13. Interfaces

- 13.1 within studios and within broadcasting transmitters (Study Groups 10 and 11).
- 13.2 within transmission links other than in the B-ISDN (CMTT).
- 13.3 between studios and transmission links and between transmission links and broadcasting transmitters (CMTT), see Note 2.
- 13.4 within the B-ISDN including the B-ISDN user-network interface (Study Group XVIII).

Note 1. – The basic representations of sound and vision signals by digital codes and the associated quality requirements are studied and recommended by Study Groups 10 and 11. Irreversible transformations of these basic representations into other representations required for transmission should be subject to subjective and objective evaluation criterias defined by Study Groups 10 and 11 before their recommendation by the CMTT.

Note 2. – Digital interfaces for sound-programme and television signals.

Digital interfaces between studios and transmission circuits may involve either serial or parallel transmission of information and, in general, may involve processing for bit-rate reduction and formatting to suit the transmission circuit. These stages are illustrated in the following diagrams.



The interface at the output of the studio may be either A, B or C. Depending on the requirements of the transmission links, the block between interfaces C and D may, or may not, be required.

The process of bit-rate reduction shown between interfaces B and C may be carried out in the studio prior to parallel/serial conversion. This block may also include some processes for error protection.

Taking account of the above, Study Groups 10 and 11 (for sound and television respectively) should have the prime responsibility for interfaces A and B but the CMTT should have the opportunity to comment on them before final agreements are reached. Similarly, for interface C, the CMTT should have the prime responsibility but Study Groups 10 and 11 should have the opportunity to comment before final agreements are reached. In the case of interface D this should be studied with close coordination between the CMTT and CCITT Study Group XVIII but Study Groups 10 and 11 should have the opportunity to comment before final agreements are reached. This interface will also be of interest for access to the ISDN. Prime responsibility for interface D is in CCITT Study Group XVIII.

DECISION 71-1

**CONTINUATION OF STUDIES ON COMPATIBILITY BETWEEN THE
AERONAUTICAL RADIONAVIGATION SERVICE IN THE BAND 108-117.975 MHz,
THE AERONAUTICAL MOBILE (R) SERVICE IN THE BAND 117.975-137 MHz
AND THE FM SOUND BROADCASTING STATIONS IN THE BAND ABOUT 87-108 MHz**

(1985-1989)

Pursuant to the XVIIth Plenary Assembly of the CCIR (Düsseldorf, 1990), studies relating to this question will be dealt with by Study Group 12.

This Decision is published separately and appears in the CCIR publication entitled "Compatibility between the broadcasting service in the band of about 87-108 MHz and the aeronautical services in the band 108-137 MHz".

DECISION 72-1*

DATA BROADCASTING SERVICES

(1986,1989)

CCIR Study Groups 10 and 11,

CONSIDERING

(a) that data broadcasting services on terrestrial and satellite channels either associated with sound or vision signals or exclusively, are under development in many countries;

(b) that data broadcasting services, such as teletext, programme identification in television and radio data services in sound broadcasting are already implemented in many countries;

(c) that new high-quality sound and vision services should be accompanied by data services with appropriate improved quality;

(d) that the CCITT is studying similar services in telecommunication networks,

DECIDE

1. that Joint Interim Working Party 10-11/5 should continue its work within the general terms of reference of CCIR Study Groups 10 and 11 and with the following specific terms of reference:

1.1 to carry out studies on the presentation layer requirements of some non-Latin based alphabets and non-alphabetic writing systems as indicated in Recommendation 653;

1.2 to carry out studies on the presentation layer requirements for data services associated with HDTV signals (see Report 1225);

1.3 to propose definitions of data broadcasting services in television and sound broadcasting channels, whether terrestrial or satellite**;

1.4 to study the data coding for these services, keeping coordination with JIWP 10-11/3, which is responsible for the accommodation of data signals in terrestrial and satellite broadcasting channels**;

* This Decision should be brought to the attention of the CCITT, the IEC and the ISO. This topic should be studied jointly with Study Group 10.

** The Study Programmes derived from Question 29-2/11, Study Programmes 44J/10, 46H/10, 47A/10 and 51D/10 as well as Study Programme 2N/10 and 11 are relevant.

- 1.5 to define appropriate service quality criteria and assessment methods;
 - 1.6 to study the requirements for international exchange and system transcoding;
 - 1.7 to study the requirements for equipment interface standardization;
 - 1.8 to maintain contact with the CMTT, the appropriate CCITT Study Groups, the IEC and the ISO;
2. that JIWP 10-11/5 should, as far as possible, work by correspondence; however, it may meet upon proposal of its Chairman, following consultation with the Chairmen of Study Groups 10 and 11 and the Director of the CCIR;
3. that the Chairman and the composition of JIWP 10-11/5 shall be as shown in Annex I.

ANNEX I

The following Administrations and International Organizations have indicated that they wished to participate in the work of Joint Interim Working Party 10-11/5:

Administrations:

Germany (Federal Republic of)
Australia
Brazil (Federative Republic of)
Canada
China (People's Republic of)
Spain
United States of America
France
Iran (Islamic Republic of)
Italy
Japan
Netherlands
United Kingdom
Sweden
USSR
Yugoslavia (Socialist Federal Republic of)

International Organizations:

EBU
OIRT

Chairman of Joint Interim Working Party 10-11/5:

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DECISION 76-1

SATELLITE NEWS GATHERING (SNG)

(Question 13/CMTT, Study Programmes 13H/CMTT and 22D/CMTT)

(1987-1989)

CCIR Study Groups 4, 10, 11 and CMTT,

CONSIDERING

- (a) that satellite transmissions using transportable or portable earth stations is an invaluable and at times the only viable solution to the timely transmission of television news from remote locations;
- (b) that throughout the world, where news events take place, standardized and uniform technical and operating procedures should be established to ensure prompt activation of SNG service;
- (c) that there was a unanimous Recommendation of the Fifth World Conference of Broadcasting Unions in Prague, 1986, on the use of international auxiliary broadcast frequencies and transportable earth stations;
- (d) that the special characteristics needed for transportable or portable transmitting earth stations may necessitate acceptance of performance objectives differing from those specified in Recommendation 567 for general purpose satellite television connections;
- (e) that there will be a requirement to provide, on the same uplink service, auxiliary circuits for programme and technical coordination, and for the management of the transportable earth station interface;
- (f) that SNG services from any specific location will be classified as occasional and/or temporary;
- (g) that the very nature of SNG requires that earth stations be activated in an expedient manner, the philosophy of which is not compatible with the long advance notice periods normally established for fixed permanent services;
- (h) that it is expected in the future that HDTV transmissions could originate from portable uplink facilities;

(j) that the ITU Constitution states in its Preamble: "...fully recognizing the sovereign right of each State to regulate its telecommunication... ."

DECIDE

1. that Joint Interim Working Party JIWP CMTT-4-10-11/1 be continued, to prepare, within the Terms of Reference of Study Groups CMTT, 4, 10 and 11, an overall strategy for SNG transmissions, proposals intended to solve the technical, operating and organizational aspects associated with the use of transportable or portable transmitting earth stations for SNG;
2. that the Terms of Reference of the JIWP should be as follows:
 - 2.1 to define the technical quality specifications for the programme video signal acceptable for SNG;
 - 2.2 to define the number and the technical quality specifications for the programme audio signals acceptable for SNG;
 - 2.3 to define the number and the technical quality specifications of auxiliary circuits required for SNG operations;
 - 2.4 to study uniform technical parameters for SNG that may be applicable on a geographically wide scale;
 - 2.5 to study uniform operating procedures for SNG that may also be applicable on a geographically wide scale;
 - 2.6 to study the overall transmission and performance objectives for HDTV transmission by portable satellite earth stations for SNG;
 - 2.7 to determine the technical characteristics of the specific equipment required to meet the objectives in 2.6;
 - 2.8 to identify the operational requirements related to HDTV transmission by portable satellite earth stations for SNG;
 - 2.9 to prepare draft Recommendations on the study items above;
 - 2.10 to investigate and prepare a Report on means to simplify the procedures required to obtain, as expeditiously as possible, temporary authorization to operate SNG facilities;
3. that, in accordance with §2.3.3 of CCIR Resolution 24, Study Groups CMTT, 4, 10 and 11 are jointly entrusted with the responsibility of the work of the JIWP, and in accordance with §2.3.9 of Resolution 24, CMTT is designated to coordinate the work of the JIWP. The JIWP shall submit its reports and results of its work to joint meetings of these Study Groups, when possible;
4. that the work of the JIWP should be completed in the course of the current Study Period; a report and the results of its work shall be submitted to the Final Study Group Meetings in 1993;
5. that the JIWP should, as far as possible, work by correspondence; however it may meet when this is considered necessary by its Chairman, by the Chairmen of Study Groups CMTT, 4, 10 and 11, and by the Director, CCIR;

6. that the JIWP will be chaired by:

Mr. Joseph A. COLSON (NANBA)
Chairman, NANBA Technical Committee
CTV Television Network Limited
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Toronto, CANADA M4Y 1T5

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and the Vice-chairmen will be:

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DECISION 78*

ACOUSTICAL QUALITY PARAMETERS FOR LISTENING CONDITIONS

(1987)

CCIR Study Group 10,

CONSIDERING

- (a) that the unification of closely specified listening conditions at the studio end is an essential prerequisite for a well-defined subjective assessment and quality control, in the interest of the international programme exchange;
- (b) that knowledge about acoustical quality parameters, especially sound-field parameters, is insufficient at the present time to standardize in the CCIR reference listening conditions (requirements for control and listening rooms, studio monitor loudspeakers etc.) with the necessary accuracy and with the appropriate measuring conditions (see Study Programme 50D/10);
- (c) that in the Nordic countries and in the OIRT member countries positive results have been reached following the guidelines for reference listening rooms set out in [CCIR, 1986-90a] based upon considerations presented in [CCIR, 1986-90b],

DECIDES

- 1. that an Interim Working Party 10/9 should be set up within the general terms of reference of Study Group 10 and with the following specific terms of reference:
 - 1.1 to study the acoustical quality parameters and values which are necessary to set up reproducible conditions for (reference) control and listening rooms and studio monitor loudspeakers;
 - 1.2 to define measuring methods for the checking and the alignment of the acoustical quality parameters of control and listening rooms and studio monitor loudspeakers;
 - 1.3 to present its Report to the next Meeting of Study Group 10;
 - 1.4 to maintain contact with the IEC;

** This Decision should be brought to the attention of the IEC.

2. that IWP 10/9 should, as far as possible, work by correspondence, however, it may meet upon the proposal of its Chairman, following consultation with the Chairman of Study Group 10 and the Director of the CCIR;
3. that the Chairman and the composition of IWP 10/9 shall be as shown in Annex I.

ANNEX I

The following administrations and international organizations have indicated that they wished to participate in the work of IWP 10/9:

Administrations

Federal Republic of Germany
United States of America
France
Japan
Netherlands
German Democratic Republic
United Kingdom
Sweden
USSR

International Organizations

EBU
OIRT

Chairman of IWP 10/9

Tom MAGCHIELSE (NOS-NOB)
Technical Research and Development
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REFERENCES

CCIR Documents

[1986-90]: a. 10/54 (German Democratic Republic); b. 10/53 (German Democratic Republic).

DECISION 79-1

CALCULATION OF THE RADIATION PATTERNS OF LF, MF AND HF ANTENNAS

(1987-1989)

CCIR Study Group 10,

CONSIDERING

- (a) the studies indicated in Study Programmes 44H/10 and 45F/10 with respect to the originating Questions 44/10 and 45/10;
- (b) Resolutions Nos. 516 and 514 of the WARC HFBC-87;
- (c) CCIR Resolutions 24 and 33;
- (d) the extensive documentation on HF-antenna characteristics already available;
- (e) the desirability of completing work on transmitting antennas in band 7 (HF);
- (f) the need to extend the documentation to include transmitting and receiving antennas in bands 5 (LF) and 6 (MF) and receiving antennas in band 7 (HF),

UNANIMOUSLY DECIDES

1. that Interim Working Party 10/1 established by Circular-letter CE 10/1255 of 2 August 1984, reconvened within the general terms of reference of Study Group 10 should by this Decision, have its mandate renewed to carry out the following studies:
 - 1.1 to examine the formulae used to derive the gain and directivity patterns for LF, MF and HF transmitting and receiving antennas and to consider the effect of imperfect and irregular ground;
 - 1.2 to evaluate the radiation patterns of new types of antennas not covered by CCIR studies in the periods from 1982 to 1990;
 - 1.3 to consider the most appropriate presentation of the calculated results;
 - 1.4 to consider the most economic and effective means of publishing the results and to propose a procedure for their inclusion in the appropriate CCIR publications;

- 1.5 to study the most suitable procedure for the provision of relevant antenna data for inclusion in the IFRB technical standards after approval by the CCIR;
- 1.6 to propose the most effective way of updating the published CCIR antenna data;
- 1.7 to consider the most appropriate designations for different types of LF, MF and HF antennas;
2. to produce computer-based procedures in a standardized form giving the gain and directivity patterns of transmitting and receiving antennas in common use in bands 5 (LF), 6 (MF) and 7 (HF). The procedures for HF antennas should be suitable for use with computer-based ionospheric propagation prediction programmes;
3. that IWP 10/1 should propose amendments to all relevant Study Group 10 texts or prepare new ones as appropriate;
4. that IWP 10/1 should submit a report to the next meetings of Study Group 10;
5. that IWP 10/1 should as far as possible work by correspondence but, that it may meet on the proposal of its Chairman, following consultation with the Director of the CCIR;
6. that IWP 10/1 should give priority to completing the work on transmitting and receiving antennas used in HF-broadcasting.
7. that the Chairman and the composition of IWP 10/1 shall be as shown in Annex I.

ANNEX I

1. The following administrations, international organizations and recognized private operating agencies have indicated that they wish to participate in the work of IWP 10/1:

Administrations:

Germany (Federal Republic of)
Brazil
Canada
China
United States
Finland
France
India
Italy
Norway
New Zealand
Netherlands
United Kingdom
Sweden
Switzerland
USSR
Vatican City State
Yugoslavia
a representative of Study Group 3
a representative of Study Group 6

International organizations and recognized private operating agencies:

ABU
Deutsche Welle
NDR
EBU

2. Chairman of Interim Working Party 10/1:

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Federal Republic of Germany

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DECISION 87 *

DETERMINATION OF THE COORDINATION AREA

(Appendix 28 of the Radio Regulations)

(1989)

This text may be found in the Annex to Part 2 of Volumes IV and IX.

*) According to the decision of Chairmen and Vice-Chairmen's Meeting (Geneva, 4-6 July, 1990) the tasks of JIWP 2-4-5-8-9-10-11/1 on the determination of the coordination area are transferred to Study Group 12 for study by Task Group 12/3.

DECISION 94

MULTI-CHANNEL SOUND SYSTEMS

(ESPECIALLY SUITED TO ACCOMPANY HIGH-DEFINITION
AND ENHANCED TELEVISION SYSTEMS)

(1989)

CCIR Study Group 10

CONSIDERING

- (a) that investigations into the requirements for sound accompanying HDTV are proceeding in several countries;
- (b) that the results of this work are referred to in CCIR Report 1072;
- (c) that it is widely recognized that a two-channel sound system has serious limitations, particularly with the larger screen sizes envisaged for HDTV;
- (d) that further investigations about multi-channel sound reproduction not associated to television are in progress;
- (e) that the requirements of cinema presentation may differ from those that apply in the home;
- (f) that broadcast HDTV signals, and those delivered by other media, should be capable of giving appropriate sound quality with a wide range of domestic loudspeaker configurations, including compatibility with two-channel stereophonic and monophonic listening;
- (g) that Joint Interim Working Party 10-11/3 has been requested, as part of its continuing activities, to study the means by which HDTV signals may be broadcast;
- (h) that in this work, JIWP 10-11/3 is advised about the necessary characteristics of HDTV video signals by IWP 11/6;
- (j) that IWP 10/9 is gathering information about optimum listening conditions;
- (k) that JIWP 10-11/6 is expert in methods of subjective quality assessment;
- (l) that IWP 11/7 has requested information about the characteristics of sound signals accompanying television signals over contribution links;
- (m) that in considering multi-channel sound, it is desirable to separate the requirements of production, delivery and presentation, though these are mutually interacting;
- (n) that JIWP 10-11/4 is formulating Recommendations for the recording of television pictures with associated sound,

DECIDES

1. that an Interim Working Party 10/12 be established within the general terms of reference of Study Group 10:
 - 1.1 to establish optimum arrangements for HDTV sound production and recording;
 - 1.2 to establish optimum arrangements for sound production for HDTV cinema viewing conditions;
 - 1.3 to establish a set of domestic loudspeaker arrangements for HDTV viewing, having different levels of performance, in which the signals required for the simpler arrangements (including monophonic and two-channel stereophonic listening) are readily derived from those envisaged for the more complex arrangements, and vice-versa;
 - 1.4 to determine, in accordance with 1.3, the number and characteristics of the sound signals that should be broadcast with HDTV pictures, or associated with the delivery of HDTV pictures on other media;
 - 1.5 to determine to what extent the conclusions obtained in the above investigations are applicable to the sound signal when not associated to a television picture;
 - 1.6 to determine the values of the technical parameters that might be chosen to improve audibility of the hearing impaired;
2. that IWP 10/12 should liaise where appropriate with IWPs 10/9 and 11/7 and with JIWPs 10-11/3, 10-11/4 and 10-11/6;
3. that IWP 10/12 should, as far as possible, work by correspondence, however, it may meet upon the proposal of its Chairman, following consultation with the Chairman of Study Group 10 and the Director of the CCIR;
4. that IWP 10/12 should present a draft Recommendation or Recommendations by no later than the 1991 Interim Meeting;
5. that the Chairman and the composition of IWP 10/12 shall be as shown in Annex I.

ANNEX I

The following Administrations and International Organizations have indicated that they wish to participate in the work of IWP 10/12:

Administrations:

Germany (Federal Republic of)
 Australia
 Canada
 United States
 France
 Italy
 Japan
 Netherlands
 German Democratic Republic
 United Kingdom
 Sweden
 USSR
 Yugoslavia (Socialist Federal Republic of)

International Organizations:

CBS
OIRT
Philips International

Chairman of IWP 10/12:

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Audio Production and
Reproduction Systems Section
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DECISION 95*

SUBJECTIVE ASSESSMENT OF SOUND AND TELEVISION PICTURE QUALITY

(1989)

CCIR Study Groups 10 and 11

CONSIDERING

- (a) that Recommendation 500 has established some methods of test for assessing the subjective quality of television systems;
- (b) that Report 1082 describes new methods and approaches to picture quality assessment, Report 1205 describes advanced methods for the presentation and analysis of test results, Report 405 presents further information about subjective tests, Report 1206 presents information about subjective picture quality in digital television, and Report 1216 presents information about picture quality assessment in high-definition television;
- (c) that Recommendation 654 establishes reference relationships between subjective quality and objective impairments for the more common forms of impairment in current NTSC, PAL and SECAM systems while Report 959 tabulates measured values for particular systems;
- (d) that Report 313 contains a lengthy bibliography of published work and CCIR contributions on subjective assessments;
- (e) that the introduction of new kinds of television signal processing such as digital coding and bit-rate reduction, new kinds of television signals using time-multiplexed components and new services such as enhanced television and HDTV, may require changes in the methods of making subjective assessments;
- (f) that significant progress has been made in the measurement of human perception;
- (g) that Recommendation 562 of Study Group 10 has established some methods for assessing the subjective quality of sound;
- (h) that the introduction of new sound broadcasting systems may require advanced development of the existing methods of subjective assessments;
- (j) that there is a common core of assessment methodology for sound and vision;

*This Decision supersedes Decision 66 which is hereby deleted.

(k) that some assessments may involve both sound and vision as components of a single service;

(l) that both sound and vision methodology will benefit from a cross flow of ideas,

DECIDE

1. that a Joint Interim Working Party 10-11/6 should be formed within the general terms of reference of Study Groups 10 and 11, and with the following specific terms of reference:

1.1 to review Questions 3-2/11 and 14-1/11, and their associated Study Programmes, in order to:

- identify those tasks which have been completed, and
- define new areas where studies are required;

1.2 to review the texts in Volume XI, Part 1 of the Recommendations and Reports of the CCIR which are concerned with the subjective evaluation of picture quality with a view to:

- consolidate and eliminate repetitive material,
- improve clarity and brevity,
- make them more comprehensible to non-specialists,
- provide a framework for the introduction of new material without reducing the clarity of the complete text;

1.3 to examine current trends in television systems described in CONSIDERING (e) and determine what changes, if any, will be required in the methodology of subjective testing to accommodate these trends and prepare any draft revision of Study Programmes which may be appropriate;

1.4 to collect, collate and harmonize contributions to studies on the methodology of subjective testing of new systems such as those described in CONSIDERING (e);

1.5 to prepare regularly updated draft Reports on the progress of such studies;

1.6 to review Study Programme 50C-1/10 in order to identify those tasks which have been completed, and define new areas where studies are required;

1.7 to review Recommendation 562-2 with a view to its consolidation and development in the light of recent developments and knowledge;

1.8 to examine current trends in sound broadcasting systems and determine what additions or changes, if any, will be required in the methodology of subjective assessments to accommodate these trends, and prepare draft new texts as appropriate;

1.9 to maintain liaison with IWP 10/12 in order to advise on subjective assessments of multi-channel sound systems, including those suited to high-definition television and enhanced definition television;

1.10 to prepare regularly updated Reports and Recommendations to Study Groups 10 and 11.

ANNEX I

The following Administrations, International Organizations and Recognized Private Operating Agencies have indicated that they wish to participate in the work of Joint Interim Working Party 10-11/6. These include those administrations who were members of IWP 11/4 and those who have expressed an interest in participating in the new JIWP studies concerning sound quality.

Administrations:

Germany (Federal Republic of)
Australia
Canada
China (People's Republic of)
United States of America
France
Italy
Japan
Netherlands
German Democratic Republic
United Kingdom
Sweden
Switzerland
USSR
Yugoslavia (Socialist Federal Republic of)

International Organizations and Recognized Private Operating Agencies:

Broadcasting Corporation of New Zealand (BCNZ)
EBU
Nippon Hoso Kyokai (NHK)
Norddeutscher Rundfunk (NDR)
OIRT
Zweites Deutsches Fernsehen (ZDF)

Chairman of Joint Interim Working Party 10-11/6:

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Geneva, Switzerland

Telephone: +41 22 798 7766
Telefax: +41 22 798 5897
Telex: 415 700 ebu ch

assisted by a Vice-Chairman:

Miss B. JONES
Science and Technology Department
National Association of Broadcasters
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United States of America

DECISION 96*

CCIR STUDIES TO BE CARRIED OUT FOR SUBMISSION TO
THE WORLD ADMINISTRATIVE CONFERENCE (WARC-93)
DEALING WITH MATTERS CONNECTED WITH
THE HF BROADCASTING SERVICE

(1989)

CCIR Study Group 10,

NOTING

(a) that the ITU Plenipotentiary Conference (Nice, 1989) in Resolution No. [PL-B/1] resolved that a World Administrative Radio Conference for Dealing with Matters Connected with the Broadcasting Service in the HF Bands be held in Geneva during the first quarter of 1993 for a period of four weeks; and

(b) that the same Resolution asks the ITU Administrative Council to establish the agenda for the Conference, taking into account the Resolutions and Recommendations of the WARC HFBC-87 relating to the HFBC Planning System and procedures.

CONSIDERING

(a) that the HFBC-87 revised or adopted a number of Recommendations and Resolutions, with explicit instructions to the CCIR:

- i. Recommendation No. 503 (WARC-79) concerning HF broadcasting, with a specific invitation to the CCIR to carry out further studies with a view to updating CCIR Recommendations 328, 332 and 205 concerning (1) out-of-band spectrum provisions, (2) receiver selectivity and (3) synchronized-frequency transmitter operation, respectively;
- ii. Recommendation No. 510 concerning planning parameters for the double-sideband (DSB) system in the HF bands allocated exclusively to the broadcasting service, with an invitation to the CCIR to continue to study the values of the planning parameters listed in the Annex to Recommendation No. 510;
- iii. Recommendation No. 514 concerning improvements to the propagation prediction method to be used for the HF bands allocated exclusively to the broadcasting service, with an invitation to the CCIR to undertake studies of the propagation prediction method adopted by the Conference and to recommend both improvements in the method and later, if necessary, an improved method to be used in the future;

* This Decision should be brought to the attention of Study Groups 1 and 6.

- iv. Recommendation No. 515 concerning the introduction of transmitters and receivers capable of both double-sideband (DSB) and single-sideband (SSB) modes of operation, with an invitation for the CCIR to complete its studies into receivers for SSB;
 - v. Recommendation No. 517 concerning relative RF protection ratio values for single-sideband (SSB) emissions in the HF bands allocated exclusively to the broadcasting service, with an invitation to the CCIR to continue to study the values of relative RF protection ratio for the different cases and frequency separations covered in the Annex to Recommendation No. 517;
 - vi. Resolution No. 514 concerning the procedure to be applied by the IFRB in the revision of the relevant parts of its technical standards used in the HF bands allocated exclusively to the broadcasting service, which resolves that the IFRB review its Technical Standards relating to the technical parameters of HF broadcasting following each CCIR Plenary Assembly, in consultation with administrations, and with specific interaction with the CCIR;
 - vii. Resolution No. 516 concerning antennas to be used for the planning of the HF bands allocated exclusively to the broadcasting service, with an invitation for the CCIR to continue to update the CCIR special publication on HF Antenna Diagrams and for the IFRB to base its Technical Standards for HF antenna characteristics on the CCIR special publication on Antenna Diagrams and information supplied by administrations;
- (b) that the studies called for in iii. above are in progress in IWPs 6/1 and 6/15;
 - (c) that the studies called for in vii. above are in progress in IWP 10/1;
 - (d) that no recent contributions have been received by the CCIR concerning the other requested studies;
 - (e) that the present schedule of the Conference does not permit approval of the CCIR consolidated report to the HFBC-93 during scheduled Study Group Meetings,

DECIDES

- 1. that Interim Working Party 10/10 be established with the following terms of reference, to provide updated or new information, as requested by the HFBC-87, concerning:
 - 1.1 out-of-band radiation provisions (Recommendation 328);
 - 1.2 receiver selectivity (Recommendation 332);
 - 1.3 synchronized-frequency transmitter operation (Recommendation 205);
 - 1.4 planning parameters for double-sideband systems, as listed in the Annex to Recommendation No. 510 (HFBC-87);
 - 1.5 values of relative RF protection ratio for the different cases and frequency separations covered in the Annex to Recommendation No. 517 (HFBC-87);

2. that although Recommendation 640 is stating SSB reception specifications, further studies called for in CONSIDERING (a) iv. above should be carried out;
3. that IWP 10/10 should identify such other material that may be deemed useful to the Conference;
4. that IWP 10/10 complete its work on the studies called for in vii. above in due time to include the results in the CCIR report to the HFBC-93;
5. that IWP 10/10 establish appropriate liaison with IWPs 10/1, 6/1 and 6/15 to assure that their reports are prepared in a coordinated editorial fashion for inclusion in the CCIR report to the HFBC-93;
6. that Study Group 1 be invited to appoint a Rapporteur to carry out the necessary liaison work related to DECIDES 1.1 above, with a view to revise the relevant provisions of Recommendation 328, and include the appropriate results in the consolidated CCIR report to HFBC-93;
7. that the IFRB be invited to provide IWP 10/10 with information and software relevant to the studies to be carried out;
8. that IWP 10/10, in preparing its report, should take into account the results of the work of JIWP 10-3-6-8/1;
9. that IWP 10/10 shall prepare a consolidated CCIR report to the HFBC-93, including the results of the above work as well as the reports of IWP 10/1, and Study Groups 6 and 1 with only necessary editorial adjustments;
10. that IWP 10/10 shall complete its work and that its report be submitted to the Chairmen of Study Groups 10, 6 and 1 for approval of the respective sections for which their Study Groups are responsible, in consultation with the Director, CCIR, so that the consolidated CCIR report to the HFBC-93 may be submitted to the Secretary-General as a Conference document at least 10 months before the opening of the Conference;
11. that IWP 10/10 be established without delay to facilitate the completion of its work in the time scale outlined above;
12. that IWP 10/10 shall carry out its work by correspondence, however, it will hold at least one organizational meeting and one additional meeting to finalize its work. Other meetings may be called if considered essential by the Chairman, following consultation with the Chairman of Study Group 10 and the Director, CCIR;
13. that the Chairman of this Interim Working Party will be:

Mr. Warren G. RICHARDS
Deputy Director
Office of International Radio
Communications (CIP/IRC - Rm. 6317)
Vice-Chairman, IWP 10-11/1
US Department of State
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assisted by the following Vice-Chairman:

Mr. G. THIAM
Technical Director
Office de Radiodiffusion Télévision
du Sénégal
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Sénégal

Telephone: 23 63 49
Telex: 51 634 sg

Administrations:

Germany (Federal Republic of)
Austria
Canada
United States of America
Finland
France
India
Italy
Japan
Netherlands
United Kingdom
USSR
Yugoslavia (Socialist Federal Republic of)

International Organizations and Recognized Private Operating Agencies:

Deutsche Welle
EBU
NDR

DECISION 97

SHARING CRITERIA BETWEEN BROADCASTING
AND FIXED AND MOBILE SERVICES IN THE BAND 2-30MHz

(1989)

CCIR Study Groups 10, 3, 6 and 8,

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, *inter alia*, the Resolutions and Recommendations of the WARC HFBC-87 relating to frequency allocations;

(b) that the WARC HFBC-87 (Resolution No. 516, Recommendation Nos. 503, 510, 514, 515, 516 and 517) has invited the CCIR to carry out certain studies in preparation for the Conference;

(c) that the WARC MOB-87 requested the CCIR in Resolution No. 319 to study the technical issues involved in the establishment of sharing criteria between the maritime mobile and fixed services in the 4 000-4 063 kHz and 8 100-8 195 kHz frequency bands including the possibility of using other emissions in the maritime mobile service by ship stations;

(d) that the CCIR has already developed basic concepts and recommended technical criteria for frequency sharing which may facilitate the work of the Conference;

(e) that a CCIR report which will be prepared by a Joint Interim Working Party (JIWP WARC-92) comprised of representatives from Study Groups 1, 2, 3, 4, 5, 6, 8, 9, 10 and 11 should be submitted to administrations [ten months]* in advance of the Conference, i.e. about [March 1991]*;

(f) that the Interim Meeting of Study Groups 10, 3, 6 and 8 for the study period 1990-1994 will be held after March 1991;

(g) the options offered by Resolution 24-6 for CCIR preparatory work for conferences;

*Subject to final decision on JIWP WARC-92.

(h) the combined interest of Study Groups 10, 3 and 8 in HF allocation matters, and the expertise which Study Group 6 can give with regard to ionospheric propagation in the HF bands,

UNANIMOUSLY DECIDES

1. that Joint Interim Working Party 10-3-6-8/1 should be established with the following terms of reference:

1.1 to develop more accurate sharing criteria between broadcasting, fixed, mobile and amateur services in the band 2-30 MHz;

1.2 to report to JIWP WARC-92 by [the end of 1990]*;

2. that the Chairman and composition of JIWP 10-3-6-8-/1 shall be shown in Annex I;

3. that the JIWP should as far as possible work by correspondence, but that it may meet on the proposal of its Chairman, following consultation with the Director of the CCIR.

ANNEX I

1. The following administrations and international organizations, and recognized private operating agencies have indicated that they wish to participate in the work of JIWP 10-3-6-8/1:

Administrations:

Germany (Federal Republic of)
Saudi Arabia
Australia
Brazil
Canada
China (People's Republic of)
United States of America
France
India
Italy
Japan
Netherlands
Norway
United Kingdom
USSR
Yugoslavia

*Subject on final decision on JIWP WARC-92.

International Organizations and Recognized Private Operating Agencies:

Deutsche Welle
EBU
IATA
NDR
ICAO

2. The Chairman of the Joint Interim Working Party will be:

Mr. J. FINNIE
Group Leader: Sound Radio
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The Vice-Chairmen of the JIWP will be:

Study Group 3: Mr. N. YAMASAKI
Vice-Chairman S.G.3, CCIR
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Satellite and Radio Communications Department
Transmission Systems Engineering Headquarter
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Study Group 6: Mr. D.G. COLE
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Study Group 8: Mr. W.J. HELWIG
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DECISION 98**

LOW BIT-RATE DIGITAL AUDIO CODING SYSTEMS

(1989)

The CCIR,

CONSIDERING

- (a) that recent progress in digital audio coding techniques allows for a very considerable degree of bit-rate reduction to be achieved while maintaining high quality;
- (b) that information gathered by the CCIR on various aspects of digital audio coding systems is contained in Reports 799, 953, 1068, 1070, 1095, 1236 and 1199;
- (c) that applications exist for low bit-rate audio coding systems for digital audio broadcasting, television sound (especially HDTV and EDTV) and for storage of signals;
- (d) that the studies concerning digital techniques in sound are in progress both in Study Groups 10 and the CMTT, being matters related to studio installations and broadcasting dealt with in Study Group 10 and matters related to transmission dealt with by the CMTT,

DECIDES

- 1. that Joint Interim Working Party 10-CMTT/1 be established within the general terms of reference of Study Groups 10 and CMTT:
 - 1.1 to define the sound quality and other requirements for each application, including:
 - studio installation including recording,
 - transmission,
 - broadcasting from terrestrial transmitters and from satellites;
 - 1.2 to define bit-rate reduction techniques that would satisfy the quality level and other requirements for each of the applications listed in 1.1 while making efficient use of storage or transmission media;

*This Decision should be brought to the attention of the IEC and ISO.

1.3 to combine in a single CCIR Report the results of studies on digital audio coding systems which are now dispersed in a number of Reports;

1.4 to present a draft Recommendation by the 1991 Interim Meeting;

2. that JIWP 10-CMTT/1 should liaise, where appropriate, with JIWP 10-11/6 (former IWP 11/4), IWP 10/12 (multi-channel sound systems) and JIWP 10-11/3;

3. that JIWP 10-CMTT/1 should, as far as possible, work by correspondence, however, it may meet upon the proposal of its Chairman following consultation with the Chairmen of Study Groups 10 and the CMTT and the Director of the CCIR;

4. that the Chairman and the composition of the JIWP 10-CMTT/1 shall be as shown in Annex I.

ANNEX I

The following Administrations and International Organizations have indicated that they wish to participate in the work of JIWP 10-CMTT/1:

Administrations:

Germany (Federal Republic of)
Australia
Canada
Denmark
United States of America
France
Italy
Japan
German Democratic Republic
United Kingdom
Sweden
Switzerland
USSR

International organizations:

EBU
OIRT

Chairman of JIWP 10-CMTT/1:

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Vice-Chairmen of JIWP 10-CMTT/1:

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Telex: 52 15 605

DECISION 99

SYSTEM DESIGN FOR HF BROADCASTING

(1989)

CCIR Study Group 10,

CONSIDERING

- (a) Question 44/10 —→ and Study Programmes 44F/10 and 44N/10;
- (b) the specific needs of the broadcasting service to provide satisfactory coverage throughout an intended service area;
- (c) the lack of adequate information on HF system design in CCIR texts,
- (d) the necessity to take into account both wanted and interfering signals,

DECIDES

1. that an Interim Working Party, 10/11, be set up within the general terms of reference of Study Group 10;

1.1 to prepare a Report for Study Group 10 on system design for HF broadcasting including:

- methods of satisfying the wide range of differing requirements;
- reduction of interference levels as a means for improving the efficiency of spectrum utilization;

1.2 that the Report mentioned in 1.1 should utilize the propagation prediction method given in Report 894 together with any relevant planning parameters given in Recommendations and Reports of Study Group 10;

2. that IWP 10/11 should report on its progress to the next Interim Meeting of Study Group 10;



3. that the Report mentioned in 1.1 should be available for consideration by the next Final Meeting of Study Group 10.

ANNEX I

1. The following Administrations, International Organizations and Recognized Private Operating Agencies have indicated that they wish to participate in the work of IWP 10/11:

Administrations:

Germany (Federal Republic of)
Austria
United States of America
India
Italy
France
Japan
Netherlands
United Kingdom
USSR
Yugoslavia

International Organizations and Recognized Private Operating Agencies:

EBU
Deutsche Welle
OIRT

2. The Chairman of the Interim Working Party will be:

Mr. K. HUNT
Technical Centre
European Broadcasting Union
Ancienne Route 17A
Case Postale 67
CH-1218 Grand Saconnex
Geneva, Switzerland

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