THE INTERNATIONAL TELEGRAPH AND TELEPHONE
CONSULTATIVE COMMITTEE
(C.C.I.T.T.)

IIIrd PLENARY ASSEMBLY

GENEVA, 1964

RED BOOK

VOLUME V bis

Telephone transmission quality, local lines and telephone sets

SERIES P RECOMMENDATIONS

STUDY GROUP XII QUESTIONS

Published by the
INTERNATIONAL TELECOMMUNICATION UNION
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CONTENTS OF THE C.C.I.T.T. BOOKS STILL APPLICABLE FOLLOWING THE THIRD PLENARY ASSEMBLY (1964)

A. RED BOOK

Volume V — Recommendations (Series P) and Questions (Study Group XII) relative to telephone transmission performance and apparatus.

Volume V bis — Additions and amendments to Volume V following the IIIrd Plenary Assembly.

B. BLUE BOOK

Volume I — Minutes and reports of the IIIrd Plenary Assembly of the C.C.I.T.T.
— Resolutions and Opinions issued by the C.C.I.T.T.
— List of Study Groups and Working Groups for the period 1964-1968.
— Summary Table of Questions under study in 1964-1968.
— Recommendations (Series A) relative to the organization of the work of the C.C.I.T.T.
— Recommendations (Series B) and Questions (Study Group VII) relative to means of expression.

Volume II — Recommendations (Series D) and Questions (Study Group III) relative to the lease of circuits.
— Recommendations (Series E) and Questions (Study Group II) relative to telephone operation and tariffs.
— Recommendations (Series F) and Questions (Study Group I) relative to telegraph operation and tariffs.

Volume III — Recommendations (Series G, H and J) and Questions (Study Groups XV, XVI and C) relative to line transmission.

Volume IV — Recommendations (Series M and N) and Questions (Study Group IV) relative to transmission maintenance of international lines, circuits and chains of circuits.

Volume VI — Recommendations (Series Q) and Questions (Study Groups XI, XIII and B) relative to telephone signalling and switching.

Volume VII — Recommendations (Series R, S, T, U) and Questions (Study Groups VIII, IX, X and XIV) relative to telegraph technique.

Volume VIII — Recommendations (Series V) and Questions (Study Group A) relative to data transmission.

Volume IX — Recommendations (Series K) and Questions (Study Group V) relative to protection against disturbances.
— Recommendations (Series L) and Questions (Study Group VI) relative to the protection of cable sheaths and poles.

Each volume contains extracts from contributions received dealing with the subject of the volume concerned and the interest of which is such as to warrant publication.

NOTICE

The present publication modifies and completes Volume V of the *Red Book* (New Delhi, 1960). The following recapitulative table indicates the texts in Volume V [1] which remain in force and, at the same time, serves as the table of contents of Volume V *bis*.

It has been indicated (immediately after the titles of Recommendations or Annexes) whether the texts are new ones approved by the Plenary Assembly of Geneva, 1964, or are texts modified at the same period. Texts without any such an indication date from at least as far back as the Plenary Assembly of New Delhi, 1960, when Volume V was divided into numbered recommendations; however, certain of these texts may be even older.

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[1] Subject to the corrections listed in the *Corrigendum to Volume V,* to be found at the end of this publication.

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RECAPITULATIVE TABLE OF CONTENTS OF VOLUMES V AND V bis OF THE C.C.I.T.T. RED BOOK

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SECTION 8

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[1] This bibliography was not kept up to date after 1960.

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PART I

SERIES P RECOMMENDATIONS

Quality of telephone transmission; local telephone installations and networks

SECTION 1

GENERAL RECOMMENDATIONS ON THE TRANSMISSION QUALITY FOR AN ENTIRE INTERNATIONAL TELEPHONE CONNECTION

RECOMMENDATION P.11 (modified in Geneva, 1964)[1]

REFERENCE EQUIVALENTS

A. Definition

National sending and receiving reference equivalents should be those calculated at the virtual switching points of the international circuit; that is to say, at points a and b of Figure 1 (for a country of average size).

The virtual switching points of an international four-wire telephone circuit are fixed by convention at points of the circuit where the nominal relative levels at the reference frequency are:

-3.5 dbr or -4.0 dNr, sending

-4.0 dbr (-4.6 dNr), receiving

The nominal transmission loss of this circuit at the reference frequency between virtual switching points is therefore 0.5 db or 6 cN.

Note. — The relative level at a given point of a four-wire circuit is determined by reference to the specifications of the transmission system on which the circuit is set up, the performance of the system (noise, crosstalk, limiting, linearity, etc.) being evaluated at a point of zero relative level. For example, the nominal mean power of signals during the busy hour, at a point of zero relative level, is indicated in Section 1 of Recommendation G.223.

[1] Series G recommendations cited in the present recommendation will all appear in Volume III of the C.C.I.T.T. *Blue Book* (Geneva, 1964).

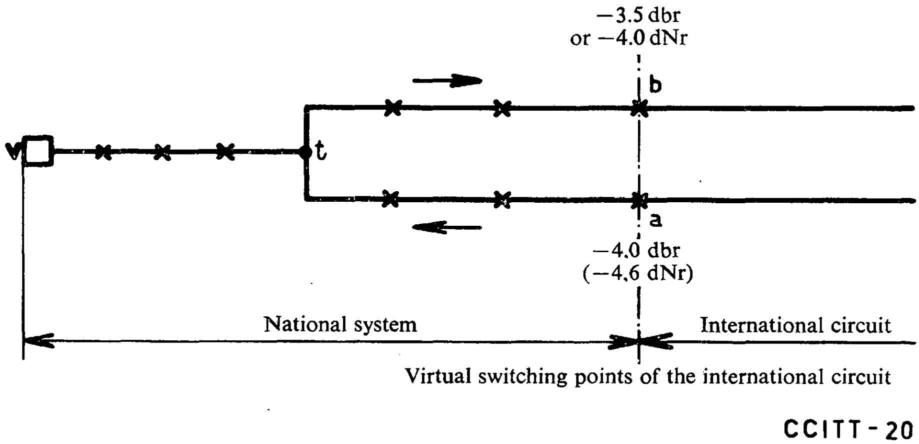


FIGURE 1. — **Definition of the virtual switching points**

B. MAXIMUM REFERENCE EQUIVALENTS

In the new transmission plan, the total reference equivalent between two subscribers is not strictly limited. Its maximum will be derived from all the various recommendations indicated below.

a) *Maximum nominal sending and receiving reference equivalents*

Provisionally, national sending and receiving systems used to set up 95 % of actual outgoing or incoming calls in an average-size country (see Recommendation G.101 B-b), should individually meet both the following requirements:

— the nominal reference equivalent of the sending system between a subscriber and the first international circuit should not exceed 20.8 db (24 dN);

— the nominal reference equivalent of the receiving system between the same two points should not exceed 12.2 db (14 dN).

In a large country, these limits shall be, respectively: 21.3 db (24.6 dN) and 12.7 db (14.6 dN) if a fourth national circuit is part of the four-wire chain,

or 21.8 db (25.2 dN) and 13.2 db (15.2 dN) if five national circuits form part of the four-wire chain.

In Figures 2 and 3, the numbers in rectangles are figures recommended by the C.C.I.T.T. The others are given only as examples of possible arrangements, subject to Recommendation G.122.

Note 1. — It may well be found impossible to meet the above limits straightaway in some existing networks constructed in accordance with old recommendations issued by the C.C.I.F. (Volume IV of the C.C.I.F. *Green Book*—first part, Section 1.1), but an attempt should be made to abide by them when the networks are reorganized or when telephone sets of a new type are introduced.

Note 2. — The 95% limit is provisional, and it is very desirable to use a higher percentage when planning new networks.

Note 3. — The nominal reference equivalents given for national systems include the systematic differences between the performances of the subscriber's set at the sending and receiving ends and their nominal values; however, they do not include the variations of loss with time in the various parts of the national system, nor fortuitous variations of the reference equivalents assessed by subjective methods.

b) *Practical limits of the reference equivalent between two operators or one operator and one subscriber*

These limits are being studied for the new transmission plan; the values hitherto recommended are given in Volume V of the *Red Book* on page 10 in Note 1 and in applying them Note 2 of the same text should be borne in mind.

The values for the complete connections shown in the table on page 9 of Volume V of the *Red Book* are not applicable to the new transmission plan.

c) *Nominal equivalent of the international chain*

The nominal loss between virtual switching points, for each international circuit, should in principle be 0.5 db or 6 cN at 1000 c/s or 800 c/s. However, some circuits can be operated with higher losses. (See Recommendation G.131 B-a.)

As far as transmission is concerned, there is no strict limit on the number of international circuits which may be interconnected in tandem, provided each of them has a nominal loss, between the virtual switching points, of 0.5 db or 6 cN in the transit condition and provided there is four-wire interconnection. Naturally, the fewer the number of interconnected circuits the better the transmission performance is likely to be. (See Recommendation G.101 C.)

d) *Nominal reference equivalent of a complete connection*

The C.C.I.T.T. Laboratory has ascertained the attenuation to be inserted between a local sending and a local receiving system to obtain an over-all reference equivalent of 36 db. In this test one, two or three A.R.A.E.N. 300-3400 c/s filters, identical with that which forms part of the S.R.A.E.N. (*Red Book*, Volume V, page 67) were inserted into the line connecting the two local commercial systems.

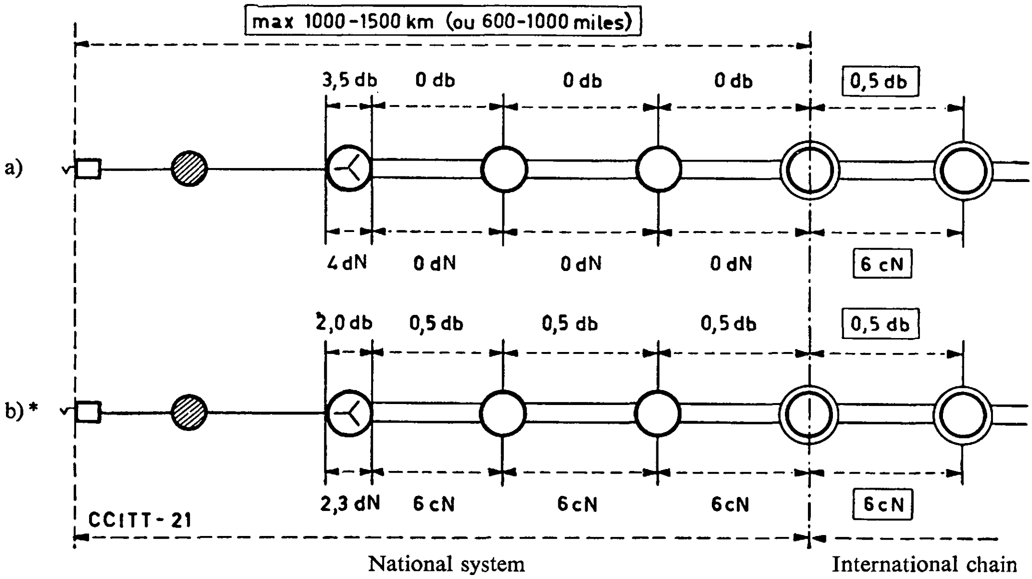
The frequency-loss characteristic of each filter meets the requirements of Graph No. 2-B of Recommendation G.232; the set of three filters in series conforms to Graph No. 1 in Recommendation G. 132 (Figure 12 of Volume III of the *Blue Book*), showing the objective for a chain of 12 carrier circuits in tandem.

The sending and receiving reference equivalents of the local systems were also determined, using the usual method.

A preliminary examination of these tests shows that the reference equivalent, corresponding to a complete connection, appears to be satisfactorily represented by the sum of the sending and receiving reference equivalents of the local systems measured separately and of the 800-c/s equivalent of the chain of long-distance circuits.

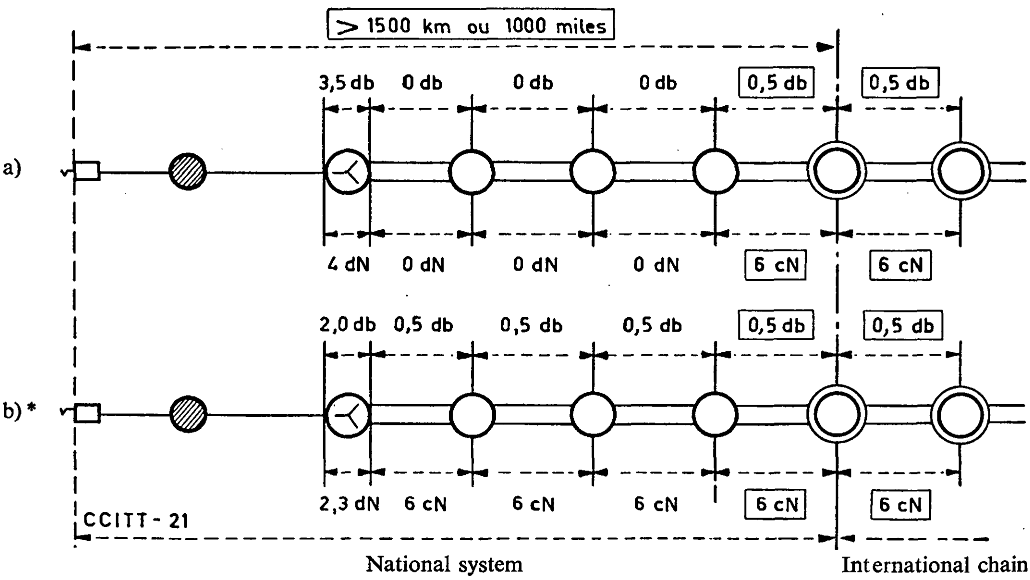
e) *Variations in time and effect of circuit noise*

The nominal reference equivalents given for national systems include the systematic differences between the performances of the subscriber's set at the sending and receiving ends and their nominal values; however, they do not include the variations of loss with time in the various parts of the national system, nor fortuitous variations of the reference equivalents assessed by subjective methods. Recommendation G.151 C sets forth the objectives



Maximum reference equivalents for 95% of calls: Sending **20.8 db (24 dN)**; Receiving **12.2 db (14 dN)**

Figure 2. — **Distribution of equivalents for an international call, in a country of average size**



Maximum reference equivalents for 95% of calls: Sending **21.3 db (24.6 dN)**; Receiving **12.7 db (14.6 dN)**

Figure 3. — **Distribution of equivalents for an international call, in a large country**

* The division of nominal transmission losses is theoretical and can readily be achieved by means of pad-switching, for example.

recommended by the C.C.I.T.T. in connection with variations in transmission losses of international circuits and national extension circuits in relation to the nominal values.

According to the results of measurements supplied by one Administration, the reference equivalent of its transmitting system rises by an average of 7 cN per annum, a regular increase due to ageing of the microphone. This point is being studied by the C.C.I.T.T. (Question 1/XII, part b).

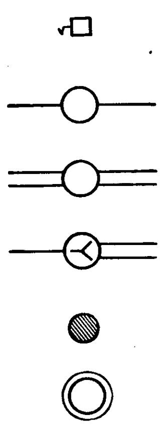
Annex A (Part 2 of this Volume) gives information on the statistical variations of reference equivalents.

Annex B (Part 2 of this Volume) mentions the effect on transmission quality of these variations in the equivalent and of the recommended limits for circuit noise.

C. Minimum reference equivalents

Administrations must take care not to overload the international transmission systems ifthey reduce the attenuation in their national trunk network. This aspect of the problem must be studied separately before any precise recommendation can be prepared. In some countries a very low sending reference equivalent may occur if unregulated telephone sets are used. Nor should the speech power applied to the international circuits by operators' sets be excessive.

Legend for Figures 2 and 3



Subscriber's set

Two-wire switching exchange

Four-wire switching exchange

Exchange with terminating unit [1]

Local exchange

International exchange

[1] A switchable pad may also be introduced at that point to compensate for losses on the two-wire side, provided that the limits given in Recommendation G.122 A for stability and attenuation are respected.

D. Assessment of the reference equivalents in a national system

Administrations and private operating Agencies can use various methods to see that the limits on reference equivalents are not exceeded. Thus, for example, artificial networks can be set up representing the main combinations of subscriber's commercial telephone apparatus, subscribers' lines, junction lines and local and trunk exchange equipments, each of these networks representing a complete national sending or receiving system, which would be compared, in a voice-ear test, with the New Master System for the determination of reference equivalents (N.O.S.F.E.R.) or with a working standard system already compared with N.O.S.F.E.R. or S.F.E.R.T.

Another way would be merely to measure the reference equivalent of the telephone apparatus in certain specific circumstances. To this reference equivalent would be added the systematic difference between the actual sensitivity of the particular subscriber's telephone apparatus and the nominal value of this sensitivity, the image attenuation (calculated or measured at 800 c/s or at another suitable frequency) of the subscribers' lines, junction lines and toll or trunk circuits connecting this apparatus to the international centre, and the composite attenuation (measured or calculated at 800 c/s for a non-reactive resistance of 600 ohms) of the exchange equipments used in the connection between this apparatus and the international centre (including the equipment of the exchange serving the subscriber and that of the international centre).

In any event, however, these calculations ought to be checked by a voice-ear test on the artificial networks representing the most typical complete national sending and receiving systems.

Note 1. — The C.C.I.T.T. is considering how the reference equivalent of a subscriber's line should be calculated when it has to be ascertained independently of the over-all reference equivalent of the local system (subscriber's line plus subscriber's set). This is necessary when the second of the above procedures is used.

Note 2. — The N.O.S.F.E.R. has replaced the Master Reference System (S.F.E.R.T.), used in the old C.C.I.T.T. Laboratory before transfer to the new I.T.U. building. It, and other reference systems, are described in Recommendation P.42, *Red Book*, Volume V.

RECOMMENDATION P.12 (modified in Geneva, 1964)

ARTICULATION REFERENCE EQUIVALENT (A.E.N.)

The transmission quality of international telephone calls will always be satisfactory if the reference equivalent limits fixed in Recommendation P.11 are respected together with the limits fixed in Volume III of the *Blue Book* for noise, crosstalk, etc., and if, in addition, use is made of telephone sets of modern types which have satisfactory sensitivity/frequency characteristics and efficient anti-sidetone arrangements.

Administrations wishing to make a thorough study of the transmission quality of their national sending and receiving systems could be guided by the A.E.N. method described below.

A. *Definition of the articulation reference equivalent (A.E.N.)*

Articulation reference equivalent (A.E.N.) (G.B.) [Equivalent articulation loss (Am.)—Affaiblissement equivalent pour la netteté (A.E.N.) (F)]

If articulation tests are made under specified conditions alternately on a telephone system to be tested and on the " reference system for the determination of A.E.N." (S.R.A.E.N.) with different values of line attenuation, up to the point where values of articulation on both systems are substantially reduced, then the results of these tests may be recorded in the form of curves showing the variation of sound articulation against attenuation. The value A_1 of the attenuation of the system under test, and the value A_2 of the attenuation of the S.R.A.E.N. at a fixed value 80% sound articulation can then be determined.

$(A_2 - A_1)$ is by definition equal to the *articulation reference equivalent* (A.E.N.).

B. *Calculation of the nominal articulation reference equivalent of a national sending or receiving system* [1]

The nominal A.E.N. of a national sending or receiving system is the sum of the following quantities:

1. The nominal A.E.N. (average value in service) of the local system;
2. The nominal A.E.N. of the connection between the local exchange and the international exchange (average value in service).

The articulation reference equivalent, in service, of the connection between the local exchange and the international exchange is equal to the sum of the following numbers [2]:

— the equivalent of the trunk circuits between the last trunk exchange and the international exchange, measured at 800 c/s, increased by the transmission impairment due to bandwidth limitation (see Recommendation P.13 below) when these circuits have an attenuation/frequency distortion greater than that which is allowed in the recommendations of the C.C.I.T.T.;

— the average articulation reference equivalent of the toll circuits given by the following expression:

$$i = K \times L$$

where

i = average A.E.N. in decibels or nepers,

L = length of the toll circuit in kilometres,

K = coefficient which depends on the type of toll circuit considered, in decibels per kilometre or nepers per kilometre (see the Remark below),

— the mean A.E.N. of each intermediate exchange. The A.E.N. resulting from the insertion of a circuit element which, in accordance with the recommendations of the C.C.I.T.T., effectively transmits frequencies from 300 to 3400 c/s, can be calculated by taking the arithmetic mean of the four values of insertion loss (or gain) of the element considered measured at 500, 1000, 2000 and 3000 c/s and expressed in decibels or nepers. Until there are more accurate values of this rating available, as will result from any measurements that Administrations may make in this respect, a provisional value of 1 db or 1 dN for each exchange introduced into the connection will be used.

Note 1. — Circuit noise which is within the limits fixed by C.C.I.T.T. recommendations is not taken into account.

Note 2. — The " composite attenuation " of the lines connecting the international exchanges to the local exchanges should be such that the reference equivalent of the national sending system and the reference equivalent of the national receiving system remain within the limits considered compatible with good telephone transmission.

[1] It is agreed for international purposes that the result obtained by this calculation B represents the magnitude of the articulation reference equivalent for a national transmitting or receiving system. This number is called the nominal articulation reference equivalent, to distinguish it from the articulation reference equivalent measured on the complete national sending or receiving system.

[2] Articulation tests have shown that the A.E.N. can be calculated approximately for such a link, in the manner shown above.

C. *Determination of A.E.N.*

The reference system for the determination of the A.E.N. (S.R.A.E.N.) and the method of determining the A.E.N. of commercial telephone systems at the C.C.I.T.T. Laboratory are described in Recommendations P.44 and P.45.

D. *Nominal A.E.N. values for the national sending system and the national receiving system*

By way of information, it is pointed out that Administrations using the A.E.N. method consider it very desirable that national sending and receiving systems used to set up 90% of actual outgoing or incoming calls should individually meet both of the following requirements:

— the nominal A.E.N. of the national sending system should not exceed 24 db or 2.8 N;

— the nominal A.E.N. of the national receiving system should not exceed 18 db or 2.1 N.

Note 1. — The values (24 db and 18 db) given above for the national sending and receiving systems refer to the two-wire terminals of the international circuit, whereas the reference equivalents recommended in Recommendation P.11 refer to the virtual switching points of the international circuit. These A.E.N. values do not include the probable variations, as a function of time, of the equivalents of the trunk circuits which form part of the national system.

Note 2. — These values apply to the A.E.N. values deduced from the values measured for a local system at the C.C.I.T.T. Laboratory, as described in Recommendation P.45 with, in particular, 60 db or 7 N room noise at the receiving end for commercial systems and an electrical background noise (having a psophometric e.m.f. of 2 millivolts) injected into the input of the receiving system of the S.R.A.E.N.

Note 3. — The A.E.N. method does not make allowance for the effect of sidetone on subscribers' speech power.

Administrations or private operating Agencies wishing to prepare transmission plans for their national network, on the basis of " transmission performance rating ", will find in Annex 1, Part II of Volume V of the *Red Book*, information on the corrections to be made to the values of A.E.N. to allow for sidetone at the sending end.

REMARK

Average A.E.N. of toll circuits

A toll circuit may be considered as a quadripole inserted between the impedance of the first trunk circuit, seen through the switchboard (or switches), and the impedance of the local system (feeding bridge + subscriber's line + subscriber's apparatus).

For a given frequency, the loss introduced by such a circuit is represented by its " composite attenuation " [1] which is the sum of the image attenuation of the circuit itself and of the other terms representing all the effects due to reflections introduced by mismatch between the image impedance of the circuit and the impedances of the terminations defined above.

According to tests made by the British Administration, the A.E.N. due to the reflections can be represented by the arithmetic mean of the reflection losses measured at frequencies of 500, 1000, 2000 and 3000 c/s.

The transmission performance rating of an unloaded line is measured by its image attenuation at 1500 c/s and this is approximately equal to the arithmetic mean of the image attenuations at the four frequencies quoted above [2].

[1] In practice, instead of using the composite attenuation, insertion loss may be used.

[2] The attenuation of a non-loaded cable circuit is proportional to the square root of the frequency. The frequencies 500, 1000, 2000, 3000 c/s are in the ratio 1, 2, 4, 6, and their square roots in the ratio 1, 1.41, 2, 2.45 of which the arithmetic mean is 1.72, i.e. almost the square root of 3; therefore this mean corresponds to a frequency of $3 \times 500 = 1500$ c/s.

Therefore, the A.E.N. of the toll circuit may be obtained directly, taking account not only of the effect due to the image attenuation but also of the effect of reflections, by taking the arithmetic mean of the composite attenuations measured at the four frequencies referred to above.

As the impedance of the local systems varies widely, it is not possible to define a single value for the average A.E.N. for a toll circuit, but only an average value obtained by taking the arithmetic mean of several values of the A.E.N., measured under several terminal conditions (see " C.C.I.F.—1952/1954—4th S.G.—Document No. 32 ", Annex).

For each type of toll circuit (defined by the electrical characteristics of the circuit), the average A.E.N. is proportional to the length of the circuit, the ratio being *easily determined* when three or four values of the A.E.N. are known. It is given by the formula:

$$i = K \times L \tag{1}$$

where

i = average A.E.N. in decibels or nepers;
L = length of toll circuit in kilometres;
K = coefficient, which depends on the type of toll circuit considered, in decibels per kilometre or in nepers per kilometre.

To determine, once and for all, the different values of the coefficient K, the composite attenuation of three or four different lengths of each type of toll circuit used in a particular network (if necessary using artificial lines) can be measured; for this purpose the technique described in Document 32 referred to above (see also Annex 2 to Question No. 10 in Volume I *ter* of the *Yellow Book* of the C.C.I.F., page 400), and one of the methods of measuring of the composite attenuation described in Supplement No. 1, Part III of Volume IV of the *Blue Book*, can be used.

From equation (1) the value of the average A.E.N. may be calculated for any length and any type of toll circuit in the national network considered.

RECOMMENDATION P.13 (modified in Geneva, 1964)

TRANSMISSION IMPAIRMENTS AND NOISE

A. Transmission impairment

a) *due to bandwidth limitation (cut-off impairment) effectively transmitted by the trunk circuit*

Observations have been made in the United States of America of the repetitions during conversations and articulation measurements have been made in various national laboratories as well as in the C.C.I.T.T. Laboratory. The results obtained permit the mean curve given in Figure 4 to be plotted showing the impairment due to cut-off frequency by a trunk circuit.

The equation to this curve is $y = 2\,(3.7 - f)^2$, where y is the transmission impairment (in decibels) due to the limitation of the frequency bandwidth effectively transmitted, and f is the frequency (in kc/s) for which the loss of the circuit exceeds its loss at 1000 c/s by 10 decibels.

Note. — The cut-off impairment for a chain of national trunk circuits or for a connection between two international exchanges made up of several international circuits is not obtained by adding the individual impairments. It is necessary to consider the impairment for the circuit which transmits effectively the narrowest band of frequencies.

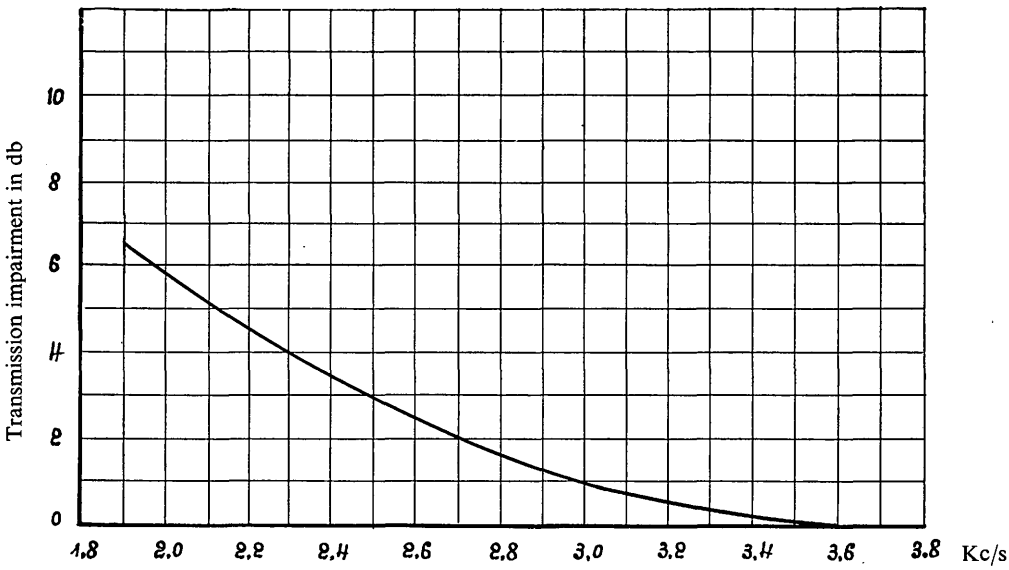


FIGURE 4. — **Transmission impairment due to bandwidth limitation (cut-offimpairment)**

Note. — The frequencies shown on the abscissa are the maximum frequencies effectively transmitted according to the definition adopted in the United States of America, i.e. those for which the attenuation is greater by 10 db than the attenuation at 1000 c/s.

b) *due to room noise*

The method of measuring A.E.N. takes account of 60 db of room noise [1] (Hoth spectrum) at the receiving end; information regarding the method of evaluating the " impairment due to room noise" used in the United States of America is given in Annex 3, Part II of Volume V of the *Red Book*.

B. EFFECT OF CIRCUIT NOISE

The C.C.I.T.T. recommends that the mean value, expressed in decibels and taken over a large number of world-wide connections, (each including six international circuits), of the distribution of one-minute mean values of noise power of the connections, should not exceed —43 dbm0p or —5 Nm0p referred to the input of the first circuit in the chain of international circuits.

Annexes B, C and D of Part II of this Volume describe how the C.C.I.T.T. made allowance for the effect of noise on transmission performance in planning the international network.

[1] The power density spectrum of the room noise used in A.E.N. measurements is given in Figure 5. The following articles give information on room noise at locations where commercial telephone sets are located:

1. A Room Noise Survey of Business Subscribers' Telephone Locations. *B.P.O. Research Report,* No. 8990—1935.
2. Room Noise at Telephone Locations. D. F. SEACORD, *Electrical Engineering,* Part 1, *58,* 255, 1939.
3. Room Noise Spectra at Subscribers' Telephone Locations. D. F. HOTH, *Journal of the Acoustical Society of America, 12,* 499, 1941.

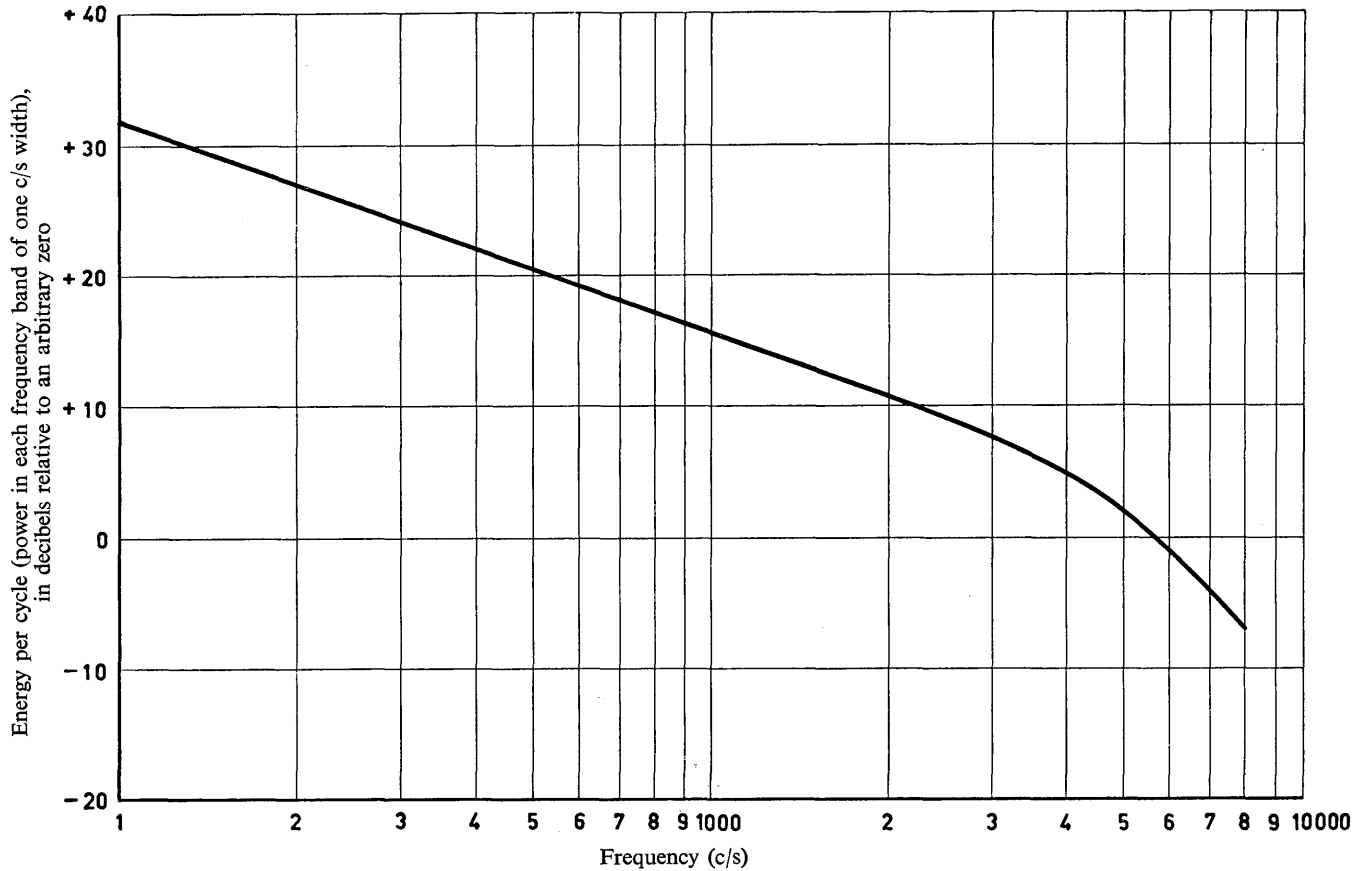


FIGURE 5. — **Power density spectrum of the room noise produced in the listening cabinet of the C.C.I.T.T. Laboratory**

This curve conforms to the mean power density spectrum of noise observed in locations where telephone sets are situated, published by Hoth.

The procedure does not make explicit use of any transmission impairment due to circuit noise.

By way of information, the method used in the United States to fix objectives for circuit noise is described by D. A. Lewinski, in an article entitled: A New Objective for Message Circuit Noise (*Bell System Technical Journal*, Volume XLIII, pages 719-740, No. 2, March 1964).

Note. — Annex 2 to the second part of Volume V of the *Red Book* is out-of-date and should be deleted.

RECOMMENDATION P.14 (Geneva, 1964)

MEAN ONE-WAY PROPAGATION TIME

A. Limits for a connection

It is necessary in an international telephone connection to limit the propagation time between two subscribers. Recent tests have shown that international connections probably will not cause adverse subscriber reaction due to the combined effect of delay and echo suppressors if the mean one-way propagation time [1] is increased from near zero to about 150 ms. As the propagation time is increased beyond 150 ms, subscriber difficulties increase, and the rate of increase of difficulty rises up to and including the maximum one-way propagation time tested, namely 400 ms.

The C.C.I.T.T. therefore provisionally recommends the following limitations on mean one-way propagation times when echo sources exist and echo suppressors are used:

a) Acceptable without reservation, 0 to 150 ms;

b) Provisionally acceptable, 150 to 400 ms. In this range connections may be permitted, in particular when compensating advantages are obtained;

c) Provisionally unacceptable, 400 ms and higher. Connections with these delays should not be used except under the most exceptional circumstances.

Until such time as additional significant information permits Administrations to make a firmer determination of acceptable delay limits, they should proceed with caution and take full account of the data in Annexes E and F (Part II of this volume) in selecting, from alternatives, plans involving delays in range b) above.

[1] Mean of the times in the two directions of transmission.

B. Values for circuits

In the establishment of the general interconnection plan within these limits the one-way propagation time of both the national extension circuits and the international circuits must be taken into account.

a) *National extension circuits*

The main arteries of the national network should consist of high-velocity propagation lines. In these conditions, the propagation time between the international centre and the subscriber farthest away from it in the national network will probably not exceed:

$12 + (0.0064 \times \text{distance in miles})$ ms

or $12 + (0.004 \times \text{distance in kilometres})$ ms.

Here the factor 0.0064 (or 0.004) is based on the assumption that national trunk circuits will be routed over high-velocity plant (155 miles/ms or 250 km/ms). The 12-ms constant term makes allowance for terminal equipment and for the probable presence in the national network of a certain quantity of loaded cables (e.g. three pairs of channel translating equipments plus about 100 miles (160 km) of H 88/36 loaded cables. For an average-sized country the one-way propagation time will be less than 18 ms.

b) *International circuits*

International circuits will use high-velocity transmission systems; the one-way propagation times, or velocity, that should be assumed for planning purposes are:

1. *Terrestrial lines* (including submarine cables)

   100 miles/ms (160 km/ms).

This propagation velocity includes an allowance for terminal and intermediate multiplex equipment likely to be associated with a transmission line.

2. *Satellite links*

The mean one-way propagation times between earth stations for two illustrative single-hop communication satellite systems are:

Satellite at 8700 miles or 14 000 km altitude	110 ms
Satellite at 22 500 miles or 36 000 km altitude	260 ms

The one-way propagation times do not include any allowance for the distance from the earth stations to locations where the satellite circuits can either be extended on other international transmission systems or switched to other national or international circuits. These additional times should be taken into account for planning purposes. The practical distances between earth stations depend not only on the altitude of the satellites but also on the orbits and positions of the satellites relative to the earth stations. Exact account should be taken of these parameters in particular applications.

Note. — The propagation time referred to above is the group delay as defined in the I.T.U. *List of Definitions of Essential Telecommunication Terms* (Definition No. 04-17); the numerical values are calculated at a frequency of about 800 c/s.

RECOMMENDATION P.15 (modified in Geneva, 1964)

GROUP DELAY DISTORTION

The permissible differences for a world-wide chain of 12 circuits, each on a single group connection, between the minimum group delay (throughout the transmitted frequency band) and the group delay at the lower and upper limits of this frequency band are indicated in the table below:

	Lower limit of frequency band	Upper limit of frequency band
	ms	ms
International chain	30	15
Each of the national four-wire extensions	15	7.5
On the whole four-wire chain	60	30

Typical group delays at various frequencies for a chain of 12 circuits in tandem are given in Recommendation G.232 C (*Blue Book*, Volume III).

SECTION 2

JUNCTION CIRCUITS AND TOLL CIRCUITS, LOCAL EXCHANGES AND MANUAL TRUNK EXCHANGES

RECOMMENDATION P.21

APPLICATION OF C.C.I.T.T. RECOMMENDATIONS ON TELEPHONE PERFORMANCE TO NATIONAL NETWORKS [1]

The different parts of a national network likely to be used for an international connection should meet the following general recommendations:

1. The national sending and receiving systems should satisfy the limits recommended above, in:

— Recommendation P.11 as regards reference equivalent;

— Recommendation P.15 as regards group delay distortion.

They should also satisfy the limits recommended in the following Recommendations of Section I, Volume III of the *Blue Book*:

— Recommendation G.122 as regards balance return loss and transmission loss;

— Recommendation G.123 for circuit noise.

Note. — Reference should also be made to Recommendations P.12 and P.13.

2. Long-distance trunk circuits forming part of the main arteries of the national network should be high-velocity propagation circuits which enable the limits fixed in Recommendation P.14 above to be respected. They should conform to Recommendations G.151 and G.152 of Volume III of the *Blue Book*.

Loaded-cable circuits should conform to Recommendation G.124 and carrier systems over very short distances to Recommendation G.125.

3. National trunk circuits should have characteristics enabling them to conform to Recommendations G.131, G.132 and G.134 in Section 1 of Volume III of the *Blue Book* as regards the other characteristics of the four-wire chain constituted by the international telephone circuits and the national trunk extension circuits.

The C.C.I.T.T. is at present studying (in its Study Group XII) the particular conditions which these junction circuits and toll circuits should satisfy in addition to the general conditions recommended above.

The following Remark contains some information about methods which may be applied to the junction and trunk networks of a country in order to ensure satisfactory quality for national calls,

[1] Terminology and references brought up to date by the Secretariat after the 1964 Plenary Assembly.

it being understood that the recommendation of the C.C.I.T.T. relating to reference equivalent (Recommendation P.11 above) is satisfied for international calls. Information on methods of planning national networks is also given for information purposes in Chapter V, "Transmission", of the *Manual on National Telephone Networks for the Automatic Service.*

REMARK

Information about the organization of a national telephone network

a) *General organization and nomenclature* [1]

In order to avoid using exclusively national terms for the various classes of exchange (such as regional centre, district office, nodal exchange, etc.) the exchanges will be designated in this Remark as follows in ascending order of rank:

LOCAL EXCHANGE: exchange to which subscribers are connected.

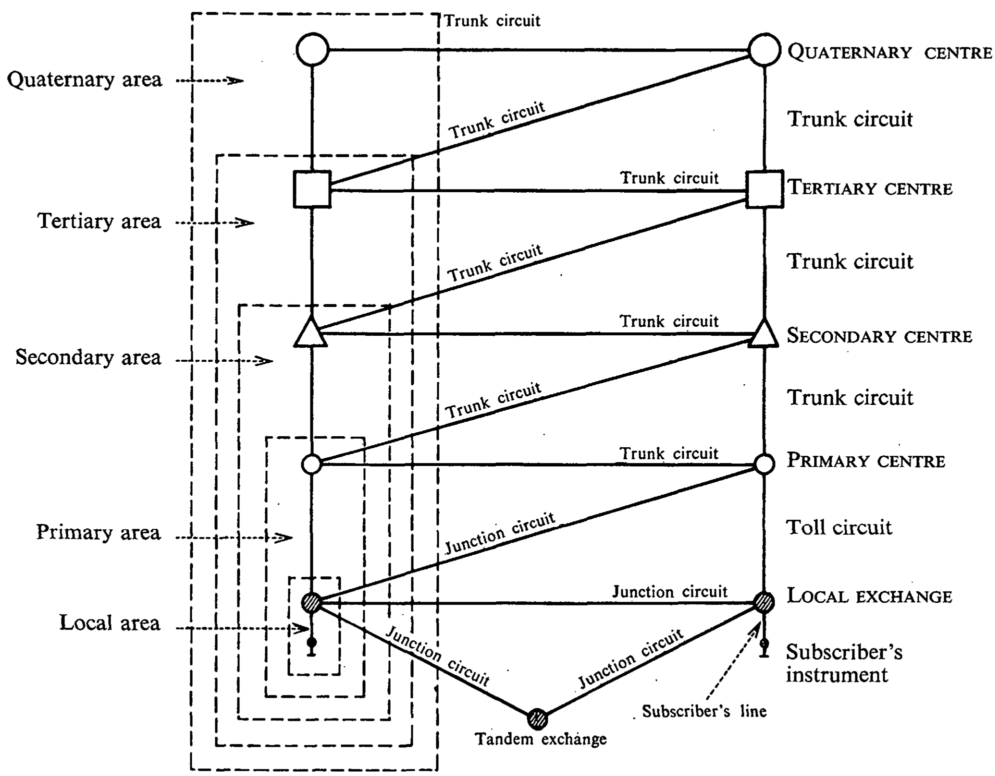


FIGURE 6. — **Designation of exchanges and circuits in a typical national network**

[1] This paragraph is identical with the corresponding passage in Chapter V, " Transmission ", of the *Manual on National Telephone Networks for the Automatic Service.*

TANDEM EXCHANGE: an exchange used for connecting local exchanges within a metropolitan network.

PRIMARY CENTRES: centres to which local exchanges are connected and via which long-distance connections are established.

SECONDARY CENTRES: centres to which primary centres are connected in order that long-distance connections might be established.

TERTIARY CENTRES:
QUATERNARY CENTRES:
QUINARY CENTRES:
} if required these can be defined in an analogous way to secondary centres.

The circuits connecting the subscribers' telephone instruments to the local exchange are SUBSCRIBERS' LINES *(lignes d'abonné)* which, together with the local exchange, form the LOCAL AREA *(zone locale)*. The circuits interconnecting local exchanges, either directly or via tandem exchanges, are JUNCTION CIRCUITS *(circuits de jonction)*. The circuits between the local exchange and the primary centres are TOLL CIRCUITS *(circuits locaux)*. All other circuits are called TRUNK CIRCUITS *(circuits interurbains)*.

The group of local exchanges (and their surrounding local areas) served by a primary centre together with the toll circuits is termed a PRIMARY AREA *(zone primaire)*. Similarly, the group of primary areas served by a secondary centre together with the trunk circuits is termed a SECONDARY AREA *(zone secondaire)*. Areas of higher rank are analogously defined.

All these terms are illustrated in Figure 6.

b) *Calculation of limits within a primary area*

Within each primary area, the resistance of the lines is limited by switching conditions and, for central battery operation, by current supply requirements of the telephone sets.

Similarly, within each primary area, the transmission loss of the circuits is limited by the requirements of the universal national service.

The limit of the sum of the transmission losses of the circuits may be determined as follows:

1. A decision is made as to how account is to be taken of the variations with time of the transmission loss of the circuits, the frequency bandwidth transmitted over the network, the noise in the network, the losses due to passages through exchanges, junction losses between the various parts of the trunk circuits as well as the quality of the telephone sets.

2. The nominal transmission loss is fixed for the circuits between higher centres and the nominal transmission loss for circuits fixed between primary and secondary centres. It is pointed out that the extension of the zone of use of four-wire circuits as nearly as possible to the primary centres, and/or the introduction of four-wire through switching, allow an increase in the transmission loss available within the primary areas. On the other hand, it should be verified that the arrangements made are compatible with the C.C.I.T.T. recommendations applicable to the chain formed by the international telephone circuits and the national trunk circuits concerned in an international telephone call (see the above Recommendation).

3. The available equivalent under the conditions thus defined is calculated within each primary area by considering a call between any two subscribers.

4. It must be decided how account is to be taken within the primary areas of the passages through the local exchanges, the junction losses between the various parts of the local end, the terminating conditions of the toll circuits and the subscribers' lines in loaded cable, losses in the repeating coils inserted on open-wire lines.

5. The allowable equivalent or transmission losses are calculated for the lines and circuits in each primary area. The frequency or frequencies to be taken into consideration in the case of unloaded cables should also be decided.

c) *Improvement of performance in existing networks*

Within existing telephone networks it is important to improve the transmission quality for unfavourably situated telephone sets which handle considerable traffic and especially international traffic. Several methods can be used for this purpose, for example:

1. Repeaters may be used on the subscribers' lines, the junction circuits in the networks of large towns and the toll circuits.

Note. — These repeaters may be either two-wire repeaters of the standard type or negative impedance repeaters (two- or four-wire). In each case it should be verified that the stability of the transmission remains adequate.

2. The transmitting and receiving insets may be graded in several qualities and the better insets may be fitted in the telephone sets served by lines having the greatest attenuation and vice versa.

Note. — Annex 4 modified (Part II of this volume) reproduces information on the constitution of the national networks of different countries.

RECOMMENDATION P.22, SECTIONS 3 AND 4

(See Volume V of the *Red Book*, pages 19 to 118)

SECTION 5

OBJECTIVE MEASURING APPARATUS

RECOMMENDATION P.51 (modified in Geneva, 1964)

ARTIFICIAL VOICES; ARTIFICIAL MOUTHS; ARTIFICIAL EARS

Although experiments directed towards replacing the human mouth in telephonometric measurements, for example by gramophone records associated with a loud-speaking telephone receiver or by a loud-speaking telephone receiver fed with a mixture of frequencies, have not yet reached the stage at which an internationally specified mechanical arrangement could be agreed which would allow the human mouth to be safely replaced, and although, for the present, it is considered essential that all telephonometric measurements at the C.C.I.T.T. Laboratory should continue to be made with the human mouth and ear, it is desirable to study the problem of designing apparatus for telephonometric measurements such that, in the future, it should be possible to make these measurements without recourse to the human mouth and ear.

In order to guide Administrations and private operating Agencies in this study, the original articles concerning the problem of the artificial ear are indicated in the bibliography below.

Meanwhile, it goes without saying that Telephone Administrations and private operating Agencies can, if they wish, use in the future devices which they may have been able to construct for large-scale testing of telephone apparatus supplied by manufacturers, provided that the results obtained with these devices are in satisfactory agreement with results obtained by real voice-ear methods.

Note 1. — The Plenary Assembly at Copenhagen in 1936 considered that it would be of interest to deal separately with the design, on the one hand, of an artificial speech source (words recorded on a gramophone record or a mixture of pure tones having the same effect as the human voice) and, on the other, of apparatus for producing a defined acoustic field according to certain specified conditions which will reproduce artificially a human mouth. The term " artificial voice " may be used for the former and " artificial mouth " for the latter.

Note 2. — Pending the standardization of an artificial ear for general use, the C.C.I.T.T. recommends the provisional adoption of a " reference coupler " which will be used by the Administrations and private operating Agencies taking part in C.C.I.T.T. work. The dimensions of the " reference coupler " are given in the appendix below.

Note 3. — The general question of artificial voices, mouths and ears continues to be studied by the C.C.I.T.T.

Note 4. — Annexes 8 to 16, Part II of Volume V of the *Red Book* describe, for information, the artificial mouths and ears used by the Administrations of France, Italy, Federal Republic of Germany, the United Kingdom of Great Britain and Northern Ireland, Switzerland, Czechoslovakia and U.S.S.R., by the Chile Telephone Company and by the F.A.T.M.E. Society (Italy). Annex G (Part II of Volume V*bis*) describes the artificial mouth used by the Swedish Administration.

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APPENDIX

Reference coupler recommended by the C.C.I.T.T.

Until a standard artificial ear is generally adopted, the C.C.I.T.T. recommends the use of a provisional " reference coupler ".

The object of this decision is simply to permit comparison between the results of objective measurements made on telephone receivers in the C.C.I.T.T. Laboratory and in the national laboratories. As this is a provisional decision, the easiest procedure is to take as a reference coupler one that has the simplest construction and which has been the subject of a detailed specification. It is therefore proposed to adopt the coupler used in the United States of America and also in many other countries of the world by telephone Administrations and manufacturers. It is pointed out that, with this coupler, certain precautions must be taken in the application of telephone receivers with a very small ear-cap.

This coupler has been derived from the type 1 coupler specially designed for the measurement of response curves of receivers having hard ear-caps (see the extracts from the standard defining this coupler in Annex 17, Volume V of the *Red Book*). The characteristics of this type 1 coupler are recalled in Figure 7 [1].

This theoretical coupler, defined by its geometrical dimensions, satisfies the following conditions:

— all the walls are considered as being completely hard, i.e. all the walls (including the microphone diaphragm) offer a very high acoustic impedance with respect to that of the gas (air) filling the cavity;

— the receiver considered, applied to the upper surface of the coupler, corresponds to a type of receiver whose ear-cap has no cavity (see Figure 7);

— the nominal volume of the cavity thus defined (without leaks) is 6 cm³, made up by the following three component volumes:

$$V_1 = 1.745 \text{ cm}^3$$
$$V_2 = 3.723 \text{ cm}^3$$
$$V_3 = 0.534 \text{ cm}^3$$

Volume V_3 represents the volume of the cavity of a type WE 640 AA microphone, in respect of which it has been assumed that the diaphragm impedance was infinite (see Standard ASA Z 24.9.1949, point 3.2.1).

[1] Figures 7, 8 and 9 of this volume replace Figures 26 and 27 of Volume V of the *Red Book*.

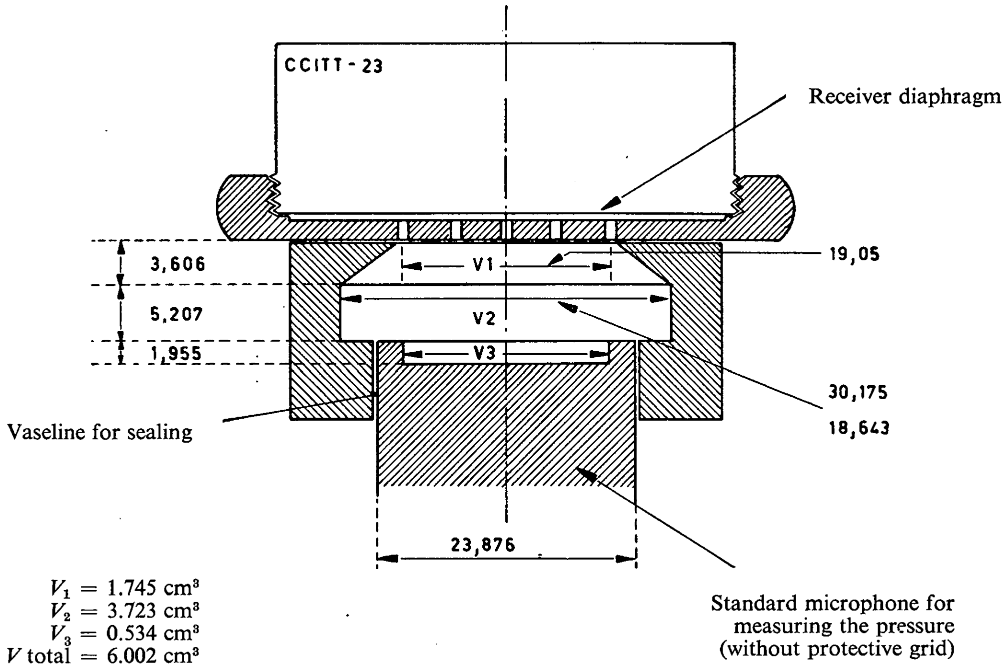


FIGURE 7. — **Type 1 coupler used with telephone receivers**

Notes. — 1) The dimensions are in millimetres.

2) These dimensions are given in inches, with tolerances, in Figure 2a, Annex 17 of Volume V of the *Red Book.*

In the standard, it is specified that in practice when an actual (electrostatic) microphone is used, it may be necessary to alter the height of the cylindrical part (volume V_2, nominal height 5.21 mm or 0.205 inch) to allow for the following factors:

— equivalent volume of the microphone ΔV (about 0.200 cm³ for the microphone WE 640 AA, and 0.150 cm³ for a microphone of the type B and K No. 4132);

— volume of the microphone cavity V_3, when its dimensions differ from the nominal values (d = 18.64 mm or 0.734 inch, h = 1.95 mm or 0.077 inch);

— where appropriate, the volume occupied by the protective grid;

— possible correction of the volume to allow for changes of atmospheric pressure due to altitude.

Thus, in the definition of the theoretical coupler, the *microphone is considered to be without a protective grid*, the latter being used only in the case of free field measurements or recordings, for example.

Examination of the C.C.I.T.T. reference coupler shows that (see Figure 3 of Annex 17 and Figures 8 and 9 below) the total volume of the coupler proper can be decomposed as follows (see Figure 8):

V'_2 volume of the cylindrical cavity 4.075 cm³;

V'_3 volume of the microphone cavity 0.534 cm³ (microphone WE 640 AA or B and K No. 4132, the latter being fitted with its adapter ring DB 0111);

ΔV equivalent volume of the microphone (0.200 or 0.150 cm³).

The total effective volume of this coupler is therefore 4.809 cm³ with microphone WE 640 AA, or 4.759 cm³ with the microphone B and K No. 4132.

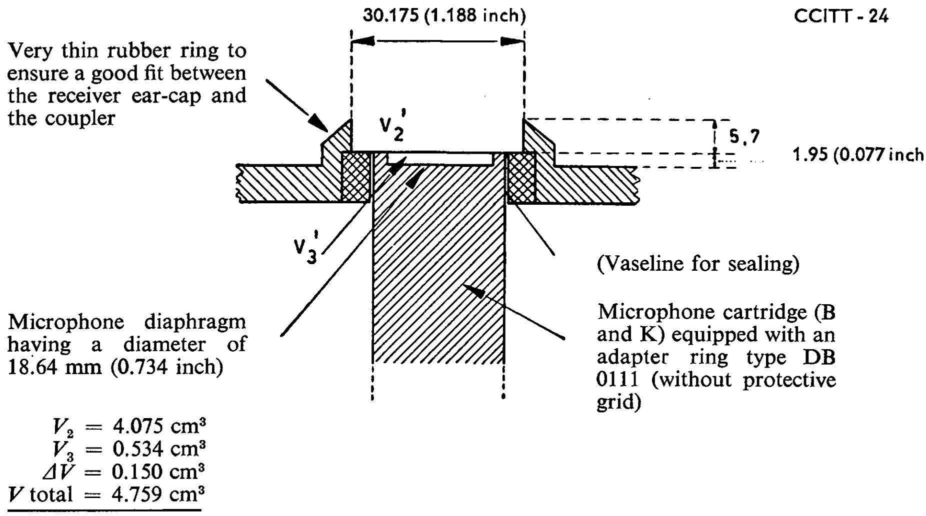


FIGURE 8. — **C.C.I.T.T. provisional reference artificial coupler**

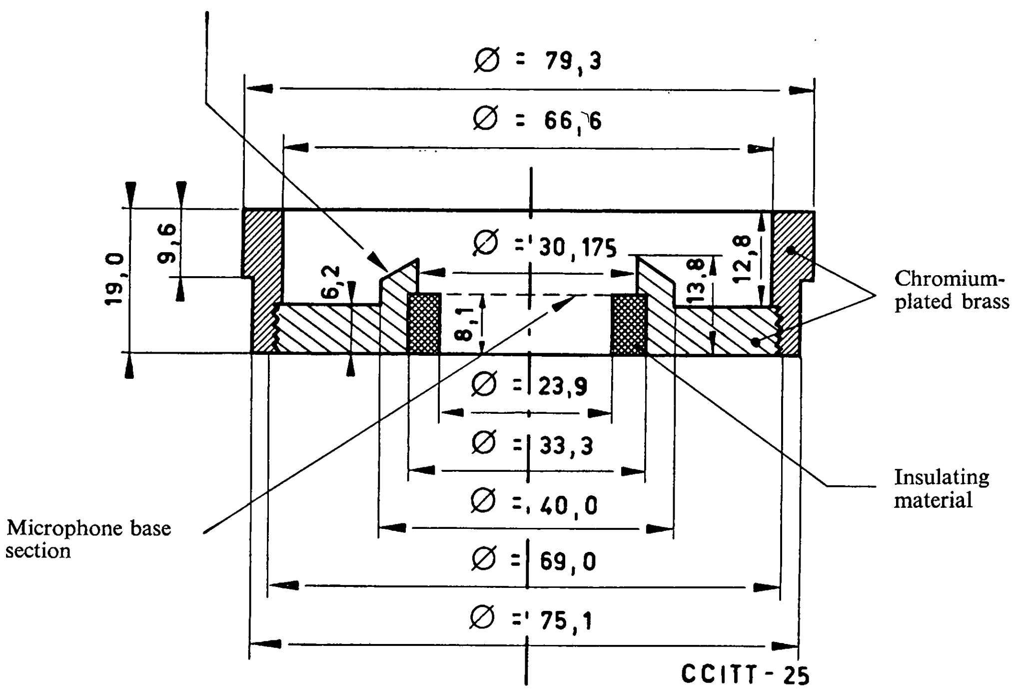


FIGURE 9. — **Exact dimensions of the C.C.I.T.T. provisional reference artificial coupler**

Note. — All dimensions are expressed in millimetres.

To this volume it is necessary to add the volume of the receiver ear-cap cavity contained between the reference plane of the coupler and the surface of the receiver ear-cap; this volume varies with the shape of the ear-cap. The total volume of the chamber cannot therefore be kept constant at the nominal value of 6 cm^3.

The C.C.I.T.T. reference coupler is used as shown in Figure 8, with the particular microphone used in the C.C.I.T.T. Laboratory (*B and K No. 4132, equipped with a DB 0111 adapter ring (as described in Figure 8) without protective grid.*

Administrations may use any microphone they prefer, but they are reminded that if other types of microphone are used the height of the cavity will have to be changed. The following dimensions must, however, be observed: diameter, 30.175 mm, total volume, 4.759 cm^3, taking into account the equivalent volume of the microphone.

OTHER RECOMMENDATIONS OF SECTION 5, SECTIONS 6 AND 7

(See Volume V of the *Red Book*, pages 122 to 158.)

RECOMMENDATION P.62, MEASUREMENTS ON SUBSCRIBERS' TELEPHONE EQUIPMENT, Volume V of the *Red Book*, page 137, *paragraph C.1:* " Objective measurement of the reference equivalent (sending and receiving) of subscribers' telephone equipment "; Annex G (Part II of this volume) describing an apparatus used by the Czechoslovak Administration should be mentioned.

RECOMMENDATION P.63 (modified in Geneva, 1964)

METHODS FOR EVALUATING TRANSMISSION QUALITY ON THE BASIS OF OBJECTIVE MEASUREMENTS

These measuring methods are being studied by the C.C.I.T.T.

Methods which have been used by the Swiss Administration and the U.S.S.R. Administration are described in Annexes 30 and 31 (Part II of Volume V of the *Red Book*).

A new method for the determination of transmission performance is being studied by the American Telephone and Telegraph Company. The basic measuring system is described in Annex 1 to Question 15/XII (Part III of this volume).

SECTION 8

MEASUREMENTS FOR MAINTENANCE OF SUBSCRIBERS' TELEPHONE EQUIPMENT AND FOR FACTORY ACCEPTANCE TESTING

RECOMMENDATION P.81 (modified in Geneva, 1964)

MAINTENANCE OF SUBSCRIBERS' EQUIPMENT

To ensure good transmission on international connections, the C.C.I.T.T. recommends periodical testing of each subscriber's equipment.

Different procedures exist for making check tests from the exchange of subscribers' stations under working conditions by means of subjective or objective measurements.

The most important of these are the following:

1. *Subjective measurements.* — a) Quick conversation test; b) Complete telephonometric test.

2. *Objective measurements.* — It is possible to envisage maintenance also based on the procedures used for factory acceptance testing. (This form of maintenance does not involve the exchange.)

1. *Subjective measurements*

a) *Quick conversation test.* — This method is used mainly in the United States (by the American Telephone and Telegraph Company) and in Switzerland. Furthermore, in Great Britain, the transmission quality of telephone equipment in public call-boxes and subscribers' stations having extension sets is assessed by means of a conversation with the exchange test desk clerk [1].

b) *Complete telephonometric test.* — This method seems to be no longer used.

2. *Objective measurements*

a) *Measurements made from the test desk.* — In Switzerland the checking of transmission quality is made subjectively by an exchange of conversation with the test desk which deals with fault control for each terminal trunk exchange. From this test desk, measurements and checks of subscribers' lines can also be made from the point of view of insulation, loop resistance, transmission of dialling impulses, etc.

[1] The British Administration considers that the cost of applying preventive maintenance to other types of telephone sets would not be justified.

b) *Electrical measurements of a general nature.* — To check the transmission quality of subscribers' telephone equipment in service, the Dutch Administration uses the same measuring equipment and methods as for factory testing; nevertheless it must be understood that the permissible limits are somewhat larger. These measuring methods are described in Recommendation P.82.

c) *Use of special measuring equipment for checking telephone equipment.* — The Federal German Administration uses the following methods.

The testing of telephone equipment to check the transmission quality of subscribers' telephone equipment in service is applied mainly to the measurement of microphone and receiver capsules, because their transmission quality depends very much on the material used and the quality of manufacture. Specifications have been fixed for microphone and receiver capsules against which they are checked by means of the equipment for objective measurement of reference equivalents described in Annex 28, Part II of Volume V of the *Red Book.*

The equipment for the objective measurement of reference equivalents enables the reference equivalents of microphone and receiver capsules to be measured. For microphone capsules, non-linear distortion and microphone noise are measured at the same time as reference equivalent by means of the modulation products. Furthermore, it is possible to check the " sensitivity-frequency " characteristic by means of a visual display.

The microphone capsules are divided according to their sensitivity into groups in steps of 3.5 db and the receiver capsules in steps of 2.5 db. These groups correspond for microphone capsules to values of sending reference equivalent 8 to 4.5 db, 4.5 to 1 db and 1 to −2.5 db and, for receiver capsules, to values of receiving reference equivalent 0 to −2.5 db, −2.5 to −5 db and −5 to −8 db. This allocation into groups is then used to associate the capsules with corresponding groups of subscribers' lines (loop resistance 0 to 250 ohms, 250 to 500 ohms and 500 to 750 ohms—see Part II of this volume, paragraph X of Annex 4.

For this allocation the capsules are stamped with the figures I, II, or III. Thus it is possible not only to compensate for too high values of reference equivalent of subscribers' lines but also, on replacement of capsules when the telephone set is repaired, to make sure that the capsules have not been changed after being put into service. For this reason the lineman who is dealing with the location of faults must always have with him some capsules of the various groups; capsules which are removed from subscribers' sets are checked at the headquarters stores depot by the equipment for the objective measurement of reference equivalents so as to determine whether they are still serviceable.

The measurement and grouping of microphone and receiver capsules with the aid of the equipment for the objective measurement of reference equivalents were introduced several years ago in the Federal German Posts and Telecommunications Administration. Each headquarters has at its telecommunications stores depot one such measuring equipment operated by non-specialist female staff. The measuring precision is so high that when the same capsule is measured with a different measuring equipment the differences are less than 1 db. The grouping of capsules and their correct allocation to the telephone sets can, so far as present experience has shown, be done without difficulty. They are considered by the telephone service staff, particularly the officers on fault location duties, as a great step forward because they are able to ensure that, by means of this grouping, the variations of receiving loudness can be compensated for different lengths of subscriber's line. A large percentage of capsules in service (about one-third of the microphone capsules and one-sixth of the receiver capsules) had to be replaced, which resulted in a great improvement in transmission quality. It was noted that most of the microphone capsules in service did not correspond to the present conditions. This also applies for receiver capsules, but to a lesser extent.

The American Telephone and Telegraph Company plans to assess subscriber sets with an electroacoustic rating system (E.A.R.S.).[1] This system is used in the laboratory to determine relative ratings of telephone set designs and local transmission plans which correlate with subjective loudness ratings. There are no present plans for using this system to evaluate subscribers' sets on an in-service basis for maintenance purposes.

RECOMMENDATION P.82 (modified in Geneva, 1964)

FACTORY ACCEPTANCE TESTING OF SUBSCRIBERS' EQUIPMENT

The methods used in various countries are described below for information.

Denmark

In addition to inspection and mechanical examination, the equipment is given the following transmission test:

The handset is placed in a support containing a sound source (artificial mouth) and a microphone (artificial ear).

With an 800-ohm generator connected to the terminals of the equipment, the acoustic pressure produced by the telephone receiver is measured and this appears on a cathode-ray oscillograph as a function of frequency over the frequency band 300-3400 c/s. In this way a simultaneous check is provided of the receiver capsule and the electrical receiving circuit.

A feeding bridge and a line impedance of 800 ohms are connected to the terminals of the equipment and the voltage at these terminals is measured while a constant acoustic pressure of 20 dynes per square centimetre, provided by the sound source, is applied to the microphone. The voltage obtained appears on a cathode-ray oscillograph as a function of frequency over the frequency band 300-3400 c/s. In this way a simultaneous check is provided of the microphone capsule and the electrical sending circuit.

The oscillograph is provided with a transparent scale on which are drawn the limit curves for sending and receiving, i.e. the mean curves ± 2 db.

[1] See Annex 1 to Question 15/XII in Part III of this volume.

UNITED STATES OF AMERICA

In addition to measurements on the various component parts of the telephone equipment, the principal measurements made upon subscribers' telephone equipment in the factory by the American Telephone and Telegraph Company are the following:

1. Once the assembly of the handset is complete:

a) both the shape and the level of the " sensitivity-frequency " characteristics of the microphone and receiver are determined by means of a cathode-ray oscillograph on the screen of which curves corresponding to the tolerance limits are drawn;

b) the d.c. resistance of the carbon microphone is measured for which upper and lower limits have been established.

c) the impedance of the varistor used to protect the receiver is measured at 60 cycles and compared with an established upper limit.

2. When the telephone set is completely assembled:

a) the ringing is tested, a given input voltage being applied;

b) to check the circuit continuity rather than to detect faulty components, a howling sound is applied to the microphone by an acoustical path in order to excite it and the following measurements are made:

1) the output voltage across an artificial line representing the subscribers' line,

2) the acoustic pressure produced by the receiver and transmitted by the side-tone path.

c) a test of the isolation-to-ground of the telephone circuit is made using a breakdown voltage of 500 volts d.c.

The tests and measurements described above are made on all sets and not on a sampling basis.

FRANCE

The French P.T.T. Administration has studied and prepared a set of apparatus for:

- the checking and maintenance of telephone sets on the subscribers' premises;
- bulk factory acceptance tests of consignments of subscribers' sets, conforming to an accepted type, submitted by manufacturers for the approval of the Administration;
- maintenance in regional exchanges.

Descriptions of these various types of apparatus are given in paragraphs II, III and IV of Annex 27, Part II of Volume V of the *Red Book*.

NETHERLANDS

The Dutch Administration has put measuring equipment at the suppliers' disposal by means of which they are required to examine the sensitivity of each microphone and receiver capsule delivered to this Administration.

In addition it is necessary to measure the resistance of each microphone capsule when white noise of spectrum restricted to the band 300-3400 c/s is applied to the microphone in an acoustic chamber. The microphone is connected to an electrical circuit which, for both a.c. and d.c., is equivalent to the average conditions obtained when the microphone is connected in the telephone network. The d.c. resistance is also measured in this condition at the current which would apply in practice. The noise voltage produced by the microphone is measured by means of a d.c. voltmeter connected in a Graetz circuit. The voltmeter indicates approximately the r.m.s. value.

For measuring the telephone receiver, the reciprocity principle is used by applying the white noise to the receiver, acoustically, and measuring the voltage across the receiver.

In this case too, the receiver is connected in a circuit which has the same nominal impedance as that of normal telephone equipment.

The levels measured in this way yield a statistical distribution and the Administration requires that no microphone or receiver capsule may be accepted which departs more than ± 3 decibels from the mean. The absolute level of the mean is also fixed by the Administration.

So far as the " sensitivity-frequency " characteristic is concerned the manufacturers are required to guarantee, for each capsule, that this complies with the tolerances specified in the Administration's standard. Experience has shown that the Dutch Administration can confine itself to checking from time to time by sampling whether the relevant clauses concerning the " sensitivity-frequency " characteristic are being observed. In general, the Administration uses the same measuring equipment for checking as is used in the factory. The measuring equipment used by the manufacturer for final checking in the factory must have been approved by the Administration. Furthermore the Administration has reserved itself the right to make measurements on the microphones and receivers in the factory.

The transmission characteristics of each induction coil must be guaranteed by the manufacturer. He can conduct his checking during manufacture in a manner approved by the Dutch Administration.

Federal Republic of Germany

For tests made from the transmission point of view, the Federal German Posts and Telecommunications Administration uses, for the acceptance of subscriber's telephone equipment by its telecommunications stores depots, the equipment for objectively measuring reference equivalent described in Annex 28, Part II of Volume V of the *Red Book*. It has been possible to prove that, in the case of good manufacture, there are scarcely any faults in assembling telephone equipment. It is therefore sufficient to make random tests at the time of acceptance. Nevertheless, on delivery all microphone and receiver capsules are again measured and grouped as described below. Furthermore, all reconditioned telephone equipment must be tested, but this is an easy matter because only a small number of items is generally involved.

When testing telephone equipment the mean sending and receiving loss is measured between the frequency limits of 200 and 4000 c/s. The resistances of line, receiver and microphone are each replaced by a 600-ohm resistor.

United Kingdom of Great Britain and Northern Ireland

General. — The processes of manufacture and the measurements made by the manufacturer are liable for inspection at any time by the Inspection Branch. Acceptance measurements are made on every piece of equipment manufactured or on samples chosen at random at the discretion of the Inspection Branch. The nature of the acceptance measurements is determined by agreement between the purchasing authority and the manufacturer before the contract is placed.

Electro-acoustical measurements on telephone microphones and receivers. — Each manufacturer is required by the Post Office to equip himself with measuring equipment of an approved design. This employs specified bandwidths of continuous-spectrum noise. To ensure that each manufacturer used the same testing signal, these bands of noise have been recorded by the Post Office in the form of optical soundtracks on a glass disc; each manufacturer is supplied with discs which are positive prints from the master negative.

For testing microphones, the appropriate noise signal is fed into an artificial mouth (see Annex 11 of Volume V), the output level of which is adjusted to a specified value with the aid of a probe microphone. The carbon microphone under test is given a conditioning treatment, placed in a standardized position in front of the artificial mouth, and the output voltage across a standard circuit is observed by means of a voltmeter. Special steps are taken to check the pressure calibrations of all the probe microphones in use at the various factories.

The voltmeter indicates true r.m.s. values, is scaled in decibels relative to 1V, has an integration time of 1.4 seconds (see Annex 18) and its reading, when a sinusoid is applied, is independent of frequency over the range 300-3400 c/s.

The noise signals used are as follows. First, the over-all sensitivity of the microphone is measured by applying wide-band noise (300 c/s-3400 c/s); for the new type of microphone (the Transmitter Inset No. 16), the permitted tolerance is ± 2 db on the specified value. Secondly, a measurement is made with each of three narrow bands, to check that the frequency response is within tolerance.

For testing telephone receivers, the noise source feeds the receiver being measured which is placed on an artificial ear (see Annex 11); the output voltage of the artificial ear is measured across a standard circuit by means of a voltmeter. For the type of subscriber's telephone receiver normally manufactured, acceptance measurements are specified with three narrow bands of noise.

Complete telephone sets are not measured for performance. As all the component parts have been separately tested before assembly, only a simple check to ensure that the telephone does work is considered necessary.

Switzerland

Subscribers' equipment and spares purchased by the Swiss Telephone Administration are acceptance-tested. This work is entrusted to the stores testing section of the Research and Testing Division. Bulk tests are made with appropriate measuring apparatus. Some components are furnished to the manufacturers of telephone apparatus after being tested by the PTT, for example, the capacitance and insulation of condensers, the handset and various cords. The return speed and impulse ratio of dial contacts and the short-circuit contacts are tested in a few seconds with SC12/SC14 (Sodeco) equipment.

The subscriber's equipment (without handset) is acceptance-tested with TLPG3 (Zellweger) measuring apparatus in a minute or so. This test covers insulation, loudness of the bell, impedance for the circuit conditions applying when calls are being received, composite attenuations for sending, receiving and sidetone, at 400 and 1600 c/s, if necessary for two different feeding currents; the check may also be extended to cover high-frequency interference suppression, the busy impulse for party lines and any auxiliary circuits in the subscriber's equipment.

Microphones and earphones are checked rapidly with the KP51/MPG12 (Autophon/Zellweger) measuring apparatus described in Annex 29, Part II of Volume V of the *Red Book*. Checks are made of the reference equivalent and the frequency curve as well as the resistance and noise level of the microphones and the centring of the transmission system in respect of the earphones.

Manufacturers of telephone equipment and components use similar measuring apparatus.

Apparatus returned as faulty by the operating services is tested in the same way and by the same testing section as that delivered by the suppliers.

PART II

ANNEXES TO RECOMMENDATIONS

ANNEX A

(Geneva, 1964; quoted in Recommendation P.11)

STATISTICAL DISTRIBUTION OF REFERENCE EQUIVALENTS OF INTERNATIONAL CALLS COMPLYING WITH THE OLD C.C.I.T.T. TRANSMISSION PLAN AND THE NEW ONE

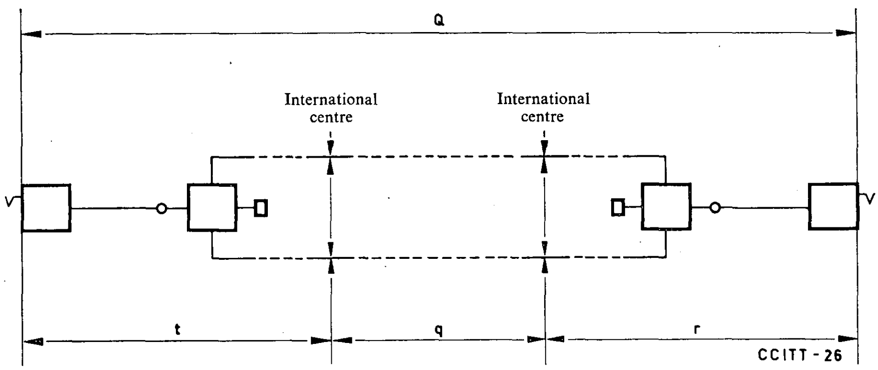


FIGURE 1

1. *General*

1.1 *Definitions of the variables considered* (see Fig. 1)

Q = over-all reference equivalent for an international call

t = reference equivalent of the national sending system } including the loss of the four-wire terminating set

r = reference equivalent of the national receiving system } including the loss of the four-wire terminating set

q = attenuation (between virtual four-wire switching points) of the chain of international circuits with four-wire interconnection.

Indices

x_e = recommended upper limit for the variable x

x_m = median value } of the normal law to which the distribution law of the variable x in the range of figures considered may be assimilated

σ_x = standard deviation } of the normal law to which the distribution law of the variable x in the range of figures considered may be assimilated

1.2 *Relations*

For a communication, $Q = t + q + r$ (relation between the actual magnitude of the variables); according to Recommendation P.11 B.d), this relation is sufficiently accurate for practical purposes, if q is taken as the value at 800 c/s.

For one category of calls, characterized by laws of statistical distribution:

$$Q_m = t_m + q_m + r_m$$

$$\sigma_Q^2 = \sigma_t^2 + \sigma_q^2 + \sigma_r^2$$

Furthermore, each of the standard deviations σ_t, σ_q, σ_r must in principle be decomposed into two, in accordance with the formula:

$$\sigma^2 = \sigma'^2 + \sigma''^2$$

in which σ' makes allowance for variations as from one call to another of the nominal value of the quantity considered (variation bound up with planning methods for σ'_t and σ'_r).

σ'' makes allowance for variations in time of the quantity in question with respect to its nominal value, and it is assumed that these two types of variations are independent.

In long-distance circuits, σ' is small and can be neglected in view of the other deviations.

1.3 *Application*

In what follows, we apply these relations to two transmission plans recommended or envisaged by the C.C.I.T.T. and briefly designated as:

Plan A — Old recommendations of the C.C.I.F. (*Green Book*, Volume III*bis*, maintained on a provisional basis in Volumes III and V of the C.C.I.T.T. *Red Book*).

Plan N — New plan contained in Part I, Section I of Volume III of the *Blue Book* and Volume V*bis* of the *Red Book*.

2. *Plan A*

2.1 *National reference equivalents*

The limits shown in Volume III*bis* of the *Green Book* are referred to the two-wire terminals in the international exchange. Hence, the loss of the four-wire terminating set must be added to these limits, giving (as in the former Recommendation P.11 of Volume V of the *Red Book*):

$$t_e = 18.2 + 3.5 = 21.7 \text{ db or } 21 + 4 = 25 \text{ dN}$$

$$r_e = 13.0 + 3.5 = 16.5 \text{ db or } 15 + 4 = 19 \text{ dN}$$

These figures are very close to those recommended in the *White Book* of the C.C.I.F. (Volume III, Budapest, 1934, page 77). According to the *White Book*, it seemed desirable, when setting up new networks or changing existing ones, to try to keep below 3 db (3.5 dN) of the limits for some 90% of subscribers.

The few data to hand in connection with the statistical distribution of reference equivalents in national networks show a high degree of dissymmetry. However, in that part of interest to us, that is to say, for subscribers approaching the limit, we can assume that the law of distribution is a normal law with a standard deviation σ of at least 3 db (3.5 dN). Since 90% of subscribers means a deviation of 1.28 σ, the above condition will be satisfied with a small margin (an improvement of about 4 db instead of 3 db) if deviations of more than about 2.58 σ can be neglected (corresponding, in accordance with the normal law, to 99.5% of subscribers), i.e. assuming that

$$t_m = t_e - 2.58\,\sigma_t \qquad r_m = r_e - 2.58\,\sigma_r$$

$$\text{i.e. } t_m = 14 \text{ db or } 16 \text{ dN} \qquad r_m = 8.8 \text{ db or } 10 \text{ dN}$$

Of course, these figures are not to be confused with the mean or median of the true distribution of t or r.

2.2 *Equivalent of the four-wire chain*

It can be assumed that q_m is the same as the nominal equivalent, i.e., 0 db (0 N) between four-wire extremities.

2.3 *Variations with time*

In principle, variations in the national circuits should be included in t_e and r_e. Had that been the case, there would have been no difficulty about taking in the new plan as *limits* for the *nominal* reference equivalent (including four-wire terminating set) *the old limits, including variations with time* (four-wire terminating set excluded). Since it seems that many countries have not kept such a margin, it was not taken into account in calculating t_m and r_m. Nevertheless, allowance will have to be made for variations that actually occur and the figures that were recommended in the *Red Book*, Volume III, page 17, can be adopted, i.e., for each national system, $\sigma = 2$ db (2.3 dN).

Measurement results previously analysed by Study Group 4 show that $\sigma = 2$ db (2.3 dN) for each of the two interconnected international circuits (three circuits could be interconnected only when $\sigma \leqslant 1.5$ db, which would give more or less the same over-all variation). The normal law is still valid for deviations not exceeding about 2.5 σ.

2.4 *Results*

From the formulae in paragraph 1.2, we obtain:

$$Q_m = 14 + 0 + 8.8 = 22.8 \text{ db} \quad \text{or} \quad 16 + 0 + 10 = 26 \text{ dN}$$

$$\sigma^2 = 3^2 + 2^2 + 2 \times 2^2 + 2^2 + 3^2 = 34 \text{ db}^2$$

$$\sigma = 5.8 \text{ db (6.7 dN)}$$

3. *Plan N*

3.1 *National reference equivalents*

For a country of average size, the limits are:

$$t_e = 20.8 \text{ db} \quad \text{or} \quad 24 \text{ dN}$$

$$r_e = 12.2 \text{ db} \quad \text{or} \quad 14 \text{ dN}$$

These " limits " are applied only to a certain percentage of *actual international calls* affecting each particular country; the exact proportion should be *at least 95%*. If this condition is only just satisfied, we have

$$t_m = t_e - 1.64\,\sigma_t = 20.8 - 1.64 \times 3 = 15.8 \text{ db} \quad \text{or} \quad 24 - 1.64 \times 3.5 = 18.3 \text{ dN}$$

$$\text{similarly, } r_m = 7.2 \text{ db} \quad \text{or} \quad 8.3 \text{ dN}$$

As an example, the calculation has also been made assuming that the percentage of calls satisfying the limits would be 97.5% or 99.5% respectively.

3.2 *Equivalent of the four-wire chain*

It is assumed that the most complicated chain comprises six circuits to be operated with a nominal equivalent other than zero (intercontinental circuits, other international circuits, or certain national circuits in a large country).

Since the stability of such a chain requires that attenuation distortion be strictly limited, the reference equivalent of the circuits can be taken as being equal to the loss at 800 c/s.

Then:

$$q_m = 6 \times 0.5 = 3.0 \text{ db} \quad \text{or} \quad 6 \times 6 = 36 \text{ cN}$$

3.3 *Variations with time*

It is assumed, as for the calculations of stability (Recommendation G.131, Volume III of the *Blue Book*) that all the circuits, whether national or international, of the four-wire chain satisfy the future objective, i.e. $\sigma = 1$ db or 12 cN.

(Annex A)

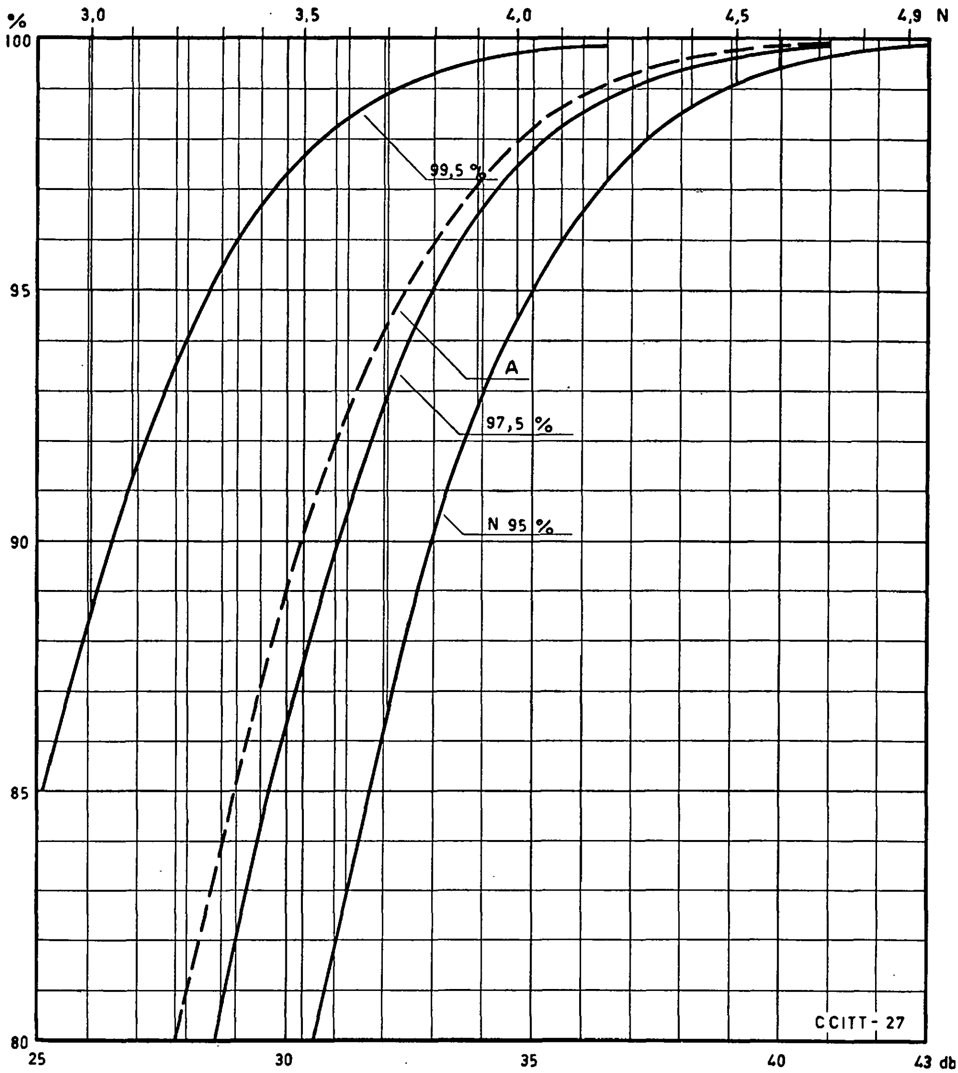


FIGURE 2. — **Percentage of subscriber pairs (plan A) or of calls over connections of six international circuits (plan N) for which the reference equivalent does not exceed the figure shown in the abscissae**

— — — —	A (old plan)	$t_e \leqslant 21.7$ db $r_e \leqslant 16.5$ db	for 99.5% of all subscribers
————	N (new plan)	$t_e \leqslant 20.8$ db $r_e \leqslant 12.2$ db	

Note. — For plan N the parameter of each curve is the percentage of calls affecting a given country for which the values of t_e and r_e are respected in that country.

3.4 *Results*

For the complete connection, we obtain

$$\sigma^2_Q = 3^2 + 12 \times 1^2 + 3^2 = 30 \text{ db}^2$$

$$\sigma_Q = 6.3 \text{ dN } (5.5 \text{ db})$$

Q_m is given, for the various cases, by the following table:

	Calls over connections affecting a given country and satisfying the send and receive limits t_e and r_e					
	95% $K=1.64$		97.5% $K=1.96$		99.5% $K=2.58$	
	db	dN	db	dN	db	dN
t_m	15.8	18.3	14.8	17	12.5	14.3
r_m	7.2	8.3	6.2	7	3.9	4.3
Q_m	26.0	30.2	24.0	27.6	19.4	22.2

$$t_m = t_e - K\sigma_t$$

$$r_m = r_e - K\sigma_r$$

$$Q_m = t_m + r_m + q_m$$

where K is the factor by which the standard deviation must be multiplied to obtain the required percentile for a Gaussian distribution.

In constructing the above table, the values of t_e and r_e were taken from section 3.1, $\sigma_t = \sigma_r$ were taken as 3 db (see section 2.1) and q_m is taken from section 3.2.

The percentages of calls over connections with the values of Q_m shown above and the value of σ_Q given in section 3.4, and for which Q may take various values, given in Figure 2.

4. *Conclusions*

Figure 2 summarizes the results of the preceding calculations. It thus shows the percentage of the combinations of pairs of subscribers (for Plan A) or of international calls (for Plan N) for which a given value of reference equivalent is not exceeded.

ANNEX B

(Geneva, 1964; quoted in Recommendations P.11 and P.13)

EFFECTS OF CERTAIN LEVELS OF CIRCUIT NOISE ON PERCENTAGES OF UNSATISFACTORY CALLS

This annex gives a summary of the results provided by Study Group XII during the period 1961-64 for the assistance of Special Committee C. The information was afterwards examined and commented upon by Study Group XVI and the additional results requested are included here.

The original study was devoted to the following five cases and information was supplied by the Federal German Republic, the United Kingdom, the L.M. Ericsson Company and the American Telephone and Telegraph Company.

FIGURE 1. — **Old plan for 2500 km**

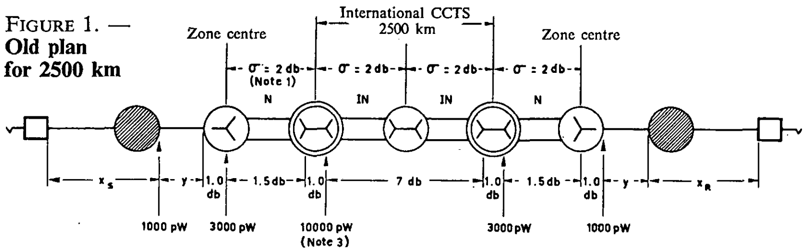


FIGURE 2. — **New plan for 2500 km**

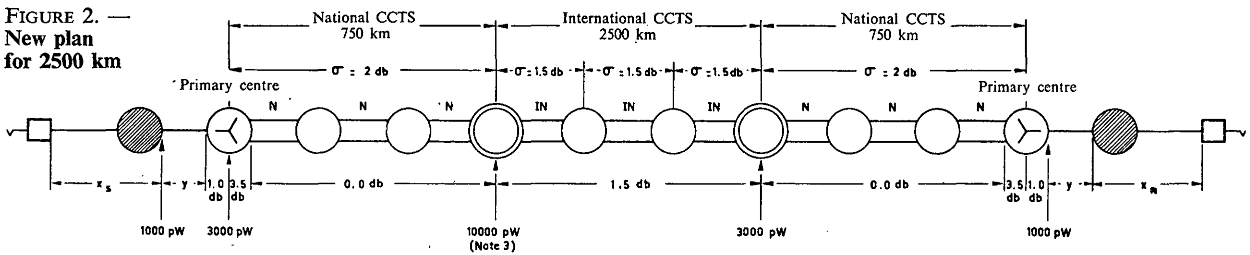
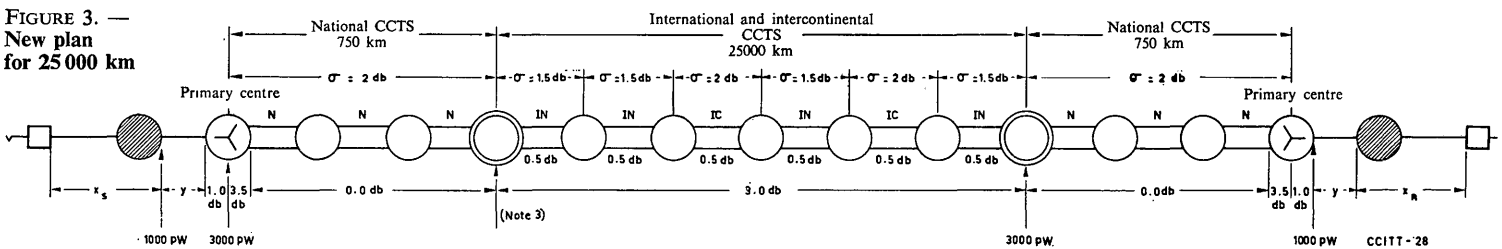


FIGURE 3. — **New plan for 25 000 km**



50 000 pW Case 3
100 000 pW Case 4
2 000 000 pW Case 5 Actual power (improved by B.P.O. compandor to an equivalent subjective noise power of 50 000 pW).

Hypothetical national, international and intercontinental circuits. Connections for the calculation of "Poor + Bad" opinion scores

Case 1. Old recommendations of the C.C.I.F. (maintained on a provisional basis in the New Delhi, 1960, versions of Recommendations P.11 and G.111). Over-all reference equivalent distributed as shown in Figure 2 of Annex A. Circuit noise allocated as shown in Figure 1 of the present annex. These recommendations were applied for connections up to 2500 km.

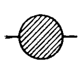
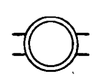
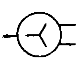
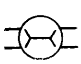
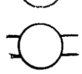
Case 2. New recommendations which were being studied at the time of this work but which now form the Geneva, 1964, Recommendations P.11 and G.111, applied to connections up to 2500 km in length. Circuit noise allocated as shown in Figure 2 of the present annex.

Cases 3, 4 and 5. As Case 2 but applied to connections up to 25000 km in length. These cases differ in respect of the circuit noise allocations (see Figure 3 of the present annex). The distribution of over-all reference equivalents was as shown in Figure 2 of Annex A (for 97.5% of outgoing or incoming international calls). These are based on realistic estimates of performance attainable in the future for 25 000-km connections.

In all these five cases an allowance has been made of 4 db impairment [1] for the combined effects of the attenuation/frequency distortions of the national and international circuits connected in tandem.

Later work, based only on the assessment data of the United Kingdom, took into consideration the following additional cases; for these the allowance for attenuation/frequency distortion was reduced to 2 db.

Legends and notes for Figures 1, 2 and 3

	Local exchange		International exchange
	Exchange with terminating unit	N	National
	Exchange with two-wire switching	IN	International
	Exchange with four-wire switching	IC	Intercontinental

1. Variation of loss with time is shown thus: I← $\sigma = 2$ db →I (in Fig. 1 the two international circuits are assumed to have a normal distribution with a standard deviation $\sigma = 2$ db for the individual circuits. Three international circuits could be interconnected only if the individual standard deviations are $\sigma \leqslant 1.5$ db).
2. Each of the the noise powers shown is referred to the zero relative level point of the circuit (or chain of circuits) concerned.
3. Noise power at all points is considered constant except at the international switching point where standard deviations $\sigma = 0$, $\sigma = 3$ db, $\sigma = 5$ db have been taken.
4. The points to which the various sources of noise have been referred are appropriate to transmission in the direction from left to right.

---

[1] In this annex the effects of attenuation/frequency distortion have been assumed equivalent to the insertion of a certain fixed additional loss, independent of the over-all reference equivalent. This is a simplification; a different, more realistic, method of allowance is to consider the additional loss as dependent upon the reference equivalent, decreasing in a certain manner as the reference equivalent increases. The latter method, explained in Reference [3], was used for treating examples, given in Reference [2], which relate to conditions very similar to those treated in this annex.

Case 1b. As Case 1 but distribution of national reference equivalents in accordance with statistics on actual international calls collected on its own network by the British Post Office during 1953 [1], when the old transmission plan was in force.

Case 3a. As Case 3 (except for the allowance for attenuation/frequency distortion).

Case 3b. As Case 3 but the standard deviation of the variations of the six links of the international circuits was reduced to 1 db for each circuit and the mean loss of this chain of circuits was reduced from 3 to 0 db.

Case 3c. As Case 3 but the standard deviation of the variations of the six links of the international circuits was reduced to 1 db for each circuit.

Cases 3a and 3b were also calculated using data provided by the American Telephone and Telegraph Company.

Only certain results of these studies have been selected for reproduction here.

Figure 4 illustrates the method of calculation, taking Cases 1b, 3a and 3b as examples. Each curve of the family indicated by (1) shows the distribution of over-all reference equivalent for the respective three cases. The curves indicated by (2) represent the percentages of customers who

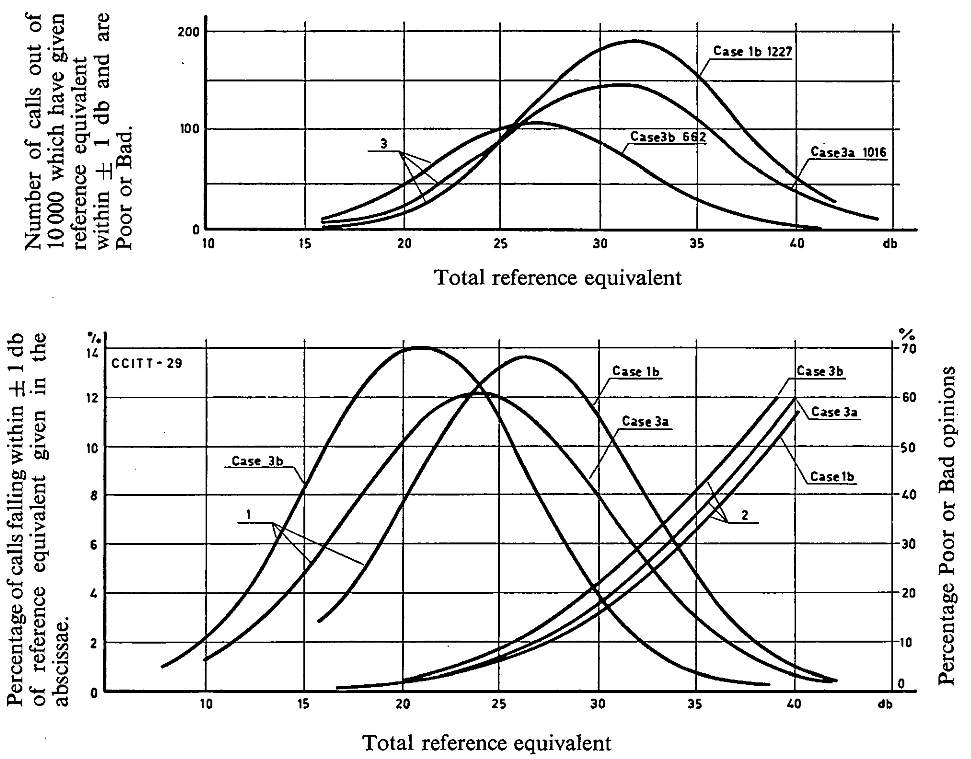


FIGURE 4

[1] The results are believed to be typical because the reference equivalents of the telephone sets involved, namely 12 db sending and 1 db receiving for the maximum permissible subscriber's line, lie very close to the medians of those of the sets of other Administrations; the circuit losses involved are also known to be fairly typical.

would regard as poor or bad a connection having the value of over-all reference equivalent shown in the abscissae and other particulars applicable to the three respective cases being considered. Il will be seen that values of reference equivalent in the range 25-40 db occur rather less frequenthy on connections according to the new plan (Cases 3a and 3b) than according to the Old Plan (Case 1b). On the other hand the percentage of poor or bad opinions is greater for the former because the circuit noise level is higher.

The combined effect of respective curves (1) and (2) has been expressed as follows: Each member of the family indicated (3) is the result of multiplying the ordinates of the respective curves (1) and (2) and represents the total number out of 10^4 calls which both lie within the range ± 1 db of the value of reference equivalent shown and are unsatisfactory. The integral of each respective curve indicated (3) gives the total number of calls out of 10^4 which are poor or bad.

The results for Cases 1, 2, 3, 4 and 5 are unduly pessimistic owing to the excessive allowance made for attenuation/frequency distortion and so they are not reproduced here. The following table shows the results for Cases 1b, 3a, 3b and 3c. The results for Case 5 (compandors associated with very high noise levels, see Fig. 3) would, making the smaller allowance for attenuation/frequency distortion, give the same result as Case 3a. This assumes compandors having the performance of those used by the United Kingdom.

Total percentage of poor or bad calls

		%
Case 1b	(Old Plan)	12.3
Case 3a	(New Plan, 25 000 km, original loss and standard deviation values)	10.2
Case 3b	(New Plan, 25 000 km, reduced loss and reduced standard deviation values)	6.6
Case 3c	(New Plan, 25 000 km, original loss and reduced standard deviation values)	9.2

The following conclusions may be drawn from the above table of results:

a) The degree of satisfaction obtainable on 25 000-km connections designed for a noise level of -43 dbm0p (Fig. 3) will, even without further reduction in loss, be slightly better than that experienced under the Old Plan (1953) on 2500-km connections designed for a noise level of -50 dbm0p (Fig. 1).

b) The effect of reducing the over-all loss of connections is very much greater than that of reducing the standard deviations although the latter will also improve the margin against instability and echo.

c) Although the situation is improved by reductions in loss and in values of standard deviations, there is no reason to be satisfied with the quality of service thus obtained and every possible effort should be made to obtain an even greater improvement and to avoid the introduction of all other unnecessary degradations.

REFERENCES

[1] ANNEX A (above): Statistical distribution of reference equivalents of international calls complying with theold C.C.I.T.T. transmission plan and the new one.

[2] WILLIAMS, H.: Over-all survey of transmission performance planning. *Proc. I.E.E.*, 1964, III, page 727.

[3] RICHARDS, D. L.: Transmission performance assessment for telephone network planning. *Proc. I.E.E.*, 1964, III, page 931.

(Annex B)

ANNEX C

(Geneva, 1964; quoted in Recommendation P.13)

NOTE ON THE ADDITION OF NOISE POWERS IN AN INTERCONTINENTAL CONNECTION

Study Group XII has made calculations assuming various values for the noise referred to the zero level point of the first circuit [1] of the 25 000-km intercontinental chain of circuits shown in Figure 3 of Annex B above.

For specification purposes the noise levels require translation into noise rates for each of the six individual circuits of the intercontinental chain of circuits.

For a simple calculation, assume that the six circuits are of equal length and each produces noise of power W (pW).

The individual noise contributions will now be referred to the receiving end of the 25 000-km chain of circuits. The noise in the first circuit is attenuated by 6×0.5 db, that of the second circuit by 5×0.5 db. Similarly for the remaining circuits.

The total noise received at the end of the 25 000-km chain of circuits in Figure 3 (Annex B) will then be $4.08 \times W$. Clearly the equivalent noise at the input end of this chain of circuits would be 3 db higher or $8.16 \times W$ (pW) but this value is actually given as 50 000 pW (Case 3), therefore $W = 6130$ pW but it was assumed that the circuits were of equal length and hence of length $\frac{25\,000}{6}$ km = 4165 km. The noise rate therefore will be $\frac{6130}{4165}$ pW/km or 1.47 pW/km.

The allowable noise rate in the case of these circuits must thus not exceed approx. 1.5 pW per km in order that the conditions of Case 3 shall be met.

ANNEX D

(Geneva, 1964; quoted in Recommendation P.13)

DETERMINATION OF A LIMIT FOR CIRCUIT NOISE ON AN INTERNATIONAL CIRCUIT USED FOR PUBLIC TELEPHONY AND NOT FITTED WITH A COMPANDOR

(Information supplied by the United Kingdom)

The problem

International telephone circuits introduce noise which can impair conversation. Compandors can sometimes be fitted and, in some cases, may effect an improvement. Several such circuits, some fitted with and some without compandors, might be connected in tandem to form a world-wide international telephone connection; a problem then arises of setting realizable and practical limits for the noise in each of the international circuits, so as to ensure that telephone transmission will be satisfactory on the vast majority of calls made over such connections.

[1] A zero level point of a circuit is to be understood in the present context to be the two-wire terminals of a hypothetical terminating set connected to the input terminals of the circuit in question.

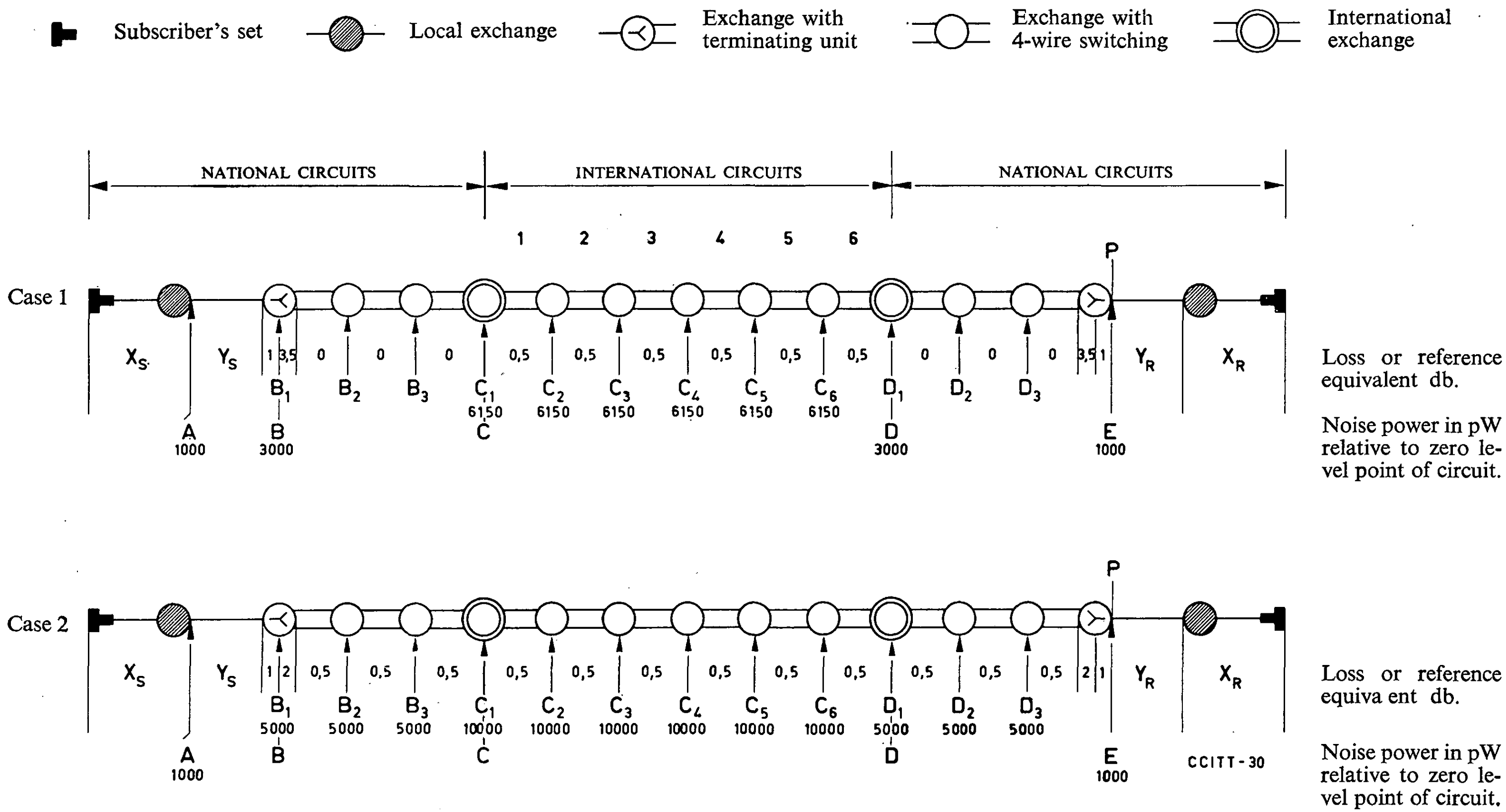


FIGURE 1. — **Details of connections for Cases 1 and 6**

Note. — The points to which the various sources of noise have been referred are appropriate for transmission from left to right.

TABLE 1

Classes of connections and associated noise powers

Case	Source or derivation	Equivalent noise power on (a) the sixth international circuit for Cases 1 to 5 (b) the fourth and sixth circuits for Cases 6 to 10		Equivalent noise power at position C of Figure 1 due to all six international circuits	
		pW	dbm0p	pW	dbm0p
1	As Figure 5 of Reference [1], and Annex B Figure 3 Case 3, but the noise due to the six international circuits is distributed uniformly among them (see Figure 1 for full details)	6 150	−52.1	50 000	−43.0
2	As Case 1 but noise due to sixth circuit is increased from 6150 pW to 25 000 pW	25 000	−46.0	83 600	−40.8
3	As case 1 but noise due to sixth circuit is increased from 6150 pW to 63 100 pW	63 100	−42.0	151 000	−38.2
4	As Case 2 but with a compandor fitted on sixth circuit	34 (see Appendix)	−74.7	39 100	−44.1
5	As Case 3 but with a compandor fitted on sixth circuit	158 (see Appendix)	−68.0	39 200	−44.0
6	Assumed adverse world-wide connection (see Figure 1 for full details)	10 000	−50.0	81 500	−40.9
7	As Case 6 but noise due to fourth and sixth circuits is increased from 10 000 pW to 25 000 pW	25 000	−46.0	129 300	−38.9
8	As Case 6 but noise due to fourth and sixth circuits is increased from 10 000 pW to 63 100 pW	63 100	−42.0	250 800	−26.0
9	As Case 7 but with compandors fitted on fourth and sixth circuits	34 (see Appendix)	−74.7	49 700	−43.0
10	As Case 8 but with compandors fitted on fourth and sixth circuits	158 (see Appendix)	−68.0	50 100	−43.0

The specific aspect of the problem treated here is to determine a limit of circuit noise, for any given circuit forming part of a connection, above which it would be desirable to fit compandors. Clearly the unnecessary insertion of compandors is not only uneconomical but may actually be deleterious to the quality of transmission performance.

It is emphasized that the noise limits considered here are applicable only for telephony; limits for other services such as telephone signalling, data transmission and voice-frequency telegraphy are not considered here.

The method

Design objectives laid down in respect of circuit noise present on various types of line plant will ensure that the transmission performance averaged over all circuits on a route will be satisfactory. There will, however, in practice be a small proportion of circuits that do not meet the current design objectives in respect of noise. The reasons may be economic or due to aging, repairs etc. If the amount of circuit noise is only slightly in excess of what is desired, it is best to accept the situation; if the excess is very large, it would be worth introducing a compandor on the circuit. The maximum circuit noise level at which it would be best to accept the situation and not fit a compandor is capable of calculation and the following information shows how this may be done.

The principles by which an assessment may be made of the transmission performance of a given class of telephone connections are described in Reference [1]; those principles have been used in all the calculations referred to in this Annex. Certain examples of the application of the principles are given in Reference [2]. The final result is expressed quantitatively as an estimate of the percentage of " unsatisfactory " calls made over a connection in the class under consideration; an " unsatisfactory " call in this context is defined as one that evokes the opinion " poor " or " bad ".

Classes of connection

Two groups of connection are considered, each group consisting of five cases. Figure 1 gives for the first case of each group (that is, Cases 1 and 6) details of numbers and types of circuits, losses, reference equivalents, and noise powers. The source or derivation of all ten cases is included in Table 1.

Losses and reference equivalents

Suitable values for the sending and receiving reference equivalents X_S and X_R and for the junction Y_S and Y_R are given in Table 2 below. (These values have been obtained from Table 3 of Reference [1]).

TABLE 2

Nominal over-all reference equivalent	Sending reference equivalent X_S	Receiving reference equivalent X_R	Junction losses $Y_S = Y_R$
db	db	db	db
28	9	0	3.5
24	8	−1	2.5

Circuit noise reaching the customers

The level of circuit noise reaching the customer depends not only on the levels of noise indicated in Figure 1 and Table 1 for each of the ten cases but also on the losses and reference equivalents. It is convenient to refer the combination of all circuit noises to the point P because this value is almost independent of the nominal over-all reference equivalent. It is then very simple to allow for the receive end loss and receiving reference equivalent, X_R and X_R.

Impairments other than loss and noise

The effects of bandwidth limitation or attenuation/frequency distortion have been allowed for in the present study by assuming an impairment of 2 db (at a nominal over-all reference equivalent of 32 db), for each of the ten cases of Table 1.

Allowance for the deleterious effects of a compandor when it is used under conditions of relatively low levels of circuit noise requires some explanation.

When the circuit noise (in the absence of a compandor) is very high, say greater than −30 dbm0p the formula given in the appendix for the effective subjective noise improvement applies without any further allowance. When there is very little circuit noise present a compandor will degrade the connection because of the presence of the combined effects of mismatch of compressor and expander, and various linear and non-linear distortions. The precise magnitude of this effect is difficult to determine and, in any case, depends upon the design and construction of the compandor. Results available for the British Post Office compandors described in Reference [3] suggest that an impairment of 1 to 1.5 db for a single compandored circuit would be a reasonable estimate. It may also be reasonably assumed that two compandored circuits in tandem (separately compandored) would result in twice these values. Accordingly Cases 4, 5, 9 and 10 have each been treated in two ways:

(1) on the basis of 1 db impairment per compandored circuit in the connection and

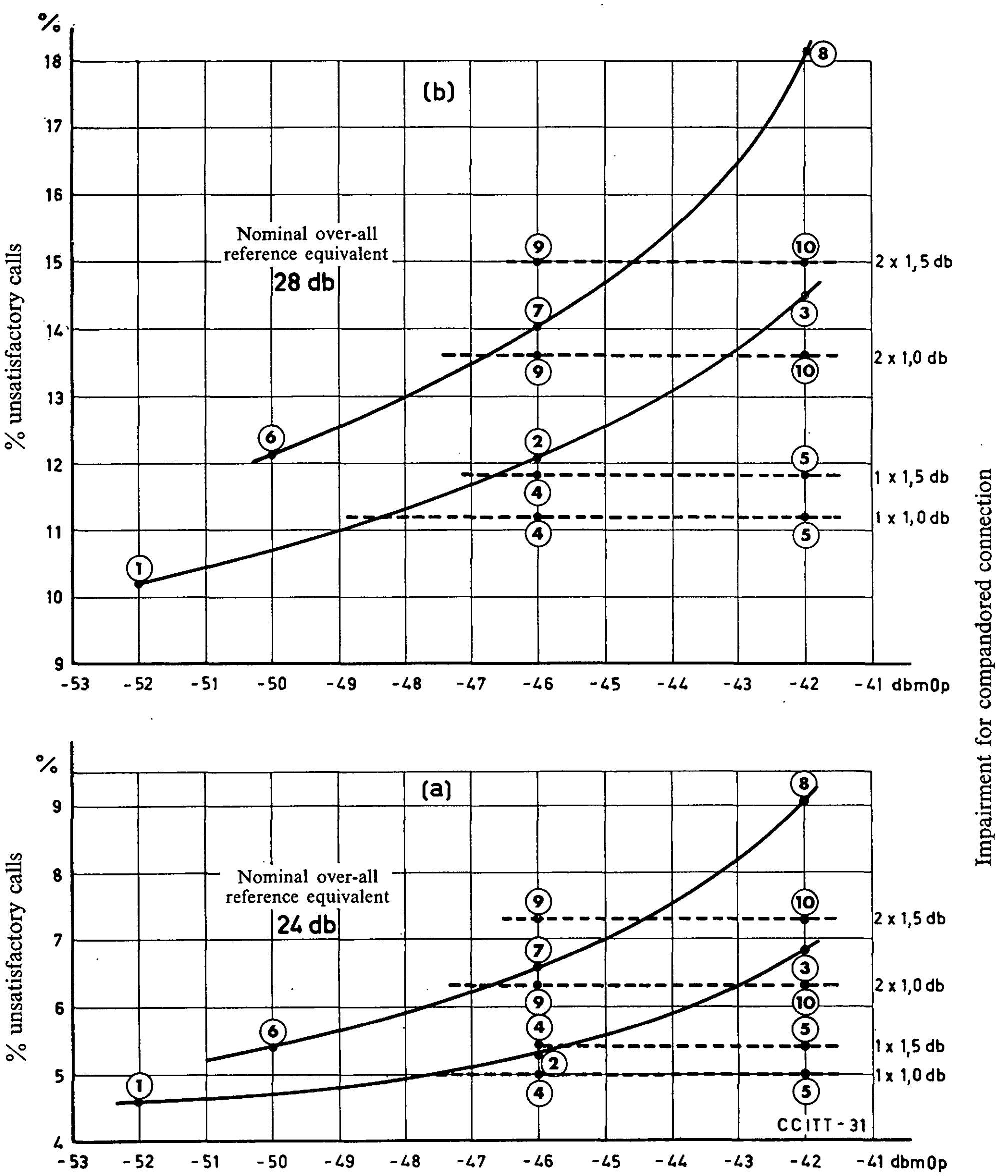
(2) on the basis of 1.5 db impairment per compandored circuit in the connection.

It has been assumed that the impairments on a connection due to the effects referred to in this section are additive and that the collective impairment can be treated, so far as calculation of percentage of " unsatisfactory " calls is concerned, as if it were merely bandwidth limitation or attenuation/frequency distortion.

Summary of results of calculations

The results of the calculations in accordance with the foregoing description are summarized in graphical form in Figure 2, which shows the percentages of " unsatisfactory " calls as functions of the level of circuit noise in the sixth circuit (Cases 1-5), and in the fourth and sixth circuits (Cases 6-10), of the international portion of the connection. The results for nominal over-all reference equivalents of 24 and 28 db are shown. Appropriately-shaped graphs have been drawn, with each graph connecting points relating to the same circuit arrangement. The intersection of graphs that refer to the same circuit arrangement but relating respectively to cases without compandor and those with compandor gives the circuit noise level at which the change should theoretically be made. Taking the more pessimistic estimates for the distortion impairment of compandors (that is, 1.5 db per compandored circuit) these values are as follows:

Nominal over-all reference equivalent	Circuit arrangements	Circuit noise level for equal performance without or with compandor
db		dbm0p
24	Figure 1 and Cases 1-5	−45.6
28	Figure 1 and Cases 1-5	−46.6
24	Figure 1 and Cases 6-10	−44.5
28	Figure 1 and Cases 6-10	−44.5



——————— Relates to connections without compandors

- - - - - - - Relates to connections with compandors

Circuit noise level (dbm0p) due to sixth circuit (cases 1-5) or due to sixth and fourth circuits (cases 6-10). Number in ring (e.g. ②) indicates case number of Table 1.

FIGURE 2. — **Percentages of unsatisfactory calls as function of the level of circuit noise**

Conclusions

Bearing in mind that compandors have other undesirable features besides those mentioned above, such as increasing the effects of variations in circuit loss, introducing possible troubles when the circuits are used for transmitting signalling, telegraphy or data, etc. it is considered that the introduction of a compandor would not be justified unless the circuit noise on that circuit exceeds a value of about −44 dbm0p. The following rule is therefore suggested:

" If the hourly-mean psophometric line noise power level of an international circuit substantially longer than 2500 km (e.g. 5000 km or more) is less than −44 dbm0p, no compandor is necessary. If the noise power level is greater than −44 dbm0p a compandor should be fitted ".

If the circuit noise (without a compandor) is very high the circuit may become suitable for telephony when a compandor is fitted but, because the compandor confers no advantage to telephony signalling, the noise level may be so high that the circuit would not be suitable for this service. It is tentatively suggested that an over-all upper limit of circuit noise should be fixed at about −33 dbm0p which should confer a margin on the design limit for international signalling. The following addition to the above rule is therefore suggested.

" It is tentatively suggested that, if the line noise power is greater than −33 dbm0p, the circuit, even though compandored, should not be allowed to form part of a six-circuit connection ".

REFERENCES

[1] Richards, D. L.: Transmission performance assessment for telephone network planning. *Proc. I.E.E.*. Volume III, No. 5, May 1964.

[2] Annex B (above): Effects of certain levels of circuit noise on percentages of unsatisfactory calls.

[3] Thomson, D.: A compandor using junction transistors. *Proc. I.E.E.*, 1959, 106B Suppl. 16, p. 619.

Appendix (to Annex D)

Equivalent subjective noise level of compandored circuits

The subjectively equivalent circuit noise level of a circuit fitted with a compandor was calculated by the following means.

In reference [1] it is stated that the effective subjective noise improvement obtained by fitting a compandor is given by:

$$\frac{S}{6} + \frac{2N}{3} - \frac{5U}{6}$$

where S (dbm0) is the speech level, as mean power while active, referred to a point of zero relative level; N (dbm0p) is the noise level on the circuit without the compandor, referred to a point of zero relative level; and U (dbm0) is the unaffected level, referred to a point of zero relative level.

Taking $S = -18$ and $U = -6$, the equivalent subjective noise improvement = 28.7 db for Cases 4 and 9 and 26.0 db for Cases 5 and 10. These values reduce the circuit noise levels respectively of −46 and −42 dbm0p to subjectively equivalent values of −74.7 and −68.0 dbm0p (34 and 158 pW).

ANNEX E

(Geneva, 1964; quoted in Recommendation P.14)

EVALUATION OF THE EFFECTS OF LONG DELAYS AND ECHO SUPPRESSORS IN TELEPHONE COMMUNICATIONS

(Contribution by the Delegation of the United States of America)

I. Introduction

During the period from January 27 to April 25, 1964, tests to determine the effects of group propagation time on international telephone calls were conducted in co-operation with the British and French Telephone Administrations, and with the knowledge and advice of the Federal Communications Commission, the National Aeronautics and Space Administration, and the Communications Satellite Corporation. In this test, data were obtained by interviewing more than 3000 telephone users after the completion of calls between New York and either London or Paris over underseas cable circuits with various amounts of added delay. This report summarizes the results from these interviews, together with the most significant findings of a series of previous experiments conducted by the Bell Telephone Laboratories with users of its P.B.X. at Murray Hill, New Jersey. The purpose of the report is to make available information now considered most meaningful in judging the impairment of telephone communications caused by group delay.

The group delays of interest are those involved in telephone connections spanning up to half the circumference of the world. Even if reasonably direct paths along the surface of the earth are provided, the round-trip delays can be as long as 250 to 300 milliseconds [1]. The use of orbiting satellite repeaters will introduce a very significant increase in path length. The actual delay introduced will depend on the positions of the ground stations relative to the satellite at any given time and is not uniquely defined by the height of the orbit. For the purposes of illustration, an average round-trip delay of 220 milliseconds can be assumed for a 6000-mile (about 9500 km) high " medium " orbit and 520 milliseconds for a 22 000-mile (about 36 000 km) high " stationary " orbit. These are the orbits of principal interest at the present time. Table I illustrates the delays that could be encountered on a 12 000-mile connection by five possible combinations of facilities.

TABLE I

Illustrative round-trip delays using alternative methods of providing a connection spanning a 12 000-mile great circle distance

Connection includes	Distance spanned on earth		Round-trip delay
	Terrestrial portion	Satellite portion	
	Miles	Miles	ms
No satellite	12 000	—	240
1 medium hop	7 500	4 500	370
2 medium hops	3 000	9 000	500
1 stationary hop	6 500	5 500	650
2 stationary hops	1 000	11 000	1060

[1] Throughout this discussion, the delay values are for a round-trip between ends of a connection.

It shows the range of delays of principal concern in planning the tests reported here and provides perspective for the interpretation of the results presented.

While attention is focused on the delay of these connections, it must be kept in mind that the acceptability depends on the integrated effects of many factors. There is the effect of the delay itself in increasing response times and double talking, and slowing down the interchange of information. The necessary presence of an echo suppressor introduces interacting effects depending on the ability to interrupt, the amount of residual echo and the speech mutilation caused by interruptions. The importance of these effects depends very strongly on the dynamics of particular conversations. The need for an evaluation with natural conversations under realistic circuit conditions led first to tests with normal users of the Murray Hill P.B.X. and then to tests with users of underseas cable circuits. The underseas cable tests appear most significant in terms of the size of the sample of users and the realism of the test conditions and will be discussed first.

II. Underseas cable tests

a) *Test description*

Two non-TASI cable circuits from New York to Paris and two from White Plains to London were equipped with special echo suppressors and added delay. The special echo suppressors were used in place of the 1A and, in London, the 6A echo suppressors regularly used on the circuits. Delay, using Echo-Vox delay units, was placed in the outgoing path from the U.S.A. as shown in Figure 1. Four values of total round-trip circuit delay were used: 90 ms, 300 ms, 600 ms and 800 ms [1].

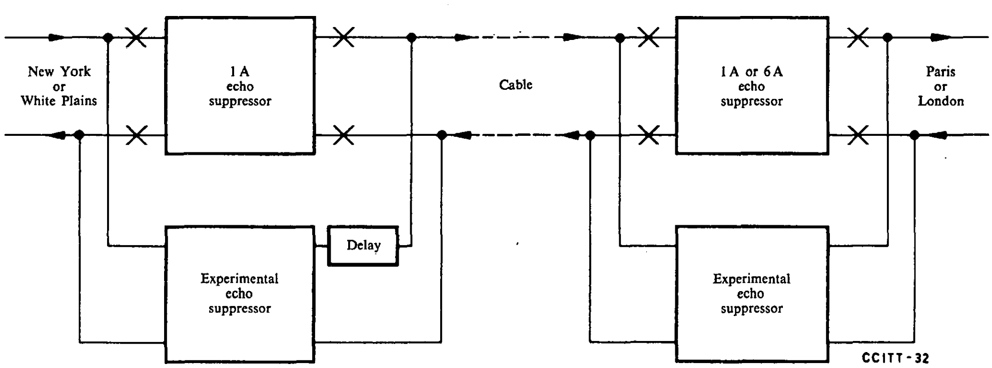


FIGURE 1. — **Placement of echo suppressors and delay**

[1] The 90 ms minimum value is the sum of the 70 ms inherent round-trip delay of the cable plus 20 ms, which is the minimum delay introduced by the Echo-Vox. The Echo-Vox delay units were used on all calls including the control, minimum delay, calls.

Two experimental echo suppressors, designated the B and the L, and modified British 6A echo suppressors were used on the White Plains to London circuits; the B, L and modified Bell System 1A echo suppressors were used on the two circuits from New York to Paris. The echo suppressors were rotated on the circuits so as to present circuit bias of the results.

The four circuits had essentially the same transmission characteristics such as bandwidth, noise and transmission loss. The 3 db down points of the bandwidth were at about 300 and 3100 c/s. The average noise level was about 42 dbrnc at the 0 transmission level point [1]. The circuits were given special access codes and were used as first choice circuits for outgoing calls from New York City to London and to Paris during the hours of 9.00 a.m. and 2.00 p.m., New York time, on weekdays. The traffic placed over the circuits was representative of the total traffic occurring during the test hours.

The operators using the circuits had no knowledge of the echo suppressors or delays being used, and they were given no special instructions other than those on the use of the specially coded circuits. The operators noted on the call ticket the access code used in addition to the normal information on party identification, call timing, etc.

The parties using the circuits were called back and interviewed by specially trained personnel in New York, London and Paris as soon as possible after call completion. This usually took less than 10 minutes. The general form of the questionnaire is shown in Figure 2. Every effort was made to use the same or equivalent forms and techniques in New York, London and Paris. For the data presented here, no person was interviewed more than once. The interviewers had no knowledge of the circuit conditions employed.

Each call was checked for transmission quality, and vu meter readings of the near and far end talkers were taken by U.S.A. plant personnel. The circuit noise and loss were measured prior to the test period each day. The circuits remained stable throughout the tests with few adjustments required.

[1] "Dbrnc" is "dbrn" as measured with a Western Electric 3A noise measuring set with C message weighting.

INTERVIEW QUESTIONNAIRE

1.	I am calling from the telephone company. We are making a study of the quality of our overseas connections. Do you have a moment to answer a few brief questions?	YES NO
2.	Our records show that you made/received an overseas telephone call to London/Paris/ from New York a short while ago. Is that correct?	YES NO
3.	Did you or the person you talked to have any difficulty talking or hearing over that connection? (If difficulty, probe and record verbatim—distinguish between called and calling parties).	NO DIFFICULTY SOME DIFFICULTY
4.	Which of these four words comes closest to describing the quality of that connection: Excellent, Good, Fair, or Poor?	EXCELLENT GOOD FAIR POOR
5.	About how many times a month or a year do you talk on the telephone with someone in London/Paris/New York? (If first call, skip to Question No. 7).	BY MONTH BY YEAR
6.	On your usual call to London/Paris/New York which of these four words comes closest to describing the quality of the connection: Excellent, Good, Fair, or Poor?	EXCELLENT GOOD FAIR POOR
7.	Do you have any other comments you would like to make about your overseas telephone service?	YES NO

FIGURE 2

b) *Echo suppressor description*

The pertinent parameter values of the echo suppressors used in the test are shown in Table II.

Of the four echo suppressors used, the modified 6A and 1A echo suppressors are very similar. Both combine the suppression control and differential action into one circuit [1]. This does not allow a loud talker to remove suppression caused by a weak talker until the suppression hangover has expired. The modification in the 1A consisted of a 10-db higher sensitivity at 1000 c/s and increased suppression loss. The modified 6A had a changed sensitivity frequency characteristic. This change was observed by the British Post Office to produce suppression for lower level speech sounds with no increase in noise operation.

[1] For a more complete description of the modified 1A and the B echo suppressors, see reference [1] at the end of this report.

The L echo suppressor is distinguished by its action during double talking, or the period when both parties to a conversation are simultaneously speaking. When this occurs the suppression is removed immediately and 6 db attenuation is inserted in both the receive and transmit paths to reduce the echo which is unsuppressed at this time. The hangover on the break-in circuit is about 30 ms. Another unique feature of this echo suppressor is the hysteresis effect present in the suppression and break-in circuits. The signal level required to release from the suppression mode is about 9 db below that required to operate.

The B experimental echo suppressor is similar to the L in that suppression is removed immediately when double talking is detected. It is, however, unique in the long break-in hangover time (200 ms) and in the insertion of a speech compressor during double talking to reduce the echo.

TABLE II

Characteristics of the echo suppressors (nominal or typical measured values)

	1A	6A	B	L
Suppression				
Operate sensitivity [a]	−41	−26	−31	−32
Release sensitivity [a]	−44	−29	−33	−41
Pick-up time [b]	10	4	8	1
Hangover time	50	50	50	30
Bandwidth	Peaked at 1000 c/s 20 db down at 500 and 2000 c/s	Peaked at 3500 c/s 7 db rise from 1000-2000 c/s	Flat over the voice band	Flat over the voice band
Break-in				
Operate sensitivity	−41	−26	−26	−36
Release sensitivity	−44	−29	−27	−41
Pick-up time	50 [c]	50 [c]	10	1
Hangover time	20	50	200	30
Bandwidth	Same as for suppression			

	Loss during double talking
1A	None
6A	None
B	6 db attenuation in transmit path. Speech compressor in the receive path with 0 db attenuation for a −40 dbm signal and 18 db attenuation for a 0 dbm signal at the zero relative level point.
L	6 db attenuation in both the receive and transmit paths.

[a] All sensitivity values are given in terms of the power level of a 1000 c/s signal in dbm at the zero relative level point.

[b] All times are in milliseconds and are measured using a 1000 c/s signal level 3 db above the operate sensitivity.

[c] In the presence of a suppressing signal in the receive path, break-in does not occur until the suppression hangover time expires.

c) *Results*

General. — The primary measures of the subjective effects of transmission delay are theans wers to Questions 3 and 4 of the interview which concern difficulty in talking or hearing and ask for a quality rating. These questions were asked of persons in four different " cities ", i.e. those who received calls in London and Paris and those who made calls in New York terminating in London and Paris; who used circuits with one of four different echo suppressors: 1A, 6A, B and L; and who had one of four values of delay: 90, 300, 600 and 800 ms. About 60 interviews were completed for each of the combinations of city, echo suppressor and delay. A total of about 3000 interviews resulted.

Percentage difficulty. — The over-all results of the answers to Question 3, plotted as a percentage of interviews reporting difficulty, are shown in Figure 3. Each point represents about 750 completed interviews [1]. The results for all cities and echo suppressors are combined for each delay. Figure 3 shows that difficulty in talking or hearing increases with delay and that the slope of this function is greater for higher delays.

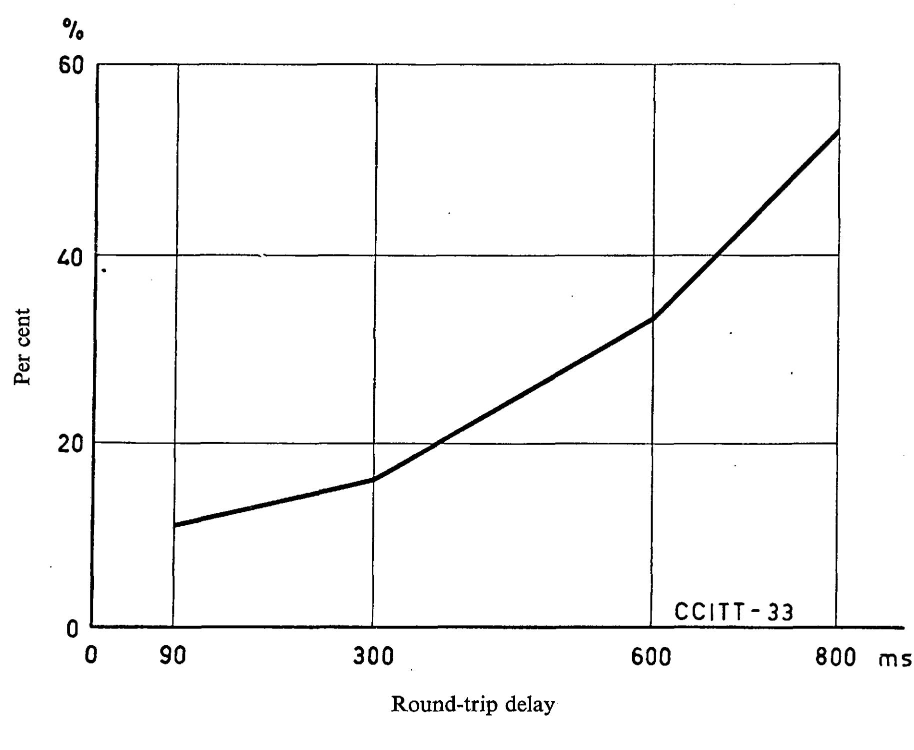


FIGURE 3. — **Percentage of interviews reporting difficulty—combined over cities and echo suppressors**

[1] Of approximately 3000 completed interviews, some 300 are interviews on calls where both the United States of America and European parties were interviewed and both had difficulty, 500 are interviews where both parties were interviewed and one had difficulty, 1000 are interviews where both parties were interviewed and neither had difficulty, and 1200 are interviews where only one party to a call was interviewed.

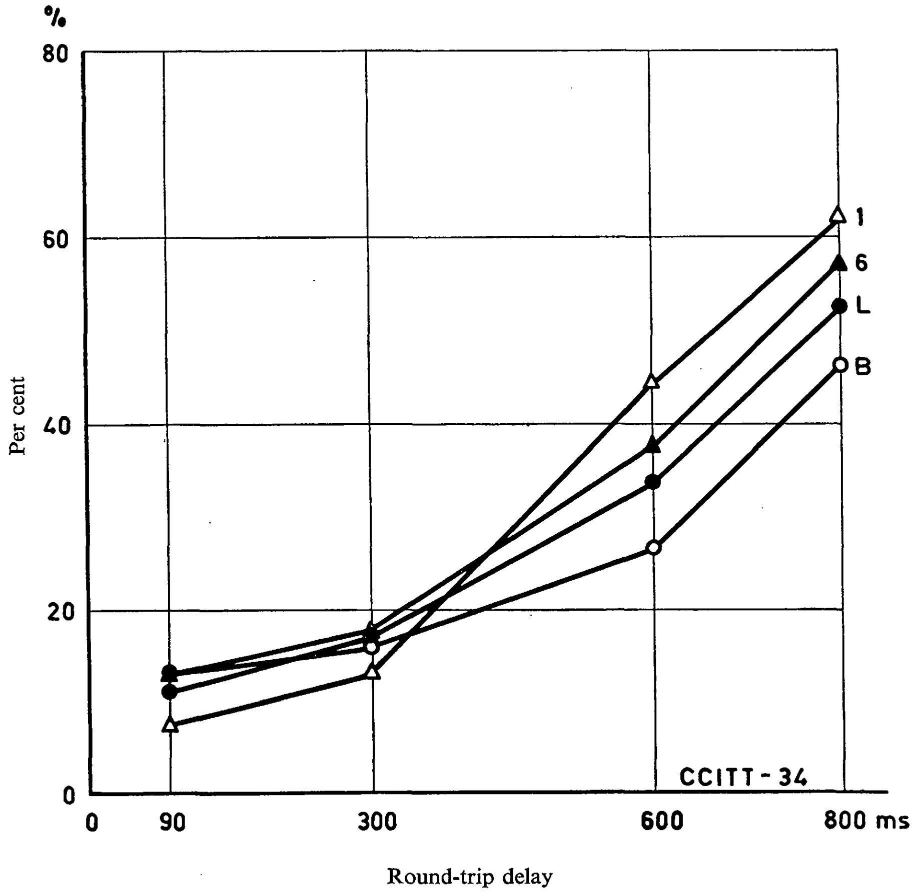


FIGURE 4. — **Percentage of interviews reporting difficulty—comparison of echo suppressors—combined over cities**

Figure 4 shows the percentage difficulty broken down for each echo suppressor separately. If the values shown in Figure 3 are taken as the true mean of the population of all calls, then all but one of the values shown in Figure 4 fall within the 95% confidence limits on the mean. The implication is that the differences among echo suppressors must be interpreted with caution since it is conceivable that the differences are due to chance sampling. It will be shown later that there is a more significant difference among echo suppressors in the specific types of difficulties they produce.

The differences between cities, when combining over echo suppressors, are small, irregular and not very significant. The similarity of the results from each of the cities at each delay holds not only for the percentage difficulty but also for the rating of quality (Question 4). Because the location of the interview is not an important factor, data from all cities are combined in the data analysis.

Rating. — The rating of transmission quality (Question 4) as a function of delay is shown in Figure 5. Data from all suppressors are combined. The percentage of Excellent ratings decreases almost linearly with delay while the percentage of Poor plus Fair ratings closely resembles the percentage difficulty curve (Figure 3).

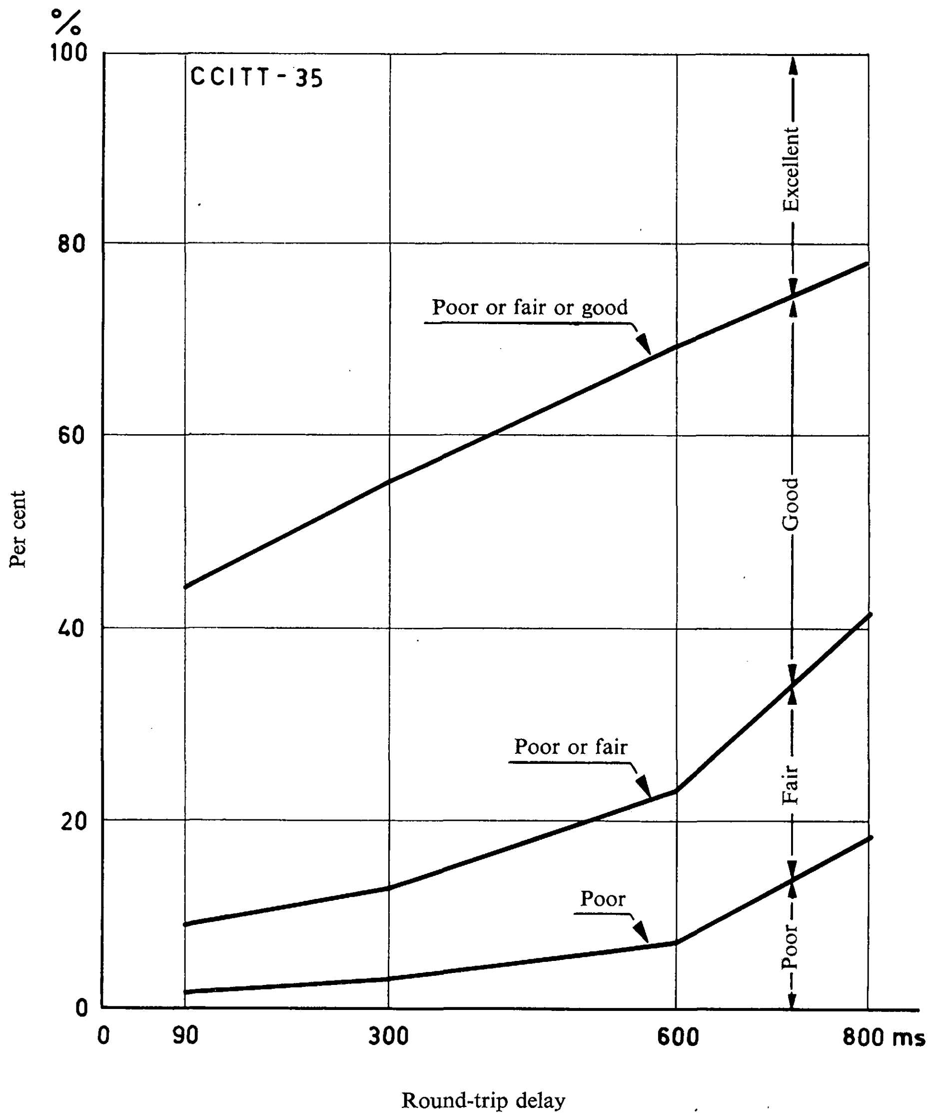


FIGURE 5. — **Rating of the calls—combined over cities and echo suppressors**

In Figure 6, the percentage of interviews reporting a Poor or Fair response to Question 4 is shown for each echo suppressor separately. The Fair plus Poor curve is similar in shape to the percentage difficulty curves of Figure 4. As in Figure 4 the differences among suppressors might conceivably be due to chance, but taken together the results at 600 and 800 milliseconds on both figures suggest strongly that the B and L suppressors produce somewhat less difficulty than the 1A and 6A.

(Annex E)

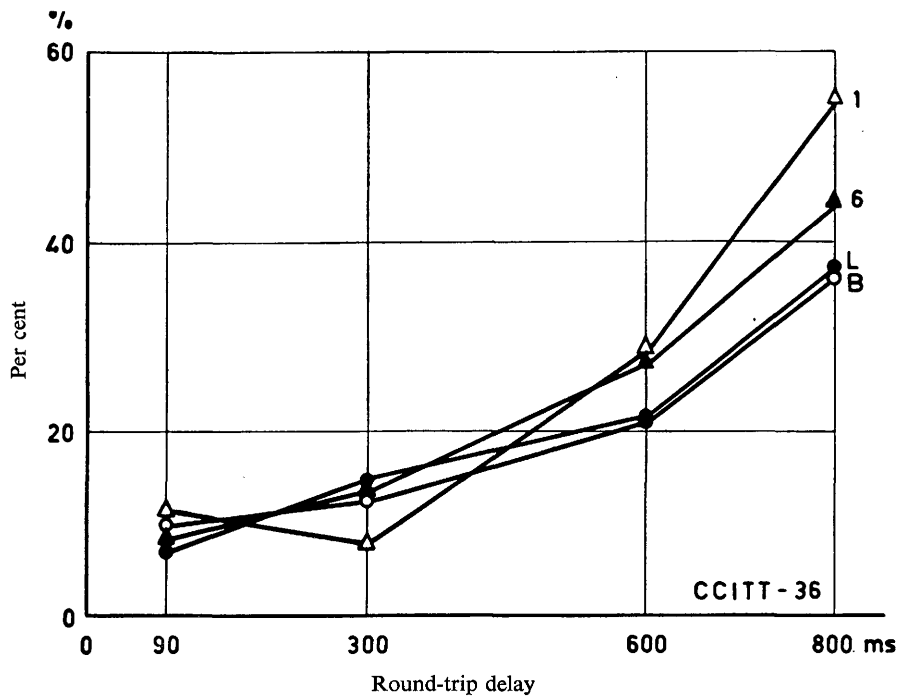


FIGURE 6. — **Percentage of persons interviewed who rated the call Poor or Fair—comparison of echo suppressors—combined over cities**

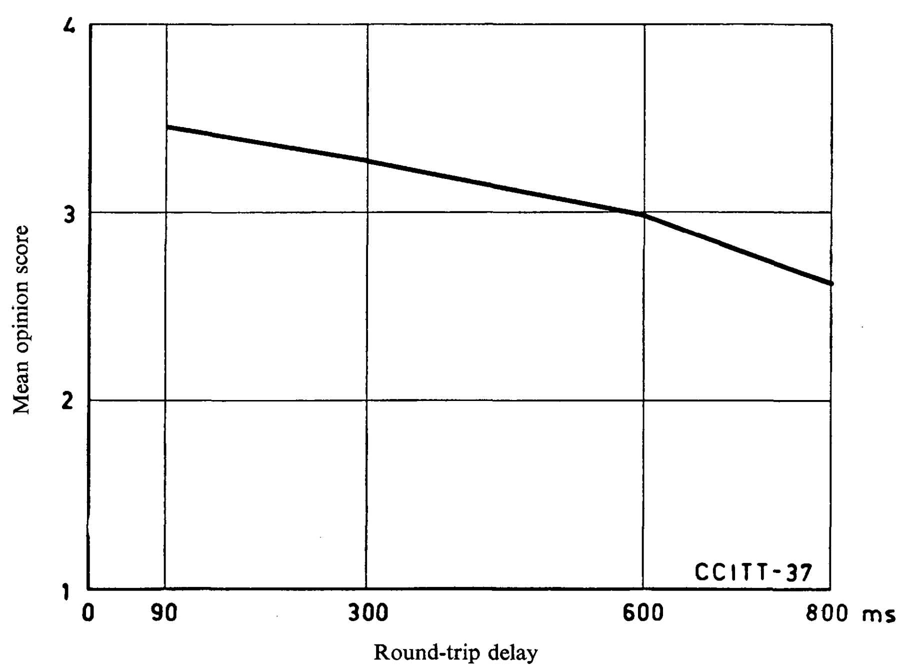


FIGURE 7. — **Mean opinion score of the calls—combined over cities and echo suppressors**

Mean opinion score. — The mean opinion score (MOS) of the interviews is shown in Figure 7 for all suppressors and in Figure 8 for each suppressor separately. The MOS is the weighted average of the responses to Question 4: Excellent is weighted 4, Good 3, Fair 2, and Poor 1. As would be expected, the MOS reflects much the same findings as the Poor plus Fair curves of Figure 5, inverted in direction and reduced in range.

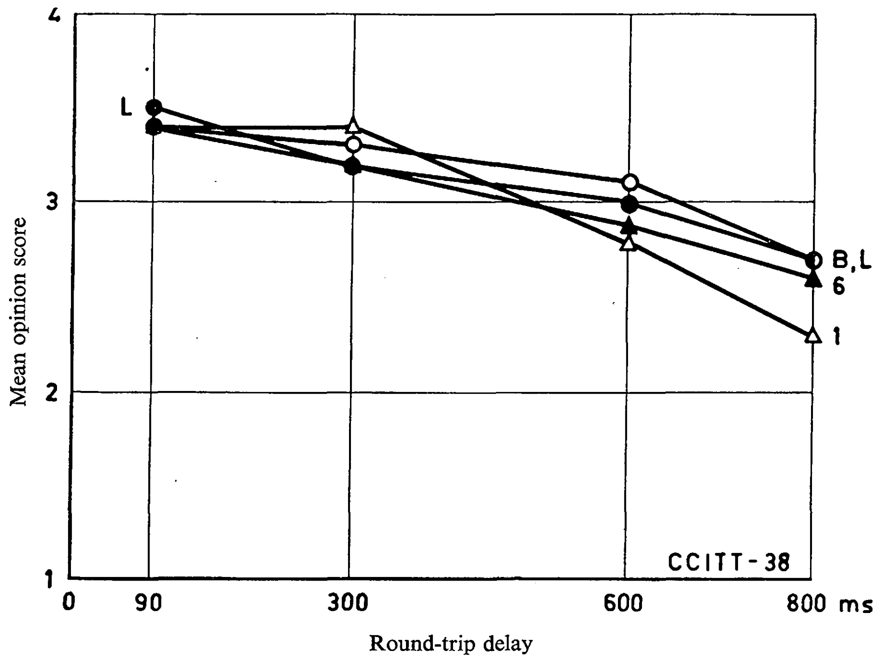


FIGURE 8. — **Mean opinion score of the calls—comparison of echo suppressors— combined over cities**

Customer flash. — As on all overseas calls, the users of the test circuits could request another circuit by flashing the overseas operator. The percentage of customer flashes increased from less than 1 % at 90 ms to about 4 % at 800 ms. The small number of customer flashes makes a further break-down by test conditions unwarranted.

Comments. — Those persons who replied " yes " to Question 3 also described the difficulty they experienced. The difficulties, as stated in the users' own words, are not always easy to classify. A three-man committee composed of persons from the Bell System, National Aeronautics and Space Administration, and Communications Satellite Corporation, without any knowledge of the echo suppressor or delay relating to the comments, classified them into 18 different types. The percentage, by type, of the total comments for all delays is shown in Figure 9. Occasionally more than one type of comment resulted from one interview.

The comments made by the users are also shown divided into comments they made about their difficulties (near end) and comments made about difficulties experienced by the other party (far end).

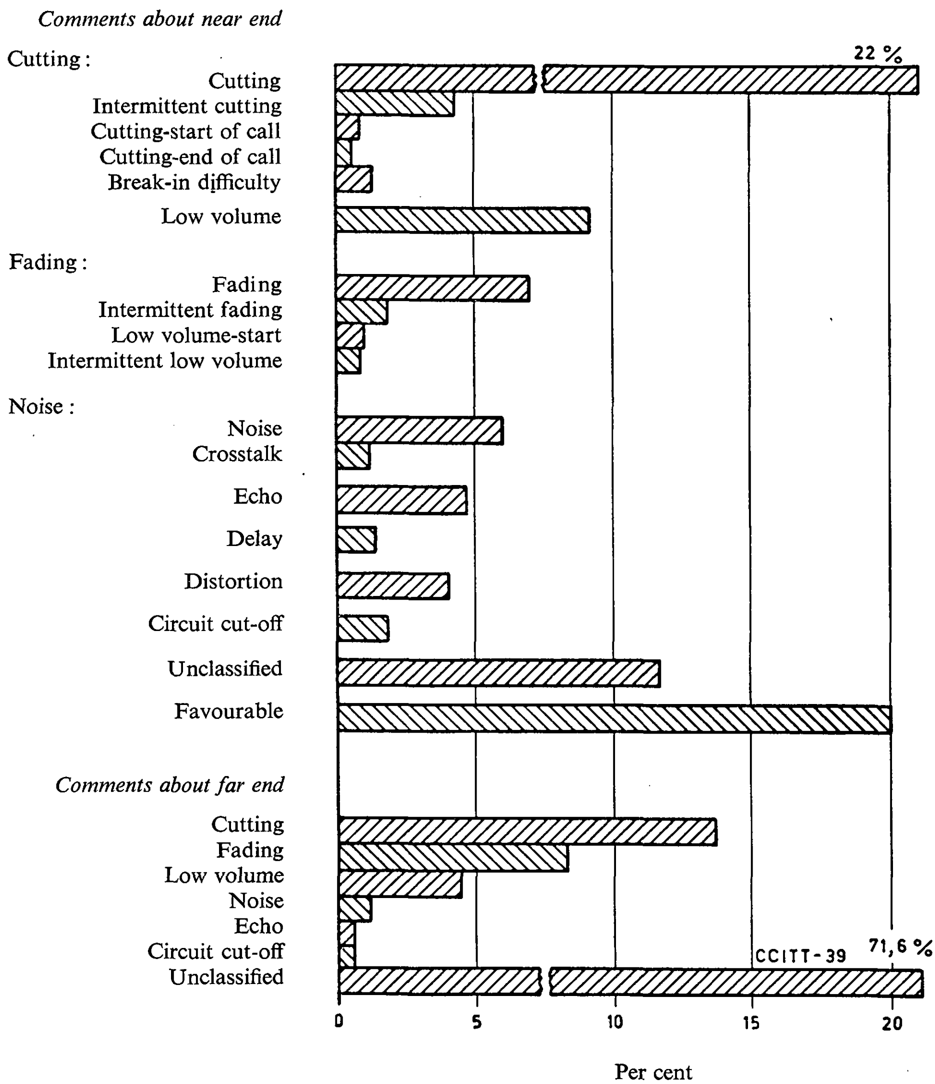


FIGURE 9. — **Percentage of total comments by type**

The relationship of comments to echo suppressors is shown in Figure 10. The types of comments are those which are grouped in Figure 9. For each comment, four points are shown—one for each delay. The points represent the percentage of *interviews* which both had a " yes " answer to Question 3 *and* which had a comment of the type shown, except for the favourable comment points. In all but one case, these came from interviews in which no difficulty was reported.

Some things are readily apparent from Figure 10. First, favourable comments generally decrease with delay; second, comments on cutting generally increase with delay; third, echo and delay comments are too few to chart trends. There is no clear pattern to most of the rest of the comment types. There is a distinct difference in the number of comments on cutting for B compared with the other echo suppressors [1]. It is believed that this is due to the difference in break-in hangover times and the insertion of the speech compressor during double talking. Although the comments on cutting are less for B, the total number of the other types of comments are somewhat more than for the other echo suppressors.

C — Cuttings	E = Echo	U = Unclassified
V = Low volume	D = Delay	O = Far end (all types)
F = Fading	M = Distortion	G = Favourable
N = Noise and crosstalk	T = Circuit cut-off	

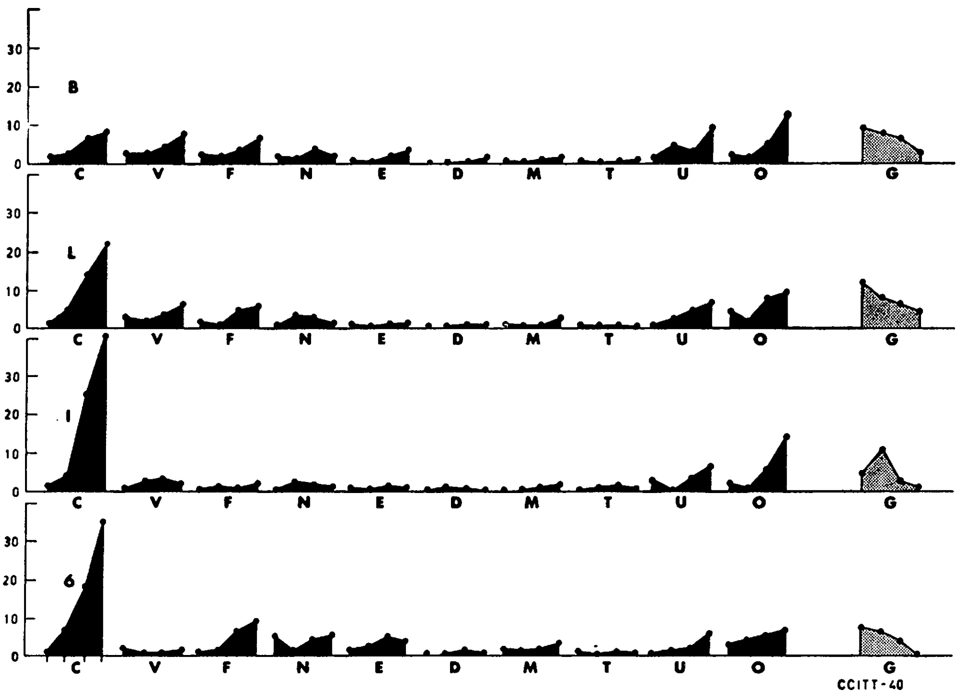


FIGURE 10. — **Comment types by echo suppressor combined over cities**

Summary of results. — Although not all measures of the subjective reaction to transmission delay have been exhaustively examined, some tentative conclusions can be drawn:

1) There appears to be no value of delay which provides a clear division between acceptable and unacceptable service. Rather, all measures of degradation increase smoothly with delay, and furthermore the rate of increase also increases with delay.

[1] The relative significance of the various categories of comments has not been established.

2) As the delay increases, user comments indicate a difference in the conversational effects produced by the various echo suppressors.
3) The experimental L and B echo suppressors showed an advantage for the longer delays, but no single suppressor showed a clear advantage over the others for all delays tested.

III. Murray Hill P.B.X. tests

a) *Test description*

A service simulation and monitoring facility, called Sibyl, is available at the Murray Hill Laboratory to permit the insertion of test conditions in the otherwise normal telephone connections through the Murray Hill P.B.X. [2]. Any of the P.B.X. extensions can be intercepted at the local distributing frame and routed through this facility. In the case of experiments with delay, echo and echo suppressors, the appropriate simulation apparatus was introduced in one call at a time in the Sibyl test room. This was done in a way which did not give the test subject a cue that a particular call was being handled in a special way. The arrangement of the simulation apparatus including hybrids, echo suppressors and delay units is shown in Figure 11. In the particular experiment to be reported—Experiment 3 described by Riesz and Klemmer [3]—the echo suppressor was the experimental B unit described above.

The subjects participating in the experiment were told that some of their calls would be routed over an experimental circuit. They were instructed that if they found any circuit " unsatisfactory " for normal telephoning they could dial the digit " 4 " without hanging up or breaking the connection and the standard circuit would be restored.

A total of 56 test subjects continued through the whole test and are included in the results reported here. After a week of testing at 50 milliseconds delay, the group was divided into four sub-groups of about equal size and calling rate and these where exposed to the sequence of delays shown in Table III. With this sequence, the subjects in a sub-group were never exposed to a delay longer than that assigned from the fifth to the fourteenth week. This avoided any possible interaction which might otherwise inflate the rejection rates at the lower delays.

Table III

Sequence of delays for each sub-group

(Table entries are round-trip delays in ms)

	Week of experiment															
Sub-group	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
1	50				600											
2		200														
3			400													
4			600													

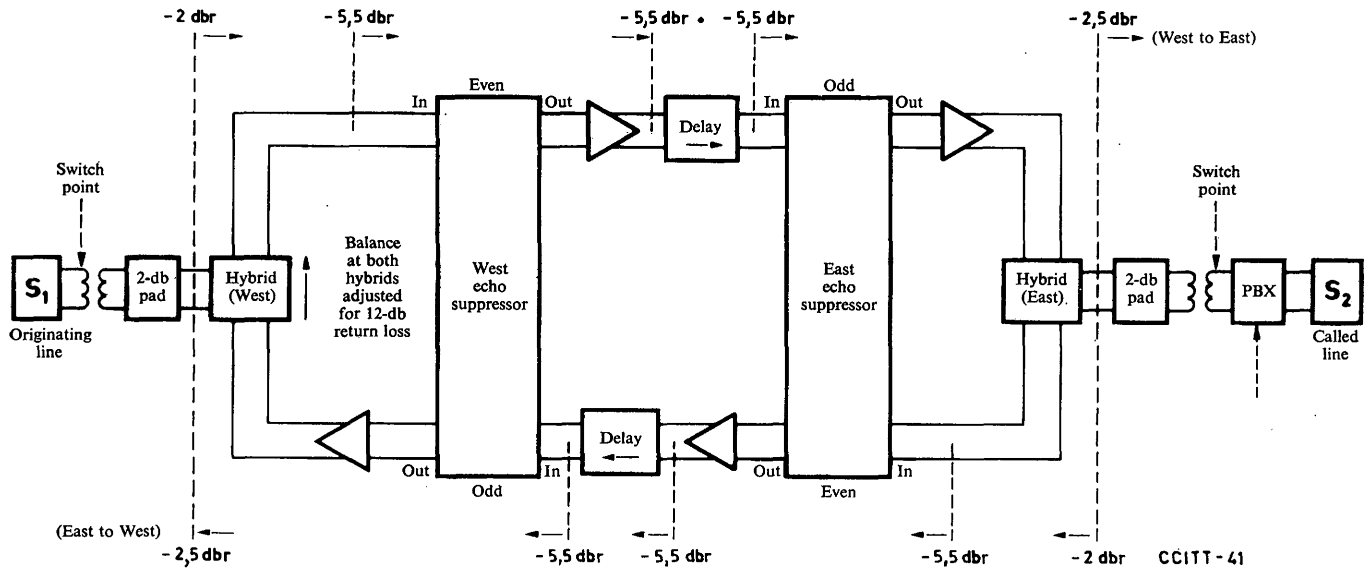


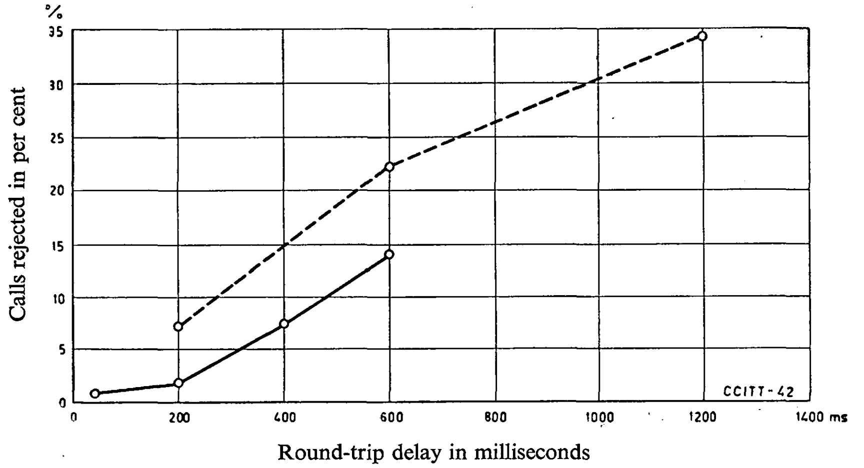
FIGURE 11. — **Satellite simulator circuit showing method of introducing delay and echo suppressors**

The level at the two-wire connection of the hybrid is arbitrarily designated as —2 db test level. The levels at other points use the same reference. The over-all loss between switch points is 4.5 db.

b) *Results*

The principal results of the experiment are shown in Figure 12, where rejection rate is plotted as a function of delay. In the solid curve the 50, 200 and 400 milliseconds points include data only from sub-groups 1, 2, and 3, respectively. The results from weeks 5-16 for sub-groups 1 and 4 were combined to obtain the 600-millisecond point. The rejection rate rises almost linearly from 2% at 200 milliseconds to 14% at 600 milliseconds.

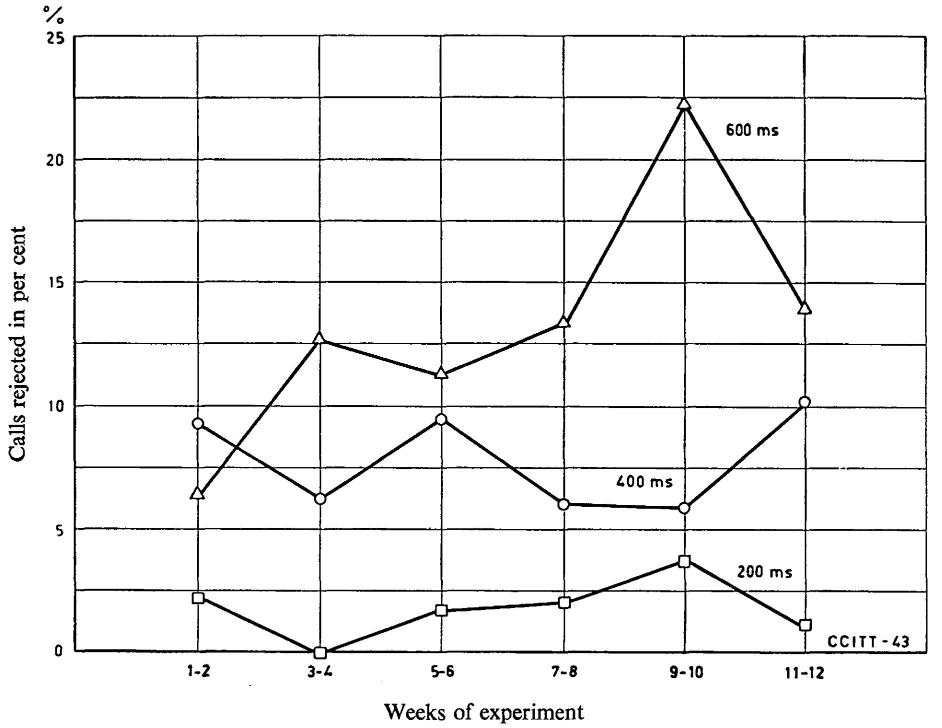
The combined sub-groups 1 and 4 were given an added week of mixed 200, 600 and 1200 milliseconds delay, and the dashed curve plotted in Figure 12 shows the results. The 34% rejection rate at 1200 milliseconds is consistent with a linear extrapolation of the solid curve, but the rejection rates at 200 and 600 milliseconds have been increased.



Each point represents separate groups of 12 to 25 subjects, except that the 50-ms group (group 1) was later given 600 ms delay and is included with the subjects initially assigned to 600 ms (group 4). The dashed line shows the results of exposing this combined group (1 and 4) to mixed delays for one week. Each point of the solid curve represents 376-774 calls. Each point of the dashed line represents 53-68 calls.

FIGURE 12. — **Rejection rate as a function of delay**

Even before an interaction of delays was introduced in the last week of the test, there was a trend toward increasing rejection rate with time. The test data are plotted in Figure 13 as a function of weeks of experience, with separate curves for 200, 400 and 600 milliseconds. The first point on each curve represents the first two weeks each group was at its maximum delay, and so forth,



Data from all groups combined as from the start of their respective weeks at the maximum delay to which each was exposed (200, 400, 600 ms). The added week of mixed delays to 1200 ms is not included. Each point represents 47 to 163 calls

FIGURE 13. — **Rejection rate as a function of weeks at maximum delay**

The 600-millisecond curve is quite erratic, but it does suggest that users become less tolerant of the effects of delay as they gain experience. Inconclusive though the results may be, they raise doubts about the use of results based on a single exposure of uninformed subjects to predict long-term acceptability of long delays.

It is interesting, but not particularly surprising, that the rejection rates obtained in this experiment are much greater than the customer flash rate which occurred in the underseas cable tests. The rejection process is easier and has benefits well known to the test subject. On the other hand, the percentage rejected is about half the " yes difficulty " interview responses in the underseas cable tests. This suggests that the rejections did not occur in every case of difficulty on a P.B.X. connection, and is consistent with the results of interviews of P.B.X. customers immediately after a sample of calls. Nearly half of those who did not reject a test call at 600 milliseconds noticed something different about that call, such as echo, chopping, noise, low volume or delay.

IV. Limitations and applications

Every effort was made in the planning and execution of both the underseas cable and Murray Hill P.B.X. tests to obtain realistic and unbiased results. Nevertheless, each has inherent weaknesses, which must be recognized in interpreting and applying the results.

A major factor in the P.B.X. tests is that it was practical to obtain data from only a small sample of users. In addition, these users were not exposed to the same range of important transmission parameters as expected in real international connections. For example, the return losses were set uniformly at 12 db, a value one standard deviation below the average return loss expected in the United States. The noise was negligibly low and the loss from the talkers to the echo suppressors was asymmetrical. These same transmission conditions can occur in the field but the test conditions represented a more critical than average situation. This fact, together with the relative ease of rejecting the test condition to obtain a very good local connection, leads to doubts about applying the results to overseas connections.

The underseas cable tests were designed to obtain data from a large sample of users on actual overseas connections, thereby overcoming the more serious limitations of the P.B.X. data. As far as is known, this information is the most extensive and realistic that has been obtained on the subject of delay. However, it too has limitations that must be recognized. As explained in the test description, user reaction was determined by interview only after the first exposure to any one of the test conditions. With the limited number of simulated satellite circuits available to serve traffic, it was not practical for any significant number of users to become experienced in the use of long delays through multiple exposures. It was also considered unwise to inform the public that some overseas calls were being served differently from others. Hence, from the users' standpoint the situation was quite different from that of the P.B.X. tests and from what will prevail when customers know that both cable and long-delay satellite circuits are being used to complete their calls.

In both tests, the basic standard of comparison available to the users was a high-quality circuit—a nearly normal P.B.X. connection or an underseas cable connection between major cities. The use of this comparison is consistent with the C.C.I.T.T. goal of improving the transmission quality of international telephone service. However, the resulting data do not permit a direct prediction of relative customer acceptance where the only alternative to a long-delay circuit is one inferior in other respects; for example, a high-frequency radio circuit.

A complication to be expected in the future use of long-delay circuits is the need for extensions beyond the gateway cities in order to serve a large portion of the total commercial traffic. Under present operating arrangements, these extensions will frequently include additional pairs of echo suppressors. It is known that tandem pairs of echo suppressors are undesirable with normal delays; it is expected that the undesirable effects will be increased when long delays are involved. This complication was avoided in the tests reported here, and no data now exist for predicting what will happen to circuit acceptability. The same can be said for the mixed combinations of suppressors and non-uniform characteristics of national extensions that can occur when the ends of circuits are controlled by different Telephone Administrations.

In summary, present data are not considered completely adequate for a confident prediction of customer acceptance when many long-delay satellite circuits are put into service on routes which parallel underseas cable circuits. Further tests should be conducted to evaluate the remaining factors. In the meantime, the data do make evident delay ranges where the effect of delay is either negligibly small or intolerably large. Administrations and operating Agencies can weigh this evidence along with other transmission performance characteristics, economic considerations, and the need for diverse paths to serve given routes in selecting among alternative facility plans.

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ANNEX F

(Geneva, 1964; quoted in Recommendation P.14)

TRANSMISSION PERFORMANCE OF TELEPHONE CONNECTIONS HAVING LONG PROPAGATION TIMES

(Results of real-traffic tests conducted by the British Post Office)

CONTENTS

1. Introduction

Mean one-way propagation times appreciably greater than 150 ms will occur on some world-wide telephone connections over submarine cable systems and on certain contemplated satellite systems. Conversations over such connections can be degraded on account of the following factors:

a) The propagation time itself, which can disrupt the normal alternation of talking and listening roles;

b) Echo, imperfectly suppressed by the echo suppressors, which may be objectionable while talking;

c) Mutilation of speech due to functioning of the echo suppressors causing unnecessary interruptions, which can be objectionable while listening.

These three factors are being studied by laboratory tests, both separately and in combination with each other, when associated with various circuit losses, noise levels, distortions and return losses. Results so far have shown that, even in the absence of factors b) and c), the presence of a mean one-way propagation time of even only 250 ms can readily be detected in conversation by many pairs of subjects. The confused situations that develop depend a great deal upon the nature of the conversation and the personal characteristics of the subjects. Under certain laboratory conditions, and with 250 ms mean one-way propagation time, a confused situation developed, on average, every two minutes, the rate of incidence increasing roughly proportionally as the propagation time was increased. Additional information is given in reference [5].

The above-mentioned factors will, of course, be present in addition to the normally encountered amounts of loss, noise and distortion and their presence is particularly important in the study of the effects of factor b). Recent work on the suppression performance of echo suppressors has shown that the detectability of imperfectly suppressed echo, when the mean one-way propagation time exceeds about 150 ms, is related to the quantity E which is defined as the amount of additional return loss, equal at all frequences, needed to render the imperfectly suppressed echo just undetectable. E is given by the expression:

$$E = T_N - N + K_S - b - w$$

where T_N is the maximum level of white noise in the telephone channel that will just fail to cause the echo suppressor to operate in the absence of any signal in the opposite direction;

N is the level of noise actually present on the circuit and audible to the talker. Both T_N and N are expressed in db relative to one milliwatt and referred to the respective points of zero relative level; psophometric weighting is applied. If room noise is not reasonably low an allowance may have to be made for it;

K_S is a number, referred to as the " suppression coefficient ", that depends upon the characteristics of the echo suppressor (including the sensitivity frequency characteristic of the control path and the operate time) and, to some extent, upon the speech level. For British Post Office echo-suppressors 6A (which comply with C.C.I.T.T. Recommendation G.151); K_S is about 10 db for the most unfavourable level of speech;

b is the return loss suitably averaged over the frequency band (the values in the region of 2500 c/s up to the upper edge of frequency range transmitted by the channel are particularly important);

w is the transmission loss at 800 c/s of the circuit in which the echo suppressors are fitted; w is measured between two-wire points of this circuit.

Application of equation (1) to the conditions that are likely to apply in practical connections shows that many situations will occur when the imperfectly suppressed echo will be readily detectable. Of course echo that is detectable is not necessarily objectionable but, when E is as large as 20 db, it is likely that a serious proportion of subjects would find the condition objectionable. Experiments have shown that K_S can be reduced (thereby improving the suppression) by adopting a more nearly level, or even rising somewhat with increase of frequency, frequency characteristic for the control paths.

An observer monitoring conversations over laboratory circuits containing echo suppressors is very much aware of the occurrence of mutilations, as many as 8 per minute being noticeable (considering speech from both ends); the subjects taking part in the conversation, however, seem to be less aware of these, no doubt due to many occurring while a subject is himself talking and therefore not attending carefully to what reaches his ear. Nevertheless there is no doubt that echo-suppressor mutilations are readily detectable under practical conditions when the one-way propagation time is of the order 150 ms or more and, furthermore, these can have a serious effect upon the customers' difficulties.

It is not possible, on practical telephone connections, to render factors b) and c) completely undetectable except by use of entirely separate paths for the two directions from customer to customer. The problem remains one of achieving the best possible compromise between the effects of these two factors because, to a considerable extent, improving factor b) will tend to make c) worse and vice versa. Unfortunately the relative importance of the two factors cannot readily be determined purely from laboratory tests; recourse must be had to studies with real traffic. Clearly great care is needed in selecting the conditions for conducting such tests because misleading results will be obtained if these differ too much from those under which it is desired to predict the performance. The real traffic tests so far completed by the British Post Office are discussed separately below.

2. Results of tests using real traffic

2.1 *Calls between the United Kingdom and Australia*

The tests, in co-operation with Overseas Telecommunications Commission (Australia) and the Australian Post Office were commenced on January 3rd, 1964, using traffic over London to Sydney and London to Melbourne circuits. Three test conditions were applied to each of the through-group circuits Nos. 19, 20 and 21; the conditions were reallocated to the circuits differently each day so that any difference between the circuits would not bias comparisons between test conditions. The three test conditions were as follows.

A. Circuit as engineered with normal echo suppressors, the mean one-way propagation time being about 130 ms;

B. Circuit with the normal echo suppressors replaced at each end by modified suppressors having greater threshold sensitivities and different control-path frequency characteristics;

C. As B, but with 200 ms additional propagation time in each direction (making about 330 ms total mean one-way propagation time).

2.1.1 *Selection of calls*

Calls outgoing from the United Kingdom were selected as follows:

All U.K.-Australia callers were first connected to a booking operator who recorded on a ticket the particulars of the required call, including the called number and the time the call was required. Before releasing the caller with advice that he would be recalled, the booking operator confirmed with a time assignment operator that a " call vacancy " existed at the required time. If there was no vacancy the booking operator arranged an alternative time with the caller.

The tickets were then sent to the supervisor of the Australian Section, who extracted and directed to an observation supervisor all tickets for calls originating in the London director area and destined for Sydney or Melbourne numbers. The observation supervisor allocated the calls to the three special circuits and noted on the tickets the appropriate access code and, if required, the special Sydney area code to obtain certain selected routings in Australia.

Calls outgoing from Australia were also used and these were dealt with in an analogous manner.

2.1.2 *Observations*

The following observations were made:

a) Monitoring by specially trained operating staff of calls while in progress. These observers note the occurrence of requests for repetitions, adverse (and other) comments on transmission quality, confused situations due to the length of the propagation time and other relevant information, finally expressing an opinion on the over-all quality of the conversation according to the scale " excellent ", " good ", " fair ", " poor " or " bad ". These observations were made in London and in Sydney.

b) Engineering measurements which included speech levels (both directions being measured in London and in Sydney), noise levels and traces with a multiple pen recorder that enabled echo, mutilation and the structure of the conversation to be studied; in particular these enabled occurrences of confusion due to the length of the propagation time and the precise location of the mutilations to be studied.

In London, the monitoring observers were drawn from a team of six assistant supervisor telephone operating staff whose members were allocated to the three test circuits in a non-systematic manner, changes being made several times per day at suitable intervals in accordance with normal operating duty procedure. In Sydney, observers were drawn from operating staff experienced in assessing the quality of overseas circuits. These observers recorded their observations on special report sheets in accordance with instructions and indicated simultaneously with keys the occurrence of certain defined events. They were also provided with a key that would enable the extra propagation time (if present) to be removed if very severe confusion was thought to be due to its presence.

After each call over one of the circuits under test the controlling operator passed the ticket to the monitoring observers involved who extracted information concerning the call and recorded it in the report sheet. Calls were thus identifiable by the circuit number and the time of commencement of conversation. The monitoring observer was not made aware of the particular test condition in force on a given circuit. The allocation of the three test conditions to the three circuits was changed daily at 1400 GMT in conformity with a Latin square experimental design.

2.1.3 *Results*

The period January 3rd to January 15th was occupied in checking the observation equipment and in gaining experience in running the tests. Tests were run from January 16th to February 14th when certain changes in the tests gear were made and an opportunity taken to trace and correct certain echo suppressor faults; this period of testing will be referred to as Period 1 and contained 424 calls handled by the monitoring operators, 327 of the calls having been subjected also to the engineering observations. Testing was recommenced on March 9th and continued to March 25th when a break was made for Easter. It was intended to resume immediately after the holiday but two cable faults prevented starting again until April 13th. On April 22nd certain changes were made in the echo suppressors in Sydney to reduce their fault liability and this date is taken as the close of Period 2 which extended from March 9th. 439 calls were handled by the monitoring observers during Period 2 of which 269 were also subjected to the engineering observations. The period April 22nd to May 26th (referred to as Period 3) was devoted to further testing to determine whether the changes in the echo suppressors had any effect on the results. A cable fault interrupted this period from April 30th to May 13th.

In the following tables the test conditions and periods of testing are referred to as described above.

TABLE 1

Opinions by monitoring observers

(a) In London

Condition	A			B			C		
Period	1	2	3	1	2	3	1	2	3
Number of calls	161	149	80	155	165	86	108	123	53
Opinion:									
Excellent	15	5	0	12	3	5	3	3	1
Good	110	107	59	103	122	59	56	64	27
Fair	30	36	17	33	32	18	40	40	15
Poor	5	1	3	6	5	3	8	9	7
Bad	1	0	1	1	3	1	1	7	3
Mean score [1]	2.83	2.77	2.68	2.77	2.71	2.74	2.48	2.38	2.30
%G+F+P+B	90.7	96.6	100.0	92.3	98.8	94.2	97.2	97.2	98.1
%F+P+B	22	25	26	26	24	26	45	46	47
%P+B	3.7	0.7	5.0	4.5	4.8	4.7	8.3	13.1	18.9
%B	0.6	0	1.3	0.7	1.8	1.2	0.9	5.7	5.7

(b) In Sydney

Condition	A			B			C		
Period	1	2	3	1	2	3	1	3	3
Number of calls	125	114	97	139	121	72	97	110	45
Opinion:									
Excellent	4	8	2	2	5	3	0	2	4
Good	101	93	74	107	95	58	52	71	24
Fair	16	9	19	24	19	10	39	28	16
Poor	4	3	2	6	1	1	3	3	1
Bad	0	1	0	0	1	0	3	6	0
Mean score [1]	2.84	2.91	2.78	2.76	2.84	2.88	2.44	2.55	2.69
%G+F+P+B	97.4	93.0	97.9	98.6	95.9	95.8	100.0	98.2	91.1
%F+P+B	13	11	22	22	17	15	46	34	38
%P+B	2.6	3.5	2.1	4.3	1.6	1.4	6.2	8.2	2.2
%B	0	0.9	0	0	0.8	0	3.1	5.4	0

[1] Mean scores were calculated by scoring the opinions 'excellent', 'good', 'fair', 'poor' and 'bad', respectively, 4, 3, 2, 1 and 0 taking arithmetic means.

TABLE 2

Incidence of repetitions, confusions and mutilations

Event	Repetitions [1]			Confusions [2]			Mutilations [3]		
Condition	A	B	C	A	B	C	A	B	C
Period	2	2	2	1	1	1	2	2	2
All calls, first six minutes (or less)									
Number of calls	151	165	123	90	81	63	91	92	77
Events per six minutes	1.04	1.33	1.85	0.14	0.13	0.33	11.8	10.1	13.3
Calls lasting at least six minutes									
Number of calls	89	94	52	Not counted			42	55	29
Events per six minutes	1.01	1.26	1.69				11.7	10.6	11.3

[1] By both customers. — [2] Considering the whole conversation. — [3] U.K. customer's speech.

TABLE 3

Distributions of durations of calls

The percentages of calls lasting at least T minutes

Period 2

Condition	A	B	C
	%	%	%
$T =$ 1	100	100	100
2	98	98	98
3	95	93	94
4	76	74	68
5	68	63	56
6	60	58	44
7	47	45	36
8	38	35	27
9	33	30	22
10	25	24	18
11	21	17	15
12	18	16	11
13	17	12	10
14	16	11	8
15	14	10	7

2.1.4 *Comments on the results*

Considering the percentages F+P+B, the results given in Tables 1(a) and 1(b) show that London and Sydney monitoring observers produce somewhat different results. The difference between the results for the three periods 1, 2 and 3 are negligible so far as the London observers are concerned and, although somewhat greater, are still not statistically significant for the Sydney observers. This enables the results for the three periods to be pooled and the following table shows this for the criteria % F+P+B and % P+B.

		Condition		
		A	B	C
% F+P+B	London	24	25	46
	Sydney	16	19	39
% P+B	London	2.8	4.7	12.3
	Sydney	3.0	2.7	6.4

It is clear from this table that the longer propagation time of Condition C has seriously degraded the performance. The expected improvement due to modifying the echo suppressors (comparison of Condition B with Condition A) has not been realized; this may well be due to the fact, afterwards established, that the modification had increased the fault liability of the suppressors. The same conclusions may be drawn from the repetition counts given in Table 2.

The incidence of confusion shown in Table 2 does increase markedly at the greater propagation time of Condition C but the incidence is far less than was experienced in the laboratory tests referred to in section 1; it may be assumed that this effect was not a major cause of difficulty.

The occurrence of mutilations is very frequent but, in these tests, the incidence is only about one half that experienced in the laboratory tests. As found also in those tests, the incidence of mutilation does not increase significantly with the added propagation time but the trouble caused may well increase because the separation in time, to the subject, between the speech that gives rise to them and the mutilations themselves becomes greater. This would render them more noticeable.

It will be seen from Table 3 that the durations of the longer propagation time calls tend to be less, the duration of a median call under Condition C being about 18% less than that under Conditions A or B. It seems very likely that the customers are terminating their calls earlier when they experience more difficult circuit conditions.

The repetitions, taking all calls, are slightly more frequent than those obtained from calls lasting at least six minutes; this would indicate that the incidence during shorter calls is greater than that during longer ones. This effect has been found in previous studies of repetitions and is believed to be due to the customers becoming more accustomed to the circuit conditions. The effect is not noticeable to a significant extent in the mutilations.

2.2 *Calls from New York to London*

The object of these tests was mainly to obtain information based on customer interviews about difficulty experienced with long-propagation-time circuits. The tests were conducted by the A.T.T. Co. in co-operation with the British Post Office and the French P.T.T. The interview results are reported elsewhere and only the results of monitoring observations are treated here.

Pilot tests commenced on January 27th, 1964, propagation times up to 600 ms (mean one-way) being employed. It was decided from the results to make a more detailed investigation using mean one-way propagation times up to 400 ms only. These tests continued from February 10th to April 24th and covered about 1600 calls altogether. Only the U.S. to U.K. calls are considered here; the full test results are given in reference [3].

2.2.1 *Engineering arrangements for tests*

The tests were conducted over two New York-London (via White Plains), non-TASI, circuits routed over cable TAT3. Three different echo suppressors were used, the changes being made daily. The types were B (described as B-H in reference [9], another experimental suppressor described as type L, and British Post Office Type 6A modified by changing the frequency response of the control paths to peak at 3500 c/s, rising about 7 db from 1000 c/s to 2000 c/s.

The propagation time was increased from the normal value by adding delay at White Plains, the values of total mean one-way propagation time being 45 ms (the normal value plus the least setting possible of the delay device used), 150 ms, 300 ms and 400 ms, changes being made daily. Only calls originating in New York and terminating in the London director area were used.

2.2.2 *Results by monitoring observers*

The opportunity was taken in London only of employing the specially trained operating staff used for the tests described in section 2.1.2. Their procedure was exactly the same, the staff being employed on the Australia circuits morning and evening and on the U.S. circuits in the afternoon.

Figure 1 shows the opinions expressed by the monitoring observers, curves (a), (b), (c) and (d) showing the percentages " bad ", " poor or bad ", " fair or poor or bad " and " good or fair or poor or bad ", respectively. The results for the United Kingdom to Australia tests have also been plotted for comparison.

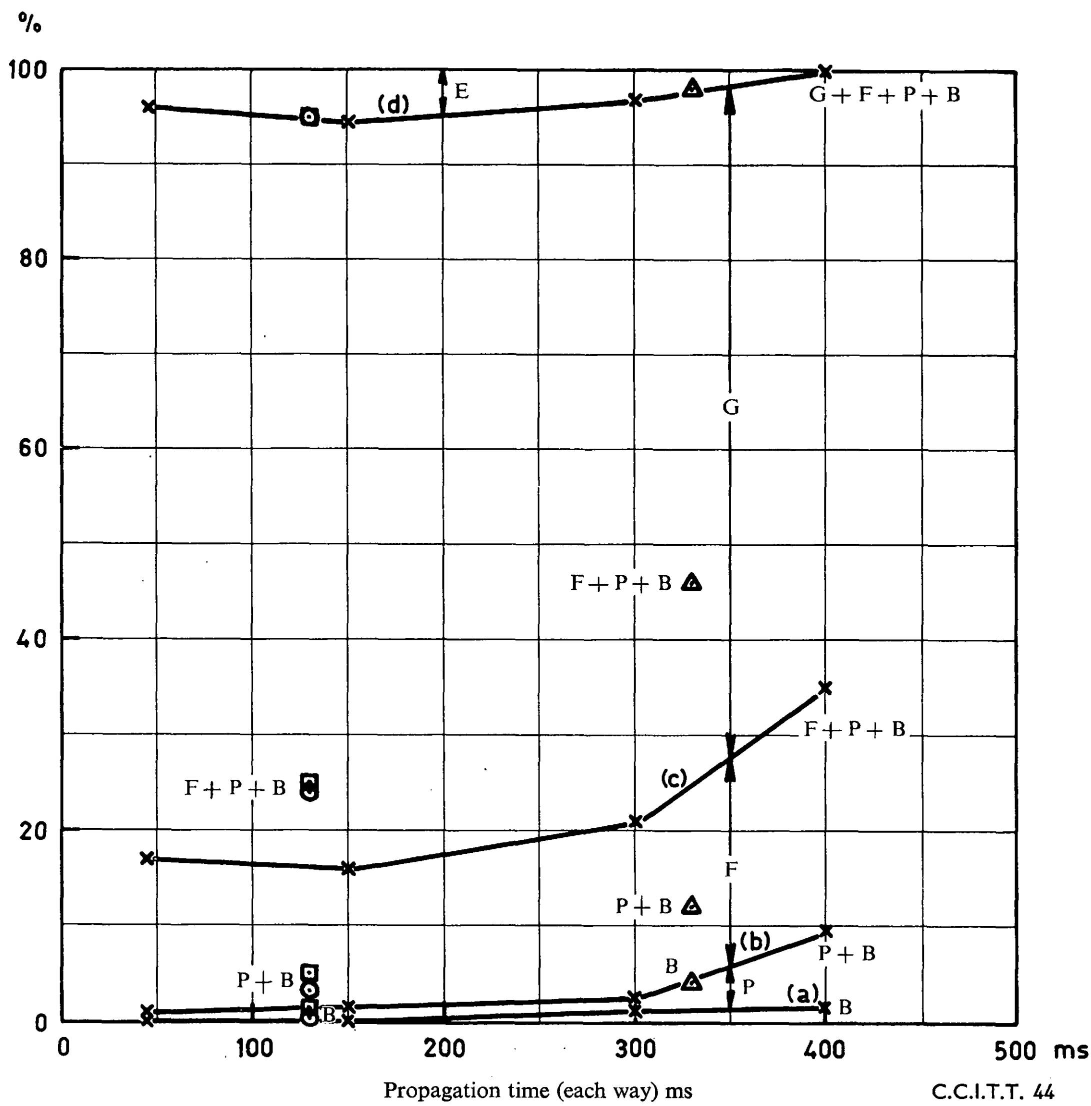
The incidence of requests for repetitions was as follows:

TABLE 4

Request for repetitions on calls between New York and London

Propagation time, ms	45	150	300	400
All calls, first six minutes (or less)				
Number of calls	191	191	178	189
Repetitions per six minutes	0.47	0.49	0.56	0.78
Calls lasting at least six minutes, first six minutes				
Number of calls	122	128	122	119
Repetitions per six minutes	0.47	0.40	0.52	0.66

Unlike the calls to Australia (Table 3), the transatlantic calls do not tend significantly to become shorter in duration as the propagation time increases; on the other hand the distribution of durations is different, the median call being appreciably longer. Table 5 gives the distribution pooled over echo suppressors.



✕ = United States—United Kingdom
○ = United Kingdom—Australia, A
□ = United Kingdom—Australia, B
△ = United Kingdom—Australia, C

(A, B, C: Mean of periods 1, 2 and 3)

FIGURE 1. — **Monitoring observers' opinions**

2.2.3 *Comments on the results*

The results from the U.K. to Australia tests have been plotted with those of the transatlantic tests on Figure 1, using the London observers' results so that the same team of observers applied. The results show worse performance at 330 ms on the U.K. to Australia circuits than on trans-

TABLE 5

Distribution of durations of calls between New York and London

The percentages of calls lasting at least T minutes

Propagation time, ms	45	150	300	400
	%	%	%	%
$T =$ 1	100	100	100	100
2	96	99	97	97
3	90	95	90	90
4	84	88	82	81
5	74	78	75	73
6	64	67	69	64
7	56	60	59	56
8	47	56	54	48
9	40	52	48	41
10	36	46	43	38
11	30	40	36	33
12	28	34	31	28
13	26	29	27	25
14	23	26	25	22
15	18	23	22	20

atlantic circuits at the same propagation time; at 130 ms the differences are less. This effect is also illustrated by comparing the repetition results in Tables 2 and 4. The conditions on the U.K. to Australia circuits differ in a number of respects from those on the transatlantic circuits: for example, the echo suppressors are different and the type of traffic and the customers' habits probably also differed. These considerations prevent any firm conclusions being drawn concerning the differences in performance of circuits on the two routes. They also make it impossible to determine whether the existence of alternative circuits with much lower propagation time on the transatlantic route had increased the awareness of these customers to the effects of lengthened propagation time.

The results obtained from the monitoring operators, both opinions and repetition counts, show an apparent improvement at 150 ms as compared with 45 ms; this is believed to be due to the systematic manner of allocating test conditions to the circuits which resulted in 400 ms on one circuit always being associated with 45 ms on the other. The observers may have been thereby biassed in such a way that they lowered their opinions on the shorter propagation time circuit and raised them on the longer, having experienced both on the same day.

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ANNEX 4

(modified in Geneva, 1964; quoted in Recommendation P.21)

METHODS APPLIED BY VARIOUS ADMINISTRATIONS IN INLAND LOCAL AND TRUNK NETWORKS, WITH A VIEW TO PROVIDING SATISFACTORY PERFORMANCE FOR NATIONAL CALLS

(it being understood that the C.C.I.T.T. Recommendation relative to reference equivalents is satisfied for international calls)

I. Argentine Republic
II. Australia
III. Canada and United States of America
IV. Finland
V. France
VI. Italy
VII. Japan
VIII. Norway
IX. Netherlands
X. Federal Republic of Germany
XI. United Kingdom
XII. Singapore
XIII. Sweden
XIV. Switzerland
XV. Czechoslovakia

I. Contribution by the Argentine Administration

Argentine is taking part in the work of the C.C.I.T.T.; its network will be brought into line with the standards laid down by that body for interconnections with networks of other countries and will be adapted to its national geographic and economic conditions.

The stability of the systems comprised in the network will finally determine the effective maximum attenuation between two subscribers. Preliminary analysis, based on normal equipment, shows that the maximum attenuation will not exceed 32 db between the most separated subscribers, the C.C.I.F. limits regarding connection loss, reflection, balance, etc., being respected.

II. Contribution by the Australian Administration

1. *Introduction*

The transmission plan described in this paper has recently been decided upon following a review of the previous plan. This review was made for the following reasons:

a) The Australian Administration has decided to adopt the crossbar system of automatic switching in place of a system using step-by-step switches for local exchanges and motor uniselector semi-automatic trunk exchanges. Trunk line switching at new installations will be carried out generally by the " true four-wire " method but two-wire switching will be used in some cases; up to the present time the " tail-eating four-wire " or two-wire switching methods have been in use.

b) It was decided to examine the necessity for taking " talker echo " into account in the design. The Australian network is composed almost entirely of high velocity circuits and echo has never been taken into account in the transmission design in the past and no trouble has been known due to this cause. However, a study of the practices of other Administrations, particularly those of the American Telegraph & Telephone Co., as published in 1953, indicated deficiencies in echo performances by those standards.

The changes which were made from the previous plan and the reasons for the design adopted are briefly set out in paragraph 3.

2. *Transmission plan*

2.1 The transmission plan is based on a reference system known as the " over-all transmission standard ", which consists of two telephone instruments with local lines and feeding bridges joined by an attenuator of 600-ohm impedance and 15-db attenuation. This is shown in Figure 1. The network design is such that transmission is worse than the reference system for only a very small percentage of connections, and for a large proportion of connections is appreciably better than the reference system. However, little expense is incurred to provide plant to meet the latter objective.

For design purposes the network is divided into three parts:

a) Local line network (subscribers' exchange lines);

b) Large metropolitan junction networks;

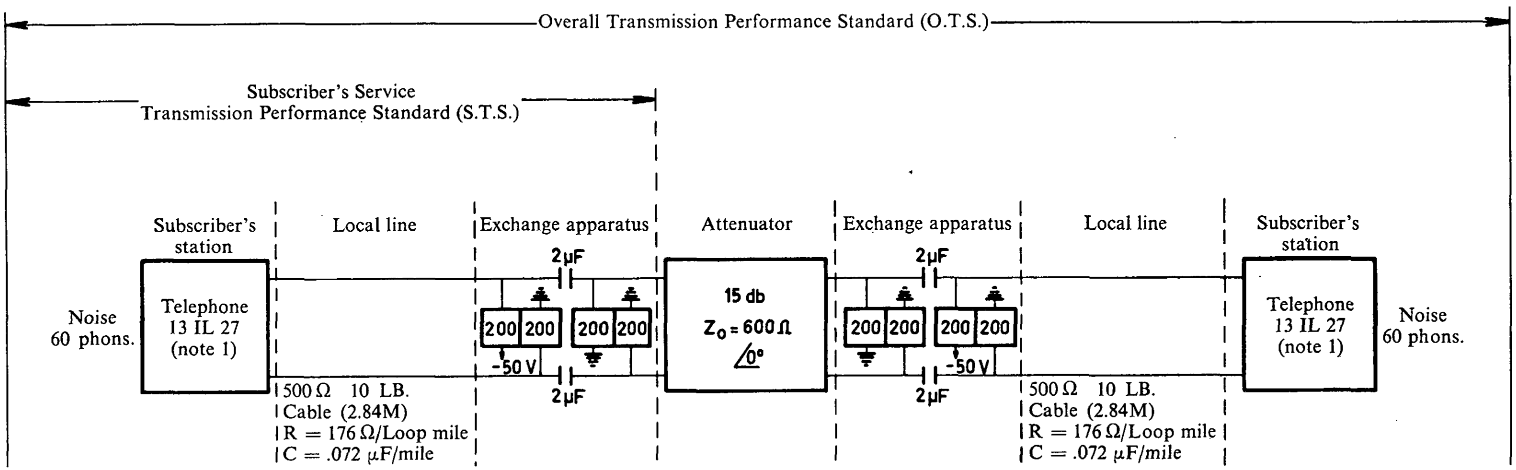
c) Long-distance network.

2.2 *Local line network*

The local line network comprises the lines from the subscribers' premises to the local exchanges; it is designed so that the transmission performance is at least equal to that of the local line section of the reference system. The telephone instruments included in the reference system are of less efficient type than those now being purchased and higher resistance limits now being applied in plant provision take account of this higher efficiency. The limits are, however, not as great as would be possible with a network composed substantially of the new type telephone instruments, due to the need for compatibility, and it is proposed to review these limits in the future.

The limits for the types of cable commonly in use are as follows:

Type of cable	Resistance limit (ohms)	Distance (km)
4 lb per mile (0.4 mm) 0.080 μF/mile	1150	4.23
6.5 lb per mile (0.51 mm) 0.072 μF/mile	920	5.5
10 lb per mile (0.63 mm) 0.072 μF/mile	770	7.05
20 lb per mile (0.90 mm) 0.072 μF/mile	610	11.3



C.C.I.T.T. 45

Note 1. — Telephone 13 IL 27 refers to handset type telephone having the following transmission components:

Transmitter inset	No. 13
Receiver type	IL
Anti-sidetone induction coil	No. 27

FIGURE 1. — **(Australia)**

There is also a signalling limit of 1000 ohms which is the controlling factor in the case of 4-lb and 6½-lb cables.

2.3 *Junction networks*

Special transmission plans are prepared for each of the large metropolitan junction networks rather than the application of the more general rules applied in the long-distance network as a whole. The trunk network design covers all circuits including junctions in other areas.

The essential requirements of the metropolitan junction network design are:

a) the attenuation between any two exchanges in the network shall not exceed 15 db, including the losses of intermediate exchanges;

b) the total attenuation from any exchange in the network to trunk exchange (which is a main trunk centre) shall not exceed 7.5 db, again including the losses of intermediate exchanges and the trunk exchange;

c) the loss of direct junctions between exchanges, i.e. junctions which can be switched only to subscribers at each terminating exchange, shall not exceed 12 db.

These figures of 15 db and 7.5 db are not divided to allow a fixed portion for particular classes of route, but allowances for each junction route are determined by considering the network as a whole and distributing the allowable total loss over the various junction groups in accordance with the most economical design as determined by the distance of each junction group. Losses are calculated from the attenuation at 1.6 kc/s for unloaded cable and 1.0 kc/s for loaded cable; most junctions are in cable loaded with 88 mH at 6000 feet (1.83 km) spacing. No allowance is made in the practical design procedures for losses due to impedance mismatch.

In practice, the permissible losses from tandem to terminal exchanges vary from, approximately, 3 db to 5.5 db, tandem to tandem junctions from 1 db to 6 db and trunk exchanges to tandem exchange from 1 db to 3 db. However, with exchange equipment being provided in the future it is proposed to provide direct junctions from the trunk exchange to all terminal exchanges. Amongst other reasons, direct trunk junctions are desirable to simplify stability pad arrangements since junctions to tandem exchanges are in general of lower loss than 3 db.

2.4 *Trunk line network*

Trunk exchanges are classified according to their position in the switching scheme as follows:

a) main trunk centre;

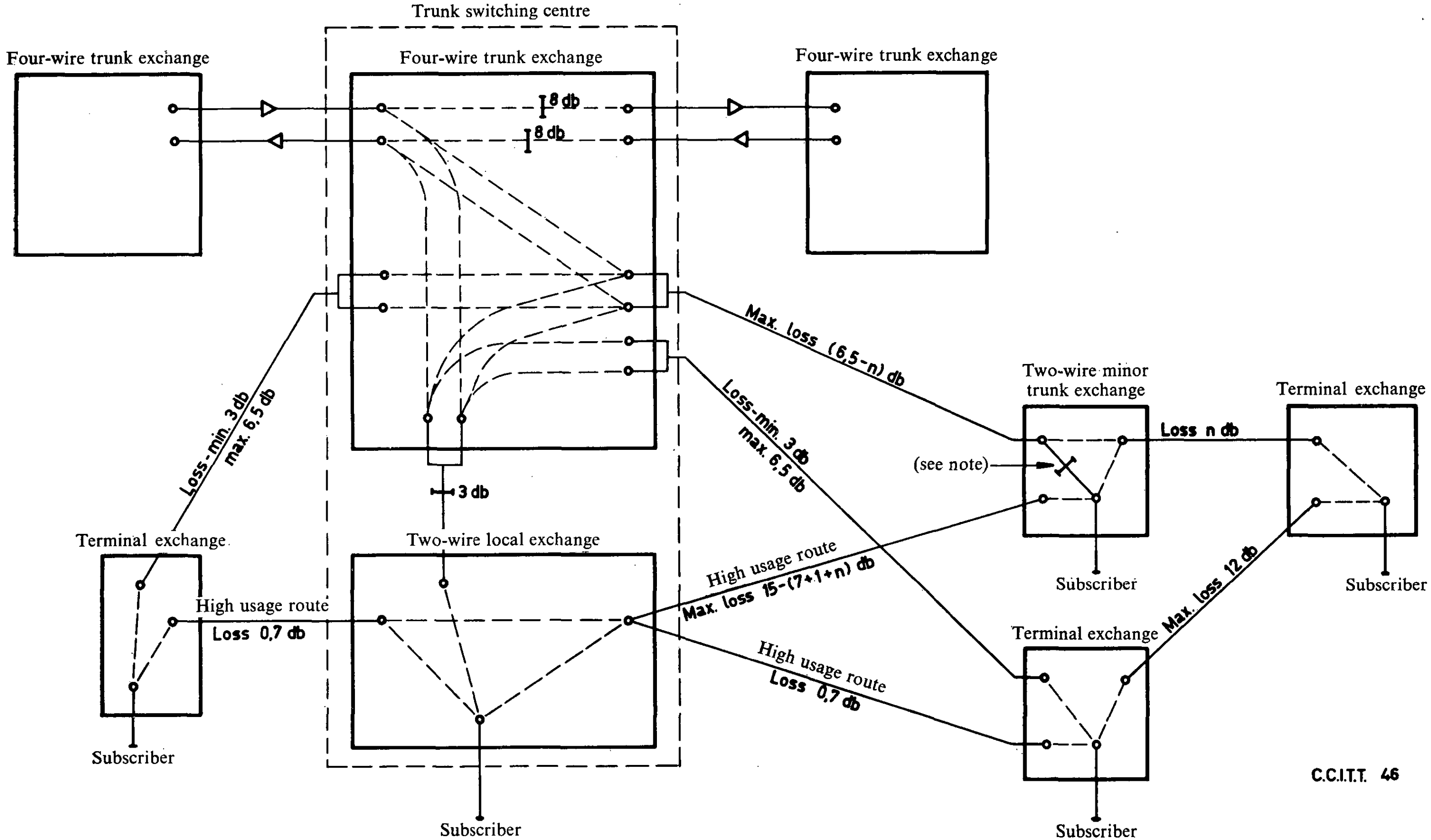
b) primary trunk centre;

c) secondary centre;

d) minor trunk centre;

e) terminal exchange.

The design is based on the use of four-wire low-loss circuits between main trunk centres, primary centres and secondary centres. Four-wire low-loss circuits are also provided in general from minor trunk centres to higher order switching centres except for the cases described below. Circuits terminating at terminal exchanges from exchanges of any other class may have a loss of 6.5 db (6 db line loss plus 0.5 db per matching transformer where required).



Note. — Pad required only if loss to switching centre is less than 3 db.

FIGURE 2. — **Circuit and pad losses terminal exchange links (Australia)**

The four-wire low-loss circuits interconnecting the higher switching centres are operated at zero loss in the transit condition with the following exceptions:

a) Circuits between main trunk centres in adjacent States are operated at a nominal loss of 0.5 db in the transit condition. The lengths of these circuits vary from 800 km to 1000 km except for the two special cases where echo suppressors are proposed to be used. Between non-adjacent States the losses are greater with a maximum of 1.5 db except where echo suppressors are proposed when the loss of 0.5 db;

b) Echo suppressors are proposed for the longest circuits only—those from Perth and Darwin to the other main trunk centres. The lengths of these circuits vary from 2500 km to 5500 km;

c) Other four-wire low-loss circuits, i.e. from main trunk centres to lower ranking centres, are to be operated at 0.5 db loss where the circuit length is between 550 km and 1200 km and at 1.0 db loss where the circuit length exceeds 1200 km.

Trunk circuits between terminal exchanges and the next higher trunk switching centres may have losses up to 6.5 db.

In cases where the full allowance of 6.5 db is not required between a minor trunk centre and any of its terminal exchanges two-wire circuits may be provided to the next higher switching centre provided that the total loss of the two circuits does not exceed 6.5 db.

Where direct junctions are provided between terminal exchanges the limiting loss is 12 db. In other cases of direct " high usage " routes circuit losses may be such that over-all losses terminal exchange to terminal exchange do not exceed 15 db including 0.5-db allowance for each intermediate switching point.

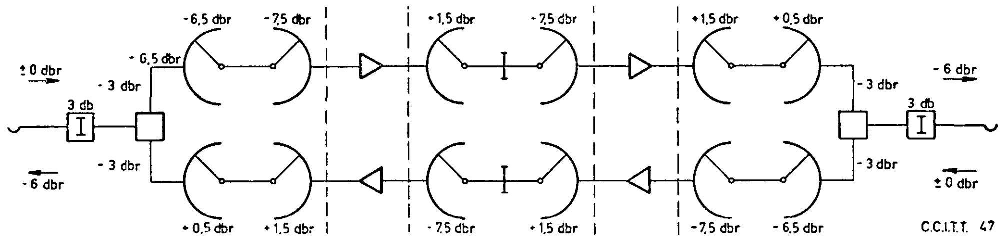
The minimum loss between terminal exchanges, including the local exchanges associated with trunk switching centres, is 6 db for connections including four-wire circuits. That is provided by the inclusion in circuit of 3 db stability pads in the two-wire/four-wire termination or by the inclusion of circuits with a loss of at least 3 db between the two-wire/four-wire terminations and subscribers' lines.

This requirement is achieved by including pads in the two-wire/four-wire terminating sets connecting the four-wire trunk exchange with the two-wire local exchange. These pads are in circuit for all connections made by this path including those between subscribers of the local two-wire exchange and two-wire circuits terminating on the four-wire exchange.

In this latter case, i.e. on connections between subscribers and two-wire circuits, pads are not necessary and are, in fact, undesirable. This disadvantage is avoided with present step-by-step switching equipment by segregating switches and circuit groups in the exchange trunking. Facilities to remove the pad in this case are unduly costly with the crossbar equipment and the loss is accepted, since only traffic which overflows from direct two-wire trunk circuits suffers the additional loss. The arrangements are illustrated in Figure 2.

The nominal impedance presented at the line terminals of the two-wire/four-wire terminations of zero loss circuits is 600 ohms.

Four-wire zero-loss circuits are lined up to zero loss between two-wire points of the circuits; the nominal loss of 6 db is achieved by 3-db pads between switchboard and two-wire/four-wire termination making the relative level at the two-wire terminals of the two-wire/four-wire termination -3 db for each direction of transmission. Relative levels at various switching points are shown in Figure 3. These levels include allowance for exchange equipment losses of 1 db.



Pad loss 8 db
Equipment loss 1 db

Note. — The levels shown apply with zero transit loss of four-wire circuits; where a loss of, for example, 0.5 db applies, this loss is included in the four-wire circuit.

FIGURE 3. — **Relative transmission levels at trunk exchange switching points (Australia)**

3. *Changes from previous plan*

The new plan described is very similar to the plan which has been in use for a number of years and is still generally in use. Four-wire circuits were switched by means of tail-eating connections at semi-automatic trunk exchanges; or by two-wire connections. 3-db switchable pads included in the line and net connections of the two-wire/four-wire termination serve both for level control purposes in tail-eating connections and for stability control. In the case of two-wire extensions to subscribers the pads are in circuit but are switched out where trunk or junction circuits with a loss greater than 3 db are connected.

However, difficulties arise in practice in achieving all requirements for two-wire switching conditions and in recent installation arrangements have been made to include pads in the path through the exchange by segregation of trunking groups into those requiring pads and those which do not. These pads must be included in such a way that they are not in circuit during signalling where this affects signalling line resistance limits.

In determining facilities to be provided by the crossbar switching system consideration was given to abandoning the use of pads for stability purposes by operating four-wire circuits at a lower gain over the " true four-wire " part of the circuit in order that sufficient loss existed between the two two-wire terminations to ensure adequate stability. This would be either to provide a loss between two-wire terminations of 4 db with a single four-wire link increasing with the number of links (loss = $(3+n)$ db where n is the number of links) or 7 db irrespective of the number of links. The 4-db arrangement has the advantage of providing lower loss on the more numerous connections with a small number of links, and a relatively uniform margin of stability with varying number of links.

A loss of 7 db is achieved by adjusting the loss between the four-wire switch banks to zero and with the two-wire/four-wire terminating sets providing the loss; this method has the advantage that no pads are required in four-wire to four-wire connections and further simplifies switching arrangements.

However, both these methods reduce the loss which may be allocated to the two-wire end links for a given over-all loss limit. This of course increases the cost of these circuits, and whilst it may be argued that transistorized VOF amplifiers can be added where necessary at relatively small cost, particularly since they may be included in the two-wire/four-wire termination which is necessary with true four-wire switching, the additional costs involved make this solution of doubtful value for general application.

4. *Echo control*

The Australian Administration has never used echo suppressors and this has not given rise to complaints from subscribers. The Australian network is, and always has been, composed almost entirely of high velocity circuits and very few calls exceed a distance of 4300 km. However, calls of 9000 km are possible and for the echo delays of approximately 100 m/s involved echo control is essential for satisfactory conversation.

Echo suppressors are therefore desirable for the longer connections since excessive attenuation would be necessary to provide satisfactory echo conditions in these cases, but for a large part of the network it has been decided to add loss to the extent given in paragraph 2.4 a) and c). This loss is insufficient to provide echo conditions desirable but it is expected that the deficiency in echo performance will not be serious and is preferred to further increasing the transmission loss. This is discussed fully in reference [2].

5. *Impedance compensation*

The maintenance of adequate stability and echo conditions is dependent on the impedance presented by two-wire terminal exchange links providing return losses of a similar order to those provided when the 3-db pads are in circuit. It is known that this condition is not met by the present Australian network, where large numbers of the two-wire links are provided by cable pairs loaded with 88 mH inductors at a spacing of 1.83 km and terminated with end sections of 0.91 km. Matching transformers of 2/1 impedance ratio are necessary at the point of connection to the trunk system but a system of compensation to improve the return loss at the high and low ends of the voice-frequency band will also probably be necessary.

REFERENCES

[1] Huntley, H. R.: Transmission Design of Intertoll Trunks. *B.S.T.J.* XXXII, 5 September 1953, page 1024.

[2] Kitchenn, R. G.: Stability and Echo in the Trunk Network. *Telecommunication Journal of Australia,* Volume 13, No. 1, June, 1961, page 49.

III. Contribution by the Telephone Association of Canada and by the American Telephone and Telegraph Co.

North-American practice for transmission requirements of the national network

This text outlines briefly the transmission practices at present in use on the North-American continent. Special mention is made of the major differences compared with transmission practices followed in other continents.

By way of general orientation the major differences between this method and other methods are:

1. The expression of transmission of other than " reference equivalent ";
2. The abandonment of the limiting loop concept in transmission design;
3. The substantial independence of the " loop and telephone set " transmission performance and that of the other links of a connection.

Transmission principles

The basic principle has been to give the subscriber the kind of transmission he himself finds satisfactory. Methods of doing this have changed from time to time, but fundamentally this approach to transmission design is based not on what subscribers will tolerate, but rather on what they prefer.

The general philosophy being applied in current reviews of transmission objectives is to provide a quality of service that will be rated by customers as GOOD in at least 95% of connections; FAIR in no more than 5% and POOR in a negligible percentage of cases due to specific trouble conditions. Guidance in achieving this goal is obtained by tests in which sample groups of subjects are asked to appraise a range of impairment conditions using categories of the GOOD, FAIR, POOR variety. Such subjective test results have begun to influence the design of the North-American network and they have been used to evaluate the present performance. As an example, it has been shown that the general service objective has not yet been fully achieved with respect to received talker volumes.

Philosophies as to how to apportion the permissible transmission losses among the various links of a connection have varied over the years. At one time, for example, there was an assigned value for " loop and trunk " portions; and extensive studies were made in individual cases to determine the most economical division of the value between loops and trunks.

The general method of transmission design outlined below is based on:

1. great improvements in the efficiency of the telephone set;
2. decreased cost of providing transmission facilities between local exchanges and for long-distance connections;
3. considering both cost and time of utilization of individual links.

Subscribers' lines (loops)

1. The line conductors are designed on the basis of assigning the minimum amount of copper necessary to meet the limitations on d.c. resistance imposed by the signalling, pulsing, and supervision requirements of the central office switching circuits.

For the most widely used types of switching systems, satisfactory operation will be obtained with external loop d.c. conductor resistance up to 1200 to 1300 ohms. This is approximately equivalent to the lengths shown in the following tabulation:

Length of pair (feet)	Gauge	Diameter (mm)	Pounds per conductor mile
15 000	26	0.40	4.1
24 000	24	0.51	6.5
38 000	22	0.64	10.3

2. In order to reduce the spread in transmission performance between the longest loops and the average loops, loading coils generally of 88 mh are introduced at 6000 feet spacing on all loops longer than 18 000 feet. The first loading coil is located 3000-feet from the central office so as to ensure maximum effectiveness in connections between loaded loops and between loaded loops and trunks.

3. Also in the interest of reducing the spread in transmission performance between the longer loops and the average, modern high efficiency type sets are assigned to loops over 10 000 feet in length, shorter lengths being assigned this type or older types. Instrument zoning practice of this kind ensures the most effective use of the two types of sets having significantly different transmission performance. Such zoning would not be required if only the high efficiency set were used.

4. Bridged taps are kept to a minimum and in any case are usually limited to a maximum of 6000 feet.

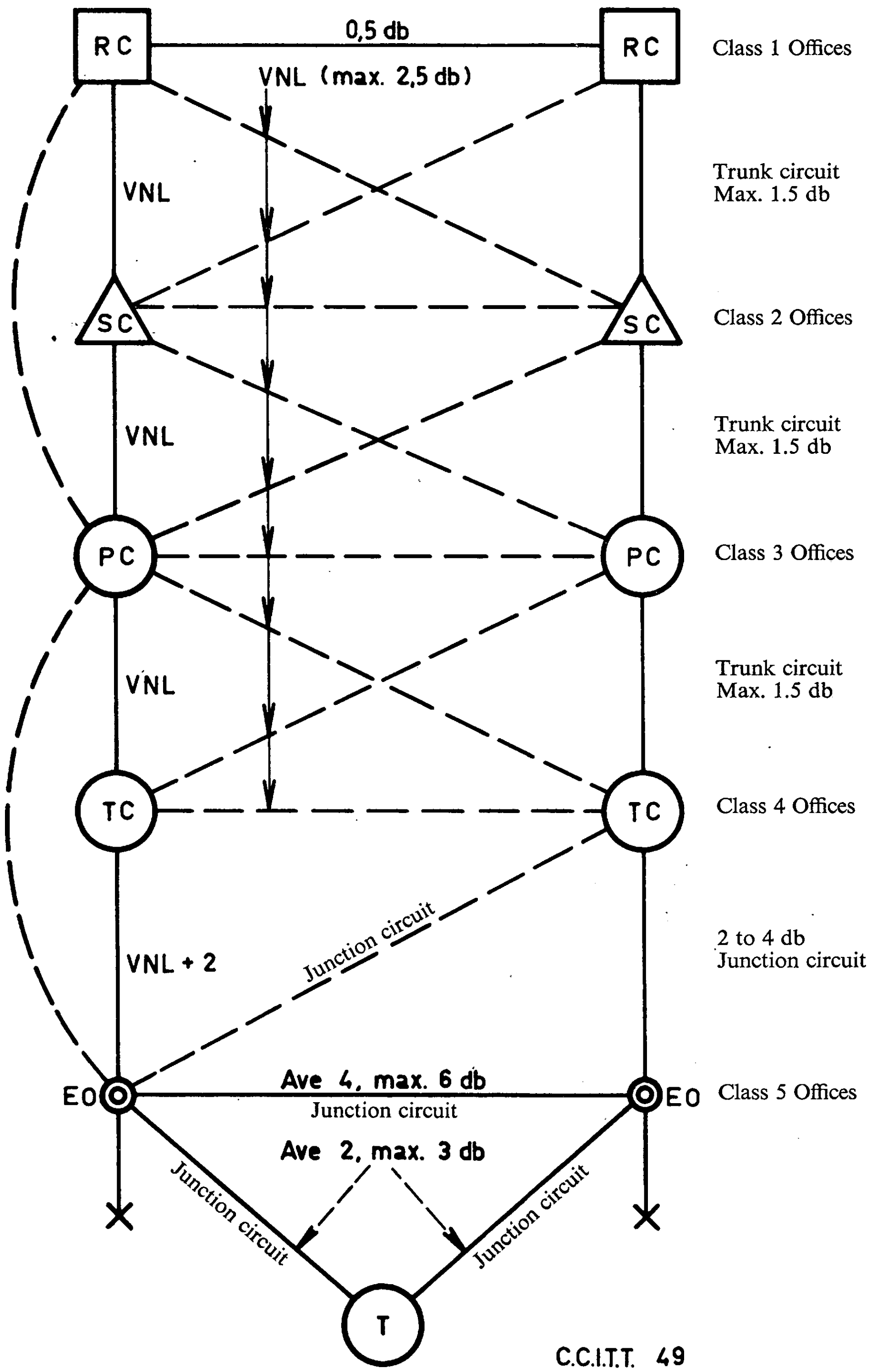
With subscriber plant designed on this basis, calls between any two telephones connected to the same switching office would be rated GOOD with respect to volume by virtually all subscribers, and hence would be in line with the broad over-all objective.

This method obviously greatly simplifies engineering [1]. Obviously, too, it might need modification with changes in switching systems, telephone apparatus, and signalling requirements or customer demand for improvement in service. Its development involved trial application to various types of central office areas as regards length and distribution of loops, the comparison of the effective transmission performance with that of previous methods, judgment as to how much to allow for losses in trunks. The allowance leaned heavily on the trend toward very low losses in these other links, and specifically the more widespread use of carrier in very short facilities.

Direct and tandem trunks

These are facilities to provide connections between two local offices. Whether the connection is via a single direct trunk or two interconnected tandem trunks the 1000-c/s loss objective is 4 db average with a maximum loss of 6 db. This is shown diagrammatically on Figure 4. For convenience the local (end) offices are shown in two different regional areas. The same objectives apply regardless of location, whether in different areas, in a multi-office exchange area, or in a large metropolitan area. The small permissible spread between average and maximum should be noted.

[1] See:
— Simplified Transmission Engineering in Exchange Cable Plant Design. *Communication and Electronics (A.I.E.E.)*, November 1954.
— Transmission Economics. *Telephony*, 23 August 1958.
— Subscriber Loop Design. *Telephone Engineer and Management*, 15 September 1961.



X = Subscriber
EO = End Office (local office)
T = Tandem office

TC = Toll Center
PC = Primary Center
SC = Sectional Center
RC = Regional Center

——— Final route
— — — — High usage route

FIGURE 4. — **(North America)**

Toll connecting trunks

These are facilities between local offices and long-distance switching systems. They are operated at losses between 2 and 4 db. The objective is to provide a loss of V.N.L.+2 db with a maximum of 4 db. V.N.L. (or via net loss) is defined as the lowest loss in db at which it is desirable to operate a trunk facility considering limitations of echo, crosstalk, noise and singing. From a knowledge of the various types of facilities employed, it has been possible to specify the V.N.L. [1] for each type on a per-mile basis. A minimum loss of 2 db is required (which in the case of short trunks is provided by inserting a resistive attenuator) to ensure stability of over-all connections.

Long-distance circuits

These are the circuits between switching centres in the continent-wide long-distance network. The losses of final trunk groups are the V.N.L. losses, i.e., the smallest loss permissible in each case from standpoint of echo, singing and crosstalk. A further requirement of 1.5 db maximum is imposed so as to ensure low loss; high velocity facilities are assigned for this service. Interregional high usage trunks (see Figure 4) are permitted to have a maximum loss of 2.5 db.

Thus it may be seen that the objective for the average loss in a connection between two local offices is about 4 db if the connection is via direct or tandem trunks; or somewhat greater if via long-distance circuits, this varying with the number of links. In 83% of calls only a single long-distance circuit is involved; in 15% two; and in three or more in the remaining 2%.

Discussion

As indicated, in North America, the transmission design of subscribers' loop and telephone sets has been largely divorced from that of trunks, and both are such that the concept of national transmitting and receiving " reference equivalents " now used in the C.C.I.T.T. is not applied. While this method was formerly used in North America it was abandoned when evolution of new telephone instruments decreased its meaning. For example, it was discovered that the very real quality improvement due to introducing a substantially flat frequency response in telephone instruments was not reflected in the loudness balance measurements of a reference equivalent. To circumvent this an effective loss method was introduced in which comparisons between the various loops and the various sets were made on the basis of the rate of repetitions asked for by subscribers during normal conversations.

Effective loss objectives were then established for the subscriber loop plant and data relating effective losses to loop make-up and sets were used for practical design purposes. The resistance design method now used, described above (see section 1), was later conceived as a more simple and practical method to use to stay within the effective loss limits and at the same time to meet loop signalling limitations.

Rating on the basis of repetitions has grown to be a rather insensitive measure of performance, and effective losses are by nature difficult to relate to reference equivalents. On the North-American continent, it again seems practical to revert to the principle of loudness ratings since telephone instruments and trunks are now of a uniformly high quality. The design of trunks

[1] For a detailed discussion of V.N.L., see: Transmission Design of Intertoll Telephone Trunks. *Bell System Technical Journal*, September 1953.

has become a matter of 1000-c/s loss since this quantity is an adequate measure of the effect of modern facilities on transmission performance. An electro-acoustic rating system [1] which measures loudness loss objectively is being studied in order to put the evaluation of loops and sets on a similarly objective basis. It is expected that such ratings can be related to reference equivalents assuming the use of similar high quality telephone instruments in both cases. It is appreciated that such a relation will not take into account the difference in performance due to sidetone, frequency response and similar factors which the basic rating systems do not evaluate.

IV. Contribution by the Telephone Company of Helsinki (Finland)

According to the new transmission plan, the sum of the reference equivalents of the national parts would be 2.4 N + 1.4 N = 3.8 N. Part of this should be allocated to the national lines and exchanges between the end of the international circuit and the local exchange. For the conditions which at present exist in Finland, the magnitude of this portion must be estimated at least at 1.15 N, so that the sum of the sending and receiving reference equivalents of the subscriber system will be 3.8 N − 2.3 N = 1.5 N.

To keep the cost of construction of the subscriber network reasonable, the telephone set must be sufficiently sensitive. Likewise, the requirements imposed upon the sensitivity of the telephone set by the maximum total attenuation must be taken into account.

To a certain extent, the drawbacks resulting from a high sensitivity of the telephone set can be avoided through employing automatic sensitivity regulation, as is the case, e.g., with our new telephone set. This, among other things, reduces the risk of too small reference equivalent between any two subscribers. In the following, attention will be devoted to this point as well as to the maximum permissible attenuation.

To characterize the reference equivalent of a local connection, the combined sending and receiving reference equivalent of the telephone set and the subscriber line is represented in Figure 5 as a function of the length of the subscriber line, assuming that use is being made of a telephone provided with automatic regulation. In the figure, the sending reference equivalent increases upwards and the receiving reference equivalent downwards. Owing to the variations of the properties of the telephone set the reference equivalent varies over the shaded areas. The reference equivalent between two subscribers is represented by the vertical distance between the points in these areas.

The magnitude of the maximum reference equivalent q_{max} (Figure 5) was estimated above at 1.5 N. This may be divided into three parts, viz., a minimum reference equivalent between the two subscribers, q_{min}, and the sending and receiving reference equivalent of the telephone and the subscriber line. In Figure 5, these two parts are represented by the shaded areas.

Judging from the experience acquired with the telephone sets provided with automatic regulation, the standard deviation of the combined reference equivalent of the telephone set and the subscriber line in the receiving direction can be assumed as 0.15 N for usable lengths of subscriber lines. This value includes the effects of imperfection in the regulation. Correspondingly, the standard deviation for sending has been estimated at 0.25 N. The distribution is assumed to be normal.

Granting that the comparative attenuations for sending and receiving permissible in a local system may be exceeded by a 5% probability and that the telephone set is dimensioned in the

[1] See: A Revised Telephone Transmission Rating Plan. *Bell System Technical Journal*, May 1955.

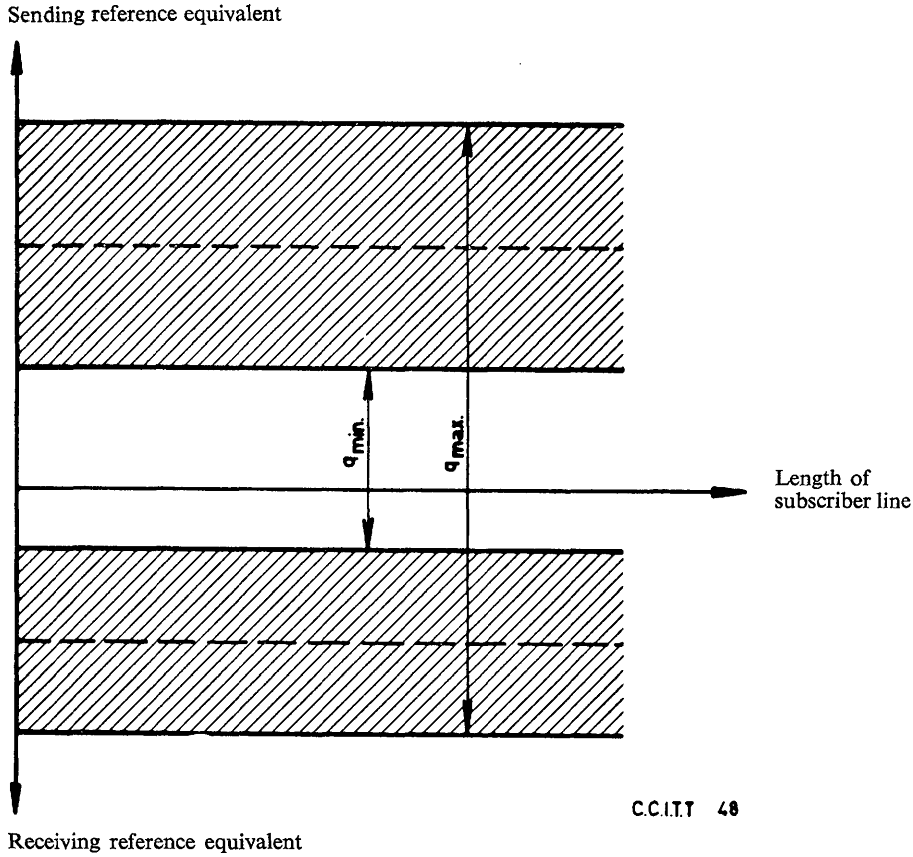


FIGURE 5. — **(Finland)**

most advantageous possible way, the mean of the reference equivalent of local calls will be $q_m = 1.5\ \mathrm{N} - (0.25\ \mathrm{N} + 0.4\ \mathrm{N}) = 0.85\ \mathrm{N}$ (Figure 5).

The standard deviation for the local calls can be estimated at $\sqrt{0.15^2 + 0.25^2}\ \mathrm{N} = 0.29\ \mathrm{N}$. Hence, in 12% of the local connections the reference equivalent can be estimated to be less than 0.5 N and, in 2%, less than 0.25 N.

The above values of standard deviation presuppose automatic sensitivity regulation operating relatively efficiently if it is sought to attain the maximum subscriber-line length (1000-ohm loop resistance). In estimating the standard deviations, no allowance is made for the effect of the ageing of the capsules, of which we have no experience.

The new transmission plan obviously makes it necessary to impose fairly rigid tolerance requirements upon the sensitivity of the telephone set as well as upon the properties of the automatic regulation if it is sought to attain the smallest possible reference equivalent between any two subscribers. According to the estimates described above, it may in many cases be markedly below the value regarded as the most advantageous. Since sufficiently sensitive capsules are available today, the possibilities of realizing the new transmission plan with the aid of a more sensitive telephone can be considered to depend chiefly upon how small the standard deviation of the subscriber system can be kept, or, in other words, upon the extent to which the properties of the capsules vary and upon how efficiently the automatic sensitivity regulation operates.

V. Contribution by the French Administration

1. *General*

The new transmission plan adopted by the French Administration takes account of the C.C.I.T.T. recommendations concerning the international telephone service, and particularly of Recommendation G.111 (or P.11), which established practical limits for the nominal reference equivalents of national sending and receiving systems.

As regards national communications, the following rules have been accepted:

1) The over-all reference equivalent should be less than 4.15 N for 90% of calls between any two subscribers in the network;

2) The over-all reference equivalent should exceed 4.60 N in quite exceptional cases;

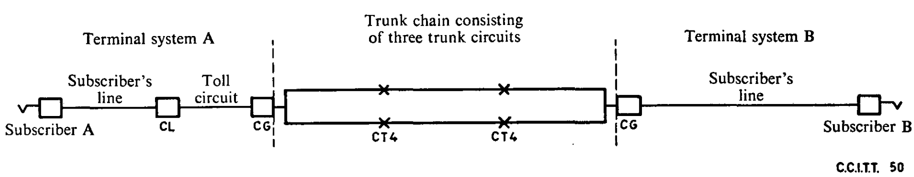
3) The nominal values of the equivalents at 800 c/s of the various circuits comprising the four-wire chain shall be selected in such a way as to ensure the stability of the chain for a negative variation of the over-all loss equal to three times the standard deviation.

By definition:

the *terminal system* is the unit formed by the subscriber's set, the subscriber's line, the local exchange(s), the toll circuit(s) and primary centre; the *trunk chain* is the unit constituted by the trunk circuits and the transit exchanges situated between the two primary centres.

Figure 6 shows, as an example, a diagram of a trunk call broken down into its three component parts.

The new plan gives separate definitions for nominal reference equivalents of the trunk circuit and the terminal system (sending and receiving).



CL = local exchange
CG = primary centre
CT4 = transit centre with four-wire switching

FIGURE 6. — **(France)**

2. *Reference equivalent of the trunk chain*

The reference equivalent of the trunk chain was assumed to be equal to the sum of the equivalents at 800 c/s of the various circuits, possibly increased by the losses due to passage through transit centres equipped with two-wire switching (CT2).

The following values were taken:

1. The nominal attenuation of a chain of interconnected four-wire circuits, as measured between two-wire input and output points, is equal to:

$$4+0.5\,n \text{ dN}$$

   n being the number of four-wire trunk circuits in the chain (in practice n is between 1 and 6).

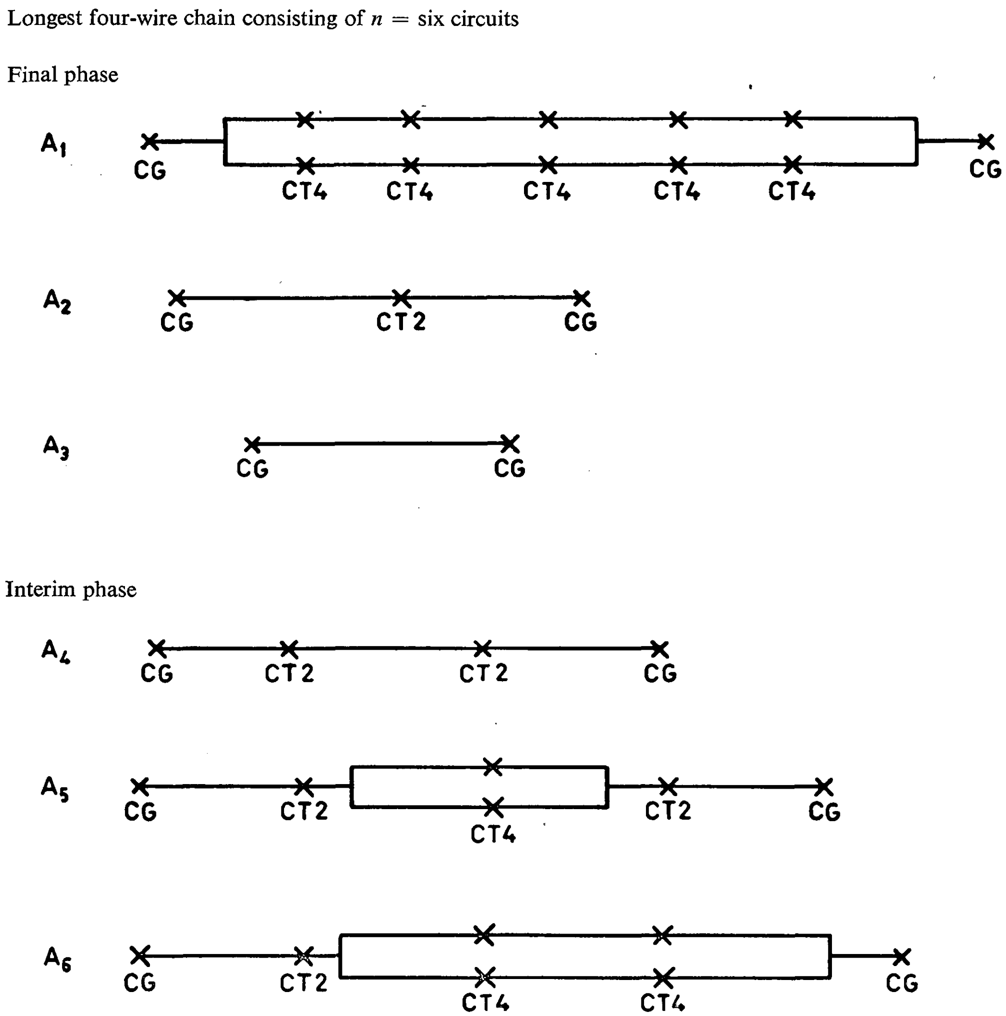
2. The trunk circuits switched on a two-wire basis at both ends are adjusted to 3 dN (four-wire circuits) or 5 dN (two-wire circuits).

3. The equivalent at 800 c/s of a circuit interconnecting two primary centres which do not provide transit facilities should not exceed 1 neper.

4. The insertion loss of a CT2 is assumed to be equal to 15 cN.

The various types of trunk chains being planned are shown in Figure 7. When four-wire switching will have been installed in all the regional transit centres and in the busiest departmental transit centres, only cases A_1, A_2 and A_3 will have to be taken into account. During the interim period, cases A_4, A_5 and A_6 will also be taken into account.

The standard deviation of the variations in the equivalent of a circuit is assumed to be 12 cN for circuits on primary groups with automatic regulation, and 17 cN for other circuits.



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CG = group centre
CT2 = transit centre with two-wire switching
CT4 = transit centre with four-wire switching

FIGURE 7. — **Trunk chains. (France)**

3. *Sending and receiving reference equivalents of the terminal system*

The terminal system reference equivalent is obtained by adding the equivalent at 800 c/s of the toll circuit(s), the loss through centres (other than the centre serving the subscriber's installation) and the reference equivalent of the subscriber's set. This last value is given by a nomogram described in the appendix.

Planning rules for local networks tend to establish the nominal maximum reference equivalent of the terminal system at 1.6 N for the sending end, and at 0.6 N for the receiving end. This value will only be attained for the least favoured subscribers in the group.

Further, the possible variations of the reference equivalent of the terminal system (sending or receiving) have been estimated at ± 2 dN about its nominal value, due:

— to dispersion and variations with time of the sensitivity of the microphone and receiver insets, and

— to fluctuations in battery voltage.

Where appropriate some account should be taken of variations in the equivalent at 800 c/s of repeatered toll circuits.

4. *Over-all results*

To verify that the rule of 4.15 N applies to 90% of calls, it appears necessary to study a statistical model of the network. The basic data are being collected and detailed information will be supplied later.

It is comparatively simple, on the other hand, to verify that in almost all possible calls, the reference equivalent is less than 4.6 N.

For each of the cases previously mentioned, Table I shows maximum estimates for the reference equivalent of calls. The assumptions on which this calculation is based are:

a) That each of the two subscribers is one of the least favoured subscribers in his group (reference equivalent of the terminal system 1.6 N sending and 0.6 N for receiving).

TABLE I

Type of trunk chain	Reference equivalent (in N)					
	Nominal of trunk chain	Nominal maximum of terminal system		Nominal total	Total (including positive variations)	
		Sending	Receiving		$\sigma=0.12$ N	$\sigma=0.17$ N
A_1 ($n = 3$)	0.55	1.6	0.6	2.75	3.57	3.75
A_1 ($n = 5$)	0.65	1.6	0.6	2.85	3.79	4.01
A_1 ($n = 6$)	0.70	1.6	0.6	2.90	3.90	4.14
A_2	1.15	1.6	0.6	3.35	4.09	4.23
A_3	1	1.6	0.6	3.20	3.84	3.94
A_4	1.60	1.6	0.6	3.80	4.62	4.80
A_5	1.80	1.6	0.6	4	4.88	5.08
A_6	1.20	1.6	0.6	3.40	4.28	4.48

b) That each of these figures has been increased by 0.2 N, which represents the maximum variation of the equivalent in the unfavourable sense.

c) That the positive variation in the loss of a chain of n trunk circuits has been estimated at $2\sigma\sqrt{n}$ ($\sigma = 0.12$ or 0.17 N) (with 2% probability that this will be exceeded for a Gaussian distribution of the equivalents about the mean).

Clearly, this procedure for calculating the reference equivalent of a call is very conservative and the probability that it will be exceeded is almost nil.

It can be seen that the condition of 4.60 N is maintained with a comfortable margin in the case of configurations relating to the final phase of application of the new plan. However, the distribution is occasionally exceeded, in cases A_4 and A_5, which correspond to the very beginning of the interim phase.

Appendix

(to the contribution of the French Administration)

Reference equivalent of a subscriber's local circuit

The method of calculation will be applied to the case of a BCI-U43 telephone set but it is also applicable to other types of subscriber sets, and in particular to the new S63 set.

The reference equivalents for the various constituent elements of the subscriber system represented in Figure 8 are calculated by simple formulae deduced from the results of subjective voice-ear measurements. Expressed in nepers these equivalents are respectively:

0.08 for the feeding bridge;

$\frac{0.08}{1000}(x + x_a)$ ohms for the line and the additional resistance of the telephone set;

$4.45 - 0.9 \log_e I_{mA}$ for the BCI-U43 telephone set, when sending (without additional resistance);

-0.3 for the same set when receiving.

The reference equivalent of the system has been obtained by adding the reference equivalents of its different elements, i.e.:

$$e = 4.45 - 0.9 \log_e \frac{48\,000}{360 + x + x_a} + \frac{0.8}{1000}(x + x_a) + 0.08$$

$$r = -0.3 + \frac{0.8}{1000}(x + x_a) + 0.08$$

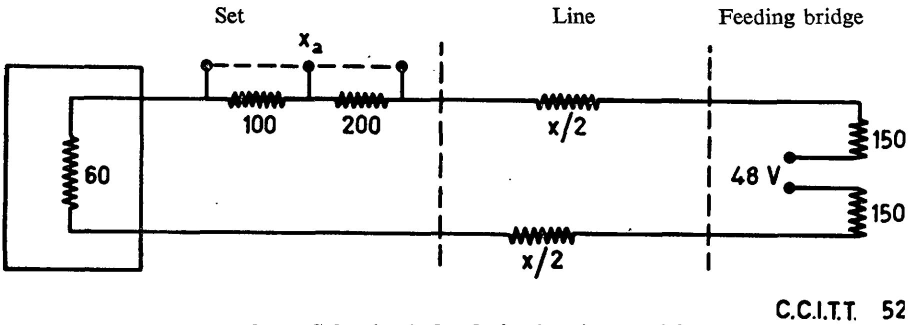


FIGURE 8. — **Subscriber's local circuit using set BCI-U43. Values of d.c. resistances in ohms (France)**

Figure 9 shows the variations of these two equivalents as a function of the variable $x + x_a$, i.e. of the line resistance increased by any additional station resistance. The sending and receiving equivalents of any subscriber system can thus be obtained in a simple manner, whether the line is of uniform or mixed composition. If the total available equivalent is taken as a starting-point, the maximum line resistance authorized can be read off from the nomogram.

The simplicity of this nomogram is mainly due to the formula adopted for the reference equivalent of the line, which depends only upon the line resistance.

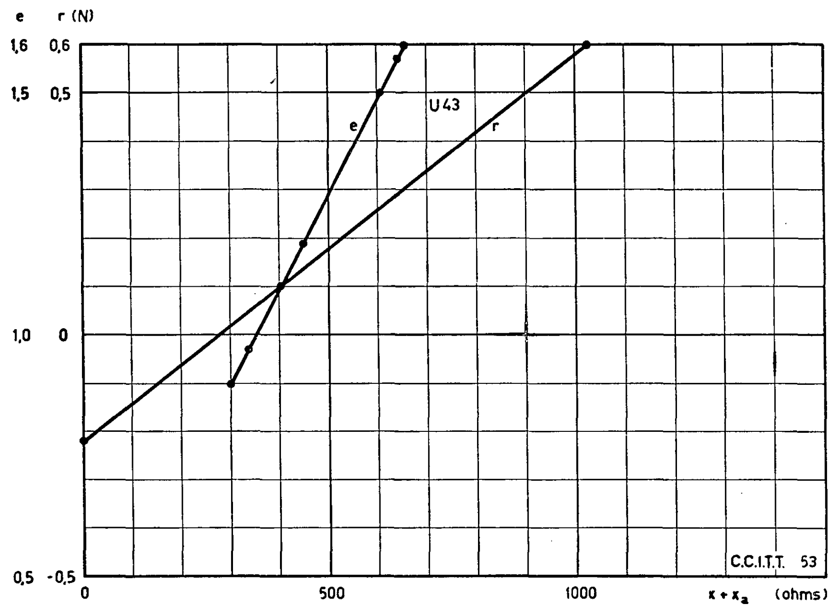


FIGURE 9. — **Nomogram for calculating a subscriber system (France)**

VI. CONTRIBUTION BY THE ITALIAN ADMINISTRATION

Introduction

The transmission plan in the Italian telephone network forms part of the " National telephone regulatory plan " (P.R.T.N.) which has been prepared with the aim of giving the Italian network a rational configuration in keeping with the principles of modern technique.

The plan caters for a division of the network into three parts:

a) District network (star layout) comprising lines connected to a district centre (CD);

b) Secondary network (star layout) comprising connecting lines between CDs and "regional centres " (CC);

c) Primary network (mesh layout), interconnecting the CCs.

The country has been divided into 220 districts grouped in 21 regions, of which the CCs are generally the regional capitals. Of these, Rome and Milan, which are also equipped as international transit centres, are called " national centres (CN) ".

Each district is divided into sectors and the sectors into local networks. This arrangement was devised in order to concentrate:

traffic of an exclusively local interest within the sectors;

traffic for zones influenced by geographical factors, commercial relations, etc. within districts; and

essentially regional traffic within the regions.

The regional centres constitute nodal sorting centres for medium and long-distance traffic (between the various regions).

Transmission plan

The Italian automatic trunk network is now being established in consideration of the following attenuation plan:

a) the nominal equivalent (at 800 c/s) between two district centres must not exceed 0.8 N, including any attenuation in intermediate transit circuits;

b) the nominal equivalent of the circuit from the telephone station to the district centre must not exceed 1.1 N;

c) the attenuation, at 800 c/s, between automatic exchanges (excluding CCs) should not exceed a total of 0.5 N;

d) the sum of the sending and receiving reference equivalents of subscribers' stations must not exceed a total of 0.7 N.

By adding a mean equivalent variation of ± 0.3 N for the chain of circuits serving two CDs, an attenuation of 4.5 N is reached, which is less than the C.C.I.T.T. limit of 4.6 N.

However, it should be observed that the value of 4.5 N is reached only in a very small percentage of cases, as it is seldom that maximum values in one and the same connection will occur simultaneously.

The circuits in the primary network are all four-wire, chiefly of the carrier type. In the regional centres, the circuits are interconnected by selectors for four-wire circuits; four-wire connection is similarly used for interconnection with circuits of the secondary network or the district network (which may be two or four-wire, whereas the modern trend is towards the constitution of four-wire carrier circuits). This helps to contribute corresponding improvement in transmission, particularly where it is necessary to use an alternative route; this is the method indicated in Volume VI of the *Red Book*, page 180, paragraph b).

REFERENCES

[1] Piano Regolatore Telefonico Nazionale. *Gazetta Ufficiale*, 30 December 1957, No. 321.

[2] Forno, A.: Servizio Telefonico in Italia. *National Association of Italian Engineers and Architects*, Milan, 17-20 June 1962.

VII. Contribution by the Nippon Telegraph and Telephone Public Corporation (Japan)

1. The transmission quality between two subscribers is evaluated by means of the nominal A.E.N. *(affaiblissements équivalents pour la netteté)*.

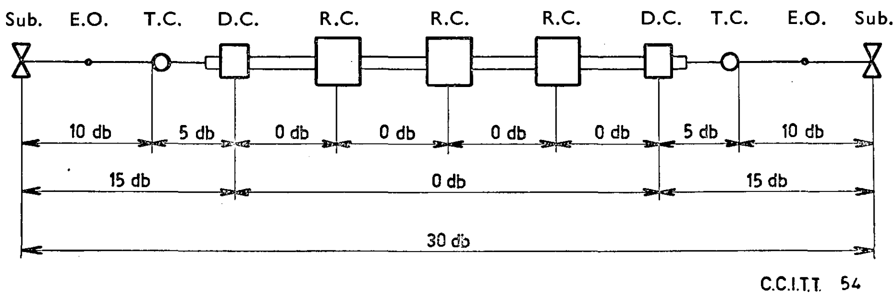
2. The objective for transmission quality is such that any connection between any two subscribers in this country should have a transmission quality of better than 49 decibels expressed in the A.E.N.

3. On the basis of the above objective, transmission qualities are assigned to trunk circuits and subscriber lines in a way to be described later. The transmission loss values assigned to subscriber lines, however, may be exceeded for 10% of subscribers of an exchange for economic reasons.

4. The transmission quality of 49 decibels is divided into:

Telephone set	10.1 db
Impairment due to band limitation	0.5 db
Impairment due to noise	2.9 db
Level variation	4.6 db
Current supply loss	1.0 db
Permissible transmission loss	29.9 db to 30 db

5. The permissible transmission loss is divided among the various classes of lines as shown in Figure 10.



Note. — The four-wire switching facilities should be provided to achieve the above transmission plan in every one of regional and district centres.

FIGURE 10. — **(Japan)**

6. The transmission performance of an exchange is also expressed in A.E.N. It is composed of not only the attenuation losses due to switching equipment and office cables but also reflection losses and noise impairment.

The maximum permissible transmission quality expressed in terms of A.E.N. is assigned to a local exchange as follows:

1 db for a local exchange in multi-exchange areas;

0.5 db for a local exchange in single-exchange areas.

These values were determined on the basis of the actual situation for existing exchanges.

VIII. CONTRIBUTION BY THE NORWEGIAN ADMINISTRATION

The main network will be a combined star and mesh network with up to five different types of exchanges, see Figure 11. (Tielines are established where profitable.)

There will be four-wire through switching in the central parts of the network. The network is based on distributed loss, with a fixed loss of 0.4 N in the terminations and 0.1 N in each link.

It is technically and economically an advantage to use overhead plastic-cable instead of open-air lines to subscribers far from the local exchange. For that reason, the future network planning will be based upon a bigger allotment of reference equivalence for the local network than has so far been usual.

This will be achieved by extensive use of four-wire switching exchanges, transistorized repeaters in two-wire trunk lines, loading of long subscriber lines and introducing of transistor-telephones.

(Annex 4)

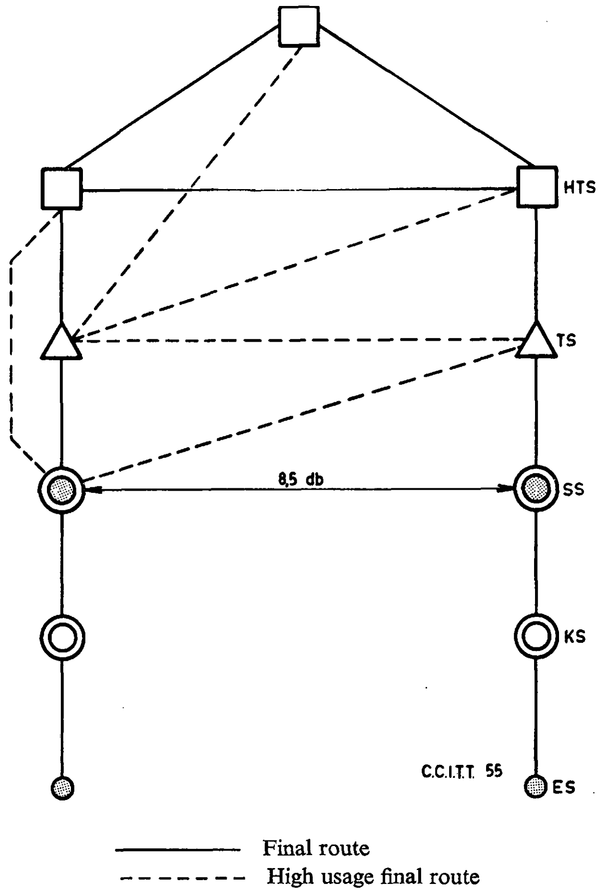


FIGURE 11. — (Norway)

IX. CONTRIBUTION BY THE NETHERLANDS ADMINISTRATION

In drawing up a national attenuation plan, the Netherlands Administration took the following principles as a basis:

1) Total subscriber-to-subscriber attenuation (excluding telephone stations) should not exceed 30 db.

2) This attenuation of 30 db must be apportioned as economically as possible between the various parts of the network.

Explanatory remarks

The over-all standards observed for transmission performance and attenuation are:

10 db	=	excessively loud (undesirable);
10-20 db	=	excellent quality (but too expensive for general application);
20-30 db	=	good quality;
30-40 db	=	inadequate quality;
More than 40 db	=	conversation nearly or quite impossible.

Apportionment of attenuation

As far as apportionment of attenuation is concerned, the Netherlands network is subdivided into:

Interregional networks (between regions);
Primary networks (between regional exchange and nodal exchanges);
Secondary networks (between nodal exchange and local networks);
Local networks.

Traffic efficiency is at its lowest in the local network, because provision has to be made for one pair of wires for each individual subscriber's line.

In all circumstances, the most economical course will be to tolerate the greatest possible degree of attenuation in the local network. Clearly, this means that less attenuation will be available for the others, and hence attenuation in the interregional networks will be very small indeed.

In the present attenuation plan, the maximum tolerable figures for attenuation are:

Interregional network	0 db
Primary network	0 db
Secondary network	3 db
Local network	5 db

The maximum attenuation between two subscribers is then 27 db (including the attenuation of the telephone exchanges and feeding current loss).

Transmission performance in a local network

By reason of the low trunk attenuation (see above), an attenuation of 5 db at 1600 c/s between a connection exchange and the subscriber's set can be tolerated at present.

This attenuation can be ensured by the use of non-loaded cables 5 km (about 3 miles) long, with conductors measuring 0.5 millimetre in diameter), provided there is a telephone of standard type at the end of the circuit. Owing to its inductive impedance, this apparatus will favourably affect attenuation. Although the characteristic attenuation of a cable 3 miles long at 1600 c/s is some 7.3 db, it will be enough if provision is made for a 5-db attenuation if the telephone set is of the standard type.

The tolerable resistance of the pairs, loop-measured, is 1000 ohms at the most. Hence, in drawing up the plan for local cables, it will suffice to observe the following two rules:

a) a loop resistance of 1000 ohms;
b) a cable length of 5 km (about 3 miles).

These standards can be met by use of cables with a diameter of 0.5 millimetre. Formerly trunk circuits took up a considerable proportion of the total available attenuation, and so attenuation less considerable than at the present time could not be tolerated. This led to the use of short cables and to the creation of numerous local exchanges, each of which served a relatively small area.

Hence, in only 1500 instances (out of a million main connections) has the above length (3 miles) been exceeded. In exceptional circumstances special measures are required (for instance, use of cables with conductors of greater diameter).

Sometimes it will suffice to create more satellite exchanges.

By analogy with the trunk-cable network, the network interconnecting the satellite exchanges and that linking these exchanges to the trunk exchanges will have to meet the following requirements:

Satellite exchange-satellite exchange: 6 db maximum, at 1600 c/s;
Satellite exchange-trunk exchange: 3 db maximum, at 1600 c/s.

The interconnecting cables are mostly coil-loaded.

In general, the diameter of the conductors is 0.8 mm, the self-inductance of the coils is 65 mH, and the coil-spacing 3 km (just under two miles). Generally speaking, it would seem well to specify the following points for the transmission characteristics of a telephone exchange:

1. Attenuation as a function of frequency.
2. Crosstalk, namely:
   2.1 between neighbouring equipments;
   2.2 in rack cabling;
   2.3 in cables between selectors;
   2.4 between contact banks;
   2.5 between the two complete circuits which are considered to affect each other most unfavourably.
3. Balance with respect to earth (also at frequencies lower than those of the audio range, to include effects of hum attributable to the mains).
   3.1 for each individual item of equipment (input and output);
   3.2 all the equipment required for setting-up a complete circuit (input and output).
4. Noise and hum.
5. Clicks.
6. Signalling levels.

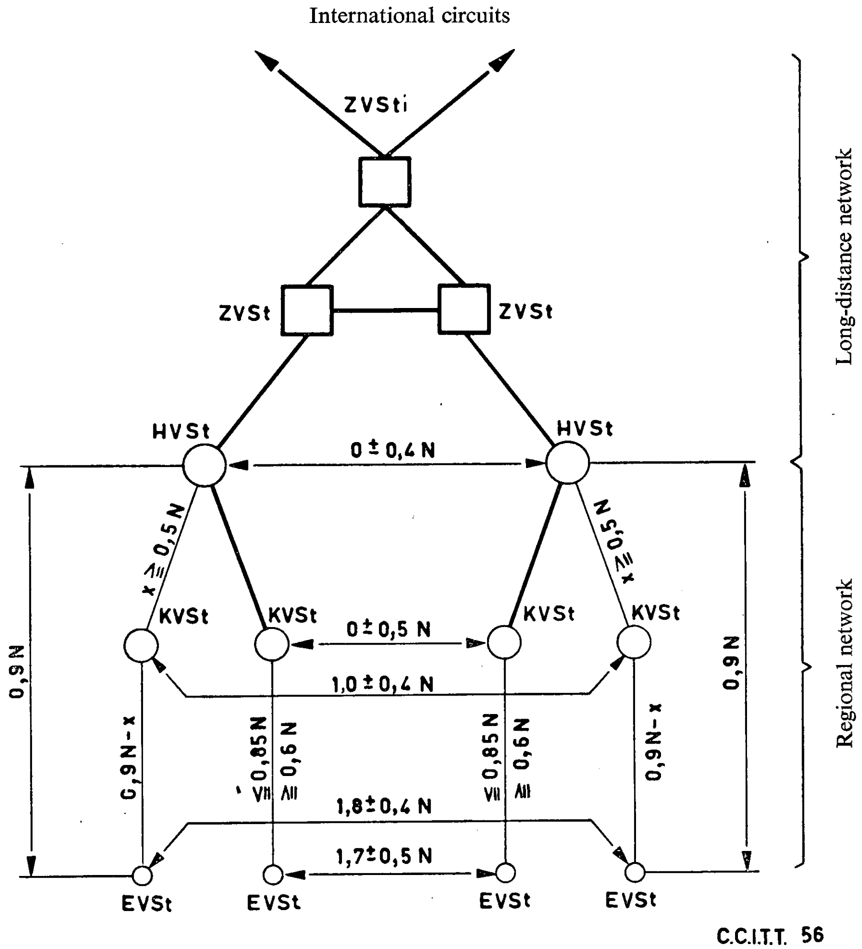
X. Contribution by the Administration of the Federal Republic of Germany

The automatic trunk telephone network of the Federal German Administration has meanwhile been extended in accordance with the attenuation plan published on page 189 of the *Red Book*, Volume V. The new designations of the switching centres are indicated on Figure 12.

The long-distance network now has only four-wire circuits, which are interconnected by selectors for four-wire circuits in the tertiary centres (ZVSt) and the secondary centres (HVSt). The circuits connecting the secondary centres (HVSt) and primary centres (KVSt) are also almost entirely four-wire circuits, but in addition to the primary centres with selectors for four-wire circuits, there are also primary centres (KVSt) with selectors for two-wire circuits. For four-wire interconnection in a primary centre (KVSt), all the two-wire circuits connecting it to a local exchange are equipped with a terminating set. Balancing is adjusted in each terminating set to the circuit concerned. The attenuation of 4 dN of the terminating set can be suppressed by putting out of circuit the attenuation lines of 4 dN at the end of the four-wire circuit that is to be connected (interconnection with attenuation compensation).

All the attenuation values indicated in Figure 12 apply to sections and relate to the interconnection points at the selector contacts. In addition to the nominal attenuation values, the diagram also shows the variations which would be produced allowing for an attenuation variation of ± 2 dN for each circuit with repeaters and quadratic addition of the attenuation variations at all the circuits of a chain circuit. The maximum equivalent (including these attenuation variations) between two local exchanges (EVSt) is 2.2 nepers. Since all the four-wire trunk circuits are connected in series with an insertion loss of 0 neper and since the tertiary centre with international lines (ZVSti) is reached from each primary centre (KVSt) by three circuit sections at the very most, the Federal German attenuation plan also conforms to the new C.C.I.T.T. recommendations relative to a new international interconnection plan.

The maximum sending and receiving reference equivalents of a subscriber line are respectively 1.2 N and 0.2 N with respect to the local exchange (EVSt). These values can be exceeded by 0.3 N



———— Two-wire circuit	━━━━ Four-wire circuit

EVSt	Local exchange	ZVSt	Tertiary centre
KVSt	Primary centre	ZVSti	Tertiary centre with international lines
HVSt	Secondary centre		

FIGURE 12. — **Attenuation diagram for the automatic trunk network (Federal Republic of Germany)**

for extensions. Since the attenuation of the section between a local exchange (EVSt) and a four-wire interconnection point of the chain of trunk circuits is 0.9 N, 2.4 N and 1.4 N are obtained for the maximum national sending and receiving reference equivalents respectively, which conforms to Recommendation P.11 for national sending and receiving reference equivalents.

The network of circuits indicated on Figure 12 will be completed by a network of direct circuits. Direct circuits between switching centres will be established in all cases where a sufficient volume of telephone traffic justifies the provision of special systems of circuits. The maximum equivalent of 2.2 N between local exchanges (EVSt) will likewise not be exceeded for telephone calls set up over these direct circuits.

The Federal German Administration is considering the following reference values, in addition to the existing recommendations of the C.C.I.T.T., as regards the operation from the transmission aspect of new equipment in local exchanges and for subscriber lines permanently connected to them:

a) The insertion loss of a local exchange (including feeding bridge) should not exceed 15 cN at 800 c/s.

b) The crosstalk attenuation between two circuits should not be less than 8 nepers.

c) The balance attenuation of the switching equipment should, in the speaking position, be at least 5 nepers over the limits (300 to 3400 c/s) of the transmission band.

d) The psophometric voltage between the two wires should not, at the exchange distribution point, exceed 0.35 mV (weighted with filter A) in the case of unconnected external lines and with the internal line terminated by 600 ohms.

e) The maximum loop resistance of subscribers' lines, including telephone sets, may attain 1500 ohms.

f) A value of 20 000 ohms may be allowed as the lower limit for insulation resistance of subscribers' lines and 50 000 ohms for the lines connecting the exchanges within a locality.

g) For the junction lines, the alternating current induced continuously along the line a value of 15 $V_{r.m.s.}$ may be allowed. Efforts should be made to attain 20 $V_{r.m.s.}$ where the frequency of the inducing alternating current is 16⅔ and 50 c/s. In the case of peaks of short duration (less than 400 ms), exceeding 40 $V_{r.m.s.}$, existing communications cannot be freed and there must be no false occupation or false metering.

A steady value of 65 $V_{r.m.s.}$ is admissible for the induced voltage on subscribers' lines affected by a.c. interference.

h) Allowable variations of the 60 volts supply voltage are included between +6 V and −3 V. In the event of mains failure, the voltage may drop to 54 V for less than 1 second without interrupting the calls in course.

Classification of microphone and receiver insets into groups

The reference equivalent of a telephone set depends greatly on the sensitivity of the transmitter and receiver insets. The sensitivity of the transmitter inset also depends greatly on the magnitude of the feeding current. In this connection large differences may be noted, not only between insets of different manufacture, but also between insets of the same manufacture; these differences may be as large as 1 neper, or even larger than 1 neper in the case of poor manufacture. Even in the case of good manufacture, a greater tolerance must be left to the manufacturer if the production is to remain economical. Hence, it can easily be seen that insets may be classified according to their sensitivity and that the more sensitive insets may be used for the longer subscriber lines and that the less efficient insets may be used for the shorter lines.

It has been observed that, up to a line resistance of 500 ohms, the reference equivalent is essentially determined by the d.c. resistance, whatever the diameter of the conductors. If the resistance is greater than 500 ohms the capacity of the cable becomes more evident. Nevertheless, since, in the case where the line resistance is greater than 500 ohms, uniform use is made of conductors of the same diameter, there exists in a general way a definite relation between the d.c. resistance and the reference equivalent, i.e. the reference equivalent of a subscriber's line may be denoted by its d.c. resistance. The increase in attenuation due to the subscriber's line is greater for sending than for receiving because the feeding current losses are added to the line attenuation. The increase in the reference equivalent due to the subscriber's line approximately corresponds to the equivalent of the subscriber's line measured at 1300 c/s, which agrees satisfactorily with the results of A.E.N. determinations on unloaded cables for which a comparison frequency has been fixed of 1300 to 1700 c/s.

It is possible to compensate for the reference equivalent of the local system by a corresponding grading of the lines and of the transmitter and receiver insets. In this way, it is possible to obtain a considerable reduction in the variations of attenuation. A grading of transmitter insets in grades of 0.4 neper and of receivers in grades of 0.3 neper is sufficient in practive. Hence, it is possible to obtain the arrangement of gradings of lines and insets as shown in the table below.

Line grading with total resistance in ohms	0 to 250	250 to 500	500 to 750
Transmitter insets	Grade I	Grade II	Grade III
E.R.E. [1] (nepers)	0.9 to 0.5	0.5 to 0.1	< 0.1
Receiver insets	Grade I	Grade II	Grade III
E.R.R. (nepers)	0 to −0.3	−0.3 to −0.6	−0.6 to −0.9
Local system			
E.R.E. (nepers)	0.5 to 1.25	0.5 to 1.25	< 1.25
E.R.R. (nepers)	−0.3 to +0.2	−0.4 to +0.2	−0.4 to +0.2

[1] E.R. = Reference equivalent
E = Sending
R = Reception

With this grading, the following results are obtained: for a local telephone system, i.e. from the telephone set up to and including the local exchange, the reference equivalent lies between the limits of 0.5 neper and 1.25 neper (average 0.9 neper) and the receiving reference equivalent between the limits of −0.4 neper and +0.2 neper (average −0.1 neper). These values are in good agreement with the results of measurements made on the German local systems at the C.C.I.T.T. Laboratory, i.e. 1.25 neper for the sending reference equivalent and −0.2 neper for the receiving reference equivalent. The variation of the reference equivalent of a local system, for sending and for receiving, is even then relatively large, even with compensation for attenuation, and it is, in most cases, larger than the difference between the A.E.N. value and the value of the reference equivalent.

Since the microphone and receiver insets are both easily changed, the method based on the compensation of the attenuation can be easily applied in practice without need for additional staff and without extra expense. Before use, the microphone and receiver insets are measured by the reference equivalent measuring apparatus used by the German Federal Telephone Administration and are marked with the grading corresponding to their efficiency. In addition to the reference equivalent, other characteristics are measured which have an influence on the articulation. The maintenance technician dealing with faults has with him some microphones and receiver insets of different gradings and fits into the subscriber's telephone apparatus insets of the grade corresponding to the total resistance of the subscriber's line.

It is recommended that the quality of the microphone and receiver insets be verified every two years.

If only a small tolerance is allowed for the reference equivalent of the microphone and receiver insets it is also possible to obtain economically a compensation of the attenuation of the local system by grading the diameter of the conductors according to the length of the subscriber's line.

XI. Contribution by the United Kingdom Administration

Limits applied in national trunk and local networks

As a result of subjective tests made by the United Kingdom Administration some years ago it was decided that the transmission loss at 800 c/s introduced between minor exchanges should not exceed 20 db [1]. Switching and reflection losses are included in this allowance. The reference

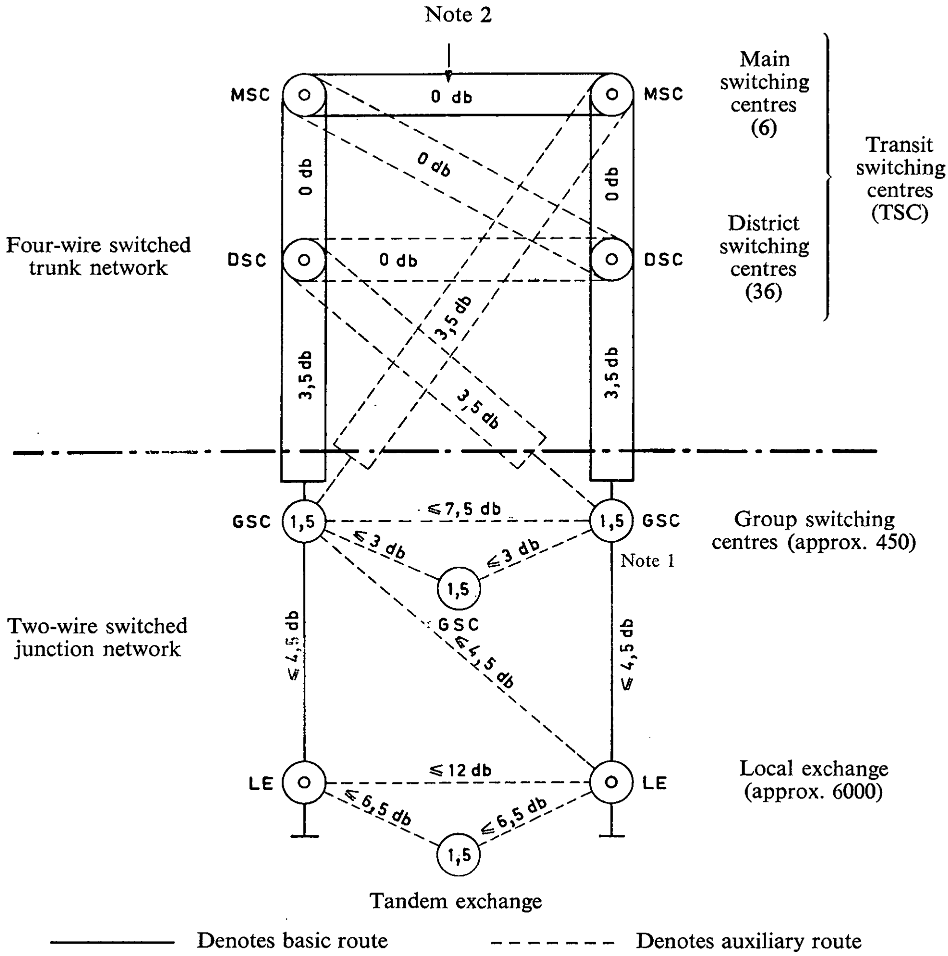


Figure 13. — **New trunk switching and transmission plan (United Kingdom)**

Notes.

1. Nominal exchange loss. This is 0 db at terminal minor exchanges because the exchange loss is included in the allowance for the subscriber's set and subscriber's line.
2. Nominal loss of the circuit.
3. If one or the other of the circuits leaving the group switching centre is repeatered it is adjusted to a nominal loss of 3 db.

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[1] This assumes no significant attenuation/frequency distortion. In the case of unloaded audio cable the attenuation at 1600 c/s is used for planning purposes.

equivalent of two limiting subscribers' lines and instruments currently in use in United Kingdom local networks is 13 db (adding together the separately measured sending and receiving reference equivalents). Thus the total maximum reference equivalent for a national connexion may be quoted as 33 db. (N.B.: The concept of " reference equivalent ", although useful for describing a network, was not used by the United Kingdom Administration for planning the network.)

The way in which the 20-db loss between minor exchanges is apportioned among the various classes of circuit is shown in Figure 13. It is patently desirable to be able to assign as much loss as possible to the unamplified junction network (minor exchange-group switching centre) where a reduction in allowable loss of even 1 db means the expenditure of much capital, and to assign as little as possible to the mostly amplified trunk network where amplifier gain must be provided anyway and extra gain costs very little. However, a limit is set by the balance return loss that can be guaranteed at the terminating set of the trunk circuit at the group switching centre which is a two-wire switching point. In order to avoid an expensive scheme of impedance correction at all group switching centres throughout the country and also to avoid having to use echo suppressors, the loss of the trunk network measured between two-wire points was fixed at 7 db. This loss is independent of the number of trunk circuits, thereby avoiding pad switching. The transmission loss of each of the two junction circuits in a maximum connection is planned to be 4.5 db which corresponds to about 12 miles (19 km) of unamplified, loaded audio cable (88 mH at 1.83 km spacing, 0.9 mm diameter copper wire).

There thus remain 4 db for switching and reflection losses of which 3 db are shown in Figure 13 as two switching-point losses at group switching centres. Reflection losses are significant at the higher audio frequencies on unloaded audio cables because of impedance mismatch at the two-wire switching point. There are no significant switching and reflection losses in the trunk network because this is four-wire switched as indicated and most of the circuits are routed on carrier systems and present a good, designed impedance at the audio switching point.

A useful general reference is:

Tobin, W. J. E.: and Stratton, J.: A new switching and transmission plan for the inland trunk network. *P.O.E.E.J.*, Volume 53, Part 2, July 1960.

Limits applied in local exchanges

The fundamental principle is that the transmission performance of any subscriber connected by a subscriber's line to a local exchange should be equal to or better than a minimum standard represented by a given combination of a subscriber's station, a subscriber's line and the exchange feed bridge. The design of the local network is such that most subscribers get transmission performance well above the minimum.

The minimum standard and the method of determining the relative transmission performance for sending and receiving are described in detail in Annex 1 of Part II of Volume V of the *Red Book*.

XII. Contribution by Singapore (Singapore Telephone Board)

C.C.I.F. standards; maximum attenuation of junction links 8 db; maximum sending and receiving for local lines 10 and 5 db respectively.

XIII. CONTRIBUTION BY THE SWEDISH ADMINISTRATION

The setting-up of our national transmission plan for a fully-automatized telephone network has been governed essentially by the following factors:

1. Reference equivalent for sets provided at present is 3 ± 2 dN for sending and −5 ± 2 dN for receiving.

2. Division of the country into exchange areas, numbering areas and transit routing areas.

Note. — Telephone apparatus now being made has reference equivalents of 1 ± 2 dN (sending) and −6 ± 2 dN (receiving). It is intended to use these values in future.

3. The existing line network, in particular the long-distance lines and such shorter cables as were already included in automatized areas at the moment of preparation of the plan. In this respect, our initial situation in preparing for long-distance automatization was unfavourable, seeing that the existing plants to be considered in certain decisions were of comparatively great extent. They included mainly coil-loaded cables, but the construction of a carrier frequency cable network had begun and was assumed to prevail in future extensions of the long-distance line system.

4. Existing automatic exchange equipments. Such equipments were already in service, at the moment of planning, in many centres of numbering areas and in many auxiliary exchanges, all of them of the two-wire type. In the building-up of transit centres for the automatic long-distance service, on the other hand, the designers had full liberty as far as technical features of transmission were concerned.

An aim pursued in planning for a national automatic network has, of course, been to fulfil, for all domestic and international calls, the C.C.I.T.T. requirement of a reference equivalent of 4.6 nepers and, if possible, to achieve a lower value (see an article by G. Swedenborg: " A survey of the development of telephone apparatus from the speech transmission aspect " appearing in *Tele*, English edition No. 1/1953. To achieve this object at reasonable cost, amplification should be used to the largest possible extent, within the limits set by echo and singing phenomena. The solution arrived at in this respect in described by B. Bjurel, H. O. Björk and E. Waldelius in another article published in the aforesaid issue of *Tele*: " Technical viewpoints respecting automatization of trunk traffic ", pp. 21-24. By way of a summary, the principal items of the transmission plan in question are given below:

a) Four-wire pad switching is used in the switching centres to which the numbering area centres (NAC) are attached.

b) To obtain the simplest possible level conditions, all links of a connection NAC-NAC have the nominal attenuation 0 in transit service. A further 0.3 or 0.5 neper will be added at each terminal point, depending on the type of the component links, so that in the most unfavourable case, the total attenuation NAC-NAC will attain nominally 1.0 neper (in the case of carrier frequency only 0.6 neper).

c) The attenuation over the line from a subscriber to an NAC must not exceed 1.5 neper, including the current feed attenuation (1.2 neper from NAC to subscriber). The attenuation in local and toll networks respectively are thus added; their mutual proportion is decided in each case on the basis of economical considerations. In this way it is possible, for instance, to achieve economies in the local network, when the distance between the local exchange and the NAC is small.

d) Taking into account the tolerances applicable to the performance of telephone sets and the variations in time of line attenuations, the maximum reference equivalent for national calls will be about 4 nepers. In the majority of cases, however, transmission conditions for interurban calls will be considerably more favourable. Further improvement is however desirable, especially in the international service, where the fulfilment of the C.C.I.T.T. requirements respecting the reference equivalent of the national sending system presents certain difficulties.

Note. — The Swedish Administration is prepared to introduce the over-all reference equivalent of 4.15 nepers as soon as this is accepted by the C.C.I.T.T., provided that the values chosen for the reference equivalent (sending and receiving) are 2.4 and 1.4 nepers respectively.

Transmission performance of local exchanges

1. The highest admissible operating attenuation in a local exchange, measured between the test jacks of the main distribution frame for incoming and outgoing lines is 2 dN at 300 c/s and 1 dN in the range 800-3000 c/s.

2. The lowest admissible crosstalk attenuation between any two call connections in a local exchange is shown in Figure 14.

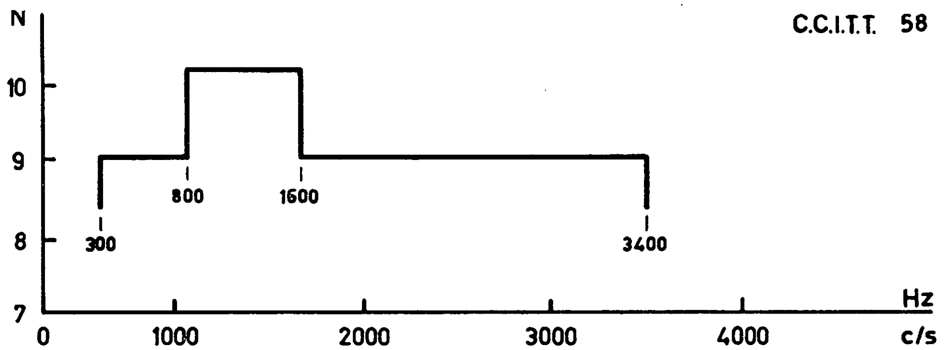


FIGURE 14. — (Switzerland)

The crosstalk attenuation is measured at the test jacks of the main distribution frame, the circuits being terminated by a resistance of 600 ohms.

The crosstalk attenuation between different exchange apparatus should be proportioned so that the values indicated above for the whole exchange are obtained. This implies, as a rule, that the crosstalk attenuations to be chosen between different exchange components must be about 1 N higher than the values shown on the diagram.

XIV. CONTRIBUTION BY THE SWISS ADMINISTRATION

The transmission characteristics of overhead and underground lines in local and regional systems (including cables between exchanges, equipment and exchanges inserted therein) must meet certain clearly defined desiderata with regard to:

— equivalent;
— attenuation distortion;
— signalling and feeding currents;
— crosstalk, and
— noise voltage.

The attenuation plan, represented in the diagram of Figure 15, represents the basis of the " Instructions for the planning of local and regional systems as a function of transmission performance "; it shows the maximum equivalents for 800 c/s. Apportionment of attenuation among the local and regional systems is dependent on financial considerations.

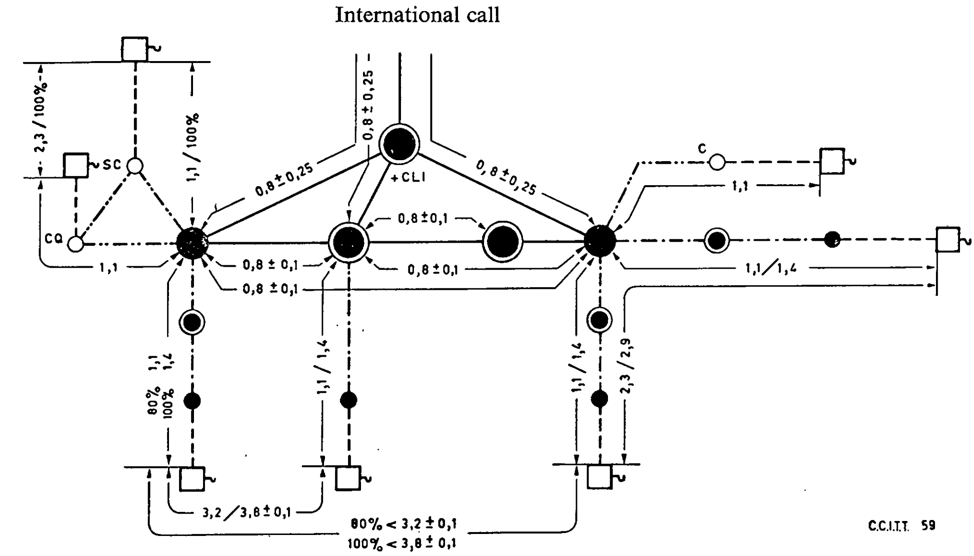


FIGURE 15. — **Apportionment of the equivalent for trunk, regional and local connections (Switzerland)** (maximum values in nepers for 800 c/s)

Legend

CLI		International terminal exchange
⬤ CNI	Trunk exchange 0.1 N	Trunk intermediate exchange with trunk terminal exchange and main exchange of a group of networks
● CTI		Trunk terminal exchange with main exchange of a group of networks
◉ CN		Intermediate exchange
● CT		Terminal exchange
○ C, CQ, SC		Local exchange, district exchange and sub-exchange
———————		Trunk line
— · — · —		Regional line
—· ·—· ·—		Inter-exchange line
- - - - - - -		Subscriber's line

In theory, the maximum resistance of a subscriber's line is 700 ohms. The attenuation distortion (over the band from 300 to 3400 c/s and relative to the equivalent at 800 c/s) in each local system must not exceed 0.5 N in a trunk line or between two regional systems. If this proves impossible, the subscribers' lines must be given a certain amount of coil-loading.

Near-end crosstalk should be more than 8.5 N for 1600 c/s.

For brevity, further details will be given about attenuation distortion, signalling and feeding currents, crosstalk and noise voltage.

	Comparative table	Send	Receive
A	Trunk exchange + terminals of a subscriber's set on an intermediate or terminal exchange in this regional system: maximum equivalent for 100%/80% of the subscribers in each local exchange network	1.4/1.1 N	1.4/1.1 N
B	Trunk exchange + terminals of the subscriber's set in the local network of this exchange: maximum equivalent for 100% of the subscribers in this network	1.1 N	1.1 N
C	Reference equivalent for the subscriber's set (minimum feeding current, without subscriber's line)	0.4 N	0 N
D	Effect of attenuation distortion from the subscriber's line and loss due to mismatching	0.2 N	0.2 N
E	Attenuation in the trunk or international exchange	0.1 N	0.1 N
F	Trunk line	0.8 N	
	Reference equivalents		
	National trunk connection between two subscribers in regional systems: Σ ACDEF	4.6/4.0 N	
	C.C.I.T.T. recommendation [1]	4.6 N	
	National trunk connection between two subscribers in the local networks of these trunk exchanges: Σ BCDEF	4.0 N	
	C.C.I.T.T. recommendation [1]	4.6 N	
	International connection for a subscriber of Group A: Σ ACDE	2.1/1.8 N	1.7/1.4 N
	C.C.I.T.T recommendation	2.1 N	1.5 N
	International connection for a subscriber of Group B: Σ BCDE	1.8 N	1.4 N
	C.C.I.T.T. recommendation	2.1 N	1.5 N

[1] Same recommendation as for an international connection.

In the above table, reference equivalents appearing in unfavourable operating circumstances are compared with the figures recommended by the C.C.I.T.T. Hence, the maximum reference equivalent of a national trunk call might attain 4.6 N.

However, within the local system of a regional zone, 80% of subscribers' lines will have a lesser attenuation, and 100% of the subscribers' lines in the local network of a trunk exchange have an attenuation of less than 4 N.

Local or regional calls give substantially lower values of attenuation. Statistics show that only rarely are the limit figures reached. Attenuation conditions are even better than this for international calls.

This attenuation plan was drawn up taking account of the existing telephone network, technical and financial considerations and the possibility of improvement. One improvement, now being made, consists in the prolongation of four-wire connection through the trunk exchanges. In this way, it will be possible, by compensating the attenuation, to maintain the equivalent of 0.8 ± 0.1 N up to the nodal and terminal exchanges linked directly to the trunk exchanges. The gain thus obtained increases transmission performance.

Similar action is under consideration for the local systems of trunk exchanges. In addition a new telephone, with automatic adjustment of sensitivity, such as would to some extent offset the attenuation of the subscriber's line, is being studied. These improvements would make it possible to conform to the values laid down in the new transmission plan now being considered by the C.C.I.T.T.

XV. Contribution by the Czechoslovak Administration

The transmission plan for the national telephone network was devised on the basis of the following principles:

a) the line equivalent for a long-distance call should be not more than 3.6 N (figure 16);

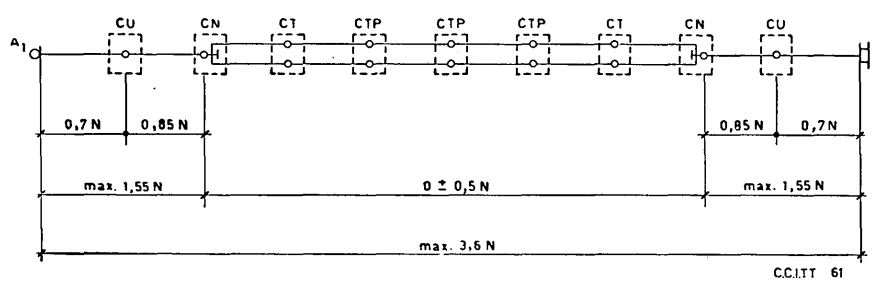
b) the relative equivalents of the microphone and the receiver should be 6 and 0 dN at the present time. Later it is intended to have relative equivalents of 3 dN and −3 dN respectively.

c) the call in the nodal exchange is set up as a two-wire call; the call from the nodal exchange to a higher long-distance level is set up on a four-wire connection;

d) four-wire metallic lines will in future (if possible) be connected on a four-wire basis;

The quality of a telephone call is assessed according to the clarity, volume (loudness) and reliability of a connection. The final transmission quality is given by the following conditions:

- the telephone band must be transmitted in the range from 300 to 3400 c/s;
- the psophometric electromotive force of a telephone circuit should not exceed 2 mV at a point of −0.8 N relative level when measured with a psophometer;
- the feeding bridges for a local connection should not cause an attenuation of more than 15 cN.



CU = trunk exchange
CN = " nodal " exchange
CT = transit exchange
CTP = main transit exchange

Figure 16. — **Reference equivalent between two subscribers (Czechoslovakia)**

ANNEX G

(Geneva, 1964; quoted in Recommendation P.51)

ARTIFICIAL MOUTH USED BY THE SWEDISH ADMINISTRATION

TECHNICAL INFORMATION AND DATA FURNISHED BY THE SWEDISH ADMINISTRATION

Design of the artificial mouth

The artificial mouth consists basically of the following:

— a pressure generator (artificial mouth and head);

— a power amplifier, with variable gain.

Correcting networks are inserted in the amplifier circuit in order to compensate for the linear distortion of the pressure generator.

The entire mechanism can be connected to a complex sound generator (with or without a weighted spectrum) or to an oscillator.

Figure 1 below shows the essentials of the transmission chain. The whole equipment is in the form of elements mounted on a rack, the artificial head being independent. The head may be placed on a special support so that it can be tilted in such a way that, when a sending system is being measured, the telephone handset is in the specified position (see Recommendation P.45, B, c, on page 71, Volume V of the *Red Book*).

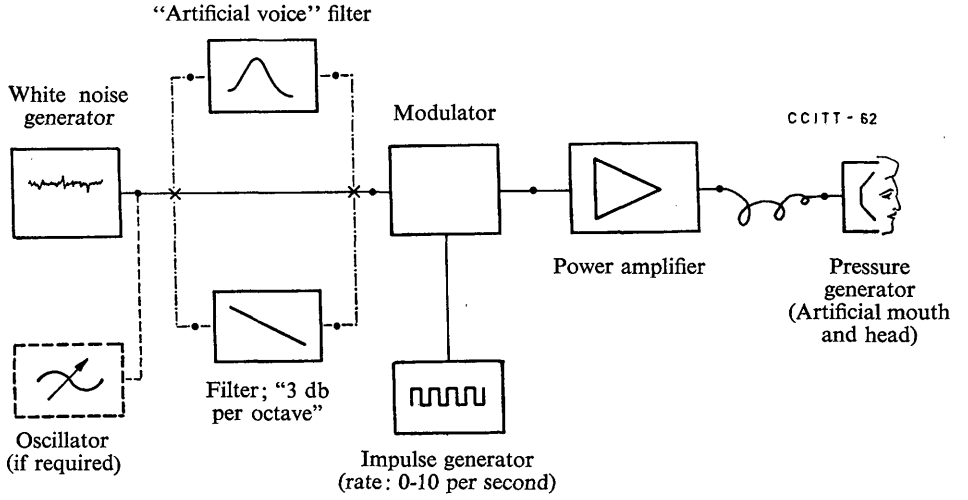


FIGURE 1. — **Schematic diagram of the artificial mouth of the Swedish Administration**

1. *Pressure generator (artificial mouth and head)*

The artificial mouth is formed by a loudspeaker situated inside a cavity, of which the output orifice constitutes the artificial lips; these imitate the shape, profile, and dimensions of real lips. The artificial mouth is thus an acoustic radiator capable of producing a complex sound or a pure tone in front of a microphone. Since a basic feature of an artificial mouth is the fact that the characteristics of the acoustic field it creates are to be comparable to those of the sound field produced by the human mouth—mainly at a short distance from the lips—the human nose and chin have been imitated in the profile of the artificial head. The head itself is made of lightweight metal, the surface being covered with a resilient plastic skin to simulate the surface absorption of the human head. A diagram of the artificial head is given in Figure 2.

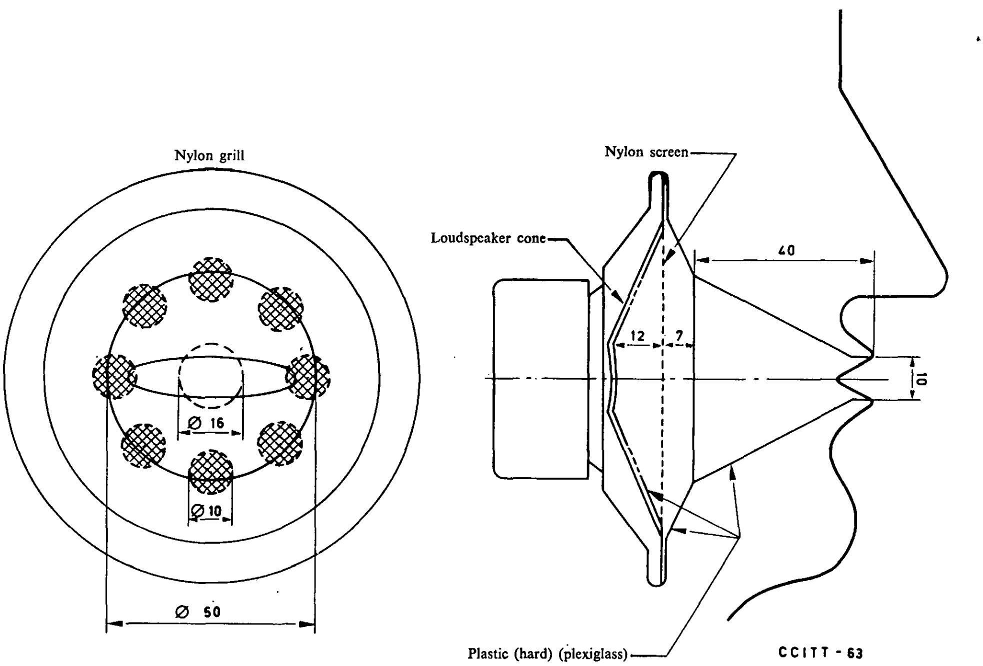
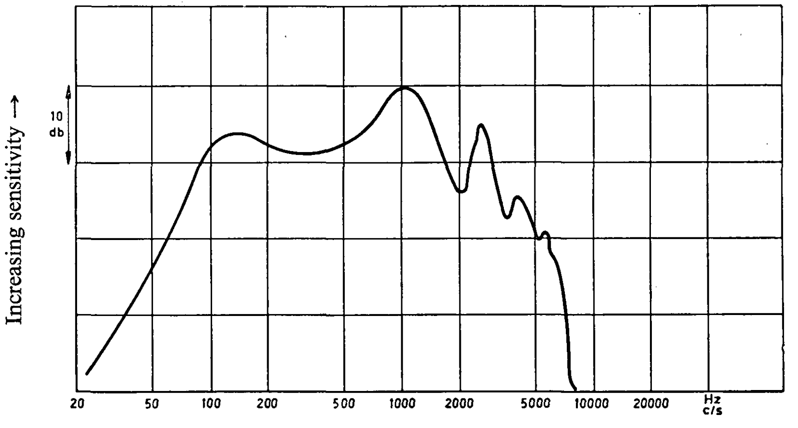
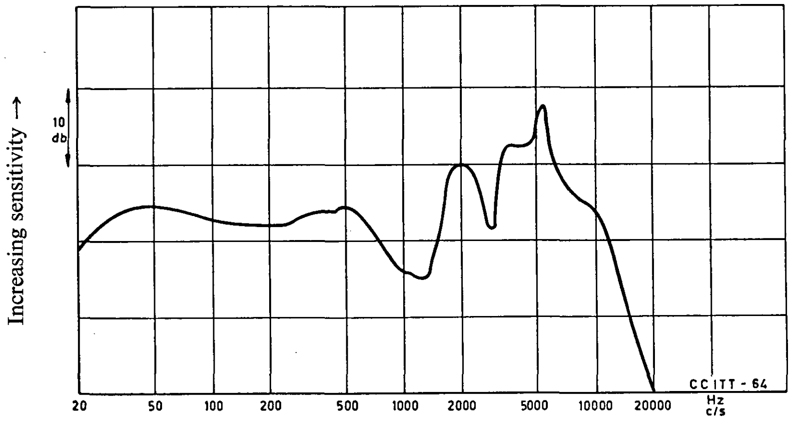


FIGURE 2. — **Diagram of the artificial mouth and head of the Swedish Administration**

The sound source is the loudspeaker. Its cone is connected with the output orifice (artificial lips) by means of a funnel-shaped cavity within which screens (or netting) of fine nylon cloth are placed to produce a particular acoustic attenuation. The shape of the cavities and their acoustic attenuation were determined empirically to obtain, in front of the artificial lips, a " pressure " characteristic for a constant applied voltage, which is independent (to some extent) of the frequency. Further adjustments to achieve a pressure response curve independent of the frequency (tolerances ± 2.5 db from 100 to 6000 c/s) are made by appropriate electric equalizer networks inserted in the power amplifier. The cavity, situated behind the loudspeaker cone, is filled with a plastic foam material. Figure 3a shows the constant-voltage " frequency-pressure " characteristic at the loudspeaker input, the pressure being measured at a point on the axis of the output orifice at a distance of 2 cm.



a) the pressure generator (artificial mouth and head)



b) the power amplifier

FIGURE 3. — **Frequency response curve with constant input voltage**

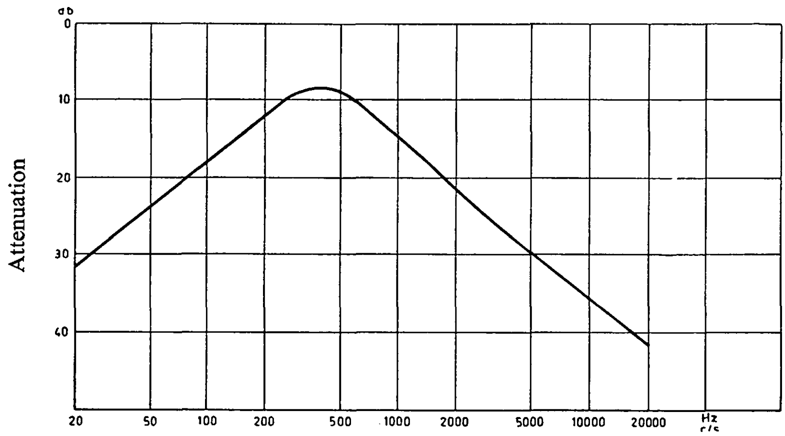
2. *Power amplifier*

The amplifier input (input impedance 27 000 ohms) may be connected to a source of complex sound (white noise or weighted noise) or to an oscillator. The output is directly connected to the loudspeaker input (impedance 20 ohms). Four equalizer networks eliminate irregularities in the response curve of the pressure generator. The amplifier output transformer is tuned to provide an increased gain at high frequencies; at the input, a T network provides a predetermined loss at frequencies in the region of 300 c/s.

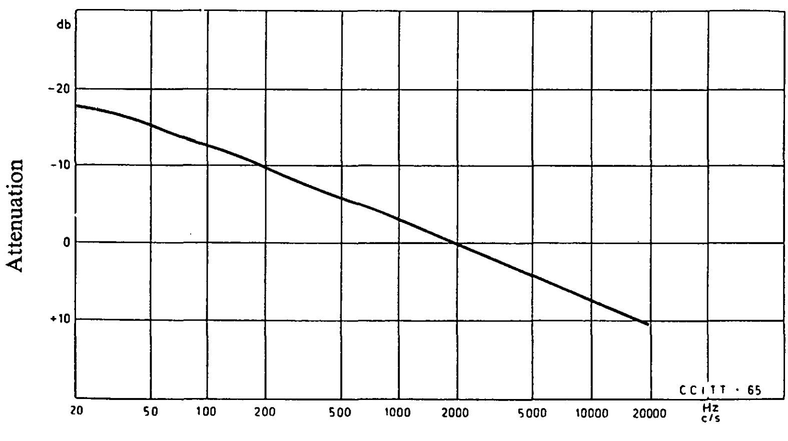
The frequency response characteristic of the amplifier is illustrated by the curve in Figure 3b. The amplifier gain can be adjusted as required by means of a potentiometer. When the amplifier gain is adjusted to its maximum value, an acoustic pressure of 10 dynes/cm^2 can be obtained at a distance of 2 cm from the artificial lips which is free from appreciable non-linear distortion, i.e. the non-linear distortion factor is less than 1 % throughout the useful frequency range. For the same gain adjustment, if the voltage at the amplified input is increased, it is possible to obtain much higher pressures of the order of 100 dynes/cm^2; in this case, the non-linear distortion factor is about 3 % (except for the frequency 2000 c/s, where $k = 8\,\%$).

3. *Noise generator and filters*

A generator with a continuous and uniform spectrum (white noise) may be applied at the power amplifier input either directly or through particular filters. A filter, called a " voice filter ", makes it possible to obtain a sound signal with a spectral composition similar to that of the average human voice. With a second filter, the " attenuation frequency " characteristic of which increases by 3 db per octave, we can obtain a constant power level in each individual band of an octave or fraction of an octave filter, during spectral analysis of white noise. The " attenuation frequency " characteristic curves are shown in Figures 4a and 4b, respectively.



a) the spectrum of the artificial voice



b) a loss of 3 db per octave

FIGURE 4. — **Response curve of filters**

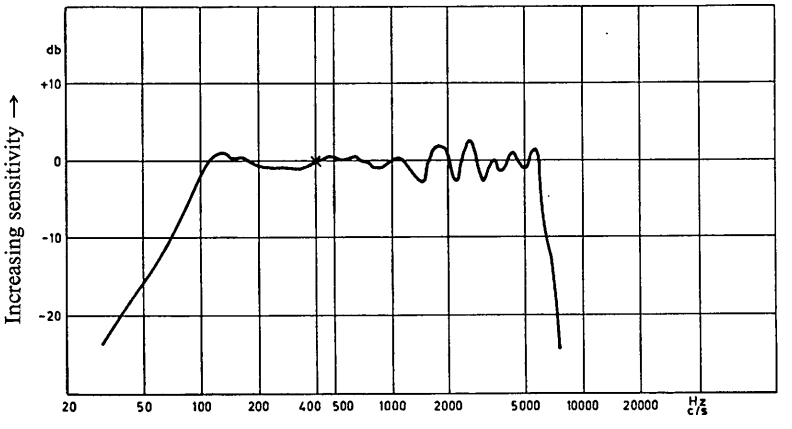
4. *Impulse generator-modulator*

Experience shows that, if results are to correspond closely with those obtained from tests with the human voice, besides using an artificial voice the emitted signal must be made discontinuous in order to simulate the human voice more effectively. For this purpose an impulse generator-modulator permits a modulation with a rhythm adjustable from 1 to 10 per second.

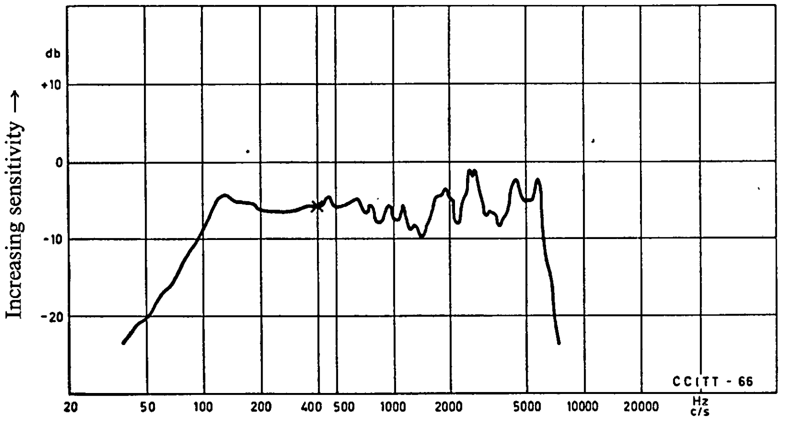
The function of the modulator is to interrupt the signal periodically and to obtain a series of rectangular impulses. This is done by electronic control of an attenuation by a symmetrical square wave produced by a free-running multivibrator. A rhythmic emission results, comparable to the emission of syllables during normal speech.

The complete equipment thus provides an artificial mouth which readily gives the normal speech power for voice/ear measurements. It was designed for a frequency range of from 100 to 6000 c/s.

Figures 5 a and b show the characteristic " sensitivity-frequency " curves of the amplifier/artificial mouth equipment taken at a point on the axis 2 and 4 cm from the artificial lips, respectively; the voltage applied at the amplifier input is independent of the frequency.



a) at a point 2 cm from the artificial lips



b) at a point 4 cm from the artificial lips

FIGURE 5. — **Frequency-response curve of amplifier/artificial mouth equipment with constant voltage at the power amplifier input (relative values—reference frequency 400 c/s)**

APPENDIX

Measurements carried out by the C.C.I.T.T. Laboratory

Measurements were made by the C.C.I.T.T. Laboratory; the results are given below. The tests were carried out in the Laboratory anechoic chamber.

The equivalent lip position was considered to be directly at the orifice of the artificial mouth (see point 1 above).

Figure 6 illustrates schematically the experimental arrangements.

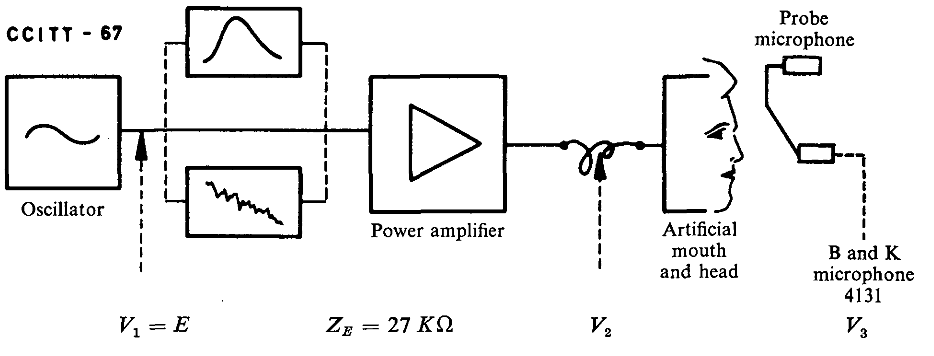


FIGURE 6

The angle of incidence of the microphone was 0°, i.e., the plane of its diaphragm was parallel to the wave planes; the free-field frequency response of this type of microphone 4131 is thus largely independent of the frequency throughout the useful frequency band 100-10 000 c/s.

Measurement results

Tables 1 to 6 and the curves in Figures 7 to 12 show the characteristics of the artificial mouth.

TABLE 1

Sensitivity-frequency characteristic values for the artificial mouth of the Swedish Administration for a pressure of 5 dynes/cm² at 1000 c/s at a distance of 5 cm

(db relative to 1 volt)

c/s	V_1	V_2	V_3	Values of V_3 referred to 1000 c/s
100	−30.6	− 4.9	−39.0	− 5.3
200		− 5.2	−36.7	− 3.0
300		− 3.9	−36.9	− 3.2
400		− 3.2	−35.9	− 2.2
500		− 3.3	−35.4	− 1.7
600		− 4.3	−34.7	− 1.0
700		− 6.0	−34.8	− 1.1
800		− 8.9	−35.3	− 1.6
900		−10.7	−34.8	− 1.1
1000	−30.6	−11.5	−33.7	0
1200		−12.5	−35.8	− 2.1
1500		− 7.7	−36.2	− 2.5
1800		+ 0.5	−32.7	+ 1.0
2000		+ 2.5	−35.8	− 2.1
2500		− 1.0	−32.8	+ 0.9
3000		− 6.7	−37.0	− 3.3
3500		+ 3.9	−36.7	− 3.0
4000		+ 4.6	−34.9	− 1.2
4500		+ 4.4	−33.6	+ 0.1
5000	−30.6	+ 6.2	−35.9	− 2.2
6000	−30.5	+ 6.9	−34.9	− 1.2
7000	−30.4	+ 0.4	−46.4	−12.7
8000	−30.3	− 1.4	−63.6	−29.9
10000	−30.2	− 3.2	−66.7	−33.0

TABLE 2

Sensitivity-frequency characteristic values for the artificial mouth of the Swedish Administration for a pressure of 15 dynes/cm² at 1000 c/s at a distance of 5 cm

(db relative to 1 volt)

c/s	V_1	V_2	V_3	Values of V_3 referred to 1000 c/s
100	−30.6	+ 5.1	−29.1	− 5.4
200		+ 4.8	−27.2	− 3.5
300		+ 6.1	−27.1	− 3.4
400		+ 6.8	−25.9	− 2.2
500		+ 6.7	−25.3	− 1.6
600		+ 5.7	−24.7	− 1.0
700		+ 4.0	−24.7	− 1.0
800		+ 1.1	−24.9	− 1.2
900		− 0.7	−24.7	− 1.0
1000	−30.6	− 1.5	−23.7	0
1200		− 2.5	−25.6	− 1.9
1500		+ 2.3	−26.1	− 2.4
1800		+10.5	−23.9	− 0.2
2000		+12.5	−25.7	− 2.0
2500		+ 9.0	−22.6	+ 1.1
3000		+ 3.3	−27.2	− 3.5
3500		+13.9	−28.0	− 4.3
4000		+14.6	−24.8	− 1.1
4500		+14.4	−24.2	− 0.5
5000	−30.6	+16.2	−26.0	− 2.3
6000	−30.5	+16.9	−27.4	− 3.7
7000	−30.4	+10.4	−38.4	−14.7
8000	−30.3	+ 8.6	−54.3	−30.6
10000	−30.2	+ 6.8	−56.7	−33.0

TABLE 3

Sensitivity-frequency characteristic values for the artificial mouth of the Swedish Administration for a pressure of 1.5 dynes/cm² at 1000 c/s at a distance of 5 cm

(db relative to 1 volt)

c/s	V_1	V_2	V_3	Values of V_3 referred to 1000 c/s
100	−30.6	−14.9	−51.7	− 8.0
200		−15.2	−46.2	− 2.5
300		−13.9	−47.1	− 2.4
400		−13.2	−46.1	− 2.4
500		−13.3	−45.6	− 1.9
600		−14.3	−44.8	− 1.1
700		−16.0	−44.7	− 1.0
800		−18.9	−45.2	− 1.5
900		−20.7	−44.9	− 1.2
1000	−30.6	−21.5	−43.7	0
1200		−22.5	−45.7	− 2.0
1500		−17.7	−46.4	− 2.7
1800		− 9.5	−42.4	+ 1.3
2000		− 7.5	−45.7	− 2.0
2500		−11.0	−42.9	+ 0.8
3000		−16.7	−46.8	− 3.1
3500		− 6.1	−46.4	− 2.7
4000		− 5.4	−44.7	− 1.0
4500		− 5.6	−43.4	+ 0.3
5000	−30.6	− 3.8	−45.9	− 2.2
6000	−30.5	− 3.1	−44.6	− 0.9
7000	−30.4	− 9.6	−55.9	−12.2
8000	−30.3	−11.4	−73.6	−29.9
10000	−30.2	−13.2	−76.4	−32.7

TABLE 4

Variation of the sound pressure with distance along the axis of the artificial mouth of the Swedish Administration

(distance *d* cm) (reference point 5 cm) (5 dynes/cm² at 1000 c/s)

c/s	5 cm	2 cm	4 cm	10 cm	20 cm	33.5 cm	60 cm	100 cm
100	0	+7.6	+1.0	−7.2	−13.8	−15.2	−23.0	−27.3
200	—	+7.5	+2.1	−6.3	−12.6	−15.4	−23.6	−28.4
500	—	+7.3	+2.1	−6.2	−12.4	−15.2	−22.9	−27.2
1000	—	+7.1	+2.0	−5.8	−11.9	−15.6	−21.4	−25.1
2000	—	+6.8	+2.1	−5.8	−12.1	−15.4	−21.4	−26.1
3000	—	+6.4	+2.0	−4.7	−10.1	−12.8	−19.3	−23.8
4000	—	+5.9	+1.6	−5.6	−11.6	−14.4	−20.2	−25.3
5000	—	+5.7	+2.5	−4.9	− 9.6	−13.1	−18.7	−23.7

(Annex G)

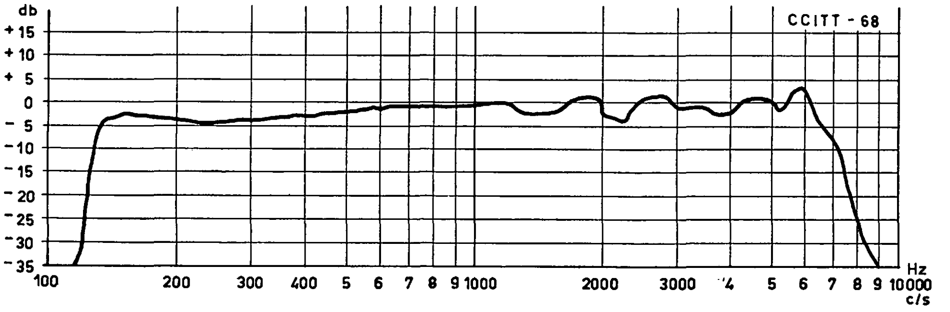


FIGURE 7a. — Sensitivity-frequency characteristics of the artificial mouth of the Swedish Administration (for a pressure of 5 dynes/cm² at 1000 c/s at a distance of 5 cm) measured with a Brüel and Kjaer recorder

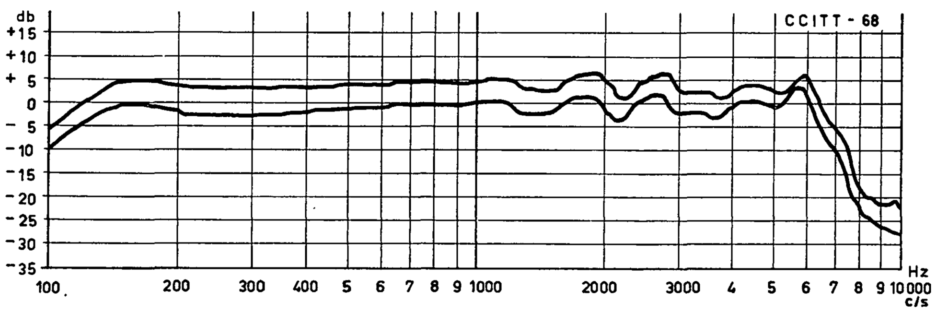


FIGURE 7b. — Sensitivity-frequency characteristics of the artificial mouth of the Swedish Administration (for a pressure of 5 dynes/cm² at a distance of 5 cm as point of reference), measured at a distance of 2 and 4 cm, respectively, from the artificial lips

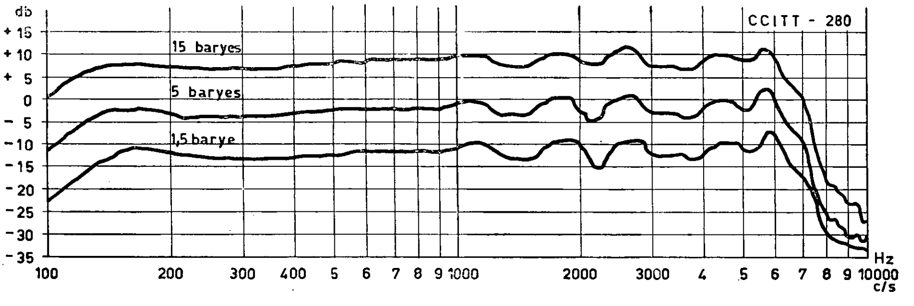


FIGURE 8. — **Sensitivity-frequency characteristics of the artificial mouth of the Swedish Administration (for pressures of 5 dynes/cm², 15 dynes/cm² and 1.5 dyne/cm² at 1000 c/s, at a distance of 5 cm) measured with a Brüel and Kjaer recorder**

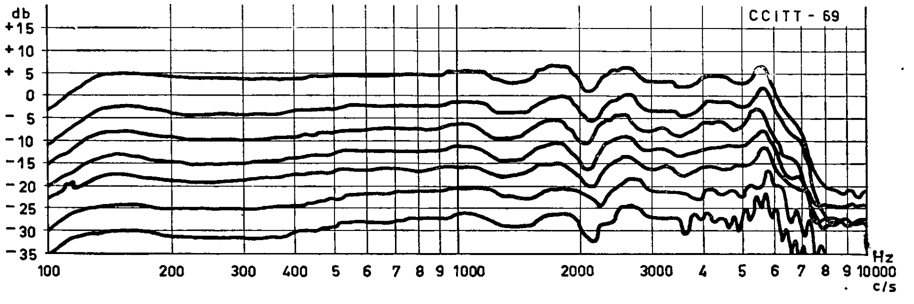


FIGURE 9. — **Sensitivity-frequency characteristics of the artificial mouth of the Swedish Administration (for a pressure of 5 dynes/cm² at 1000 c/s at a distance of 5 cm as point of reference), measured at distances: 2, 5, 10, 20, 33.5, 60 and 100 cm**

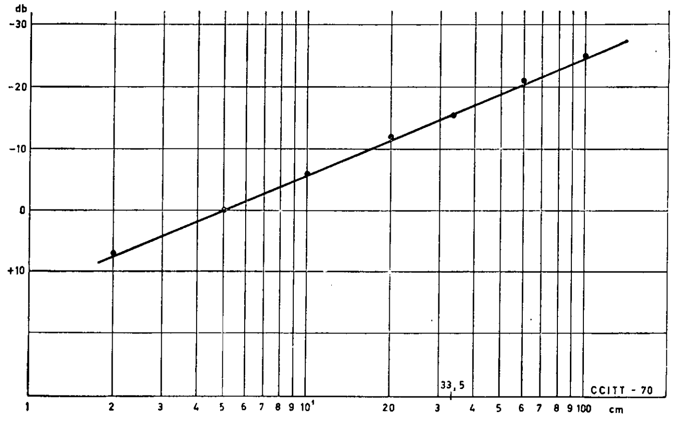


FIGURE 10. — **Variation of the sound pressure with distance along the axis of the artificial mouth of the Swedish Administration at 1000 c/s**
(Reference point at a distance of 5 cm)

TABLE 5

Characteristic values for the determination of the horizontal polar diagram of the artificial mouth of the Swedish Administration at a distance of 5 cm

(angle relative to the axis of the mouth = $\pm \alpha$)
(values in db, referred to the point defined by $\alpha = 0°$)

c/s	$\alpha = \pm 30°$			$\alpha = \pm 60°$			$\alpha = \pm 90°$		
	+	−	Mean	+	−	Mean	+	−	Mean
100	+0.2	−0.4	−0.1	+0.5	+0.1	+0.3	+0.6	+0.6	+0.6
200	0	+0.3	+0.1	+0.5	+0.7	+0.6	+0.8	+0.8	+0.8
500	−0.1	+0.2	0	+0.5	+0.5	+0.5	+0.5	+0.6	+0.5
1000	0	+0.3	+0.1	+0.4	+0.5	+0.4	+0.5	+0.5	+0.5
2000	−0.3	0	−0.1	−0.1	−0.1	−0.1	−0.3	−0.6	−0.4
3000	0	+0.3	+0.1	−0.4	−0.3	−0.3	−0.6	−0.9	−0.7
4000	−0.5	0	−0.2	−1.3	−0.8	−1.0	−2.5	−2.2	−2.3
5000	+0.1	−0.1	0	−0.7	−1.8	−1.2	−3.0	−5.1	−4.0

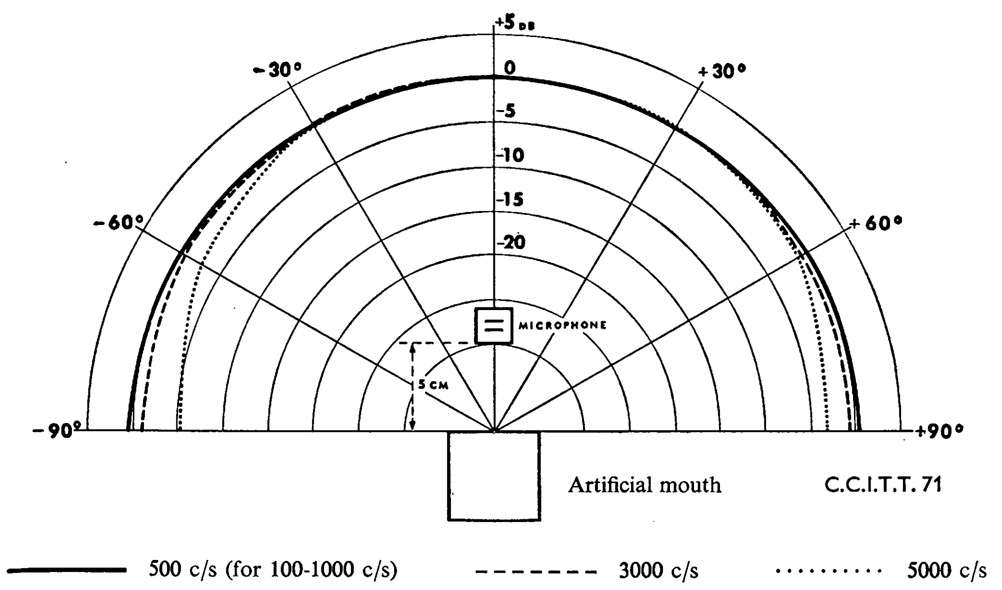


FIGURE 11. — **Horizontal polar diagram of the artificial mouth of the Swedish Administration**

TABLE 6

Characteristic values for the determination of the vertical polar diagram of the artificial mouth of the Swedish Administration at a distance of 5 cm

(angle relative to axis of the mouth = ± β)
(values in db, referred to the point defined by β = 0°)

c/s	β = ± 30°			β = ± 60°			β = ± 90°		
	+	−	Mean	+	−	Mean	+	−	Mean
100	+0.9	+0.8	+0.8	−0.2	+0.5	+0.1	−0.7	+0.2	−0.2
200	−0.1	+0.5	+0.2	−0.8	+0.3	+0.2	−1.1	−0.2	−0.6
500	−0.1	+0.3	+0.1	−0.8	+0.2	+0.3	−1.2	−0.3	−0.7
1000	+0.1	+0.4	+0.2	−0.5	+0.2	−0.1	−1.0	−0.5	−0.7
2000	−0.1	+0.3	+0.1	−0.8	−0.2	−0.5	−1.0	−1.3	−1.1
3000	−0.4	+0.6	+0.1	−1.6	0	−0.8	−2.2	−2.6	−2.4
4000	−1.4	+0.6	−0.4	−2.5	+0.5	−1.0	−2.5	−2.0	−2.2
5000	−0.2	+0.1	0	−1.1	−0.8	−1.0	−1.6	−3.1	−2.3

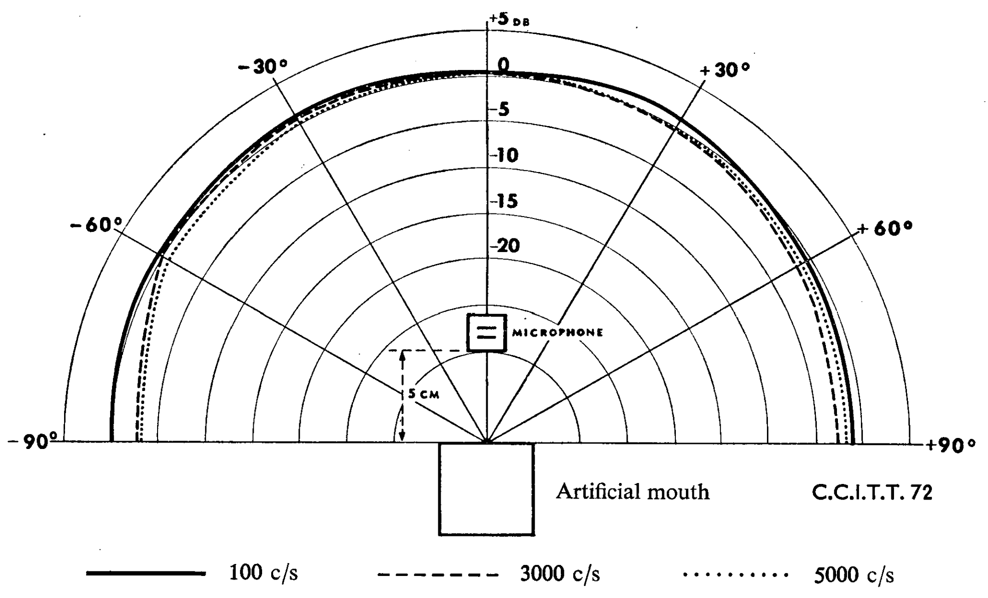


FIGURE 12. — **Vertical polar diagram of the artificial mouth of the Swedish Administration**

ANNEX H

(Geneva, 1964; quoted in Recommendation P.62)

PORTABLE TELEPHONOMETER FOR THE OBJECTIVE MEASUREMENT OF REFERENCE EQUIVALENTS USED BY THE CZECHOSLOVAK ADMINISTRATION

To measure the reference equivalent of subscribers' telephones the Communications Administration and the telephone industry of Czechoslovakia use two types of telephonometric equipment designed and made in that country. For more exact measurements, use is made of an equipment consisting of an audio-frequency generator, an artificial mouth, an artificial ear, a reference equivalent meter and an audio-frequency response tracer. This equipment is not essentially different from a similar equipment of the German and Swiss Administrations and described in Annexes 28 and 29, Volume V of the *Red Book*.

For simpler measurements of the sending and receiving reference equivalent of telephone apparatus, a portable telephonometer is used in Czechoslovakia. The portable telephonometer is built into a portable case with the dimensions $28 \times 21.5 \times 12.5$ cm ($11 \times 8.5 \times 5$ in.) and a weight of 6.5 kgs/14 lbs.

Simulation of the human voice is given by a noise signal made up of two rapidly alternating signals, one of which has a spectrum in the frequency band below 1 kc/s and the other in the frequency band above 1 kc/s. An instrument radiating this signal, compared with an instrument with a constant noise spectrum, has the advantage that it assesses the transmission performance of the electroacoustical transducers (fitted into the telephone apparatus) in two frequency bands, which contribute almost equally to the value of the resulting reference equivalent. Another advantage of this design is that the microphone insets do not change their sensitivity during the measurement.

When measuring microphones, the signal is amplified and reproduced by the artificial mouth and the electrical signal from the telephone apparatus passes through a transmission bridge to the input of the reference equivalent meter.

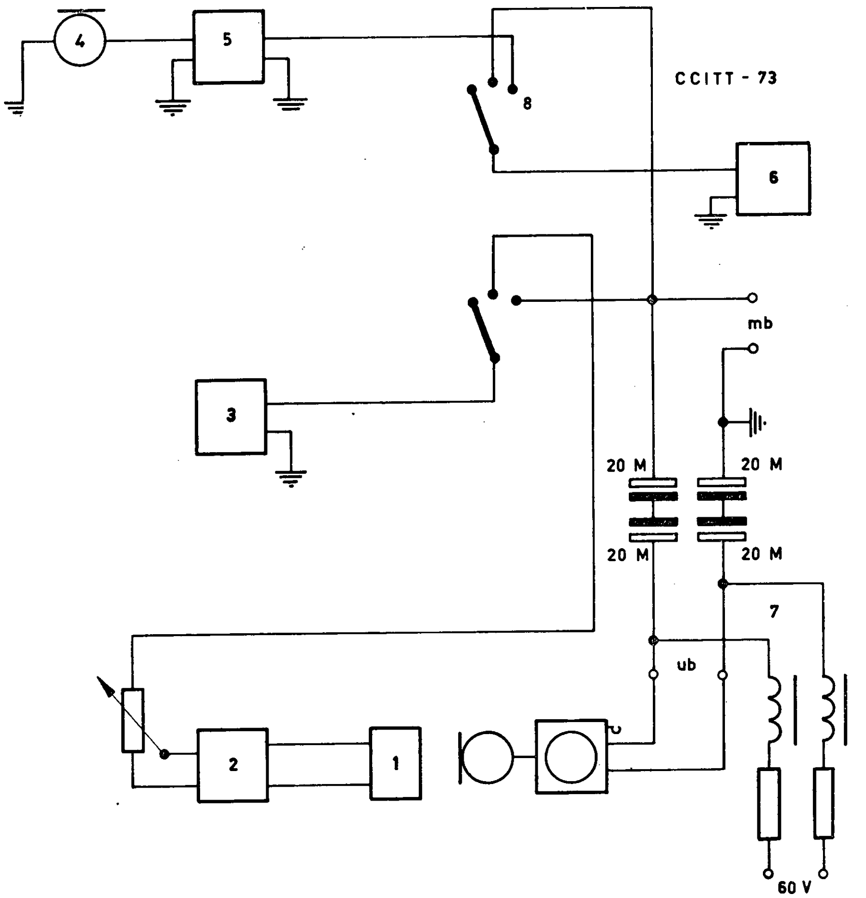
For measuring receivers, the electrical signal from the noise generator is applied to the telephone set being measured. The acoustical signal from the receiver under examination is picked up by the artificial ear and, after suitable amplification, is applied to the input of the reference equivalent meter.

On the front panel of the portable telephonometer there is the indicating meter (microammeter), the main switch and two other switches, by means of which all the functions of the instrument may be selected for measuring, calibration and battery test.

The interconnection of the individual parts of the telephonometer during measurement is shown in Figure 1.

The voice simulator is composed of a white noise generator, a frequency separator and an electronic switch. The output signal of the noise generator is divided by the frequency separator into two frequency bands with a dividing frequency of 1 kc/s. The electronic switch alternates these two bands about 15 times a second.

The artificial mouth has a cylindrical form of about 10 cm (4 in.) diameter. The diameter of the output orifice is 2.5 cm (1 in.). The sound generator itself is formed by a loudspeaker mounted into the artificial mouth. The resonance peaks of the artificial mouth-frequency response are attenuated by acoustic impedance and this response is further corrected in the amplifier of the



Pure resistances
of 150 ohms

1 = artificial ear
2 = artificial ear amplifier
3 = voice simulator
4 = artificial mouth
5 = artificial mouth amplifier
6 = reference equivalent meter

7 = feeding bridge
8 = main switch
mb = terminals for connecting a local battery telephone
ub = terminals for connecting a central battery telephone

FIGURE 1. — **Diagram of the transmission chain of the portable telephonometer (Czechoslovakia)**

artificial mouth, so that the maximum deviation in the frequency band 250-4000 c/s does not exceed ± 3.5 db. Before starting the measurements the artificial mouth has to be connected to the telephonometer. In the final position the orifice of the artificial mouth and the artificial ear have positions which conform to the dimensions of the standard European head.

The form of the artificial ear is in agreement with that recommended by C.C.I.F., XVIIth Plenary Assembly, Geneva (see Recommendation P.51, Volume V of the *Red Book*). In the inside there is an electromagnetic earphone inset serving as a microphone. The sensitivity frequency characteristic is modified by negative feedback in the artificial ear amplifier. This is permanently built into the telephonometer case. A fixing stirrup is used for applying the handset to the artificial ear.

The part of the equipment used for reference equivalent evaluation contains two amplifier stages and a copper oxide rectifier, a capacitor and a microammeter constituting the measuring circuit. The same ammeter is used as a low-frequency voltmeter for the adjustment of the correct voltage value (285 mV) at the output of the voice simulator generator and for testing the individual parts of the equipment.

The electronic part of the telephonometer is made by the printed-circuit technique and uses Czechoslovak-made transistors and diodes.

The power supply consists of a four-cell battery (over-all voltage 6 V) located inside the instrument. This internal supply is used for field measurements, for maintenance purposes, etc. For continuous use of the equipment in one place, an external 6 V power supply is recommended. The correct polarity of an external current supply is indicated by a signal lamp.

Technical characteristics

General	
dimensions	12.5 × 21.5 × 28 cm (5 × 8.5 × 11 in.)
weight	6.5 kg (14 lbs)
power supply	four cells of 1.5 V (complying with the Czechoslovak standard CSN 364171)
accuracy	±2.5 db
Voice simulator	
output voltage	285 mV
input impedance	600 ohms ± 10%
Artificial mouth with amplifier	
frequency response	250-4000 c/s ± 3.5 db
voltage gain	about 12 db
maximum electric power output	50 mW
Artificial ear with amplifier	
volume of artificial ear	about 4 cm^3
voltage gain of amplifier at 200 c/s	about 80 db
output impedance	600 ohms ± 10%
Reference equivalent meter	
input impedance	about 10 000 ohms or 600 ohms ± 10%
range	+15 to −6 db

PART III

QUESTIONS CONCERNING TELEPHONE TRANSMISSION QUALITY AND LOCAL NETWORKS ALLOCATED TO STUDY GROUP XII IN 1964-1968

Question 1/XII — Reference equivalents of national systems in the new transmission plan

(continuation of Question 1 of Study Group XII, 1961-1964)

a) What is the percentage of international calls for which it seems possible to satisfy the limits of 24 dN and 14 dN for the national reference equivalents?

Note. — A minimum of 95% is provisionally recommended in Recommendation P.11.

b) How should allowance be made for variation with time of the equivalents of national circuits (other than four-wire circuits interconnected to the international chain on a four-wire basis), and for the ageing of microphones and telephone receivers?

Note. — The following Annex shows the results of a study made by the Swedish Administration on the ageing of microphones and receivers.

c) What is the minimum value [1] to be recommended for the nominal reference equivalent of the national sending system?

d) What maxima and minima values should be advocated for the nominal reference equivalent of a connection comprising an operator's telephone set?

ANNEX

(to Question 1/XII)

The ageing of telephone microphones and receivers

(Contribution by the Swedish Administration)

The Swedish Administration has made an investigation of the transmission qualities of 946 used telephone sets. These sets have been in use for a longer or shorter time, and at least since

[1] The documentation collected by Study Group XVI concerning minimum reference equivalents will be included in the documentation of Study Group XII in 1964-1968.

1953, when the present model (BC 560) was introduced. The sets have been chosen from a total of 3000 which were replaced by sets with push-button dials in the town of Nynäshamn.

The samples were selected on the basis of the following principles:

a) the telephone set had to be of type BC 560;

b) the set, microphone, receiver or dial were not maintenance exchanged;

c) the age of the telephone sets could be determined.

In practice, however, it was difficult always to fulfil these requirements. First, about one-third of the telephone sets were equipped with handsets of an older type (the so-called Class II), containing microphones and receivers with other qualities than the present type (Class I). Secondly, condition b) was not fulfilled in some cases, which caused difficulties in estimating the age of the microphones, for which a marking of production date was not introduced until January 1961.

For all telephone sets the sending and receiving reference equivalents were measured in the Siemens O.B.D.M. apparatus. Also, the microphone resistance was assessed. For 100 telephone sets the frequency response curve of microphone and receiver was recorded, and for 30 carbon microphones the microphone resistance as a function of d.c. was measured. Certain measurements of the properties of dials have also been carried out but these are outside the scope of the present report.

The results are best displayed in Figures 1 to 9. Values of sending reference equivalent given in this report have been corrected by the difference between subjective and objective measuring values which has been found to exist for the handsets in use.

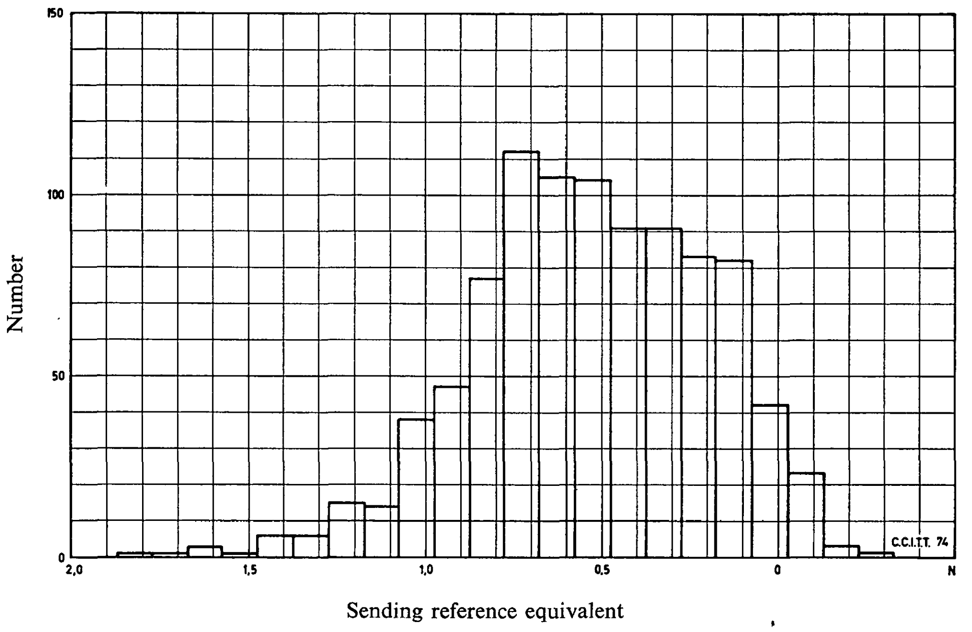


FIGURE 1. — **Distribution of sending reference equivalent for 946 used telephone sets (Classes I and II)**

Figure 1 shows the distribution of the sending reference equivalent of all measured telephone sets, while Figure 2 shows the distribution only of those sets containing the modern microphone type FL 5213 (Class I). The curve in Figure 3 is the integrated function of Figure 2. The variation with time is seen from Figure 4.

Figure 5 shows the distribution of the receiving reference equivalent of all telephone sets and Figure 6 only of those containing modern receivers (Class I). The variation with time is seen from Figure 7. Finally, Figure 8 shows the distribution of the d.c. resistance of the microphones FL 5213 and Figure 9 the corresponding variation with time.

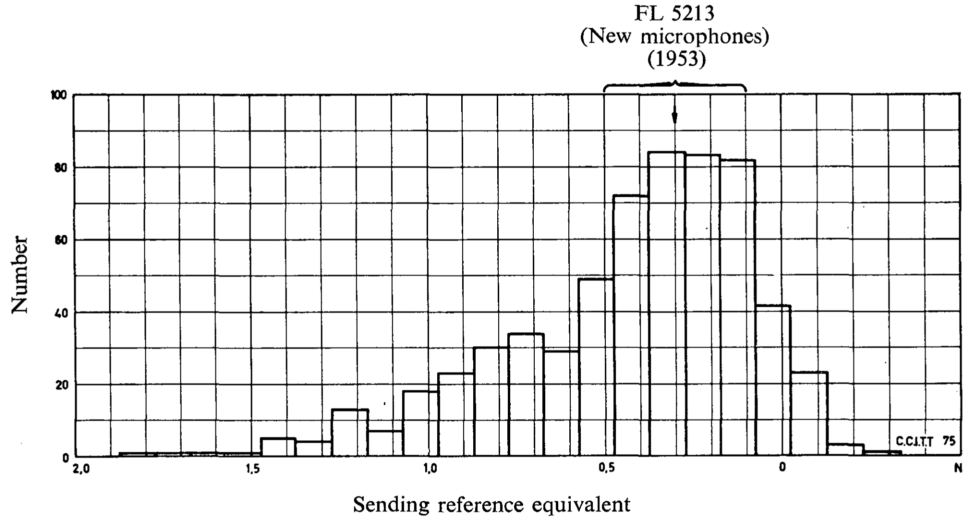


FIGURE 2. — **Distribution of sending reference equivalent for 606 used telephone sets (Class I; microphone type FL 5213)**

In order to facilitate a discussion of the results displayed in the diagrams, the table shown below has been compiled, giving the mean values, as measured at different times, of the reference equivalent and microphone resistance for new telephone sets (A) and for the used sets dealt with in this study (B).

	A			B
	1953		1963	Class I
	Class I	Class II	(Class I)	
Sending	+3 dN	+2 dN	+1 dN	+4 dN
Receiving	−5 dN	+1 dN	−6 dN	−6 dN
Microphone resistance	350 ohms	350 ohms	350 ohms	460 ohms

The tolerances are ±2 dN for Class I and ±4 dN for Class II. For column B, the standard deviation amounts to 3.5 dN for sending, somewhat less for receiving, and 145 ohms for the resistance.

The different nominal values for sending are given in Figure 3. It will be seen from this that 38 % of the microphones have reference equivalents worse than was permitted for new ones in 1953. 65 % lie outside the tolerance limit for 1963, a fact which is partly explained by a gradual improvement of this microphone type during the relevant 10-year period, as may be seen from the above table (the mean value has changed by 2 dN).

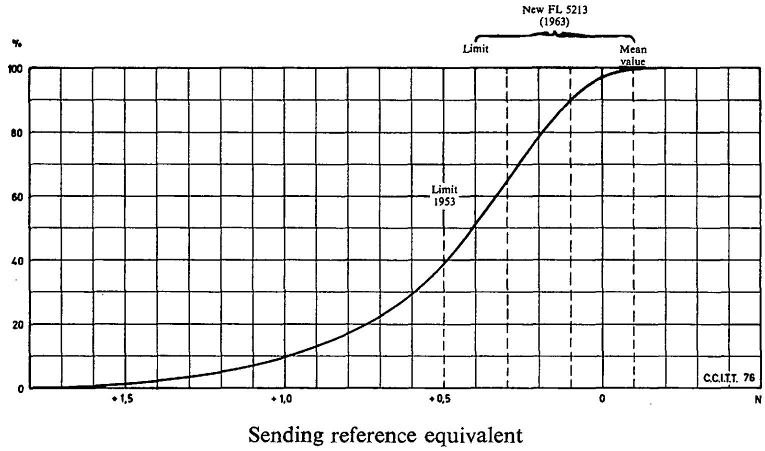


FIGURE 3. — **Cumulative distribution corresponding to the distribution of Figure 2**

Figure 4 shows how the sending reference equivalent depends on the age of the microphone. The scale indicates the manufacturing year. During the last 10-year period there is a trend of deterioration by 7 cN a year, reaching a plateau after about four years. The marked displacement of the curve around 1957 is probably explained by a change of the type of carbon granule used in 1956. The number of microphones measured which date from that period is too small to make the sudden change quite clear. Another thing that may be noticed is that the distance between the quartiles increases with time.

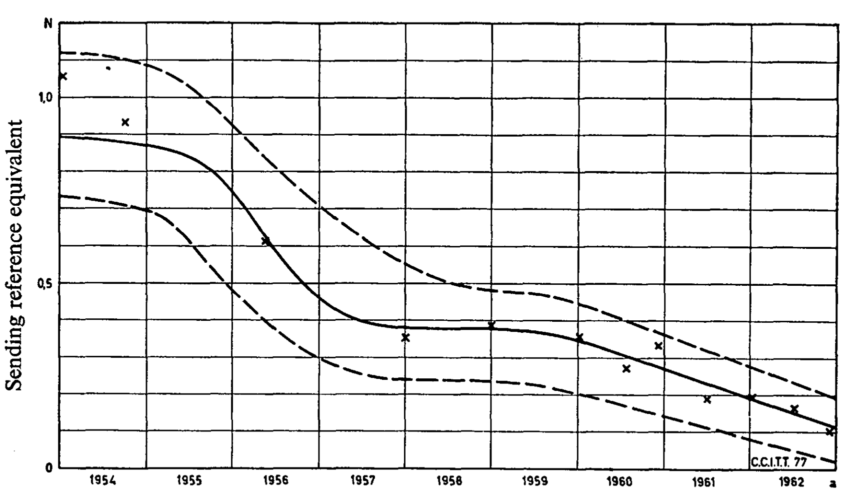


FIGURE 4. — **Dependence of sending reference equivalent of 606 used telephone sets (class I) on manufacturing year of microphone (type FL 5213)**

Dashed lines indicate quartiles (25% and 75% of cumulative distribution)

Note. — The carbon granule type was changed in 1956.

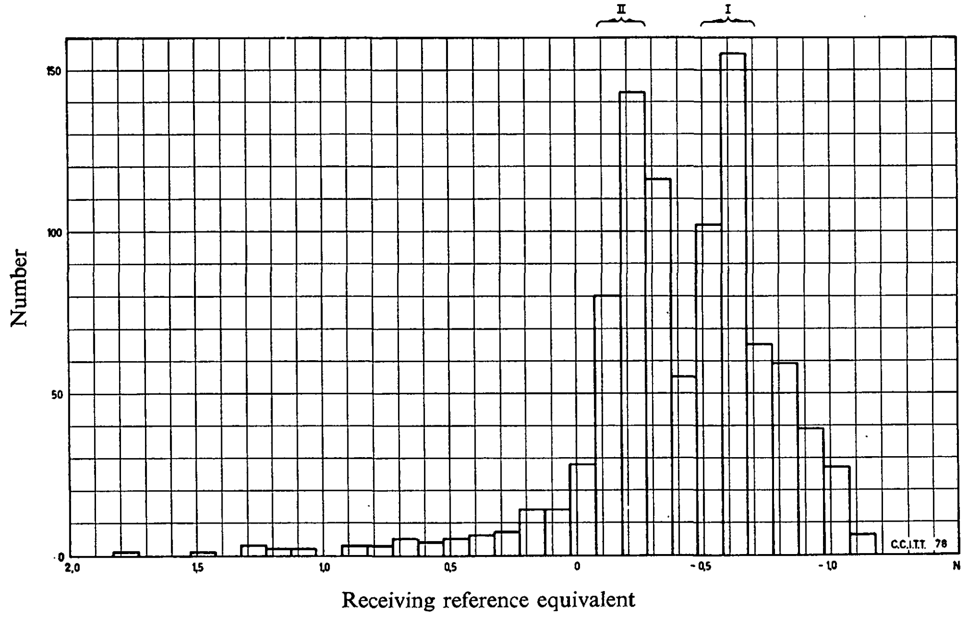


FIGURE 5. — **Distribution of receiving reference equivalent for 946 used telephone sets (Classes I and II)**

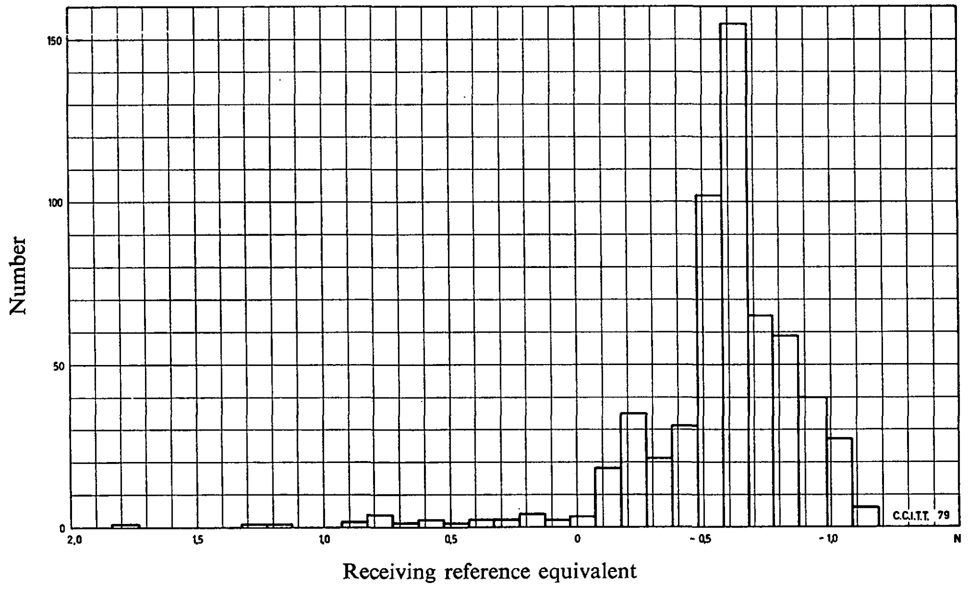


FIGURE 6. — **Distribution of receiving reference equivalent for 582 used telephone sets (Class I; several receiver makes)**

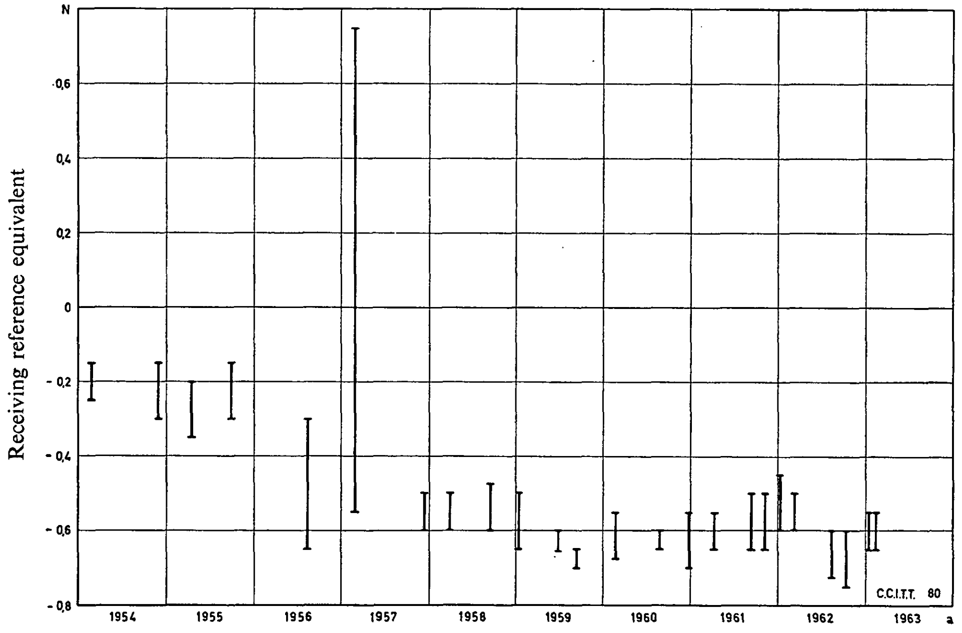


FIGURE 7. — **Dependence of receiving reference equivalent of 582 used telephone sets (Class I) on manufacturing year of receiver**
(Ends of bars indicate quartiles)

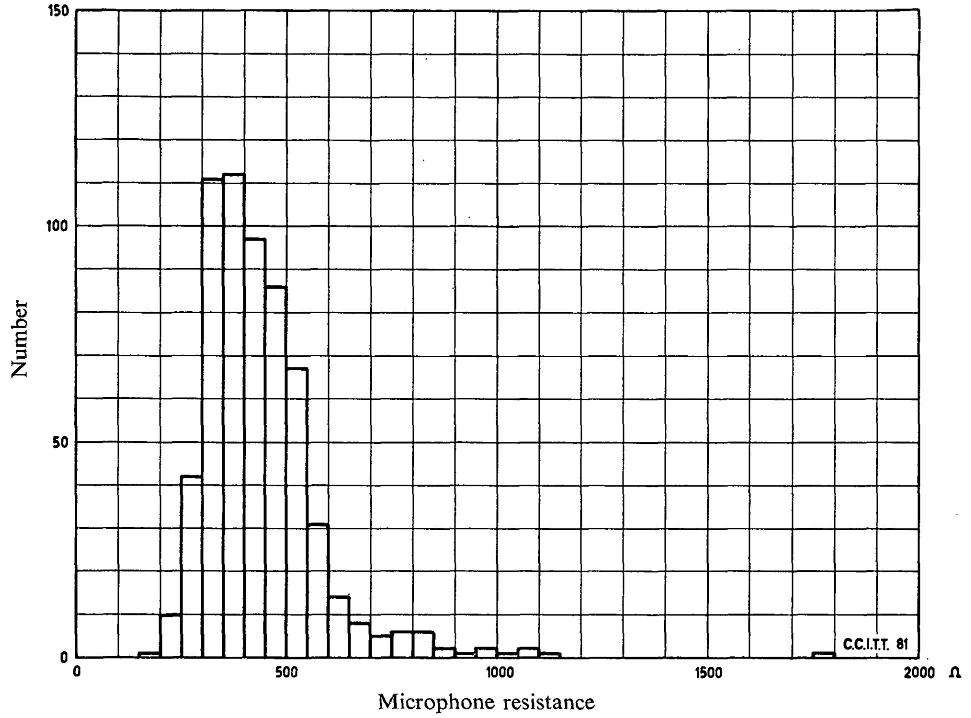


FIGURE 8. — **Distribution of the resistance of 606 used carbon microphones (type FL 5213)**

The microphone resistance (measured under normal speech conditions) exhibits a more continuous alteration with time, but, here too, the rate of change is greatest at the beginning (Figure 9). The fluctuations around the mean curve (which is drawn by eye) are not random but can be deduced from corresponding variations in the manufacture quality of microphones, as has been proved by means of acceptance test statistics. The dispersion increases with time similarly as was the case for the sensitivity.

The receiving reference equivalent does not show any appreciable deterioration with time. Variations during the last few years can be ascribed to changes in the production. Before 1957, a different type of receiver was used and in changing to a new type different constructions were tried, which explains the exceptionally high dispersion at that time.

It is seen from Figure 5 that the reference equivalent distribution for all receivers together is a superposition of the two distributions for Class I and Class II. 21% of the Class I receivers have a reference equivalent worse than the limit set in 1953. This is chiefly caused by faults in the earlier types of receivers.

Summarizing, it may be said that the investigation has shown that, if only the mean values are regarded, there is almost no change in the receiving properties with time, while there is a trend for gradual deterioration of the sending properties of the telephone sets. In both cases, a comparatively small number of transducers are found of which the properties have grown much worse. Therefore, the distributions become more asymmetrical with time, showing a tail towards higher reference equivalent values, without changing the mean value of the distribution very much. At the same time, the most serious faults are eliminated by maintenance repairs as soon as they are noticed by the subscriber. It should also be remembered that the continuous improvement of the transmission properties of telephone sets makes it very difficult to isolate possible ageing effects of this kind with great accuracy.

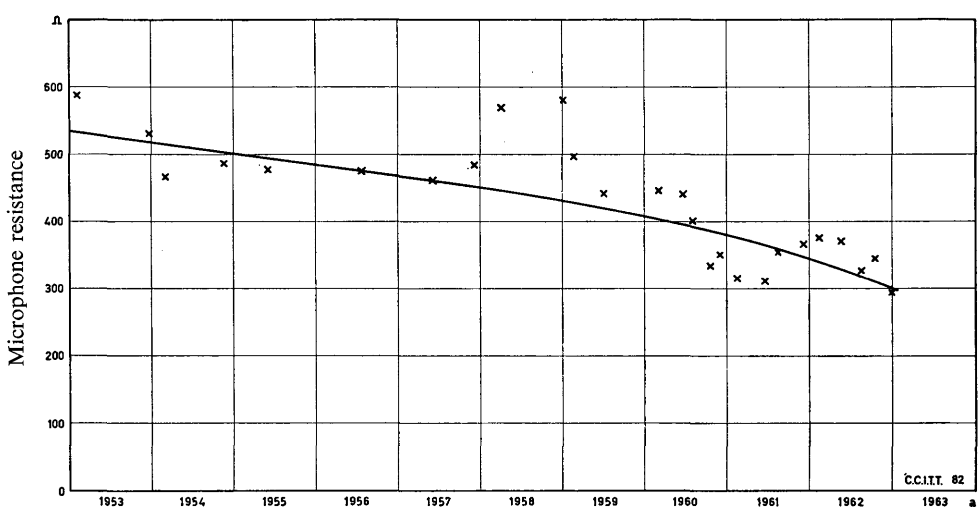


FIGURE 9. — **Dependence of microphone resistance on manufacturing year**

Question 2/XII — Measurement and limits of reference equivalent for sidetone

(continuation of Question 7 of Study Group XII, 1961-1964)

a) What is the method used by your Administration to study and measure the anti-sidetone properties of subscribers' telephone sets for speech and room noise?

b) What range of values, according to this method, do you consider as satisfactory for the sensitivity of the sidetone path of the subscriber's telephone set?

Note 1. — Although Administrations are free to choose their experimental method, the assessment of the effect produced by the anti-sidetone device in the subscriber's telephone set should in principle be based on opinions. The opinion tests should be carried out in a normal room with a room noise of 50 db (Hoth spectrum). It is possible to use several room-noise values and to show the variation of the results in terms of the room noise. The weighting curve of the noise meter used to measure the level of the room noise should also be shown.

Administrations are asked to include, as an annex to any contribution they submit, a detailed description of the instructions given to the subjects taking part in the opinion tests.

Note 2. — Administrations are recommended to express the results in terms of parameters characterizing the sidetone channel and the transmission channel; these parameters, expressed in decibels, are obtained by comparisons based on loudness and related to the acoustic link corresponding to a one-metre air path in a free field.

Note 3. — Annex 1, below, gives information supplied by Administrations on the methods that they use and on the range of values of sidetone sensitivity that they consider to be satisfactory for subscribers' telephone sets.

For information, the principle of the experimental method used by the Italian Administration is given in Annex 2.

ANNEX 1

(to Question 2/XII)

Some Administrations use special methods to characterize the anti-sidetone properties of subscribers' sets; they have provided the following information about the range they consider adequate for the sensitivity of the subscriber's telephone sidetone channel.

The American Telephone and Telegraph Company, in the light of various tests, feels that as far as possible the attenuation of the sidetone channel should lie between 8 and 15 decibels, measured by the E.A.R.S. system.

The Polish Administration measures the electrical attenuation between the microphone terminals and those of the receiver, the telephone line terminals being closed with a standard artificial line. It considers that 2 nepers is a minimum for the sidetone attenuation.

The Federal German Administration, in the light of opinion polls, finds that it would be well to increase the sidetone reference equivalent, but that there is no point in going beyond 18 db or so.

The United Kingdom Administration has undertaken various tests and has reached the conclusion that the efficacity of the sidetone channel should not exceed about +25 db, this efficacity being expressed with respect to one metre of air, and that in the majority of calls it should be at least a few decibels below this value.

ANNEX 2

(to Question 2/XII)

Tests conducted by the Italian Administration to assess side effect from opinion polls

(See the *Red Book*, Volume V, pages 540 to 542.)

Question 3/XII — Measurement of the disturbing effect of clicks

(continuation of Question 3 of Study Group XII, 1961-1964)

What methods and what apparatus should be recommended for the objective measurement of a quantity representative of the subjective disturbing effect (on the quality of telephone transmission) of a rapid series of clicks of small amplitude arising in a local telephone network: e.g. disturbances produced by the telex service on a telephone call, noise due to the normal operation of switching apparatus or disturbances due to the existence of faulty switching apparatus?

Note 1. — The study of this question should be undertaken by exploring, in particular, the following two possibilities:

a) how to avoid an overloading of the psophometer if it be used as a measuring instrument.

b) how to use the C.I.S.P.R. procedure in telephony (see, especially, Appendix 2 to Annex 3).

Note 2. — Administrations are invited to carry out tests based on the opinions of users. By way of example, a method is described in Annex 4.

The tests should be carried out:

1. with the exponentially decreasing type of pulse having a time constant of 0.1 ms and 1 ms;
2. for pulses which occur at the rate of 10, 100, 300, 500 and 700 times per minute;
3. for a reference equivalent of the complete connection of (20), 30, (40) db.

The following information should be supplied concerning the results of the tests:

a) the curves showing: rate of occurrence $\times$ intensity $\times \frac{1}{60}$, as a function of frequency, using the mean scores 3, 2, and 1.5 parameter;

b) the psophometric electromotive force of the click peaks, measured with the psophometer specified by the C.C.I.T.T., as a function of frequency for the mean scores 3, 2 and 1.5
(the attention of Administrations is drawn to the fact that the inertia of the psophometer needle has an influence on the correlation between its indications and the mean score for pulses which repeat less than 150 times a minute);

c) the indication given by the apparatus specified by the C.I.S.P.R. for the mean scores 1.5, 2 and 3 (details of this specification are given in Annex 5).

Note 3. — For information, Annexes 1 to 4 give results of tests carried out by various Administrations and based on the opinions of typical telephone users.

Note 4. — Annex 5 gives the specifications for the apparatus recommended by the Comité international spécial des Perturbations radioélectriques (C.I.S.P.R.) for measuring pulses or series of pulses.

Note 5. — Annex 6 indicates the importance of the overload factor of the meauring apparatus, when it is desired to measure clicks correctly.

Note 6. — Measurement of clicks for data transmission forms the subject of Question 7/C.

Note 7. — Annex 7 describes an apparatus for measuring both steady and impulsive noise and developed by the Australian Administration.

ANNEX 1

(to Question 3/XII)

Measurement of the disturbing effect of clicks

(Contribution by the Japanese Administration)

(See the *Red Book*, Volume V, pages 543 to 547.)

ANNEX 2

(to Question 3/XII)

Contribution by the Federal German Administration

(See the *Red Book*, Volume V, pages 547 to 555.)

ANNEX 3

(to Question 3/XII)

Measurement of the disturbing effect of impulsive noise

(Contribution by Siemens and Halske, A.G.)

(See the *Red Book*, Volume V, pages 555 to 563.)

ANNEX 4

(to Question 3/XII)

Contribution by the Czechoslovak Administration

(See the *Red Book*, Volume V, pages 563 to 571.)

ANNEX 5

(to Question 3/XII)

Extracts from the specifications for the C.I.S.P.R. measuring equipment for frequencies between 0.15 and 30 Mc/s

(See the *Red Book*, Volume V, pages 572 to 576.)

ANNEX 6

(to Question 3/XII)

New contribution by the Federal Republic of Germany

In measuring the disturbance caused by clicks, it should be borne in mind that the maximum instantanous voltage may be much greater than the peak voltage of a sine wave of equal r.m.s. value. It is possible, therefore, that the psophometer may be overloaded during such measurements, although the needle by no means moves to the end of the scale. The person using the psophometer cannot be aware of this overloading.

C.C.I.T.T. Study Group XII has therefore raised the problem: " How can overloading of the psophometer be avoided during measurements? ". The following comments have been made:

Experience shows that the overload reserve of 2.5 times laid down in the C.C.I.T.T. specification for a psophometer (*Red Book*, Volume V, page 129) is far from adequate. We have therefore introduced a new psophometer with an overload factor of about 25 (corresponding to 28 db, valid for measuring the r.m.s. value). The integration time of the psophometer is not changed.

ANNEX 7

(to Question 3/XII)

Measurement of the disturbing effects of clicks

(Contribution by the Australian Administration)

1. *General comments on noise, crosstalk and interference problems*

A significant amount of work has been carried out by the Australian Administration over the last ten years in respect of measurement of noise. More work has to be done to complete the investigations, but based on the results available, the following indicates thepresent views.

When considering the amount of noise, crosstalk and interference permissible on a circuit (whether it be a carrier-derived circuit, a physical circuit or a subscriber's circuit) it is necessary to consider not only the interference in the audio-band, but also interference at frequencies above and below this band, since interference at these frequencies may prevent the satisfactory operation of an over-all connection, consisting of a tandem connection of circuits, even when the individual circuits considered in isolation perform satisfactorily.

It has been found appropriate to consider noise (noise in this context including crosstalk, clicks and all other types of interference) in the three frequency bands:

a) Noise at very low frequencies, say from 30-150 c/s. Such noise may originate from the d.c. supplies to exchanges and other types of equipment, interference from distribution systems for electric power and traction and from numerous other sources. When circuits having excess low-frequency noise are connected up to carrier derived circuits, the low-frequency noise may prevent the carrier equipment from operating properly (notably interfere into pilot channels and out-of-band signalling channels, the frequency of these facilities normally being within 150 c/s of the carrier).

b) Noise at high frequencies, say in the range from 5 kc/s to 250 or 500 kc/s. Such noise may originate from a large variety of sources such as from medium-wave broadcasts, industrial apparatus of many types and, last but not least, from a variety of " crosstalk " paths within stations, line-crosstalk such as crosstalk from the HF-line of a (short-distance) frequency division or time-division multiplex system into normal physical lines of the local and junction network. When circuits having excess high-frequency noise are connected to carrier-derived circuits, the high-frequency noise may penetrate the filters at the input to the modulator and may appear (by a process of " leakage " and/or by a process of unwanted modulation) at the output of the modulator. Those products which, at the output of the modulator, exhibit frequencies inside the channel band will pass through the modulator band-pass filter, and will finally appear at the far end as audio noise.

c) Noise at audio-frequencies, which may interfere with any type of communication.

The noise in the three frequency bands may be in the form of a " steady " signal (such as sine waves or white noise) or in the form of clicks (impulses). The latter type is the type which usually causes erratic performance of data transmission circuits, voice-frequency telegraph circuits, out-of-band signalling channels and high-speed inter-register signalling.

The noise in the three bands may be measured at any desired time by a single instrument. This instrument should comprise a weighting network associated with a suitable indicating device, such as a meter and/or counter for impulse noise. Our proposals for the weighting network and the indicating device are given below.

2. *Weighting characteristics*

The permissible level of low-frequency and of high-frequency noise (as described in 1 a) and b) above) has been found in practice to be determined by the characteristics of the carrier equipment rather than by its effects on the user. This Administration has, based on a large number of tests on carrier equipment of sound modern design, devised a weighting curve of such characteristics that it is possible by a single measurement to ensure that the noise in each of the three bands is within satisfactory limits.

The weighting characteristics of the network coincides with the psophometric weighting in the 150-4700 c/s band. Below and above this band the weighting characteristic must be approximately complementary to the appropriate characteristics of the carrier equipment, leading to the following weighting:

30-150 c/s = 29.0 db (the C.C.I.T.T. psophometric weighting at 150 c/s is 29.0 db)

4.7-(250)-500 kc/s = 29.4 db (the C.C.I.T.T. psophometric weighting at 4.7 kc/s is 29.4 db).

A network exhibiting characteristics very closely approximating the idealized characteristic is shown in the appendix of this contribution, and it is proposed that the desired weighting characteristics be defined as the characteristics exhibited by this network.

This weighting characteristic is being used to specify noise and crosstalk requirements by the Australian Administration.

The proposed weighting curve would be used for assessing the noise performance of circuits which may be extended or tandem switched to carrier derived channels. The proposal to use this type of weighting does not imply that the present C.C.I.T.T. psophometer should not be used for the measurement of noise performance apparent to subscribers; indeed the weighting curve proposed for the new instruments has been chosen to agree as far as possible with the C.C.I.T.T. weighting curve to allow the maximum amount of correlation between readings on the two instruments.

3. *Indicating device*

Investigations directed towards the determination of the most suitable type of indicating (and/or pulse counting) device are as yet not completed. However, the following comments serve to indicate our present thoughts on the matter.

It is desirable that the indicating device should be capable of indicating the basic noise level (that is, the predominant noise level existing over most of the period of time over which the measurement is extended) and of indicating, for instance on a counter, the number of times the impulse noise exceeds a predetermined level.

The present concept of the indicating device is based on a device with properties similar to those of a peak programme voltmeter (such as for instance the United Kingdom peak programme meter described in C.C.I.T.T. *Red Book*, Volume III, Annex 49) with a charge time in the order of 2.5 ms. However, in order to enable impulse noise to be determined the device should be equipped with a second output arrangement designed to drive a counting device which would determine the number of times the impulse noise exceeds a preset level (this second output arrangement appears to be similar in concept to the Western Electric Type 6A Impulse Counter, as described in the *Bell Laboratories Record*, March 1963). In order to cater adequately for impulse noise interfering for instance with high-speed data transmission over telephone channels, the charge time constant of the output which drives the impulse counter should be in the order of 0.25 ms (in lieu of 2.5 ms used in the output which drives the indicating meter).

The discharge time-constant of the impulse counter output should be reasonably short, say in the order of 0.5 second, in order to provide good resolution between successive impulses. The discharge time constant of the indicating meter output should be reasonably long, say 2 seconds, in order that the indication given may be reasonably independent of the ballistic properties of the indicating meter provided a fast acting meter is used.

4. *Noise objectives*

The noise objectives when using the noise measuring instrument described would be those currently recommended as objectives when using the C.C.I.T.T. psophometer. The operate level of the impulse counter would be set at, for instance, 30 db above the steady noise limit and the maximum rate of impulse counting could be specified as, for instance, 30 counts per any half-hour.

5. *Conclusion*

The existing C.C.I.T.T. psophometer was not intended for measurements of impulse noise, whereas the proposed instrument is intended to be suitable for the measurement of impulse noise as well as " steady " noise.

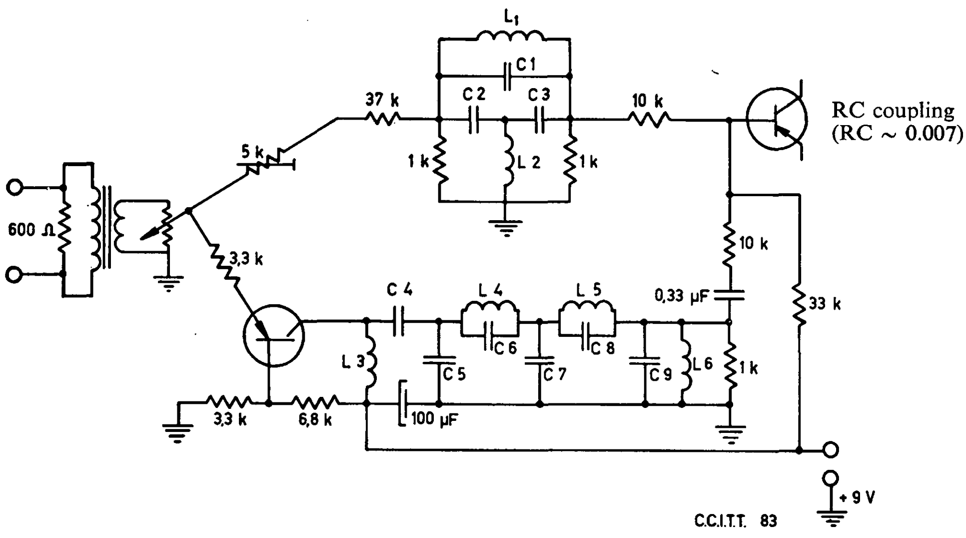
The weighting characteristic has been arranged to meet the needs of present-day telephone networks employing a large proportion of carrier-derived channels.

Being capable of measuring both impulse and steady noise, the instrument should be capable of measuring the types of noise affecting telephone users as well as the types affecting voice-frequency telegraphy, data transmission and signalling.

It may be of interest to note that the proposed instrument would facilitate the use of recording instruments for read-out, since the time constants are controlled in the electrical section of the instrument rather than in the mechanical meter movement as in the case of the psophometer. These recordings could be used for statistical analysis along the lines of C.C.I.T.T. Recommendation G.222.

APPENDIX (to Annex 7)

Circuit diagram of weighting network for noise measurements



A. *Components of reactance networks*

1. *Capacitors*

$C_1 =$ 9.80 nF m/s	$C_4 =$ 0.327 μF p.e.	$C_7 =$ 53 nF m/s
$C_2 =$ 28.5 nF m/s	$C_5 =$ 0.103 μF p.e.	$C_8 =$ 12.5 nF m/s
$C_3 =$ 28.5 nF m/s	$C_6 =$ 18.8 nF m/s	$C_9 =$ 15.8 nF m/s

m/s = mica and/or styroseal p.e. = polyester

2. *Inductors*

	Inductance	Core type ferroxcube	Air gap approx.	Turns/mH approx.	Number of turns approx.	B and S wire gauge
L_1	57 mH	K3000-62	0.23 mm	60	461	31 d.e.
L_2	24 mH	K3000-62	0.23 mm	60	304	29 d.e.
L_3	280 mH	K3001-00	0.58 mm	51	860	31 d.e.
L_4	46 mH	K3000-62	0.23 mm	60	412	30 d.e.
L_5	32 mH	K3000-62	0.23 mm	60	348	30 d.e.
L_6	770 mH	K3001-00	0.58 mm	51	1420	33 d.e.

d.e. = double enamel

Resonant frequencies

$L_1 // \left(C_1 + \dfrac{C_2\, C_3}{C_2 + C_3} \right) = 4.30$ kc/s

$L_2 // (C_2 + C_3) = 4.30$ kc/s

$L_4 // C_6 = 5.43$ kc/s

$L_5 // C_8 = 7.90$ kc/s

Tolerances

Capacitors and inductors $= \pm 1\%$
Resonant frequencies $= \pm 30$ c/s

B. *Transfer functions*

1. *Current transfer function of variable-loss network*

$$M(p) = \frac{\text{input current}}{\text{output current}} =$$

$$= \frac{(p^2 + 0.252\,p + 0.071)\,(p^2 + 0.98\,p + 0.99)\,(p^2 + 2.15\,p + 12)\,(p^2 + 2.64\,p + 18)}{0.142\,p^3\,(p^2 + 5.45^2)\,(p^2 + 7.9^2)}$$

where $p = jf/f_0$ and $f_0 = 1$ kc/s

2. *Transfer function of phase-shift network*

$$\Phi\,(p) = \frac{p^2 + 3.31\,p + 18.5}{p^2 - 3.31\,p + 18.5} \qquad p \text{ as above}$$

Resonant frequency

$f_r = 4.30$ kc/s

Note. — The variable resistor in the network branch containing the phaseshift network should be so adjusted that, at constant input voltage level of the complete noise measuring equipment, its reading at 5.5 kc/s is 29.2 db below that at 800 c/s.

C. Measured voltage transmission loss values of complete noise measuring network

(The network includes an RC coupling with $R \times C = 0.007$ ohm $\times$ farad as referred to on the schematic.)

f	A	A_1	f	A	A_1
c/s	db	db	kc/s	db	db
20	45.4	29.1	3.3	24.1	7.8
30	45.0	28.7	3.5	25.4	9.1
45	45.1	28.8	3.8	28.5	12.2
70	45.4	29.1	4.0	31.4	15.1
100	45.9	29.6	4.2	34.8	18.5
120	45.9	29.6	4.4	38.6	22.3
150	44.6	28.3	4.6	42.5	26.2
170	42.0	25.7	4.7	44.2	27.9
200	37.1	20.8	4.8	45.6	29.3
250	30.6	14.3	5.0	46.2	29.9
300	26.6	10.3	5.2	46.0	29.7
350	24.2	7.9	5.5	45.5	29.2
400	22.6	6.3	6.0	45.2	28.9
500	20.3	4.0	6.5	45.4	29.1
600	18.6	2.3	7.0	45.7	29.4
700	17.3	1.0	8	46.0	29.7
800	16.3	0	9	45.9	29.6
900	15.7	−0.6	10	46.1	29.8
1000	15.5	−0.8	12	46.0	29.7
1100	15.5	−0.8	14	46.1	29.8
1200	15.7	−0.6	16	46.2	29.9
1300	16.1	−0.2	18	46.2	29.9
1400	16.5	0.2	20	46.2	29.9
1600	17.3	1.0	25	46.3	30.0
1800	18.2	1.9	30	46.3	30.0
2000	18.9	2.6	40	46.4	30.1
2200	19.5	3.2	50	46.4	30.1
2500	20.3	4.0	60	46.5	30.2
2700	21.0	4.7	80	46.6	30.3
3000	22.3	6.0	100	46.6	30.3

A_1 = loss referred to 800 c/s

Question 4/XII — Effect of circuit noise on transmission performance

(continuation of Question 4 of Study Group XII, 1961-1964)

What is the family of curves which gives the noise transmission impairment as a function of the indications given by the psophometer standardized by the C.C.I.T.T. and for different values of the equivalent of the chain of national international circuits used?

Note 1. — The study should be pursued with particular attention to the following points:

a) to find a simple method whereby, given the results obtained for a particular type of telephone set, it would be possible to calculate the results that would be obtained in the case of a telephone set having a different sensitivity. In this connection, attention is drawn to the article by F. Markman, published in *Ericsson Technics*, No. 2, 1960. It is desirable that Administrations which may put forward such a method carry out some tests in order to verify its validity.

b) Administrations which have carried out opinion tests are asked to give the values of the speech volumes measured in these tests, if they did, in fact, measure them.

Note 2. — Annex 4 gives information on the correspondence between the readings of the latest American psophometer (in dbrn), those of the previous model (in dba) and those of the C.C.I.T.T. psophometer.

Note 3. — The method used by the United Kingdom Administration is described in the following article: RICHARDS, D. L.: Transmission Performance Assessment for Telephone Network Planning. *I.E.E. Proceedings*, pages 931-940, May 1964.

Comments by S.G. 12 and Annexes 1 to 3

(see *Red Book*, Volume V, pages 576 to 600).

ANNEX 4

(to Question 4/XII)

Contribution by the American Telephone and Telegraph Company

Information now used by the A.T. & T. Co. in assessing noise impairment is given in an article by D. A. Lewinski in the *Bell System Technical Journal*, March 1964 [1]. In this article, noise is expressed in terms of readings with C-message weighting on the 3A noise meter now used in the United States. Because the weighting differs from that associated with the older 2B noise meter and the C.C.I.T.T. 1951 psophometer, the relationship among measurements with these instruments is influenced by the spectrum of the noise measured. If one milliwatt of white noise in the band 300-3400 c/s is applied to each, the following readings are obtained:

3A noise meter (C-message weighting)	88 dbrn
2B noise meter (F1A weighting)	81.5 dba
C.C.I.T.T. psophometer (1951 weighting)	−2.5 dbm

Recognizing that the relationship will change for other noise spectra, the following rounded conversion factors are proposed for practical comparison purposes:

C.C.I.T.T. 1951 weighting		3A noise meter C-message weighting		2B noise meter F1A weighting
0 dbm	=	90 dbrn	=	84 dba
−90 dbm	=	0 dbrn	=	−6 dba
−84 dbm	=	6 dbrn	=	0 dba

These conversion factors include the effect of the difference between the reference frequencies used (800 c/s in the C.C.I.T.T. psophometer, 1000 c/s in the American noise meters).

Detailed information concerning the noise meters used in the United States is referred to in references [2] and [3] below.

REFERENCES

[1] Lewinski, D. A.: A New Objective for Message Circuit Noise. *Bell System Technical Journal*, 43, March 1964, page 719.

[2] Cochran, W. T. and Lewinski, D. A.: A New Measuring Set for Message Circuit Noise. *Bell System Technical Journal*, 39, July 1960, page 911.

[3] Aikens, A. J. and Lewinski, D. A.: Evaluation of Message Circuit Noise. *Bell System Technical Journal*, 39, July 1960, page 879.

Question 5/XII — Precision of subjectively determined reference equivalents

(new question)

Considering that the planning of national and international telephone connections is based on reference equivalents which can be determined only with a certain precision,

and bearing in mind that the precision obtainable from the relevant subjective measurements is limited by certain factors such as:

1. the stability of the testing crew;
2. the variability of the telephone sets, especially their carbon microphones;
3. the method of conducting the loudness balance comparisons,

a) What is the precision obtained with the present specified method?

b) Can the precision obtained with the present specified method be improved by suitable changes in the procedure?

Note 1. — Certain tests relating to this question are included in the programme of tests by the C.C.I.T.T. Laboratory.

Note 2. — Annex 6, Part II of Volume V of the *Red Book*, describes one statistical method for analysing loudness balancing results and for expressing the precision of the results.

Note 3. — Annex 7 of Volume V of the *Red Book*, describes one alternative method for conducting loudness balancing measurements.

Question 6/XII — Subscribers' tolerance of echo and propagation time

(continuation of Question 6 of Study Group XII, 1961-1964)

a) In the light of the results of opinion tests, what is the mean curve indicating subscribers' tolerance of echo when modern telephone sets are used, and what is the standard deviation of this tolerance's variation, for individual subscribers, in relation to this mean curve?

Note. — Owing to the important effect of the sidetone of telephone sets where echoes are concerned, this question should be studied in conjunction with Question 2/XII.

b) When using the echo tolerance curve, how should the echo balance return loss be calculated from the variation curve of the balance return loss as a function of the frequency?

c) What is the maximum propagation time for which echo suppressors complying with Recommendation G.151, Volume III of the *Red Book*, would be satisfactory? The effects of practically occurring ranges of relevant factors such as transmission losses, return losses, circuit noise levels and types of subscribers' sets and lines, should be considered.

d) What transmission performance can be expected from telephone connections having mean one-way propagation times of the order 150 ms and upwards? The following transmission factors are likely to affect the performance and appropriate ranges of their magnitudes should be considered:

1. The presence of several interconnected circuits, each having a separate pair of echo suppressors. The cases of similar types of suppressors and of different types in the several circuits require consideration;
2. The presence of appreciable end-delay;
3. Losses on the two-wire extensions at each end of the four-wire part of the connection;
4. Return losses of various magnitudes and loss-frequency characteristics at the two ends;
5. Circuit noise level, including the effects of very low levels achieved by companding;
6. Echo suppressors of different types at the two ends of the international circuit.

Note. — Annex 1 reports the situation regarding this question in 1964.

ANNEX 1

(to Question 6/XII)

Study Group XII's observations in 1964

When telephone connections of increasingly long propagation times are considered, certain problems arise that are peculiar to certain ranges of propagation time. Under practical conditions that apply in public telephone networks, speech signals will be reflected from the remote end and be returned to the customer while he is talking; these will appear to him as an augmentation of sidetone or as an echo. If the propagation time is very small the amount by which the reflected signals must be attenuated to be unobjectionable can readily be made sufficient; the maximum propagation time for which this is the case is the subject of Recommendation G.121, B (Volume III of the *Red Book*) and the present question part a) is concerned with the revision of this information. To use such information, it is necessary to express return losses which are frequency-dependent by a single figure suitably averaged over the transmitted frequency band; this problem is the subject of part b) of the present question.

When the propagation time exceeds the value discussed above, it is necessary to furnish the circuits with echo suppressors and the type described in Recommendation G.151 will be satisfactory up to a certain propagation time which it is not, at present, possible to define; part c) of the present question is concerned with the definition of this value of propagation time. Recent tests have shown that the performance of circuits equipped with improved types of echo suppressor and having mean one-way propagation times of the order of 150 ms is practically as good as that of circuits otherwise identical but having a very short propagation time; part d) of the present question is concerned with the further study of problems associated with the use of connections having propagation times of this order and upwards.

DOCUMENTATION AVAILABLE FOR THE STUDY OF QUESTION 6/XII

Part a. *Subscribers' tolerance of echo for delays which do not render the use of echo suppressors necessary*

Study Group XII is not yet in a position to recommend that a new curve be used to show the minimum permissible attenuation of echo as a function of delay. Hence Study Group XVI may, for the time being, go on using the old American Telephone and Telegraph Company curve on which Recommendation G.121, Volume III of the *Red Book*, was based.

Study Group XII has received a number of contributions which are reproduced in Annex 2, below.

Part b. — *Calculation of balance return loss as regards echo*

Study Group XII has a contribution by N. V. Philips' Telecommunicatie Industrie which describes a method of calculating return loss as regards echo. This contribution is reproduced in Annex 3 below.

Parts c and d. — *Subcribers' tolerance of high propagation times, such as to necessitate the use of echo suppressors, and of residual echo*

Recommendation P.14 contains provisional limits for the propagation time in an international communication.

Information regarding the study of these parts of the question will be found in Annexes E and F (Part II of this volume) and in Annex 4 below.

ANNEX 2

(to Question 6/XII)

Subscribers' tolerance of echo for delays which do not render the use of echo suppressors necessary

(information supplied by various Administrations)

I. CONTRIBUTION BY THE AMERICAN TELEPHONE AND TELEGRAPH COMPANY

Two sets of data on the tolerance of observers to echo have been obtained from tests spaced by a period of years [1], [2]. The important results are summarized by means of the following table.

The new data show a greater sensitivity to echo than the old data, which continue to be used as the basis for the assignment of circuit losses in the North-American network. There is no real urgency to change the original design rules because they include more than adequate margin against low terminal return losses and loop and trunk losses at the talking end.

The whole approach to the assignment of circuit losses is under study in the light of new information on terminal return losses and loop and trunk losses as they actually occur in practice as well as new data on echo tolerance. An important consideration will be the possible use of equalizer networks to improve terminal return losses. However, when all factors are considered, it is not expected that there will be a large change in trunk loss objectives.

Round-trip delay (milliseconds)	Minimum permissible loss in echo path			
	Old		New	
	Average observer	Standard deviation	Average observer	Standard deviation
	db	db	db	db
0	1.4	2.5	−2.0	5.0
20	11.1		13.2	
40	17.7		21.3	
60	22.7		26.8	
80	27.2		31.2	
100	30.9		34.8	

REFERENCES

[1] CLARK, A. B. & MATHES, R. C.: Echo Suppressors for Long Terminal Circuits. *Bell Telephone Laboratories*, Monograph 165, presented at A.I.E.E., June 1925.

[2] PHILLIPS, G. M.: Echo and its Effects on the Telephone User. *Bell Laboratories Record*, 32, August 1954, pages 281-284.

II. CONTRIBUTION BY THE JAPANESE ADMINISTRATION

1. *Introduction*

This report concerns studies of the relationship between the delay-time and the necessary echo-path attenuation for Japanese-language communications. The results show practical echo-path attenuation on the basis of the speaker's echo. They also show the echo-path attenuation for which the echo is hardly perceptible.

2. *Tests made*

Figure 1 gives a block-schematic diagram of the test circuit. A magnetic tape-recorder (MTR) was used as a delay-device, and a variable attenuator (ATT) was used to adjust the echo-path attenuation. The telephone set used is a No. 4A automatic telephone set, a standard model in Japan. The subjects were 14 men with normal hearing. The place in which the test was made was an office room.

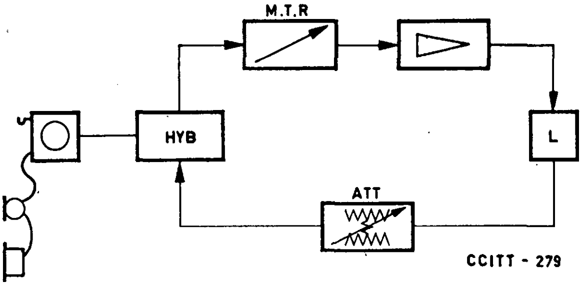
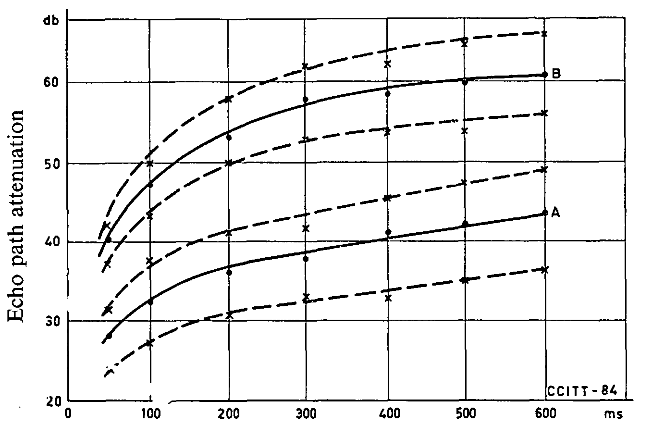


FIGURE 1. — **Block-schematic diagram of the test circuit**

Several values of delay-time between 50 ms and 600 ms were selected. For each delay-time, a subject was required to adjust the attenuator according to his own auditory sensation, thereby making it possible to find the echo-path attenuations for which the echos are imperceptible or perceptible but not objectionable to him. Incidentally, the subjects repeated the same experiment with four telephone sets to minimize variations caused by the sets. The telephone transmitter varies in characteristics when it is moved. An appropriate current was therefore set flowing through the transmitter to eliminate this effect.

3. *Test results*

The results are shown in Figure 2.



Curve A: Perceptible but not objectionable Curve B: Hardly perceptible

FIGURE 2. — **Echo-path attenuation for delay time (Japan)**

Curve A shows delay time and average echo-path attenuation for which an echo just becomes intolerable to the subjects. The dotted lines above and below the curve denote the standard deviation of variations depending on the subjects.

Curve B shows, in a similar manner, the average echo-path attenuation for which an echo is just perceptible for each delay-time.

The echo-path attenuations increase rapidly up to 100 ms, but thereafter increase but slowly and the variations in the measured values are spread with an almost normal distribution, the standard deviation of which is approximately 5 db.

By using the observed average echo-path attenuation and the standard deviation shown by curve A, the necessary echo-path attenuation can be obtained for any desired grade of service from the point of view of echo. Subtraction of the two-wire reflection loss at the distant end from this value makes it possible to find the necessary attenuation to be given by the echo suppressor.

4. *Conclusion*

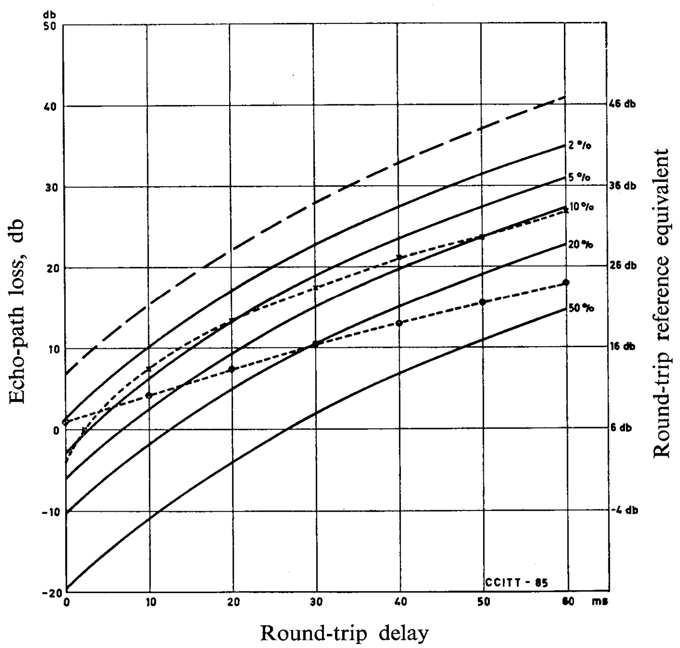
When Japanese-language communications are carried out under the aforementioned operating conditions, echo-path attenuation of about 37 db is required, to fulfil the condition that 86% of the speakers will perceive the echo but do not consider it to be objectionable, for a round-trip delay of about 100 ms in the earth satellite communication of low altitude for example. With stationary satellite communication in which the round-trip delay is about 600 ms, the value is about 48 db.

III. Contribution by the United Kingdom Administration

The results of tests carried out by the British Post Office have been published (reference [3]) and Figure 3 gives a summary of the more important conclusions.

The telephone sets used were of the BPO 300-type extensively in service at the time of the tests (1956-58). The echo-path attenuation was expressed in terms of the round-trip path loss beyond the talker's local exchange and this loss is given in the left-hand scale of Figure 3; the right-hand scale gives the round-trip reference equivalent from the talker to his local exchange, round the echo path and back to him from his local exchange. Clearly it is the right-hand scale that, strictly, should be used in comparing these results with those obtained with other types of telephone set, having different send and receive reference equivalents. Nevertheless the two curves reproduced from references [1] and [2] have been superposed on the BPO results for comparison using the left-hand scale because the appropriate reference equivalents were not known.

It will be seen in Figure 3 that the median value of loss at which untrained subjects will object to echo is 26 db less than the median value at which an experienced testing crew can just detect the presence of echo. The dispersion of tolerance to objection by the untrained subjects corresponded to a standard deviation of 10 db and this enabled the other percentile values shown to be expressed;



x Results published by Bell Telephone Laboratories • Present C.C.I.T.T. Recommendation

— — — — Median threshold of detection for experimental crew

———— Median and other percentiles for objection to echo by untrained subjects (standard deviation = 10 db)

Figure 3. — **Relationships between echo-path loss and round-trip delay for various percentages of subjects who would find the echo objectionable (United Kingdom)**

the dispersion of the threshold of detectability by the trained crew was much less and corresponded to only a very few db.

Some information is given in reference [3] on the effect of varying the sidetone level of the telephone set used by the talker. In that publication some of the information is expressed in terms of "air-to-air sensitivity, db relàtive to a one-metre airpath"; for comparison with other results it may be assumed that 0 db on that basis is equivalent to an over-all reference equivalent of about 30-33 db.

REFERENCES

[1] PHILIPS, G. M.: Echo and its Effects on the Telephone User. *Bell Laboratories Record*, 1954, *32*, page 281.

[2] *C.C.I.F. Green Book*, 1956, III, Annexes, pages 8-9.

[3] RICHARDS, D. L. and BUCK, G. A.: Telephone Echo Tests. *Proc. Inst. Elec. Engrs.*, 1960, *107*, Part B, page 553.

ANNEX 3

(to Question 6/XII)

Weighting for balance return loss as regards echo

(Contribution by N. V. Philips)

1. *General*

The balance return loss a_R of the two-wire line impedance Z and the balancing network W, at the terminating set, generally is a function of frequency:

$$a_R\,(f) = 20 \log \left| \frac{W + Z}{W - Z} \right| \text{db} \tag{1}$$

The criterion for stability is the minimum value of the sum of balance return loss and circuit loss in a four-wire loop.

As a measure of the annoyance due to echo this minimum value is not representative. The distribution of the energy of telephone speech over the frequency band must be taken into account, as well as a weighting for the aural effect of disturbing components of various frequencies. It would be profitable to have both effects incorporated into one weighting function, by means of which a weighted return loss a_{RW} can be determined from a given $a_R\,(f)$.

An application of such a weighted return loss can be found in Recommendation G.121. The curve on page 15, *Red Book*, Volume III, gives the minimum permissible value for transmission equivalent for connection without echo suppressors as a function of the propagation time.

From Annex 1 in Volume III, p. 325, it appears that the assumed echo-balance return loss (11 db) is expressed as a *weighted mean power ratio over the band 500-2500 c/s.*

What is meant by this weighting can be found in Contribution COM XVI—No. 18 of 1961-1964, reading as follows:

"In the reply given in 1960 to the question which preceded Question 1/XVI [1], the possibility was envisaged of defining the echo-balance return loss later on by means of a psophometer and a generator of clearly defined power spectrum. If any Administrations wish to carry out measurements applying such a definition, it would be desirable for them to use the

[1] Reproduced in the *Red Book*, Volume III, page 501.

conventional telephone signal already recommended by the C.C.I.T.T. for other applications, as defined in the Note to Recommendation G.222, paragraph 5.2 (*Red Book*, Volume III, page 52 and Figure 14, page 51)."

Note: These weightings approximate to those employed by the Telephone Association of Canada in Contribution COM XII N° 30 of 1961-1964.

2. *Question*

The question arises whether or not Study Group XII can agree with this definition of weighted echo-balance return loss.

This could perhaps be discussed in connection with Questions 19/XII and 6/XII.

a) As echo-effects occur only on long connections, whose frequency band is limited to 300-3400 c/s, it is logical to consider the echo-balance return loss over this band only.

A possible reduction to the band 500-2500 c/s will be discussed in paragraph 5.

b) The spectral energy distribution as a function of frequency not only depends on human speech, but also on the frequency response of the subscriber set and circuits.

Taking into consideration the widespread tolerances of sets and circuits it is necessary for practical applications to agree about a conventional distribution.

The conventional spectral energy curve of telephone speech, according to *Red Book*, Volume III, page 51, seems to be useful indeed for this application.

c) It is a question whether the weighting network of the psophometer, intended as a weighting for the aural effect of disturbing components of various frequencies, also weights the components of an echo correctly.

d) The aural loudness impression at normal speech level is related to the sound pressure p as $p^{2/3}$ or $p^{1/2}$; but for lower levels, as echos may be considered, this relation changes to p or p^2. As sound pressure is proportional to voltage V, the latter (p^2) means a loudness impression proportional to V^2 or power, in accordance with the evaluation by the psophometer.

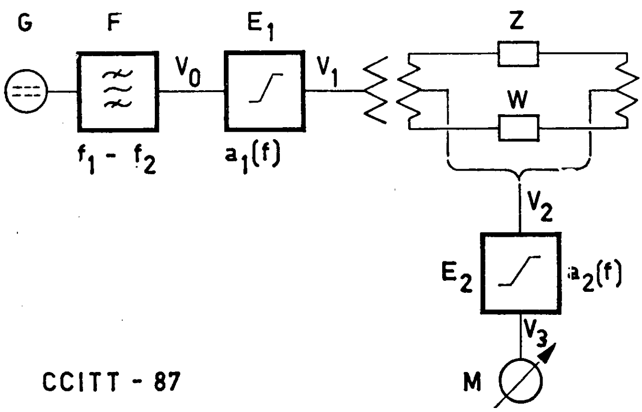


FIGURE 1

3. *Weighting function*

Pending the opinion of Study Group XII about the question of the preceding paragraph 2, we derived a weighting function a_W (f) that permits the calculation, in a simple way, of the weighted echo-balance return loss a_{WR} in cases where the return loss a_R (f) as a function of frequency is known.

The derivation, based on the definition from the *Red Book*, Volume III, page 501, and an example of application are given below.

3.1 *Definition*

From the measuring arrangement, envisaged in Contribution COM XVI No. 18, 1961-1964, a weighting function can be deduced.

The impedances Z and W are in a bridge (see Fig. 1) for which:

$$20 \log \left| \frac{V_1}{V_2} \right| = a_R \text{ db}$$

This bridge is supplied with noise from the generator G, the bandpass filter $F\,(f_1 - f_2$ c/s) and the weighting network E_1 with an attenuation $a_1\,(f)$.

The weighting network E_2 with attenuation $a_2\,(f)$ and the voltmeter M form together the psophometer.

We suppose the noise voltage and the sensitivity of the psophometer during a calibration with $Z = \infty$ or $Z = 0$ ($a_R = 0$) to be so adjusted that the meter M reads an arbitrary value α_0 db. In the measuring position, with Z in the bridge, the meter reading will be reduced to α_1 db.

The weighted echo-balance return loss, expressed as a weighted mean power ratio over the band $f_1 - f_2$, now simply equals:

$$a_{RW} = \alpha_0 - \alpha_1 \text{ db} \tag{2}$$

Both reading α_0 and α_1, and thus a_{RW}, can be calculated from Figure 1.

The noise power after F amounts to

$$P_0 = \frac{V_0^2}{R}$$

The frequency band $f_1 - f_2$ is divided in n equal small bands Δf

$$\Delta f = \frac{f_2 - f_1}{n}$$

In such a small band $(\Delta f)_i$ the meter voltage V_3 amounts to

$$V_{3i} = V_0 \times 10^{-0.05\,(a_1 + a_2 + a_R)_i}$$

The meter reading α_1 is proportional to the logarithm of the mean power, i.e.

$$\alpha_1 = 10 \log \frac{1}{f_2 - f_1} \sum_{f_1}^{f_2} \frac{V_{3i}^2}{R}\,\Delta f = 10 \log \frac{P_0}{n} \sum_{i=1}^{n} 10^{-0.1\,(a_1 + a_2 + a_R)_i}$$

And for $a_R = 0$ we have:

$$\alpha_0 = 10 \log \frac{P_0}{n} \sum_{i=1}^{n} 10^{-0.1\,(a_1 + a_2)_i}$$

$$a_{RW} = \alpha_0 - \alpha_1 = -10 \log \frac{\sum_{i=1}^{n} 10^{-0.1\,(a_1 + a_2 + a_R)_i}}{\sum_{i=1}^{n} 10^{-0.1\,(a_1 + a_2)_i}} \tag{3}$$

We now introduce the weighting function $a_W\,(f)$, which is entirely a function of a_1 and a_2, only in such a way that:

$$\sum_{i=1}^{n} 10^{-0.1\,a_{Wi}} = 1$$

as given by (4) and (5)

$$10^{-0.1\, a_{Wi}} = \frac{10^{-0.1\,(a_1+a_2)_i}}{\sum_{i=1}^{n} 10^{-0.1\,(a_1+a_2)_i}} \tag{4}$$

$$a_{Wi} = a_{1i} + a_{2i} + 10 \log \sum_{i=1}^{n} 10^{-0.1\,(a_1+a_2)_i} \text{ db} \tag{5}$$

The expression (3) can now be simplified by substituting (4) in it.

$$a_{RW} = -10 \log \sum_{i=1}^{n} 10^{-0.1\,(a_{Ri}+a_{Wi})} \text{ db} \tag{6}$$

3.2 *Calculation of* a_{Wi}

The weighting $a_1\,(f)$ according to the *Red Book*, Volume III, page 51, is defined by:

$$100 < f < 500 \qquad a_1 = 14.3 \log \frac{500}{f}$$

$$500 < f < 1000 \qquad a_1 = 0$$

$$1000 < f < 3000 \qquad a_1 = 33.5 \log \frac{f}{1000}$$

$$3000 < f < 5000 \qquad a_1 = 64 \log \frac{f}{1700}$$

The psophometer weighting $a_2(f)$ is given in the *Red Book*, Volume III, pages 54-55.
With this data, the weighting function a_{Wi} is calculated in Table 1. The 500-2500-c/s band has been divided into $n = 10$ bands of $\Delta f = 200$ c/s wide.

The 10 mid-frequencies f_i of each of these bands are then 600 ... 2400 c/s.

From the fourth column it is seen that $\sum_{i=1}^{10} 10^{-0.1\,(a_1+a_2)_i} = 4.00$

The weighting function a_W according to (5) thus becomes simply:

$$a_W = a_1 + a_2 + 6 \text{ db} \tag{7}$$

and is given in the fifth column.

TABLE 1

f	a_1	a_2	$10^{-0.1\,(a_1+a_2)}$	a_W
600	0	2.0	0.63	8.0
800	0	0	1.00	6.0
1000	0	−1.0	1.26	5.0
1200	2.7	0	0.54	8.7
1400	5.0	0.9	0.25	11.9
1600	6.9	1.7	0.14	14.6
1800	8.6	2.4	0.08	17.0
2000	10.2	3.0	0.05	19.2
2200	11.6	3.5	0.03	21.1
2400	12.8	4.0	0.02	22.8
			4.00	

4. *Application*

As an example of the method discussed above, the weighted echo-balance return loss a_{RW} has been calculated for a case in which the return loss a_R as a function of frequency was known (a_R according to Annex 4 to Question 19/XII, case A_1).

f	a_W	a_R	$10^{-0.1\,(a_R+a_W)}$
600	8.0	16.5	$3.54 \cdot 10^{-3}$
800	6.0	14.8	8.35 »
1000	5.0	13.1	15.50 »
1200	8.7	11.7	9.12 »
1400	11.9	10.3	6.03 »
1600	14.6	9.1	4.28 »
1800	17.0	8.3	2.95 »
2000	19.2	7.5	2.14 »
2200	21.1	7.0	1.55 »
2400	22.8	6.5	1.17 »
			$54.63 \cdot 10^{-3}$

$$\sum_{i=1}^{10} 10^{-0.1\,(a_R + a_W)_i} = 54.63 \times 10^{-3}$$

$$a_{RW} = -10 \log 54.63 \times 10^{-3} = \underline{12.6 \text{ db}}$$

5. *Choice of the band limits*

Table 1 is based on the band 500-2500 c/s. To prove the rightness of these limits, another weighting function $a'_W\,(f)$ was calculated for the band 300-3300 c/s.

It was then found that the frequencies in the bands 300-500 c/s and 2500-3300 c/s count for 4% only. Thus one can expect that the return loss in these bands will generally have little influence on the weighted echo-balance return loss a_{RW}.

To prove this, the weighted echo-balance return loss has again been calculated for the same $a_R\,(f)$ as mentioned in the application, paragraph 4, but now as a weighted mean power ratio over the band 300-3300 c/s; the difference with the band 500-2500 c/s amounts to only 0.04 db (12.62 — 12.58 db).

For extremely unfavourable cases, in this respect, where the return loss drops 8 db from 500 to 300 c/s and from 2500 to 3300 c/s the difference between the two band limits remains limited to some tenths of db.

These small differences mean that the choice between the two bands considered can be done on a conventional basis.

In the event that the weighted echo-balance return loss is defined as a weighted mean power ratio over the band 500-2500 c/s, it can be calculated by means of the expression (6) and the weighting function a_W, according to Table 1.

ANNEX 4

(to Question 6/XII)

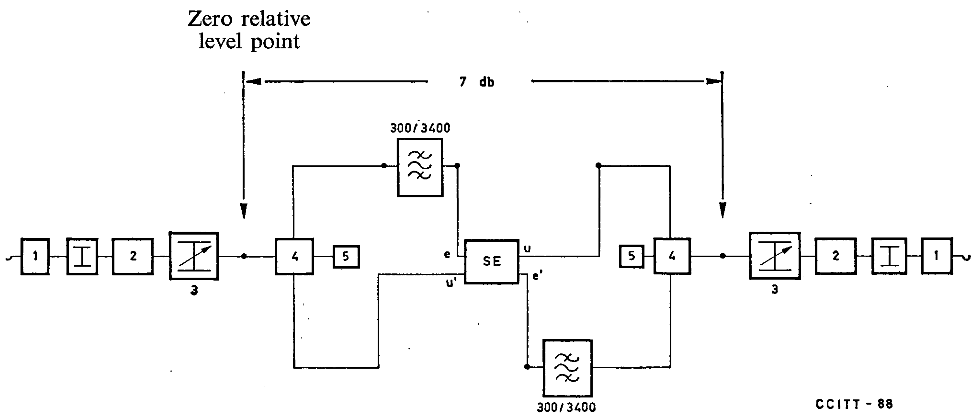
Subscribers' tolerance of very long propagation times

(Contribution of the Italian Administration)

First series of tests

A series of opinion tests was carried out on international circuits using modern telephone sets which were subjected to group delay (equal in both directions) of 0.3 and 0.6 second. To ascertain the effect of group delay, the same circuits were tested with negligible group delay.

The testing arrangement, which is shown in Figure 1, represents a typical international call in which the chain of four-wire circuits has a fixed equivalent of 7 db (towards the two-wire terminals).



1 = telephone set
2 = feeding bridge
3 = adjustable attenuator
4 = terminating set
5 = balancing network
SE = delay unit

FIGURE 1. — **Test circuit for the first series of tests**

In Figure 1, SE denotes an apparatus arranged especially for these tests which basically (see Figure 2) is a " looped " magnetic tape recorder-reproducer. It has two channels, i.e. two groups of erasing circuits (E), of recording circuits (R) and of reproducing circuits (P), all operating at once.

By adjusting the distance between the magnetic heads *P* and *R*, we obtain two adjustable delay lines between *e* and *u* and between *e'* and *u'*.

In the test diagram mentioned above, therefore, the effects of group delay on both directions of the call were analysed. The tests were made without room noise.

Particular care was paid to the balancing of the four-wire terminating sets so that only the effect of long group delay, and not echo side-effects, was produced.

The various conditions created during the tests were as follows:

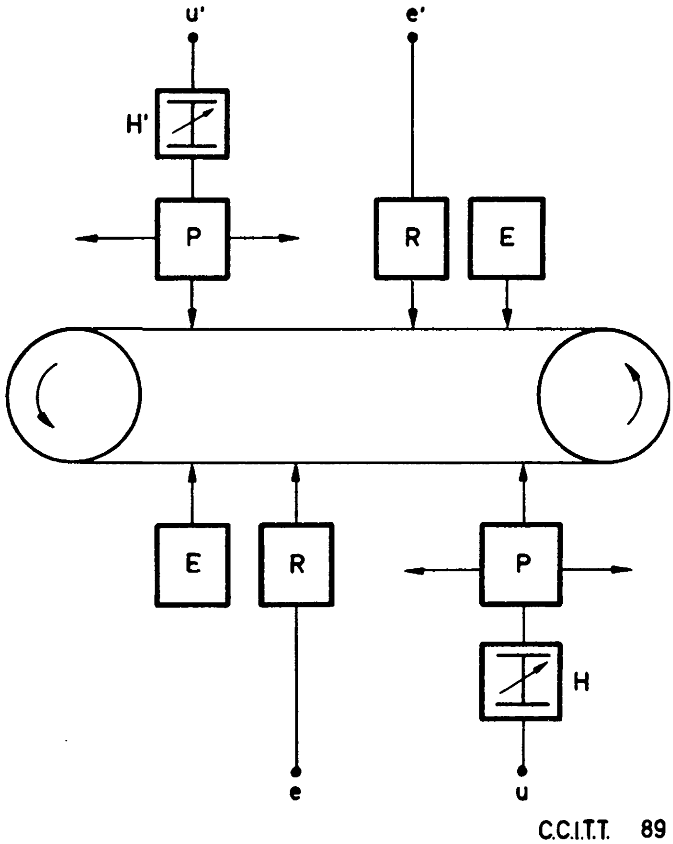


FIGURE 2. — **Two-channel delay unit**

1. The values +20, +30 and +40 db were adopted as the reference equivalent of the entire connection (by adjusting attenuators 3 of Fig. 1).

2. The values 0, 0.3 and 0.6 seconds were adopted as the group delay. The results obtained are reproduced in Figure 3, which indicates that it is possible to ensure the same average performance on a circuit with long-group delay as on a circuit with no group delay whatever, by reducing the reference equivalent as required.

As may be seen in Figure 3, this reduction may be considered as having a value of approximately 10 db to 3 db.

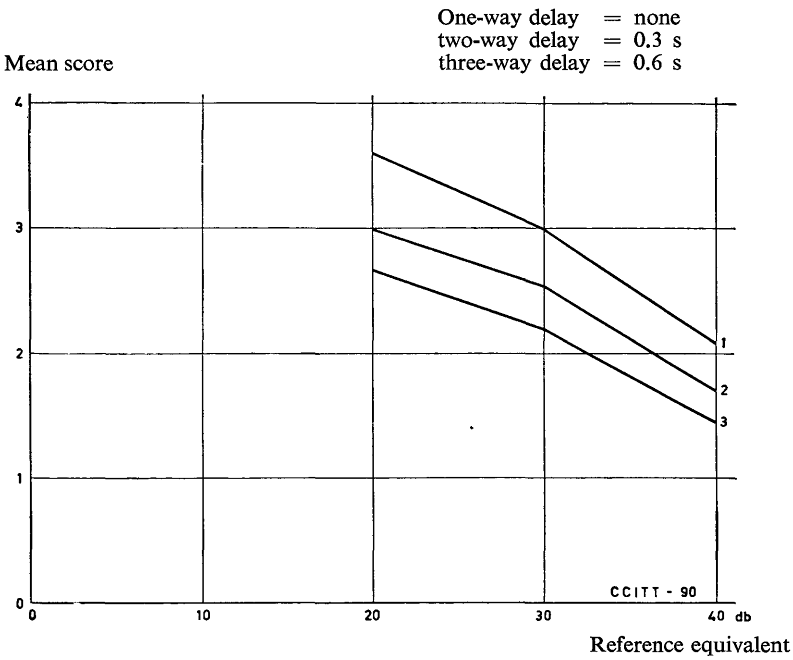


FIGURE 3

Second series of tests

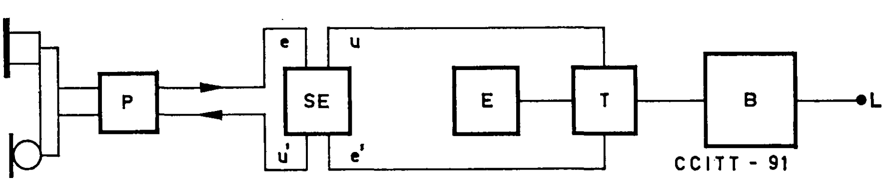
In connection with calls affected by very long group delays a series of tests, of which preliminary results are given above, have been carried out for the purpose of ascertaining the inconvenience caused in calls by the delay itself—that is to say in circumstances in which the secondary effect of echoes was reduced to a negligible minimum by very careful balancing of the terminal equipment.

The tests, the results of which will be given below, were carried out on actual outgoing calls from a telephone station in the normal course of the telephone user's work.

The delay (the same in both directions) was obtained by replacing the two-wire terminal between the apparatus and the private automatic telephone by a four-wire line with a terminal set, to which a two-channel tape recorder was connected capable of providing the required delay in both directions (see Figure 2).

The arrangement of the apparatus for the tests is shown in Figure 4.

At the start of the conversation no information was given to the second speaker about the experimental nature of the circuit.



P = subscriber's set
SE = delay unit
T = terminating equipment
E = terminating set
B = private automatic exchange
L = subscriber's line

FIGURE 4. — **Circuit for the second series of tests**

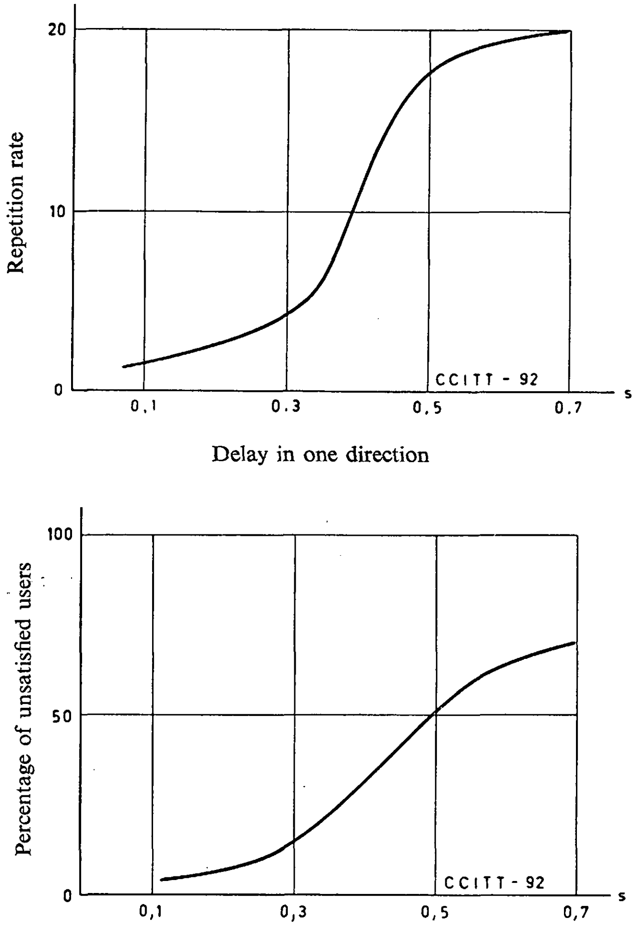


FIGURE 5

During the call, the user of the fixed station took note of repetitions and impediments which arose during the call.

At the end of the call, the user of the fixed station asked the second user for his opinion of the circuit quality—to be expressed either as " good " or " bad ".

The delays used (the same in both directions) were 0.1 s; 0.3 s; 0.5 s; 0.7 s.

The results obtained for 20 calls, with each delay applied at random, are shown in Figure 5, in the form of the repetition rate (number of repetitions in 100 seconds) as a function of the delay and also the percentage of unsatisfied users as a function of the delay.

These results show that a remarkable intolerance to delay occurs quite abruptly for delays which exceed 0.4 to 0.5 second.

Question 7/XII — Determination of transmission quality by objective measurements

(continuation of Question 7/XII, 1961-64)

a) What criterion should be adopted to determine equality of performance of different transmission systems?

b) What practical method, based on objective measurements, should be specified for the determination of the transmission performance of local telephone systems?

Note 1. — The following Annex 1 gives directives on the method to be followed in studying this question.

Note 2. — Annex 2 gives a method, based on information theory, of calculating transmission quality from the results of objective measurements.

Note 3. — Annex 31 (Part II of Volume V of the *Red Book*) describes an objective method, known as the tonality method, for measuring articulation, evolved by the U.S.S.R. Administration.

ANNEX 1

(to Question 7/XII)

Directives for the study of Question 7/XII

1. The first essential of a method for measuring the quality of a telephone circuit is that it should give results which correspond to the experience of any user employing the telephone for the purposes of everyday life. Furthermore, the method should be both simple and practical.

2. There are four types of measurement that may be considered here:

a) Assessment of quantities, such as loudness made with people speaking, listening and carrying on a conversation;

b) Assessment of quality (i.e. including the effects of noise, distortion, sidetone, echo, etc.) made with people speaking, listening and carrying on a conversation;

c) Measurement of quantities, such as loudness, using instruments, including artificial voices, mouths and ears;

d) Measurement of quality (i.e. including the effects of noise, distortion, etc.) using instruments including artificial voices, mouths and ears.

The main advantages of measurements c) and d) are rapidity, precision and low cost.

3. The only method of type a) accepted by the C.C.I.T.T. is that of reference equivalents, which does not perhaps fully satisfy the first condition mentioned in paragraph 1. The questions of a reasonably realistic speaking distance and vocal level are among those which deserve attention.

4. Since it seems probable that methods of type c) will replace those of type a), thorough studies of them are called for; Question 15/XII has been set in this connection. For this purpose, satisfactory answers to Questions 8/XII and 12/XII are needed.

5. Since it seems probable that methods of type d) will replace those of type b) in certain respects, it would be useful to collect information on the best and most widespread methods of type d).

ANNEX 2

(to Question 7/XII)

Calculation of telephone transmission performance information theory

by J. Lalou, C.C.I.T.T. Counsellor

Summary

Main symbols used

$A \quad = \int_B a(f)\, \mathrm{d}\, U$

$a \quad =$ loss in db or nepers

$\alpha \quad = \dfrac{f - f_i}{f_i}$

$B \quad =$ frequency bandwidth of a channel, in c/s

$C \quad =$ capacity of the condenser in a resonant circuit, in farads

C_i, $C(f)$ = weighting coefficient in the Δf_i band or $\mathrm{d}f$

Δf_i = $f_i - f_{i-1} \approx f_{i+1} - f_i$

E = $\int_B e(f)\,\mathrm{d}U$

e = sensitivity of a sending system, in db with respect to 1 mV/dyne/cm^2 (or 1 V/millibar)

F = frequency band occupied by noise, in c/s

f = variable frequency, in c/s

f_i = centre frequency of a filter of rank i, in c/s

H = information (negative entropy), in bits

I = channel capacity, in bits per second

J = information rating, in decibels

K = critical bandwidth, in c/s

L = self-inductance of a resonant electric circuit, in henrys

l = circuit noise level, in dbm

l_s = spectrum power of circuit noise at ear input

m = number of filters in the receiver

N = $\int_B n_s(f)\,\mathrm{d}U$

$\overline{N}$ = average power (in time) of noise

N_s = average spectrum power of noise in a band of 1 c/s or hearing threshold for a sound with a continuous spectrum

n = mean opinion score

n_s = $10 \log_{10} N_s$

ν = exponent occurring in the expression for loudness

ξ = $2\,Q\alpha$

P_0 = reference acoustic power $\left(10^{-16} \dfrac{\mathrm{W}}{\mathrm{cm}^2}\right)$

p = probability of a symbol among $\dfrac{1}{p}$ equiprobable symbols

Q = Q factor of a resonator

q = objective reference equivalent

R = $\int_B r(f)\,\mathrm{d}U$

R_i = resistance of a resonant electric circuit in ohms

r = sensitivity of a receiving system, in db with respect to 1 dyne per cm^2/mV (or 1 millibar/V)

$\overline{S}$ = mean power of wanted signal

S'_i = mean power of the disturbing signal in channel i

S_s = spectrum power of wanted signal in a band of 1 c/s

s = $10 \log_{10} S_s$

T, t = duration of transmission (in seconds)

τ = time constant

$U(f)$ = weighting function

u_i = electric voltage at frequency f_i

$V(x)$ = loudness (subjective, in sones)

$v(x)$ = $10 \log_{10} V(x)$

$X(f) = \frac{S_s}{N_s}$

$x = 10 \log_{10} X$

$Y_i(f-f_i) \equiv y_i(f)$

$y(f)$ = function characterizing the selectivity of a filter

Z = impedance of a resonant electric circuit

$\omega = 2\pi f$

1. *Principle of the calculation method*

The capacity of a channel in the presence of noise, i.e., the maximum information flow which it is capable of transmitting with an arbitrarily low error rate, is generally calculated using formula [1]:

$$I = B \log_2 \left(1 + \frac{\overline{S}}{\overline{N}}\right) \tag{1}$$

Comparing this formula with the general formula [2]:

$$H = \log_2 \frac{1}{p} \tag{2}$$

in which we take for $\frac{1}{p}$ the number of equiprobable symbols transmitted *per second*, we may write:

$$I = \log_2 \left[\left(1 + \frac{\overline{S}}{\overline{N}}\right)^B\right]$$

$\frac{1}{p}$ is then the number of combinations of a code with B units and $1 + \frac{\overline{S}}{\overline{N}}$ significant conditions.

We may imagine that B pulses per second are transmitted each of duration $T = \frac{1}{B}$ and with $1 + \frac{\overline{S}}{\overline{N}}$ amplitudes discernible for reception in the presence of noise, so that we may assume, with an arbitrarily low error probability, that the power of the signal, in the absence of noise, would have one of the values 0, $2\,\overline{N}$, $4\,\overline{N} \ldots 2\,\overline{S}$, of which the mean value is equal to $\overline{S}$.

This physical interpretation is very suggestive in the case of actual pulse transmission—for instance, in data transmission—but does not, when applied to telephony, bring out any clear relation to the properties of the human voice and ear. Let us consider a hypothetical transmission system comprising:

a) a generator capable of transmitting sinewaves at m frequencies $f_1 \ldots f_i \ldots f_m$ and a variable level;

b) the electroacoustic transmission system under study;

c) a receiver containing m filters with $f_1 \ldots f_i \ldots f_m$ as the centre frequencies.

This system may be regarded as consisting of m channels in parallel. So long as these are independent, the capacity of channel i can be calculated by a formula similar to (1)

$$I_i = \frac{1}{t_i} \log_2 \left(1 + \frac{S_i}{N_i}\right)$$

where

S_i is the mean power of the signal at frequency f_i,

t_i is the time required for the signal to attain a sufficient amplitude for measurement (and for the preceding signal to revert to a negligible amplitude),

N_i is the mean power of the over-all background noise picked up by the filter.

If the filter's selectivity is characterized by a function $y_i(f)$, proportional to the power received at frequency f and equal to 1 for $f = f_i$, we shall have

$$N_i = \int_B N_s(f)\, y_i(f)\, \mathrm{d}f$$

To increase the total capacity, a large number of frequencies such as f_i have to be used; since the filters have a finite selectivity, interference is produced between the signals transmitted in neighbouring channels. If the filter's selectivity is sufficient and the signal power does not vary abruptly with frequency, signal S_i can be interfered with only by the two adjacent signals. The over-all mean power of the interfering signal is then

$$S'_i = S_{i-1}\, y(f_{i-1}) + S_{i+1}\, y(f_{i+1})$$

If we put $S'_i = \lambda N_i$, the signal will have as the peak value $2\lambda N_i$ and may mask the λ values $0.2\, N_i, \ldots 2(\lambda - 1)\, N_i$ of S_i, thereby reducing the value of I_i to $\frac{1}{t_i} \log_2 \left(1 + \frac{S_i}{N_i} - \frac{S'_i}{N'_i}\right)$. Since account has been taken of mutual interchannel interference, the total capacity is $I = \Sigma\, I_i$.

If it is also admitted that the frequencies and levels vary progressively and that the frequencies $f_i \ldots$ are in close proximity, then we have substantially

$$f_i - f_{i-1} = f_{i+1} - f_i = \Delta f_i \qquad \frac{S_{i-1} + S_{i+1}}{2} = S_i$$

and if $Y(-\Delta f) = Y(\Delta f)$, the total capacity can be written as:

$$I = \sum_{i=1}^{m} \frac{1}{t_i} \log_2 \left[1 + \frac{S_i}{N_i} [1 - 2\, Y(\Delta f_i)]\right] \tag{3}$$

2. *Application to the human voice and to the auditory system*

To go beyond this the filters' characteristics must be defined. We shall proceed with the calculation to cover the case of a receiver with simple resonators which would seem to represent well enough the selectivity properties of the human auditory system [1]. We shall suppose that a generator possessing certain statistical properties of the human voice excites the receiver through the transmission system under study.

The following must not in any way be regarded as a theory of the hearing *mechanism*. The purpose is merely to compare the final result of the calculation with the results obtained by conventional methods for the assessment of transmission performance.

2.1 *Determination of the function Y*

Measurements and calculations made at three frequencies [3] have shown that the selective properties of the auditory system correspond, with a sufficient degree of accuracy, to those of a set of simple resonators tuned to the frequencies to be received.

Without attempting to define the mode of operation of such a resonator, we may liken it to its electrical counterpart, characterized by the parameters R_i, L, C and with an impedance

$$Z = R_i + \mathrm{j}\left(L\,\omega - \frac{1}{C\omega}\right)$$

By putting: $2\pi f_i = \omega_i = \frac{1}{\sqrt{LC}}$ and $Q = \frac{L\omega_i}{R_i}$ we obtain

$$y(f) = \frac{1}{1 + Q^2 \left(\frac{f}{f_i} - \frac{f_i}{f}\right)^2} \tag{4}$$

[1] This system includes the ear, together with the components of the nervous system which take part in the hearing process (including the brain centres). The over-all response will be considered throughout.

Putting $\dfrac{f}{f_i} = 1 + \alpha$, we have

$$Y(\alpha) = \frac{1}{1 + Q^2 [1 + \alpha - (1 + \alpha)^{-1}]^2} \tag{5}$$

If α is small in respect to unity, we may write:

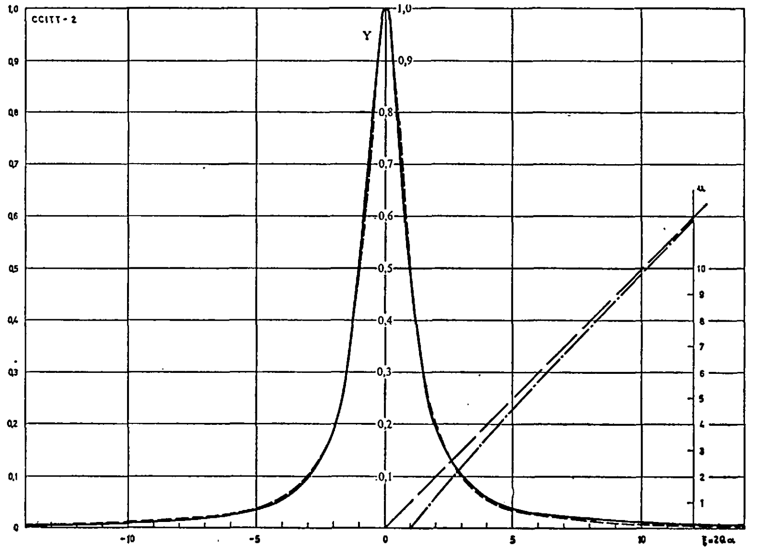
$$Y(\alpha) = \frac{1}{1 + Q^2 (2\alpha - \alpha^2 + \alpha^3 \dots)^2}$$

and, whatever the order of magnitude of $Q\alpha$, α^2 and the following terms in brackets may be neglected in view of 2α. We then have an approximate expression:

$$Y(Q\alpha) = \frac{1}{1 + 4\, Q^2 \alpha^2} \tag{6}$$

which is only a function of the variable $\xi = 2Q\alpha$ and where Q no longer appears as a parameter, and thus a single selectivity curve in normalized units.

As an example, the curve delineated by a continuous line in Figure 1 represents $Y(\xi)$ according to formula (5), for a typical value $Q = 25$ and the curve shown by a broken line the approximate formula (6). It can be seen that the divergence between these two curves is very slight for all values of ξ for which Y has an appreciable value.



———— $Y(\alpha) = \dfrac{1}{1 + Q^2 \left(1 + \alpha - \dfrac{1}{1 + \alpha}\right)^2}$ $(Q = 25)$

-------- $Y(\xi) = \dfrac{1}{1 + \xi^2}$

—·—·—·— $u = \xi \left(1 - \dfrac{2}{1 + \xi^2}\right)$

— — — — $u = \xi$

FIGURE 1. — **Selectivity curve**

2.2 *Relation between the Q-factor and the critical bandwidth*

The Q factor can be calculated from the measured value of the critical bandwidth K. FLETCHER ([4], Chapter 10) has defined this by the following two properties:

If a noise with a uniform spectrum in the frequency band F and nil outside this band masks one pure sound whose frequency is situated in the middle of this band, the power of the masked sound is equal to the noise power integrated in a band of width:

$$F \text{ if } F \leqslant K$$

$$K \text{ if } F > K.$$

Then, if f_1 and f_2 are the limits of band F, the expression of this power, according to equation (4), is

$$\overline{N} = \int_{f_1}^{f_2} \frac{N_s}{1 + Q^2\left(\dfrac{f}{f_i} - \dfrac{f_i}{f}\right)^2}\, \mathrm{d}f$$

N_s being constant, the bandwidth in which the spectrum components of the noise are power additive is $\dfrac{\overline{N}}{N_s}$. Noting that $\mathrm{d}f = f_i \mathrm{d}\alpha = \dfrac{f_i}{2Q}\,\mathrm{d}\xi$, the following approximate expression can be used:

$$\frac{\overline{N}}{N_s} = \frac{f_i}{2Q} \int_{\xi_1}^{\xi_2} \frac{\mathrm{d}\xi}{1 + \xi^2}$$

whence

$$\frac{\overline{N}}{N_s} = \frac{f_i}{2Q} \Big[\operatorname{arctg} \xi\Big]_{\xi_1}^{\xi_2}$$

as $\xi_1 = -\xi_2$

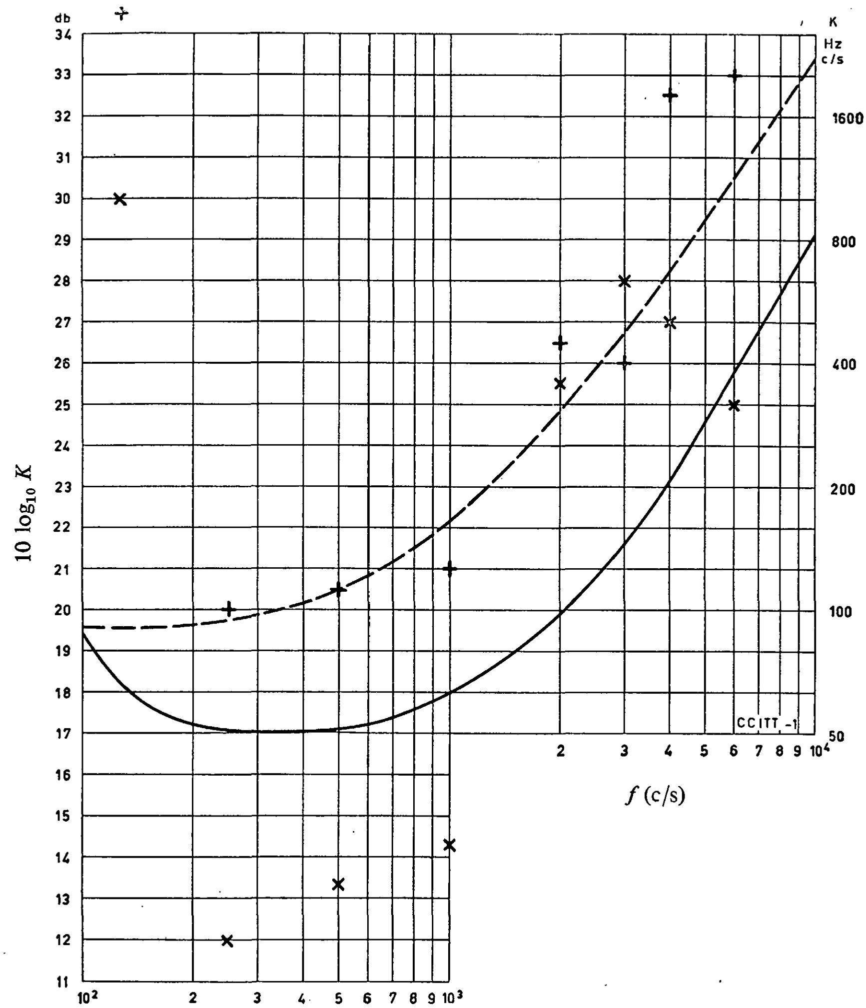
$$\frac{\overline{N}}{N_s} = \frac{f_i}{Q} \operatorname{arctg} \xi_2 \tag{7}$$

If ξ_2 is small, arctg ξ_2 can be replaced by its principal part, i.e., $\xi_2 = 2Q\dfrac{F}{2f_i}$ and we find: $\dfrac{\overline{N}}{N_s} \approx F$ which shows, according to the first property of the critical bandwidth, that the level of the masked pure sound is in fact equal to $\overline{N}$. According to the second property, K must be the limit of $\dfrac{\overline{N}}{N_s}$ when F becomes large, i.e.,

$$K = \frac{f_i}{Q}\frac{\pi}{2} \tag{8}$$

This reasoning is not rigorous since when $Q\alpha$ becomes very great, α cannot remain small compared with unity. In fact arctg ξ tends towards the limit $\dfrac{\pi}{2}$ so rapidly that, for the usual values of Q, the error due to this approximation may be ignored.

Figure 2 shows the values of K found by different authors. These values differ to an appreciable extent but their variations with frequency show the same general trend. The final formulae which will be obtained subsequently are such that, if the values of Q at various frequencies are multiplied by a constant factor, the total information is itself multiplied by a constant factor — a fact which is of little importance in practice. There is thus no risk that the choice of one or other of these sets of values for K will entail the introduction of a serious error.



———— *Fletcher* [4], p. 101, applicable to monaural listening, according to [5]

— — — — *Zwicker, Flottorp* and *Stevens* [6]

Values measured by *Lehmann* [7] with noise bands one octave wide; level of unfiltered noise
(40 db +)
(70 db ×)

FIGURE 2. — **Critical bandwidth (K) as a function of frequency, according to various authors**

Formula (8) has been used to calculate the values of Q corresponding to the values of K given by FLETCHER (Figure 2). These values are shown in Figure 3. It will be seen that at very low frequencies, Q assumes very small values; consequently, the use of the approximate formula (6) and of formula (8) derived therefrom is hardly justifiable. Nevertheless, these very low frequencies play an insignificant part in the total information, for various reasons:

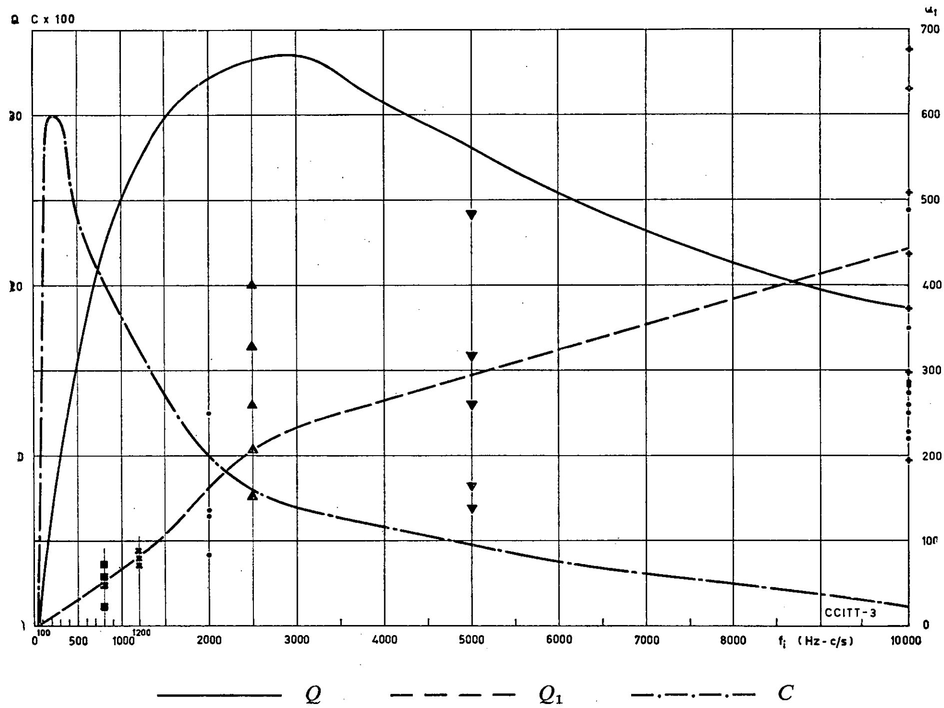


FIGURE 3

— the weighting coefficient tends to zero (see paragraph 2.4);

— the sensitivity of the ear is low;

— in practice, these frequencies are always greatly attenuated by some part of the transmission system.

The values of Q given in Figure 3 may thus be regarded as sufficiently accurate.

The critical bandwidth depends strictly on the level of the signals [7]; nevertheless, because of the divergencies mentioned above between the values measured in different laboratories, it has not been feasible to take account of this effect.

2.3 *Determination of the optimum spacing of the signal frequencies*

The human voice cannot emit simultaneously a large number of independent frequencies. Nevertheless, since we are endeavouring to assess the maximum information which can be transmitted to the auditory system by a sound source possessing some of the characteristics of the voice, we may seek to establish Δf_i in order to have the maximum of the expression (3). The spectrum of the source considered must, of course, correspond to the same power distribution as the mean spectrum curve of the human voice; thus, we will assume that for each pure sound of frequency f_i, inside a free band Δf_i, we have $S_i = S_s \cdot \Delta f_i$. For each frequency f_i, t_i is fixed and the logarithm, which is usually in the neighbourhood of $\log_2 \left[\frac{S_i}{N_i}[1 - 2\,Y(\Delta f_i)]\right]$ must be maximized. When spectrum powers are introduced, $S_i = S_s \cdot \Delta f$, and, provided that Q is sufficiently large (very selective filter) and that there are no noise components at other frequencies much higher than N_s, we have $N_i = N_s \cdot K$.

It also seems desirable to contract the frequency spacing in the high-selectivity regions (high Q factor). This condition will be fulfilled if the same value of $\xi = 2\,Q\dfrac{\Delta f}{f}$ is taken for all the Δf_i. Moreover, the total number m of frequencies is inversely proportional to the common value chosen for ξ (the values of Q_i being fixed).

We are therefore led to choose ξ in such a way as to maximize:

$$\frac{f}{\Delta f}\log_2\left[\frac{S_s}{N_s}\frac{\Delta f}{K}\,[1 - 2\,Y(\Delta f)]\right]$$

or

$$\frac{1}{\xi}\left[\log_2\frac{S_s}{\pi\,N_s} + \log_2 u\right]$$

putting

$$u = \xi\left(1 - \frac{2}{1 + \xi_2}\right)$$

It is always necessary to take $\xi > 1$, if the information transmitted at each frequency is not to be nil and this information quantity increases with ξ. The curve representing $u\,(\xi)$, plotted on Figure 1, shows a point of inflection at $\xi = \sqrt{3}$; u then increases more slowly and tends to ξ. Numerical calculations, made for several values of $\dfrac{S_s}{N_s}$ that are encountered in practice, show that a maximum is reached for a value of ξ which depends to some extent on $\dfrac{S_s}{N_s}$, but is always of the order of 2. When ξ becomes sufficiently large, $\log_2 u$ is greater than the constant term in the term between brackets; the quantity whose maximum is being sought then behaves as $\dfrac{1}{\xi}\log_2\xi$ and therefore tends to zero.

Hence, it seems reasonable to assume $\xi \approx 2$ as the value near to the optimum, or even $\xi = \dfrac{2\,\pi}{3}$ which corresponds to $\Delta f = \dfrac{2}{3}K$.

The corresponding value of y is 0.2. Formula (3) can consequently be replaced by:

$$I = \sum_{i=1}^{m}\frac{1}{t_i}\log_2\left(1 + 0.4\frac{S_s}{N_s}\right) \tag{9}$$

If we put: $C_i = \dfrac{1}{t_i\,\Delta f_i}$, we may write

$$I = \sum_{i=1}^{m} C_i\frac{10}{3.01}\log_{10}(1 + 0.4\,X)\,\Delta f_i \tag{10}$$

The Δf_i exactly fill the total frequency band B, with neither spaces nor overlapping; provided that they are small, the preceding expression can thus be replaced by the integral

$$I = \int_B C(f)\,10\log_{10}(1 + 0.4\,X)\,df \tag{11}$$

2.4 *Determination of the duration t and of the weighting function*

This determination amounts to that of t_i and Δf_i, or of t_i and Q_i. Since Q has already been determined above, it remains to determine t_i.

The transients in the equivalent resonant electric circuit are determined by the function $\exp\left(-\frac{R_i t}{2L}\right)$; in other words, the time constant of this circuit is

$$\tau = \frac{2L}{R_i} = \frac{2Q}{\omega_i} = \frac{Q}{\pi f_i} = \frac{1}{2K}$$

t_i must be of the order of τ, i.e. $t_i = \mu\,\tau$. It will be seen that C_i is a constant $C_i = \frac{3}{\mu}$, this result being independent of Q. In particular, if Q is very large, the resonant circuits approximate to perfect filters with a bandwidth K; if in formula (10) the spacing Δf is replaced by K, the coefficient 0.4 must be replaced by 1 for the perfect filter, when formula (1) is found again, which is indeed the case if $t_i = \frac{1}{K}$. Thus we should have $t_i = 2\,\tau_i$ and a value of Q can be derived from a measurement of t_i:

$$Q_1 = \frac{\pi}{2} f_i\, t_i \tag{12}$$

Now, Pumphrey and Gold [8] have determined Q from the properties of the ear in the transient state, which is the same as calculating Q_1, and have found much higher values than those obtained from the critical bandwidths, as can be seen in Figure 3. A phenomenon other than transients in the resonators thus intervenes in the auditory system to increase t_i.

Q_1 must then be considered as a determination of t_i, the true Q being

$$Q = \frac{\pi f_i}{2K}$$

By eliminating f_i and t_i between the expressions of C_i, Q and Q_1, we find

$$C_i = \frac{3}{2}\,\frac{Q}{Q_1}$$

By including the numerical coefficient 0.33 (due to the introduction of decibels instead of base 2 logarithms) in the weighting coefficient, we have

$$C(f) = 0.5\,\frac{Q}{Q_1} \tag{13}$$

It is this formula which was applied, taking for Q the values (Fig. 3 above) deduced from the critical bands shown by Fletcher, and for Q_1 the values measured by Pumphrey and Gold [8]. The dots in Figure 3 represent these latter values; we see that they are fairly scattered. Actually, they were obtained with only two subjects, one of which gave values only at 10 kc/s; these latter values (represented by the sign ●) have been eliminated since, to judge the general trend of the variation of Q_1, it seemed preferable to use the results of the same subject at all the frequencies.

The simplest assumption to explain the difference between Q_1 and Q is to assume that a constant delay d is introduced (probably by the nervous system), when we should have

$$t = 2\frac{Q}{\pi f} + d$$

whence

$$Q_1 = Q + \frac{\pi\, d f}{2}$$

In a somewhat mediocre fashion, such a relation represents the results of the experiments. We used the curve traced freehand by eye on Figure 3, between the representative points of the results of experiment; it would be desirable to have more precise results.

By means of formula (13) values of $C(f)$ represented on Figure 3 were calculated. The values in Figures 4a and 4b were deduced therefrom by graphical integration.

$$\int_0^f C(f)\,\mathrm{d}f = 700\,U(f) \tag{14}$$

with U (10 000) = 1 by convention.

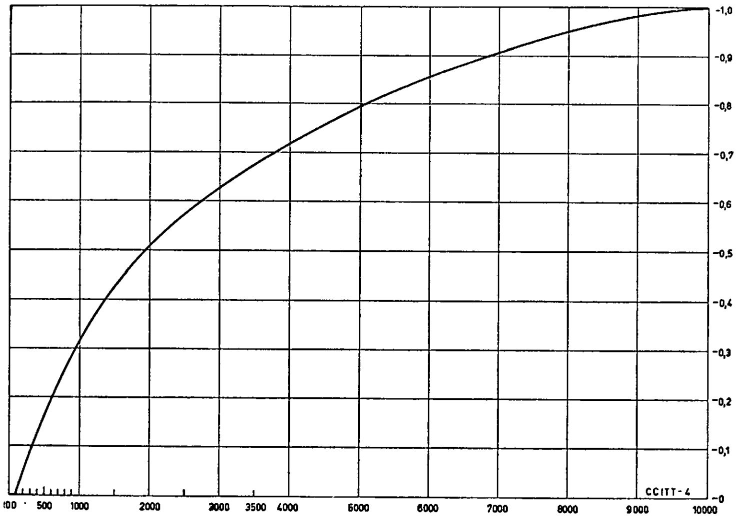
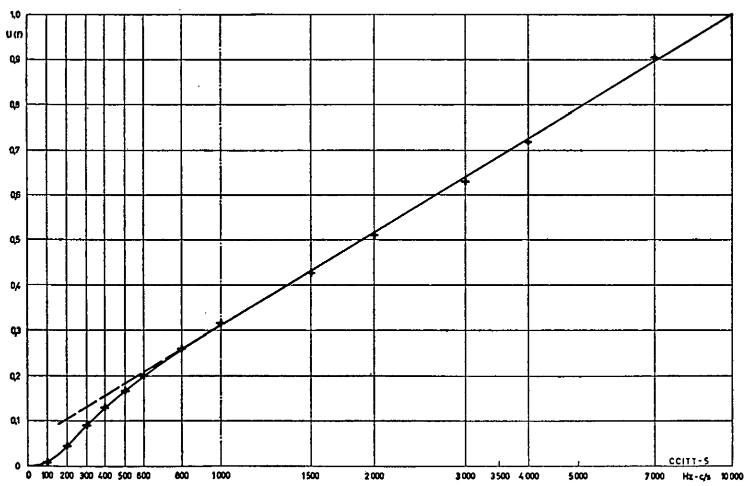


FIGURE 4a. — $U(f)$ **with linear frequency scale**



+ Calculated points

FIGURE 4b. — $U(f)$ **with a logarithmic scale of frequencies above 1000 c/s**

2.5 *Final formulae*

According to (11) and (14), the capacity I of a telephone channel is

$$I = 700\,J \tag{15}$$

where the numerical coefficient is the product of 1/3 (conversion factor for decibels into base 2 logarithms) by 2100 c/s, which can be regarded as the " equivalent bandwidth " of the whole audible frequency spectrum; incidentally this coefficient is not known with great precision; J, which we shall call " information rating ", is expressed in decibels and is expressed:

$$J = \int z\,(f)\,\mathrm{d}\,U \tag{16}$$

where the integral is extended over the audible frequency spectrum;
$U\,(f)$ is given by the curve in Figure 4 (a or b);
at each frequency we have:

$$z = 10 \log_{10} (1 + 0.4\,X) \tag{17}$$

Figure 5 represents the variation of z as a function of x, x being the signal/noise ratio or (in the absence of noise) the level of the signal above the hearing threshold in the band considered.

When x is greater than about 15 db, we have:

$$z = x - 4 \text{ db} \tag{18}$$

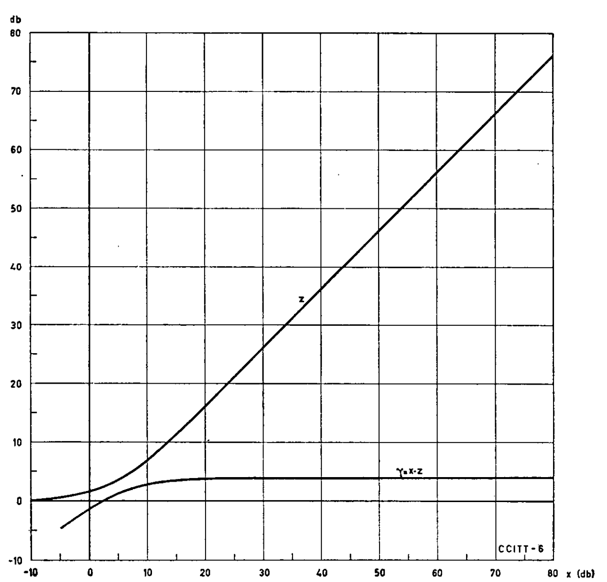


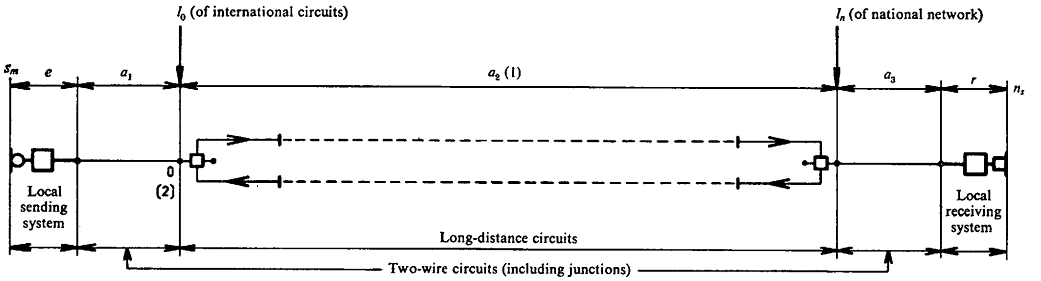
FIGURE 5

Appendix 1 shows an example of the practical application of these formulae and curves.

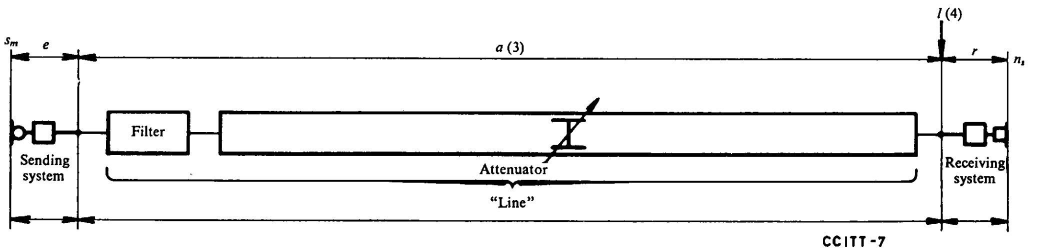
3. *Network planning method*

3.1 *Principle*

The basic problem in the planning of telephone networks lies in allocating to the different parts of a connection, such as that represented in Figure 6a, ratings whose sum gives the over-all



a) Long-distance telephone call



b) Laboratory model equivalent to a) *(for one direction of transmission)*

Notes: (1) $a_2 = 7 + 0.5\, n$ db if there are n international circuits
(2) o is the zero relative level point for the first international circuit
(3) $a = a_1 + a_2 + a_3$ (with corrections for reflection losses, if necessary)
(4) l corresponds to the sum of powers corresponding respectively to $l_0 - a_2 - a_3$ and to $l_n - a_3$.

FIGURE 6. — **Examples considered for the planning method**

rating for the connection. Since the laws for the addition of line attenuations are known, we may confine ourselves to considering the laboratory model in Figure 6b.

Let:

s_1 be the spectrum power level of speech sounds at 1 m from the speaker's mouth;

d $= s_m - s_1$ (s_m being the spectrum power on the microphone of the sending system);

e the sensitivity of the local sending system, in db relative to 1 mV/dyne per cm^2 (or 1 V/millibar);

a line attenuation;

r the sensitivity of the local receiving system, expressed in units corresponding to those used for e;

n_s the spectrum power level of the noise at the entry to the listener's ear.

For the complete connection, we have at each frequency:

$$x = s_1 + d + e - a + r - n_s \tag{19}$$

and according to formula (17)

$$z = s_1 + d + e - a + r - n_s - \gamma \tag{20}$$

where γ is a corrective term which, according to formula (18), amounts to 4 db when x is sufficiently large.

By integrating over the whole of the band B where z has an appreciable value we obtain:

$$J = S_1 + D + E - A + R - N - \Gamma \tag{21}$$

where J is the information rating of the connection, given by formula (16)

$$S_1 = \int_B s_1(f)\,\mathrm{d}\,U$$

and the other quantities are defined by similar formulae.

According to formula (17) or Figure 5, z is very small as soon as $x < 0$. Hence, we can, without appreciable error, identify band B with the band where $x \geqslant 0$. Over a large part of this band, x is sufficiently large (if J corresponds to an acceptable quality) and γ is substantially equal to 4 db; over the remainder of this band, γ can vary only between $+4$ and -1.5 db. It may therefore be expected that the mean of γ throughout the transmitted band is about 2 to 3 db; in what follows, a mean of 2.5 db will be assumed. We then have:

$$\Gamma = 2.5 \int_B U(f)\,\mathrm{d}f \quad \text{db} \tag{22}$$

Consequently, when B is determined, Γ is a constant; the same applies to S_1 (limited to B); if J_0 corresponds to the minimum transmission performance required of a connection, we must have

$$D + E - A + R - N \geqslant J_0 - S_1 + \Gamma \tag{23}$$

The second member being a constant, this inequality defines a method for planning networks.

We have thus defined ratings applicable to part of the connection—or to the effect of noise and of the hearing threshold on the receiving system. However, the magnitudes of these ratings depend on band B transmitted by the complete connection. In practice, this band is determined essentially by the filters of the long-distance carrier circuits. At the limits of the frequency band effectively transmitted (in the C.C.I.T.T. sense) by the chain of long-distance circuits, a is about 9 db higher than the values obtained in the central part of this band and it grows rapidly towards the frequencies that are lower or higher than each of the limits. This cut-off effect is accentuated towards the lower frequencies by the increase in the room noise level which passes between the receiver earcap and the ear (or, in the absence of room noise, by the rapid rise in the hearing threshold) and towards the higher frequencies by the decrease in s_1 and e (see the curve of the conventional telephone signal, *Red Book*, Volume III, page 51 or, to be more precise, *Blue Book*, Volume III, Recommendation G.227). In practice, B will therefore be very close to the frequency band effectively transmitted by the chain of long-distance circuits; as this band forms the subject of recommendations or current studies by the C.C.I.T.T., there will be only a small number of situations to be considered for the calculation of sending and receiving system ratings, in the presence of noise having limit values.

3.2 *Practical example*

The curves representing, as a function of the frequency, the variables defined in paragraph 3.1 for $a = 30$ db at the centre of the transmitted band and $l = -60$ db have been plotted in Figure 7, on the basis of the information supplied in [13]. The abscissae scale in this figure is proportional to $U(f)$ (scale of ordinates in Figure 4). In the connection under consideration, the line filter is the S.R.A.E.N filter (*Red Book*, Volume V, page 67), for which the effectively transmitted frequency band ranges from about 200 to 3500 c/s. According to Figure 4, $U(200) = 0.05$ $U(3500) = 0.67$ i.e. if this band were transmitted perfectly, a gain of 10 db throughout the band would increase J by 6.2 db. If we compare two by two the values of Table 4 in Appendix 1 corresponding to an acceptable quality, we find that this gain increases J by 6.0 db on the average. This latter result will be obtained by dividing by 20 the sum of values obtained for 12 frequencies corresponding to equal increases of $U(f)$ in the band under consideration. These frequencies are indicated in the first columns of Table 1. Using this table, the following ratings were calculated for $l = -60$ dbm and a line attenuation of 30 db:

$$S = 33.2 \qquad E = 16.6 \qquad A = 18.4 \qquad R = -16.6 \qquad N = 8.1 \qquad \text{db}$$

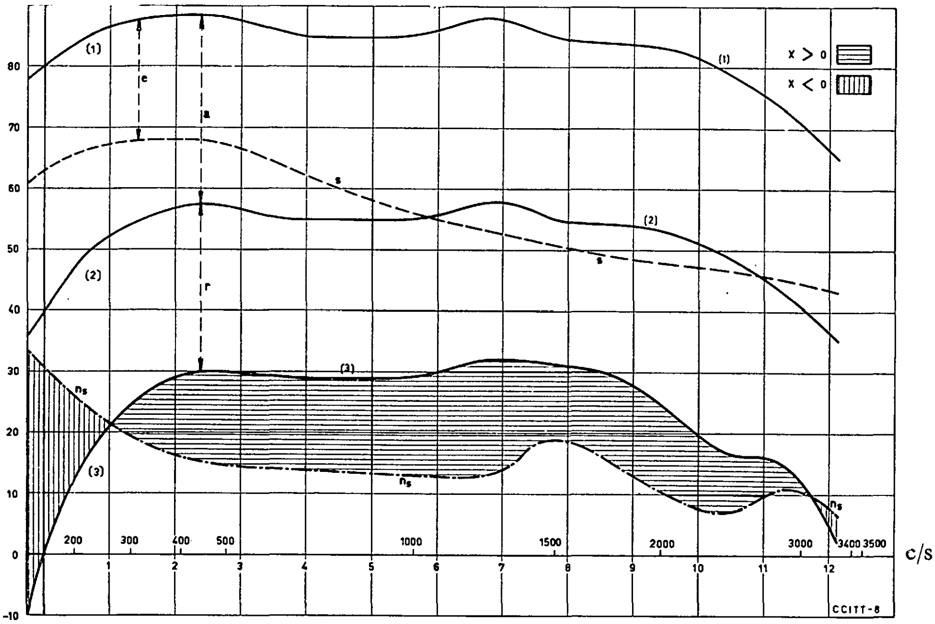


FIGURE 7. — **Example of the addition of components of** x (0 db corresponds to 10^{-16} W/cm²)

The numbered curves represent the levels at the input to (1) the line, (2) the receiving system (3) the ear

TABLE 1

Calculation of information ratings for the various parts of a connection

Line attenuation 30 db

Frequency		Speech spectrum at the micro-phone	Sending sensitivity	Line attenuation	Receiving sensitivity	Value of n_s for I =			
No.	Approxi-mate value					$-\infty$	−70 dbm	−60 dbm	−50 dbm
	c/s	$S_m = s_1 + d$	e	a	r	n_s	n_s	n_s	n_s
1	250	67.3	19	34.5	−31.5	21	21	21	22
2	380	68	20.5	32	−27.5	15	15.5	16	21
3	540	66.5	21	31	−27	11	12	14.5	21.5
4	700	62	22.5	30	−26	6.5	7.5	14	23
5	880	58	26.7	30	−26	3	5.5	13	23
6	1060	55	31	30	−26	1	4.5	13	23
7	1280	52.5	35	30	−26	9.5	10.5	13.7	23
8	1530	50.3	34	30	−24	16	16.5	18.5	26
9	1840	48.5	35.3	30	−26	5.5	7.5	13	23
10	2220	47	34	30	−31	−4.5	−0.5	7.5	18
11	2650	46	29.7	30	−29.5	−4	1.5	9.5	19
12	3200	43.5	23.5	30	−31.5	−2.5	0	8	18.5
	Total	664.6	332.2	367.5	−332.0	77.5	101.5	161.7	261.0
	Total/20	$S_m = 33.2$	$E = 16.6$	$A = 18.4$	$R = -16.6$	$N = 3.9$	$N = 5.0$	$N = 8.1$	$N = 13.0$

All the numbers which appear in this table (except for the first two columns) are in decibels.

According to the assumptions made, E is positive and R negative (see Volume V of the *Red Book*, page 615); A is positive since it is an attenuation and not a gain. Moreover, we have $\Gamma = 2.5 \times 0.6 = 1.5$ db according to formula (22).

Moreover, S_m corresponds to a mean pressure of 11 dynes/cm² ([13], para. 5.2.1) hence $d = 21 + 9.5 = 30.5$ db at any frequency (9.5 db being the nominal acoustic attenuation of A.R.A.E.N. at low frequencies) and $D = 0.6\,d = 18.3$ db.

We therefore obtain by addition, according to formula (21):
$J = 33.2 + 16.6 - 18.4 - 16.6 - 8.1 - 1.5 = 5.2$ db, whereas the value calculated in Appendix 1 for the complete connection is $J = 5.7$ db.

Table 2 gives a comparison between the values of J obtained by addition and directly for all the noise values considered in the table and for three line attenuations.

TABLE 2

Comparison between the values in db of J obtained by adding the individual ratings and the values (in brackets) calculated for complete connections in Appendix 1

l (dbm)	Line attenuation (db)		
	10	20	30
$-\infty$	21.4 (21.9)	15.4 (16.0)	9.4 (9.7)
-70	20.3 (20.6)	14.3 (14.3)	8.3 (8.45)
-60	17.2 (17.35)	11.2 (11.2)	5.2 (5.7)
-50	12.3 (12.0)	6.3 (6.6)	0.3 (2.14)

It can be seen that the error of addition is not more than 0.6 db and is therefore less than the sum of errors to be expected in the numerical calculations or the measurements—except for a condition corresponding to an inacceptable transmission quality. In actual fact, the minimum transmission quality accepted at present by the C.C.I.T.T. is attained, for the telephone sets considered, in the following situation:

a line attenuation of 30 db, i.e. a total reference equivalent of 36 db (*Red Book*, Volume V, page 583);

$l = -60$ dbm, a value obtained by assuming in Figure 6a that $a_1 = a_2 = a_3 = 10$ db (to satisfy the reference equivalent limits), $l_0 = -43$ dbm0 (objective recommended for a world-wide connection), $l_n = -53$ dbm (or 5000 pW, a value rather on the high side). For this situation we have the following information ratings:

Local sending system $D + E$	=	34.9 db
Attenuation of the two-wire circuits and the terminating set: $10 + 3.5$ db $\times 0.6$	=	$-$ 8.1
National sending system		+27 db
Local receiving system $R - N$	=	-24.7
Two-wire circuits and terminating set		$-$ 8.1
National receiving system		-33 db

If it is made a condition, in planning national networks, of not falling below the above mentioned limits for partial information ratings and if the noise limits are respected [1], the information rating of any connection will be equal to, or higher than, the limit if the mean speech power for all the subscribers is that which was assumed; this limiting value of J is equal to about 5 to 6 db.

The information rating may therefore be used for planning networks on the basis of limiting over-all values, which can be apportioned among the component parts of a network. The value of $R - N$ considered in this example assumes that the subscriber line is of a given length. For other lengths, it is not sufficient to allow for the variation in sensitivity; the sensitivity of the sidetone path must also be re-measured and a check must be made to see whether it leads to an increase in N.

3.3 *Application to international telephony*

Everything that has just been said about the example in paragraph 3.2 can be extended to international telephony, the transmitted frequency band being specified.

If the nominal band from 300 to 3400 c/s were transmitted perfectly, a factor $U\,(3400) - U\,(300) = 0.57$ would have to be introduced. By comparison with the case mentioned in paragraph 3.2, it may be expected that a factor of about 0.55 will be found in practice (or an information rating of 5.5 db per 10 db of distortion less gain over this band). To calculate J, we must then divide by 20 the sum of values calculated at 11 frequencies with equal spacing between 300 and 3400 c/s on the abscissa scale of Figure 7; frequencies Nos. 2 to 12 of Table 1 reasonably satisfy this condition [2].

The study of Question 1/XV should help to specify the band transmitted in a worldwide connection which may include 3-kc/s channel equipment and improved 4-kc/s equipment.

3.4 *Measurement methods*

If the sensitivity e of the sending system is determined in the free acoustic field, d corresponds to the ratio between the pressure at the position of the microphone (in its absence) and the pressure at a point 1 metre from the lips; we then have, substantially, regardless of the frequency:

$$d = 20 \log_{10} \frac{100.6}{\delta + 0.6}$$

δ being the distance (in cm) between the lips of the speaker and the diaphragm of the microphone.

It is also possible to determine $d + e$ directly, at each frequency, with an artificial mouth which has been calibrated at a distance of 1 metre.

During these measurements all due precautions should be taken to ensure that a carbon microphone is in the same operating conditions as when it is excited by the voice.

For the receiving system, a semi-subjective method could be applied to determine x directly. The measurement method would be similar to that described for the application of the " tonality " method (*Red Book*, Volume V, page 520) although the subsequent calculations would be different.

Once an artificial ear has been defined by the C.C.I.T.T., it will be possible to use it to measure r and the noise transmitted by sidetone. Measurement of the noise transmitted between the ear and the receiver earcap is a more delicate matter ([13], section 4). It will then be possible to calculate the masking effect of the various noises on a typical hearing threshold curve.

Instead of measuring the sensitivities at various frequencies and of weighting them by calculation, it would be possible to design apparatus similar to the apparatus for the objective measurement of the reference equivalent, which sweeps the transmitted frequency band in accordance with this weighting law. The instrument should then give an indication proportional to the logarithm of the voltage measured at each frequency (see for example the *Book of Annexes* to Volume IV of

[1] For these conditions, it is impossible to encounter at the same time a value of l higher than -60 dbm and attenuations near the limiting values.

[2] If one wishes to change the convention made with regard to formula (14) in such a way that J is equal to the distortion less gain in the band of 300-3400 c/s, it suffices to take the average of the values for these 11 frequencies or, more simply, for the following nine frequencies: 400, 600, 800, 1000, 1250, 1600, 2000, 2500 and 3150 c/s.

the C.C.I.F. *Green Book*, page 84), or be preceded by a logarithmic amplifier (see for example Volume V of the *Red Book*, pages 502 to 503). Use of such an apparatus to measure N can be contemplated, but it should then include a generator to introduce (at least at the high frequencies) an artificial noise having the spectrum curve of the hearing threshold; furthermore, the introduction of the noise between the receiver earcap and the ear would have to be simulated.

4. *Comparison with existing theories or results*

4.1 *Calculation of the information*

The direct application of formulae (1) or (2) to the calculation of the capacity of a telephone channel leads to a well-known paradox [9].

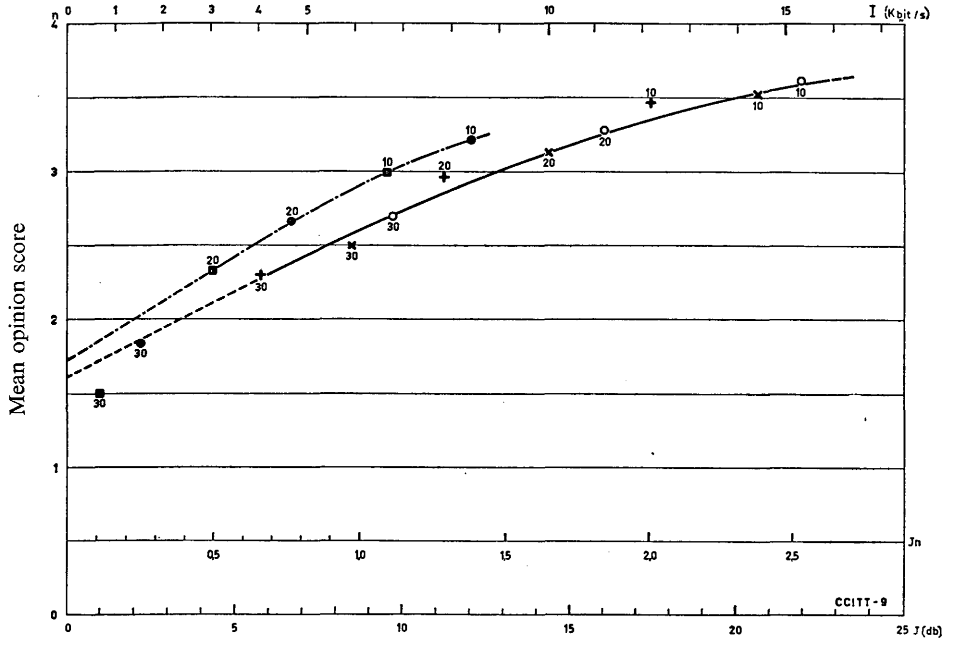
a) If a calculation is made of the information corresponding to the meaning of the phonemes transmitted, we find a very low value of the order of 50 bits per second. This is confirmed by experience [10]. In fact, such a method of calculation is tantamount to considering as equal the quantities of information transmitted by a telephone channel and a telegraph channel which transmit the same words. Such a point of view may be useful in the study of " vocoders " or similar devices but it is very far from the concept of " telephone transmission performance ".

b) If a calculation is made of the information corresponding to the physical characteristics of the signals transmitted, we find a value which is expressed in kilobits per second. We know [11] that this rate of information varies in the same direction as the transmission performance, evaluated by the customary processes; nevertheless, the relation between these quantities has not yet been established. Moreover, the calculation of the rate of information presented certain difficulties. Formula (1) does not take into account distortion in the transmission system or variations in the sensitivity of the auditory system with frequency. To arrive at a more accurate calculation, account has been taken of the differential sensitivity of the auditory system, that is to say of the smallest differences discernible in the amplitude or the frequency [12] but these differences are determined in laboratory experiments of fairly long and ill-defined duration, with the result that we find ourselves very far from the normal conditions for the flow of information in a telephone call. For the sake of comparison, we have found in para. 2.3 that the optimum spacing Δf of the frequencies was equal to two-thirds of the critical bandwidth, that is to say about 13 times the minimum frequency difference which is perceptible ([4], p. 171). For $Q = 10$ to 20, $\frac{\Delta f}{f}$ is of the order of a semi-tone to a quarter-tone. Moreover, the calculation methods employed hitherto did not allow for the masking effect between different components of the transmitted signal.

4.2 *Relation between the information rating and the results of opinion tests*

The information rating was calculated for a system formed of two commercial sets (sending and receiving) of the type whose objective characteristics were measured by the United Kingdom Administration, for various junction losses and circuit noises. Assuming that the same type of set was used in the opinion tests made by that Administration, we have shown on Figure 8 the mean opinion score (according to Figure 3b, page 585 of Volume V of the C.C.I.T.T. *Red Book*) as a function of the information rating calculated for the same conditions.

If the correlation between these two variables is perfect, all the points must lie on the same curve. We have plotted in Figure 8 a logistic curve (with a simple equation) which, on the whole, passes close to the points corresponding to moderate noise levels (-60 dbm or lower). The point situated on the extreme left on the full line portion of this curve corresponds to the noise and reference equivalent limits recommended by the C.C.I.T.T., as indicated in para. 3.2. The four points corresponding to higher noise levels with a moderate line attenuation (10 or 20 db), are all above this curve and are very well aligned on another logistic curve (chain-dotted); it is probable



$$n = \frac{1}{0.25 + \exp\left(-J_N - 1.0\right)} \qquad n = \frac{1}{0.25 + \exp\left(-1.25\, J_N - 1.1\right)}$$

Level of circuit noise (dbm): ○ $-\infty$ × -70 + -60 ● -50 ■ -45

(The number written next to each point is line attenuation, in db)

FIGURE 8

that for these points the speech power, in the opinion tests, was higher than the general average. It can, moreover, be seen from Figure 6a that a quite exceptional combination of attenuations is required to obtain such values of I, if C.C.I.T.T. objectives for circuit noise are reached.

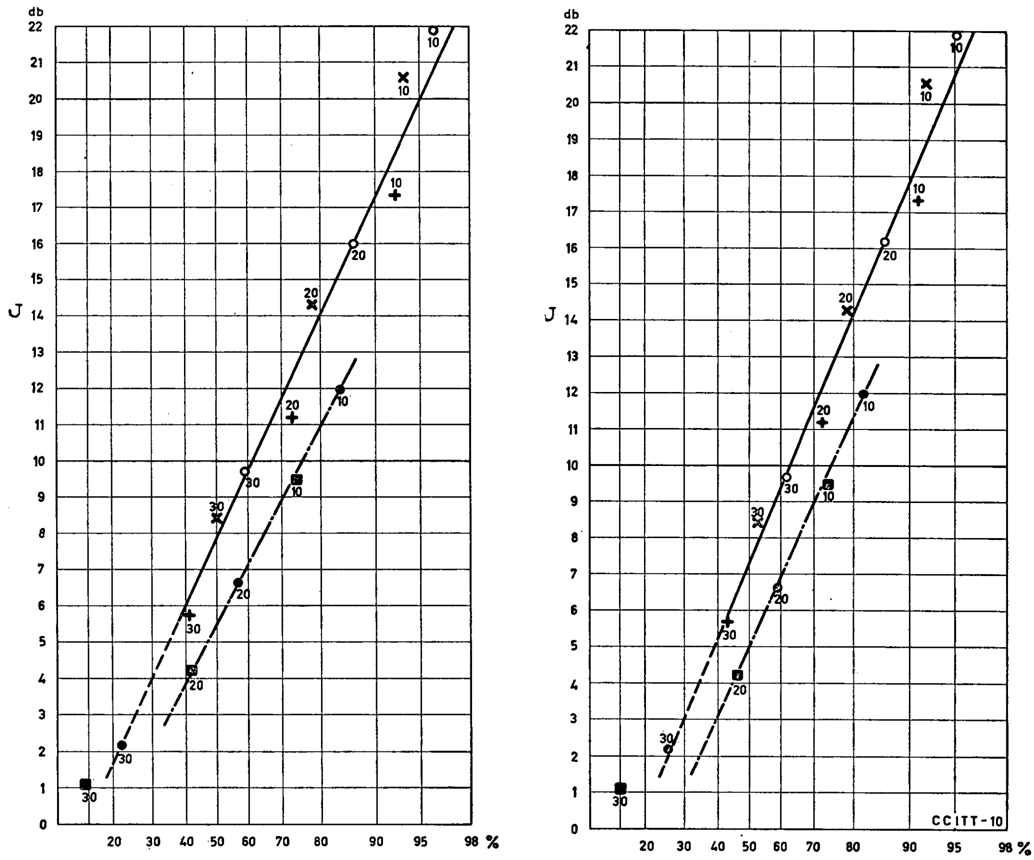
On the contrary, the points corresponding to an attenuation of 30 db and to high noise levels are situated on the extension of the full line curve, or below it. In these conditions, it is probable that the necessary effort to increase the speech power becomes conscious and that users are inclined to consider the connection hardly satisfactory. In any case, these points correspond to a transmission performance inferior to that resulting from existing C.C.I.T.T. recommendations.

We are therefore led to the conclusion that, according to the data assembled in one country for one type of telephone set, there is a good correlation between the information rating and the mean opinion score when the mean sending speech power is known. This result is all the more remarkable in that the information rating has been calculated for a one-way transmission path and that the opinion tests are carried out for two-way calls. However, account has been taken only of factors (attenuation and noise) which obviously affect transmission, and even then in a single direction. The effect of factors that are essentially related to bilateral conservation (such as group delay and echo) would require another study.

If this result is verified for other types of telephone sets, the information rating will provide a common and fixed basis for comparing the results of opinion tests carried out in different countries on different sets. It will also enable the evolution of subscribers' opinions in any one country to be studied, when the transmission performance normally provided in the telephone network is improved over the years.

The abscissae in Figure 8 show the values of J calculated for a complete connection; a sufficient degree of accuracy would be obtained in practice if J were calculated by adding the ratings of the various parts of the connection, as indicated in para. 3.2. If, then, the planning method described in that paragraph is applied to the same telephone sets as those considered here, the J values will give a true picture of transmission performance (judged according to opinion tests) for all connections in which the subscribers do not need to raise their voices to overcome an exceptionally high noise level.

Since the information rating has to be calculated for a given speech power, it seemed reasonable to study its correlation with the mean opinion score, but the latter is linked by known relations (empirical or theoretical) to the percentages of subjects which express certain opinions (C.C.I.T.T. *Red Book*, Volume V, pp. 589, 590, 599). It is thus possible to calculate the "percentage of satisfied users" which corresponds to each information rating. The results of these calculations are given in Figure 9.



(a) (b)

FIGURE 9. — **Percentage of users expressing the opinion "excellent" or "good", as a function of the information rating**

Relationship between the mean score and the percentages according to

a) the results of experiments carried out in the United Kingdom (*Red Book*, Volume V, page 589).

b) a binominal distribution of opinions (*ibid.*, page 599).

4.3 *Comparison between the present theory and the methods for calculating articulation or the reference equivalent*

A detailed comparison is given in Appendix 2, from which it can be deduced that the information rating is zero at the same time as the articulation and, in general, increases with the transmission performance, assessed by conventional criteria. This conclusion is confirmed by the results of para. 4.2.

5. *Conclusions*

5.1 *Results achieved*

We defined an " information rating " possessing the following properties:

a) It is proportional to the flow of information transmitted by the electroacoustic system under study, placed between a signal generator and receiver having certain statistical properties of the human voice and the auditory system. It is thus possible, without ambiguity, to calculate it when the physical characteristics of the electroacoustic system are known.

b) For one type of telephone set used in one country, the information rating shows a good correlation with the results of opinion tests and can thus be regarded as a good criterion of the transmission performance which could be determined in service conditions. It would be desirable to check whether the same applies for the other sets which have been submitted to opinion tests.

c) The relations found between the information rating and various quantities characterizing the articulation or reference equivalent show that, generally speaking, the information rating varies in the same direction as the transmission performance assessed according to conventional criteria —each of the latter being used in its known sphere of validity.

d) By calculating the information rating for a number of loss, noise and attenuation distortion values, we can determine with a fair degree of accuracy how it is possible to " exchange " one of these factors against another while the transmission performance of the system concerned remains constant. We can also exactly assess the effect on the transmission performance of simultaneous variations in two of these factors or in the three. Provided that the results indicated above in b) are confirmed, this method is thus more general and more reliable than the empirical method of " transmission impairments ".

e) Since the calculation based on the information theory assumes the use of an optimum coding, the result characterizes the transmission system and in no way depends on the language used, nor on the various conventions required to define the articulation and the reference equivalent. Thus the information rating seems particularly valuable for characterizing transmission performance in the international telephone service.

f) We have given, for a sending system, a line or a receiving system considered independently, information ratings which can be added together to give the rating of the complete connection with an accuracy that is adequate in practice. It is therefore possible to use the information rating for the planning of telephone networks.

5.2 *Further studies*

a) It would be advantageous to have reliable, generally agreed values for the physiological data used: power spectrum of the human voice, audibility thresholds, critical bandwidths, signal identification time. Comments from Members of Study Group XII on the provisional values adopted here will be very helpful.

b) It would be extremely helpful if every Administration which has conducted opinion tests would measure the objective characteristics of the telephone apparatus used in those tests and make calculations on the lines of the example given in Appendix 1. In this way a general relationship could be established between the information rating and the results of opinion tests.

5.3 *Possible generalizations*

a) The calculation was developed for the ease of listening with a single ear by means of a telephone receiver. Obviously the theory would apply equally in the case of free air listening, with both ears; it would suffice, in this event, to modify some of the physiological data.

b) A noise component at one frequency may be considered equivalent to a continuous noise, spread over the critical bandwidth including this frequency. Thus it is possible to calculate the effect of steady noise on transmission performance, regardless of the nature of that noise (continuous spectrum noise, rectifier noise, etc.). For the information rating one might develop a method similar to the one described for loudness in article [23].

c) When we have completely clarified the question of the time required for the hearing system to identify the sounds received, it must be possible to generalize the theory to include impulsive noises or other transients.

d) Non-linear distortion causes a number of unwanted components to appear, which depend on the level of the wanted components of the signal. The theory remains applicable provided we know how to calculate the most probable value of the various unwanted components that may affect the reception of the wanted components at each frequency—a calculation which in itself presents serious difficulties.

APPENDIX 1

Example of the calculation of J

J as been calculated for commercial sets whose characteristics are given in [13] and for various line attenuation values (with a 300-3400 c/s filter) and circuit noise levels, with a room noise of 50 db.

Table 3 shows the method of calculating x.

Columns in Table 3:

(1) number attributed to that frequency in [13]
(3) taken from column C8 of Table 15 of [13] and diminished by the line attenuation (here 30 db)
(4) n_s is the threshold of hearing in the presence of room noise, taken from B10 of Table 14 [1]
(5) = (3)−(4)
(6) = (9)−10 db
(7) power addition of (4) and (6)
(8) = (3)−(7)
(9) taken from B13 of Table 14
(10) power addition of (4) and (9); already calculated in B14 of Table 14 for −60 dbm
(11) = (3)−(10), etc.

For each value of x, the value of z (read on Figure 5) is plotted on the graph in Figure 10 where the scale of the abscissae is proportional to $U(f)$ (scale of the ordinates in Figure 4). The same calculations were made for two other line attenuation values.

[1] All the tables quoted are in reference [13].

TABLE 3

Calculation of x for determination of J

Line loss 20 db

Frequency		Speech sounds at the entry to the ear	Noise level of circuit at the input of the receiver system (dbm)													
			−∞		−70			−60			−50			−45		
No.	Value		n_s	x	l_s	n_s	x	l_s	n_s	x	l_s	n_s	x	l_s	n_s	x
(1)	(2)	(3)	(4)	5)	(6)	(7)	(8)	(9)	(10)	(11)	(12)	(13)	(14)	(15)	(16)	(17)
	c/s	db	db													
1	98	−10	33	−43	−29	33	−43	−19	33	−43	−9	33	−43	−4	33	−43
2	236	18	23	−5	−9	23	−5	1	23	−5	11	23	−5	16	23.8	−5.8
3	372	28	16	12	−1	16	12	9	17	11	19	20.75	7.25	24	24.5	3.5
4	510	30	12	18	1	12.5	17.5	11	14.5	15.5	21	21.5	8.5	26	26	4
5	646	29	7	22	3	8.5	20.5	13	14	15	23	23	6	28	28	1
6	784	29	4	25	3	6.5	22.5	13	13.5	15.5	23	23	6	28	28	1
7	926	29	2	27	3	5.5	23.5	13	13	16	23	23	6	28	28	1
8	1083	30	1	29	3	5	25	13	13	17	23	23	7	28	28	2
9	1272	32	9	23	3	10	22	13	13.5	18.5	23	23	9	28	28	4
10	1493	31	17	14	5	17	14	15	19	12	25	25.5	6.5	30	30	1
11	1752	30	10	20	5	11	19	15	16	14	25	25	5	30	30	0
12	2055	24	−2	26	0	2	22	10	10	14	20	20	4	25	25	−1
13	2411	17	−6	23	−3	−1	18	7	7	10	17	17	0	22	22	−5
14	2830	15	−1	16	1	3	12	11	11	4	21	21	−6	26	26	−11
15	3319	2	−5	7	−4	−1.5	3.5	6	6.5	−4.5	16	16	−14	21	21	−19

By graphical integration of the curves of Figure 10, the values of J indicated in Table 4 are obtained.

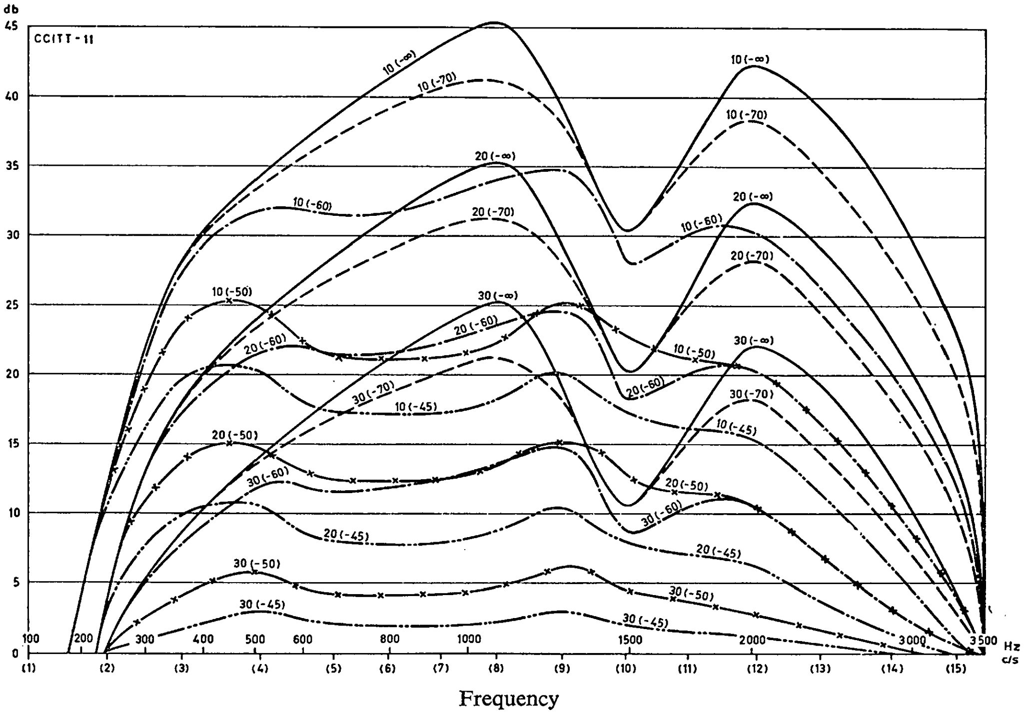


FIGURE 10

TABLE 4

Values of J (in db) for various values of the line attenuation a (in db) and of the noise level l (in dbm)

a \ l	$-\infty$	-70	-60	-50	-45
10	21.9	20.6	17.35	12.0	9.5
20	16.0	14.3	11.2	6.6	4.25
30	9.67	8.45	5.7	2.14	0.99

APPENDIX 2

Comparison with earlier theories for the calculation of articulation, loudness and reference equivalent [1]

1. *General*

Articulation is calculated through the intermediary of a quantity A, such as the band articulation ([13], [14], [15], [16]) or the articulation index [5] which possesses the following two properties:

a) it is a single valued function of the articulation (for sounds, logatoms, etc.) measured in specific conditions;

b) it can be calculated, for a system whose physical characteristics are known, by a relation of the form

$$A = \int_B U'(f)\, W(x)\, \mathrm{d}f \tag{31}$$

where x is either the signal/noise ratio or the level of the signal above the hearing threshold in the absence of noise, which corresponds to the definition of this variable given above. *By convention*, $W(x)$ varies from 0 to 1 when x ranges over the whole of the variation interval which appreciably affects the articulation and, if we consider

$$U(f) = \int_0^f U'(f)\, \mathrm{d}f$$

its maximum value (extended to the whole audible frequency band) is equal to 1. Hence, A varies from 0 to 1 when the articulation varies from 0 to 100%.

A similar formula is used for calculating the reference equivalent [17].

Now, formula (31) is of the form (16). Hence a comparison can be made of the functions of f and the functions of x corresponding to the different theories.

2. *Weighting functions as a function of the frequency*

Figure 11 represents various integral weighting functions $U(f)$ which were proposed:

1) in section 2.4 of this annex;

2) by *Richards* and *Archbold*, from consideranda bearing on the information theory, for articulation calculations;

3) by *French* and *Steinberg*, from articulation tests;

4) and 5) by the Italian Administration based on a theory of hearing. Curve (4) is applicable to articulation calculations and curve (5) to calculations of the reference equivalent.

We find that all these curves have the same general trend. Each may be represented fairly exactly by two straight-line segments, with the frequency scale adopted on Figure 11 (linear from 0 to 1000 c/s and logarithmic from 1000 to 10 000 c/s). The differences found seem to arise largely from the fact that, in certain curves, the influence of frequencies below or above certain more or less arbitrary limits [2] have been ignored. Since such frequencies always undergo serious attenuation in certain parts of the transmission system, these differences do not seem important in practice for telephony.

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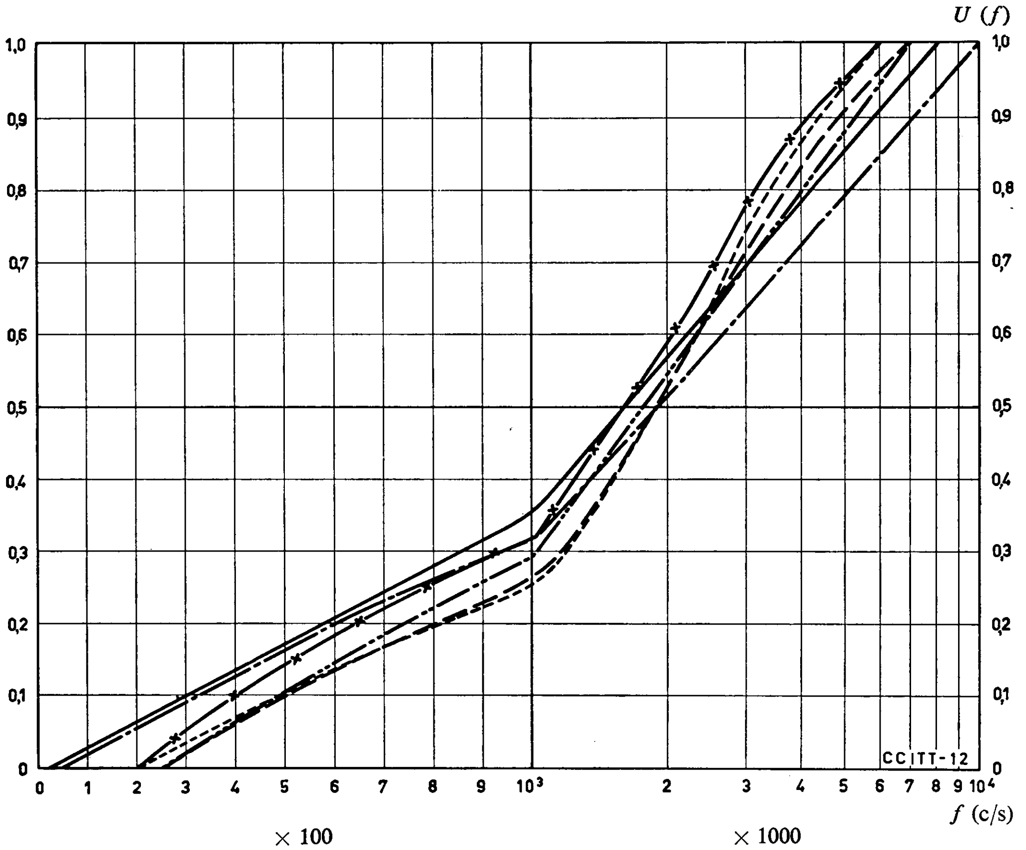
[1] Some symbols used in this appendix have different meanings from those used in the body of the Annex.

[2] For example, in Figure 11, curve (6) was calculated by limiting the ordinates of curve (1) to the 250-7000-c/s band and bringing the value for 7000 c/s up to 1.0. This curve can be compared with curve (3), which is restricted to the same band.



(1) —·—·—·— proposed curve

(2) ———————— *Richards and Archbold* [13]

(3) — — — — *French and Steinberg* [5]

(4) - - - - - - - $\int \frac{1}{\Delta f} \mathrm{d}f$ [18], [19]

(5) —×—×— $\int \frac{K}{\Delta f} \mathrm{d}f$ [18], [19]

(6) —··—··— (1) restricted to 250-7000 c/s

FIGURE 11. — **Integral weighting curves as a function of frequency**

The weighting curves used by *Fletcher* and *Galt* [20] for the calculation of articulation and by *Braun* [17] for the calculation of the reference equivalent implicitly take into account other factors and are not directly comparable to the preceding curves.

3. *W(x) function used for articulation calculations*

The " W factor " of French and Steinberg is expressed as a function of a quantity $H = B_s + p + K - \beta_0$ ([5], equation (11)). With our notations:

$$B_s = 10 \log_{10} \frac{S_s}{P_0}$$

β_0 is the threshold of hearing for a pure tone, so in the absence of noise,

$$\beta_0 - K = 10 \log_{10} \frac{N_s}{P_0}$$

p is a peak factor equal to 12 db.

Thus we must take $x = H - 12$ db

The function P defined by *Richards and Archbold* [13] and the perception coefficient P_K, used by the U.S.S.R. Administration [21], play the same role as the W factor. However, these quantities are not expressed as a function of the mean speech power; we find (see Figure 12) that they are very near if certain empirical corrections are made on the abscissae of their representative curves. Henceforward, only the W factor will be considered. Figure 13 represents the relation between W and z, obtained from Figures 5 and 12 by eliminating x.

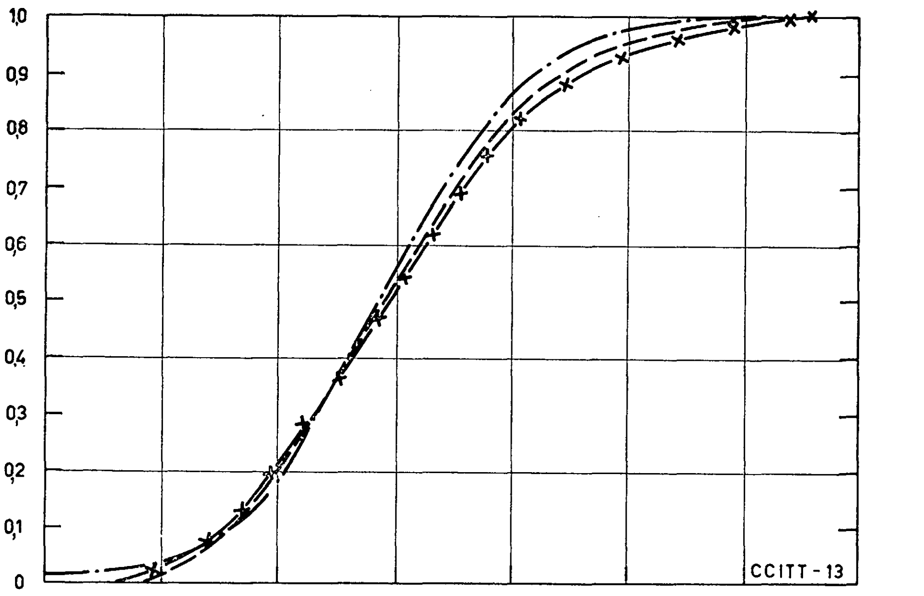


FIGURE 12. — **Variation, as a function of signal-to-noise ratio x, of various quantities used for articulation calculations**

— — — — W factor after *French and Steinberg* [5] (12 db subtracted from the abscissae)

—·—·—·— P after *Richards and Archbold* [13] (15 db added to the abscissae)

— × — × — Perception coefficient P_K, after the U.S.S.R. Administration [21] (14 db subtracted from the abscissae)

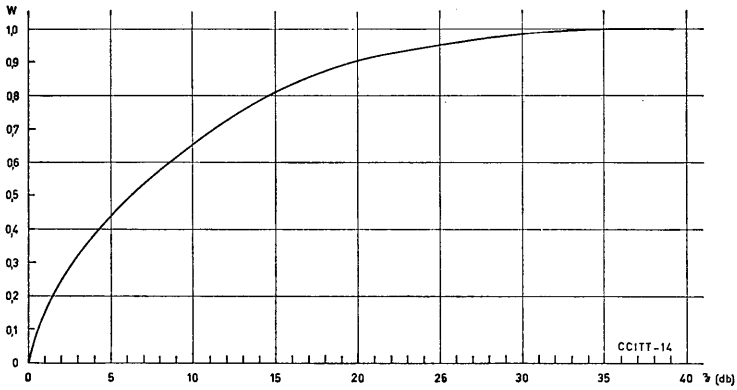


FIGURE 13

4. *Functions of x used for the calculation of loudness and the reference equivalent*

4.1 *Loudness*

For $x \geqslant 40$ db, this is given by the expression

$$V(x) = 10^{\frac{\nu}{10}(x-40)} \tag{32}$$

or in decibels

$$v(x) = \nu (x - 40) \tag{33}$$

the value of ν varying between 0.25 and 0.50, depending on the authors[1]. On Figure 14 we have shown $V(z)$ according to Figure 5 and according to the variation curve of V as a function of x after *Fletcher* ([4], p. 193), which corresponds to $\nu = 0.33$.

4.2 *Objective reference equivalent*

Braun has defined [22] by the formula

$$q = \frac{1}{2\nu} \ln \frac{u_1^{2\nu} + \ldots + (u_i^{2\nu} + \ldots\ldots + u_m^{2\nu}}{(u_1 e^{-b_1})^{2\nu} + \ldots + u_i e^{-b_i})^{2\nu} + \ldots\ldots +} \text{ nepers} \tag{34}$$

a quantity which we shall call " objective reference equivalent " of a line if the u_i are electric voltages and which would be applicable to an electroacoustic transducer with the introduction, if necessary, of acoustic pressures; in this formula, b_i is an attenuation, at frequency f_i, expressed in nepers.

The numerator of this expression represents a complex voltage measured with a rectifier apparatus with an exponent 2ν. If 2ν is suitably selected (twice the value of ν indicated in the paragraph above and corresponding to measurements of power) and if the level is sufficiently high, the loudness of the human voice will thereby be measured at the line input, provided that the u_i correspond to the mean speech spectrum and that the f_i are so spaced as to obtain the required weighting law as a function of frequency.

The denominator is the same quantity, measured at the line output. The factor $\frac{1}{2\nu}$ was chosen so as to find $q = b$, as in the case of the reference equivalent proper, if all the b_i have the same value and are equal to b.

[1] Strictly speaking, this expression is valid only at 1000 c/s, but the corrective term may be incorporated in the weighting factor as a function of the frequency.

Attenuation relative to S.F.E.R.T.

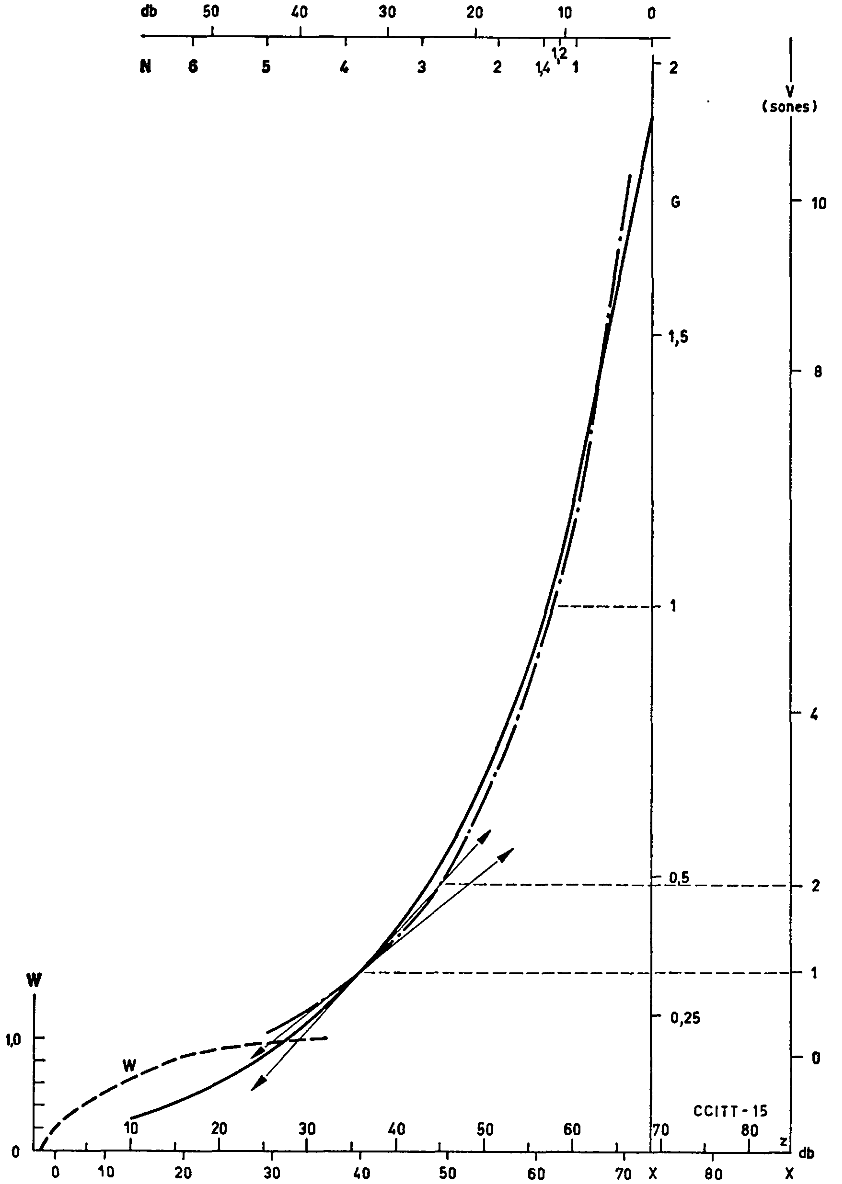


FIGURE 14. — **Contribution of a narrow band of frequencies**

—·—·—·— V to loudness
————— G to reference equivalent
— — — — W to articulation index

Measurements made on the basis of this formula gave a value [22] when the human ear was used as the measurement instrument of $2\nu = 0.66$, i.e. $\nu = 0.33$. The function $V(x)$ is thus the same as in the preceding section.

4.3 *Reference equivalent*

The reference equivalent proper is defined as a loss which, when inserted in the line of a reference system, would reduce the loudness to the level of that at the output of the system measured; the corresponding loss in a narrow frequency band is then $x_0 - x$, x_0 being the value of x corresponding, in that band, to the zero reference equivalent.

Braun [17] has shown that the reference equivalent could be calculated from the " excitation " (Erregung) function G whose values he measured as a function of what we have called $x_0 - x$. Having determined x_0 as shown in the next section, the curve in Figure 14 shows the variation of G with x and with z. It can be seen that this curve is very close to the representative curve of V, when the scales for the ordinates are so chosen as to make two points of these curves coincide.

4.4 *Calculation of* x_0

The pressure on the A.R.A.E.N. microphone is 1 dyne/cm² (i.e., 74 db in relation to 2.10⁻⁴ dyne/cm²) for the "reference speech power for A.R.A.E.N.", that is a level of 70 db for the "normal speech power for telephonometric measurements ". This pressure is deduced from the indications of the A.R.A.E.N. speech voltmeter; the average pressure is a little lower, the correction term is small for the conventional sentence used in measuring the reference equivalent and is probably of the order of 3 db. To obtain the value of S_s at this point the ordinate in Figure 2 of [5] must therefore be increased by 2 db. At each frequency the loss in A.R.A.E.N. with 30 db in line (Recommendation P.41) must be subtracted from this and 34 db (the reference equivalent for this system—technical reports Nos. 220 and 257 of the C.C.I.T.T. Laboratory) must be added in order to obtain S_s corresponding to the zero reference equivalent. N_s is given directly by Figure 5 of [5].

It may be seen that, for the most important frequencies x_0 is nearly constant and is in the region of 73 db.

Frequency (c/s)	100	200	300	500	800	1000	2000	3000	4000
Spectrum power at 1 m from the mouth db/$2 \cdot 10^{-4}$ dyne/cm²	32	35	36.5	36.5	32	29	19	13	9
id. augmented by 36 db	68	71	72.5	72.5	68	65	55	49	45
Loss in A.R.A.E.N. (db)	9.5	9.5	9.5	9.5	9	8.5	5	2	2
$(S_s)0$ db/$2 \cdot 10^{-4}$ dyne/cm²	58.5	61.5	63	63	59	56.5	50	47	43
N_s db/$2 \cdot 10^{-4}$ dyne/cm²	+20	+7	0	−10	−15	−17	−23	−29	−29
x_0 (db)	38.5	54.5	63	73	74	73.5	73	76	72

5. *Comparison of various functions of x*

When x and z are varied, beginning with very low values, the following observations can be made from Figures 13 and 14.

For $z = 0$, $W = 0$; thus, the articulation is nil. Then, with x and z increasing W increases regularly. There comes a time when W increases more and more slowly and ends by tending towards 1, while z continues to show a steady increase. This is a well-known effect: when the speech sounds received have reached a certain level, practically no further gains are made, from the standpoint of articulation, by raising this level further, although in actual fact the transmission performance continues to improve.

With x and z continuing to increase, a zone is entered in which the reference equivalent presents a more accurate picture of the transmission performance than A.E.N.; it is seen that the loudness and the excitation increase steadily with z when x is the level above the hearing threshold (in the absence of noise). Furthermore, when $x_0 - x$ lies between about 20 and 40 db (which corresponds to the usual reference equivalents for a complete system), the curves representing V and G remain fairly close to their tangents at a mean point taken as reference (see Figure 14). This explains why, in this region, it is possible, approximately, to add the reference equivalents expressed as decibels or nepers.

If there is an appreciable noise, z is always defined (x then being the signal to noise ratio), whereas the reference equivalent is no longer sufficient to characterize the transmission performance.

To summarize, if the transmission performance is assessed on the basis of conventional criteria, each applied within its own sphere of validity as determined through experience, it is found that the function of x which characterizes the quality of transmission in a narrow frequency band always varies in the same direction as z. This conclusion has a fairly general application, since W and G were determined with different telephone apparatus and with talkers using different languages, while the loudness was assessed under appreciably different listening conditions.

6. *Conclusion*

Since the curves in Figure 14 are not straight lines and since, in each case, account must be taken of attenuation distortion, there is no simple and general relation between the information rating and articulation or the reference equivalent—which, in any case, provide but imperfect assessments of the transmission performance. It may nevertheless be concluded that the information rating is nil at the same time as the articulation and, as a general rule, increase at the same time as the transmission performance.

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B.S.T.J. = *Bell System Technical Journal.*
J.A.S.A. = *Journal of the Acoustical Society of America.*
P.I.E.E. = *Proceedings of the Institution of Electrical Engineers.*
T.F.T. = *Telegraphen-, Fernsprech-, Funk- und Fernseh-Technik.*
Vol. I (Vol. V) = *Volumes of C.C.I.T.T. Red Book.*

Question 8/XII — Measurement of the sensitivity of a carbon microphone

(continuation of Question 8 of Study Group XII, 1961-1964)

The shape of the sensitivity/frequency characteristic of a sending system depends greatly on the artificial mouth used and also upon the method of measurement employed.

Similarly, the shape of the sensitivity/frequency characteristic of a receiving system depends upon the artificial ear used and also upon the method of measurement employed.

What methods of measurement should be recommended for tracing these curves and what accuracy should be recommended for these measurements?

Note 1. — By way of information, the annex below gives the principal experimental conditions which ought to be observed when measuring the "sensitivity-frequency" characteristic of a transmitter system comprising a carbon microphone.

Note 2. — The various methods described in Annexes 27 to 31 (Part II of Volume V of the *Red Book*) should also be applied to telephone systems other than those to which they have already been applied.

Note 3. — Within the framework of the study of Question 12/XII Study Group XII has prepared a programme of tests to be carried out in the C.C.I.T.T. Laboratory with various artificial mouths and ears (see Annex 2 to Question 12/XII).

ANNEX

(to Question 8/XII)

Principal experimental conditions which should be taken into account when determining the sensitivity-frequency characteristic of a sending system having a carbon microphone

(Contribution from the Administration of the German Federal Republic)

(See pages 668 and 669 of Volume V of the *Red Book*.)

Question 9/XII — Limits applied to national trunk and local networks

(continuation of Question 9 of Study Group XII, 1961-1964)

(documentary question)

What are the limits applied by your Administration to the national trunk and local networks of your country in order to ensure satisfactory quality for national calls, it being understood that the recommendation of the C.C.I.T.T. relating to reference equivalents is satisfied for international calls?

Note. — The documentation already received is mentioned in Recommendation P.21 and reproduced in the revised Annex 4 (Part II of this volume).

Question 10/XII — Increase in the sensitivity of local systems

(continuation of Question 10 of Study Group XII, 1961-1964)

(documentary question)

Considering that modern developments have enabled considerable improvements to be made in the sensitivity of telephone sets and that even further increases in sensitivity can readily be achieved, it is desirable to examine the consequences of such increases in sensitivity and the manner in which they may be turned to advantage.

If transmitting and receiving sensitivities of subscribers' apparatus are increased above levels at present prevailing, what will be the consequences in regard to:

a) the subscriber's behaviour;

b) the performance of transmission systems?

Note 1. — In order that the replies to this question may be presented in a uniform way, it is recommended that the speech levels be expressed in v.u. These levels, and the statistical distribution law at present found acceptable for sending, should be established from measurements carried out at local exchanges and/or international exchanges.

Note 2. — It is desirable to study the effect upon volume in service and its distribution when subscribers are provided with telephone sets which are more sensitive than those to which they have become accustomed. By way of information, the transmission features which are liable to be effected by the introduction of more sensitive telephone instruments are given below:

a) non-linear distortion and cross modulation in multichannel transmission systems;
b) crosstalk in cables and lines;
c) inconvenience due to the increased audibility of echo;
d) signal imitation;
e) sidetone and singing due to coupling between transmitter and receiver;
f) room noise impairment.

Note 3. — Annex 1 below deals with the effect which increased sensitivity of subscribers' telephone sets might have on the subscribers' behaviour. Section IV (contributed by the Helsinki Telephone Company) of Annex 4 (Part II of this volume) contains some comments on the use of telephone sets with automatic regulating devices.

Note 4. — Annex 2 gives the results of tests carried out by the Administration of the German Federal Republic.

Note 5. — Study Group XVI, under Question 1/XVI, is studying a single measuring method to check the volume effectively measured in an international exchange at the junction between the national and international networks.

ANNEX 1

(to Question 10/XII)

Effect of the increase in sensitivity of telephone sets on subscribers' behaviour

(See Volume V of the *Red Book*, page 671.)

ANNEX 2

(to Question 10/XII)

Consequences of increased sensitivity of microphone capsules

(Contribution by the Administration of the Federal Republic of Germany)

(See Volume V of the *Red Book*, pages 671-673.)

Question 11/XII — Statistical methods of checking subjective tests

(continuation of Question 11 of Study Group XII, 1961-1964)

(documentary question)

Collection of information concerning the application of statistical methods for the control and direction of subjective tests.

Note 1. — The method of statistical analysis used for the reference equivalent measurements in the C.C.I.T.T. Laboratory is described in Annex 6 (Volume V of the *Red Book* — Part II).

Note 2. — The method of statistical analysis which was used in the 10th series of experiments of the C.C.I.F. Laboratory and which was used to determine the A.E.N. in the C.C.I.T.T. Laboratory is described in Recommendation P.45.

Note 3. — The method of statistical analysis which was used by the C.C.I.F. Laboratory during the 8th and 9th series of experiments is described in pages 84-96 of Volume IV of the C.C.I.F. *Yellow Book* (Paris, 1949).

Note 4. — For the information of Administrations wishing to use the so-called "quantal response" (réponse par échelons), Annex 7 (Volume V of the *Red Book* — Part II) gives the plan and the statistical analysis of loudness efficiency measurements based on these methods.

Question 12/XII — Artificial voices, mouths and ears

(continuation of Question 12 of Study Group XII, 1961-1964)

(documentary question)

What general characteristics and corresponding tolerances should be fixed for artificial voices, mouths and ears?

Note 1. — Pending the standardization of an artificial ear for general use, in collaboration with Technical Committee 29 of the I.E.C., the C.C.I.T.T. has published Recommendation P.51.

Note 2. — Annexes 8 to 16 (Volume V of the *Red Book*, Part II) describe, by way of information, the artificial mouths and ears used by the Administrations or private operating Agencies of Chile, Czechoslovakia, France, the Federal Republic of Germany, Italy, Switzerland, the United Kingdom and the U.S.S.R.

Annex G (Part II of this volume) describes the artificial mouth used by the Swedish Administration.

Note 3. — Annex 1 below describes the systematic method followed by Study Group XII for solving the problem of the artificial ear used in telephone measurements.

Note 4. — Within the framework of the study of this question, Study Group XII has drawn up a programme of work for the C.C.I.T.T. Laboratory concerning the measurements to be made with artificial mouths and ears (see Annex 2 below).

Note 5. — When the essential electroacoustical characteristics of an artificial voice, mouth and ear have been standardized, directives should be drawn up for the objective determination of the "sensitivity-frequency" characteristics (response curve) of subscribers' telephone apparatus using the artificial voice, mouth and ear, so that the study of Question 8/XII can be pursued.

Note 6. — Annex 3 below gives the principal test conditions to be taken into consideration during the measurements with human and artificial mouths and ears.

Note 7. — Annexes 4 and 5 describe two methods for determining the acoustic impedance of human and artificial ears, studied by the Administration of the Federal Republic of Germany.

Note 8. — Annex 6 below reproduces a contribution by the Italian Administration concerning the study of a probe microphone. Annex 7 reproduces a contribution by the Chile Telephone Company relating to the results of preliminary measurements made for the purpose of determining the external physical form to be given to an artificial mouth.

ANNEX 1

(to Question 12/XII)

Systematic method adopted by Study Group XII for solving the problem of the artificial ear

Points for study

1. Determination of representative characteristics for the human ear.

2. Assembly of information about existing artificial ears.

3. Measurement of the characteristics of commercial telephone systems using artificial ears now available.

4. Comparison of results under 3 with the results of measurements made with real ears.

5. Standardization of artificial ears on an international basis.

6. Recommendation of a method for measuring the sensitivity/frequency characteristics of telephone receivers with equipment standardized under 5.

Clearly 1 to 4 inclusive must be completed before it is possible to deal with 5 and 6.

Comments

Comment 1. — Study Group XII asked I.E.C. Technical Committee 29 to supply the available information that had been collected on the acoustical impedance of real ears (under point 1 above). This was supplied and is assembled in COM XII—Nos. 33 and 34 (1961-1964). It would appear that the results show considerable dispersion; moreover, it is noted that I.E.C. Technical Committee 29 is still active studying this question. Study Group XII wishes to be kept informed on the progress made.

Comment 2. — C.C.I.T.T. Study Group XII has assembled a collection of all available artificial ears that are used by telephone Administrations for testing telephone sets. The C.C.I.T.T. Laboratory is in process of measuring the acoustical impedances of these artificial ears. This work is likely to extend at least to the end of 1964 but when completed will enable comparisons to be made with information under point 1.

Comment 3. — The C.C.I.T.T. Laboratory has made preparations to measure the sound pressure under the earcaps of a small number of telephone receivers applied to the various artificial ears available. If possible the frequency band will extend from 100 to 4000 c/s. Also included will be one receiver used for reference equivalent determinations (A.R.A.E.N. receiver).

Comment 4. — Corresponding measurements will be made of the sound pressures from the same receivers applied to the ears of the subjective testing team of operators at the C.C.I.T.T. Laboratory. Comparison will then be possible of the sound pressures in the artificial and real ears but, in addition, the reference equivalents will be calculated from the sound pressure measurements and compared with the reference equivalents measured subjectively in the normal manner.

When the foregoing programme of work, in the C.C.I.T.T. Laboratory as described in Annex 2 (points 1.1 and 1.2), has been completed and reliable information under point 1 is available for comparison with the results obtained under point 2, it will be possible to consider standardizing an artificial ear for measurements on commercial telephone receivers.

C.C.I.T.T. Study Group XII is adopting a similar approach to the problems associated with artificial voices and artificial mouths (see Annex 2—Points 2.1 and 2.2).

ANNEX 2

(to Question 12/XII)

Detailed programme of tests to be undertaken by the C.C.I.T.T. Laboratory under points 2 and 3 mentioned in Annex 1

1. Measurements to be made with artificial ears

1.1 *Measurement of the acoustic impedance (or acoustic load) presented by various artificial ears to telephone receivers* (point 2 of Annex 1, above)

The Working Party renewed its previous recommendation that the C.C.I.T.T. Laboratory should not undertake tests of the acoustic impedance of human ears.

The artificial ears available in the Laboratory are the following: the artificial ears of the Administrations of France, the United Kingdom (A.R.A.E.N. ear), the Federal Republic of Germany, the provisional reference artificial ear recommended by the C.C.I.T.T. (see Recommendation P.51), the reference coupler provisionally specified by the I.E.C. (coupler N.B.S. 9 A).

The measurements will also be carried out with the artificial ears of the Swedish and Swiss Administrations; these Administrations will arrange for these items to be available to the Laboratory in good time.

The range of measuring frequencies should extend, if possible, from 100 to 4000 c/s. To this end, the Federal German Administration will provide the Laboratory with supplementary elements enabling the calibrating tube to be suitably lengthened.

The measurement method, based on purely electrical measurements, adopted by the Working Group, forms the subject of Annex 4 below. A development of this method is contained in contribution COM XII—No. 79 for the period 1961-1964.

1.2 *Measurement of the sound pressure produced by a small number of telephone receivers and one of the A.R.A.E.N. receivers* (point 3 of Annex 1)

The measurements will be carried out with the receiver applied successively to:

— the ear ordinarily used by the operator during loudness balancing tests;

— the particular artificial ear being investigated.

The following artificial ears are available in the C.C.I.T.T. Laboratory:

Ears supplied by the Administrations of France, the Federal Republic of Germany, the United Kingdom (A.R.A.E.N. ear), the provisional reference artificial ear recommended by the C.C.I.T.T. (see Recommendation P.51), the reference coupler provisionally specified by the I.E.C. (coupler N.B.S. 9 A).

The artificial ears of the Swedish and Swiss Administrations will be made available to the Laboratory in good time.

The artificial ear of the Swiss Administration will correspond to the description given in Annex 12 (Part II of Volume V of the *Red Book*).

The Swedish Administration will provide the C.C.I.T.T. Laboratory with the necessary documentation.

These two Administrations will be informed of the date on which the tests will start (with an advance notice of two months).

1.2.1 *Details of the experiments*

The sound pressure will be measured by a tubular probe microphone, the tip of which will be placed near the centre of the receiver earcap and in the plane of the couplers' opening which is considered as the reference plane. The probe microphone is of the Bruël and Kjaer type (microphone with a diameter of ½ inch with a probe 1 mm in diameter, 59 mm long). The sensitivity-frequency characteristic of this probe microphone is given in Figure 1 below.

The measurements will be made at a certain number of frequencies suitably distributed over a frequency range extending, if possible, from 100 to 4000 c/s.

The voltage U_R applied to the receiver terminals will be adjusted so that the sound pressure developed by the receiver is reasonably constant and about 2 dynes/cm^2; the sound intensity level will be about +80 db above $2 \cdot 10^{-4}$ dyne/cm^2, and thus will not be too high and should not inconvenience the operator.

For example, if reference is made to Figures 2 and 3 below, the voltage E will vary with the frequency, the sound pressure developed by the receiver remaining appreciably constant; for example, E will be -40 db/1 volt for the frequency band between 100 and 1500 c/s, -30 db/1 volt for the frequency band 1800-2700 c/s and -20 db/1 volt for frequencies above 2700 c/s in the case of a commercial receiver [1], whereas in the case of an A.R.A.E.N. receiver [1], this voltage E could be kept constant and adjusted to -30 db/1 volt.

[1] The voltage response of a commercial receiver generally depends on the frequency; this is not so with an A.R.A.E.N. receiver if we consider the useful frequency band 100-4000 c/s.

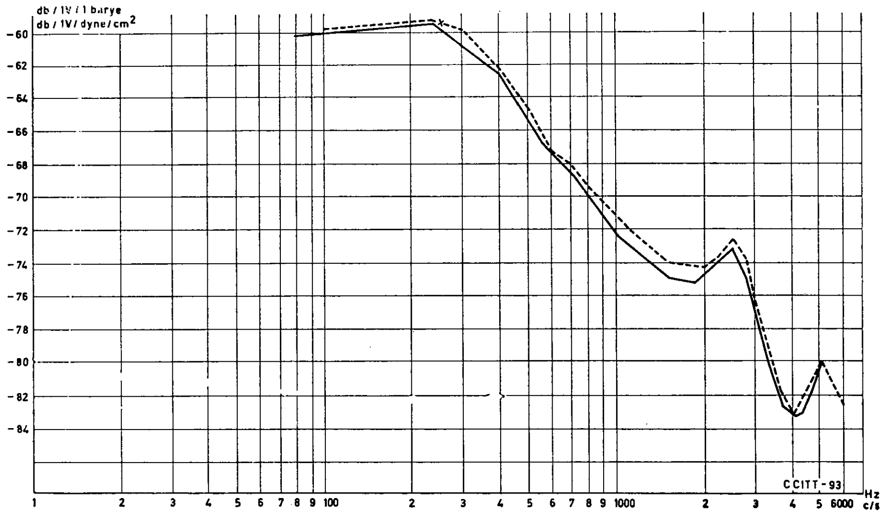


FIGURE 1. — **Sensitivity/frequency characteristic**
(microphone M 032 with a 1-mm probe, 59 mm long)

——————— measured by means of the Rayleigh disc
– – – – – – – measured in a free field
× calibration of the probe by means of the pistonphone (250 c/s)

The axial application force of the receiver on an artificial ear will be 1 kg (10 newtons), including the weight of the handset with its receiver (if the handset with the receiver is excluded the additional applied force is about 750 gr.).

As regards the application of the receiver to the human ear, care should be taken to make it fit the ear as well as possible; for this purpose a warble tone at a very low level will be injected for a few moments instead of the measuring frequency—for example, the warble tone used as a sound source in the apparatus for the objective measurement of reference equivalents used by the Federal German Administration (Annex 28, Part II of Volume V of the *Red Book*).

The circuit arrangement for use with a commercial receiver will be as shown in Figure 2 below.

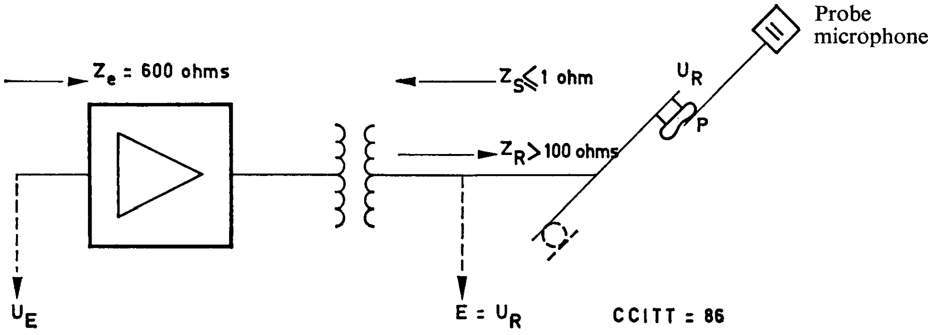


FIGURE 2

The receiver will be considered by itself, and not in conjunction with the telephone set. The amplifier gain will be adjusted to a suitable figure, so that the sound pressure developed by the receiver remains within acceptable limits (see above). Further, it is arranged that subjective measurements will be made by comparison with one N.O.S.F.E.R. channel. The reference system having a definition other than that normally considered will first be examined and the circuit shown in Figure 3 below will be set up.

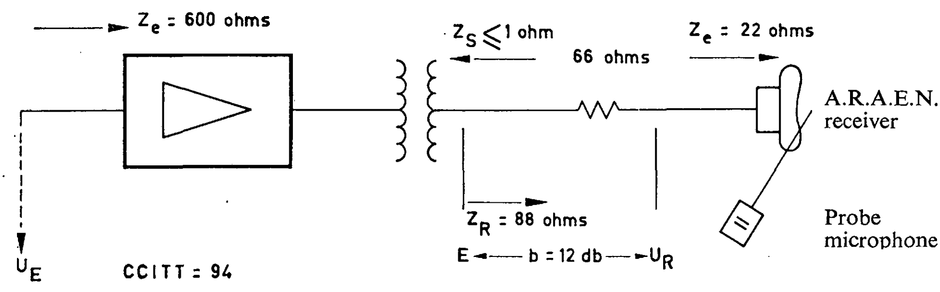
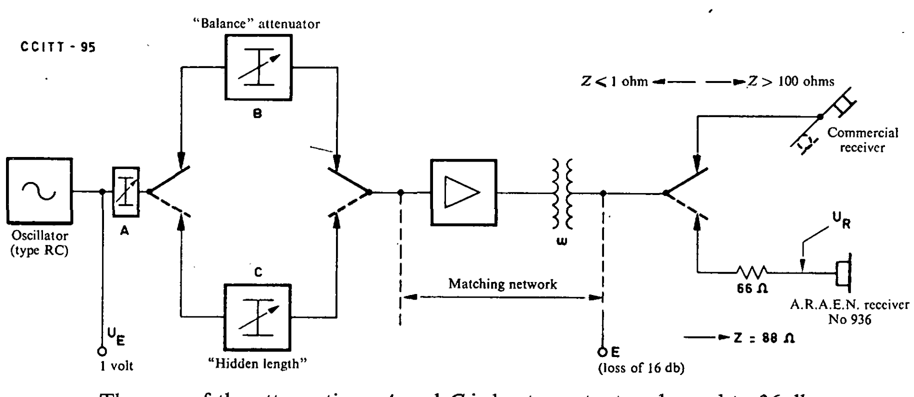


FIGURE 3

In this chain, use is made of the same elements as in the case of a commercial receiver.

Thus, if we consider these figures, the sensitivities of the two systems will (within the limits of the amplifier gain adjustment) be defined by the sensitivity frequency characteristics of the receivers themselves ($20 \log_{10} \frac{P}{U_R}$, the voltage U_R being constant and independent of frequency).

With such definitions for the measuring circuits, subjective and objective measurements may be made with circuit arrangements identical to that shown in Figure 4, and the results obtained may then be compared.



The sum of the attenuations A and C is kept constant and equal to 36 db

FIGURE 4. — **Schematic diagram of the experimental circuit for the comparison, subjectively and objectively, of the sensitivities of a commercial receiver and an A.R.A.E.N. receiver**

In the subjective comparisons of the two circuits, the sum of the attenuations $A + C$ will be kept constant (for each measurement frequency) and equal to 36 db. The value of U_E will be 1 volt. The gain (represented by a loss) of the matching network (amplifier + transformer) will be 16 db. Thus, the mean subjective sound intensity level will be roughly the same as in the subjective comparisons (voice and ear) when the reference equivalents are determined with a constant level of received speech sounds in the N.O.S.F.E.R. corresponding to an attenuation of 36 db in the N.O.S.F.E.R. line. (This experimental condition is generally defined by the symbol R36.)

The mean sensitivity of the N.O.S.F.E.R. receiving system is approximately +25 db/1 dyne/cm²/1 volt (see Table 9, p. 49, Volume V of the *Red Book*), the receiver alone having a mean sensitivity of approximately +43 db/dyne/cm²/1 volt; thus, assuming that during subjective comparisons (reference equivalents) the average speech voltage at the output of the N.O.S.F.E.R. sending system is −10 db/1 volt with 36 db in the line, the mean subjective sound intensity level developed by the N.O.S.F.E.R. receiver at the listening operator's ear will be −21 db/1 dyne/cm² [(−10) + (−36) + (+25)] (or 53 db above the reference level of $2 \cdot 10^{-4}$ dyne/cm²). In the measurements described above, we have, for $U_E = 1$ volt, a mean sound pressure level equal to:

$$0 + (-36) + (-6) + (-12) + (+43) = -21 \text{ db/1 dyne/cm}^2.$$

Then, by using an oscillator associated with a recorder, the response curves of the above-defined systems will be quickly determined. The tests will be completed by taking the measurements with a normal N.O.S.F.E.R. listening channel, keeping the same receiver as in the above tests (Fig. 3). In these last measurements, the voltage applied to the input of the N.O.S.F.E.R. receiving system will be kept constant, independent of the frequency, and adjusted to −20 db/1 volt; thus, the voltage at the receiver terminals will depend on the frequency but with appropriate corrections deduced from the " frequency-gain " characteristic of the electrical part comprised between the input terminals of the N.O.S.F.E.R. receiving system and the receiver terminals, the values measured on the basis of the reference system defined in Figure 3 should be re-obtained. The response curve of a N.O.S.F.E.R. channel will then be taken with a recorder.

With the curves obtained in this way in all the cases, the over-all sensitivity differences will be determined by the method described in Annex 29 (Part II of Volume V of the *Red Book*), between a commercial receiver, or a single A.R.A.E.N. receiver, and the N.O.S.F.E.R. receiving system. The measurements will be completed by determining (subjectively) the reference equivalent of a system defined either in Figure 2 or in Figure 3. In this measurement, the level of received speech sounds, in the N.O.S.F.E.R./N.O.S.F.E.R. system will correspond to that occurring in the case of a total reference equivalent of 36 db.

1.2.2 *Choice of handsets and receivers*

The most delicate part of the tests relates to the measurement of the sound pressure developed by a receiver. To this end, the Laboratory has examined all the available types of earcap and has chosen for these tests some which represent typical shapes, viz:

— a *flat* earcap only slightly concave (as in the handset of the French U43 telephone, and as in the telephone used by the British Post Office);

— a *shallow* earcap (type used by the Federal German and Swiss Administrations, the two being slightly different);

— a *markedly concave earcap* (as in the Ericsson telephone used by the Netherlands Administration);

— each type of ear-piece will be designated by a letter α, β, etc.

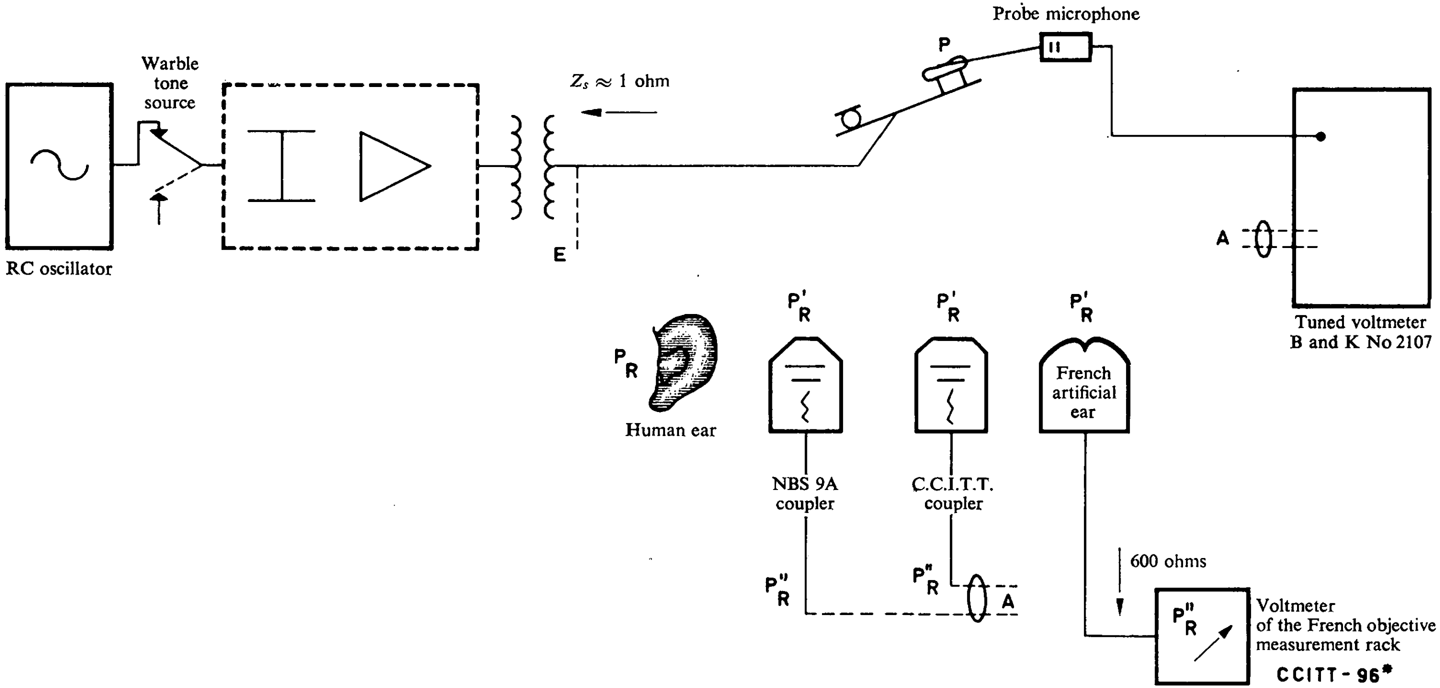


FIGURE 5

P_R = Pressure measured by the probe microphone when the receiver is applied to the human ear

P'_R = Pressure measured by the probe microphone on the artificial ear (plane of the opening of the artificial ear coupler)

P''_R = Pressure referred to the microphone of the coupler, taking into account the absolute sensitivity of the artificial ear

The Working Party considered it necessary to complete this list and to include a type of earcap having a relatively low acoustic impedance. To this end, the representative of the Chile Telephone Company proposes to provide the Laboratory with the following elements, relating to the handset type I.T.T. Kellogg 65 C:

— two handsets

— two supplementary earcaps

— one receiving inset

These requests were made for purely practical reasons. It is not easy to see beforehand how to introduce the microphone probe and the position it will take up without knowing exactly how any particular earcap would fit any particular artificial ear. With these additional items, the Laboratory can determine exactly the position of the probe and choose the diameter and shape of the most suitable probe. The tip of the probe will thus lie at the centre of the reference plane of each artificial ear investigated.

Note. — During the meeting, members of the Working Party examined the experimental precautions taken for the performance of these tests.

1.2.3 *Method of carrying out the tests and presentation of the results*

The method for carrying out the tests will be as described below and illustrated in Figure 5 (the human ear and three types of artificial ear were taken as examples). For a given measurement frequency, the operator adjusts the receiver to his or her ear as well as possible. This fit will be obtained using a warble tone [1] of relatively low level which is injected beforehand for a few moments instead of the measurement frequency. When the operator considers the fit to be satisfactory, he switches to the measurement position. The subject being tested will not read the voltmeter indications, which will be out of his sight; he should not therefore be influenced in adjusting the receiver to his ear.

First of all, the gross values of P_R will be measured; these values represent the sound pressure developed by the receiver at a point in the plane of the ear where the tip of the probe microphone is situated. These gross values will be entered in tables similar to Table 1 (α) [2]; there will be a separate table for each operator and for each type of receiver earcap. Since it is possible to calibrate the probe microphone, it will be possible to calculate the absolute value of the sound pressure (in db/1 dyne/cm^2). The corrected gross values will then be entered in tables similar to Table 1 (α).

[1] The warble tone is the sound source of the apparatus for the objective measurement of reference equivalents used by the Federal German Republic (Annex 28, Part II of Volume V of the *Red Book*).

[2] The artificial ears will be designated respectively by the letters *a*, *b*, *c* . . . *g*, for example; Tables 1 (α), 2 (α), 3 (α), relate to the type of receiver earcap α.

TABLE 1 (α)

Type of receiver:

Pressure measured by the probe microphone (allowance made for its sensitivity) when the receiver is placed against the human ear

(db relative to 1 dyne/cm²)

Measurement of P_R Operator............

c/s	Artificial ear [1]							Total	Mean
	a	*b*	*c*	*d*	*e*	*f*	*g*		
100									
200									
300									
400									
500									
600									
700									
800									
900									
1000									
1100									
1200									
1300									
1500									
1800									
2000									
2200									
2500									
2700									
3000									
3300									
3600									
4000									

[1] These values relate to the human ear; they were measured (for a given operator) before the receiver was placed on the given artificial ear (see the operating method described above).

Each artificial ear is designated by a different letter.

Then the operator will place the receiver on the artificial ear under study in a manner conforming to clearly-defined instructions; the gross values of P'_R will be noted, which represent the sound pressures developed by the receiver at a point in the reference plane of the coupler of the artificial ear close to the centre of the cavity cross-section. These values, corrected in the same way as above, will be entered in tables similar to Table 2 (α). In this case, for a given artificial ear, there will be a table for each type of earcap.

TABLE 2 (α)

Type of receiver:

Pressure measured by the probe microphone (allowance made for its sensitivity) when the receiver is placed on an artificial ear

(db with respect to 1 dyne/cm²)

Measurement of P'_R Artificial ear No. ......

c/s	Operator No. 1	2	3	4	5	Total	Mean
100							
200							
300							
400							
500							
600							
700							
800							
900							
1000							
1100							
1200							
1300							
1500							
1800							
2000							
2200							
2500							
2700							
3000							
3300							
3600							
4000							

Notes. To obtain the absolute sensitivity of the receiver, account must be taken of the voltage applied to the receiver terminals (-40 db/1 volt 200-1500 c/s band; -30 db/1 volt 1800-2700 c/s band; -20 db/1 volt $f \geq 3000$ c/s).

The corrected values could be compared with the values obtained in Table 3 (α); the latter being themselves corrected to allow for the sensitivity of the microphone included in the artificial ear.

At the same time, the readings supplied by the microphone of the artificial ear will be noted. These values, themselves corrected to allow for the sensitivity of the artificial ear in question (or of the microphone included in the coupler) will be entered in tables similar to type 3 (α). There will also be in this case, for a given artificial ear, one table for each type of receiver earcap.

When, for a given operator and frequency, the three measurements relating to one artificial ear are completed, the operator will recommence the same operations in order to carry out the measurements with another artificial ear, and so on up to the last artificial ear. The operator who has finished this measurement cycle will be replaced by another operator.

TABLE 3 (α)

Type of receiver:

Measurement of the sound pressure produced by the receiver when applied to the artificial ear (readings from the artificial ear, allowance made for its sensitivity)

(db with respect to 1 dyne/cm²)

Measurement of P''_R Artificial ear No. ......

c/s	Operator No. 1	2	3	4	5	Total	Mean
100							
200							
300							
400							
500							
600							
700							
800							
900							
1000							
1100							
1200							
1300							
1500							
1800							
2000							
2200							
2500							
2700							
3000							
3300							
3600							
4000							

See the notes to Table 2 (α).

Thus, for a given artificial ear, the values entered in the corresponding Tables 1 (α), 2 (α), and 3 (α) for example (or 1 (β), 2 (β), 3 (β)...) will be comparable and it will be possible to complete the presentation of the results with the ratios:

$$P_R/P'_R \text{ and } P'_R/P''_R$$

2. TESTS TO BE CARRIED OUT WITH ARTIFICIAL MOUTHS

2.1 *Measurement of the increase in sound pressure due to an obstacle placed before the lips of the artificial mouth*

The C.C.I.T.T. Laboratory will measure the increase in sound pressure resulting from the presence of a given obstacle placed in front of the lips of the artificial mouths at its disposal.

The artificial mouths available to the Laboratory belong to the following Administrations:

— French Administration
— Chile Telephone Company
— Italian Administration

— F.A.T.M.E. Company (Italy)
— Administration of the German Federal Republic
— The United Kingdom Administration
— Czechoslovak Administration

Moreover the artificial mouths of the Swedish and Swiss Administrations may be put at the Laboratory's disposal.

Descriptions of and information concerning these mouths are the subject of contributions COM XII—Nos. 46 and 63 (period 1957-1960).

The measurements will be made using a disk of 6.3 cm diameter placed in turn at 2 cm and 4 cm directly in front of the lips of each artificial mouth.

At each of the above two points measurements will be made of the pressure with a probe microphone, with and without the disk in position.

Response curves will be traced (using pure tones) for the two distances.

As guidance in carrying out and presenting the results, an extract from contribution COM XII—No. 57 (period 1957-1960) is given below.

It should, however, be noted that this extract refers to tests with real mouths.

2.1.1 *Measurement of diffraction effect of a disk*

Figure 6 shows the diffraction effect produced by a disk 6.3-cm diameter when placed in front of the mouth. The change in the frequency response of the diffraction effect when the obstacle is moved from 2 cm to 4 cm from the artificial mouth is seen from the difference between these two figures. The same disk was placed in front of the mouth of human subjects at corresponding distances and the diffraction effect measured for narrow bands of speech.

Note. — The artificial mouths of the Swedish and Swiss Administrations will be included in these tests; the Laboratory will inform these Administrations of the date when the tests will start, with an advance notice of two months.

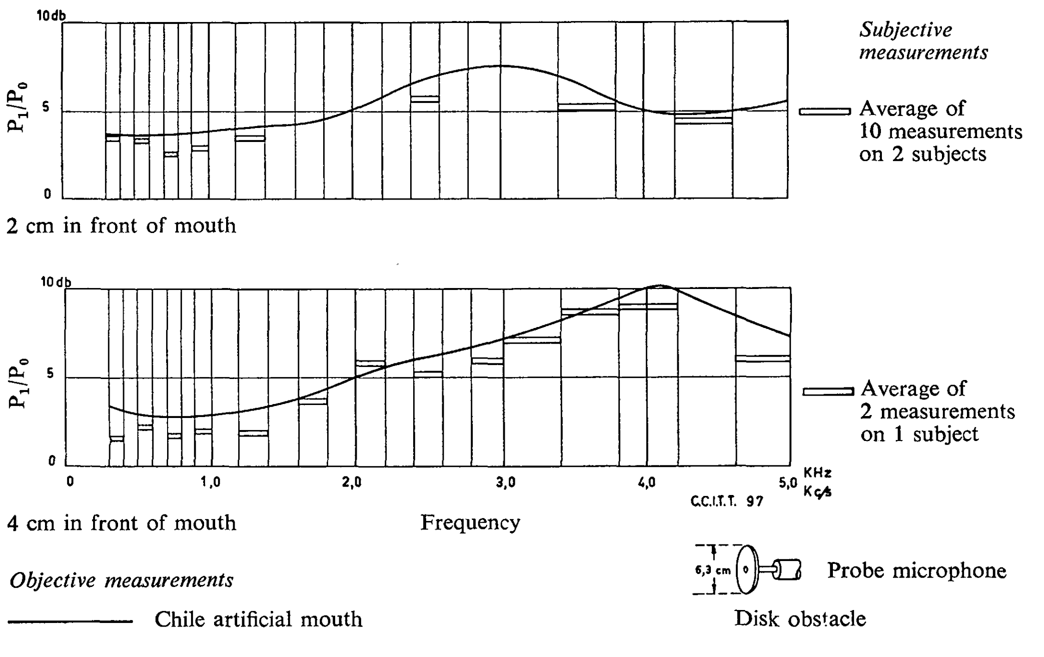


FIGURE 6. — **Increase in sound pressure P_1/P_0 due to a 6.3-cm diameter disk**

2.2 *Tests to be carried out under points 3 and 4 of Annex 1*

Determination of the sending reference equivalent of two linear microphones mounted in the handsets of the telephone systems of the following five Administrations:

France, Mexico (10th series), Norway, Sweden and Switzerland.

These microphones are:

— of the electrodynamic type (supplied by the Swedish Administration)

— of the magnetic type (supplied by the Swiss Administration).

In a letter dated 18 April 1963, the Swedish Administration promised to send a microphone (with a handset and receiver) associated with an appropriate amplifier, which, with a feeding bridge, would form a complete sending system.

Before a start is made with the subjective tests, it will be necessary to check whether the microphone insets are correctly placed in the mouthpieces of the handsets; for that purpose, the response curve of the sending systems will be measured with the aid of a particular artificial mouth.

2.2.1 *Determination of the reference equivalent by subjective comparison in the normal manner, with a talker*

For each of the two kinds of inset, two replications will be performed per type of handset. The reference equivalent will be derived from the mean of $20 \times 2 = 40$ individual talker/listener balances.

The level of speech received in the N.O.S.F.E.R./N.O.S.F.E.R. reference system will be kept constant at a value corresponding to 36 db in the line of the reference system.

Guard rings will be used, appropriate to each of the two talking distances, namely:

— that for determination of reference equivalents;

— that for A.E.N. measurements.

For each of these talking distances, the operator will adjust his talking level to the appropriate value, i.e. either the normal talking level for voice/ear tests (the sensitivity of the A.R.A.E.N. volume meter being adjusted to -10 db), or the reference talking level for A.E.N. (the sensitivity of the volume meter of the A.R.A.E.N. being adjusted to -6 db) [1].

The talking distance adopted for the reference system (N.O.S.F.E.R.) will in both instances be that defined for N.O.S.F.E.R.

2.2.2 *Determination of the reference equivalent by subjective comparison; the talker being replaced by an artificial mouth emitting speech*

The artificial mouths will be those of the following Administrations:

France, Federal Republic of Germany, United Kingdom, Czechoslovakia, and the following private Companies: Chile Telephone Company, and F.A.T.M.E. (Italy).

The artificial mouths of the Swiss and Swedish Administrations will be added to this list. To this end, the Laboratory will inform these Administrations in good time of the starting date of the tests.

[1] Previous tests have shown that to get the reference talking level for A.E.N., operators had to pronounce the conventional phrase "PARIS..." with a level 4 db higher (A.R.A.E.N. voltmeter reading) than the normal talking level for voice/ear tests.

Note. — It will not be possible to use the artificial mouth of the Italian Administration owing to its design (see Annex 9, Part II of Volume V of the *Red Book*). The artificial mouth proper (loudspeaker) is associated with a source derived from a magnetic tape on which a certain number of discrete frequencies have been recorded; the amplitudes of these frequencies allow for the linear distortion of the loudspeaker, so that the response curve of the assembly (magnetic tape plus loudspeaker) is appreciably independent of the frequency.

Initially, for each of the five operators in the team, a sample of speech will be recorded on an endless tape, namely:

— the conventional phrase " Paris-Bordeaux ... Loudun ",

— a short, well-chosen, phrase lasting roughly as long as the conventional phrase spoken in the *talker's own language* (Spanish, English, Norwegian, German).

These phrases are as follows:

English: CAN — CON — BUY — DACE — ALSO.

German: BERLIN — HAMBURG — MÜNCHEN — KOBLENZ — LEIPZIG — DORTMUND.

Spanish: PALENCIA — GIJÓN — FERROL — SANLÚCAR — LÉON — TERUEL.

Norwegian: OSLO — BERGEN — TRONDHEIM — NARVIK — ÅLESUND — VARDO.

These recordings will be made with a Studer 37 high-fidelity recorder, at the output of the A.R.A.E.N. *sending system* (sensitivity practically independent of frequency). The talking distance will be that specified for the measurement of A.E.N. (33.5 cm). This distance can be adopted for, as is shown by Technical Report No. 295 issued by the Laboratory, the effects due to the cabin can almost be neglected for this distance.

The tape-recorder output terminals will be connected (possibly by a matching network) to the " electrical input " terminals of the equipment defining the input of that particular artificial mouth (amplifier, correcting network, etc.). The point representing the " equivalent lip position " of both artificial mouths will be placed alternately in front of the N.O.S.F.E.R. microphone guard-ring and in front of the handset guard-ring. The acoustic speech pressure developed by the artificial mouth will first be adjusted in such a way that the reference deflection for either the normal talking level used in voice/ear tests or for the A.E.N. talking level can be read off from the A.R.A.E.N. volume meter.

The experimental procedure and the number of replications will be as mentioned above.

ANNEX 3

(to Question 12/XII)

Principal experimental conditions to be taken into consideration during measurements with artificial mouths and ears

(See the former Annex 1, pages 674 to 677 of the *Red Book*, Volume V.)

ANNEX 4

(to Question 12/XII)

The acoustic input impedance of the outer ear

(Communication from the Institut für Nachrichtentechnik der Technischen Hochschule, Stuttgart)

by E. ZWICKER

Introduction

The input impedance of the human ear is important not only from the physiological point of view but also for more technical reasons. When the subscriber uses the telephone, he holds the handset to his ear. The transducer in the handset, which is usually an electromagnetic or even a moving coil system, is terminated acoustically by the input impedance of the human ear. Since the transducer discharges acoustic power, and as it must function so far as possible in the best conditions (these are discussed more fully below), its terminal acoustic impedance is most important. From the technical standpoint, the input impedance in front of the ear drum, of interest to the physiologist, is unimportant. We are interested here exclusively in the terminal acoustic impedance encountered by the handset transducer, i.e. the input impedance of the ear as it affects the receiver. Handset transducers are tested with so-called " artificial " ears [1, 2] which should reproduce the input impedance of the human ear. The transfer coefficient tolerances prescribed for such tests are relatively close, and the influence exercised on this transfer coefficient by the input impedance of the artificial ear may be considerable. It is advisable, therefore, to ascertain for a large number of persons, the input impedance of the ear when using the telephone and to compare the mean values with those obtained for " artificial " ears.

The procedure used for these measurements was originated by A. E. Kennelly [3], and was developed by C. W. Kosten and C. Zwicker [4] to test the absorption capacity of absorbent materials. With this method, advantage is taken of the effect of the acoustic termination of a transducer upon its electrical input impedance.

Purely acoustic test methods are relatively inaccurate, while electrical methods have been greatly improved in the past 20 years. It would seem preferable, therefore, to use electrical rather than acoustic methods, despite the low efficiency of electroacoustic transducers.

The usual term " acoustical impedance ", as opposed to electrical impedance, is not always correct. Often, as in this article, we mean mechanical impedance, which is defined as the force/velocity ratio. Nevertheless, as the latter expression is not yet so widely used as to include the "mechanical input impedance of the ear", which is what we mean here, we shall continue to speak of acoustical impedance (cf. equations (6) and (7)).

Measuring method

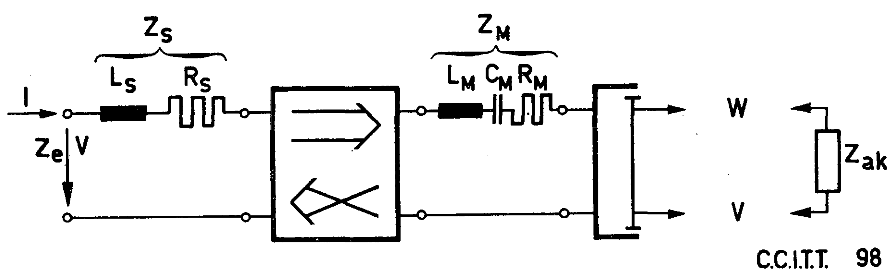
Every acoustic transducer transforms electrical power into acoustic power. The power obtained depends on the type of transducer and the conditions of sound propagation in which the acoustic power is discharged. The electromagnetic and moving coil transducers give the best results for sounds in air, whereas both the piezoelectric and electrostatic transducers function under conditions of poor matching over the average audio-frequency range.

It is a simple matter to vary the inherent resonance of a magnetic transducer and situate it at a given frequency. Efficiency is very high at the resonance frequency, so that originally it seemed best to work with magnetic transducers functioning at resonance. When the output is high, the influence of the acoustic termination on the electrical input impedance of the transducer is very considerable. Tests on these lines, however, have proved that the resonance frequency varies more with the temperature than is acceptable for the accuracy required. As regards an electromagnetic transducer, it is undesirable that this should function in practice according to a square law and it is endeavoured to seek linearity by superimposing a constant magnitude. In this way we obtain a single valued variation of the electrical input impedance in relation to the voltage.

In the tests described below, moving coil systems were preferred, electromagnetic systems being used only for comparison purposes. It was not possible to find any variation of the input impedance in relation to voltage in moving transducers. Where there was a variation in temperature, moreover, the main effect was a variation in the real component of the input impedance. It is the variation of the copper conductance in the moving coil which is important here. It is possible to compensate for this by making relative instead of absolute measurements. As the moving coil transducer is based on a linear law and it is generally accepted that its transmission characteristics are very constant, this type of transducer was used for almost all the measurements. In the light of previous experience with electromagnetic systems, special systems operating at resonance were not used.

Representation of the acoustical impedance plane on the electrical impedance plane

The moving coil transducer has an equivalent circuit [5] which is shown in Figure 1. It is based on the analogy, voltage → energy, or, current → velocity. The unit electroacoustic transducer links the electroacoustic quantities, energy W and velocity V, to the electric quantities, voltage V and current I. The series elements L_M, C_M and R_M are responsible for the inherent resonance of the system. The insertion of the gyrator corresponds to the action of magnetic fields in the conversion of electric power into acoustic power. L_S and R_S result from the inductance



$$\begin{Vmatrix} 0 & Z_{\text{gyr}} \\ \frac{1}{Z_{\text{gyr}}} & 0 \end{Vmatrix} \qquad \begin{Vmatrix} 1 & Z_M \\ 0 & 1 \end{Vmatrix} \qquad \begin{Vmatrix} \frac{1}{N} & 0 \\ 0 & N \end{Vmatrix}$$

FIGURE 1. — **Equivalent circuit of an electrodynamic transducer**

and the resistance of the moving coil. For the characteristic impedance of the gyrator we have $Z_{\text{gyr}} = B_0 \cdot l/N$, where N is the transfer coefficient of the unit transducer, B_0 the induction in the air-gap, and l the length of the conductor. Each of the iterative matrices shown in Figure 1 refers to the graphical symbols appearing above it.

The question of interest here is the variation of electric input impedance Z_e as a function of the terminating acoustic impedance Z_{ak}. By multiplying the matrices shown in Figure 1, we first obtain the iterative matrix for connection of the gyrator in series with Z_M and of the unit transducer. It is expressed as follows:

$$\begin{Vmatrix} 0 & Z_{\text{gyr}} \cdot N \\ \dfrac{1}{Z_{\text{gyr}} \cdot N} & \dfrac{N \cdot Z_M}{Z_{\text{gyr}}} \end{Vmatrix}$$

From [6], the input impedance Z_1 of the gyrator is, in the case of acoustic termination by Z_{ak}:

$$Z_1 = \frac{A_{11}\, Z_a + A_{12}}{A_{21}\, Z_a + A_{22}} = \frac{Z_{\text{gyr}} \cdot N}{\dfrac{Z_{ak}}{Z_{\text{gyr}} \cdot N} + \dfrac{N \cdot Z_M}{Z_{\text{gyr}}}}$$

By adding Z_S we obtain:

$$Z_e = Z_S + \frac{Z^2{}_{\text{gyr}} \cdot N^2}{Z_{ak} + Z_M\, N^2} \qquad (1)$$

This equation has the form

$$Z_e = L + \frac{A}{Z_{ak} + G} \qquad (2)$$

It may be concluded, from laws valid for conformal transformations [6] that the right-hand half-plane of the acoustic terminating impedance Z_{ak} becomes a circular area in the plane of the electric input impedance Z_e. The conversion of Z_{ak} into Z_e is represented by the diagram in Figure 2. Consideration of this conformal transformation produces interesting results which will not be given here in detail. We shall simply mention here that the real component of the quantity $G\,[R_e(G) = R_M \cdot N^2]$, with the quantity $A = (B_0 l)^2$, determines the diameter of the circle, while the imaginary component of $G \left[I_m(G) = N^2 \left(\omega L_M - \dfrac{1}{\omega C_M} \right) \right]$ causes the graduation of the circle to become closer or wider. The point ∞, independent of the frequency, should then always be the point in the circle that is closest to the imaginary axis. As shown by the practical measurements described below, this is not always the case, however. The equivalent diagram shown above is therefore not yet quite complete. Nevertheless, the laws of the conformal transformations are valid at a given frequency, so that the shape of the electric input impedance as a function of the electro-acoustic terminal impedance indicated in equation (2) is produced.

Figure 2 shows quite clearly that the imaginary axis in the Z_{ak} plane is changed into a circle in the Z_e plane, while any points of the right-hand plane, Z_{ak}, are changed into points on the circle in the Z_e plane. If it is further considered that it is possible to use the laws of conformal representations to obtain the complete mapping of the circular area, taking as a basis a given graduation of the circumference, the problem lies in finding the graduation of the circle, i.e. of the transformed imaginary axis. All the lines parallel with the imaginary axis or the real axis in the Z_{ak} plane are changed into circles in the Z_e plane. The graduation of these groups of circles depends on that of the external circle, which is the representation of the imaginary axis of the Z_{ak} plane.

Since A, G and L are complex magnitudes, depending on the frequency, it is easy to understand that there can be no univocal transformation except for a constant specified frequency. There is a different transformation for each frequency.

The other electroacoustic transducers can also be represented as quadripoles [5]. The relations established in equation (1) for the moving-coil transducer can also be established in corresponding fashion for the other transducers and be enunciated in the form of the equation (2). The considerations put forward are here therefore valid for all electroacoustic transducers governed by linear laws.

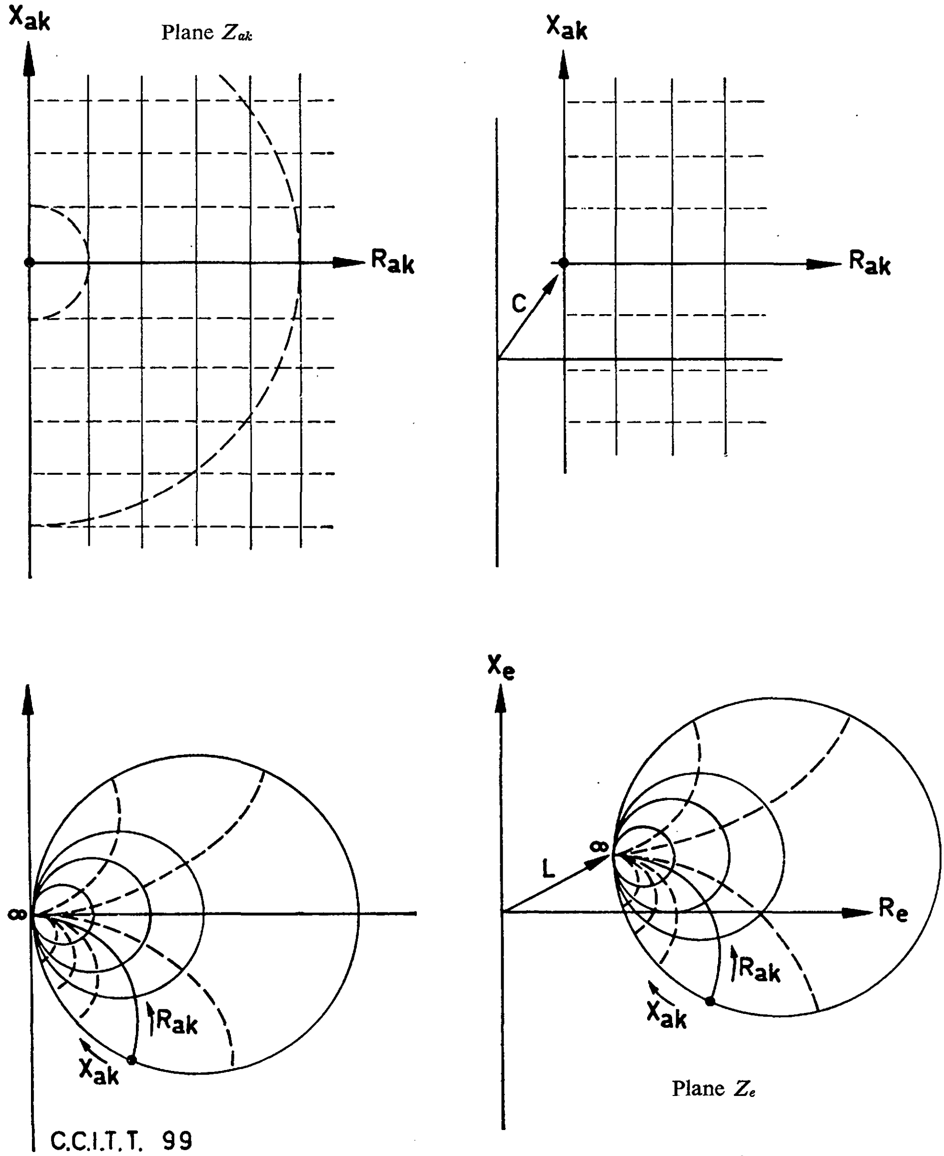


FIGURE 2. — **Transformation of the right-hand half-plane of the terminating impedance Z_{ak} into the plane of the electric input impedance Z_e of the transducer**

Graduation of the circle

The imaginary axis of the Z_{ak} plane is represented by the circle. Electroacoustic reactance with easily variable values, i.e. the imaginary axis, can be simulated with the aid of a sound tube with a rigid end wall (Fig. 3). Provided the transversal dimensions of the tube are small with respect to the wavelength, the following expression is obtained for a short-circuited acoustic tube having the dimensions indicated in Figure 3:

$$\begin{aligned} W_1 &= W_2 \operatorname{ch} j\,\beta\, l \\ V_1 &= V_2/Z \operatorname{sh} j\,\beta\, l \end{aligned} \tag{3}$$

with

$$\beta = \frac{2\pi}{\lambda} \tag{4}$$

$$\lambda = \frac{c}{f} \tag{5}$$

and

$$Z = Z_0 \cdot S \tag{6}$$

where

- β is the transmission coefficient of the tube (α ignored)
- λ is the wavelength in the tube
- c is the speed sound in the air at 20° C
- f is the frequency
- Z_0 is the characteristic impedance of air $Z_0 = 41.3 \dfrac{\text{g}}{\text{cm}^2 \cdot \text{s}}$
- S is the cross-sectional area of the tube.

$$\left(S = \pi \frac{d^2}{4}\right)$$

The result, for a tube diameter of 1.7 cm, is:

$$S = 2.27 \text{ cm}^2 \text{ and } Z = 93.5 \frac{\text{g}}{\text{s}}$$

From equations (3) and (4) we have:

$$\frac{W_1}{V_1} = Z_{ak} = Z \coth j\,\beta\, l = -j\, Z \cot \beta\, l = -j\, Z \cot 2\pi \frac{\lambda}{l} = j\, X_{ak} \tag{7}$$

or

$$\frac{X_{ak}}{Z} = -\cot 2\pi \frac{l}{\lambda}$$

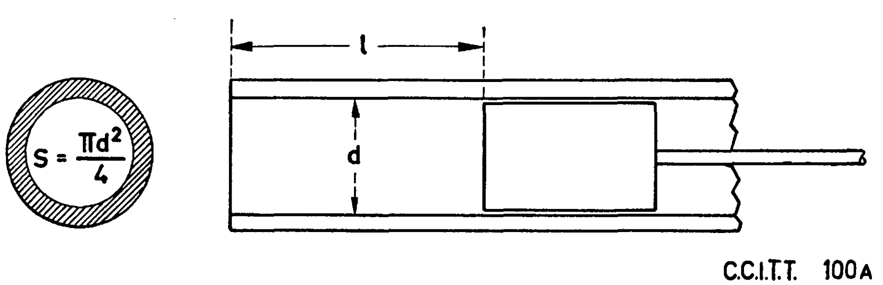


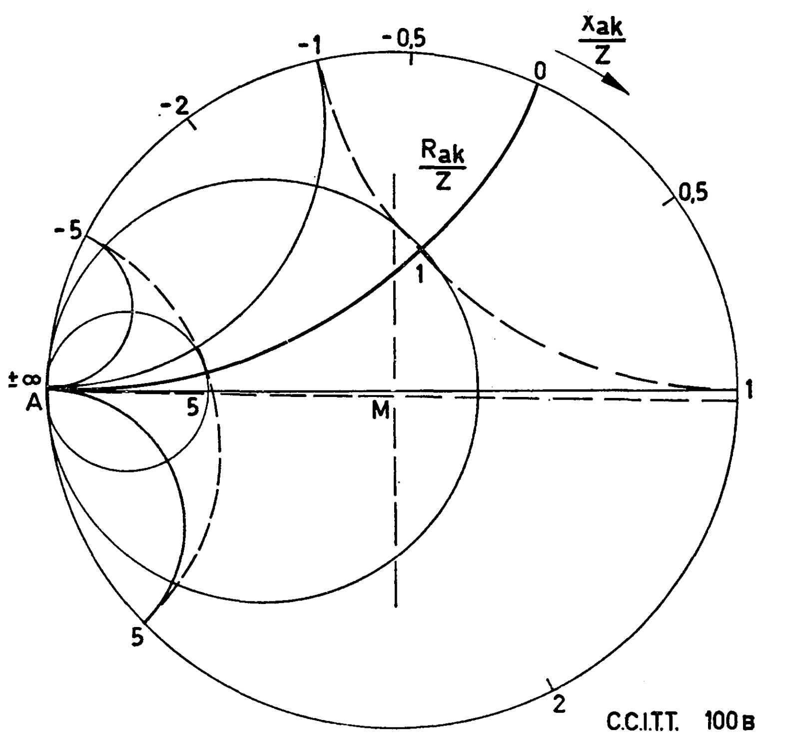
FIGURE 3. — **Sketch of the defined electroacoustic reactance, formed by a tube with a rigid termination**

From this equation, we have determined for 17 values of $\frac{X_{ak}}{Z}$ the corresponding values for l/λ taking into account equation (5). From $l/\lambda = 0$ to $l/\lambda = 0.5$, passing through $l/\lambda = 0.25$, $\frac{X_{ak}}{Z}$ goes from $-\infty$ to $+\infty$, passing through 0. In this way, the circle is traversed once in the Z_e plane and a check can be made of the extent to which the values for $l/\lambda = 0$ agree with those for $l/\lambda = 0.5$. Measurements were taken at the following values for

$$\frac{X_{ak}}{Z} = \pm\infty; \quad \pm 10; \quad \pm 5; \quad \pm 2; \quad \pm 1; \quad \pm 0.5; \quad \pm 0.2; \quad \pm 0.1; \quad 0.$$

In this way, the graduation of the circle is fixed and it is a question of fixing the groups of circles inside and of graduating them.

Figure 2 shows that all the straight lines in the Z_{ak} plane, especially the lines parallel to the real axis and the imaginary axis, are represented in the plane by a point. This point is represented by A in Figure 4. It is the first point of determination for all the circles and corresponds to the value which is obtained when measuring with $l/\lambda = 0$; 0.5; 1; 1.5 ... therefore for $\frac{X_{ak}}{Z} = \pm\infty$, i.e. open-circuit operation. When the circle and its mapping are known, the centre of the circle (M) is also known and the result is that, as second means for determining the circles $\frac{X_{ak}}{Z} =$ const., they are tangents to the straight line AM at A. Since they must also pass through the point of



Plane Z_e

FIGURE 4. — **Examples for mapping of the circular area**

the outer circle having the value $\frac{X_{ak}}{Z}$, which the whole circle must have, and since they must, at this point, cut the outer circle orthogonally, the circles $\frac{X_{ak}}{Z}$ = const. are quite clearly determined. The circles $\frac{R_{ak}}{Z}$ = const. result from the conformal transformation at the corresponding angle. Their centres must lie on the straight line AM and must also touch the circles in the Z_e plane which have resulted from the circles surrounding the zero point in the Z_{ak} plane. For the values $\frac{Z_{ak}}{Z} = 1$ and $\frac{Z_{ak}}{Z} = 5$, the auxiliary circles necessary in the R_{ak} plane in Figure 2 and in the Z_e plane in Figure 4 are shown dashed. In this way, a complete mapping of the interior of the circle is obtained, so that any electroacoustic impedance which loads the transducer can be determined by measuring the electric input impedance of the transducer.

The circuit arrangement used for the measurements and the measurement conditions

Figure 5 shows the circuit used for the measurements. A constant voltage low-frequency generator feeds the impedance bridge which is used to measure the input impedance of the transducer for different electroacoustic loads.

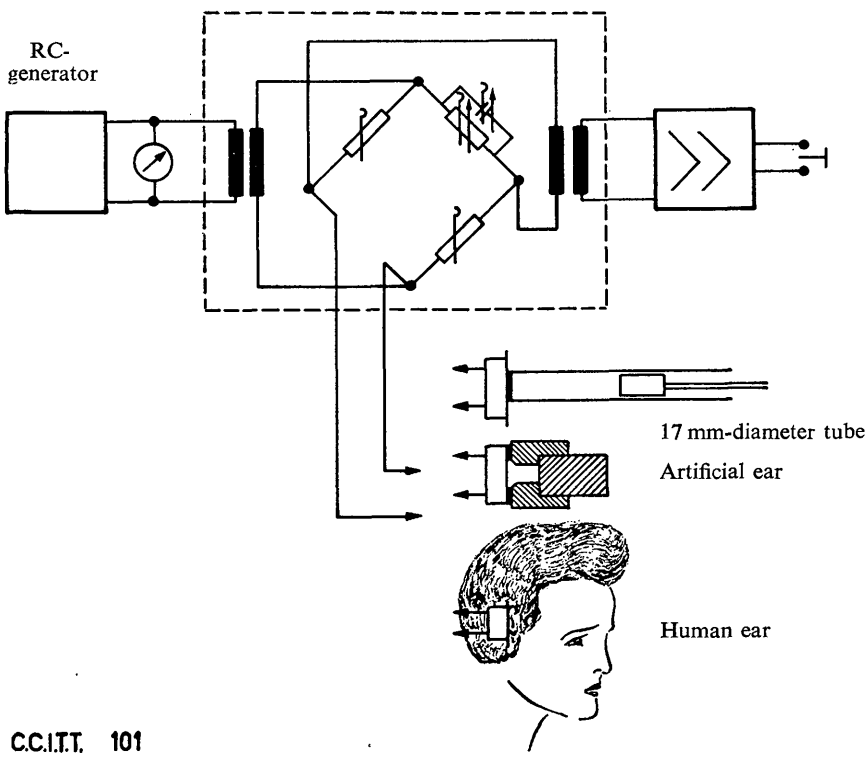


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FIGURE 5. — **Measurement circuit arrangement**

Stringent requirements are placed on the measurement bridge, at least as regards its relative accuracy. A heterodyne receiver is connected to the bridge output. The bridge is balanced using headphones.

The circular diagrams and electroacoustic impedances were measured at frequencies of 300 c/s, 500 c/s, 700 c/s, 1 kc/s 1.5 kc/s, 2.2 kc/s and 3.4 kc/s according to the frequency band transmitted in the telephone network. The voltage at the bridge input was set at 100 mV, except when special conditions were specifically indicated. The simulation of the acoustic reactance has been effected by means of steel or brass tubes 17 mm in diameter.

The following effects have been studied and, as far as possible, eliminated:

a) The radiation of sound in the air, coming from the open back in the case of acoustic transducers, has been suppressed by filling the handset in which the transducer is mounted with sand.

b) The acoustic radiation is also reduced by this precaution. Moreover, the handset used was made of material having considerable internal attenuating properties.

c) The effect of the acoustic radiation of the tubes has been reduced by bedding the tubes in sand so that they could not vibrate.

d) The possible effect of the diameter of the tube was checked by measuring the mapping of the diameter of the circle, using tubes of 12 mm, 17 mm and 35 mm diameter. It was found that the circle remained unchanged when the diameter was varied, and that its mapping became smaller or larger as prescribed, depending on the ratio of the transversal sections. Hence, with an internal diameter variation from 17 mm to 12 mm, a reduction in the ratio of two to one was noted very precisely. Only with an interior diameter of 35 mm and also only at the highest frequency used (3.4 kc/s) did changes occur in the form of the circle, showing that the hypothesis $D \ll \lambda$ was no longer fulfilled. In the evaluation, tubes 17 mm in diameter were used.

e) The position of the transducer capsule in the handset was marked and kept constant. The effect of the contact resistance between the contact springs and the capsule were studied; it had no effect on the measurements.

f) The voltage at the input of the bridge was varied from 30 mV to 10 V. For the moving coil systems with which the following results were obtained, no change occurred in the balance of the bridge; only the heating of the copper wire of the moving coil by the electric power applied produced a slow variation. The supply voltage to the bridge was kept constant at 100 mV, corresponding to a voltage of about 25 mV at the transducer.

g) The variation in the electric input impedance of the transducer as a function of the temperature is considerable in relation to the requisite measurement precision. Since the temperature of the transducer cannot be kept exactly constant, the values measured have been corrected for the temperature in question by checking the value for $\frac{l}{\lambda} = 0$, hence $\frac{X_{ak}}{Z} = \infty$, before and after each measurement of the decisive acoustic impedances. All the other measurements were referred to this value. This seems admissible because exact measurements of the circle and its mapping as a function of the temperature have shown that the circle keeps not only its diameter but also the position of its graduations and, when the temperature increases, is only displaced towards the greatest real values, hence variations occur in the real component of the input impedance of $3\,{}^\circ/_{\circ\circ}/{}^\circ\mathrm{C}$, which corresponds to the temperature coefficient of the copper.

The variation in the speed of sound as a function of the temperature had to be taken into consideration during the recording of the circles, with $\frac{\Delta c}{c\,(20^\circ\mathrm{C})} = 1.8\,{}^\circ/_{\circ\circ}/{}^\circ\mathrm{C}$ as the variation in the wavelength in the tube.

h) A study was also made to see whether the type of transducer used affected the results of the measurements. To this end measurements were made, both for the electromagnetic transducer and for the moving coil transducer, of the acoustic impedances of artificial ears. The results of the measurements agree within the measurement precision limits, so it can be assumed that the transducer itself has no influence on the result of the measurements. The results given here were obtained with a moving coil transducer. It has proved to be the most reliable and stable transducer for all the measurements.

The measurement plan

The acoustic input impedance measurements at the frequencies mentioned above were first of all carried out on 30 to 40 persons, so that 60 to 80 measurement points were available at each frequency, since the left and right ears of each person were measured. With one exception the models were men (students of about 22 to 28 years of age). About one-third wore glasses.

For the first measurement, the people were asked to hold the receiver directly and firmly to the ear as they would when telephoning. For the second measurement, which was made a few day later, the same people were asked to hold the receiver very close to the ear as they might if they were telephoning and finding difficulty in hearing because the sound was too faint. In that case also the receiver was to be placed centrally and firmly to the ear.

The measurements were taken in such a way that, after calibrating the measurement bridge, the point for $Z_{ak} = \infty$ at one frequency was measured, then the input impedance of the left ear of the person at that frequency, again for $Z_{ak} = \infty$, then the right ear was measured and finally the measurement was repeated for $Z_{ak} = \infty$. After these five measurements the frequency was then changed and the bridge was calibrated for the new frequency. Then the five corresponding measurements were taken. About half-an-hour was needed for the series of measurements with seven frequency values and five measurement points for each of them.

The results of the measurements

a) *When the receiver is applied normally*

Figure 6 is the right half-plane of the acoustic terminal impedance Z_{ak} in the electric input impedance Z_e of the electrodynamic transducer (placed in the handset) at a frequency of 1000 c/s. The 82 measurement points for 41 people are shown as black dots in the circle. Most of the measurement points are well within the limits $0.5 < \frac{R_{ak}}{Z} < +3$ and $-2.5 > \frac{X_{ak}}{Z} > -5$. It would seem that the most probable point is approximately at $\frac{R_{ak}}{Z} = 2$ and $\frac{X_{ak}}{Z} = -4$. However, a "comet's tail" of the measurement points can clearly be recognized which extends by $\frac{R_{ak}}{Z} = 8$, $\frac{X_{ak}}{Z} = 0$, to $\frac{R_{ak}}{Z} = 1$, $\frac{X_{ak}}{Z} = +2$.

Above 1 kc/s this anomaly diminishes. At 1.5 kc/s, it is just perceptible. At 2.2 and 3.4 kc/s, measurements give no more than a mass of points. But towards the lower frequencies the " comet tail " becomes steadily more evident. At 700 c/s, it reaches nearly $\frac{R_{ak}}{Z} = 0$, $\frac{X_{ak}}{Z} = 0$.

This is yet more marked at 500 c/s, while at 300 c/s, the comet tail finally turns into a " milky way". Figure 7 shows the circular diagram for 300 c/s. The measurement points, corresponding to the acoustic input impedances of the ears of the listeners are shown as black dots. Their dispersion has a systematic character which it would seem worth while studying in greater detail.

The acoustic input impedance of the ear is made up of the acoustic impedance of the eardrum, the acoustic impedance of the auditory canal and the acoustic impedance of the outer ear. It is

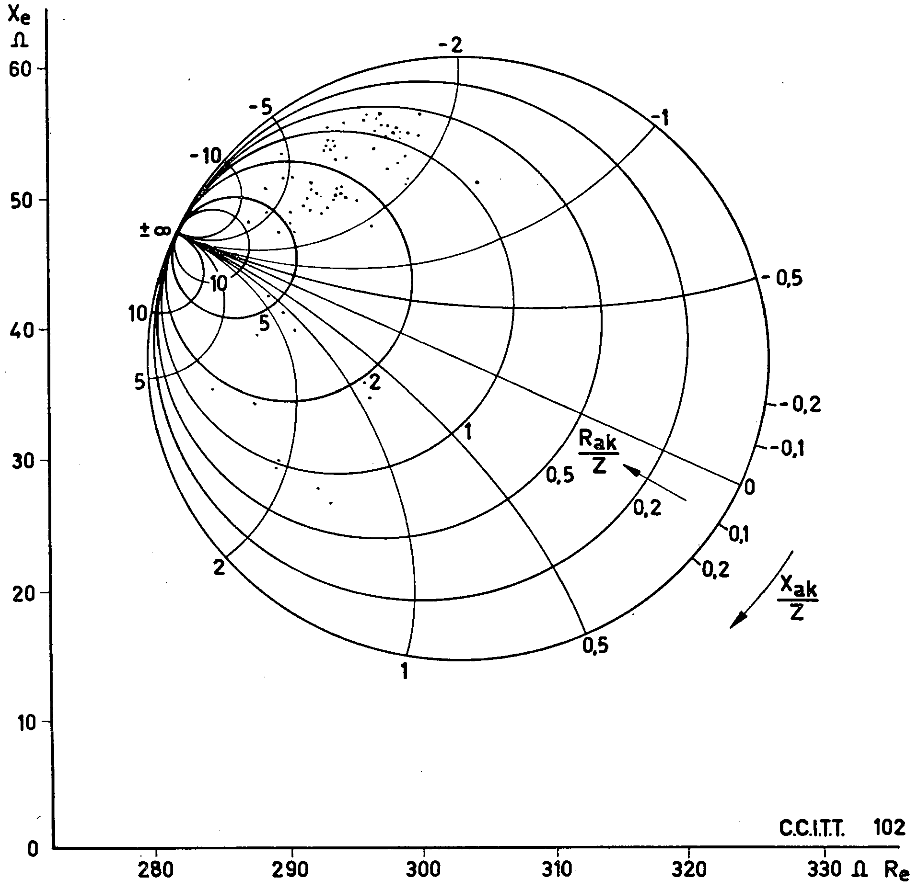


FIGURE 6. — **Representation of the right-hand acoustic half-plane in the electrical input impedance Z_e of the dynamic transducer at 1000 c/s**

The readings, taken on 41 people (82 ears), with the handset correctly applied, are shown as points

not likely that the input impedance to the eardrum or the length of the auditory canal will vary much between individuals at the same frequency. However, the form of the outer ear differs greatly between individuals. This is a factor which is decisive for the acoustic leak between the external ear and the handset, i.e., the transducer, and hence it was tempting to assume that the individual shape of the outer ear gave rise to widely varying degrees of acoustic leakage, which, particularly at low frequencies, is very important for the acoustic input impedance.

This assumption was confirmed by a series of measurements in which thin tubes, of increasing diameter, were placed between the outer ear and the handset, at the rim of the concha of the subject's ear, in such a way that an acoustic leak was artificially produced and was increased during the measurements. The results for both ears are shown in Figure 7 (circles connected by lines). The direction of the arrow shows the increasing diameter of the thin tubes, i.e. the direction of increasing acoustic leakage. In this way the importance of acoustic leakage at low frequencies was demonstrated. Corresponding measurements gave the same results at 500 c/s. For this

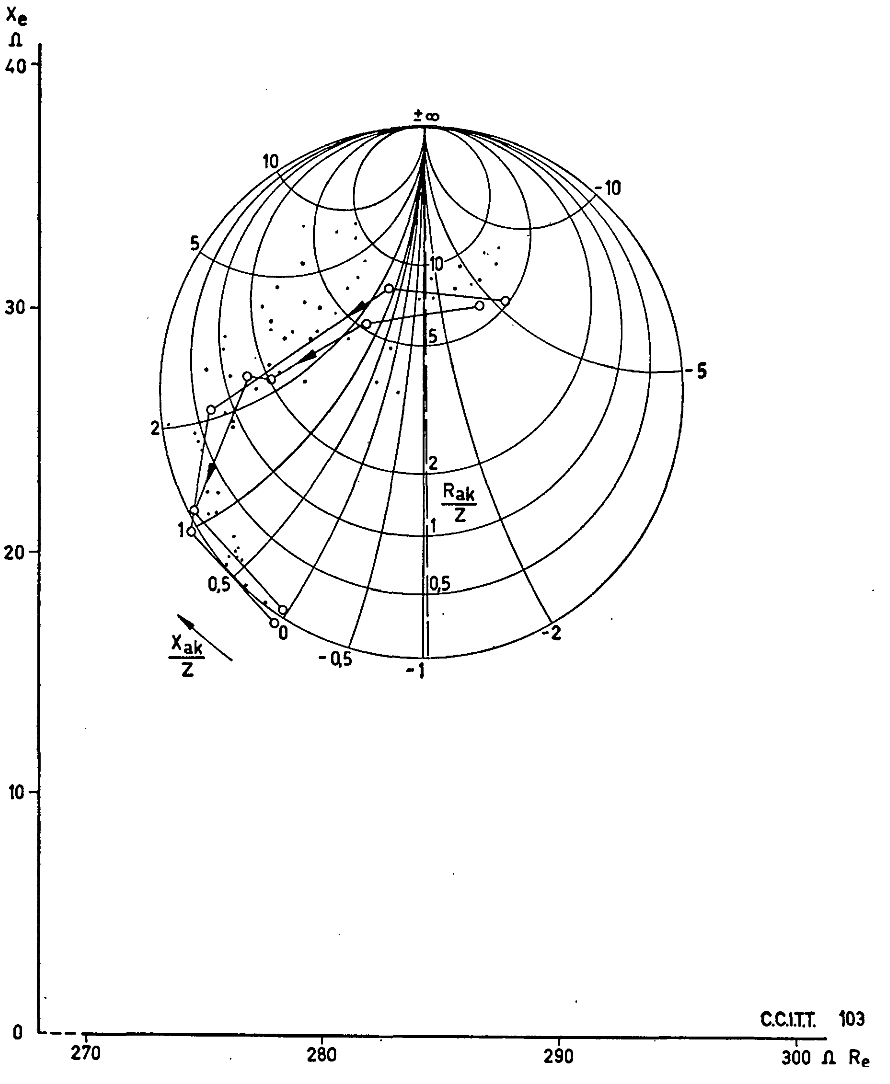


FIGURE 7. — **Representation of Z_{ak} on Z_e at 300 c/s**

The points represent readings taken from the human ear when the handset is correctly applied. The circles represent readings with leakages artificially produced between the handset and the ear. The arrows indicate the direction of increasing leakage

reason these measurements were repeated with the handset perfectly applied, in the hope of finding a clearer accumulation of measurement points.

b) *Perfect application of the receiver*

Figure 8 shows the figures obtained for 300 c/s with the handset correctly applied. It will be readily seen that the " milky way " has changed into a " comet tail " and that the effect of the acoustic leak has accordingly greatly diminished. Here we can indicate once more an area of

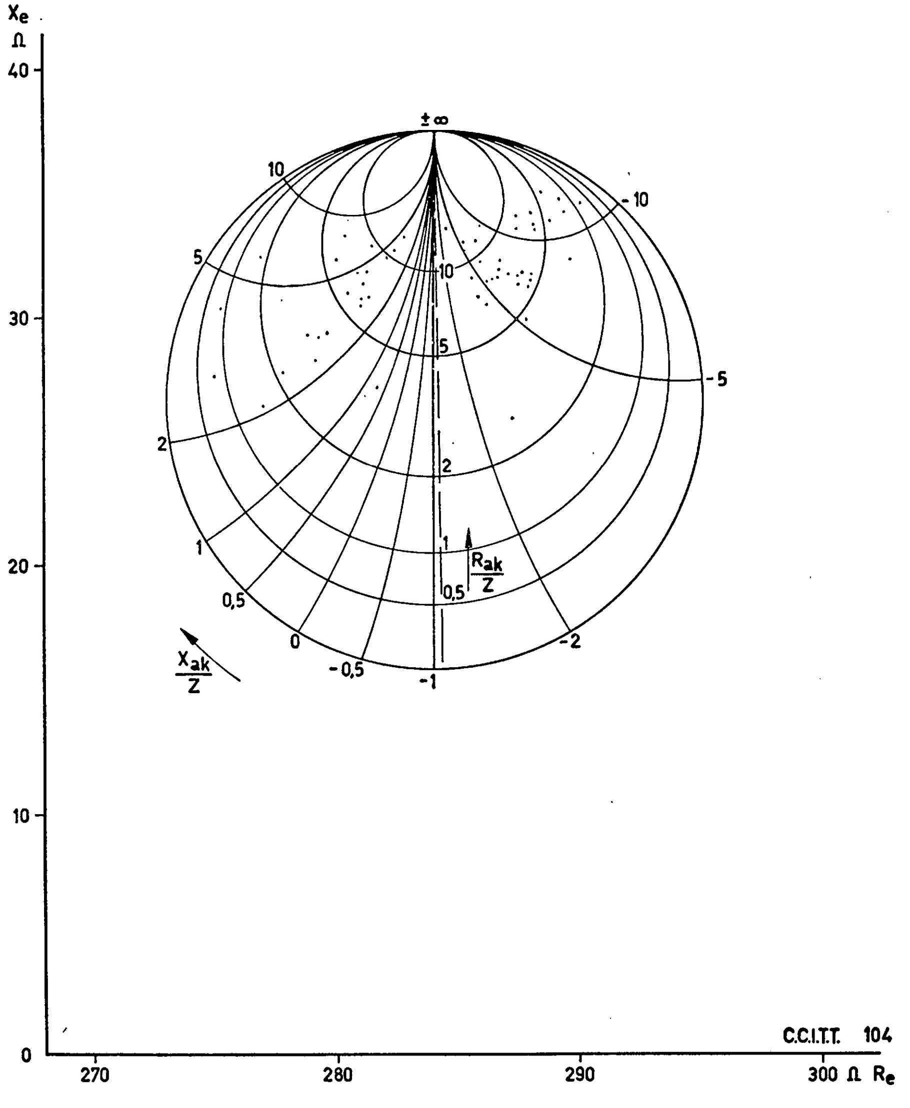


FIGURE 8. — **Representation of Z_{ak} on Z_e at 300 c/s**

The measurement points were obtained with the handset perfectly applied

accumulation at about $5 < \frac{R_{ak}}{Z} < 10$, $-5 > \frac{X_{ak}}{Z} > -10$. At higher frequencies this result appears equally clearly; the " comet tail " has practically disappeared at 500 c/s and has disappeared completely at 700 c/s.

c) *Dispersion and mean values*

Figures 9 to 11 show acoustic input impedances in the acoustic impedance plane for three frequencies, in the case of the perfectly applied receiver. In view of the considerable dispersion, we may well wonder whether the method is accurate enough, or whether inaccuracies during

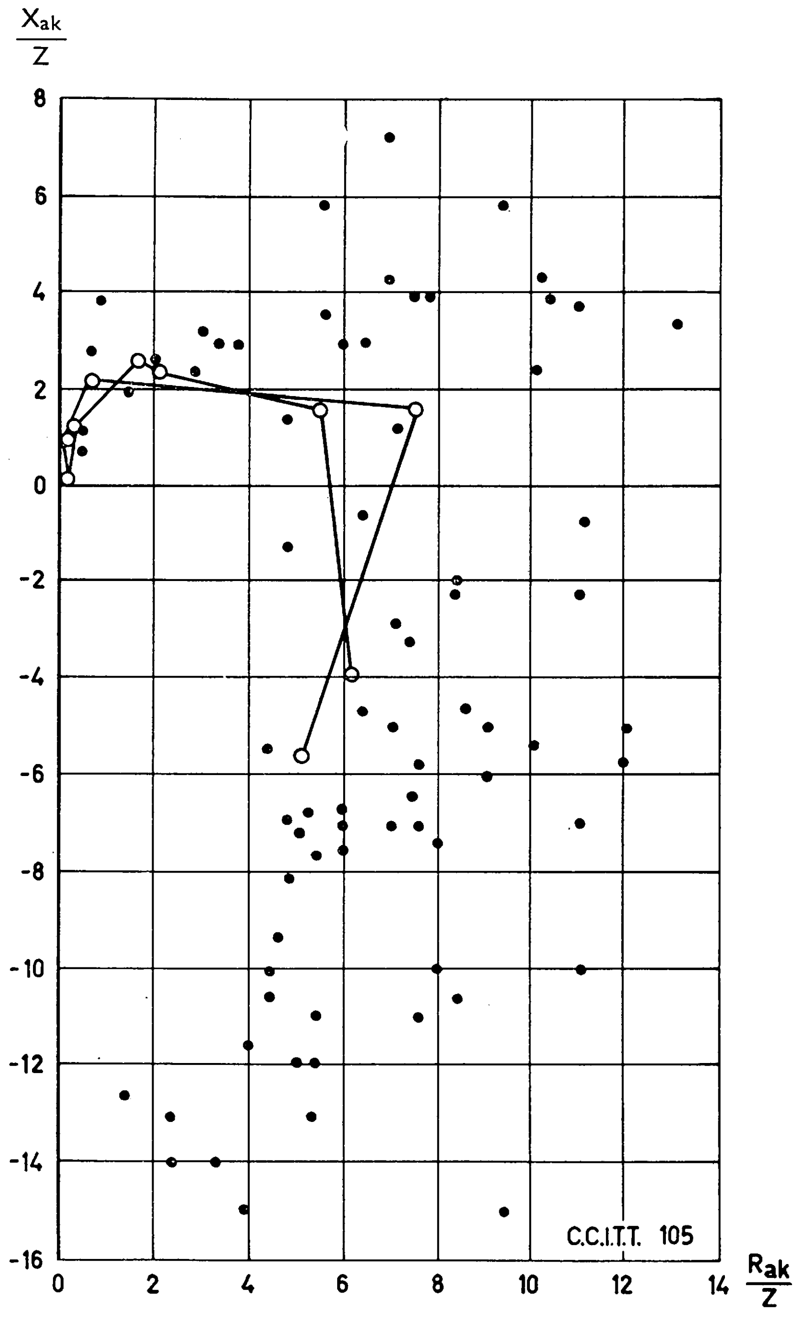


FIGURE 9. — **Acoustic input impedances at 300 c/s with the receiver perfectly applied**

The circles correspond to the values shown in Figure 7

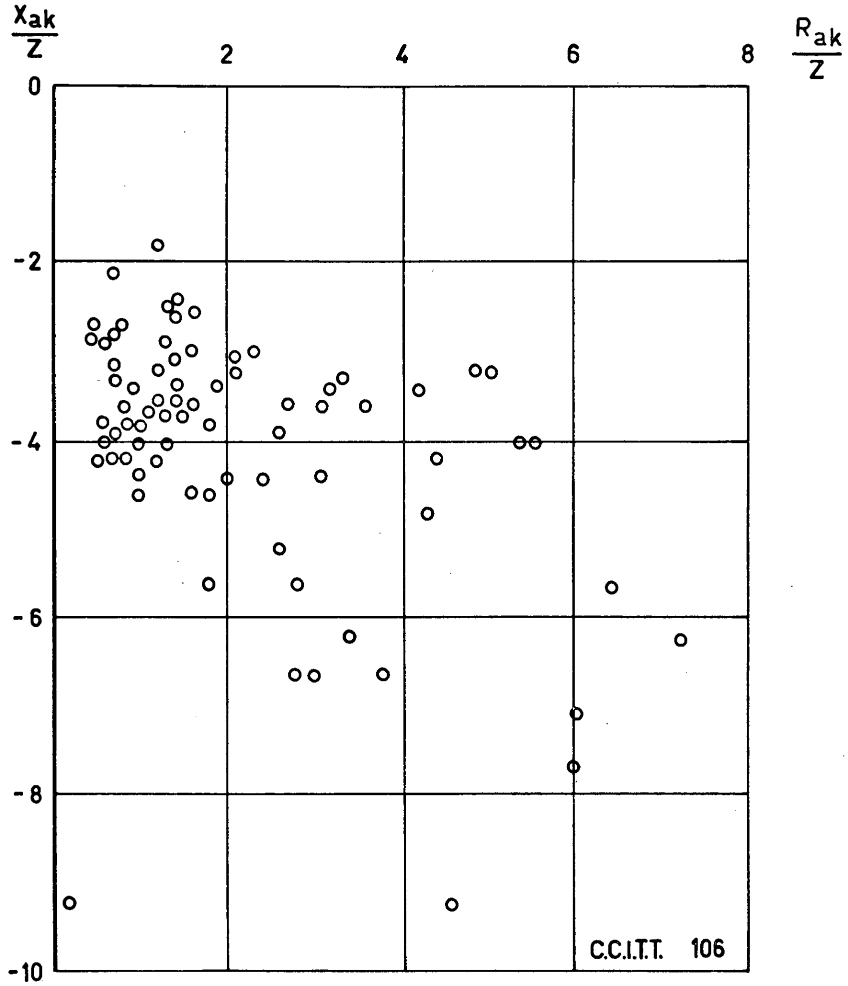


FIGURE 10. — **Acoustic impedances at 1000 c/s with the receiver perfectly applied**

bridge measurement could have caused so great a dispersion. An estimate of all influences shows that an over-all accuracy of about 3 per mille is obtained. This means that the accuracy is ± 1 ohm in the plane of the electrical impedance, assuming the resistance of the copper in the mobile coil is roughly 300 ohms. Since this accuracy holds good for the electrical plane and since there is considerable distortion in the scales of the acoustic impedances in the circular diagrams, the accuracy of the acoustic impedances largely depends on the magnitude existing in this particular instance. However, the accuracy always increases with efficiency, i.e. with the diameter of the circle. Since the diameter is smallest at 300 c/s (22 ohms, 54 ohms at 500 c/s, 64 ohms at 700 c/s, 46 ohms at 1 kc/s, 22 ohms at 1.5 kc/s, 25 ohms at 2.2 kc/s and 40 ohms at 3.4 kc/s), the accuracy is also the least. Figure 7 shows that some of the values are outside the circle $\frac{R_{ak}}{Z} = 0$, which should not, if the truth be told, be the case. The accuracy of the measurement method is clearly shown by these deviations of about 1 ohm. On the other hand, the considerable dispersion of the measured figures is due to individual anatomical variations, especially in the outer ear.

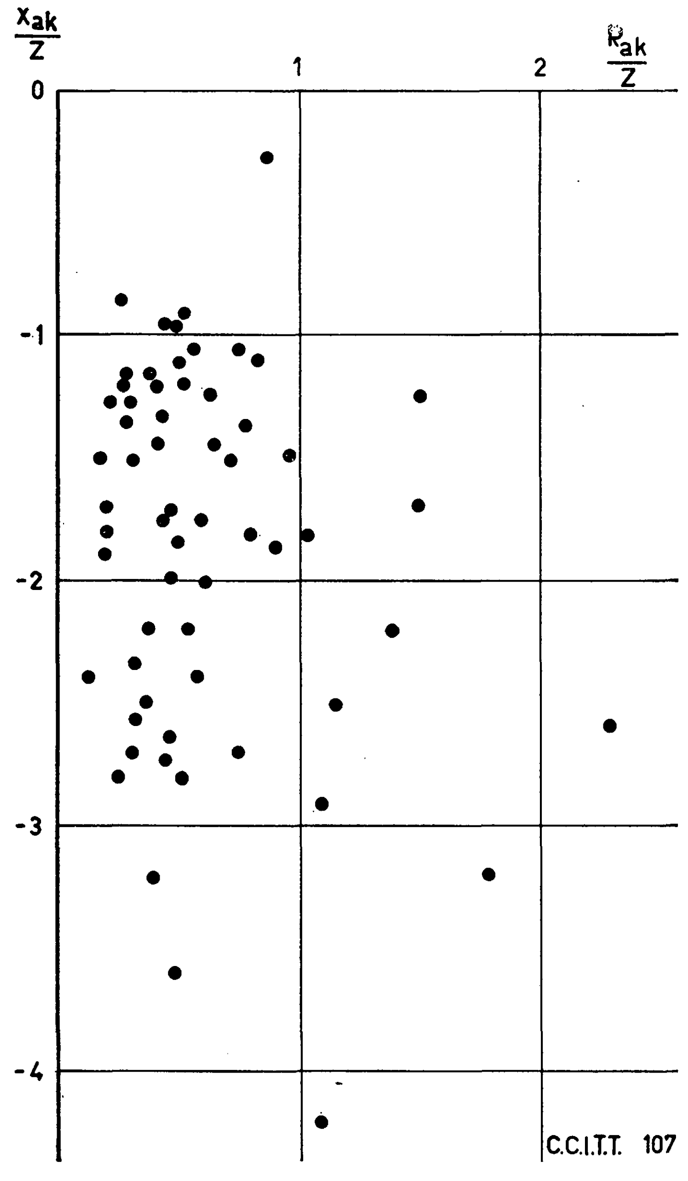
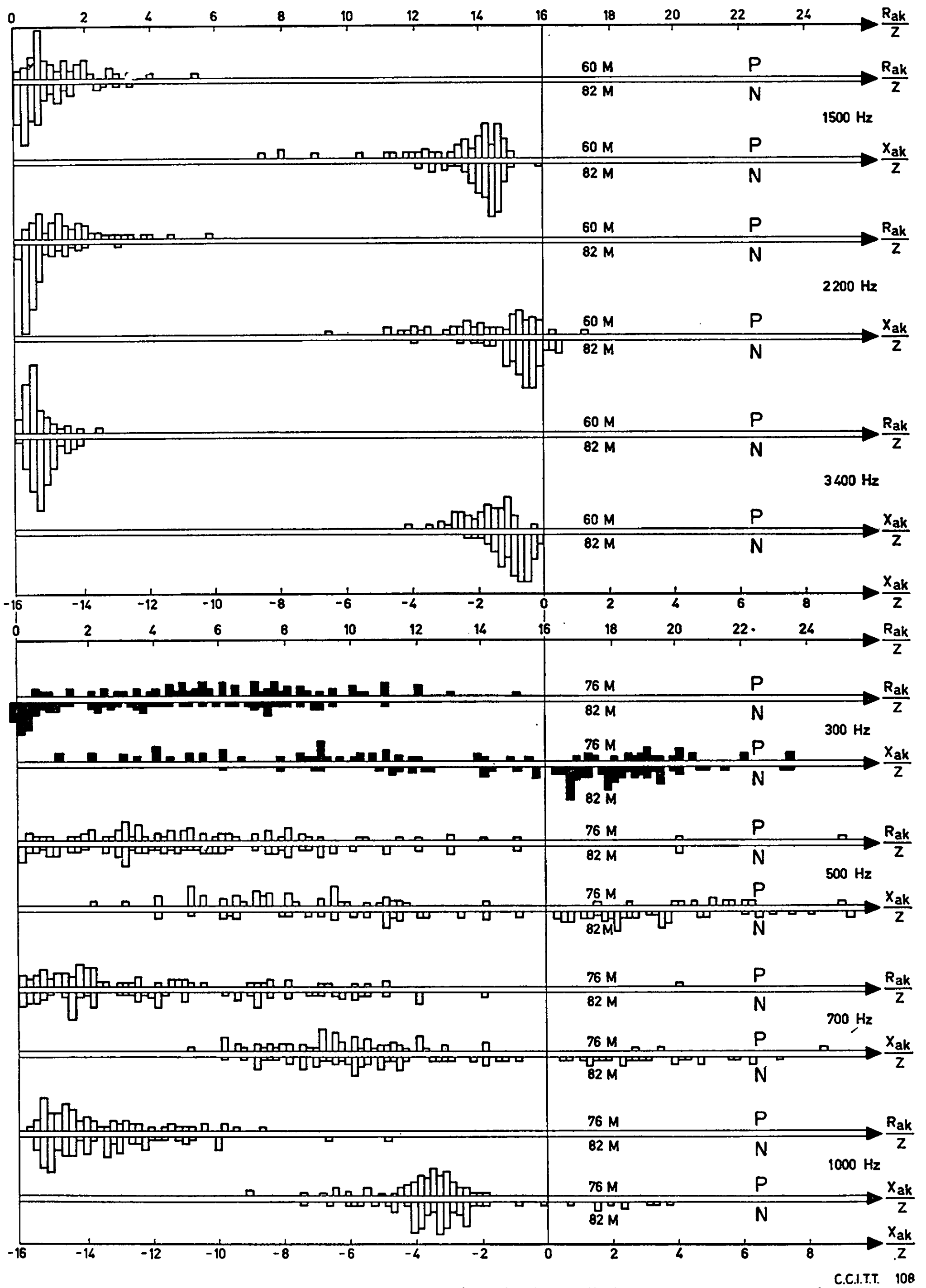


FIGURE 11. — **Acoustic input impedances at 3400 c/s with the receiver perfectly applied**

To facilitate estimation of the mean value and so as to be able to make comparison with measurements made with the receiver normally applied, the impedance components are separately shown in Figure 12, for the real and the imaginary components. The average figure obtained from Figure 12 for the handset correctly applied is shown in Figure 13 in the acoustic impedance plane as a function of frequency.

The formation of the mean value seems of little significance at the low frequencies when the receiver is normally applied. At the high frequencies, the figures for the perfectly applied receiver



M = Measurement points *P* = Perfectly applied *N* = Normal

FIGURE 12. — **The dispersion of real and imaginary components of acoustic input impedances of human ears measured with the handset normally applied and perfectly applied**

and the normally applied receiver almost overlap (see Figure 12). When the receiver is correctly applied the airgap between the handset and the ear is clearly small, and this may be considered one reason why the input impedances are also shifted to the high frequencies.

d) *Artificial ears*

Artificial ears were terminated not by the condensor microphone with a rigid termination in the particular band, but by an exactly matched brass cylinder. When the acoustic input impedance of artificial ears was being measured, the brass cylinder (weighing 600 grammes) was used to produce the application pressure. The handset, with transducer, was mounted in a fixed position with the orifice directed exactly upwards. The artificial ear was applied exactly and was centred on the brass cylinder. The artificial ear was measured in accordance with Braun [1] and the A.S.A. artificial ear [2]. Figure 13 shows the outcome. The readings obtained were noted several times to check the equipment. They are quite reproducible because, in measurement, there is no warmth transmitted by the hand and hence there is no variation in the temperature of the transducer.

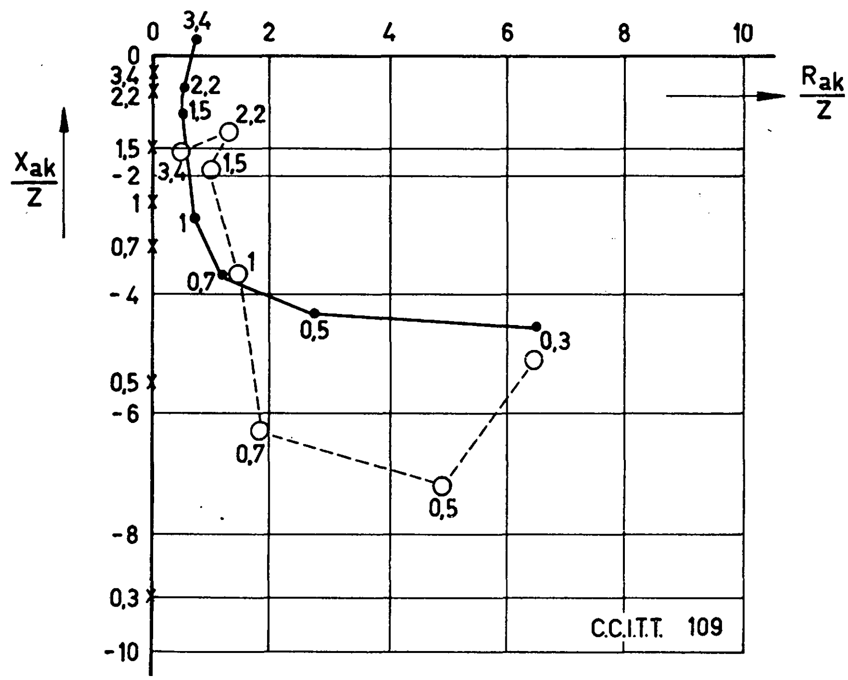


FIGURE 13. — **Comparison between the mean values (circles) obtained from measurements made with handset perfectly applied and the input impedances measured on artificial ears (... according to Braun; xxx according to A.S.A.)**

Discussion of the results obtained

The acoustic input impedance of the ear as encountered by the handset in the form of a terminal impedance during a telephone call depends to a great extent on how the subject applies the handset to his ear. Acoustic leaks are especially noticeable in the input impedance at low frequencies. Since, when the sound level is low, the user presses the handset strongly against his ear, it would seem proper to base a tolerance diagram on the input impedance measured on a receiver perfectly applied.

If the transducer used for telephony has to be checked, it would seem logical to measure its transfer coefficient and to lay down tolerances for circumstances approaching those encountered in practice. This means that with a termination in the artificial ear the transducer embodied in the handset must exert a very strong acoustic pressure at the point where the eardrum would be. The artificial ear used should be of a size such that it imitates the acoustic input impedance of the human ear. Braun's artificial ear already meets these exigencies fairly closely, while the artificial ear described by the A.S.A. has a reactance and does not seem suitable for the use described.

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ANNEX 5

(to Question 12/XII)

Contribution by the Federal Republic of Germany

The acoustic orifice of the receiver earcap is small, and varies with the type of earcap. Hence the acoustic fields differ and depend on the size and shape of the output orifices. Usually the acoustic field will not be plane (the basis on which determination of the acoustic impedance reposes). The acoustic impedance of the ear thus depends on the shape of the acoustic field produced by the receiver. Since the acoustic pressure, in the acoustic orifice of the receiver earcap or at the entry to the ear, may depend on the point at which it is measured, the acoustic impedance will vary and is by no means easy to specify.

If the transducer return current coefficient is determined by Braun's method, this uncertainty disappears. In transmission terms, the absolute magnitude of the ear's acoustic impedance is of less interest than the return loss a_{st_2} (loss due to reflection) produced by the fact that the mechanical impedance of the receiver is not matched to the ear impedance.

The relationship between the composite loss [1] a_u of an electroacoustic transducer and the image attenuation coefficient, a, can be written:

$$a_u = a + a_{st_2}$$

[1] *Note by the C.C.I.T.T. Secretariat.* — It is useful to refer to the following general definitions which appear in the I.T.U. "List of definitions of essential telecommunication terms", Part I, 2nd impression, 1961:

05.20 "composite loss"
05.09 image attenuation coefficient
05.38 reflection loss
05.25 return loss (reciprocal of the ratio (g_{F_2}) as defined here)
05.26 return loss (here a_{F_2} is the real part of the return current coefficient)

The image attenuation coefficient can be determined by Braun's method [1], mentioned above, or according to the method described below, the latter procedure being based on measurement of the return current coefficient determined at the transducer input when the transducer radiates on an infinite mechanical impedance (" rigid " termination).

The return loss a_{st_2} can be calculated from the return current coefficient $\exp(g_{F_2}) = \exp(a_{F_2} + j\, b_{F_2})$, at the transducer acoustic output. The return current coefficient represents the ratio of the acoustic terminating impedance W_2, i.e. the acoustic impedance of the ear, to the acoustic image impedance Z_2 of the transducer. This ratio is:

$$\exp(g_{F_2}) = \frac{W_2 + Z_2}{W_2 - Z_2}$$

The following relationship obtains between the reflection coefficient $\exp(g_{st_2}) = \exp(a_{st_2} + j\, b_{st_2})$ and the return current coefficient $\exp(g_{F_2})$

$$\exp(-g_{st_2}) = 1 - \exp(-2\, g_{F_2})$$

The return current coefficient can be calculated from the return current coefficient at the electric terminals of the transducer by the equation:

$$g_{F_1} = 2\, g + g_{F_2} \tag{1}$$

where $g = a + j\, b$, g is the image transfer coefficient of the transducer.

If the acoustic output of the transducer is terminated by an infinite mechanical impedance (W_2 infinite), $g_{F_2} = 0$ and hence

$$g_{F_1} = 2\, g$$

The return current coefficient at the input is then equal to the image transfer coefficient of two identical transducers mounted in tandem, which is fully confirmed by experiment.

The return current coefficient $\exp(g_{F_1})$ is defined by the relation:

$$\exp(g_{F_1}) = \frac{W_1 + Z_1}{W_1 - Z_1}$$

in which W_1 is the electric impedance of the transducer when it is terminated on the acoustic impedance W_2 of the ear and Z_1 the electric image impedance when it is terminated by an acoustic impedance Z_2. The transducer is terminated on its acoustic image impedance Z_2 when its acoustic orifice is terminated by the acoustic orifice of a second identical transducer and when the latter has its electric terminals connected to the input of a third identical transducer.

The return current coefficient can easily be determined with the help of an impedance measurement bridge as in Figure 1. The measured impedances W_1 and Z_1 may be taken as the basis for calculating the return current coefficient $\exp(g_{F_1}) = \exp(a_{F_1} + j\, b_{F_1})$. This coefficient can also be measured directly by first of all balancing the bridge when the receiver is terminated on its image impedance Z_1. With careful balancing, the voltage U_0 is very small (Figure 1). When the transducer is terminated on the acoustic impedance W_2 of the ear, the voltage U_0 will increase. If the impedance of the voltmeter is very high, we can write:

$$\exp(g_{F_1}) = \frac{U_g}{2\, U_0}$$

where U_g is the generator voltage measured at the ends of the bridge. The logarithmic ratio of the half-voltage of the generator to the voltage U_0 gives the return current loss a_{F_1} and the phase differences between U_g and U_0 gives the phase-change b_{F_1}. The measurement of large return

[1] K. Braun: *N.T.Z.* 13 (1960), p. 365. The translation of this article constitutes Annex 5 to Question 15/XII.

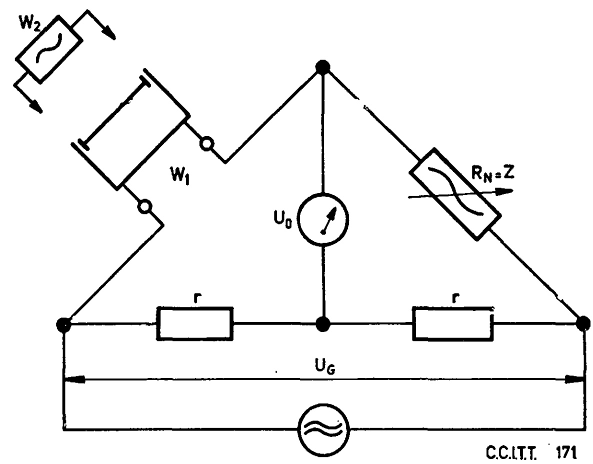


FIGURE 1. — **Diagram of the impedance measuring bridge**

current losses depends on the quality of the design of the bridge and the constancy of the temperature. The effects of temperature variations which may occur during measurements on the human ear can be largely eliminated if a check is made, before and after the measurement on the human ear, of the balance Z of the bridge. To reduce the effect of temperature variations, the lowest possible measurement voltages should be chosen, this also being more agreeable for the people taking part in the tests. The accuracy of the measurements can be increased by using an oscillator with a high frequency stability and a tuned voltmeter.

The return current coefficient g_{F_2} ascertained from the measurements for g_{F_1} and $2g$ (relation 1) will make it possible to evaluate the degree of agreement between the acoustic impedance of the human ear and that of an artificial ear, since g_{F_2} is a function of the ratio between the acoustic impedance W_2 and the image impedance Z_2 of the transducer. If, moreover, it is desired to ascertain the absolute acoustic impedance of the ear, the acoustic image impedance of the transducer must be known. For this purpose it is necessary to measure the voltage U at the electric input of the transducer and the acoustic pressure, p, on the surface of the orifice of the acoustic output of the transducer terminated without reflection by a second transducer. The acoustic image impedance is obtained by the relation

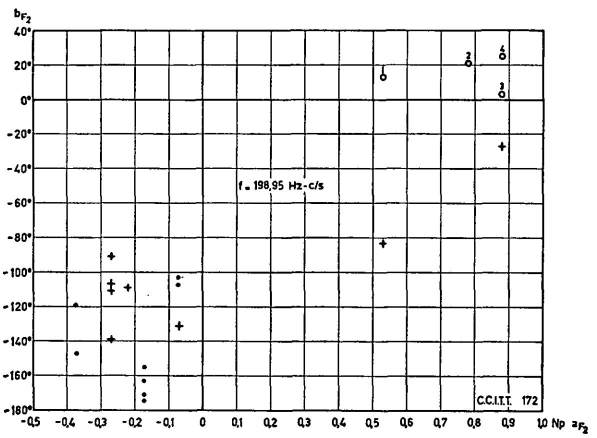
$$Z_2 = \frac{p^2}{U^2} \cdot Z_1 \exp (2g)$$

in which Z_1 is the electric image impedance, i.e., the input impedance in the case of termination without reflection, and g the image transfer exponent of the transducer. However, it is not necessary to know the absolute acoustic impedance of the ear to determine the composite attenuation. As has already been stated, the absolute acoustic impedance of the ear is not univocal because the acoustic pressure depends on the acoustic field characteristics.

With this method, and using the chosen electrodynamic receiver capsules, the return current coefficient has been determined for the acoustic impedances of real and artificial ears. The sensitivity of the receiver capsules was very high. The image attenuation was of the order of 1.3 to 2 nepers in the frequency range 200 to 3500 c/s; in such circumstances this coefficient can be fairly accurately determined. Most of the measurements were made with the receiver cap usually used by the

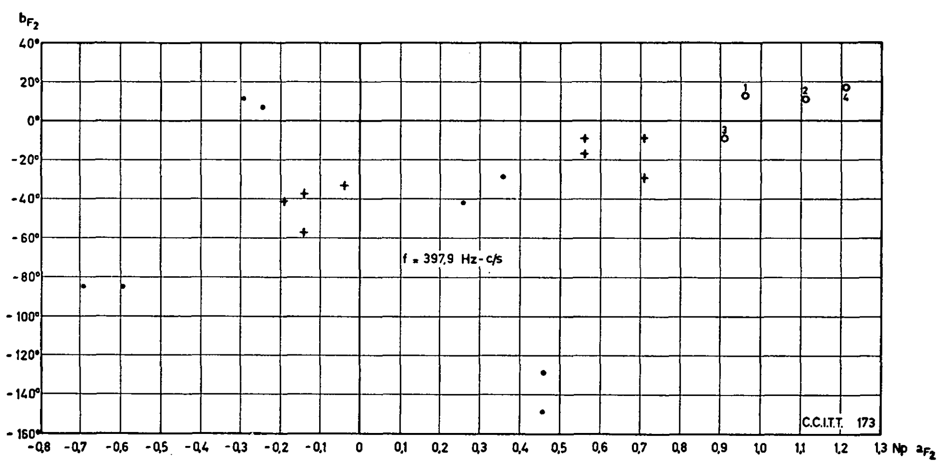
German Administration. This earcap has 8 holes of 3 mm diameter. Some of the tests were made with a receiver cap of a different type with an acoustic orifice of 7 mm diameter.

Figures 2 to 15 illustrate these results for the return current coefficient $g_{F_2} = a_{F_2} + j\, b_{F_2}$. The return current coefficient a_{F_2} is shown in nepers and the phase-change b_{F_2} in degrees. Figures 2 to 8 show the results obtained with the most common type of receiver cap (eight holes). The measurements were made with 1 woman and 3 men who applied the receiver twice, lightly or firmly, against the same ear. Measurements were also taken with four artificial ears, for comparison. The artificial ears 1 and 2 were damped, whereas 3 and 4 were not, and their design corresponded respectively to that of the N.B.S. 9 A coupler and the C.C.I.T.T. coupler.



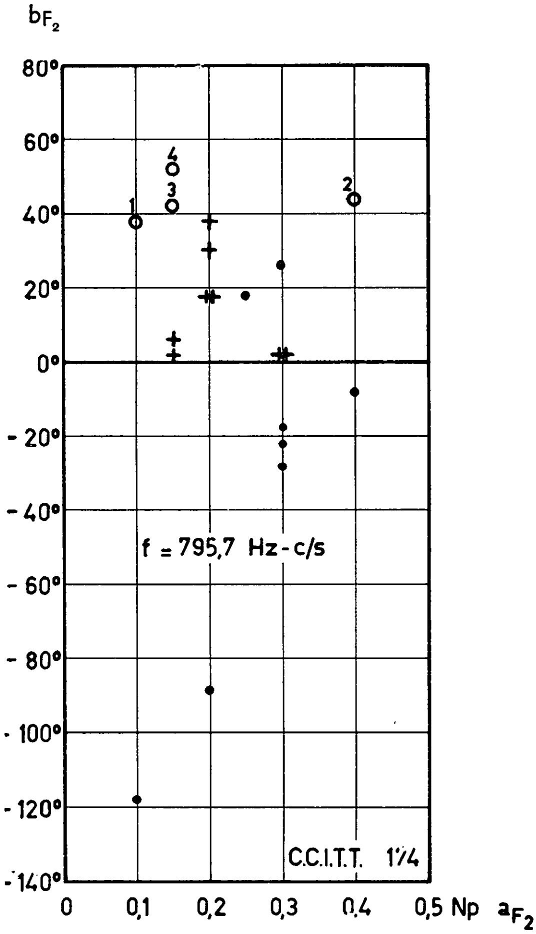
• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 2



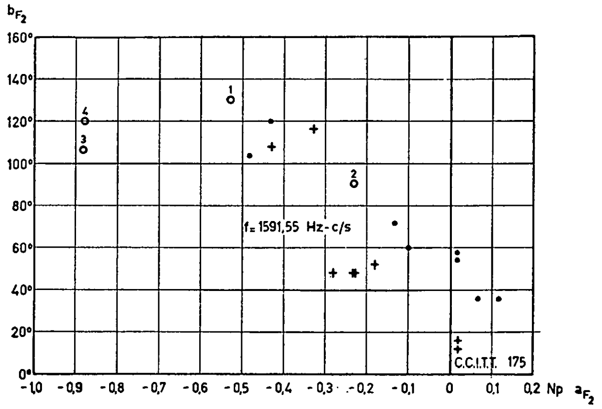
• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 3



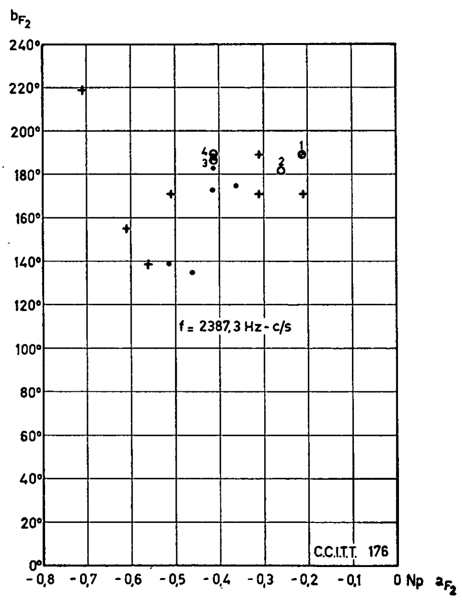
• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 4



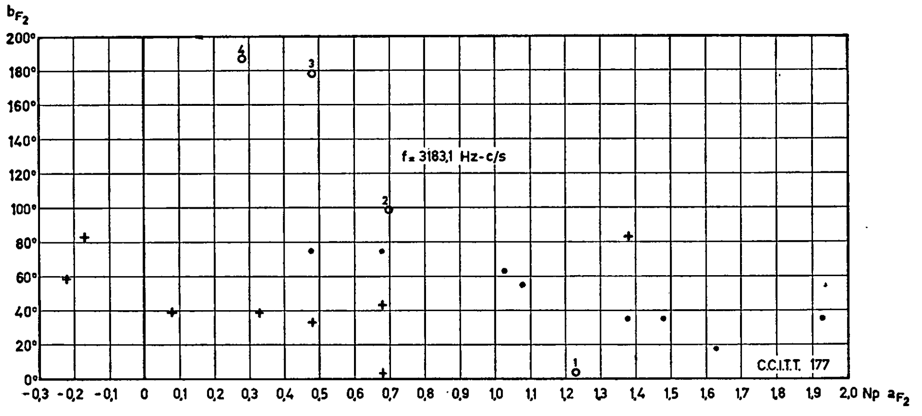
• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 5



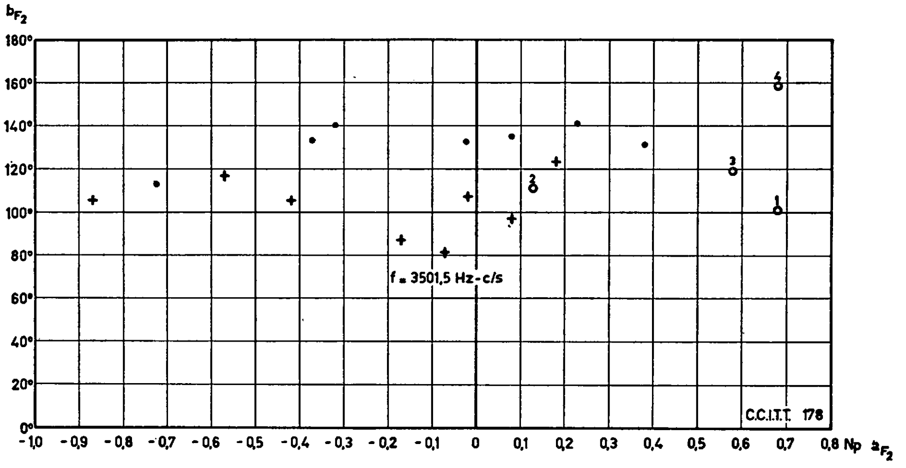
• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 6



• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 7

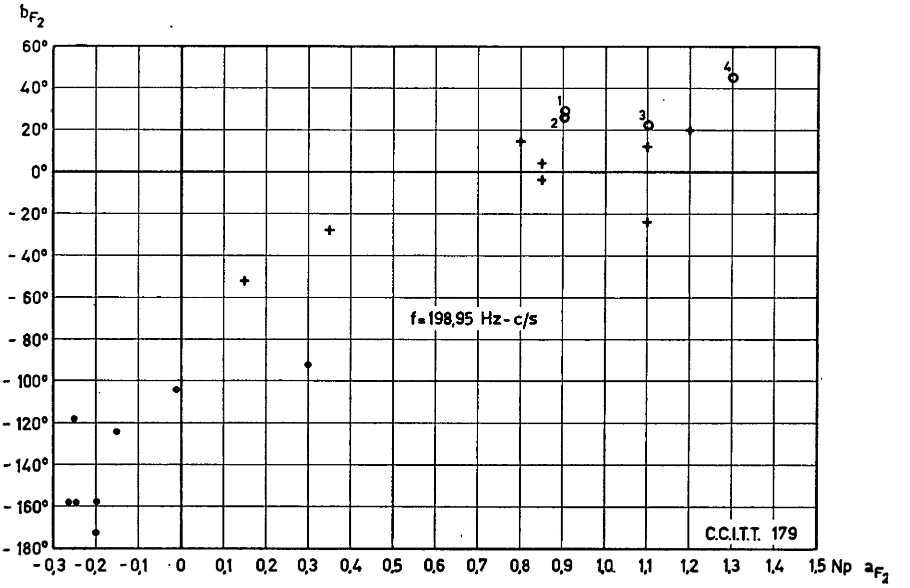


• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 8

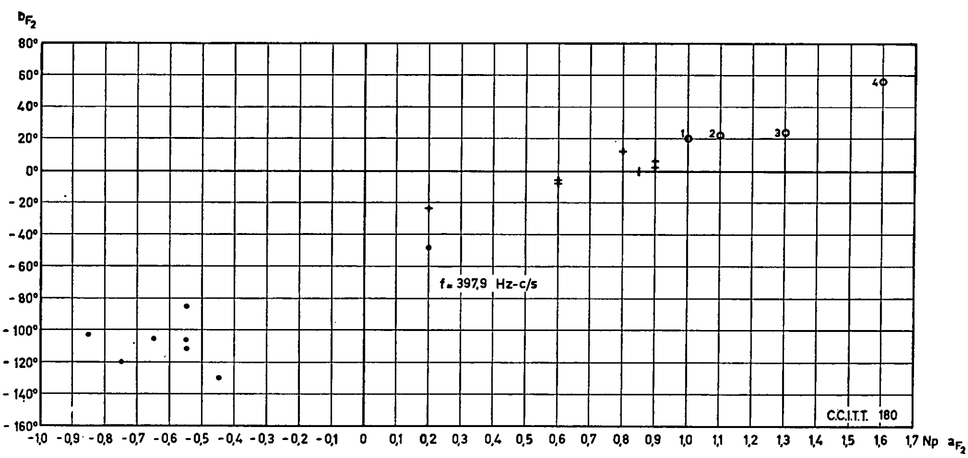
Figures 9 to 15 show the results obtained with the receiver cap with an acoustic orifice of 7 mm diameter. These tests were made with only one man and one woman; two measurements with both the right ear and the left ear were made each time with light and firm application of the receiver to the ear. The values achieved with the artificial ears were recorded for comparison.

The values obtained for the return current exponent g_{F_2} show that the acoustic impedance for the human ear is not a fixed quantity; it differs greatly from person to person. With the same person, it depends not only on the pressure with which the receiver is placed to the ear but also on the position of the receiver in relation to the entry to the ear canal. It is also affected by the shape of the ear-cap. Generally speaking, the return current loss is small, which means that the receiver is poorly matched to the ear impedance. The matching is better in the case of artificial ears. The return current loss varies greatly with the dispersion of the return current exponent. This might be one cause of the small differences in the hearing curves measured on different people.



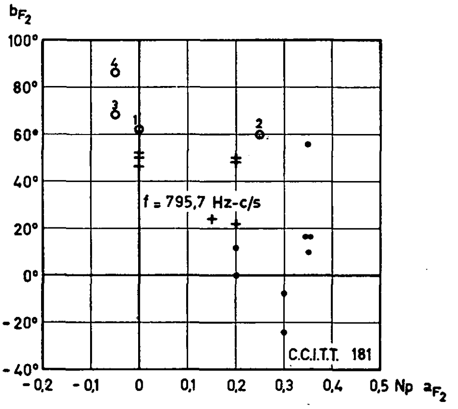
• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 9



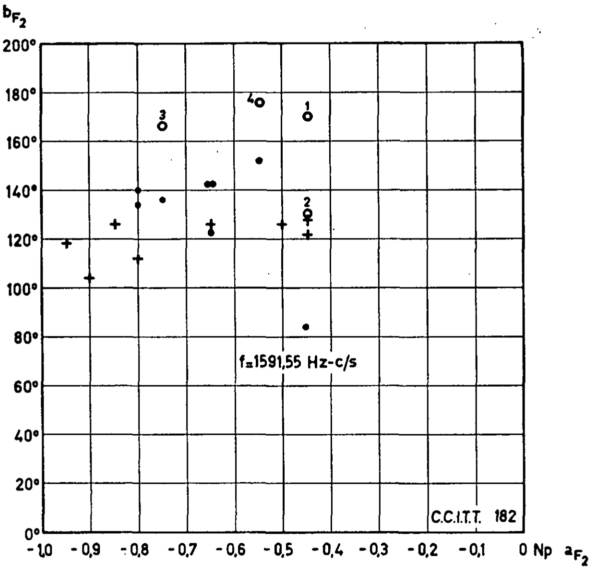
• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 10



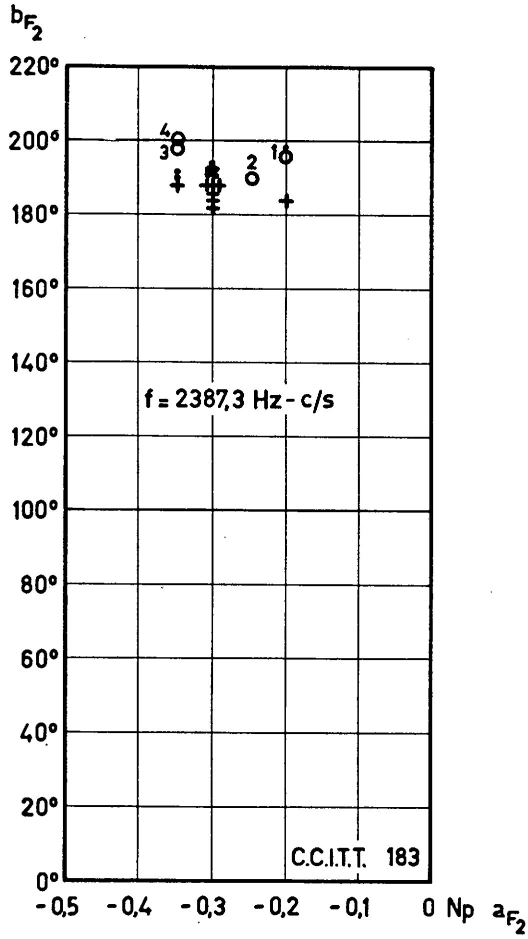
• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 11



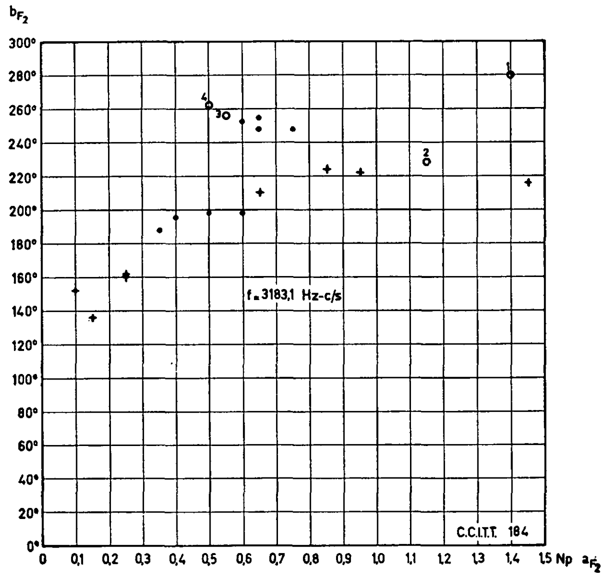
• = receiver applied lightly to the ear + = receiver applied firmly to the ear O = artificial ear

FIGURE 12



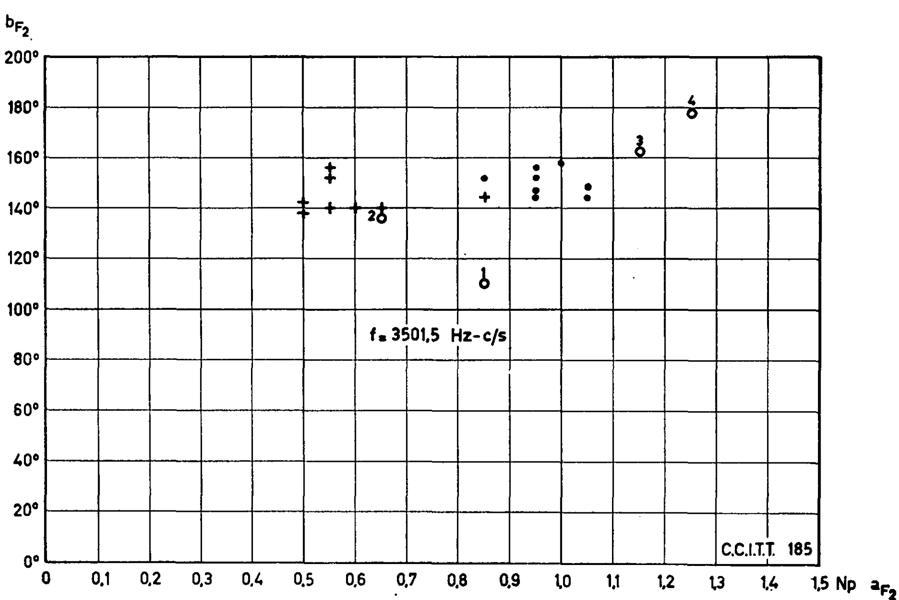
• = receiver applied lightly to the ear   + = receiver applied firmly to the ear   o = artificial ear

FIGURE 13



• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 14



• = receiver applied lightly to the ear + = receiver applied firmly to the ear o = artificial ear

FIGURE 15

ANNEX 6

(to Question 12/XII)

A study of a probe microphone

(Note by the Italian Administration)

(See former Annex 2, pages 677 to 689, Volume V of the *Red Book.*)

ANNEX 7

(to Question 12/XII)

Results of some preliminary measurements to determine the external shape of an artificial mouth which is to present the same obstacle effect as that of a human head

(Contribution by the Chile Telephone Company)

(See former Annex 3, pages 690 to 696, Volume V of the *Red Book.*)

Question 13/XII — Non-linear distortion of telephone apparatus

(continuation of Question 13 of study group XII, 1961-1964)

(documentary question)

Collection of information:

1. On the effects which the non-linear distortion of a subscriber's telephone apparatus has on the quality of telephone transmission;

2. On methods of measuring the non-linear distortion of a subscriber's telephone apparatus; and

3. On the effects of carbon microphone noise in a subscriber's telephone apparatus on the quality of telephone transmission.

Note. — The documentation collected up to the present time is contained in Annex 26 (Part II of Volume V of the *Red Book*). The following Annex reproduces a new contribution to the study of this question. For additional information, Administrations may refer to Contribution COM 12—No. 38 (period 1957-1960) which is a contribution by the Japanese Administration concerning the factors liable to affect the natural sound of the voice in a transmission system.

ANNEX

(to Question 13/XII)

Contribution by the Italian Administration

Item 1

Transmission performance tests were carried out on a complete telephone communication in which the non-linear distortion and reference equivalent could be modified at will. The circuit diagram is shown in Figure 1, where it will be seen that the carbon microphone was replaced by a receiver capsule suitably amplified and corrected.

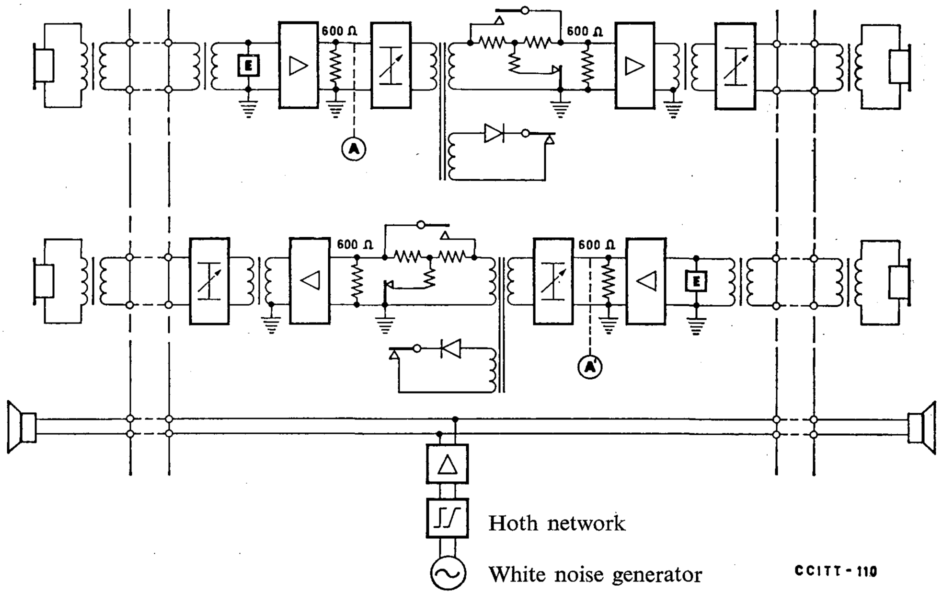


FIGURE 1

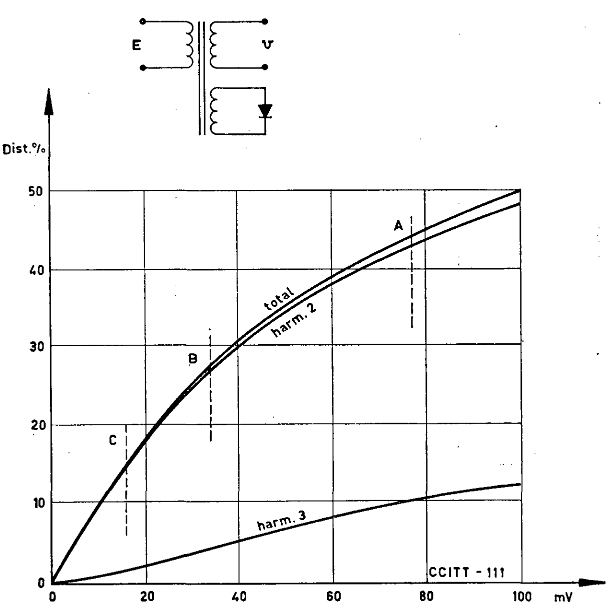


FIGURE 2

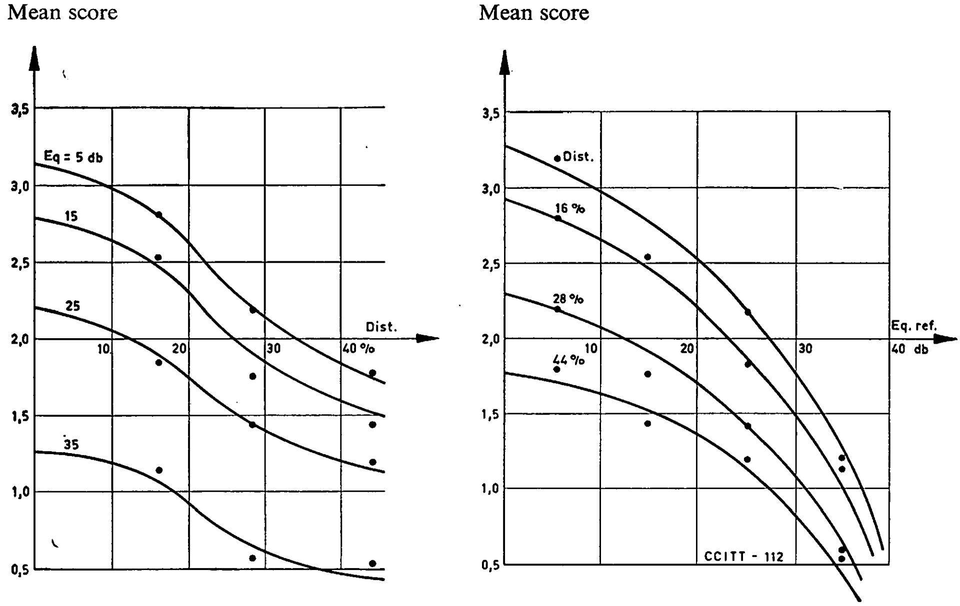


FIGURE 3

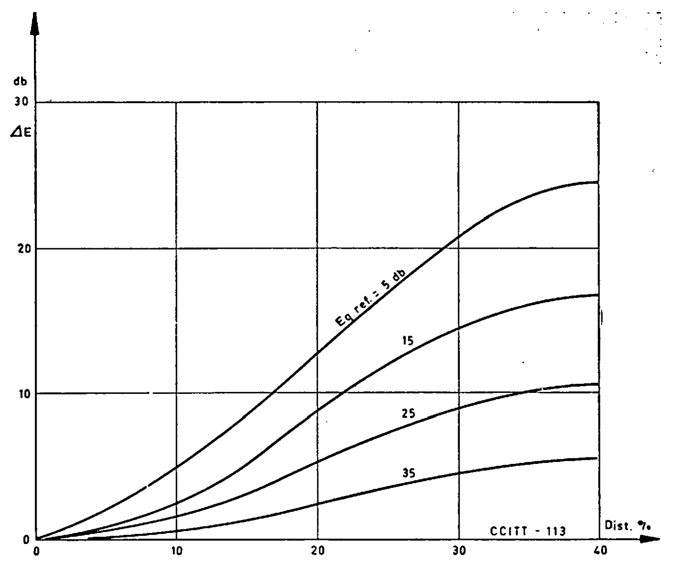


FIGURE 4

Figure 1 also shows the circuit producing the distortion, brought about with a rectifier inserted in the third winding of a transformer. The distortion depends on the level of the signal, in accordance with the curves shown in Figure 2. An attenuator at the input of the distortion circuit, with an amplifier and an attenuator at the output of this circuit, renders it possible to modify the distortion independently of the value selected for the reference equivalent of the complete circuit.

Opinion tests were performed with a 50-db room noise (Hoth spectrum), and the figures obtained are shown in Figure 3, which gives the mean score for various values of reference equivalent and distortion. Figure 4 shows the reduction in transmission performance, in db, as a function of distortion for various values of the reference equivalent.

Item 2

This point was studied with particular reference to the non-linear distortion produced by four kinds of microphone capsules in common use at the present time with a feeding current corresponding to conditions of use in a telephone set connected to a feeding bridge by a zero line.

The microphones were energized by a source of constant acoustic pressure at various frequencies; the pressures used were 10, 30, and 100 dynes/cm². The experimental circuit is shown in Figure 5.

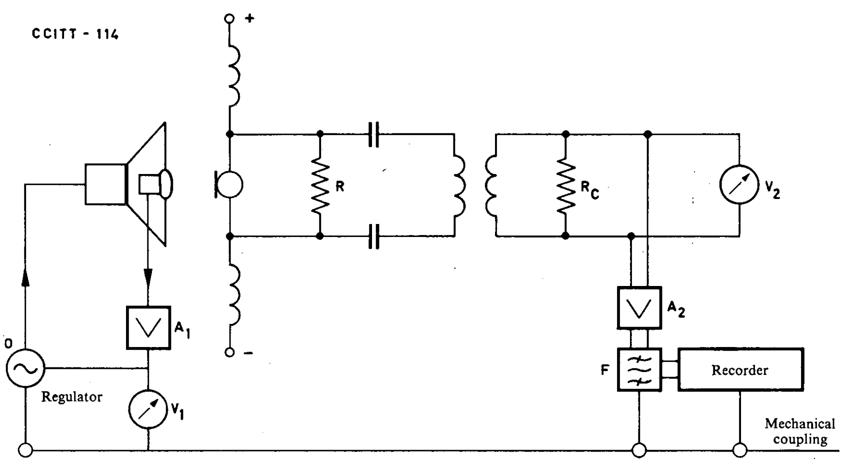


FIGURE 5

The following two methods were used:

1) Microphone energized at the fixed frequencies of 300 ,500 and 1000 c/s, care being taken to tap the microphone in a uniform manner at the beginning of the measurement which was performed as rapidly as possible. At the output of the circuit, the voltage of the fundamental and the second and third harmonics were measured.

2) Microphone energized with continuous sweeping of frequencies between 100 and 5000 c/s, lasting some 10 seconds. In this case no preliminary mechanical treatment of the microphone was necessary. The circuit output was connected to a level recorder preceded by a variable frequency band-pass filter, synchronized with the sweep rate, either in step with the sweep frequency or spaced from it by one octave or two octaves. Thus three representative curves were obtained for the fundamental, second harmonic, and third harmonic.

The figures obtained with these two methods are fairly constant and coincide fairly well for type A, while for types B, C, and D only the second method gave fairly constant results. Furthermore, type A gave higher distortion readings than types B, C, and D, which more or less coincide. The results are shown in Figure 6.

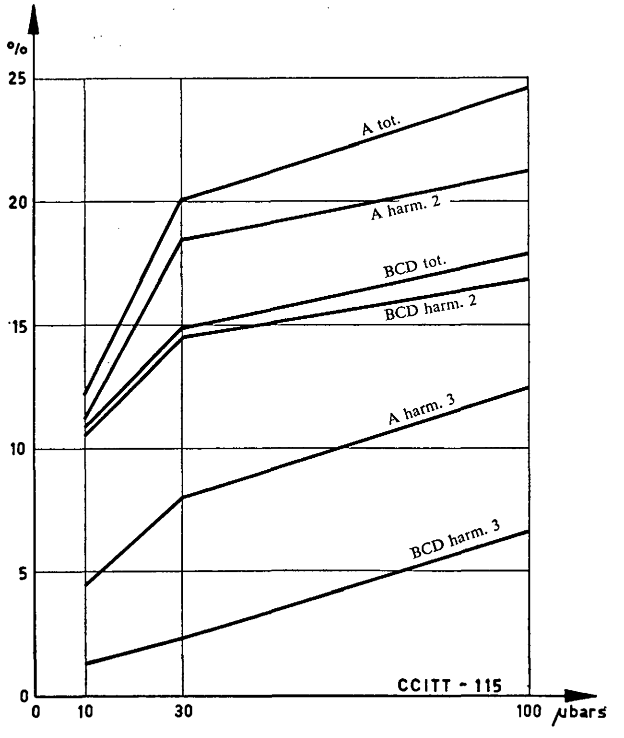


FIGURE 6

Item 3

The noise level produced by the carbon microphone in the above test conditions was measured and found to be, on the average, −76 dbmp, with a standard deviation of 2 db.

It is thought that, considering its effect on transmission performance, this noise, either on account of its level or on account of the speech level, may be regarded as a circuit noise.

In this context reference may be made to the previous contribution by Italy (Annex 1 to Question 4/XII).

Question 14/XII — C.C.I.T.T. Laboratory premises and equipment

(continuation of Question 14 of Study Group XII, 1961-1964)

a) Fitting-out of the C.C.I.T.T. Laboratory premises;

b) Equipment for the C.C.I.T.T. Laboratory.

Question 15/XII — Measurement of ratings based on loudness

(new question)

1. Several methods are used by various Administrations to measure loudness ratings in their laboratories, for example:

a) the reference equivalent defined by the C.C.I.T.T. and used in the C.C.I.T.T. Laboratory;

b) the O.B.D.M. equipment used by the Administration of the German Federal Republic and in the laboratories of other Administrations;

c) the E.A.R.S. equipment used by the A.T. & T. Company;

d) path through one metre air-path method, based on the A.R.A.E.N. and used by the United Kingdom Administration.

2. Can the results obtained from these diverse methods be related by theoretical considerations in respect of the following quantities:

a) over-all rating of a subscriber's set and line connected through an attenuator to another subscriber's set and line;

b) sending rating of a subscriber's set and line;

c) receiving rating of a subscriber's set and line;

d) sidetone rating of a subscriber's set connected to a given line?

If so, then what ought the relationships to be?

3. If theoretical relationships can be derived it would be very desirable to check that they apply accurately enough in practice. This would be best achieved by co-operation between the laboratories of Administrations and the C.C.I.T.T. Laboratory to secure measurement of various types of subscribers' sets by all these methods.

4. If no theoretical relationship can be derived, the only way of obtaining a useful empirical relationship would be by undertaking a series of measurements on different kinds of sets. A testing plan should be prepared for this purpose.

5. Study of this question should begin by assembling all available relevant information. Much of this exists in the C.C.I.T.T. documents and Administrations may have some information which has not been published.

Note 1. — By way of information, Annexes 1, 2, 3 and 4, below, describe methods by which quantities analogous to reference equivalents may be determined by means of objective measurements. In Annexes 27 to 29 of Part II of Volume V of the *Red Book*, descriptions are given of apparatus used by the Administrations of France, the Federal Republic of Germany and Switzerland, respectively, for the objective measurement of reference equivalent. Annex H (Part II of this volume) describes the portable telephonometer used by the Czechoslovak Administration.

Note 2. — Annex 5 describes a method developed by the Administration of the Federal Republic of Germany for measuring the image attenuation of microphone and receiver insets.

ANNEX 1

(to Question 15/XII)

Contribution by the American Telephone and Telegraph Company

Introduction

A previous response by the A. T. & T. Co. to this question included a published description of a revised telephone transmission rating plan based on the objective measurement of loudness [1]. It was recognized that, while other factors are important in engineering telephone sets and circuits, loudness loss is probably the best measure of transmission performance once a uniformly high standard of performance has been achieved in telephone instruments. At the time of publication of this revised plan, the method for making an objective measurement sufficiently well correlated with subjective loudness had not been fully worked out. A method has now been developed which gives very satisfactory results for the purpose of rating both through transmission and sidetone of the standard sets currently used by the Bell System. This method is similar in principle to that first proposed by K. Braun and H. Koschel [2].

This measurement method is called the Electroacoustic Rating System, (or E.A.R.S.). It is the method now used within the Bell Telephone Laboratories to evaluate the transmission of new telephone sets. While not used directly by the operating telephone Companies in the way visualized in the original article, E.A.R.S. will serve as the basis for setting standards of transmission and establishing rules for engineering and installing transmission facilities in the field. Its electroacoustic specification is briefly described below.

An electroacoustic rating system for telephone circuits

General

The measurement and rating of telephone circuits are complicated by the fact that many factors, including frequency response, sidetone, and non-linear distortion, can influence the performance from a user's viewpoint. Most rating systems have been based on the comparison of telephones and their interconnecting circuits with reference or standard circuits on a subjective basis. In the beginning, test subjects were asked to compare circuits with the reference on the basis of the loudness of speech received from a real talker. This is the way that reference equivalents are determined. As telephone instruments were improved in other respects simple loudness comparisons with the reference no longer gave a proper evaluation of the improvement. Measuring methods based on repetition counts were introduced as the means for subjective comparison. This led to the effective loss concept used in the Bell System during the transition to modern high quality telephone sets.

Now the transition is substantially complete, and impairment due to factors other than loudness has so diminished that they may be neglected in the basic rating procedure. Loudness is the main variable in plant design and it is possible to return full circle to rating by subjective comparisons with a reference circuit on the basis of loudness—to return to the reference equivalent concept.

A major difficulty with a rating plan based on an actual physical reference circuit is that the instrumentalities required eventually become obsolete and unobtainable. It is far better to specify the reference in terms of physical parameters which will always have meaning. If, in addition, comparison can be made objectively rather than subjectively, the advantages are really significant.

The freedom to return to a loudness criterion encouraged the development of a new rating plan based on an objective method of measurement. Such a plan was described in 1955 by Messrs. Blye, Coolidge and Huntley [1]. At the time that this revised rating plan was presented, the methods for objective measurement of loudness had not fully evolved. However, the basic principles were clearly stated in the 1955 paper. The following sections describe how the objective measurement method has been implemented to support the design and engineering of future telephone instrument and transmission facilities.

The rating plan

The starting point of the plan is an objective definition of an over-all rating of a telephone connection. Assuming that complex acoustic pressures and voltages can be measured in a way directly correlated with the subjective loudness sensation, the over-all rating is simply the ratio of the input acoustic speech pressure at the lips of a talker to the output acoustic speech pressure delivered by the telephone receiver to the ear of the listener. A rating of zero can be thought of an air path between lips and ear of zero length. The transmitting loop rating is arbitrarily defined as the ratio of the input speech pressure (in millibars) at the lips of the talker to the output voltage (in volts) across a 900-ohm resistive load. It follows that the receiving loop rating is the ratio of one-half the open circuit input voltage (in volts) from a 900-ohm resistive source to the acoustic speech pressure (in millibars) delivered to the ear of the listener by the telephone receiver.

The same definitions of send and receive ratings can be applied to a telephone set alone, and the results are called telephone set conversion losses. It is necessary to make such measurements over a range of line current since it will affect transmitter efficiency and equalizer losses. The rating at 100 mA is normally quoted as the rating of the set.

Subscriber line ratings are defined as the difference between loop ratings and telephone set conversion losses for the same line current. The division of loop ratings into separate set and line losses is of real value only in those special applications where the fixed relationship between loop current and line loss does not exist. In practice, it is most convenient to deal with the complete loop ratings. Experimentally determined transmitting and receiving loop ratings as a function of length and gauge of cable and loading are particularly useful in engineering studies.

In principle, the rating of a trunk is the difference between the over-all rating of a connection and the sum of the two loop ratings. For trunks which are relatively flat across the band from 300 to 3300 c/s and have an impedance close to 900 ohms, the rating can be satisfactorily approximated by the 1000-c/s loss. Significant departures from ideal can be taken into account by introducing corrections for attenuation distortion and reflection loss.

The objective measuring method

In discussing the rating plan, the assumption was made that complex acoustic pressures and voltages could be measured in a way directly correlated with the subjective loudness sensation. Loudness rating computation methods have been devised in the past which are very accurate as judged by subjective tests. These methods involve, first, dividing the total band into incremental subbands, and weighting the response of the circuit of interest in that subband by the normal speech spectrum and the known contribution of that subband to loudness. It was discovered that if the band is swept logarithmetically so that the sweep rate in c/s per second is proportional to frequency, each subband is given very nearly its proper weighting.

In the computation method, the weighted subbands are combined in a way which amounts to taking the 2.2 root of the response in the subband, summing, and raising to the 2.2 power. This same process can be accomplished physically by a fast-acting amplitude compressor with a ratio of 2.2 db to 1 db, and a long-time constant voltmeter with a suitable scale. It is this physical implementation of the computation method which serves as the basis of the design E.A.R.S.

The over-all system for determining the ratings required by the rating plan has two principal parts: an artificial voice and a compression type measuring system with both an electrical and acoustic input. In addition, a tape recording of the output of a transmitting telephone loop is used to provide the electrical drive for making receiving loop ratings.

The artificial voice is used as an acoustic source to energize the transmitter of a telephone system. It produces a logarithmic sweep frequency tone of constant reference amplitude of −42 dbmB at a specified reference point at its " lips ". The sweep is from 300 to 3300 c/s at a 6 c/s per second rate. The −42 dbmB level is used because it is about average human speech level. The voice is composed of a logarithmic sweep frequency oscillator, an artificial mouth with equalizer, an amplifier, and a transmitter conditioner. The latter is a 6-db attenuator which is switched out of the voice circuit for a few seconds as part of a " conditioning " procedure for a carbon transmitter just prior to measuring.

The measuring set is used to measure the acoustical or electrical output of an over-all or partial telephone system in terms of loudness loss. It is used also to calibrate the input to the system. The set is composed of two input circuits and a common measuring circuit. One input circuit, the acoustical, is composed of a 6-cc coupler and a condenser transmitter making an artificial ear and an amplifier. The other input circuit, the electrical, is simply a high impedance (12 000 ohms) input coil. Each of these input circuits is connected by a selector switch to the measuring circuit. The latter is composed of calibrating and measuring attenuators, amplifiers, a 2.2 to 1 compressor, and a linear integrating vacuum tube voltmeter. The voltmeter output may be bridged by a condenser to damp the 6-cycle vibrations of the indicator in order to facilitate accurate reading.

The function of the compressor is to compress all incoming signals in a ratio of 2.2 db to 1 db, corresponding to the required 2.2 root mentioned earlier. This it does with satisfactory accuracy over a wide enough amplitude range to cover the usual amplitude range of any regular telephone circuit. Beyond the working range of the compressor, the compression ratio reverts to 1:1. Therefore it is important that the level of the input signals to the compressor be regulated so that they are within the compression range.

The function of the vacuum tube voltmeter is to read the mean of the output voltages of the compressor. This means that the voltage actuating the meter indicator must vary linearly with the compressor output voltage (meter input voltage) over the whole amplitude range of this voltage. (This range, of course, is only 1/2.2 times the compressor input range.) Because of the amplitude range limitations imposed by the compressor and voltmeter, it is first necessary to select a voltmeter which fits the compressor. Secondly, in order to make an accurate calibration or measurement with the measuring circuit, it is necessary to adjust the calibrating or measuring attenuators until the voltmeter reads a certain critical value (called the reference value), which is near the centre of the satisfactory operating ranges of both the compressor and voltmeter. When this reference value is reached, the calibration is achieved or the rating measurement is read directly from the measuring attenuators. A direct-reading indicating meter for the present is impracticable for rating because of the limited useful amplitude ranges of the compressor and the voltmeter and because of slight deviations of the compression ratio from 2.2 to 1 over the compression range.

Based on this description, it should be quite evident how an over-all rating can be obtained using the measuring set. In making a receiving rating there is a question of what to use to provide the electrical input to the receiving section. A flat logarithmic sweep with an open circuit voltage of -36 dbV might be used. However, this would allow the edges of the band in typically flat

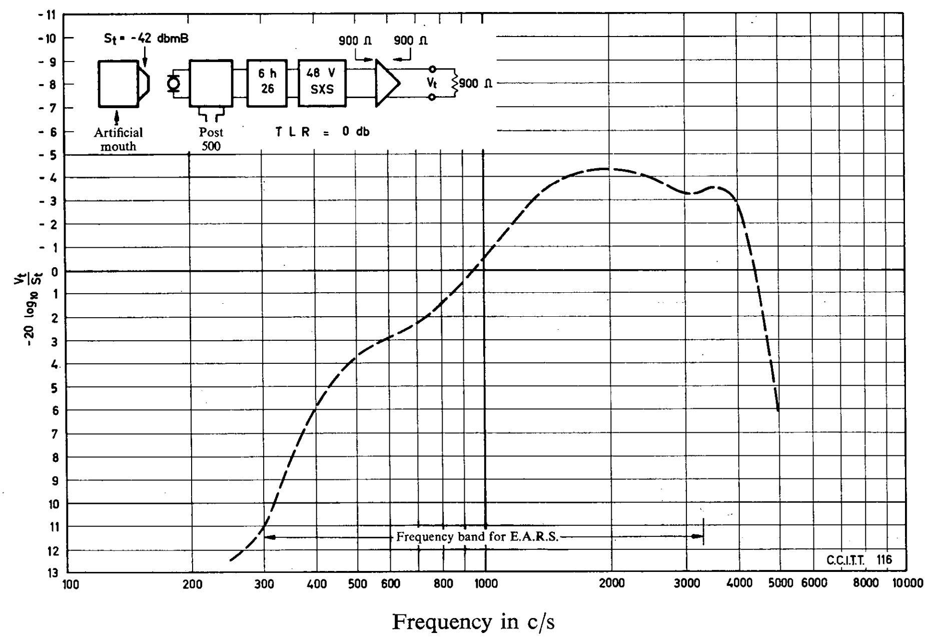


FIGURE 1. — **Response of source used for rating connecting circuit and receiving components**

receiving circuits to contribute more to loudness than they would if driven from a typically peaked transmitting circuit. Such receiving weightings added to transmitting ratings would not yield accurate over-all ratings. The solution is to drive the receiving circuit from a "typical" transmitting circuit. The transmitting circuit used and its response are shown in Figure 1. In practice, this transmitting circuit can be replaced by a tape recording to eliminate the instability of the carbon transmitter.

This principle of using a tape recording to obtain convenient and stable signals can be extended to the telephone sets included in any circuit to be rated. If the measurements are to accurately determine the effect of some line parameters, such as length of cable, the telephone set output must be quite stable. This can be accomplished by replacing the carbon transmitter with a tape recording of the output of a carbon transmitter energized by the artificial voice. In adjusting the level, due allowance must be made for the effect of line current on transmitter efficiency.

When a carbon transmitter is energized by an artificial mouth in making a measurement or preparing a tape recording, the handset is mounted in a special framework. In this framework the handset can be rotated for conditioning the carbon granules in the transmitter and then brought to rest at an accurate spacing from the artificial mouth. This spacing corresponds to the modal distance between transmitter and talker's lips for this particular handset. Further conditioning is applied by momentarily increasing the acoustic pressure by 6 db before the measurement.

Further work and application

The accuracy of E.A.R.S., as judged by the highly accurate loudness computation method mentioned earlier, is within about 1.5 db when measuring circuits with rather extreme distortion. It is quite adequate for use as a standard method of evaluating telephone sets and local transmission plans and is now being used for that purpose by the Bell Telephone Laboratories. One of the first major applications will be the characterization of the present loop transmission plant in terms of E.A.R.S. ratings. This study may suggest changes in present loop design rules to achieve a more optimum balance between cost and performance.

A refinement of the method is also being considered. The flat spectrum of the artificial voice may have a significantly different " conditioning " effect on the carbon granules of a transmitter from that of actual speech. This matter will be investigated by shaping the artificial voice spectrum and including the inverse shaping ahead of the compressor in the measuring circuit. If the results are different and significantly more consistent with subjective tests, the E.A.R.S. system may be modified. This change will have no effect on the rating plan and only a minor effect on existing rating data.

REFERENCES

[1] Blye, P. W., Coolidge, O. H. and Huntley, H. R.: A revised telephone transmission rating plan. *Bell System Technical Journal*, Vol. 34, pp. 453-472, May 1955 (reproduced in the *Red Book* of the C.C.I.T.T., Volume I, pp. 636-651, and Volume V, pp. 607-624).

[2] Braun, K. and Koschel, H.: The direct reading reference equivalent measuring set-up, and its importance in the improvement of telephone standards: *Fernmeldetechnische Zeitschrift*, Vol. 5, No. 10, 1952.

ANNEX 2

(to Question 15/XII)

Contribution by the Italian Administration to the study of objective methods of measuring reference equivalent

(See pages 624-637 of Volume V of the *Red Book*.)

ANNEX 3

(to Question 15/XII)

The " loudness function " and the calculation of relative equivalents for Japanese speech sounds

(Note by the Japanese Telephone Administration)

(See pages 637 to 662 of Volume V of the *Red Book.*)

ANNEX 4

(to Question 15/XII)

Contribution by the French Administration

(See pages 663-665 of Volume V of the *Red Book.*)

ANNEX 5

(to Question 15/XII)

Image attenuations of microphone and receiver insets

(Translation of an article by K. Braun published in *Nachrichtentechnische Zeitschrift*, 1960, No. 8, pages 365-370)

1. *Introduction*

In a telephone connection, the microphone and telephone receiver are the terminating elements in a long ladder network, made up of circuits and switching equipment. To plan and measure such circuits, we use the characteristic attenuations of the theory of quadripoles. Hence, we may readily suppose that a characteristic attenuation adaptable to the attenuation plan for the circuit could be used for the telephones too.

In 1926, the C.C.I.F. adopted a reference system called the S.F.E.R.T. (an abbreviation of the French). This may be considered as an ordinary telephone of which the transmission features are unambiguously fixed. It is used for comparison of telephone apparatus by voice and ear tests on a loudness basis. If the reference system transmits speech more loudly than does the telephone being measured, an attenuation, called reference equivalent, will have to be included in the reference system to obtain the same loudness, at the receiving end. Since a telephone is used for both speaking or sending and listening or receiving, we may distinguish sending and receiving reference equivalents. The reference equivalent can also be considered as mean value found subjectively of the transmission equivalent of the microphone or telephone receiver, relative to the S.F.E.R.T. reference standard. The mean transmission equivalent corresponds roughly to the transmission equivalent for 1000 c/s.

The reference equivalent is a relative measurement which gives no information regarding the actual attenuation of the telephone receiver. To find out what the reference equivalent of an ideal receiver would be, we need to know the image attenuation. A method is described below whereby the image attenuation and image impedances of an electroacoustic transducer can be determined. As long ago as 1943 and 1944, the author made a theoretical investigation into electroacoustic transducers considered as electroacoustic quadripoles [1, 2].

2. *Theoretical considerations*

a) *The transducer as a telephone receiver*

Here we shall consider a reversible transducer in the form of a telephone receiver inset. If the electroacoustic transducer be used as a telephone receiver (Figure 1), it is supplied with, on the electric side, electric power into its terminals, as a voltage U_1 and a current J_1 in amps. On the acoustic side, it delivers acoustic power at the output orifice as an acoustic pressure p_1 together with an acoustic flux φ_1.

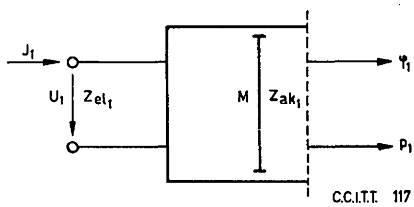


FIGURE 1. — **An electroacoustic quadripole as a telephone receiver**

Since, expressed in units of acoustic energy, the electrical power $UJ = p\ \varphi \cdot 10^{-7}$ when p and φ are measured in the C.G.S. system, i.e., p in μ bars and φ in cm^3/s, it is well to use the M.K.S.A. system, i.e. to express p in $N/\text{m}^2 = 10$ μbars and φ in m^3/s $= 10^6$ cm^3/s.

On the acoustic side, the output terminals must be replaced by the acoustic reference surface, which is limited by the edge of the receiver inset.

On the electric side, the ratio between the voltage U_1 and the current J_1 is determined by the image impedance Z_{el_1} when the output is terminated on the acoustic image impedance Z_{ak_1}. The corresponding ratio is to be understood inversely for the acoustic image impedance Z_{ak_1} as the ratio between the acoustic pressure p_1 and the acoustic flux φ_1.

Electroacoustic transducers are generally equivalent to electrically long quadripoles. Hence, their input impedances are practically independent of the output terminal impedance. In magnetic transducers (dynamic receivers and electromagnetic receivers), the electrical image impedance is determined by the ohmic resistance and the inductance of the coil, and in electrical transducers (condenser microphone, piezoelectric receiver) by the transducer capacitance. The acoustic image impedance is chiefly determined by the volume of air V_L of the cavity between the diaphragm M and the acoustic reference surface. As long as the dimensions of this cavity are small compared with the wavelength, the acoustic image impedance may be represented by:

$$Z_{ak} = \frac{\chi P}{j\, \omega\, V_1},$$

where χ = 1.4 (ratio of specific heats),

$P = 1.013 \cdot 10^6$ μbars (atmospheric pressure),

$\omega = 2\, \pi f$ (angular frequency), and V_1 the effective volume of the cavity.

As indicated in the Appendix, the volume V_1 results, roughly, from the parallel arrangement of the acoustic impedance of the cavity V_L and of the acoustic impedance of the diaphragm. Since the diaphragm impedance is purely capacitative only at low frequencies, the effective air volume decreases at higher frequencies.

The image transfer coefficient $g_1 = a_1 + j\, b_1$, of which a_1 is the image attenuation constant and b_1 the image phase constant, is given by:

$$e^{2g_1} = \frac{U_1\, J_1}{p_1\, \varphi_1} = \frac{U_1^2}{p_1^2} \cdot \frac{Z_{ak_1}}{Z_{el_1}}$$

The image attenuation a_1 is:

$$a_1 = \ln \frac{U_1}{p_1} + \tfrac{1}{2} \ln \frac{Z_{ak_1}}{Z_{el_1}} \text{ nepers or } 20 \log_{10} \frac{U_1}{p_1} + 10 \log_{10} \frac{Z_{ak_1}}{Z_{el_1}} \quad \text{db.}$$

If the telephone receiver be terminated on its acoustic image impedance Z_{ak_1}, the transfer coefficient, if a_1 be expressed in decibels, will be:

$$\ddot{u}_T = \frac{p_1}{U_1} = \sqrt{\frac{Z_{ak_1}}{Z_{el_1}}}\; 10^{\frac{70-a_1}{20}} \;\mu \text{ bars/V}$$

If we know the transfer coefficient $\ddot{u}_T$ and the image impedances, we can compute the image attenuation. The acoustic image impedance can be calculated with rough accuracy from the volume of air V_L of the cavity, from the relation:

$$Z_{ak_1} = \frac{1.42 \cdot 10^6}{\omega\, V_1}$$

If the reference surface of the acoustic output is terminated, not by the image impedance, but by a rigid partition or a condenser microphone with a very high acoustic resistance, we get the open circuit transfer coefficient:

$$\ddot{u}_{TL} = \frac{2\, p_1}{U_1} = 2\, \ddot{u}_T$$

If we know the open circuit transfer coefficient, the image attenuation and the electrical image impedance, we can calculate the acoustic image impedance Z_{ak}.

The image transfer coefficient is defined electro-acoustically as

$$U_T = 20 \log_{10} \ddot{u}_T = -a_1 + 10 \log_{10} \frac{Z_{ak}}{Z_{el}}$$

It is thus defined inversely to the usual form $20 \log_{10} \dfrac{1}{\ddot{u}_T}$, employed in transmission technique.

b) *The transducer as a microphone*

If the transducer, in the form of a receiving inset, is used as a microphone (Fig. 2), acoustic power is received at the acoustic input as an acoustic pressure p_2 and an acoustic flux φ_2, and delivered at the output terminals as a voltage U_2 and a current J_2. The corresponding image impedances of the quadripole are Z_{ka_2} and Z_{el_2}.

The image transfer coefficient $g_2 = a_2 + j\, b_2$ results from the relation

$$e^{2g_2} = \frac{p_2\, \varphi_2}{U_2\, J_2} = \frac{p^2_2}{U^2_2} \cdot \frac{Z_{el_2}}{Z_{ak_2}}$$

The image attenuation is

$$a_2 = 20 \log_{10} \frac{p_2}{U_2} + 10 \log_{10} \frac{Z_{el_2}}{Z_{ak_2}} \quad \text{db}$$

The logarithmic ratio $U_M = 20 \log_{10} \frac{p_2}{U_2}$ is denoted electroacoustically by the " transfer coefficient ". If it is terminated by its image impedance Z_{el_2}, the image transfer coefficient is

$$\ddot{u}_M = \frac{U_2}{p_2} = \sqrt{\frac{Z_{el_2}}{Z_{ak_2}}} \cdot 10^{-\frac{70 + a_2}{20}} \text{ V/}\mu\text{bar}$$

When the electrical side is open-circuited, we obtain the open circuit transfer coefficient

$$\ddot{u}_{ML} = \frac{2\, U_2}{p_2} = 2\, \ddot{u}_M$$

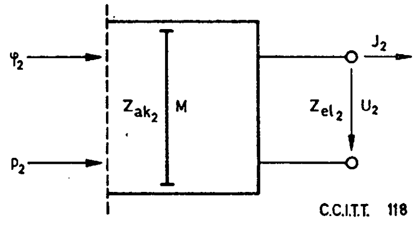


FIGURE 2. — **Electro-acoustic quadripole as microphone**

When the transfer coefficient and image impedances are known, it is possible to calculate the image attenuation. Inversely, when the transfer coefficient-image attenuation and electrical image impedance are known, it is possible to determine the acoustic image impedance.

If we calculate the transfer coefficients when the same transducer is used as a telephone receiver and as a microphone, the product of the transfer coefficient is

$$\ddot{u}_M \cdot \ddot{u}_T = 10^{-\frac{2a}{20}}$$

or the sum of the transfer indices $\ddot{u}_M$ and $\ddot{u}_T$ is equal to twice the image attenuation a.

c) *Interconnection of two transducers*

When two electroacoustic transducers are so interconnected that their acoustic orifices are in contact, we obtain an electrical quadripole the first half of which consists of a telephone receiver and the second of a microphone.

If U_1, J_1 and Z_{el_1} are the voltage, the current and the electric image impedance at the input of the quadripole, and U_2, J_2 and Z_{el_1} the voltage, the current and the electric image impedance at the output of the quadripole, its image transfer index $g_{12} = a_{12} + \mathrm{j}\, b_{12}$ is given by the relation:

$$e^{2\, g_{12}} = \frac{U_1\, J_1}{U_2\, J_2} = \frac{U_1^2\, Z_{el_2}}{U_2^2\, Z_{el_1}}$$

The image attenuation a_{12} is

$$a_{12} = 20 \log_{10} \frac{U_1}{U_2} + 10 \log_{10} \frac{Z_{el_2}}{Z_{el_1}} \text{ db}$$

The attenuation a_{12} is equal to the sum of individual image attenuations a_1 and a_2 when the acoustic image impedances Z_{ak_1} and Z_{ak_2} of both transducers are practically the same. Thus

$$a_{12} = a_1 + a_2$$

When both transducers have the same image attenuation, we have

$$a_1 = a_2 = \tfrac{1}{2}\, a_{12}$$

When the image attenuations on transducers 1 and 2 are not the same, the difference can be determined by using a third transducer, whose mode of operation and dimensions may differ from 1 and 2, since only a relative and not an absolute measurement is required.

If the third transducer is used as an auxiliary telephone receiver, transducers 1 and 2 are measured as microphones. If it is used as an auxiliary microphone, transducers 1 and 2 are measured as telephone receivers. An active transducer with amplification may also be used as an auxiliary transducer.

The difference Δa between attenuations a_{13} and a_{23} when transducer 3 is used as a microphone, or the difference Δa between attenuations a_{31} and a_{32} when transducer 3 is used as a telephone receiver, is equal to the difference between attenuations a_1 and a_2. We therefore have

$$\Delta a = a_1 - a_2 = a_{13} - a_{23} = a_{31} - a_{32}$$

We thus obtain

$$a_1 = \tfrac{1}{2}\,(a_{12} + \Delta a)$$

$$a_2 = \tfrac{1}{2}\,(a_{12} - \Delta a)$$

When the image attenuations a are known, the transfer coefficients can be calculated from the above-mentioned relations using the known image impedances. We find that the open-circuit transfer factor coincides with that determined by the reciprocity method; in this case, the attenuations a must be replaced by the logarithmic ratio of the input voltage and the input open circuit voltage, and the input voltage by the product of the current and the electrical image impedance [3, 4].

The image attenuation of the combination of transducers 1 and 2 can be found to an accuracy of 0.1 decibel by using known measurement methods (for example, with a reference circuit or a level measuring set). In this case, it is advisable to have transducer 2 terminate on a third transducer of the same type so that the impedance termination is correct.

To calculate the transfer coefficient, it is again necessary to know the image impedances. The electrical image impedance of the transducer combination may be measured with a high degree of accuracy by finding the input and output impedances. If both transducers of the same type have the same image attenuation, the input and output impedances are exactly the same.

To measure the acoustic impedance, it is advisable to use the method of parallel connection of a capacitive acoustic impedance $Z' = \dfrac{\chi P}{\omega \Delta V}$ which is inserted as supplementary volume ΔV between both acoustic image impedances $Z_{ak_1} = Z_{ak_2} = \dfrac{\chi P}{\omega V}$, where V is the volume of a transducer. The parallel connection or increase in volume produces an insertion loss Δa, which is such that

$$10^{\frac{\Delta a}{20}} = 1 + \frac{1}{2}\,\frac{\Delta V}{V}$$

From this, we obtain the volume required to determine the acoustic image impedance Z_{ak}

$$V = \frac{1}{2} \frac{\Delta V}{10^{\frac{\Delta a}{20}} - 1}$$

When the transducer is not terminated on an image impedance but on the resistance W corresponding to conditions of use, we obtain the service transfer coefficient $\ddot{u}_B$. Since transducers are usually electrically long quadripoles, the transducer output in the microphone can be replaced by a generator with its open-circuit voltage as e.m.f. and its electrical image impedance Z_{el} as internal impedance and, in the receiver, by a generator with an open-circuit acoustic pressure and acoustic image impedance Z_{ak}. For the service transfer coefficient, we obtain

$$\ddot{u}_B = \frac{\ddot{u}_L}{1 + \frac{Z}{W}}$$

where $\ddot{u}_L$ is the unloaded transfer coefficient.

3. *Results of experiment and calculations*

a) *Measurement of receiving insets*

A pair of identical moving coil receiver insets and a pair of identical electromagnetic insets with magnetic compensation were chosen for these measurements. Each unit of a pair had almost the same receiving reference equivalent and the same frequency characteristic. Both receiver insets in each pair were assembled in two special inset-holders, in a tube, in such a way that their acoustic orifices were close to each other. The electrical output of the pair was terminated by a third inset having the same electrical image impedance.

The voltage ratio between input and output of the pair of transducers was measured, using a high impedance level measuring set, as the level difference in nepers, this being equal to twice the image attenuation a of a transducer.

The difference in phase of the input and output voltage is equal to twice the image phase change b, and can be found with an oscillograph. To determine the effective volume V required to calculate the acoustic image impedance, the two receiving insets were separated from each other by a distance corresponding to the volume $\Delta V = 10$ cm³ and the insertion loss Δa was measured as the difference in level in nepers.

Figure 3 shows, for the range of telephone frequencies, the measurement results for the pair of moving coil insets; these include the image phase change b as well as the image attenuation a and insertion loss Δa.

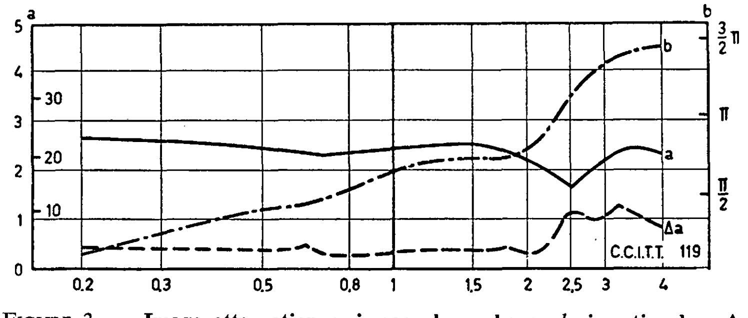


FIGURE 3. — **Image attenuation a, image phase change b, insertion loss Δa of a moving coil transducer**

Figure 4 gives the measurement results for the pair of electromagnetic insets.

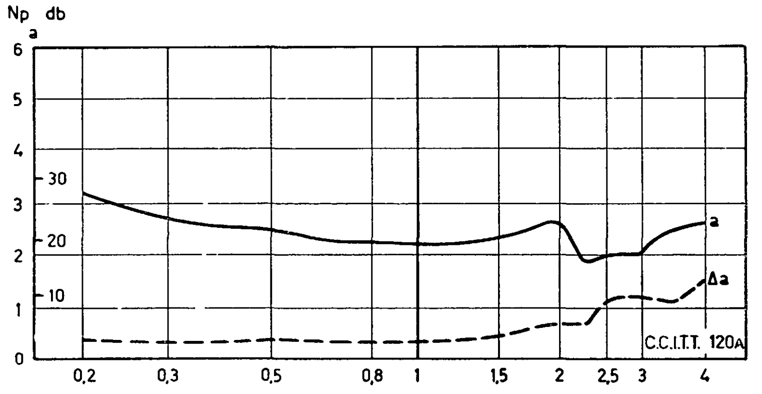


FIGURE 4. — **Image attenuation a, insertion loss Δa of an electromagnetic transducer**

The basic shape of the image attenuation curve a is similar for both transducers, although their electrical image impedance is quite different. The diaphragm resonance occurs at about 2400 c/s. The insertion loss Δa, for the moving coil transducer, is constant to about 2000 c/s and, for the electromagnetic transducer, to 1400 c/s and is equal to approximately 0.35 neper, which corresponds to an effective volume of 12 cm^3. Near the resonance frequency, the insertion loss increases. For an insertion loss of 1 neper, the effective volume is 2.9 cm^3.

The electrical image impedance Z_{el} and the acoustic image impedance Z_{ak} represent the moving coil transducer in Figure 5 and the electromagnetic transducer in Figure 6. The acoustic image impedance was calculated from the insertion loss and the supplementary volume ΔV.

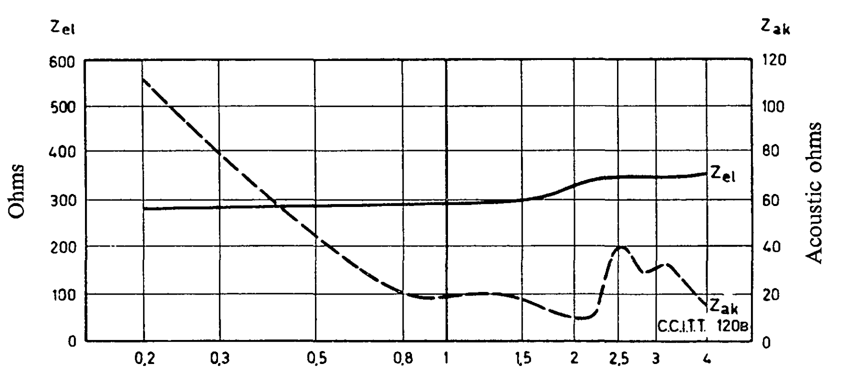


FIGURE 5. — **Electrical (Z_{el}) and acoustic (Z_{ak}) image impedances of a moving coil transducer**

For the moving coil transducer, the electrical image impedance is almost independent of the frequency, whereas for the electromagnetic transducer it increases substantially with the frequency.

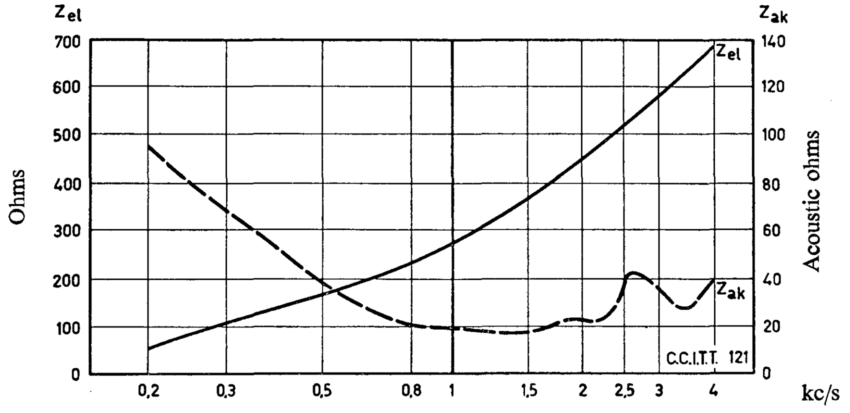


FIGURE 6. — **Electrical (Z_{el}) and acoustic (Z_{ak}) image impedances of an electromagnetic transducer**

The shape of the acoustic image impedance curve is identical for both transducers; this image impedance diminishes as a rule with the frequency, but increases again at resonance.

When the receiving reference equivalent of the receiver insets is measured not mounted in a telephone circuit but referred to 600 ohms, we obtain -0.85 neper for the moving coil inset and -1.1 neper for the electromagnetic inset. When mounted in the telephone set, the values of the reference equivalent are -0.46 neper and -0.73 neper. To compare the image attenuation with the reference equivalent, the values for 1000 c/s must be taken as a basis. For the moving coil inset, we obtain $a = 2.4$ nepers and, for the electromagnetic inset, $a = 2.2$ nepers. The difference between the image attenuation and the reference equivalent is therefore about 3.3 nepers. The receiving reference equivalent is found by subtracting 3.3 nepers from the value of the image attenuation. For an ideal transducer with an image attenuation equal to zero, the receiving reference equivalent not mounted in a telephone set would thus be -3.3 nepers.

b) *Measurement of microphone insets*

The image attenuation of a microphone inset can be measured by forming an electric quadripole with a receiver inset and the microphone inset and placing them with the acoustic orifices against each other. Since the acoustic image impedances of the microphone and receiver insets do not differ much, the image attenuation a_2 of the microphone inset can be found from the difference between the total image attenuation a_{12} of the transducer combination and the known attenuation a_1 of the receiving inset. The moving coil receiver inset is suitable for these measurements because its electrical image impedance, which is almost constant, equals 300 ohms. A modern carbon microphone with a resistance of about 60 ohms during modulation was used as the microphone inset. This microphone was fed by a distortionless feeding bridge of low attenuation.

The input and output impedances of the quadripole are 300 ohms and 60 ohms, respectively. When the input and output voltage level of the quadripole is measured, the total image attenuation a_{12} is obtained and the difference in voltage level is further adjusted by $\frac{1}{2} \ln \frac{60}{300} = 0.8$ neper. By subtracting the image attenuation a_1 of the receiver inset from a_{12}, we obtain the image attenuation a_2 of the microphone inset shown in Figure 7. The image attenuation a_2 is negative and therefore represents a gain averaging 4.4 nepers or 38 decibels.

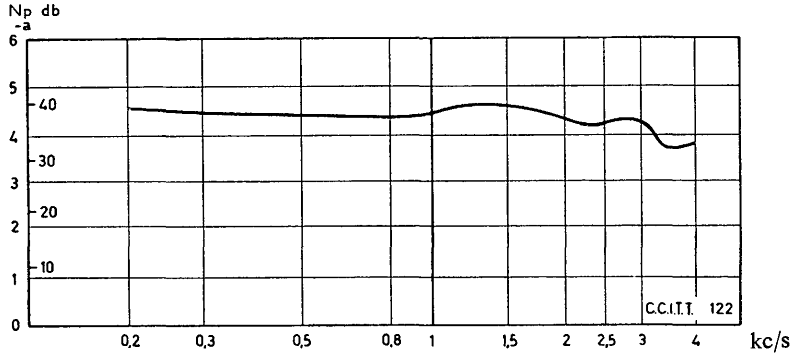


FIGURE 7. — **Gain $-a$ of a carbon microphone**

If, now, we measure the sending reference equivalent of the microphone inset, not mounted in the telephone circuit, but referred to a 600-ohm termination, a value of -1.2 neper is found. When mounted in the telephone set, the value of -0.2 neper is obtained.

The difference between the sending reference equivalent and the image attenuation is 3.2 nepers. This figure has been confirmed for other transducers, both passive and active. The sending reference equivalent is deduced from the image attenuation by adding 3.2 nepers. For an ideal passive transducer with an image attenuation of zero, the sending reference equivalent, not mounted in the telephone set, would hence be $+3.2$ nepers.

For the image transfer coefficient of the microphone inset, we get:

$$\ddot{u}_M = \frac{U_2}{p_2} = \frac{U_2}{U_1}\sqrt{\frac{Z_{el_1}}{Z_{ak_2}}} \cdot 10^{\frac{a_1-70}{20}} \text{ V}/\mu\text{bar},$$

where U_2 is the voltage at the output of the microphone inset when terminated on an internal impedance of Z_{el_2}, U_1 is the voltage at the input to the receiver inset, Z_{el_1} is the electrical impedance, and a_1 the image attenuation of the receiver inset in decibels. The acoustic image impedance is fairly steady for the receiver inset and the various microphone insets. Hence the ratio between the output voltage and the input voltage will be (within the limits of one constant) equivalent to the transfer coefficient of the microphone inset.

c) *Comparison with S.F.E.R.T.*

Using the transfer coefficient and the image impedances, the image attenuations for the S.F.E.R.T. sending and receiving systems can be calculated. The electrical impedance of the output of the sending system and of the input of the receiving system is 600 ohms. The acoustic image impedance is known from the volume. The S.F.E.R.T. microphone has, for calibration purposes, a volume of 9.1 cubic centimetres in the pressure chamber. The transfer coefficient will then be 26.6 V/μbars.

The S.F.E.R.T telephone receiver has a volume of 11.5 cubic centimetres. For calibration purposes, it is loaded with a volume of 18.3 cubic centimetres. The transfer coefficient is then 16.2 μbar/V. When loaded with its own volume, the transfer coefficient of the telephone receiver is 21 μbar/V. If, then, we make allowance for image attenuation, we find for the mean frequency of 1000 c/s an image attenuation of

$$a_S = -2.83 \text{ nepers} = -24.6 \text{ decibels for the transmitter}$$

$$a_E = +3.3 \text{ nepers} = +28.6 \text{ decibels for the receiver.}$$

The S.F.E.R.T. microphone is, when being used with speech, mounted in a holder with a cylindrical opening, causing a marked resonance at 3000 c/s. The transmission equivalent of the average field of the thus mounted microphone is increased by approximately 0.4 neper. For speech in a free acoustic field, $a_S = -3.2$ nepers should be taken as the basis for the transmitting system.

The differences of the image attenuations a for microphone and receiver insets measured by reference to the image attenuations of the S.F.E.R.T. sending and receiving system give the reference equivalents for sending and receiving. That is to say, for a microphone inset the sending reference equivalent

$$A_S = a + 3.2 \text{ nepers} = a + 28 \text{ decibels and}$$

for a receiver inset the receiving reference equivalent

$$A_E = a - 3.3 \text{ nepers} = a - 28.6 \text{ decibels.}$$

These figures agree with the reference equivalents measured for the various microphone and receiver insets.

The sum of image attenuation of the transmitting system and the S.F.E.R.T. receiving system is referred to the transfer coefficient of 4 decibels measured in the pressure chamber and the transfer coefficient of approximately zero in the free acoustic field.

The sending reference equivalent should not exceed 2.1 nepers or 18.2 decibels for the national sending system and 1.5 neper or 13 decibels for the national receiving system. This means that the image attenuation between the acoustic input of the microphone inset and the terminals of the international circuit should not exceed -1.1 neper or -10 decibels and between the international circuit terminals and the acoustic output of the receiver inset 4.8 nepers or 41.6 decibels. The image attenuation of a complete telephone link between the acoustic input of the microphone inset and the acoustic output of the receiver inset should not exceed 4.7 nepers or 40.6 decibels.

4. *Conclusions*

The conclusions which may be drawn from the above regarding measuring technique and the specification of transmission performance of microphone and receiver insets will be dealt with in a special article.

Although much progress has been made in improving the transmission performance of receiver insets, their image attenuation is still too great. In the case of very sensitive receiver insets, it is not lower than approximately 15 decibels, which should be considered extremely high in comparison with an attenuation of only 0.5 decibel for modern transducers. Although this figure could probably never be reached with a transducer, a considerable reduction in the image attenuation would nevertheless be possible.

Using ideal transducers with an image attenuation of zero as a microphone and as a telephone receiver, it should be possible in principle to retain the distribution of attenuation when planning the network. The ideal transducer has, as a microphone, a sending reference equivalent of approximately 3 nepers or 26 decibels. With speech the microphone gives a speech voltage approximately 3 nepers less, so that the ratio of signal to line noise at present tolerated is not sufficient.

With the receiver insets at present in use, acoustic termination has practically no influence on the input impedance because of the considerable image attenuation. With the ideal transducer, any variation in the acoustic termination would, however, influence the electrical input. The acoustic impedance of the ear in the telephone frequency band is capacitive. The magnetic transducer acts as a converter and on the electrical side transforms the impedance of the ear into inductive impedance while the electric or capacitive transducer leaves it, on the electrical side, as a capacitive impedance. This is the case, however, only so long as the telephone receiver is held close to the ear.

The image attenuation of a microphone is also a criterion for unhindered reception of extremely small acoustic pressures. The lower margin for an acoustic pressure of p_u, called the equivalent acoustic pressure, is determined by the voltage of background noise U_R of the microphone, on the electrical side. This equivalent acoustic pressure is

$$p_u = U_R \sqrt{\frac{Z_{ak}}{Z_{el}}} \cdot 10^{\frac{a+70}{20}} \quad \mu\text{bar/V}$$

The smaller the image attenuation a, the smaller the equivalent acoustic pressure.

APPENDIX (to Annex 5)

The cavity may, as a preliminary approximation, be replaced, according to its volume $V_L = F \cdot l$, by an acoustic quadripole with an average cross-section F and length l. Its transfer index is $g_L \approx \mathrm{j} \frac{\omega}{c} l$ and its image impedance $Z_L = \frac{\rho c}{F}$, c being the speed of sound and ρ the air density. At the end, the quadripole is terminated by an acoustic impedance Z_M of the diaphragm.

The input impedance of the quadripole is

$$Z_1 = Z_M \frac{1 + \frac{Z_L}{Z_M} \operatorname{thg}_L}{1 + \frac{Z_M}{Z_L} \operatorname{thg}_L}$$

We get $\frac{Z_L}{Z_M} \ll 1$ and $\operatorname{thg}_L \approx \mathrm{j} \operatorname{tg} \frac{\omega}{c} l \approx \mathrm{j} \frac{\omega}{c} l$ because l is small in comparison with the wavelength $\lambda = \frac{c}{f}$. We therefore obtain

$$Z_1 = \frac{1}{\frac{1}{Z_M} + \mathrm{j}\,\omega \frac{V_L}{\rho c^2}}$$

ρc^2 can be replaced by χP and the equation given above is then obtained.

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Question 16/XII — The effect of compandors on transmission quality

(continuation of Question 16 of Study Group XII, 1961-1964)

According to the results of tests based on the opinions of typical telephone users, what is the effect on transmission quality when compandors are inserted in an international telephone circuit?

Note 1. — Annex 1 describes a testing method which might be used by Administrations making such tests.

Note 2. — Annexes 2, 3, 4 and 5 give the results of tests already made by the Administrations of Italy and of the United Kingdom and France.

Note 3. — It would be very interesting to carry out tests on compandors which barely meet the new C.C.I.T.T. recommendation for compandors that are to be used on very long international circuits (document AP III/49, page 19 et seq., section C with the modifications indicated in document AP III/94, pages 10-11).

ANNEX 1

(to Question 16/XII)

Recommended procedure for tests concerning the effects of compandors

(See *Red Book*, Volume V, pages 697-699.)

ANNEX 2

(to Question 16/XII)

Comparative appreciation tests to assess the effect of compandors used in telephone systems with circuit noise

(Contribution of the Italian Administration)

(See *Red Book*, Volume V, pages 699-701.)

ANNEX 3

(to Question 16/XII)

Opinion tests on compandors

(Contribution of the Italian Administration)

1. *General*

A series of opinion tests was undertaken to assess the effect produced by a syllabic compandor as regards subscribers' tolerance of continuous spectrum noise.

To this end, a typical two-way telephone system was set up in the laboratory, the four-wire chain having a fixed equivalent of 7 db (on the two-wire side of the terminal equipment). In this, a compandor couple at zero unchanged level could be inserted at will in either direction of transmission.

With this circuit, several series of tests were made, the reference equivalent of the complete system being varied each time together with the noise level.

The outcome may be summarized as follows:

a) the effective gain introduced by the compandor varies with the absolute noise level measured at the zero relative level point. This gain is almost negligible at very low noise levels, and increases with the noise level.

Above a certain level, this gain will evidently decrease if the noise levels continue to increase.

The tests were not continued to the point where this aspect became apparent as far as this, but consideration of the behaviour of the compandor leaves no doubt about this phenomenon. It was found that for a reference equivalent of 40 db the gain could be assessed at 10 db for noise with a psophometric level of −40 dbm0, injected into a compandor-equipped circuit; for a level of −25 dbm0 this gain will be around 12 db, and for a level of −15 dbm0 the corresponding gain is about 15 db.

b) The above results will also be valid for a reference equivalent of 30 db, while for lower equivalents the gain will be a little lower, but will vary in the same manner.

2. *Results of opinion tests with circuit noise*

Opinion tests were afterwards undertaken with a telephone system having three compandor couples for both directions. This represents the maximum permissible number for an international call according to C.C.I.T.T. recommendations.

The three compandor couples were constructed by different firms, but each compressor was connected to an expander constructed by the same manufacturer.

The noise was injected with the same level on each of the three circuit sections determined by the three compandor couples.

Since, as a result of the preliminary tests, it was discovered that the response characteristics of the equipment at the various levels had a dominating effect, an attempt was made to obtain a response very close to the most unfavourable response tolerated by the C.C.I.T.T. Accordingly, the response represented in Figure 1 was obtained; this met requirements fairly well.

The unchanged compandor level was adjusted to zero.

The tests were carried out with two values of reference equivalent for the complete call (40 and 30 db) and with three psophometric levels (−40, −30, and −20 dbm0) for the noise injected in the three sections.

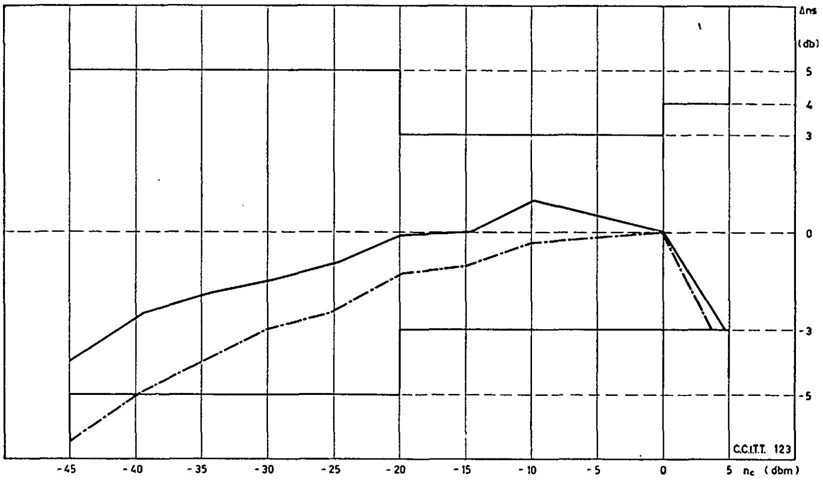


FIGURE 1

The results obtained are shown in Figures 2 and 3.

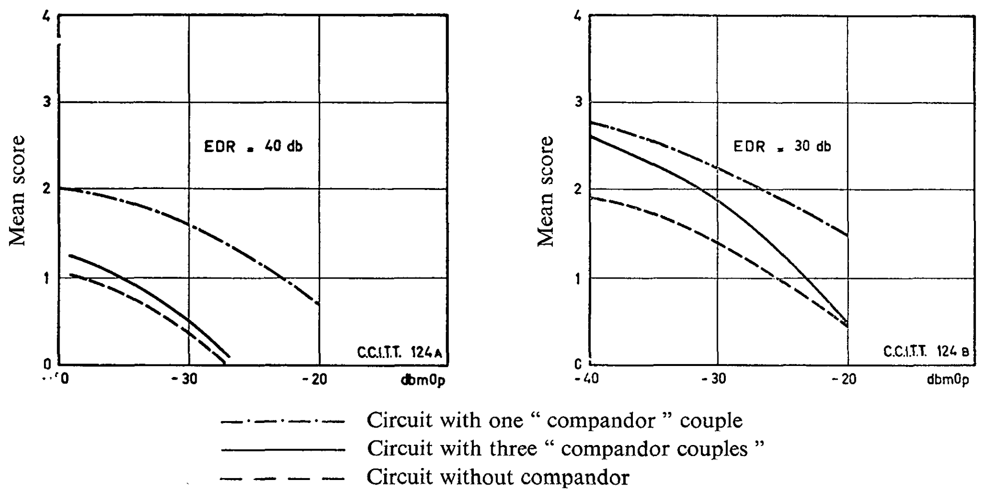
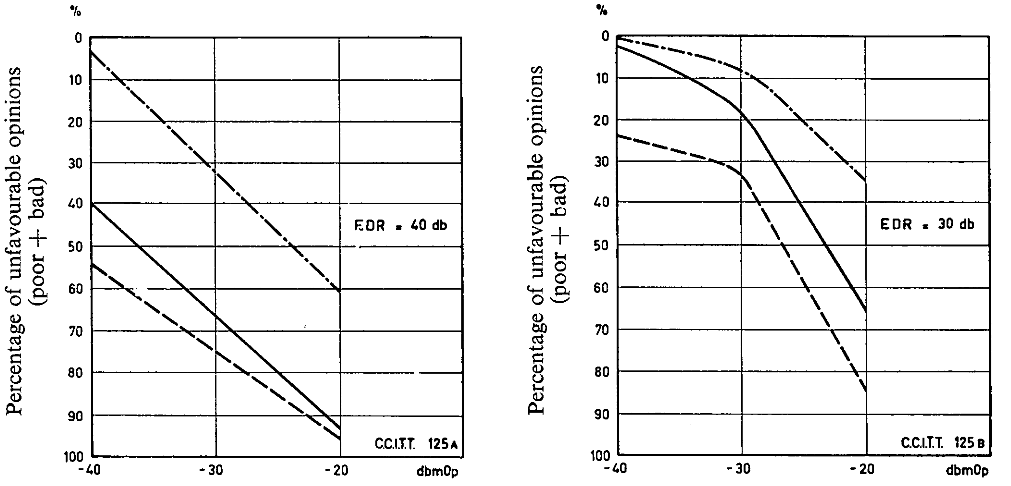


FIGURE 2

Figure 2 shows the mean score obtained for each circuit condition as a function of noise level.

Figure 3 represents the percentage of unfavourable opinions obtained for each circuit condition as a function of the noise level.

One of the three curves represented in each diagram in Figures 2 and 3 relates to a circuit without compandor, a second to a circuit with three compandor couples, and a third to a circuit with a single compandor.



—·—·—·— Circuit with one " compandor " couple
———— Circuit with three " compandor couples "
— — — — Circuit without compandor

FIGURE 3

Examination of these figures, and comparison between the circuit with three compandors and the circuit without compandor shows that the mean score obtained is in close agreement with the percentages of unfavourable opinions.

In particular, it will be seen that in a circuit with three compandors gain should be at its maximum with respect to noise for noise levels lower than those used in the tests.

It will be seen that there is a loss of gain for these noise levels when the noise level increases. Thus, with a reference equivalent of 40 db and a psophometric noise level of −35 dbm0, injected on each circuit section equipped with compandors, the gain will be about 4 db, while for higher levels the gain steadily falls off, until it is less than 1 db. Similarly, with a reference equivalent of 30 db and a noise level of −30 dbm0 the gain is about 10 db, while if the noise level is −25 dbm0 the gain will be reduced to about 4 db, and will be even lower should the noise level increase.

Furthermore, comparing Figures 2 and 3 for the case of the circuit having a single compandor, it will be observed that the results differ little from those obtained in the first series of tests, if it be borne in mind that in the second series the noise levels indicated relate to each section, while in the first series they relate to the entire circuit.

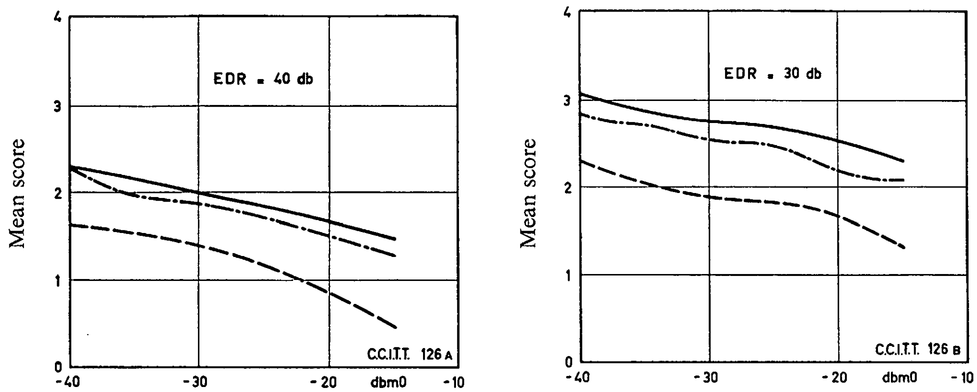
In Figures 2 and 3 it will be seen that there is a gain of approximately 17 db for reference equivalents of both 40 db and 30 db with a noise level of −25 dbm0, while with a noise level of −20 dbm0 the gain is about 12 db.

3. *Results of opinion tests on circuits suffering from intelligible cross-talk*

After the experiments on the effects of continuous-spectrum noise, opinion tests were carried out on circuits on which intelligible crosstalk was present.

To this end, use was made of a conversation recorded on magnetic tape, first directly and then through a compressor, so as to carry out the tests either with uncompressed cross-talk (crosstalk caused by a circuit without compandor) or with compressed crosstalk (between two circuits with a compandor).

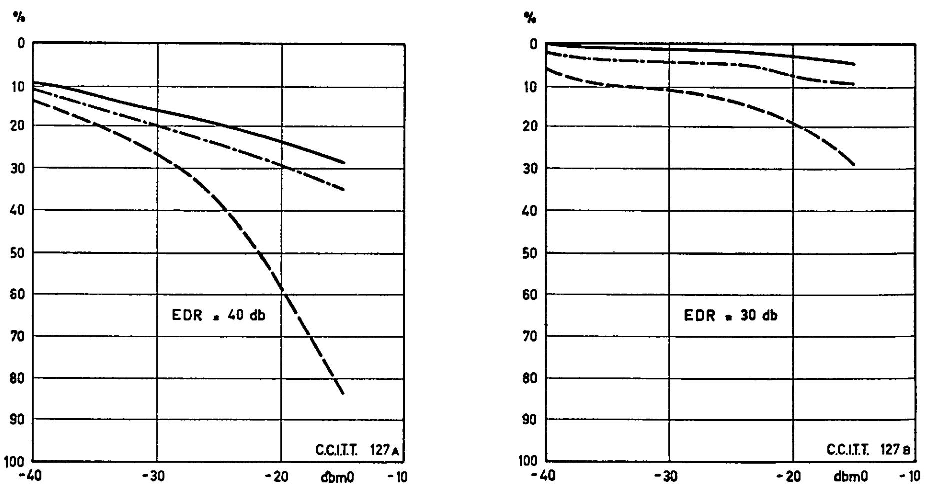
The crosstalk thus recorded was transmitted into a circuit with three compandor couples, but only in one of the three circuit sections.



——————— Compressed crosstalk } Circuits with compandors
—·—·—·— Uncompressed crosstalk }
— — — — Uncompressed crosstalk and circuit without compandor

FIGURE 4

Figures 4 and 5 summarize the outcome of these tests. Figure 4 shows the mean score, and Figure 5 the percentage of unfavourable opinions. In the same figures, the tests classified as "without compandor" were also made on the circuit with three compandors, but the crosstalk was injected at a point between an extensor and a compressor, that is to say at a point of zero-compression. Thus, as far as crosstalk is concerned, the circuit could be considered as without a compandor.



—·—·—·— Compressed crosstalk } Circuits with compandors
——————— Uncompressed crosstalk }
— — — — Uncompressed crosstalk and circuit without compandor

FIGURE 5

Before drawing conclusions from these figures, it may be mentioned that the opinions method does not seem suitable for the determination of the subscribers' tolerance of intelligible crosstalk, and in consequence other methods should be tried.

The reason is that, if we consider only those curves showing crosstalk on circuits without compandors, and if we compare them with the similar curves for noise in the preceding figures, we shall see that, making allowance for the different evaluation of levels along the abscissae, the tolerance of users to crosstalk is greater than their tolerance to continuous-spectrum noise.

This was to have been anticipated, but it does not enable us to reach the conclusion that levels of intelligible crosstalk can be higher than those of continuous spectrum noise, since the limits on intelligible crosstalk are governed by the need for keeping conversation secret.

The C.C.I.T.T., indeed, has itself recommended a limit of −50 dbm0 for noise and −52 dbm0 for intelligible crosstalk. Moreover, the opinion method seems suitable for a simple comparative assessment of the advantages to be derived from the use of compandors in connection with intelligible crosstalk.

Accordingly, it will be seen that if we consider compressed crosstalk (the most commonly encountered in practice), for a reference equivalent of 40 db, the gain produced by the compandor is of the order of 16 to 18 db, while for an equivalent of 30 db the gain will be of the order of 22 to 24 db.

These very high gains are very close to the theoretical figures derived from steady-state conditions. Further, it seems that these figures are affected only to a negligible degree by the crosstalk level.

Lastly, if we consider uncompressed crosstalk, the gain produced by the compandor is reduced from 3 to 4 db.

Hence it seems that compressed crosstalk is a little less troublesome than uncompressed crosstalk, although the average power of compressed crosstalk is a little greater.

We may therefore conclude that, as far as crosstalk is concerned, the average power would be less important than the peak power, which can be reduced by the compressor.

ANNEX 4

(to Question 16/XII)

The speech transmission performance of compandored circuits

(Contribution of the United Kingdom of Great Britain and Northern Ireland)

Introduction

Compandors are inserted in speech communication links to mitigate the deleterious effects of noise generated in the circuit between terminal equipments. A compandor consists of a 'compressor' at the input to a channel and an 'expander' at its output; the former processes the speech by reducing its amplitude range and the latter performs the inverse function and restores this range to its original value. The success of the device depends upon the fact that a given level of noise heard in a telephone conversation is more serious when it occurs during intervals when speech is temporarily absent than when it accompanies speech utterances.

Compressor and expander could be simple instantaneously non-linear devices if the transmission link between them were capable of conveying faithfully all the components, produced by the compressor, which lie beyond the frequency band of the original speech; if these are suppressed by, for example, transmission filters the re-expanded speech will be distorted. The application to normal F.D.M. systems is considered here in which syllabic compandors are used. These exploit certain properties of speech signals that enable the amplitude compression to take place with negligible extension of the frequency bandwidth.

Speech can be represented as having a ' carrier ' component, extending in bandwidth from, say, 250 to 3400 c/s but of small amplitude range and containing nearly all the information, together with an amplitude modulation component which fluctuates relatively slowly (at syllabic rate) but over a very large amplitude range. The compressor treats the two components separately reducing the amplitude range of the latter and retransmitting the new combination. The frequency spectrum of the modulating function extends only up to about 30 c/s and so the processed signal will be relatively little increased in bandwidth.

The control currents of compressor and expander correspond roughly to the modulating function. They are extracted by rectification and filtered by a circuit that largely rejects the ' carrier'. Because there is a wide interval between the frequency spectrum of the modulating function and that of the ' carrier ', this filtration can be achieved with simple RC smoothing networks. Smoothing time constants of compressor and expander are not critical and may differ in a ratio as great as $2^1/_4$ to 1 with negligible effect upon the speech. For the same reason any leakage of control current into the line from the compressor due to unbalance of the variable loss device will not be transmitted (the line equipment does not transmit below about 250 c/s) and so cannot appear as a spurious signal.

Control currents for compressor and expander are derived from the signal at opposite ends of the line. If it is sufficiently distortionless, therefore, accurate matching of compression and expansion processes is possible. Attenuation/frequency distortion, delay distortion and peak power limitation are obviously important and have greater effects than in the absence of a compandor. Suddenly applied pulses of tone, which cause surges in the compressor output, can cause trouble. With speech the effect is, under practical conditions, negligible because speech does not involve very rapid rates of rise.

Compression has the effect of raising the power level of syllables that are lower than a certain value (termed the unaffected level) and of reducing that of those above. The unaffected level can be so chosen that the long-term mean power of the speech signals is unchanged by the compression. Taking typical values, the unaffected level must be about −6 dbm0 for this to apply. The expander must also be arranged so that a test tone at the input to the circuit of −6 dbm0 will appear at the output of the expander at a level 6 db below the nominal relative level. Even when this is so, any departure from accurate compression and expansion laws could result in excessive gain at some other level. The stability margin would thus be reduced and could only be restored by increasing the nominal loss, reducing the circuit loss variations or improving the return losses. Static mismatch will also result in some distortion of the speech signals but, fortunately, this must be very severe to give rise to appreciable subjective degradation.

Effective subjective noise improvement

A compandor yields its advantage principally by inserting a relatively large loss in the expander at the relatively low levels occupied by the circuit noise. Thus, when speech is absent, the circuit noise is considerably attenuated relative to the level of the speech (when the latter appears); the result (with 2:1 companding) is an improvement in ' speech-off ' noise of $U - N$ where U is the unaffected level and N is the level of the noise; both units are in dbm0, the noise level being a psophometrically weighted value. When speech is present it will be accompanied by circuit noise which will be less attenuated than in the absence of speech because the expander loss is now controlled by the speech on the line and this is higher in level than the noise. When speech is present, therefore, the expander gives no improvement in speech-to-noise ratio; the compressor, however, raises the speech on the line by $\frac{1}{2}(U - S)$ where S is the long-term mean power of the speech while present (dbm0) at the input to the compressor.

Tests have been made at various speech levels and with several values of circuit noise in which listeners compared sentences heard over a compandored circuit with the same speech heard over uncompandored circuits. The noise was adjusted on the uncompandored circuits until they were judged subjectively equivalent to the compandored. The amount by which the noise level on the uncompandored circuit had to be reduced below that on the compandored gives the effective subjective noise improvement. The results showed that the improvement was approximately equal to two-thirds the ' speech-off ' noise improvement plus one-third the improvement in speech-to-noise ratio due to the action of the compressor. This yields the formula:

$$\text{Effective subjective noise improvement} = \frac{5U}{6} - \frac{S}{6} - \frac{2N}{3}.$$

It is sometimes desired to restrict the noise improvement and this can be done by limiting the lower end of the 2 : 1 range of the expander and allowing it to assume a linear law. If the level at which this change-over takes place is denoted by L_x (dbm0) and if L_x is higher than N, the ' speech-off ' noise improvement is reduced to $U - L_x$ and N must be substituted by L_x in the the formula. If L_x is lower than N it will have no effect. To avoid any excess gain over the compandored circuit at levels below L_x it is necessary to limit the 2:1 range of the compressor also to a value not lower than L_x. The limit of the 2:1 range of the compressor is denoted by L_c. Provided that this is done mismatch of L_c and L_x has a negligible effect on the speech quality. To reduce risk of excess gain at low levels, it is common practice to set L_c and L_x at -45 dbm0; a compandor is unlikely to be used in a circuit with lower noise level than $N = -44$ dbm0.

Taking $S = -12$ dbm0, $N = -30$ dbm0 (psophometrically weighted) and $U = -6$ dbm0, the effective subjective noise improvement is 17 db. Such a circuit would thus be equivalent to an uncompandored circuit with noise level -47 dbm0.

Residual degradation

Compandors will, in practice, be inserted only in circuits that require some noise improvement but it must be noted that the residual degradation due to the combined effects of the various mismatches and distortions mentioned above can be appreciable if the noise improvement is small. This, together with other disadvantageous factors, makes it undesirable to insert compandors when less than a certain minimum noise improvement is necessary.

A single compandor inserted in a 4-kc/s circuit will, in the absence of noise, degrade that circuit by about $1\frac{1}{2}$ db. This impairment and those mentioned below apply when the over-all connection has a reference equivalent of 20 to 25 db and may be somewhat greater for lower values and less for higher. This effect is included in the improvement given by the formula above;

if further frequency bandwidth restriction and attenuation/frequency distortion are present an additional allowance must be made. For example, one compandor inserted in a tandem connection of four 3-kc/s circuits will increase the additional impairment to 2½ db. If the four 3-kc/s circuits are compandored separately the additional impairment rises to 4 db.

Conclusions

A compressor/expander combination that remains accurately matched under all service conditions can render otherwise unacceptable line noise levels quite tolerable and the effective subjective noise improvement can be calculated by a simple rule.

Certain precautions must be taken and certain penalties endured with some tandem-connected circuits and these may limit the application of compandors to circumstances where a substantial noise improvement is required.

ANNEX 5

(to Question 16/XII)

The effect of compandors on transmission performance

(Contribution of the French Administration)

The French Administration has undertaken a number of opinion tests to assess the performance of telephone circuits using a compandor in both directions of trunk transmission.

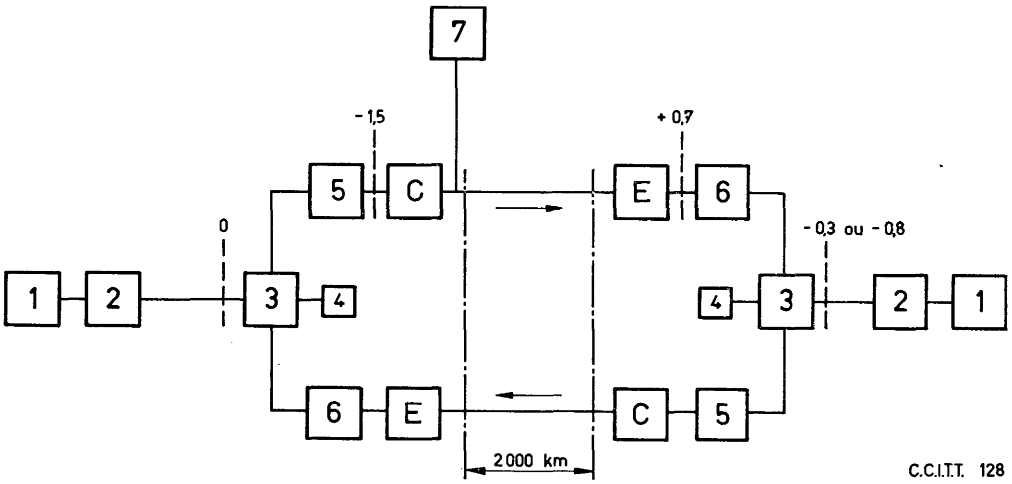
The compandors used were in accordance with the recommendations of the C.C.I.T.T.

The telephone circuit used had two local lines of medium length and a four-wire trunk circuit 2000 kilometres long, Paris-Pau-Paris. Each transmission channel therein was made up by combining two carrier channels Paris-Pau and Pau-Paris with four-wire looping at Pau. The circuit was a special one and went through no switching centre, either local or trunk.

The telephone sets used were of a type in current use in France (U43, local battery, type 328/1). The reference equivalent of such a telephone set, subscriber's line excluded, is +5.5 dN (sending) and −1.5 dN (reception).

Figure 1 shows how the connection was made up and the relative levels at various points in the trunk circuit. The total reference equivalent of the connection was successively adjusted to 1.7 and 3.6 N, in the following manner:

Sending equivalent of the telephone set	0.55	0.55
Subscriber's line and attenuator (2)	0.5	1.2
Trunk circuit equivalent	0.3	0.8
Subscriber's line and attenuator (2)	0.5	1.2
Receiving equivalent of the telephone set	−0.15	−0.15
Total reference equivalent	1.7 N (15 db)	3.6 N (31 db)



C	Compressor	3	Terminating sets
E	Expander	4	Balancing networks
1	Telephone set	7	Noise generator
2, 5, 6	Attenuators		

FIGURE 1. — **Diagram of the connection showing relative levels in nepers**

For each value of reference equivalent six circuit conditions were examined.

— inclusion and non-inclusion of compandors;

— three separate values of line noise: −6.3 NmOp; −4.6 NmOp; −3.45 NmOp, or −55 dbmOp; −40 dbmOp; and −30 dbmOp respectively.

In the first instance, this was the noise proper to the trunk circuit itself. In the other two, a Gaussian white noise was added to the inherent circuit noise by means of the noise generator 7.

Tests were made with a team of three operators who had already taken part in opinion tests and knew the scoring rules. They were not, however, specialists in telephonometry. To overcome difficulties due to the team being so small, the tests were repeated several times according to a systematic plan.

For each individual test, a particular pair of operators was used (operator *A* speaking, operator *B* listening); scores were taken on a circuit with a particular reference equivalent, in a specific circuit condition. After a brief introduction, operator *A* would read out a text of average difficulty for a couple of minutes, and operator *B* would score in accordance with the ordinary scale, from 0 to 4 according to the varying degree of attention required to follow the text. In assessing the quality of the connection, such transmission incidents (clicks, cut-offs, etc.) which might have occurred during this short interval of time were overlooked. Since one direction only was actually used during one individual test, the noise was adjusted to its nominal level only in one direction of transmission.

A full replication comprised 36 individual tests. Each pair of operators had to assess each one out of six circuit conditions in a communication with a particular reference equivalent.

One peculiarity of the test plan was that each of the three operators had to listen and score during 12 consecutive tests (six circuit conditions and two different talkers). The plan of the subsequence (12 tests) was kept secret and was often changed. This procedure, it is thought, reduces discrepancies in scores due to individual severity and it renders the differences in assessment occurring in one and the same sub-sequence more significant.

Incidentally, to reduce the range of variation of mean scores, several replications were undertaken for each reference equivalent (two replications for 1.7 N, four replications for 3.6 N), i.e. 12 assessments per circuit condition in the first instance, and 24 in the second.

Clearly, we cannot say whether the team chosen, in view of its smallness, constituted a representative sample of all telephone users.

The results of the tests are summarized in Tables I and II and in Figure 2.

Figure 2 a, relating to the connection with 1.7 N reference equivalent, shows how the curves cross at about −36 dbm0p. For a circuit with a reference equivalent of 3.6 N, it cannot be said exactly where this point will be, since the two curves are very much the same for low noise levels (Figure 2 b). However, it does seem that use of a compandor is justified as soon as the noise level reaches −40 dbm0p.

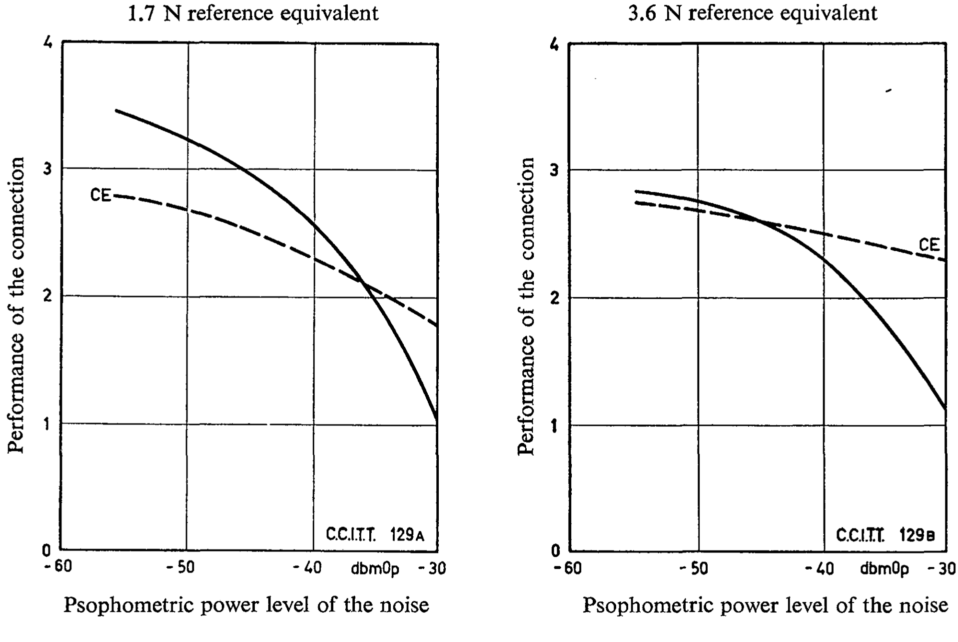


Figure 2a

Figure 2b

FIGURE 2

The tests performed by the French Administration were designed to give curves such as those in Figure 2, showing telephone performance assessment as a function of the noise level at the end of a trunk circuit taking as parameter the total reference equivalent of the circuit. Hence, the tests relating to the two connections (results in Tables I and II) were made separately, with the same team, but after fairly long intervals. It will be seen that the team of operators has not uniformly applied the scoring rules during the tests represented in Table I and during those shown in Table II. The average score is almost unchanged (2.32; 2.30), whereas it ought to be significantly lower for the second series of tests, made on a circuit with a reference equivalent significantly

TABLE I

Performance of a connection of 1.7 N reference equivalent

	R_b	C_b	B_c	R_c	C_r	B_r	mean
−55	2 4	3 4	3 4	3 3	3 4	4 4	3.42
−55 CE	2 2	3 4	2 2	2 3	3 4	3 3	2.75
−40	2 2	3 3	3 3	2 3	3 3	2 2	2.58
−40 CE	2 1	3 3	2 2	2 2	3 3	2 3	2.33
−30	1 1	1 1	1 1	1 1	1 2	1 1	1.08
−30 CE	2 2	2 3	1 1	2 2	2 1	2 1	1.75
mean	1.92	2.75	2.08	2.16	2.67	2.33	2.32

TABLE II

Performance of a connection with 3.6 N reference equivalent

	R_b	C_b	B_c	R_c	C_r	B_r	mean
−55	3 2 3 3	4 3 3 3	3 2 2 2	3 3 3 3	3 2 3 3	2 4 3 3	2.83
−55 CE	4 3 3 3	3 4 2 2	2 2 3 2	4 3 2 2	3 3 2 2	3 4 3 2	2.75
−40	3 3 3 2	2 2 2 2	2 2 2 2	2 2 3 3	1 2 3 2	3 2 3 3	2.33
−40 CE	2 2 4 4	2 3 3 2	2 2 2 3	2 2 3 3	2 2 3 3	2 2 2 3	2.5
−30	2 2 1 2	0 1 1 1	0 1 1 1	1 1 2 2	0 1 1 1	1 1 2 1	1.12
−30 CE	3 2 2 3	3 2 2 3	2 1 2 2	2 3 2 3	2 1 2 3	2 3 2 3	2.29
mean	2.67	2.29	1.88	2.46	2.08	2.46	2.30

Notes for Tables I and II

Talkers = *B C R*
Listeners = *b c r*

The figures in the first column express the psophometric voltage level at the end of the trunk circuit, in dbmOp. When this figure is followed by the letters CE, the companders are included in the circuit.

higher. Hence these results cannot be used for dual presentation (performance as a function of reference equivalent, with the noise level as a parameter).

This fact, incidentally, confirms the fluctuations in operators' judgments, and would seem to show how well justified are consecutive scorings in comparative performance tests.

Statistical analysis of the results

Statistical tests were made on Tables I and II.

Let i be the line index (for a particular circuit condition), j the column index (for a particular ordered pair of operators), and k the sequence index.

$i = 1, 2 \ldots 6$

$j = 1, 2 \ldots 6$

$k = 1, 2$ for Table I

$k = 1, 2, 3, 4$ for Table II.

We shall assume that the individual scores x_{ijk} are random variables obeying normal laws with the same standard deviation σ.

The assumption subjected to statistical test is that the mean of each of these laws of probability depends on i et j alone (circuit conditions and operator pairs), without interaction between them.

$$\mu_{ijk} = \alpha_i + \beta_j$$

and not

$$\mu_{ijk} = \alpha_i + \beta_j + \gamma_{ij}$$

Under these conditions, we can obtain two independent estimates of σ^2 from the variance between replications and the residual variance.

For Table I, we have:

$$\widehat{\sigma^2} = \frac{\sum\limits_{ijk} (x_{ijk} - x_{ij.})^2}{36} = 0.236 \text{ (number of degrees of freedom: 36)}$$

$$\widehat{\sigma^2} = \frac{2 \sum\limits_{ij} (x_{ij.} - x_{i..} - x_{.j.} + x_{...})^2}{25} = 0.354 \text{ (number of degrees of freedom: 25)}$$

For Table II:

$$\widehat{\sigma^2} = \frac{\sum\limits_{ijk} (x_{ijk} - x_{ij.})^2}{108} = 0.366 \text{ (number of degrees of freedom: 108)}$$

$$\widehat{\sigma^2} = \frac{4 \sum\limits_{ij} (x_{ij.} - x_{i..} - x_{.j.} + x_{...})^2}{25} = 0.214 \text{ (number of degrees of freedom: 25)}$$

In the customary manner, we indicate the means taken over i, j or k, either singly or in combinations by replacing the indices with dots.

If we take a threshold of significance of 1 %, application of Snedecor's test leads to acceptance of the initial hypotheses.

On the other hand, the assumption:

$$\mu_{ijk} = \alpha_i$$

(independent mean value of the operator pairs and the replication) has to be rejected. In this case a third estimate of σ^2 can be obtained by calculating the between columns variance.

$$\widehat{\sigma^2} = \frac{12 \sum_j (x_{.j.} - x_{...})^2}{5} = 1.314 \text{ (number of degrees of freedom: 5)}$$

(for Table I)

$$\widehat{\sigma^2} = \frac{24 \sum_j (x_{.j.} - x_{...})^2}{5} = 1.978 \text{ (number of degrees of freedom: 5)}$$

(for Table II)

and this last estimate differs significantly from the first two.

Suppose—and this is equivalent to adopting the least favourable estimate—that the standard deviation is 0.6 for both Tables I and II. This being so, the circuit condition means will obey normal standard deviation laws:

$$\frac{\sigma}{\sqrt{12}} = 0.17 \qquad \text{(Table I)}$$

$$\frac{\sigma}{\sqrt{24}} = 0.12 \qquad \text{(Table II)}$$

Question 17/XII — Loudspeaker telephones

(continuation of Question 17 of Study Group XII, 1961-1964)

What conditions (from the point of view of telephone transmission) should be satisfied by subscribers' telephone stations which may be used for international calls and which include loudspeakers or broadcasting type microphones with amplifiers?

Note 1. — Annex 1 below sets out the principles adopted for studying the conditions which telephone sets with loudspeakers must satisfy from the point of view of transmission performance.

Note 2. — The following two points seem to merit especial attention:

1. A method to be recommended for measuring the sensitivity of a loudspeaker telephone. (A method used by the United Kingdom Administration is described in Annex 2 below.)

2. The effect of noise and echo time in the room in which the loudspeaker telephone is installed. Should limits be laid down beyond which neither the loudspeaker nor the broadcasting type microphone should be used? Some comments on this point appear in Annex 3.

Note 3. — Annexes 4, 5 and 6 give additional information supplied by the Administrations of the German Federal Republic, Poland and Sweden.

ANNEX 1

(to Question 17/XII)

Principles adopted for studying the conditions which telephone sets with loudspeakers must satisfy from the point of view of transmission performance

(See Volume V of the *Red Book*, pages 703 and 704.)

ANNEX 2

(to Question 17/XII)

A method for measuring the sensitivity of a loudspeaking telephone set

(Contribution of the United Kingdom Administration)

The United Kingdom Administration has in hand studies which will determine the transmission performance characteristics needed for successful loudspeaking telephones. The factors being studied include the following:

1. Vocal level used while speaking into a microphone more distant than the normal handset. It has been found that subjects talk some few db louder.

2. Preferred sound pressure when listening to speech reproduced by a loudspeaker. The sound pressure needed in a subject's ear, when listening without room noise, is considerably lower for loudspeaker listening (some 20 db) than that needed when speech is reproduced from an earphone. On the other hand loudspeaker listening performance, in terms of opinion score, is much degraded by the presence of, say, 50 db room noise. Furthermore, it does not seem possible to compensate this degradation by any increase in sound pressure; the speech would first be judged too loud.

3. Preferred shape for over-all air-to-air sensitivity-frequency characteristic is also being studied. The preferred shape is by no means a level response; some 6 to 8 db per octave rise from 200 to 3200 c/s seems preferable but may not be optimum. It would seem desirable to incorporate this shaping in the sending part and leave the receiving part substantially level. This ensures satisfactory operation in combination with the usual shapes of frequency response of modern handset telephones.

If the preliminary conclusions mentioned in 1 and 2 above are substantiated, it is clearly undesirable to aim at sensitivities, sending and receiving, which are as high as those of handset telephones. Such sensitivities are, in any case, probably unattainable for a loudspeaking telephone.

The experimental results are being collected with the aid of what might be termed a " working reference loudspeaking telephone ". This comprises high-quality components, no voice switching but reasonably adequate over-all sensitivity by means of directional microphones and careful matching.

Sensitivities, sending and receiving, are being measured and expressed in terms of the corresponding parts of the A.R.A.E.N. (without filter). The principle is as follows:

High-quality recorded speech material is used. For sending, it is reproduced from an artificial mouth and the electrical output from a loudspeaking telephone is compared with that from the A.R.A.E.N. send end, comparison being made by means of a speech voltmeter which measures long-term r.m.s. (with pauses excluded). For receiving, the speech signal is injected into the loudspeaking telephone line terminals and the sound pressure at a defined point not too far from the loudspeaker compared with that which would be found in the ear of a listener had the same electrical input been applied to the A.R.A.E.N. receive end. Figures 1 and 2 show the conventional positions and distances that have been adopted. An A.R.A.E.N. send end can conveniently be used in conjunction with a speech voltmeter to measure the sound pressure from the loudspeaking telephone. All measurements are made with the loudspeaking telephones standing on a table in an ordinary room so as to correspond to normal conditions of use.

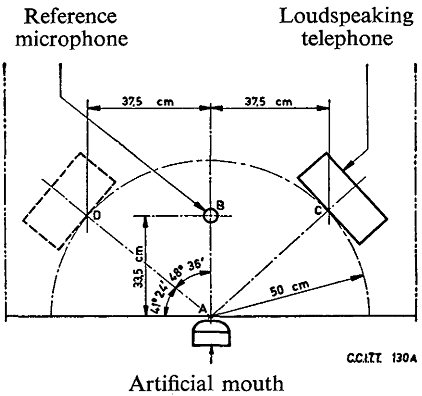


FIGURE 1. — **Sending sensitivity**

1. All dimensions are parallel to table surface.
2. The point source of the artificial mouth at A is in alignment with the table edge and 35 cm above it.

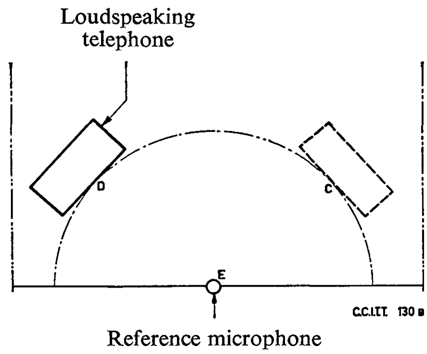


FIGURE 2. — **Receiving sensitivity**

3. The height of the acoustic screen of the reference microphone at B or E is 35 cm above the table surface.

Typical measured results are as follows:

Loudspeaking telephone	With or without voice switching	Sensitivity db relative to A.R.A.E.N.	
		Sending	Receiving [1]
A	Without	−30.0	−12.2 to +1.7
B	With	−19.0	−10.8 to +6.9
C	With	−20.0	− 6.1 to +1.1

[1] Extreme positions of manual gain adjustment which is under the direct control of the user.

Corresponding sensitivities for handset telephones with no subscribers' lines relative to the A.R.A.E.N. send and receive ends are about −9.5 db for sending and +14.5 db for receiving. (These figures were obtained in a somewhat different manner and so need to be checked; they are given as rough indications only.)

It appears from the foregoing that desirable sending and receiving sensitivities could be specified in terms of the A.R.A.E.N.

ANNEX 3

(to Question 17/XII)

Effect of room noise and reverberation time in the room in which the loudspeaking telephone set is used

I. Contribution by the United Kingdom Administration

Preferred sound pressure when listening to speech reproduced by a loudspeaker. The sound pressure needed in a subject's ear, when listening without room noise, is considerably lower for loudspeaker listening (some 20 db) than that needed when speech is reproduced from an earphone. On the other hand loudspeaker listening performance, in terms of opinion score, is much degraded by the presence of, say, 50 db room noise. Furthermore, it does not seem possible to compensate this degradation by any increase in sound pressure; the speech would first be judged too loud.

II. Contribution by the Swedish Administration

In the light of its experience with various kinds of loudspeaker telephones, the Swedish Administration considers that especial attention should be given to the effect of room noise with loudspeakers with voice-operated switching. To reduce this effect, the set must be so designed that the directional effect throws the speaker's voice into relief. Otherwise room noise will cause pointless blockages.

III. Contribution by the American Telephone and Telegraph Company

As mentioned in previous responses we do not believe that the customers' loops or other plant used in the regular switched network should be changed to fit speakerphone (distant talking and listening) set requirements. Therefore, the only logical objective is that, broadly, a speakerphone should provide the same grade of transmission as the hand telephones used by the same Administration and meet the same impedance requirements.

It was realized that the early Bell System speakerphones could not meet this objective. They employed amplification in both the transmit and receive branches of the telephone set to compensate for the distance losses of the microphone and loudspeaker. It is impractical to provide enough hybrid balance, for the wide range of impedances existing in the loop plant, to permit full compensation for these losses without creating a singing, or near singing condition, through the air path coupling the microphone and loudspeaker. Thus these sets could approach the objective only under very favourable situations. Recently the Bell System has developed a much improved speakerphone using the switched gain principle. This is proving highly satisfactory and is rapidly replacing speakerphones of the earlier type. With the new device amplification is switched, by voice control, from the receiving branch, for incoming speech, to the transmitting branch, for outgoing speech. The receiving branch amplification is under control of the user but in no case is the total gain in the loop through the transmit, receive, hybrid and coupling paths allowed to be sufficient to create singing or near singing conditions.

The new speakerphone provides transmission closely approaching the objective for favourable room conditions. It should be appreciated, of course, that any distant talking set will be less satisfactory than close talking if used in a highly reverberant room. Similarly, transmission in rooms with very high noise will be less satisfactory with a speakerphone than with a hand telephone set both because of the greater noise pickup by the microphone and because of the greater effect on received speech at practical loudspeaker levels. For this reason speakerphone installations should be limited as far as practical to favourable conditions such as exist in offices occupied by individuals of the executive or junior executive level or to home locations free from high noise levels and reverberation. Users should be instructed to speak within a reasonable distance of the microphone, say about 2 feet, for best performance. It also is important to provide a hand telephone for alternative use to ensure better privacy or to meet the needs of unfavourable conditions.

The switched gain speakerphone and its design principles are described in two papers available to C.C.I.T.T. members:

Fundamental Consideration in the Design of a Voice-Switched Speakerphone, A. BUSALA, *B.S.T.J.*, Vol. 39, p. 265.

Functional Design of a Voice-Switched Speakerphone, W. F. CLEMENCY and W. D. GOODALE, Jr., *B.S.T.J.*, Vol. 40, p. 649.

ANNEX 4

(to Question 17/XII)

Contribution by the Administration of the Federal Republic of Germany

(See Volume V of the *Red Book*, pages 705-706.)

ANNEX 5

(to Question 17/XII)

Contribution by the Swedish Administration

(See Volume V of the *Red Book*, pages 706-708.)

ANNEX 6

(to Question 17/XII)

Contribution by the Polish Administration

For information we are submitting provisional specifications for the loudspeaker subscriber's telephone station " AGAT ", intended for use in the public network of the People's Republic of Poland and accepted for international telephone calls.

1. *Introduction*

The present description gives the AGAT specifications which explicitly characterize the properties of these loudspeaker receiving devices. The parameters of these receiving devices must meet the conditions required of transistor telephone devices.

2. *Transmission parameters*

2.1 *Sending sensitivity*

The AGAT sending sensitivity is defined by the ratio between the output voltage at the terminal of a load resistance adjusted to the resistance of the apparatus and the acoustic pressure on the microphone.

The sensitivity is determined by use of a signal with a spectrum equivalent to that of the human voice.

The normal sensitivity must be such that, in the presence of the most probable acoustic pressure on the microphone in normal sending conditions, the output voltage obtained is equal to that the most likely to be observed at the terminals of ordinary telephone sets.

The sending sensitivity must be equal to $S_e = 0.17$ V/μb. (See articles [1] and [2]).

It must be possible to adjust the sending sensitivity within the limits 0 and 1 N relative to the normal value (the regulating devices are accessible to the installer and the user of the apparatus).

2.2 *Maximum effective output level*

The AGAT output level must not exceed the absolute level, independently of the pressure on the diaphragm of the microphone.

2.3 *Sensitivity-frequency curve*

The sensitivity-frequency curve at the sending end of AGAT must be within the limits of $\pm$ 0.5 N relative to the optimum curve.

The optimum sensitivity curve at the sending end of AGAT must satisfy the following conditions:

a) the form of the curve over the telephone spectrum (300-3400 c/s) must compensate the attenuatoin distortion introduced by the TKM 0.6 type of cable (typical subscriber's line) with a nominal attenuation at 800 c/s of 1 N;

b) outside the telephone spectrum (i.e. below 300 c/s and above 3400 c/s) the transmission sensitivity must fall off rapidly.

2.4 *Non-linear distortion*

The harmonic distortion coefficient throughout the passband of telephone frequencies must not exceed 2% for the normal output level.

2.5 *Background noise*

The level of background noise in the transmission channel must be at least 5.1 N below the absolute level.

3. *Receiving parameters*

3.1 *Receiving sensitivity*

The receiving sensitivity of AGAT is defined by the ratio between the acoustic pressure measured in the free field at a distance of 1 metre along the axis of the loudspeaker and the tension at the terminals of the telephone set.

The maximum sensitivity must not permit the crosstalk received by AGAT to be intelligible.

The following is proposed as the maximum sensitivity at the receiver: $S_r = 7.6$ μb/V (see article under [1]).

It must be possible to adjust the sensitivity at the receiving end within the limits -3.5 and 0 N in relation to the maximum value (adjustment is accessible to the subscriber).

3.2 *Normal loudness*

The loudness level, averaged over a long period, in a free field at a distance of 1 metre along the axis of the loudspeaker, must not be less than 61 db.

3.3 *Non-linear distortion*

At the given loudness level the harmonic distortion coefficient must not exceed 10%. Below the normal level this coefficient should diminish rapidly.

3.4 *Sensitivity-frequency curve*

The " sensitivity-frequency " curve at the receiving end of AGAT must be within the limits ± 0.5 N in relation to the optimum curve. The optimum curve must satisfy the following conditions:

a) the form of the curve over the telephone spectrum (300-3400 c/s) must be horizontal;

b) outside the telephone spectrum the receiving sensitivity must drop by at least 1.5 N per octave.

3.5 *Background noise*

The background noise of the receiving channel must correspond to an input signal with a level of about -6.7 N. It is desirable to respect tolerances of $\pm$ 0.2 N.

4. *General parameters*

4.1 *Internal d.c. resistance of the apparatus*

In order that the apparatus may be connected with a Strowger 32 AB type exchange the d.c. resistance must not exceed 400 ohms when calling and 1800 ohms during the call. After the circuit has been cut the resistance of the apparatus must not be below 100 000 ohms.

4.2 *Internal impedance of the apparatus at telephone frequencies*

The impedance of the apparatus must imitate the characteristic impedance of an average telephone line.

The return loss between the impedance of the apparatus and the characteristic impedance of the TKM 0.6 cable must not be below 1.6 N at the lower end of the frequencies 500-1500 c/s and at least 1.4 N in the rest of the telephone spectrum.

The above value must be met in any position of the volume control at the receiving end, the AGAT microphone being mechanically blocked during the measurement for a voltage level not exceeding 0 N through the subscriber's line on the exchange side to the apparatus.

4.3 *Echo attenuation*

The echo attenuation in AGAT must not be below 1.6 N in the frequency band 500-1500 c/s and 1.4 N in the rest of the telephone spectrum.

This value of attenuation must be reached in the maximum position of the volume control at the receiving end, AGAT being placed, during the measurement, in a room whose reverberation time is 1 second; a voltage level not exceeding 0 N being applied through the subscriber's line from the exchange side to the apparatus.

4.4 *Singing margin*

The singing margin of AGAT in any conditions (even the most unfavourable) must exceed 0.3 N.

During the call the singing margin must be large enough for the coupling distortion to be at an inaudible level. It is proposed that 2.5 N should be adopted as the minimum singing margin during conversation, the loudspeaker being near the microphone (at sub-critical distance). With a greater distance between the loudspeaker and the microphone the singing margin may be less but not below 1 N.

4.5 *Transient parameters*

In a voice-operated system the value required for the singing margins and the active balance return loss must be observed during variations in strength.

Self-oscillation of the system must not cause audible acoustic effects.

4.6 *Exchange feeding current and allowable loop resistance of the subscriber's line*

The apparatus must maintain the parameters indicated above for the nominal line voltage of the exchange and for the loop resistance of the subscriber's line which must not be higher than the allowable resistance for the exchange to which the apparatus can be connected.

4.7 *Allowable variation of the parameters of the apparatus with external factors*

The allowable variations in the AGAT parameters must not exceed ± 0.1 N for temperature variations within the limits 10 and 40° C; humidity: 50 and 80% and throughout a usable period of 15 years.

REFERENCES

[1] Kowalski, Zb.: Loudspeaker telephone apparatus. Requirements and possibilities. *Przeglad Telekomunikacyjny*, 1961, No. 11, p. 343-348 and No. 12, p. 371-375.

[2] Kowalski, Zb.: Comparative analysis of the properties of loudspeaker telephones. *Prace Instytutu Laczneści*, 1960, No. 4, p. 3-63.

Question 18/XII — Calculation of the reference equivalent of a subscriber's line

(new question)

Can a simple formula be given which, when transmission plans are made for local networks, can be used to calculate the reference equivalent of a subscriber's line being calculated when its value is required to be known independently of the over-all reference equivalent of the local system (subscriber's line and telphone set)?

Note 1. — Two methods have been considered for this calculation:

1) to use a simple formula which involves only the resistance of the line (see for example, in Annexes 1 to 4 below: contributions by the French, Italian and Dutch Administrations, and the Helsinki Telephone Company).

2) to increase the image attenuation of the line, as measured at 800 c/s, by a certain percentage (which for an unloaded cable is equivalent to measuring or calculating the attenuation at a frequency above 800 c/s) (see, for example in Annex 5 below, the information supplied by the Swedish Administration).

Note 2. — The effect of the loss of a subscriber's line is allowed for by the United Kingdom Administration in the manner described in Annex I (Part II of Volume V of the C.C.I.T.T. *Red Book*). Moreover, reference should be made to the following article: ANDERSON, E. W.: Local Networks. *Proceedings of the Institution of Electrical Engineers*, 111, April 1964.

Note 3. — Obviously, whichever method is used in the case of sending, account must also be taken of the feeding current loss due to the ohmic resistance of the line.

ANNEX 1

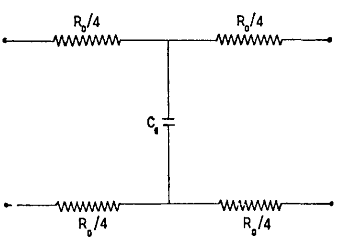
(to Question 18/XII)

Reference equivalent of a local cable line

(Contribution of the French Administration)

It is stated in Note 3, Recommendation P.11, that the reference equivalent of a subscriber line or a junction can be assimilated, at a first approximation, to the image attenuation at 800 c/s of that line. Subjective voice and ear tests carried out by the French Administration over different local cables have shown that, in actual fact, the reference equivalent differed considerably from the image attenuation.

Three types of elementary networks corresponding to 1 km of cable 0.4, 0.6 and 1 mm in diameter, were used (Figure 1). An artificial cable 1 km long was set up by putting L elementary networks of a given type in tandem.



d mm	R_0 ohms	C_0 µF
0,4	275	0,05
0,6	122	0,05
1	44	0,05

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FIGURE 1. — **Artificial cable, 1-km long with conductors of diameter *d*mm**

A block diagram of the measuring circuit is given in Figure 2. The telephone set was of the type currently used by the French Administration (BCI-U43). The battery voltage was adjusted in such a way as to supply the telephone set with a current of between 45 and 60 mA, irrespective of the over-all resistance of the artificial cable. The Primary System for Determining the Reference Equivalents had been previously calibrated in relation to the N.O.S.F.E.R.

The measurement consisted in comparing by ear the sound impressions obtained through an artificial cable with a resistance of $L\ R_0 = R$ and a resistance with a value R shunted by a capacitor of 100 µF. Because of this, the two measurements were made with constant feeding current. The difference between the balancing attenuation and the " hidden length " attenuation corresponded to the attenuation brought about by the artificial cable.

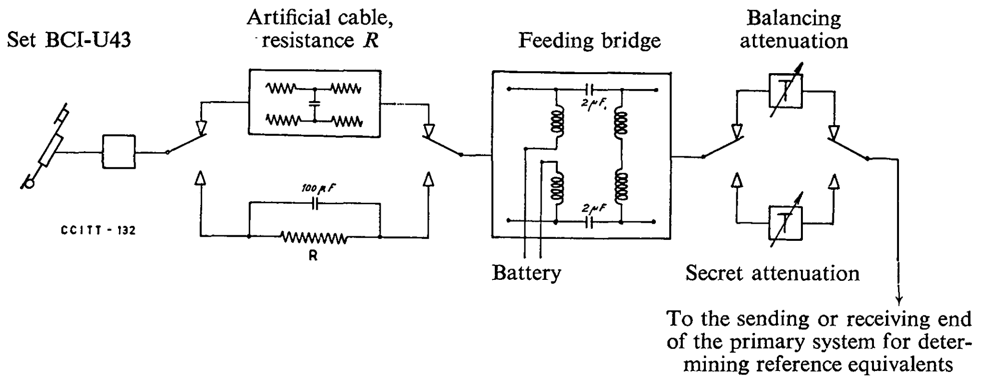


FIGURE 2. — **Block diagram for subjective measurement of the reference equivalent of a local cable**

Table I shows the results obtained through measurement of the reference equivalents of cables 0.4, 0.6 and 1 mm in diameter and 2.4, 4 km and 6 km long, for sending and reception.

TABLE I

Reference equivalent of various local cables

Diameter (in mm)	Length (in km)	Reference equivalent (in N)	
		Sending	Receiving
0.4	2	0.45	0.43
—	4	0.90	0.95
—	6	1.24	1.23
0.6	2	0.14	0.18
—	4	0.40	0.39
—	6	0.61	0.65
1	2	0.09	0.07
—	4	0.28	0.21
—	6	0.33	0.34

It may be seen from Figure 3 that the reference equivalent of a local cable depends, at a first approximation, only upon the line resistance, and may be represented, bearing in mind the margin of error inherent in all telephonometric measurements, by a simple law such as:

$$a = 0.4 \frac{R}{500} = 0.4 \frac{R_0 L}{500} \qquad (1)$$

R = line resistance in ohms
R_0 = resistance per km in ohms
L = length in km
a = reference equivalent in nepers

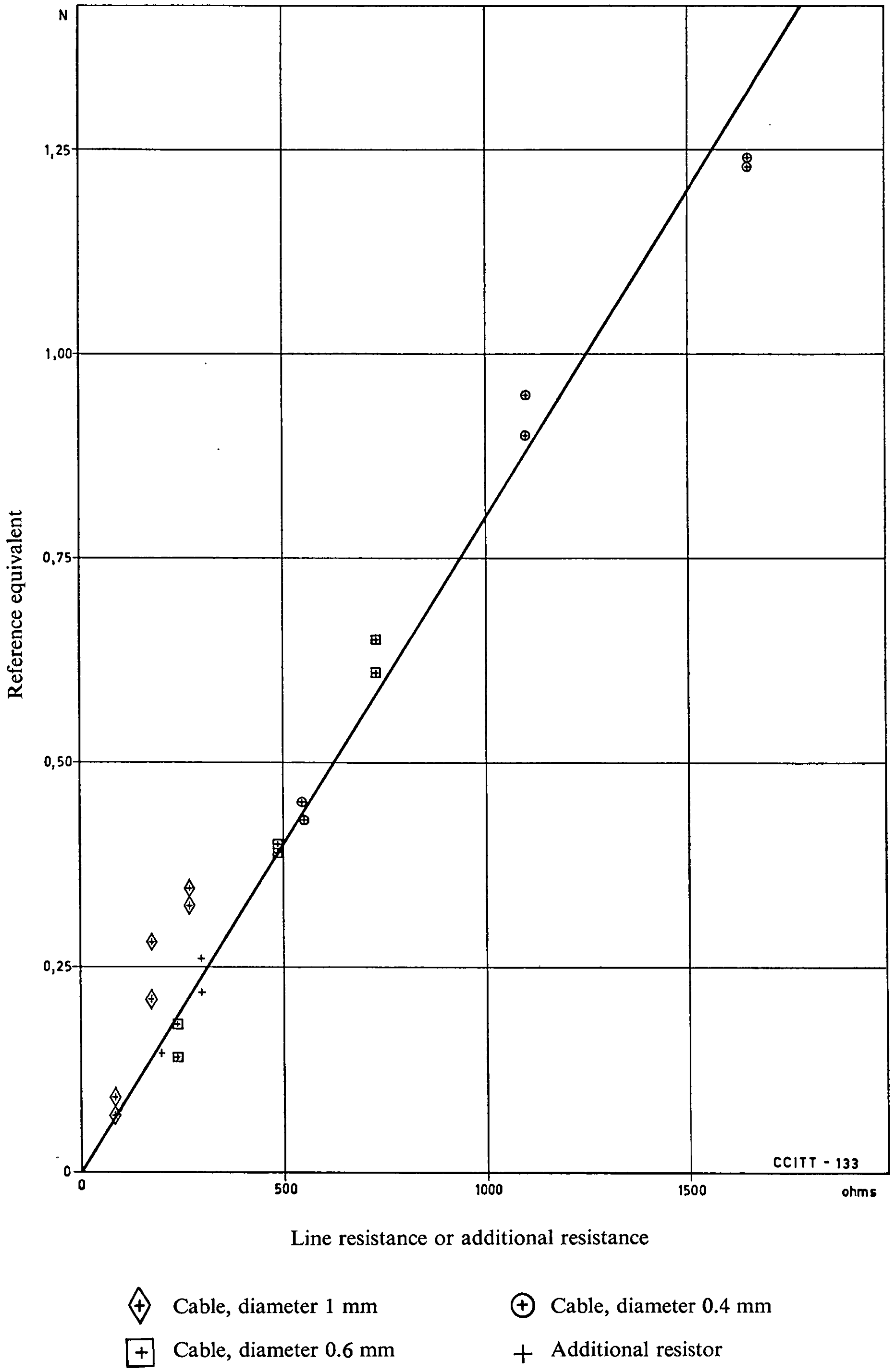


FIGURE 3. — **(France)**

The image attenuation α at angular frequency ω is of the form:

$$\alpha = \sqrt{\frac{C\,R\,\omega}{2}} = L\sqrt{\frac{C_0\,R_0\,\omega}{2}} = R\sqrt{\frac{C_0\,\omega}{2\,R_0}} \tag{2}$$

C capacitance of the line

C_0 capacitance per km

The various mathematical expressions of α show that the image attenuation α differs from the reference equivalent a in two basic respects:

1) α is proportional to the square root of the line resistance R_0 whereas a is directly proportional to it;

2) on a graph similar to that in Figure 3 where α is shown in the ordinate and R in the abscissa, the points plotted at a given frequency for local cables with a fixed diameter and variable in length, are aligned along a straight line through the origin. The lower R_0, and consequently the larger the diameter of the cable, the steeper the slope of the straight line. At a given frequency, the image attenuation over the whole of local cables of different lengths and diameters is therefore represented by a pencil of straight lines starting from the origin while the reference equivalent for the same set is represented by a single straight line with a given slope.

A law similar to type (1) may be found for the variation of the reference equivalent with line resistance, ignoring propagation phenomena and assuming the line resistance to be a single potentiometric attenuation, an assumption which involves neglecting the capacitance of the line.

Insertion of a resistance R between a source (or a receiver) with an internal impedance of $R_1 + \mathrm{j}\,X_1$ and a receiver (or source) with a resistance R_2 introduces an attenuation A

$$A = \frac{1}{2}\log_e \frac{(R + R_1 + R_2)^2 + X_1^2}{(R_1 + R_2)^2 + X_1^2}$$

that is, if

$$R < R_1 + R_2$$

$$A \simeq \frac{1}{2}\log_e \left[1 + \frac{2\,R\,(R_1 + R_2)}{(R_1 + R_2)^2 + X_1^2}\right] \simeq \frac{R\,(R_1 + R_2)}{(R_1 + R_2)^2 + X_1^2}$$

If: $R_1 = X_1 = 300$ ohms (impedance at 800 c/s of BCI-U43 set)

$R_2 = 600$ ohms, then

$$R_1 + R_2 + \frac{X_1^2}{R_1 + R_2} = 1000 \text{ ohms}$$

which is not significantly different from the proportionality factor in formula (1)

$$\frac{500}{0.4} = 1250 \text{ ohms}$$

Supplementary measurements were also carried out which furnished an estimate of the reference equivalent of the additional (unshunted) resistances of 100, 200 and 300 ohms for the BCI-U43 set. The results obtained are given in Table II as well as in Figure 3. These equivalents may be represented by the law defined in (1).

TABLE II

Reference equivalent of additional resistors in the BCI-U43 set

Resistance (in ohms)	Reference equivalent (in N)	
	Sending	Receiving
100	0.08	0.09
200	0.14	0.14
300	0.22	0.26

ANNEX 2

(to Question 18/XII)

Calculation of the reference equivalent of a subscriber's line

(Contribution by the Italian Administration)

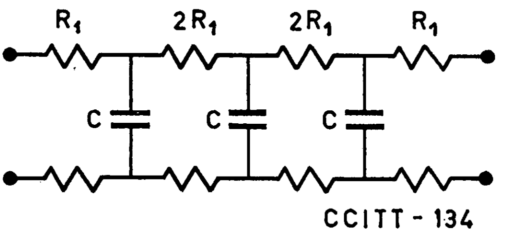
In studying this question, three types of cable pairs in current use on local systems were considered, namely pairs with wires of diameter 0.4, 0.6 and 0.9 mm.

On the basis of several statistical measurements for service conditions, the most probable values for the resistance per kilometre, R_0, and the capacity per kilometre, C_0, for lines of this kind were ascertained.

These values are:

diameter 0.4 mm	$R_0 = 274$ ohms/km	$C_0 = 40$ nF/km
diameter 0.6 mm	$R_0 = 122$ ohms/km	$C_0 = 36$ nF/km
diameter 0.9 mm	$R_0 = 53$ ohms/km	$C_0 = 41$ nF/km

Four-pole networks of the form shown below (three elementary cells in tandem) were then constructed to represent specimens of different lengths L:



$$C = \frac{1}{3} C_0 L \qquad R_1 = \frac{1}{12} R_0 L$$

FIGURE 1

Three lengths were simulated for each type of line:

diameter 0.4 250 ohms (0.91 km), 450 ohms (1.64 km), 650 ohms (2.37 km)
diameter 0.6 250 ohms (2.05 km), 450 ohms (3.69 km), 650 ohms (5.33 km)
diameter 0.9 78 ohms (1.5 km), 132 ohms (2.5 km), 210 ohms (4.0 km)

Subjective tests showed, firstly, that the reference equivalents of these artificial lines correspond (allowance being made for the inherent error margin) to the increase of receiving reference equivalent which results when they are included in a typical subscriber's system.

Furthermore, we measured the image attenuation of each artificial line, for the frequencies 300, 800, 2400, and 4000 c/s.

Table I assembles figures for reference equivalent and image attenuation, the common value of image impedance being shown between brackets.

TABLE I

Line (diameter, total resistance)	Reference equivalent (db)	Image attenuation			
		(300 c/s)	(800 c/s)	(2400 c/s)	(4000 c/s)
0.4 mm 250 ohms	+2.2	1.4 db (1600)	1.8 db (1080)	2.9 db (646)	3.7 db (486)
0.4 mm 450 ohms	+3.5	2 (1710)	2.8 (1160)	4.6 (639)	5.6 (498)
0.4 mm 650 ohms	+5.2	2.9 (1710)	4.2 (1080)	6.5 (576)	8.3 (460)
0.6 mm 250 ohms	+2.6	1.8 (1165)	2.5 (775)	4.2 (429)	4.8 (346)
0.6 mm 450 ohms	+4.4	2.6 (1220)	4 (777)	6 (470)	7.2 (370)
0.6 mm 650 ohms	+7.0	0.7 (1190)	5.4 (763)	8.6 (440)	11.2 (337)
0.9 mm 78 ohms	+0.9	0.8 (816)	1.3 (504)	2.1 (295)	2.6 (234)
0.9 mm 132 ohms	+1.7	1.4 (782)	2 (492)	3.4 (277)	4.2 (213)
0.9 mm 210 ohms	+2.6	2.3 (780)	3.1 (497)	4.8 (289)	6 (227)

It was noted that, for the types of cable investigated, a frequency can be found for which attenuation is practically the same as the reference equivalent, although this frequency varies considerably with the diameter of the conductors.

Furthermore, it was remarked that the reference equivalent of the lines in question was not related in any simple way to the line resistance.

Since the capacity per kilometre remains more or less constant with diameter, it is to be expected that for a given resistance the lines of greater diameter (i.e., the longest) will also have the biggest over-all capacity, and hence a higher reference equivalent.

A more thorough consideration of the correspondence between resistance of the line and the reference equivalents shows that the ratio between the equivalents and the resistance, multiplied by the square root of the diameter, is reasonably constant for all lengths and cable diameters (see the diagram in Figure 2).

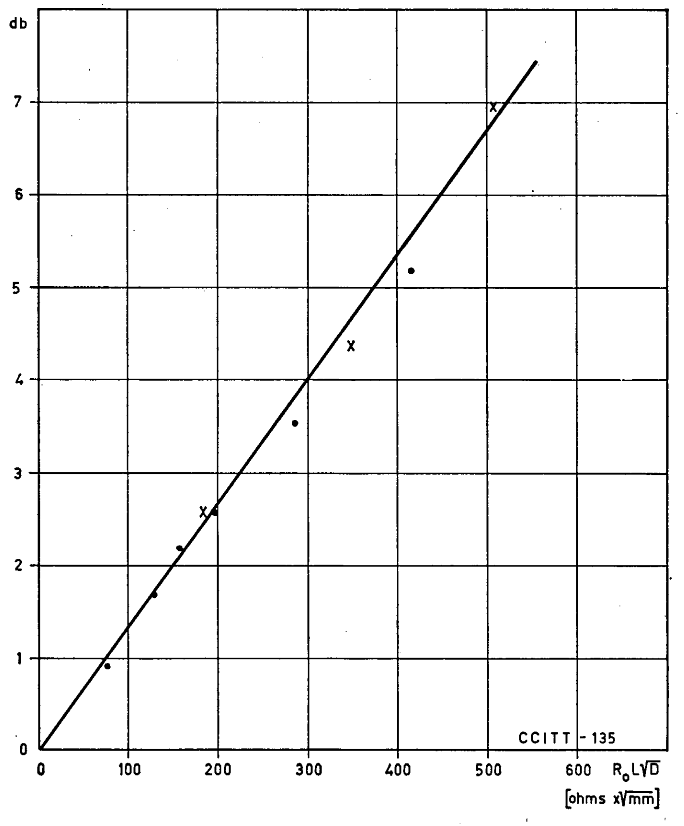


FIGURE 2. — **(Italy)**

Hence the following practical formula might be proposed for calculating the reference equivalent, E, of a subscriber's line:

$$E\,(\text{db}) = 0.0132\ R_0\, L \sqrt{D}\ (D \text{ being the diameter, in mm}).$$

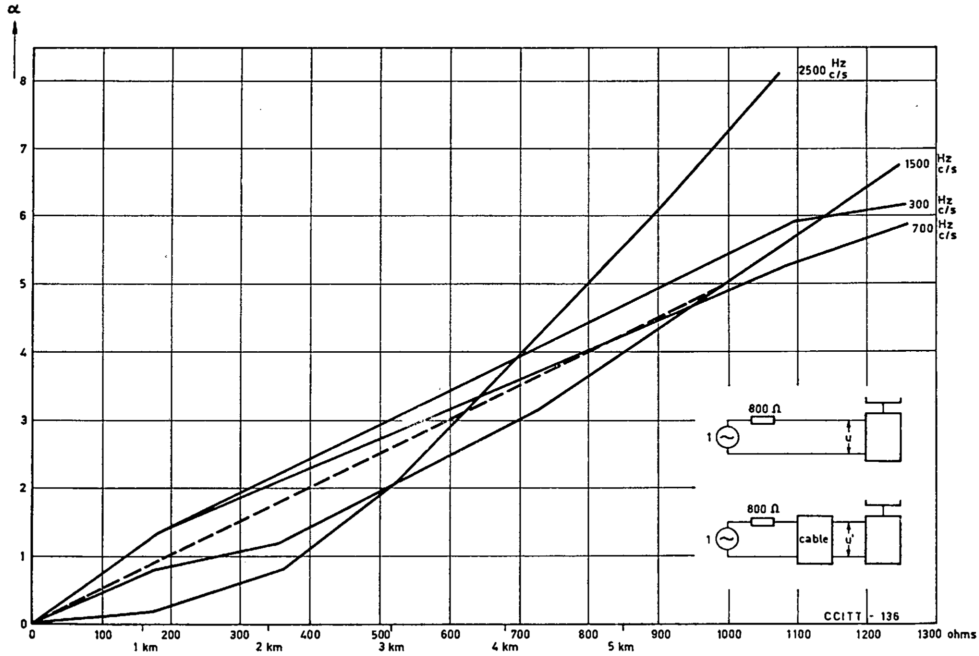
This formula has proved to be satisfactory for the cable lines in the Italian telephone system.

ANNEX 3

(to Question 18/XII)

Contribution by the Netherlands Administration

For the design of the local cable network a single rule is applied for lengths not exceeding 5 km. According to this single rule the insertion loss for the local network can be calculated at 1 db per 200 ohms loop resistance. This holds for cables whose capacity between the pairs of conductors is maximally 35 nF/km. Cables with a length of about 5 km or more show deviations from this empirical rule, which in certain configurations can be of importance. In Figure 1 the attenuations have been plotted, for different frequencies, as a function of the loop resistance. This graph, which only serves as illustration, is made for a conductor diameter of 0.5 mm. The part of the frequency band (700-1500 c/s) that is of importance for the loudness generally follows this simple rule. The subscriber set used for the measurements is of a type normally used in the Netherlands telephone service.



——— Loss for some frequencies — — — — 1 db/200 ohms (empirical rule)

FIGURE 1. — **Insertion loss of a local cable (Netherlands)**

$$\varnothing = 0.5 \text{ mm} \quad \alpha = 20 \log \left(\frac{u}{u'}\right)$$

ANNEX 4

(to Question 18/XII)

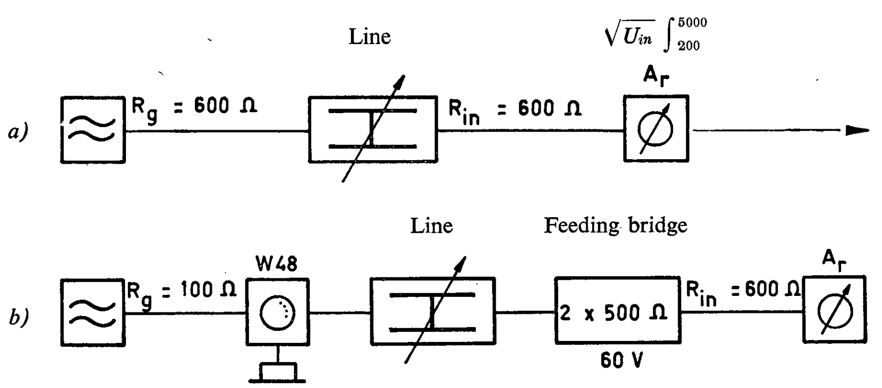
Reference equivalent of local cable lines

(Contribution by the Research Institute of the Helsinki Telephone Company)

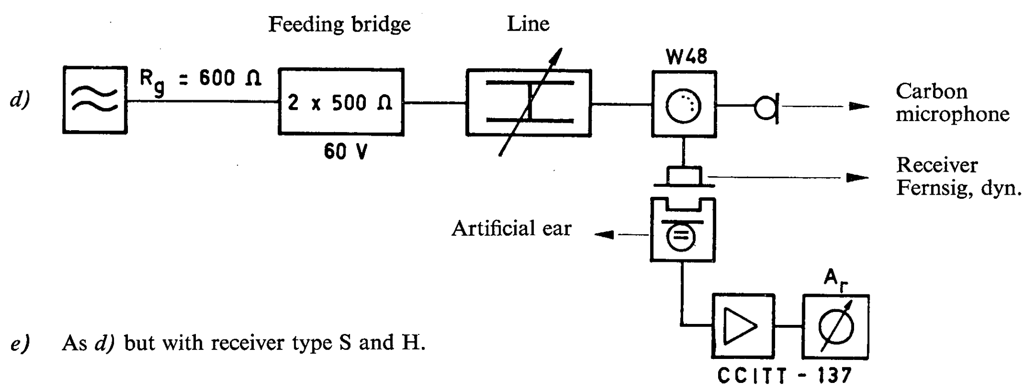
In Note 3, Recommendation P.11, it has been mentioned that as the reference equivalent of a subscriber's line the image attenuation at 800 c/s of the line can be used. However, it appears from the measurements carried out in the Research Institute of the Helsinki Telephone Company that the reference equivalent may deviate from the image attenuation.

The reference equivalents of cable lines have been measured using the objective method in accordance with the principle of K. Braun. The frequency scale of the generator is logarithmic, and the frequency band has been between 200 c/s and 5000 c/s. The voltage that comes to the indicating instrument is weighted according to the square root law, and is integrated during one sweep of the frequency band that lasts approximately 7 seconds. The pointer of the instrument stops to show the value of reference equivalent. The reading can be carried out very accurately, because the pointer stands absolutely still and does not swing as it does, e.g., in the OBDM of S & H, in which the same principle is applied, but the generator is rotating incessantly.

The reference equivalents of a cable line have been measured in five different ways. Block diagrams of these are shown in Figure 1 (*a*, *b*, *c*, *d* and *e*). Point *a*) is identical with the method



c) As *b)* but $R_g = 200$ ohms



e) As *d)* but with receiver type S and H.

FIGURE 1 **(Helsinki Tel. Co.)**

that is generally used in telephony for measuring the insertion loss between non-reactive resistances of 600 ohms, while the points *b*), *c*), *d*), and *e*) correspond to different conditions of use. Because the carbon microphone is unstable, it has been replaced in the measurements for sending by a generator the internal resistance for which is 100 ohms (point *b*). In order to find out the effect of the resistance of the microphone, a similar measurement has been made with a generator the internal resistance of which is 200 ohms (point *c*). As is generally known, the impedance of the receiver has a great effect upon the impedance of the telephone set. Because of this, the measurements for receiving have been carried out with two different receivers: point *d*), with a dynamic receiver the impedance of which is almost constant at a frequency band from 200 c/s to 5000 c/s, and point *e*), with a receiver the impedance of which increases with the frequency.

The measurements were made with three different cables, the technical values of which are given in Table 1.

TABLE 1

d (mm)	r_0 (Ω/km)	c_0 (nF/km)	Remarks
0.5	180	37	
0.6	125	37	Artificial cable (0.6 mm only)
0.8	68	37	According to measurements, $c_0 = 35$ nF/km

The results that were obtained are given in Table 2. Because the measurement carried out according to point *a*) does not correspond to common conditions of use, and the results obtained in this way differ from the others to some extent, they have not been considered when calculating the mean values. From the last column of the table it can be seen that for a_r (reference equivalent/km) a different value is obtained for different cable lengths. Thus, the reference equivalent of a local cable line (A_r) greatly depends upon the impedance matchings. In Figure 2 the reference equivalent (mean values) of the measured cable lines are shown as a function of the line resistance (R_L). From the curves drawn it can be seen that the reference equivalent of a cable line depends upon other factors, too, than merely upon the line resistance.

TABLE 2

Diameter	Length	Reference equivalent (N)							
			Sending		Receiving		Mean values	
(mm)	(km)	(*a*)	(*b*)	(*c*)	(*d*)	(*e*)	(*b, c, d, e*)	a_r (mN/km)
0.5	2	0.30	0.28	0.26	0.27	0.26	0.27	135
—	4	0.60	0.58	0.58	0.57	0.56	0.57	143
—	6	0.90	0.88	0.88	0.86	0.85	0.87	144.5
0.6	2	0.22	0.22		0.22		0.22	110
—	4	0.46	0.45		0.45		0.45	112.5
—	6	0.71	0.70		0.72		0.71	118.3
0.8	2.145	0.15	0.14	0.14	0.14	0.14	0.14	65.3
—	3.945	0.30	0.29	0.29	0.28	0.27	0.282	71.6
—	6.090	0.48	0.47	0.47	0.47	0.45	0.465	76.4

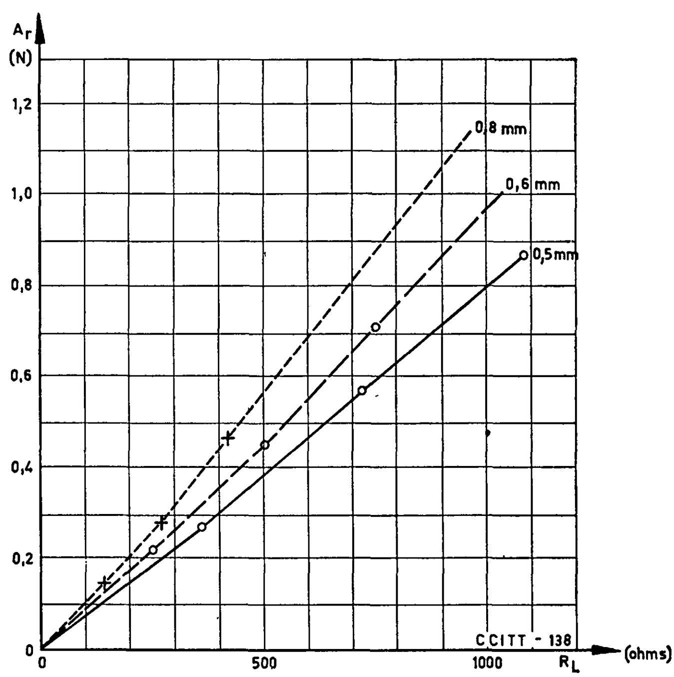


FIGURE 2. — **Reference equivalent of local cables (Helsinki Tel. Co.)**

The image attenuation (A) at angular frequency ω is obtained from the following equation:

$$A = \sqrt{\frac{R\,C\,\omega}{2}} = L\sqrt{\frac{r_0\,c_0\,\omega}{2}} = L\sqrt{\pi\,r_0\,c_0\,f} \tag{1}$$

C = capacitance of the line
c_0 = capacitance per km
R = line resistance in ohms
r_0 = resistance per km in ohms
L = length in km
f = frequency in c/s

It is always possible to find a frequency f_v, for which the calculated image attenuation is the same as the reference equivalent. For cables 0.5 mm and 0.8 mm in diameter these frequencies are approximately 1000 c/s and 790 c/s.

TABLE 3

d (mm)	f (c/s)	α (mN/km)
0.5	1000	145
0.8	790	77

Let us assume that between the frequency f_v and the diameter d there is the relation

$$f_v = k\,d^{-n}$$

According to Table 3 we obtain

$$\frac{1000}{790} = \left(\frac{0.8}{0.5}\right)^n$$

By calculation we obtain $n = 0.5$.
For the factor k we obtain the following value:

$$k = 0.5^{0.5} \cdot 1000 = 707$$

The reference equivalent can thus be calculated from the equation of the image attenuation by using the frequency f_v:

$$f_v = \frac{707}{\sqrt{d}} = \frac{1000}{\sqrt{2d}}\ \text{(c/s)} \tag{2}$$

where d = diameter in mm.

Example: a_v for 0.6-mm cable:

$$f_v = \frac{1000}{\sqrt{2 \cdot 0.6}} = 913\ \text{c/s}$$

$$a_v = \sqrt{\pi\,r_0\,c_0\,f_v} = \sqrt{\pi\,125 \cdot 37 \cdot 10^{-9} \cdot 913} = 116\ \text{mN/km} = 1.16\ \text{dN/km}$$

The value obtained fits fairly well the values in Table 2.

In order to calculate A_r with an arbitrary value of R_L, the following equation, too, can be found:

$$A_r = L\sqrt{c_0\, r_0\, \pi f_v}$$

$$f_v = \frac{1000}{\sqrt{2d}} \quad \text{(c/s)}$$

$$r_0 = \rho \frac{2 \cdot 1000}{\pi \dfrac{d^2}{4}} = \frac{8\rho}{\pi} \cdot 1000 \cdot \frac{1}{d^2} \left(\text{ohms/line km}\right)$$

$$\rho = \text{resistivity} = 0.0175 \text{ ohm } \frac{\text{mm}^2}{\text{m}} \text{ for copper}$$

$d =$ diameter in mm

$$r_0 = \frac{8 \cdot 0.0175 \cdot 10^3}{\pi} \frac{1}{d^2} = 44.5 \frac{1}{d^2} \left(\text{ohms/km of line}\right) \tag{3}$$

$$R_L = L r_0 ; \; L = \frac{R_L}{r_0} = \frac{R_L}{44.5} \cdot d^2$$

$$A_r = \frac{R_L}{44.5} d^2 \sqrt{\frac{\pi c_0 \cdot 44.5 \cdot 1000}{d^2 \cdot \sqrt{2d}}}$$

$$A_r = \sqrt{50\, c_0} \cdot R_L\, d^{3/4} \tag{4}$$

$d =$ diameter in mm

$R_L =$ line resistance in ohms

If $c_0 = 37$ nF/km, the following equation for calculating A_r is obtained:

$$A_r = 1.36 \cdot 10^{-3}\, R_L\, d^{3/4} \tag{5}$$

Table 4 gives the values of a_r, calculated according to equation (4).

TABLE 4

d (mm)	r_0 (Ω/km)	c_0 (nF/km)	a_r (mN/km)
0.5	180	37	145.5
0.6	125	37	116
0.8	68	35	76

The equations were found by making use of the values obtained in the measurements for cables 0.5 mm and 0.8 mm in diameter. But also for a cable 0.6 mm in diameter almost the same value is obtained by calculating as by measuring ($L = 6$ km). For a cable 0.4 mm in diameter the r_0 of which is 275 ohms/km and $c_0 = 50$ nF/km, one would, according to equation (4), obtain $a_r = 218$ mN/km. This value fits well to the results received in subjective tests that have been carried out in France (see Annex 1).

For short cable lengths, a smaller reference equivalent for a cable is obtained by measuring than by calculating according to equation (4). The differences are, however, so small that they are of little significance in practice.

ANNEX 5

(to Question 18/XII)

Contribution by the Swedish Administration

The Swedish Administration has recently investigated the relation between the reference equivalent and the image attenuation at 800 c/s for local cables. The investigation has comprised measurements on cables by means of Siemens OBDM-apparatus as well as graphical assessment of the reference equivalent from the frequency response curves of the cables.

The measurements have produced a value of 1.10 to 1.25 for the said relation, the value depending on the cable dimensions.

The calculations, on the other hand, have shown that for the stated frequency response a frequency 1400 c/s instead of 800 c/s must be used in order to give an attenuation equal to the reference equivalent. Since there exists no simple relation between frequency and the composite attenuation of a cable, the frequency thus calculated cannot be used to state a general relation factor as for the measured values above.

The wanted relation has proved to be independent of the length of cable.

Question 19/XII — Impedance variations in subscribers' lines and telephone sets

(continuation of Question 19 of Study Group XII, 1961-1964)

(documentary question)

a) What is the range of variation of the impedances of subscribers' lines and telephone sets?

b) What are the methods which may be applied by Administrations to reduce this range of variation?

c) Do Administrations consider that these methods should be applied to improve the balance return losses in their networks, particularly for the international calls which will be made in accordance with the new switching plan from the transmission point of view?

Note 1. — As regards point a), information is given in Annexes 1 (American Telephone and Telegraph Co.), 2 (France), 3 (Italy), 4 (Netherlands) and 5 (Federal Republic of Germany).

Note 2. — As regards point b), information is available in Annex 5 to part I of chapter V (Transmission) of the *Manual on national telephone networks for the automatic service* concerning standard methods using only linear elements. Annex 6 (Austria) describes a terminating set giving improved balance return loss by means of non-linear devices, and Annex 7 (Netherlands) indicates some results obtained in operational conditions with a terminating set of this type.

ANNEX 1

(to Question 19/XII)

Impedance variations in subscribers' lines and telephone sets

(Contribution by the American Telephone and Telegraph Company)

The characteristics of Bell System loops, terminated in typical telephone stations at the customer's end, have been investigated by using a sample of 1000 loops selected so as to be statistically representative of the plant as a whole.

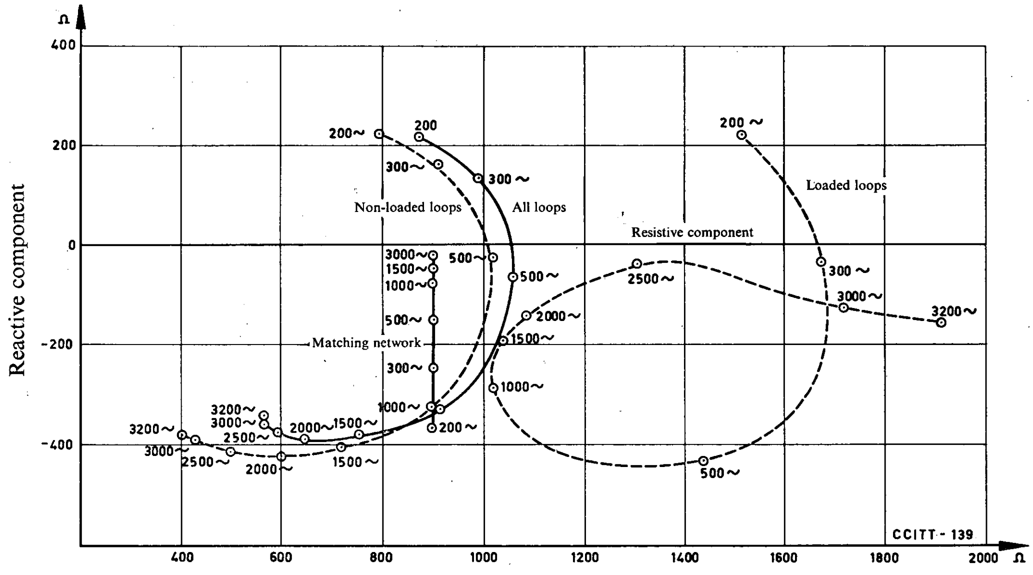


FIGURE 1. — **Input impedance of subscriber loops at the central office (does not include wiring or equipment)**

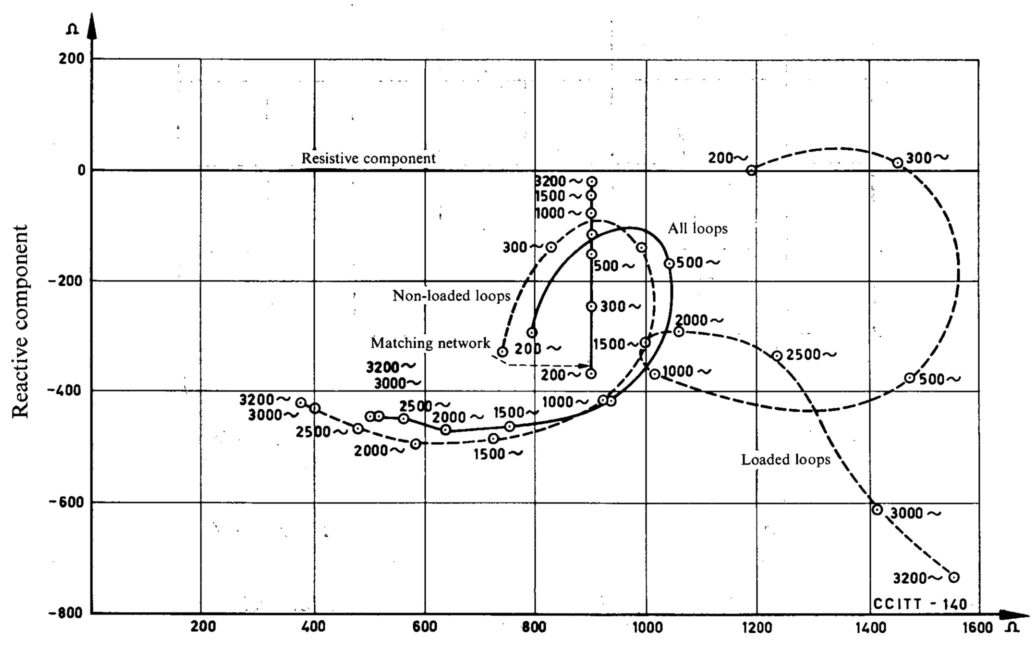


FIGURE 2. — **Input impedance of Bell System loop plant—includes No. 5 Crossbar incoming trunk circuit and central office wiring**

Figure 1 shows a plot of the mean impedance of these loops at the central office main frame for frequencies between 200 and 3200 c/s. Since the Bell System employs both loaded and non-loaded loops the impedances of the two types are shown separately as well as the mean of the composite plant. While these data show the impedance of the loop itself, the impedance as viewed through the central office equipment and wiring is more significant from the standpoint of echo control and all further data presented here will include the effects of the equipment and cabling in a typical No. 5 Crossbar Office. Figure 2 shows the plot of mean impedance seen through the office and differs from Figure 1 mostly at low frequencies. For comparison, a plot of the standard Bell System matching (balancing) network consisting of 900 ohms in series with 2.16 microfarads is also shown.

The simplest way to express the variation in impedance is in terms of return loss. Figure 3 shows the distribution of both singing and echo return loss relative to the standard balancing network. In this and other data on return loss the 3-kc figure is taken as representative of singing return loss. Echo return loss is defined here as the loss corresponding to the weighted average of the power ratios taken at the following frequencies.

500—weight = ½
1000—weight = 1
1500—weight = 1
2000—weight = 1
2500—weight = ½

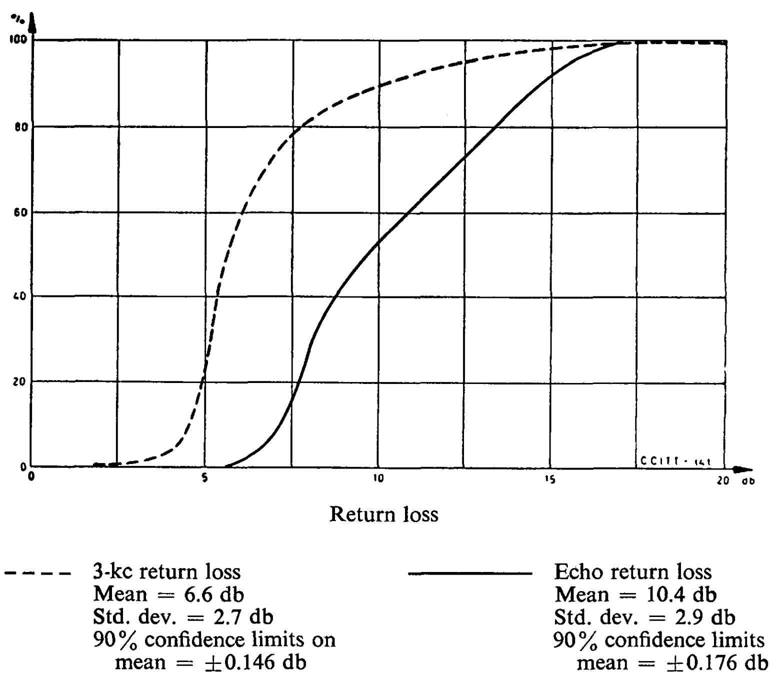


FIGURE 3. — **Echo and singing return loss of Bell System loop plant—measured against 900 ohms μF + 2.16 (includes central office equipment and wiring)**

The variation in impedance with frequency is illustrated more specifically by Figure 4 for non-loaded loops and Figure 5 for loaded loops. For each figure the loops have been classified by echo return loss. The figures show for both the real and imaginary components the range covered by the 25 per cent of the loops having the best (largest) return loss (curve A). Similar data for the best 50 per cent, best 75 per cent and all loops are also shown (curves B, C, D, respectively).

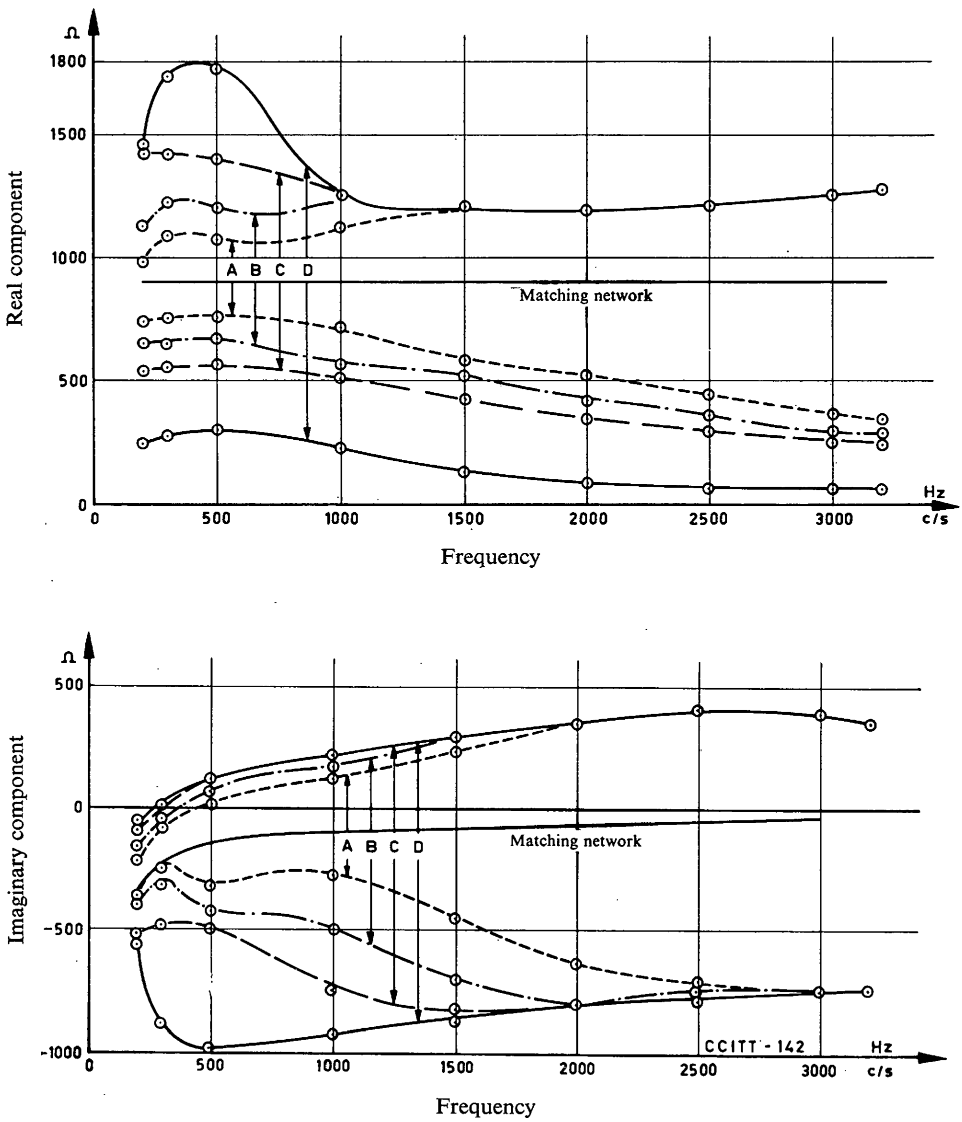


FIGURE 4. — **Distribution of non-loaded loop input impedance based on return loss performance (includes central office equipment and wiring)**

(A = Range of variation for best 25% of non-loaded loop plant, B = best 50%, C = best 75%, D = range for all 879 non-loaded loops)

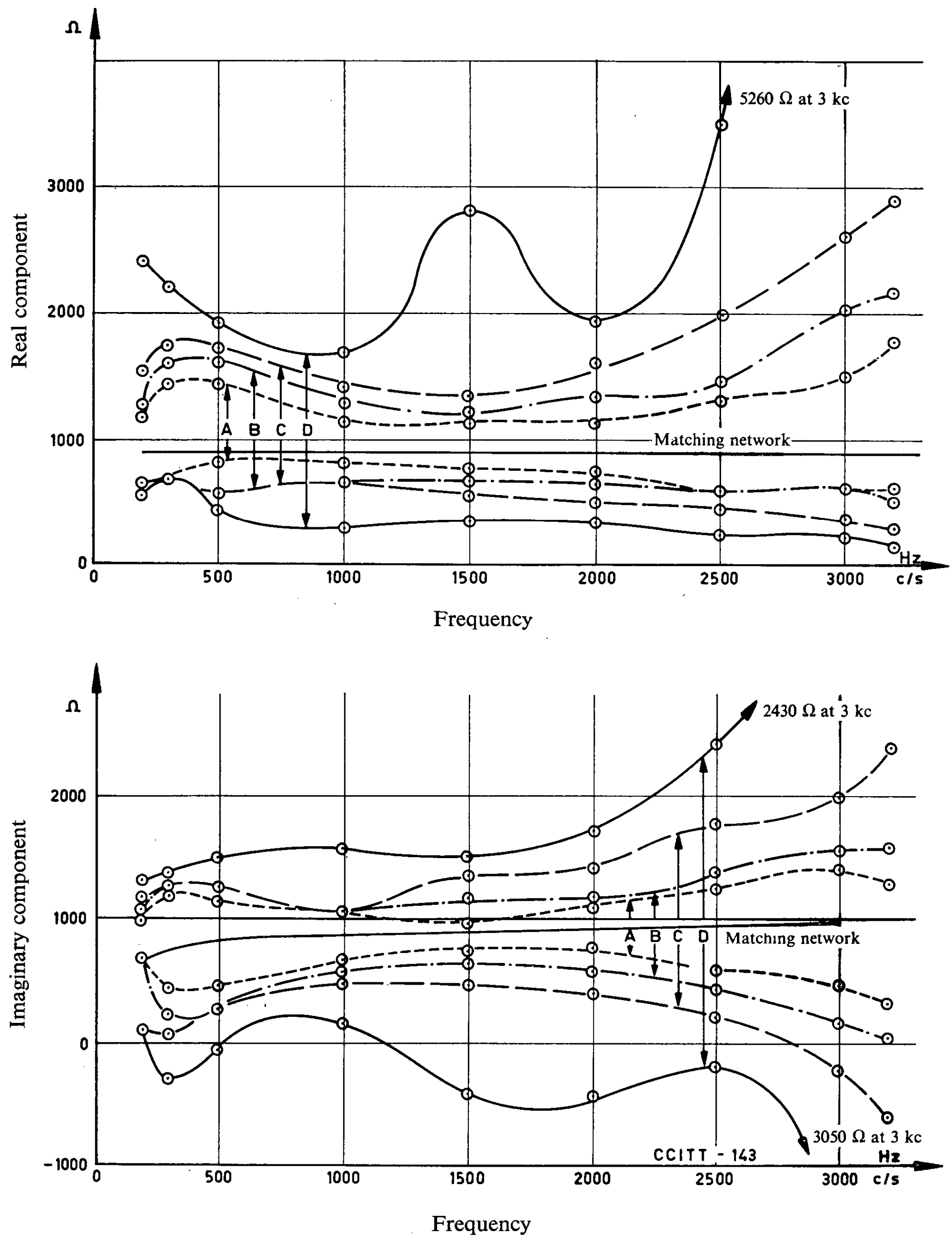
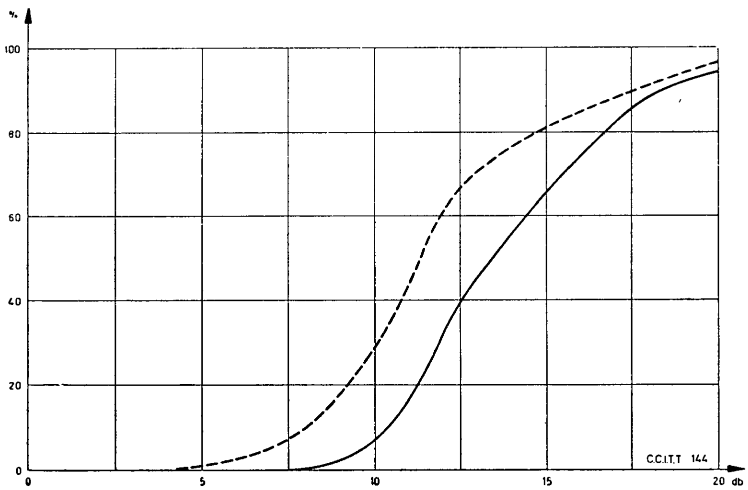


FIGURE 5. — **Distribution of loaded loop input impedance based on return loss performance (includes central office equipment and wiring)**

(A = range of variation for best 25% of loaded plant, B = best 50%, C = best 75%, D = range for all 121 loaded loops)

Figure 6 shows the improvement obtained by means of a correcting network (series insertion of a balanced 32 mH inductance in parallel with a resistance of 1650 ohms).



------ 3 kc return loss
Mean = 12.1 db
Std. dev. = 3.7 db
90% confidence limits on mean = ±0.217 db

———— Echo return loss
Mean = 14.0 db
Std. dev. = 3.3 db
90% confidence limits on mean = ±0.199 db

FIGURE 6. — **Improved echo and singing return loss due to equalizer in trunk circuit**

ANNEX 2

(to Question 19/XII)

Impedance variations in subscribers' lines and telephone sets

(Contribution by the French Administration)

The French Administration has made a number of measurements which provide precise information on the impedance presented over the audio-frequency band by a subscriber's line of variable length terminating in a station in the speaking condition.

The telephone set was of a current type used by the French Administration (U43, local battery, type 328/1).

Measurements were made at the local exchange to which the subscriber was connected. A typical subscriber's line was chosen in each of the following three classes: short lines, 0.5 kilometre long, lines of medium length, 1.5 kilometre long, and lines more than 3 kilometres in length. Three kinds of line were successively studied: underground, overhead, and mixed underground and overhead. Since the telephone sets were of the local-battery type, the distortion ordinarily introduced via the power-supply bridge was eliminated from the measurements.

For the purposes of comparison, a U43 set was measured in exactly the same circumstances and with the same equipment.

The numerical results are shown in Tables III, IV, V, and VI. Figures 1, 2, 3, and 4 correspond to these tables.

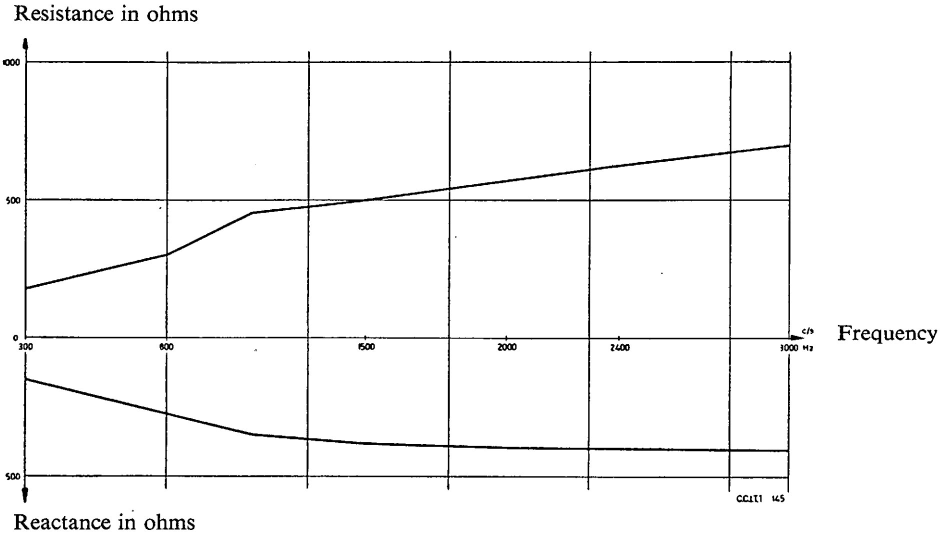


FIGURE 1. — **Impedance of a U43 telephone set**

TABLE III

Impedance of a U43 local-battery set (in the speaking condition)

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	176	145
800	296	268
1100	454	345
1500	497	376
2000	571	390
2400	628	392
3000	700	395
3400	752	389

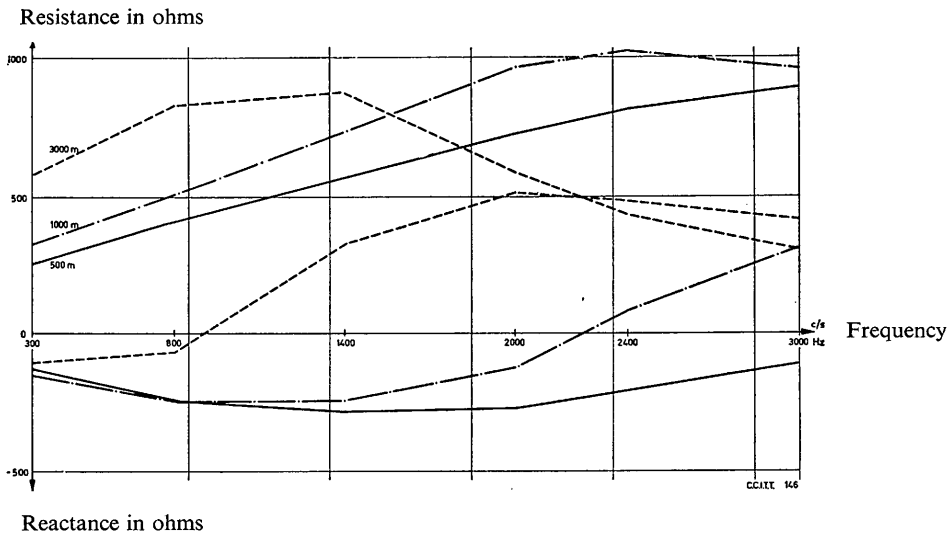


FIGURE 2. — **Impedance of a subscriber's line, in underground cable, terminated by a U43 telephone set**

TABLE IV

Impedance of a subscriber's underground line terminating in a U43 local-battery telephone (in the speaking condition)

Short line

580 m of 0.6-mm cable
Loop resistance: 82 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	255	134
800	400	252
1400	560	290
2000	720	278
2400	810	212
3000	890	113

Medium line

1 km of 0.6-mm cable
Loop resistance: 119 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	320	150
800	500	252
1400	720	248
2000	960	126
2400	1020	−75
3000	960	−300

Long line

3 km of 0.6-mm cable
Loop resistance: 358 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	564	106
800	830	65
1400	870	−318
2000	590	−505
2400	430	−484
3000	300	−415

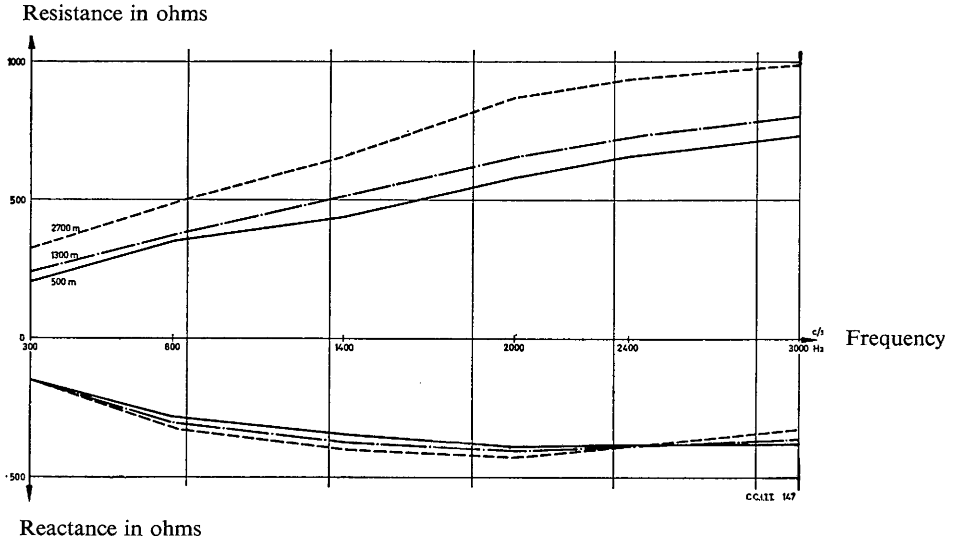


FIGURE 3. — **Impedance of a subscriber's open-wire line terminated by a U43 telephone set**

TABLE V

Impedance of a subscriber's overhead line terminating in a U43 local-battery telephone (in the speaking condition)

Short line

500 m of 11/10 wire
Loop resistance: 23 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	200	150
800	340	285
1400	440	344
2000	580	390
2400	650	378
3000	730	376

Medium line

1300 m of 11/10 wire
Loop resistance: 49 ohms

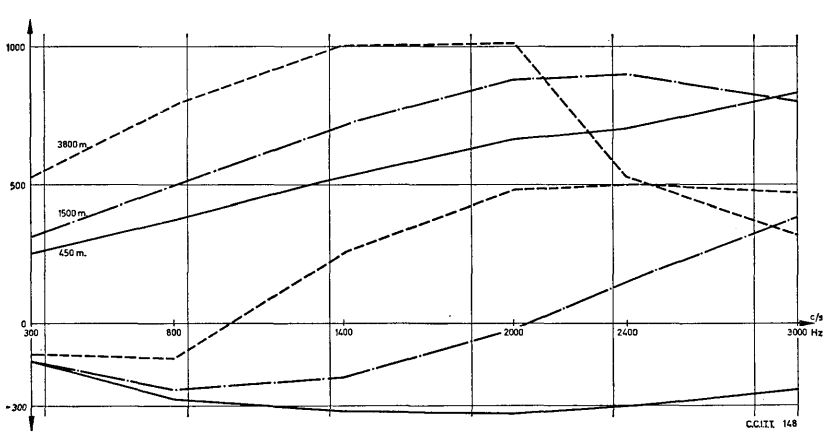
Frequency in c/s	Resistance in ohms	Reactance in ohms
300	235	153
800	365	302
1400	510	370
2000	650	403
2400	720	392
3000	800	358

Long line

2500 m of 11/10 wire
Loop resistance: 90 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	225	145
800	350	282
1400	495	354
2000	640	390
2400	770	378
3000	890	320

Resistance in ohms



Frequency

Reactance in ohms

FIGURE 4. — **Impedance of a subscriber's line—partly open line/partly cable—terminated on a U43 telephone set**

TABLE VI

Impedance of a combined overhead and underground subscriber line terminating in a U43 local-battery telephone (in the speaking condition)

Short line

450 m (half underground, half overhead)
Loop resistance: 57 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	250	140
800	370	268
1400	520	318
2000	660	328
2400	710	302
3000	830	244

Medium line

1500 m (900 m underground, 600 m overhead)
Loop resistance: 136 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	310	136
800	490	242
1400	710	203
2000	880	25
2400	900	−150
3000	800	−376

Long line

3800 m (1400 m underground cable, 1100 m overhead cable, 1300 m overhead wire)
Loop resistance: 338 ohms

Frequency in c/s	Resistance in ohms	Reactance in ohms
300	520	113
800	780	126
1400	1000	−248
2000	1010	−480
2400	525	−497
3000	322	−468

The overhead lines do not substantially change the impedance of the subscriber's set. The line loop resistance is simply added to the resistance of the telephone set itself.

But as far as underground lines and combined underground-overhead lines were concerned (the latter because of their path underground), the impedance measured at the local exchange is quite different from that of the set alone. In particular, the reactance becomes negative when the line is a kilometre or more in length and the measurement frequency is sufficiently high.

The following dispersion was noted at 2000 c/s for the nine lines investigated:

Mean resistance : 743 ohms
Maximum resistance: 1010 ohms (variation of +36%)
Minimum resistance : 580 ohms (variation of −22%)

Mean reactance : 106 ohms
Maximum reactance : 403 ohms
Minimum reactance : −505 ohms

ANNEX 3

(to Question 19/XII)

Study of the distribution in the national network of the balance return loss in local telephone networks

(Contribution by the Italian Administration)

In problems relating to telephone transmission, it is normally assumed that the impedance of local networks at the " trunk line " terminal of the feeding bridge is a pure resistance of 600 ohms.

To determine the degree of approximation of this assumption the balance return loss has been measured in local systems comprising the most commonly used telephone equipment in the national network connected to subscribers' lines of different lengths with conductors of 0.4 and 0.6 mm diameter.

To eliminate impedance variations in subscribers' lines due entirely to construction tolerances, artificial subscriber lines of different lengths were simulated by quadripoles consisting of resistances and capacities determined on the basis of the average characteristics given in the specifications.

To eliminate the effect of instability of the resistance of the carbon microphone capsules, the microphones were replaced by a fixed resistance having the same nominal resistance as the carbon capsule.

Table I shows the average values of balance return loss, for a balance of 600 ohms, as a function of the frequency and the length of the subscriber's line.

For each average value of return loss, the standard deviation σ, corresponding to the different telephone equipment, is also shown.

From Table I it will be seen that for cable lines with 0.4-mm conductors the balance return loss is inadequate for lines of more than 1 kilometre in length.

Moreover, it will be observed that for short lines the values of σ are very high.

For cables with 0.6-mm conductors the balance return loss is inadequate for the whole length of the line.

To make an appreciable increase in the balance return loss it is sufficient to use a balance consisting of a resistance of 1000 ohms and a capacity of 50 nF in parallel.

In Table II we have shown the average balance return loss with the balance mentioned in the preceding paragraph.

To achieve even higher balance return losses, for example always above 10 db, it would be necessary to introduce a hypothetical line in the exchange, on the subscriber's side, for lines of less than 1 kilometre, corresponding to a line with a length of 2 kilometres in 0.6-mm conductors (attenuation of 2 db at 800 c/s).

TABLE I

Balance return loss for a balance of 600 ohms

0.4-mm diameter conductors										
Frequency	Length (km)									
	0.25		1		2		3		4	
c/s	db	σ	db	σ	db	σ	db	σ	db	σ
300	13.95	5.28	15.35	2.44	13.03	1.63	9.86	1.16	7.96	0.79
500	17.45	4.02	17.29	2.69	12.23	2.6	8.8	1.75	7.13	0.56
800	17.7	6.15	15.35	4.22	10.38	2.08	7.86	1.15	6.35	0.72
1000	17.5	7.56	13.75	3.5	9.6	1.76	7.55	0.995	6.1	0.79
1500	15.8	7.62	11.6	3.4	8.26	1.63	7.36	0.994	5.95	0.63
2000	14.26	6.51	10.63	3.29	7.56	1.57	7.63	0.28	6.15	0.53
2500	14.1	7.13	9.7	3.02	6.93	1.49	8.01	0.86	6.45	0.46
3000	13.46	6.92	9.2	3.1	6.61	2.14	8.6	0.88	6.8	0.3
3400	13.1	6.8	8.7	2.82	6.35	1.27	9.08	0.62	7.08	0.35

0.6-mm diameter conductors										
Frequency	Length (km)									
	2		4		6		8		10	
c/s	db	σ	db	σ	db	σ	db	σ	db	σ
300	14.5	2.3	12.2	1.73	9.45	1.1	7.5	0.75	6.3	0.5
500	16.5	2.73	11.5	2.49	8.48	1.2	6.7	1.2	5.8	0.9
800	15.15	5	9.6	2.3	7.53	1.1	6.3	0.89	5.8	0.86
1000	13.28	3.5	8.8	1.64	7.2	0.98	6.3	0.82	6.06	0.42
1500	10.8	2.79	7.36	1.27	7.05	0.67	6.75	0.35	6.78	0.29
2000	9.4	2.47	6.75	1.05	7.2	0.2	7.2	0.27	7.2	0.15
2500	8.15	2.04	6.2	0.84	7.36	0.52	7.4	0.51	7.35	0.075
3000	7.46	2	5.9	0.69	7.5	0.28	7.48	0.23	7.25	1.29
3400	6.8	1.75	5.76	0.64	7.58	0.3	7.5	0.055	7.1	0.69

TABLE II

Balance return loss for a balance consisting of a 1000-ohm resistance in parallel with a capacity of 50 nF

0.4-mm diameter conductors										
Frequency	Length (km)									
	0.25		1		2		3		4	
c/s	db	σ	db	σ	db	σ	db	σ	db	σ
300	10.03	3.56	14.3	4.04	24.1	6.69	21.08	3.17	15.4	2.21
500	10.5	3.29	15.7	4.26	22.3	6.41	18.4	5.23	14.3	2.26
800	9.8	2.48	14.2	3.38	17.4	6.32	15.4	2.82	12.1	2.04
1000	9.08	2.36	13.1	3.11	15.9	4.23	15.1	2.36	12.1	1.59
1500	7.6	2.2	11.2	3	13.9	3.75	16.3	2.93	13.5	1.58
2000	6.6	2.22	10.2	3.17	13.9	3.62	20	3.5	16.2	1.9
2500	5.83	2.21	9.7	3.09	14.6	3.62	24.9	3.33	20.06	1.96
3000	5.05	2.1	8.9	2.43	15.08	2.27	27.4	4.52	25.7	5.82
3400	4.66	2.12	8.9	2.37	16.2	2.18	22.78	4.16	34.7	7.75

0.6-mm diameter conductors										
Frequency	Length (km)									
	2		4		6		8		10	
c/s	db	σ	db	σ	db	σ	db	σ	db	σ
300	14	4.1	21.8	5.82	18.7	3.6	14.06	1.61	11.1	1.645
500	17.3	5.83	21.4	2.01	17.08	3.795	12.58	1.71	10.3	1.065
800	17.1	4.48	20.01	3.88	15.1	3.07	11.8	1.405	10.4	0.675
1000	16.8	4.28	19.2	4.01	14.55	2.425	11.83	1.015	10.7	0.46
1500	17.15	4.7	17.06	3.1	14.15	1.22	12.26	0.7	11.6	0.69
2000	17.8	5.19	15.8	2.54	14.18	0.86	12.6	1.345	12.1	1.29
2500	18.7	4.52	15.2	1.55	14.15	0.51	12.76	0.51	12.5	0.1
3000	20.05	3.8	14.9	1.66	14.1	0.42	12.7	0.68	12.58	0.32
3400	21.6	4.67	14.6	1.06	12.28	0.3	12.6	0.67	12.58	0.32

The average balance return loss obtained with very short subscribers' lines and with this hypothetical line on the subscribers' side are shown in Table III.

The standard deviations, however, remain fairly high and it would be desirable to limit them by using telephone equipment with uniform electrical characteristics.

By using a hypothetical line for very short subscribers' lines there is also the advantage of improved sidetone balance and a smaller difference in levels for different lengths of subscribers' lines.

TABLE III

Balance return loss for an artificial line corresponding to a 2-km line of 0.6 mm, with 1000 ohms balance, 50 nF

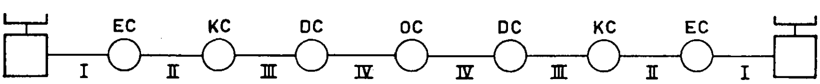
0.4-mm diameter conductors						
Frequency	Length (km)					
	0		0.25		1	
c/s	db	σ	db	σ	db	σ
300	12.7	3.447	14.2	3.864	19.2	4.705
500	15.86	4.89	18.76	6.563	21.98	2.685
800	17.2	5.11	19.36	5.505	21.3	3.89
1000	17.23	5.088	18.86	4.73	21.15	4.613
1500	17.78	5.052	19.2	4.495	19.58	4.475
2000	18.46	4.99	19.4	3.895	18.4	3.725
2500	19.5	4.647	20.01	3.194	17.4	2.666
3000	21	4.65	21.1	3.05	17.1	2.159
3400	21.98	4.085	21.4	2.295	16.58	1.546

ANNEX 4

(to Question 19/XII)

Distribution of balance return loss in the telephone network of the Netherlands

1. General structure of the telephone network



I = local cable
II = coil-loaded cable

III = Coil-loaded cable or carrier system
IV = Carrier system

EC = local exchange
KC = primary centre
DC = secondary centre
OC = international exchange

FIGURE 1

Figure 1 shows a connection that can be established, with a maximum number of links between two end exchanges. In principle it is possible to set up a connection with fewer links, e.g. the connection of Figure 2, where the link between two district exchanges and the possible overflow exchange is absent because the connection shown is established within one district.

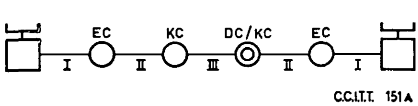


FIGURE 2

The geographic structure is represented in Figure 3.

In the drawings the connection between end exchange and subscriber has been simplified; as a matter of fact there may by a P.B.X. and/or a satellite exchange between the end exchange and the subscriber. These situations have not been included in the measurements.

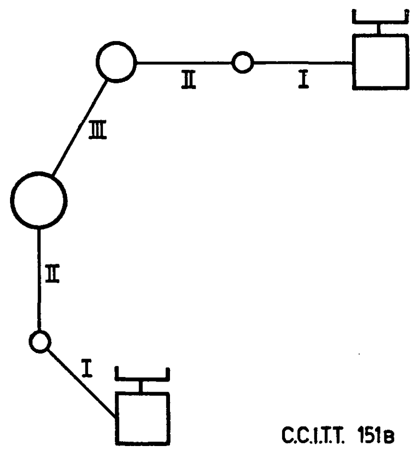


FIGURE 3

2. The points of reflection in the telephone system

The points of reflection, i.e. the points where reflection occurs, and where the echo attenuation is of essential importance, are usually the points where two transmission systems are linked, e.g. when a carrier link is connected to a coil-loaded cable or to a homogeneous cable.

If we assume that the entire telephone system is constructed in such a manner that the exchanges between carrier links have four-wire switching and that no coil-loaded cables are used between the district exchanges and the sector exchanges but only carrier links (a situation that is being aimed at), the points of reflection can be classified according to Figure 4.

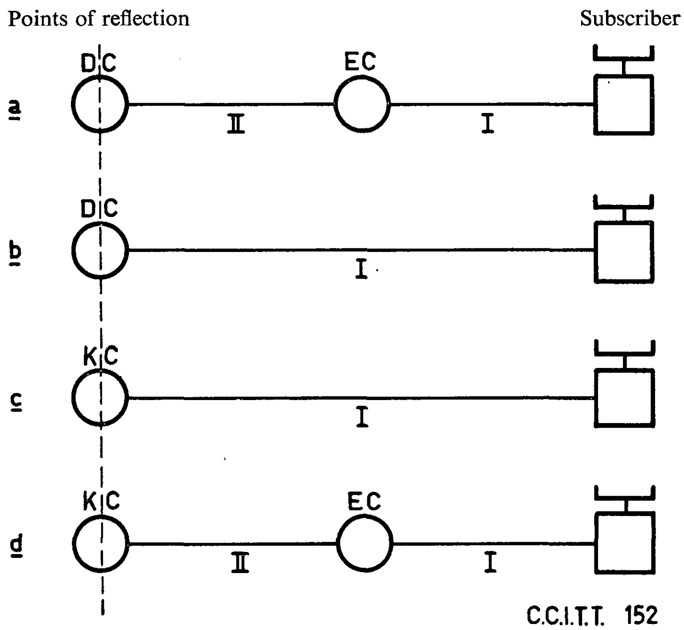


FIGURE 4

From this classification it appears that the points where the four-wire circuits are connected to the two-wire circuits (coil-loaded or homogeneous cable) have been assumed to be the points of reflection. The reflection arising when a coil-loaded cable is connected to a homogeneous cable has not been taken into account, because for the region of speech the delay of a homogeneous cable is so small that the impedance formed by the local cable and the telephone set can be considered to be one variable.

In fact this means that hybrid coils are concerned terminated by means of a homogeneous cable and a telephone set, either or not via a coil-loaded cable.

From this difference it appears that a division into two groups is to be expected for the echo attenuations; the measurements on terminals with coil-loaded cable will show a tendency towards a higher average.

For this and other reasons which will not be mentioned here, it has been investigated whether it is desirable that the circuits without a coil-loaded cable should be corrected in such a manner that they are equivalent to circuits with coil-loaded cables.

Therefore, after the echo attenuation measurements have been taken in the original situation, networks have also been provided in the circuits *b* and *c* of Figure 4 to examine the influence of these networks on the echo attenuations. These networks have been called " 3-db filters ", because they introduce the target value for the attenuation of a coil-loaded cable (3 db) and furthermore an approximately equal limitation of the bandwidth ($f_0 = 4000$ c/s).

Because a coil-loaded cable can moreover be corrected to obtain a better adjustment to the hybrid coil, together with improved echo attenuation, the series has been supplemented with circuits with so-called *m* sections, which are networks that transform, within certain limits, the impedance of coil-loaded cables beginning with half-sections into an impedance having the character of a resistance.

The 3-db filters have been inserted in the places indicated in Figure 5.

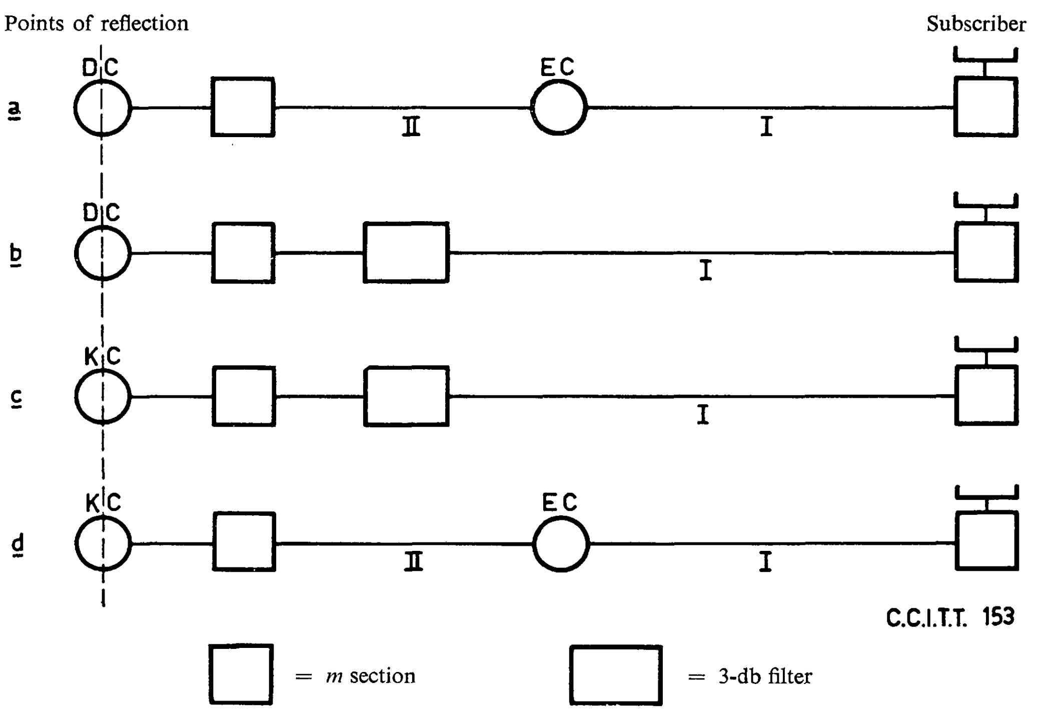


FIGURE 5

It should be observed that a difference has been made between the measurements made in a district exchange and those made in a sector exchange to examine a possible geographic influence on the local cable network. Therefore the configurations of Figure 4 *b* and Figure 4 *c* and also those of Figure 5 *b* and Figure 5 *c* are in principle the same, but in practice the results can differ.

This holds to an even greater extent for the configurations of Figure 4 *a* and Figure 4 *d* and for those of Figure 5 *a* and Figure 5 *d*. As a matter of fact the situations indicated by *a* hold for the district exchanges in one of the three local telephone Services (The Hague, Rotterdam and Amsterdam) which are connected to the end exchanges by means of coil-loaded cables, but which cables have properties that, owing to geographical circumstances, deviate considerably from the standardized cable types used between EC and KC, i.e., between the local exchange and the primary centre.

To restrict the number of measurements, the measuring of a connection between a district exchange and an end exchange via a coil-loaded cable has been abandoned because it may be expected that such a situation will deviate little from those represented in Figure 4 *c* and Figure 5 *c*.

3. Definition of echo attenuation

The reflection occurs in the hybrid coil according to the arrangement of Figure 6, which need not be explained any further. From it we can derive that the attenuation of one four-wire side to the other four-wire side is determined by:

$$\frac{e}{2u} = 2\left(\frac{Z_1+Z_2}{Z_1-Z_2}\right)\left(\frac{1+\dfrac{Z_4}{Z_3}+\dfrac{2(Z_1Z_2+Z_3Z_4)}{Z_3(Z_1+Z_2)}}{4}\right) \tag{1}$$

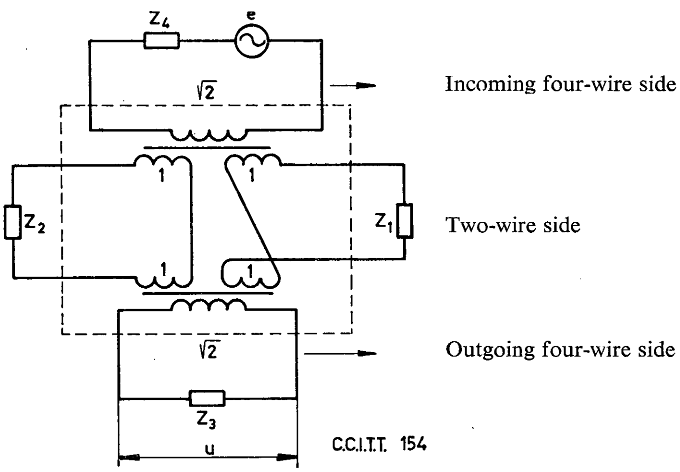


FIGURE 6

Introduction of $Z_1 \approx Z_2 \approx Z_3 \approx Z_4$ reduces this equation to

$$\frac{e}{2u} \approx 2\left(\frac{Z_1+Z_2}{Z_1-Z_2}\right)$$

This is the overflow of one four-wire side to the other four-wire side.

The echo attenuation a_e is now defined as:

$$a_e = 20\log\left|\frac{Z_1+Z_2}{Z_1-Z_2}\right| \approx 20\log\frac{e}{4u} \tag{2}$$

The measurements have been made according to the diagram of Figure 7. According to this diagram

$$a_m = 20\log\left|\frac{e}{u}\right| = 20\log\left|\frac{Z_1+Z_2}{Z_1-Z_2}\right| \tag{3}$$

where Z_2 is the balancing circuit and Z_1 the two-wire impedance.

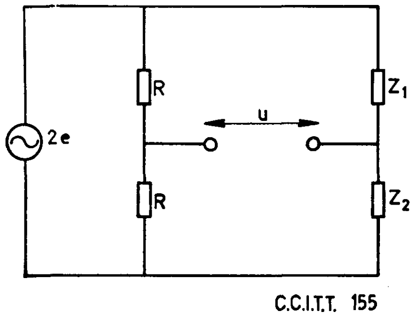


FIGURE 7

As a balance impedance a series connection of a resistor of 820 ohms ±1% and a condensor of 1 μF ± 5% has been used. This is the balance circuit normally used for hybrid coils.

4. Measurements and measuring results

The number of measurements made is represented in the table below.

Situation	Number
$A_{1,2}$	304
$B_{1,2}$	149
$C_{1,2}$	218
$D_{1,2}$	355

The numbers depend on the distribution of the subscribers over the entire telephone system.

The measurements have been made for the frequencies 300, 500, 800, 1000, 1500, 2000, 2500, 3000, 3800, 4000 and 4300 c/s.

From the results the average and the standard deviation have been derived for each group and each frequency.

The average is defined by:

$$\mu = \frac{\sum_{j=1}^{j=n} a_j}{n}$$

and the standard deviation by:

$$\sigma = \sqrt{\frac{\sum_{j=1}^{j=n} (\mu - a_j)^2}{n}}$$

The average values are shown in Figure 8 and Figure 9. In this latter figure the influence of the correction networks can be seen.

The standard deviations are represented in Figure 10 and Figure 11.

To get some insight into their nature, a distribution is shown for every situation holding for a frequency of 1000 c/s.

For the known situation it is represented in Figures 12-19 inclusive.

Figures 8-11 also contain a line marked *E*. This line represents the over-all result, in which have been incorporated situations A, B, C and D.

The distribution of the echo attenuation over the entire country for a frequency of 1000 c/s is shown in Figure 20 and Figure 21. Here the number of measurements is larger than the total of A, B, C and D because more measurements have been incorporated in it which could not be distributed over the groups A, B, C and D.

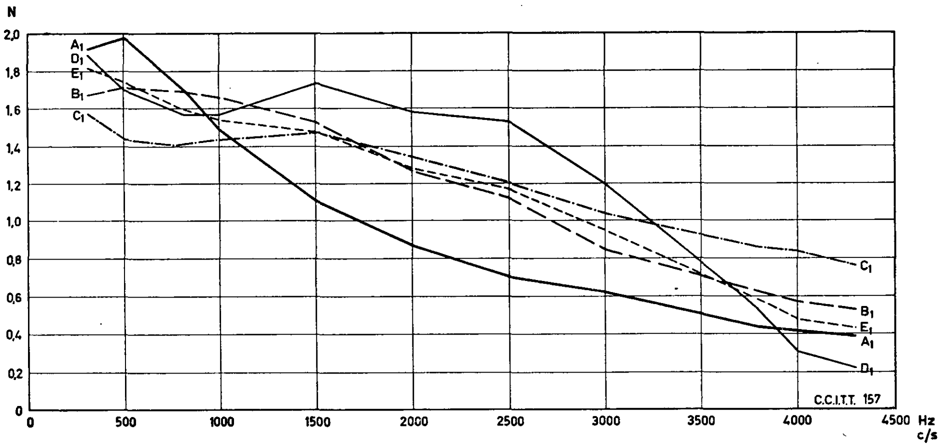


FIGURE 8. — **Mean balance return loss (without correcting networks)**

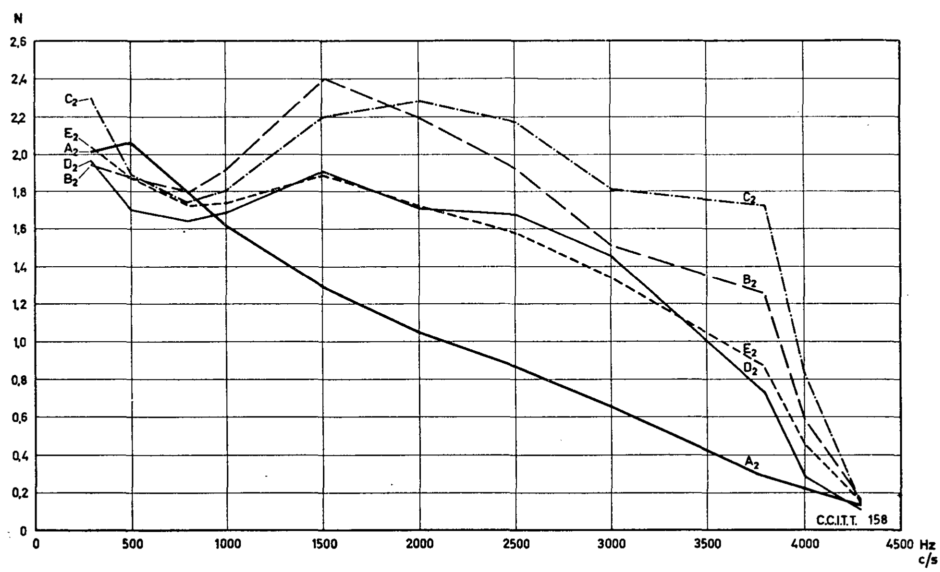


FIGURE 9. — **Mean balance return loss (with correcting networks)**

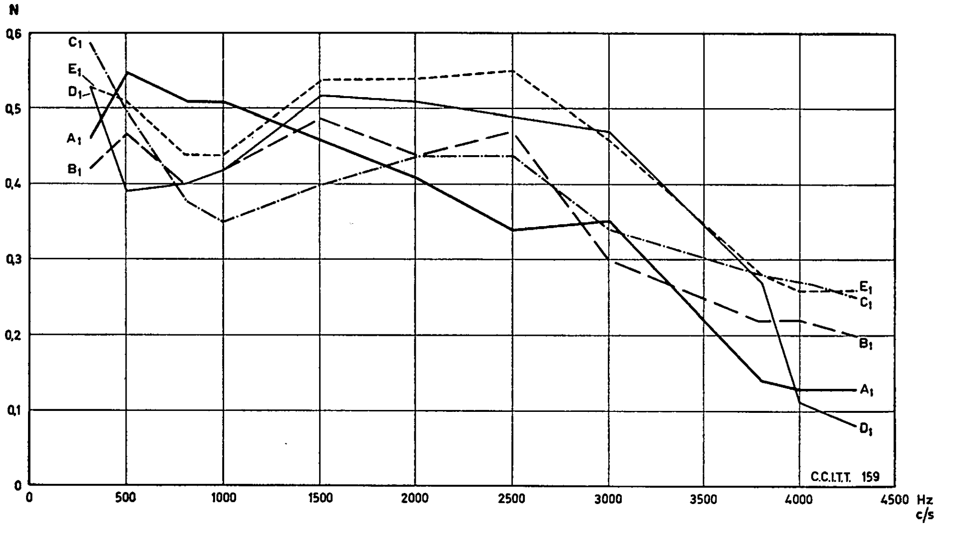


FIGURE 10. — **Balance return loss: standard deviation (without correcting networks)**

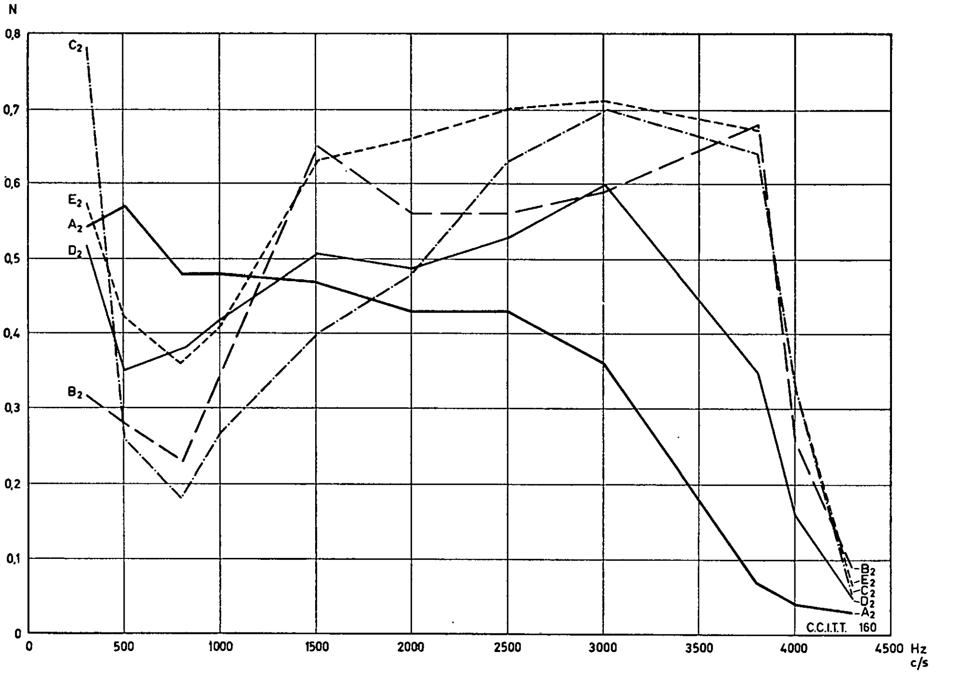


FIGURE 11. — **Balance return loss: standard deviation (with correcting networks)**

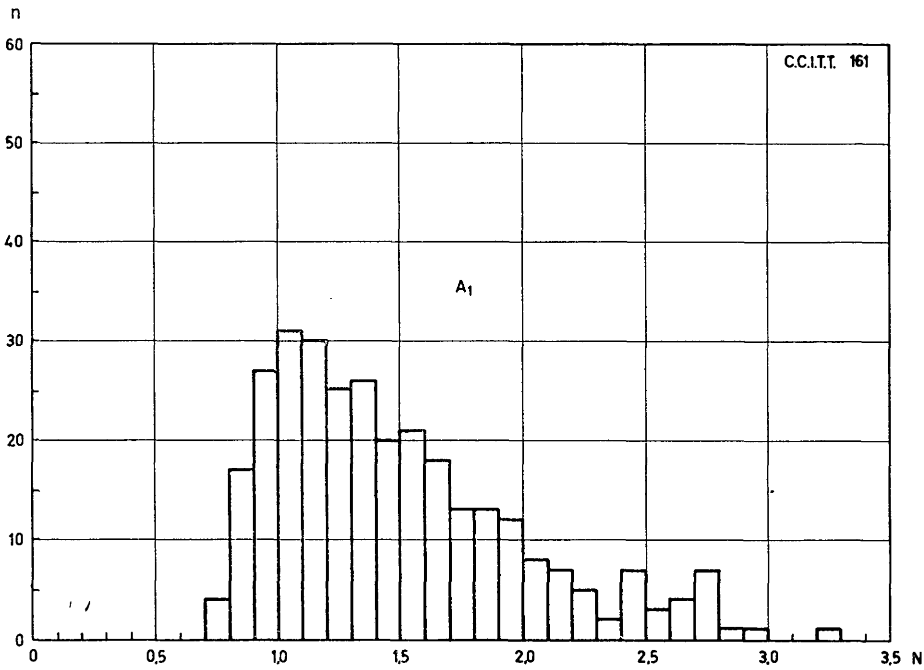


FIGURE 12. — **Balance return loss distribution for 1000 c/s**

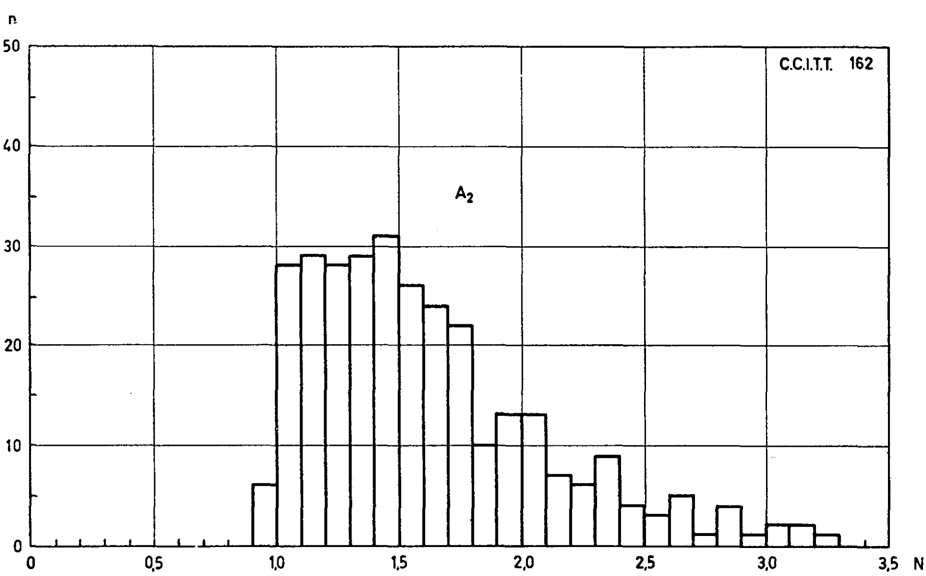


FIGURE 13. — **Balance return loss distribution for 1000 c/s**

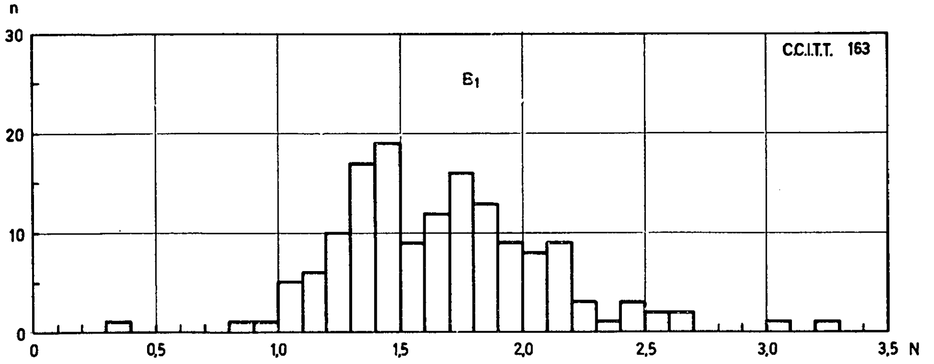


FIGURE 14. — **Balance return loss distribution for 1000 c/s**

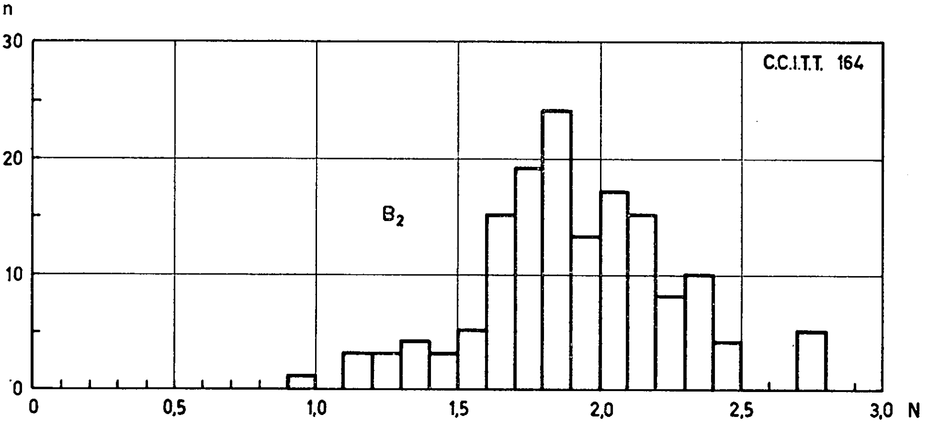


FIGURE 15. — **Balance return loss distribution for 1000 c/s**

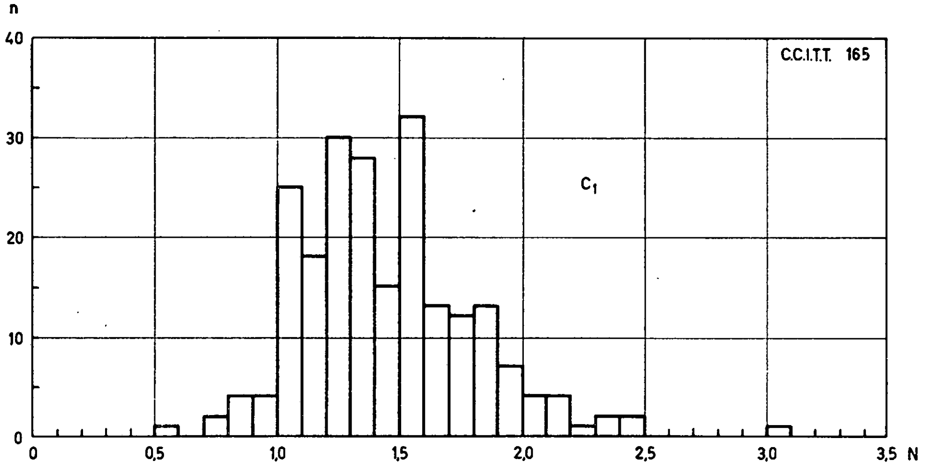


FIGURE 16. — **Balance return loss distribution for 1000 c/s**

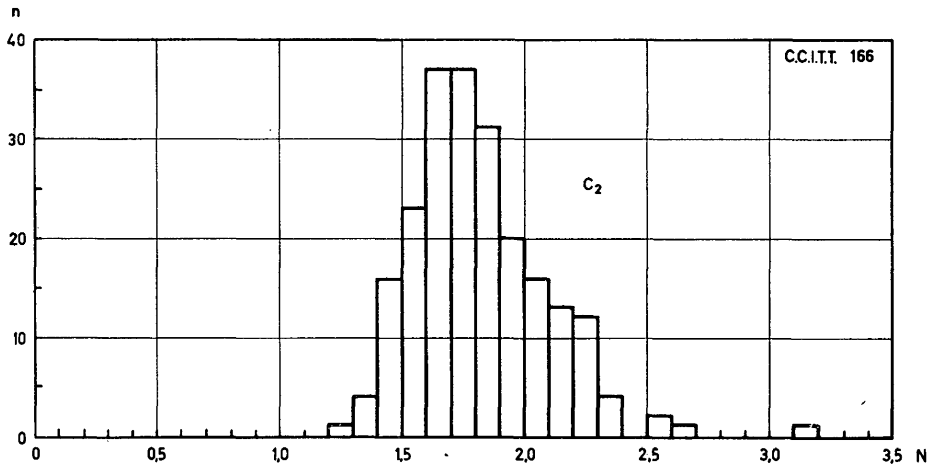


FIGURE 17. — **Balance return loss distribution for 1000 c/s**

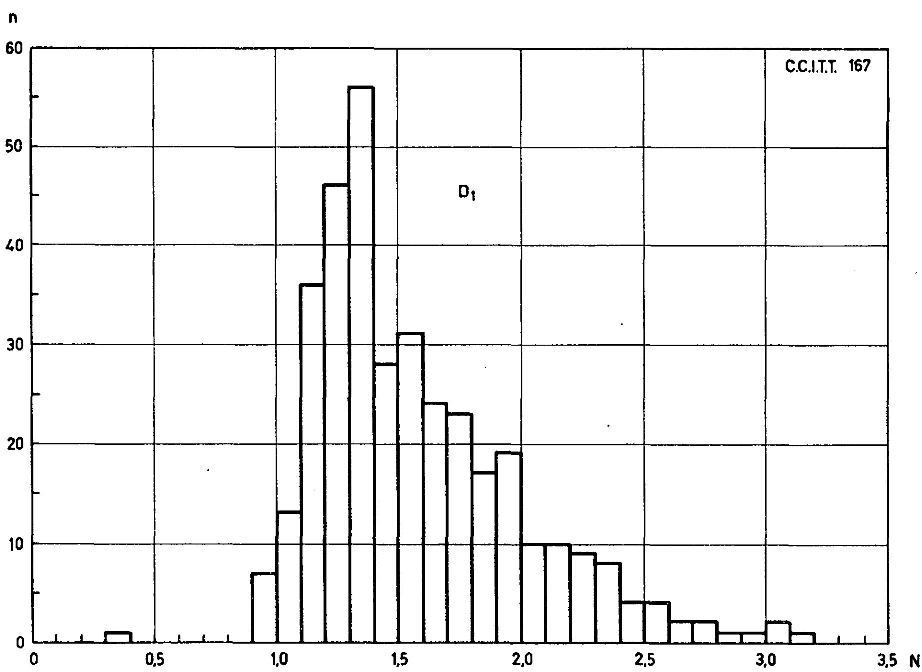


FIGURE 18. — **Balance return loss distribution for 1000 c/s**

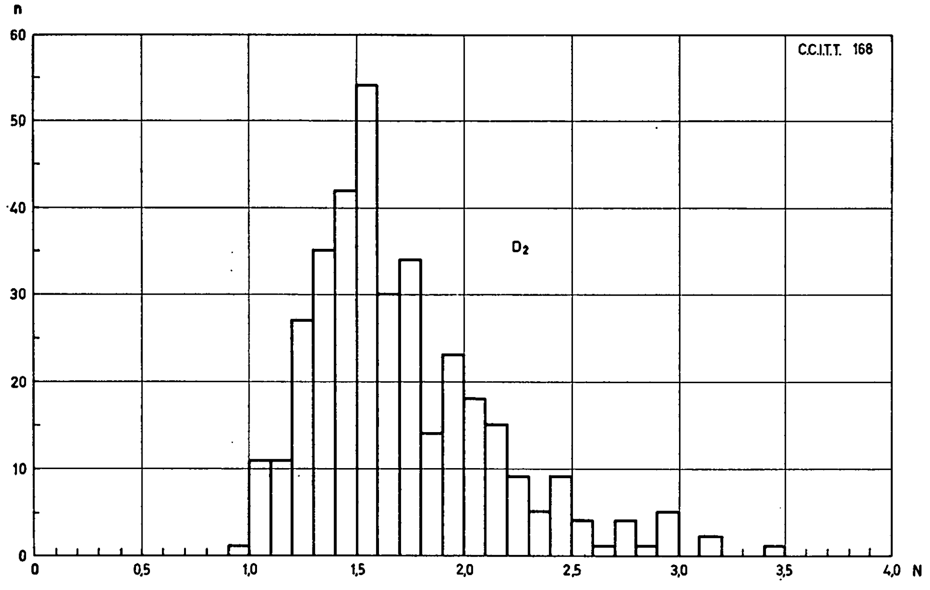


FIGURE 19. — **Balance return loss distribution for 1000 c/s**

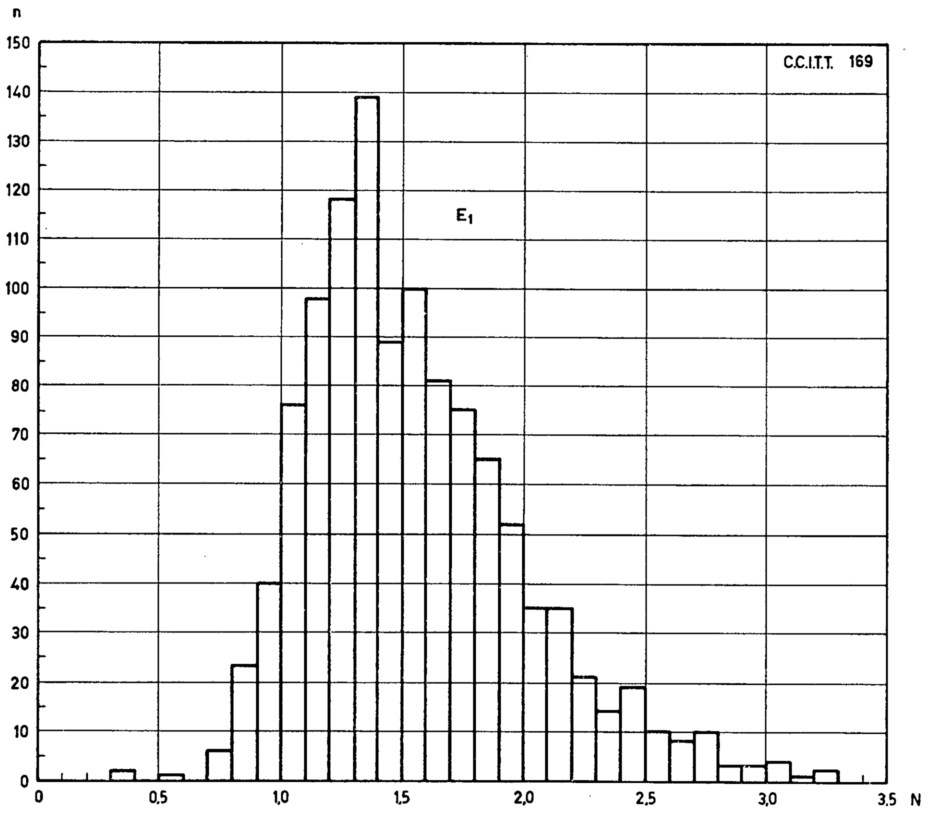


FIGURE 20. — **Balance return loss distribution for 1000 c/s**

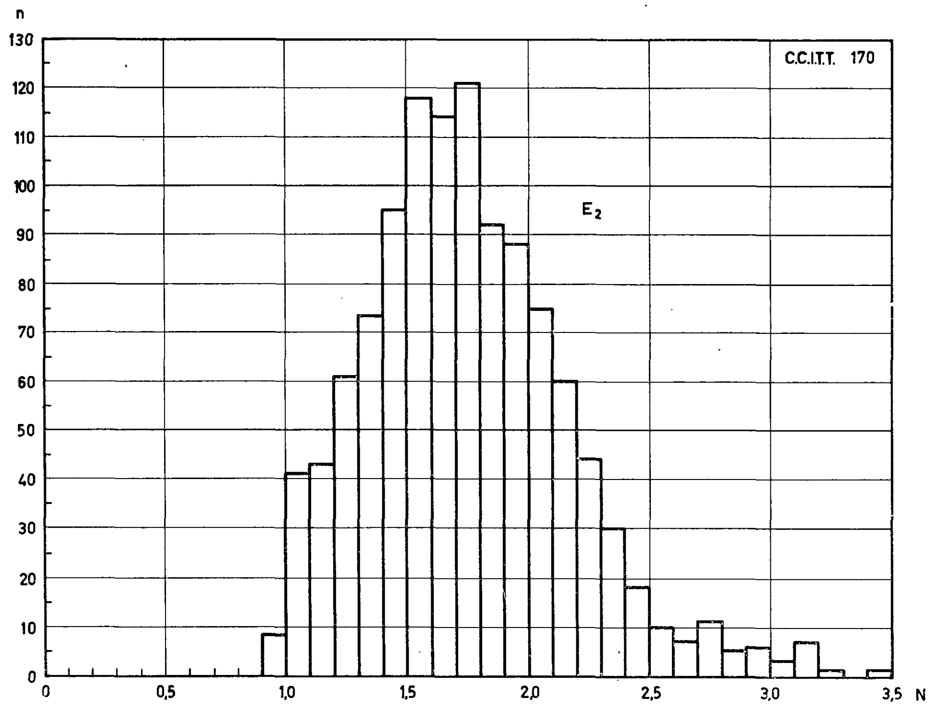


FIGURE 21. — **Balance return loss distribution for 1000 c/s**

ANNEX 5

(to Question 19/XII)

Contribution by the Federal Republic of Germany

The input impedance of a subscriber's station depends on the type of receiving capsule used. The impedance of the magnetic receiving capsule has a strong inductive component and is, for this reason, relatively small at low frequencies and increases greatly at the high frequencies of the frequency band transmitted. Similarly, the trend of the impedance of a telephone set equipped with a magnetic capsule depends greatly on the frequency, as is shown by the modulus and phase angle values of the impedance of a W48 telephone set equipped with a magnetic receiver capsule, as shown in Table 1 a).

The impedance of a dynamic receiver capsule is practically resistive. Because of the small inductive component it increases only slightly with frequency. Similarly, the impedance of a telephone apparatus equipped with a dynamic receiving capsule varies relatively little as a function of frequency, as can be seen from Table 1 b). Whereas the impedance of apparatus equipped with a magnetic receiving capsule varies between -39% and $+355\%$, in relation to a resistance of 450 ohms at 800 c/s, the same apparatus equipped with a dynamic receiving capsule varies only from -18% to $+21\%$, referred to 570 ohms at 800 c/s.

TABLE 1

Impedance of the W48 telephone set

	a) with magnetic receiver capsule				b) with dynamic receiver capsule			
f (c/s)	300	800	1600	3400	300	800	1600	3400
Z (ohms)	275	450	690	1600	465	570	625	690
argument	36°	48°	49°	47°	33°	22°	18°	23°

In transmission technique, it is preferable to base judgment on the balance return loss rather than on the impedance. In this case, it is advisable to determine the balance return loss of a subscriber's telephone system seen through the battery supply bridge in the switching centre. In this way the battery supply bridge, the subscriber's line and the telephone set are included. In general, it is customary to refer the balance return loss to a pure resistance of 600 ohms.

Tables 2 a) and 2 b) show the balance return losses for a W48 set when subscribers' lines of different diameters (0.4 and 0.6 mm) and different lengths (0.25 to 12 km) are used. In Table 2 a) the set was equipped with a magnetic receiver capsule and in Table 2 b) with a dynamic receiver capsule.

TABLE 2

Balance return loss of the input impedance in a subscriber's system with a W48 telephone set, referred to 600 ohms

	a) with magnetic receiver capsule				b) with dynamic receiver capsule			
f (c/s)	300	500	2500	3400	300	500	2500	3400
l/d								
0.25/0.4	8	8	8	8	10	15	19	17
1/0.4	11	14	7	7	14	30	12	9
2/0.4	19	31	5	5	17	18	7	6
2/0.6	12	14	7	5	14	29	10	8
4/0.6	19	34	5	4	11	12	6	5
6/0.6	19	15	5	5	14	12	5	5
8/0.6	16	8	5	6	13	10	6	6
10/0.6	12	9	6	6	11	8	6	6
12/0.6	12	7	7	7	16	7	7	7

l = line length in km. d = conductor diameter in mm.

Table 3 a) shows the balance return loss for the new W61 set equipped with a dynamic receiver capsule, for which the telephone set is better matched.

Table 3 b) shows the balance return loss for the same set, but in the switching centre a complex balance composed of a parallel connection of 800 ohms and 50 nF was taken as the basis, instead of a 600-ohm balance.

TABLE 3

Balance return loss (in db) of the input impedance in a subscriber's system with a W61 telephone set and a dynamic receiver capsule

	a) referred to 600 ohms				b) referred to 800 ohms/50 nF			
f (c/s)	300	500	2500	3400	300	500	2500	3400
l/d								
0.25/0.4	9	12	22	23	11	11	11	10
1/0.4	15	19	22	13	15	17	18	18
2/0.4	21	20	9	8	11	15	18	22
2/0.6	15	19	12	10	15	17	27	25
4/0.6	22	19	8	7	12	18	16	16
6/0.6	19	13	7	6	9	15	15	13
8/0.6	14	10	7	7	8	12	14	13
10/0.6	13	9	7	7	6	11	14	14
12/0.6	11	8	7	7	6	10	15	14

l = line length in km. d = conductor diameter in mm.

By inserting, in short subscribers' lines up to 1 km/0.4 mm diameter a pad which replaces a subscriber's line of about 2 km length and 0.6 mm diameter, we obtain the values shown in Table 4 a) in the case of a balance by 600 ohms, and the values indicated in Table 4 b) in the case of a complex balance by a parallel connection of 800 ohms and 50 nF.

TABLE 4

Balance return loss of the input impedance in a W61 subscriber's system with a pad and a line-length of 0 km

	c/s	300	500	2500	3400
a) Balance 600 ohms	db	15	19	11	9
b) Balance 800 ohms/50 nF	db	16	19	31	31

ANNEX 6

(to Question 19/XII)

Terminating set with improved attenuation characteristics

(Contribution by the Austrian Administration)

Introduction

As long ago as 1958, the Austrian Administration introduced terminal equipment with considerably greater attenuation in the return channel into the fully-automatic service, and since then such equipment has been used on a steadily increasing scale. In such equipment, the return-channel loss, i.e., the loss in the four-wire channel going from the incoming side to the outgoing side, is greater than it is for conventional four-wire terminating sets with differential transformers or with resistances.

In what follows, we shall discuss the physical principles of the new arrangement, then the practical applications, emphasizing the benefits accruing for transmission and switching techniques.

1. *Bridge arrangement with impedances independent of the impedance of the two-wire circuit*

No electric current flows through the branch $c - d$ of the bridge arrangement (see Fig. 1) for any value of the impedance z_2, if the resistances r_x, r_y, N_x and the inductances L_x and L_y have the values shown in the figure. It should be noted that the factors ρ and $(1-\rho)$ are equivalent to the ratio of the subdivision formed by the connecting point d of the secondary winding of the transformer U_3.

The complex loadings of the four elements of the bridge arrangement are as follows:

$$\left.\begin{aligned} W_{a-c} &= \rho\,(r_x + N_x + \mathrm{j}\,\omega\,L_x) \\ W_{c-b} &= (1-\rho)\,(r_x + N_x + \mathrm{j}\,\omega\,L_x) \\ W_{a-d} &= \rho\left[r_y + z_2\left(\frac{W_{3s}}{W_{3P}}\right)^2 + \mathrm{j}\,\omega\,L_y\right] \\ W_{d-b} &= (1-\rho)\left[r_y + z_2\left(\frac{W_{3s}}{W_{3P}}\right)^2 + \mathrm{j}\,\omega\,L_y\right] \end{aligned}\right\} \qquad (1)$$

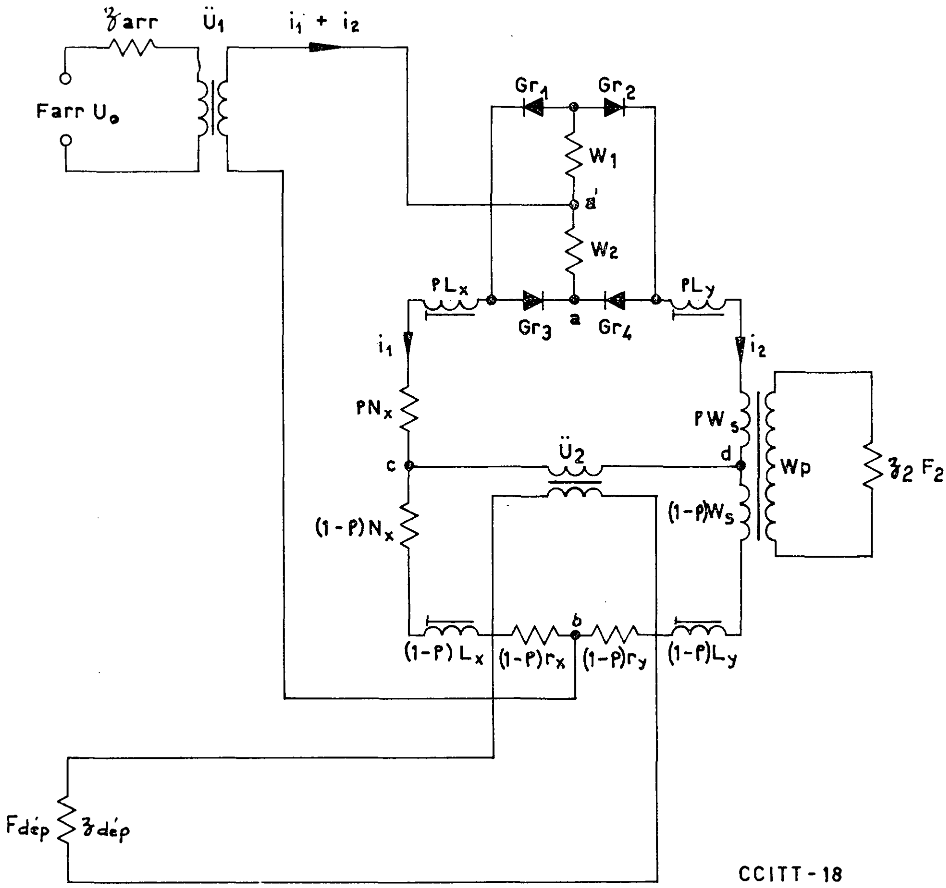


FIGURE 1. — **Bridge arrangement with branches independent of the impedance z_2**

No current flows through the diagonal $c-d$ of the arrangement provided the products of the complex loads of the opposed elements are the same.

Hence, we must have:

$$W_{a-c} \cdot W_{d-b} = W_{a-d} \cdot W_{c-b} \qquad (2)$$

If, now, we insert the figures derived from equations (1) into equation (2), it will be readily seen that electrical balance subsists for all values of the impedance z_2, and even when the current i_3 is zero, that is to say, when the primary winding of the transformer U_3 is open. If this be the case, we shall get the following for the impedances W_{a-d} and W_{d-b}:

$$W_{a-d} = \rho\,[r_y + \mathrm{j}\,\omega\,L_y] + \mathrm{j}\,\omega\,k\,\rho^2\,W_{3s}^2 + \mathrm{j}\,\omega\,k\,\rho\,(1-\rho)\,W_{3s}^2 =$$

$$= \rho\,[r_y + \mathrm{j}\,\omega\,L_y + \mathrm{j}\,\omega\,k\,W_{3s}^2], \qquad k = \frac{\mu_0\,\mu}{l}$$

$$W_{d\text{-}b} = [1 - \rho]\, [r_y + \mathrm{j}\, \omega\, L_y] + \mathrm{j}\, \omega\, k\, (1 - \rho)^2\, W_{3s}^2 + \mathrm{j}\, \omega\, k\, \rho\, (1 - \rho)\, W_{3s}^2 =$$

$$= [1 - \rho\, [r_y + \mathrm{j}\, \omega\, L_y + \mathrm{j}\, \omega\, k\, W_{3s}^2].$$

The elements $\mathrm{j}\, \omega\, k\, \rho\, (1-\rho)\, W_{3s}^2$ make allowance for the mutual inductances of the partial winding $\rho\, W_{3s}$ and the partial winding $(1-\rho)\, W_{3s}$ and vice versa, since the electrical balance is not disturbed.

Let us now write z_{arr} for the impedance of the part of a four-wire circuit transmitting the incoming current. In similar fashion we shall write z_{dep} for the impedance of the same four-wire circuit, outgoing side. In general $z_{\text{arr}} = z_{\text{dep}}$. We have kept a different expression for the two directions of transmission so as to be able to describe those two directions unambiguously. z_2 is the impedance of the two-wire circuit which is to be connected to the four-wire circuit.

The bridge arrangement, as shown in Figure 1, represents, with the resistance r_0 and the matching transformers U_1 and U_2, a four-wire terminating set which, for any value of the impedance z_2, has an infinite loss from z_{arr} to z_{dep}.

For conventional four-wire terminating sets with differential transformers or resistances, the loss in the channel from z_{arr} to z_{dep} (disconnected two-wire circuit) amounts to 0.9 N. This is only double the loss in the ordinary channel, that is to say, from z_{arr} to z_2. As soon as the two-wire circuit is connected, this loss increases by an amount equivalent to the balance return loss.

The present arrangement should terminate the four-wire circuit perfectly, if reflections are to be avoided, that is to say, the impedance must be z_{arr} for the direction of transmission F_{arr} to F_2. Hence the equation:

$$\left(\frac{W_{1P}}{W_{1s}}\right)^2 [r_0 + R_{ab}] = z_{\text{arr}} \tag{3}$$

must be respected. $\left(\frac{W_{1P}}{W_{1s}}\right)$ represents the transformation ratio of the transformer U_1 related to the terminals F_{arr}; R_{ab} stands for the impedance of the bridge arrangement for the connecting points a and b. R_{ab} is given by the following equation:

$$R_{ab} = \frac{[r_x + N_x + \mathrm{j}\, \omega\, L_x] \left[r_y + \mathrm{j}\, \omega\, L_y + z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2\right]}{r_x + r_y + N_x + \mathrm{j}\, \omega\, [L_x + L_y] + z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2} \tag{4}$$

Equation (3) gives the transformation ratio of the transformer which we shall have to maintain for matching of the impedance, the resistance r_0 and the complex loads on the bridge having been already chosen.

We shall still have to determine the loss of the four-wire terminating set for the channel F_{arr} to F_2. To this end, the reference power N_v will have to be compared with the power consumed by the impedance z_2.

For the reference power, we have the equation:

$$N_v = \frac{U_0^2}{4\, z_{\text{arr}}} \tag{5}$$

For the power consumed by the impedance Z_2, we have:

$$N_2 = U_{ab}^2 \frac{z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2}{\left[r_y + \mathrm{j}\, \omega\, L_y + z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2\right]^2} \tag{6}$$

In this formula, U_{ab} is the voltage at the points a and b of the bridge arrangement and we can easily derive it from the equation:

$$U_{ab} = U_0 \left(\frac{W_{1s}}{W_{1P}}\right) \frac{R_{ab}}{z_{\text{arr}} \left(\frac{W_{1s}}{W_{1P}}\right)^2 + r_0 + R_{ab}} \tag{7}$$

in which U_0 is the applied voltage. To calculate the impedance R_{ab} of the bridge wiring, we have equation (4). We shall then get the following for the loss α of the bridge arrangement:

$$N_v = e^{2\alpha} \cdot N_2$$

whence:

$$\alpha = \frac{1}{2} \ln \frac{N_v}{N_2} \tag{8}$$

If the currents in the branches of the bridge are equal, we shall have the special relationship:

$$\frac{N_v}{N_2} = 4 \frac{z_{\text{arr}} \left(\frac{W_{1s}}{W_{1P}}\right)^2}{z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2}$$

Assuming $z_{\text{arr}} \left(\frac{W_{1s}}{W_{1P}}\right)^2 = z_2 \left(\frac{W_{3s}}{W_{3P}}\right)^2$, we shall obtain the remarkable value of $\frac{1}{2} \ln 4 = 0.7$ N for the bridge loss.

2. *Transmission path from the two-wire incoming circuit towards the outgoing four-wire circuit with electrical balancing in the direction incoming four-wire circuit to outgoing two-wire circuit*

If we use the bridge arrangement shown in Figure 1 in the direction from the two-wire circuit towards the four-wire one, we shall have to put the voltage source in the two-wire circuit, and this will have the impedance z_2. The four-wire side F_{arr} is an undesired, but unavoidable, consumer of power; it has an impedance z_{arr}. The four-wire side F_{dep} has an impedance z_{dep}, and carries the requisite power-supply current.

We can simplify things if we replace the impedance triangle with apices a, b, and c, by a star of equivalent impedance. If, now, we introduce the abbreviations below for the impedances in the three sides of the triangle:

$$\gamma_1 = r_0 + z_{\text{arr}} \left(\frac{W_{1s}}{W_{1P}}\right)^2, \quad \gamma_2 = \rho\,[r_x + N_x + \mathrm{j}\,\omega L_x], \quad \gamma_3 = [1 - \rho]\,[r_x + N_x + \mathrm{j}\,\omega L_x]$$

the following relations will hold for the equivalent impedances in the star:

$$\delta_1 = \frac{\gamma_1 \cdot \gamma_2}{\gamma_1 + \gamma_2 + \gamma_3}, \quad \delta_2 = \frac{\gamma_1 \cdot \gamma_3}{\gamma_1 + \gamma_2 + \gamma_3}, \quad \delta_3 = \frac{\gamma_2 \cdot \gamma_3}{\gamma_1 + \gamma_2 + \gamma_3}$$

Then we can transform the circuit diagram of Figure 1 into that shown in Figure 2.

It will be readily seen that a perfectly-balanced bridge arrangement for the channel from F_{arr} to F_2, that is to say, with the impedance values shown in the diagram, will be balanced in the direction F_2 to F_{dep} as well. Here again, no current flows through the diagonal of the bridge $c-d$. However, if we insert an additional impedance in any branch of the bridge, the balance of the bridge is disturbed, and the diagonal will be traversed by the current $i_3 - i_2$.

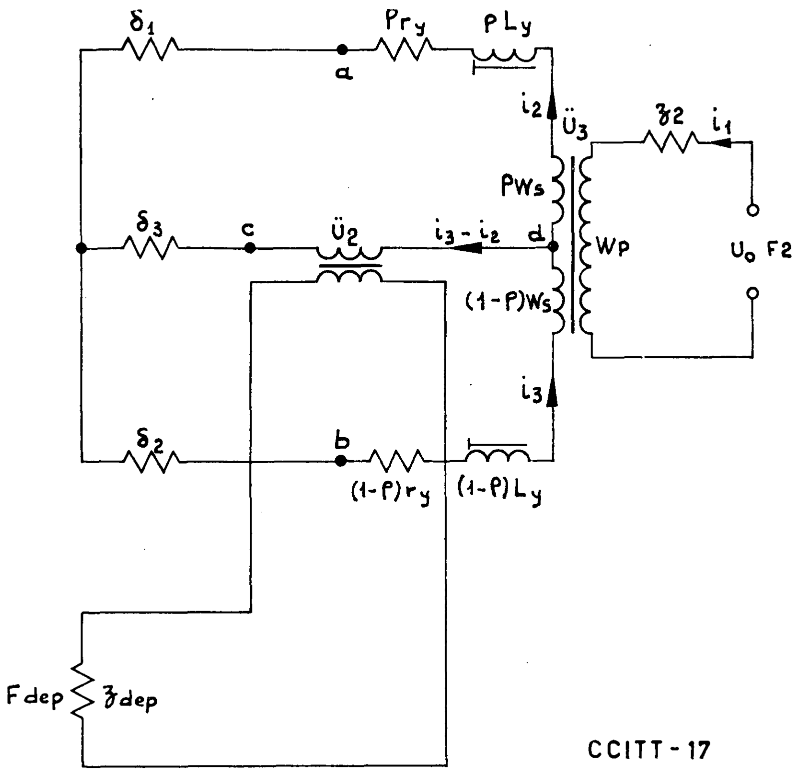


FIGURE 2. — **Equivalent diagram for bridge arrangement of Figure 1 for the path F_2 to F_{dep}**

3. *Transmission path in the direction two-wire circuit to four-wire circuit when the electrical balance is disturbed*

In currents flowing from a four-wire to a two-wire circuit, the electrical balance must be as perfect as possible, so that line stability is not interfered with and there is no possibility of singing.

In currents flowing from a two-wire to a four-wire circuit, on the other hand, there cannot be any balance, since the outgoing channel from the four-wire circuit receives its current through the diagonal of the bridge $c-d$. When the terminating set has to meet this requirement, unbalance must be produced whenever necessary by semi-conductor control devices operated by speech. These conditions are obtained in the simplest and most economical way by a diode circuit on the lines of Figure 3.

The positive half-waves of the sinusoidal currents from F_{arr} to F_2 pass through W_1, and arrive at the bridge arrangement after traversing Gr_1 and Gr_2. The negative half-waves reach the bridge through W_2, Gr_3 and Gr_4. In Figure 1, r_0 corresponds to the resistances W_1 and W_2. These resistances do not, properly speaking, form part of the bridge arrangement. The resistances in the direction of the current through Gr_1 and Gr_3 are similar to the resistance $\rho\, r_x$, and those in the direction of Gr_3 and Gr_4 to the resistance $\rho\, r_y$. They form part of the basic bridge arrangement and, to ensure symmetry, require additional adjusted balance resistances $(1-\rho)\, r_x$ and $(1-\rho)\, r_y$.

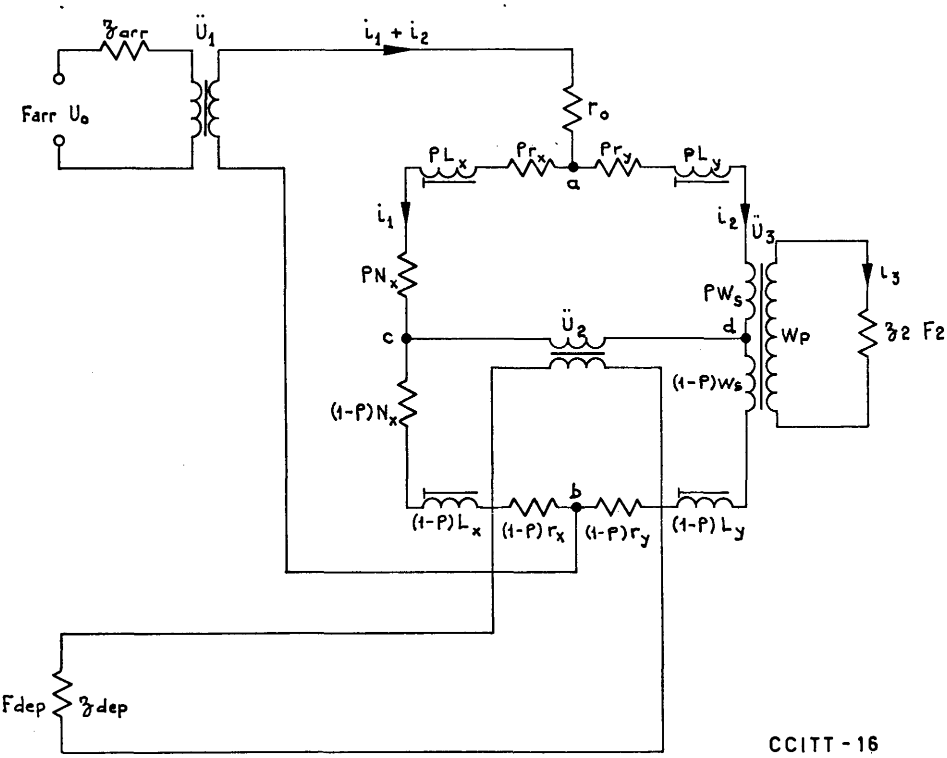


FIGURE 3. — **Diode circuit operating in conjunction with the balanced bridge arrangement**

When the diode circuit is used in the direction F_2 to F_{dep}, the position is entirely different. The positive half-waves of the sinusoidal currents pass through Gr_4, the two resistances in wires W_2 and W_1 and the diode Gr_1. The negative half-waves pass through Gr_3, the resistances in series W_2 and W_1 and Gr_2. The shunt created by the transformer U_1, and connected to the bridge points a' (displaced by W_2 in relation to point a) and b, causes an inevitable loss of power, which in turn means that the diodes Gr_4 and Gr_1, Gr_3 and Gr_2, respectively, cannot conduct currents that are exactly equal.

Now, contrary to the circuit conditions when transmission is from F_{arr} to F_2, the resistances W_1 and W_2 are in the perfectly balanced bridge arrangement and disturb the electrical balance so that the transformer U_2 receives current and the circuit z_{dep} consumes power. To compute the current $i_3 - i_2$, we should examine an equivalent circuit on the lines of Figure 2, replacing the resistance combinations δ_1, δ_2 and δ_3 by the values ε_1, ε_2 and ε_3.

The values ε_1, ε_2 and ε_3 are obtained as follows:

$$\varepsilon_1 = \frac{z_{arr}\left(\dfrac{W_{1s}}{W_{1P}}\right)^2 [\gamma_2 + W]}{z_{arr}\left(\dfrac{W_{1s}}{W_{1P}}\right)^2 + W + \gamma_2 + \gamma_3}$$

$$\varepsilon_2 = \frac{z_{\text{arr}} \left(\dfrac{W_{1s}}{W_{1P}}\right)^2 \gamma_3}{z_{\text{arr}} \left(\dfrac{W_{1s}}{W_{1P}}\right)^2 + W + \gamma_2 + \gamma_3}$$

$$\varepsilon_3 = \frac{\gamma_3 \left[\gamma_2 + W\right]}{z_{\text{arr}} \left(\dfrac{W_{1s}}{W_{1P}}\right)^2 + W + \gamma_2 + \gamma_3}$$

γ_2 and γ_3 have the meanings indicated in section 2). In accordance with the actual setting-up of such arrangement, it was assumed that $W_1 = W_2 = W$. In the calculation, account must be taken of the fact that the resistance W appears also in the branch between the connection points a' and d, so that the load in this branch is given by

$$W + \rho \left[r_y + \mathrm{j}\, \omega\, L_y + \mathrm{j}\, \omega\, k\, \rho\, W_{3s}^2\right]$$

For these currents and using the symbols of Figure 2 we find, therefore:

$$i_1 = \frac{U_0}{z_2 + R \left(\dfrac{W_{3P}}{W_{3s}}\right)^2} \tag{9}$$

$$i_2 - i_3 = U_0 \left(\frac{W_{3s}}{W_{3P}}\right) \frac{1}{z_2 \left(\dfrac{W_{3s}}{W_{3P}}\right)^2 + R} \times$$

$$\times \frac{\rho\, \varepsilon_2 - (1-\rho)\left[W + \varepsilon_1\right]}{[1-\rho]^2 \left[W + \varepsilon_1\right] + \rho^2 \varepsilon_2 + \rho\,[1-\rho]\,[r_y + \mathrm{j}\,\omega\, L_y] + \varepsilon_3 + z_{\text{dep}} \left(\dfrac{W_{2s}}{W_{2P}}\right)^2} \tag{10}$$

$$R = \frac{[W + \rho\,(r_y + \mathrm{j}\,\omega\, L_y) + \varepsilon_1]\,[(1-\rho)\,(r_y + \mathrm{j}\,\omega\, L_y) + \varepsilon_2] + [W + r_y + \mathrm{j}\,\omega\, L_y + \varepsilon_1 + \varepsilon_2]\left[\varepsilon_3 + z_{\text{dep}} \left(\dfrac{W_{2s}}{W_{2P}}\right)^2\right]}{[1-\rho]^2\,[W + \varepsilon_1] + \rho^2\, \varepsilon_2 + \rho\,[1-\rho]\,[r_y + \mathrm{j}\,\omega\, L_y] + \varepsilon_3 + z_{\text{dep}} \left(\dfrac{W_{2s}}{W_{2P}}\right)^2} \tag{11}$$

after agreeing to neglect certain quantities since $z_2 \ll \mathrm{j}\, \omega\, k\, W_{3P}^2$.

The denominator of the equation (9) is the characteristic equation for the input impedance of the arrangement. For perfect matching, therefore, we should find that:

$$R \left(\frac{W_{3P}}{W_{3s}}\right)^2 = z_2 \tag{12}$$

The useful power can be calculated from the following:

$$N_2 = [i_2 - i_3]^2\, z_{\text{dep}} \left(\frac{W_{2s}}{W_{2P}}\right)^2 \tag{13}$$

The loss of the terminating set can then be found from equation (8).

According to equation (10), $i_2 - i_3$ will also become zero when $W = 0$. The loss of the terminating set will be infinitely great because this assumption implies the condition required for absolute electrical balance of the bridge arrangement. The latter is no longer maintained when W is inserted; the loss of the terminating set steadily decreases therefore as W increases.

In the transmission direction F_{arr} to F_2, the values governing the terminating set loss show an inverse trend. When $W = 0$, the voltage in the bridge will be at its highest, so that the terminating set loss will reach its minimum according to the above-mentioned assumption. When W increases, the bridge voltage diminishes and the terminating set loss becomes greater. For certain bridge impedances there will thus be a well-defined value of W, at which the terminating set loss is the same in both directions of transmission.

If ever this value were too high and the bridge resistances could not be reduced (for example, because the diodes used prevented it), it would still be possible to obtain a lower figure for the terminating set loss. Actually it is not very difficult to reduce the loss in the F_{arr} to F_2 direction of transmission. The greater loss in the opposite direction F_2 to F_{dep} should therefore be reduced by inserting an amplifier producing unilateral gain in F_{dep}, which would give a lower figure in the direction F_{arr} to F_2.

4. *The terminating set used by the Austrian Posts and Telegraph Administration*

Figure 4 shows the terminating set used by the Austrian P. and T. Administration, which gives improved balance return loss.

The figure shows, first of all, that the conditions required for absolute electrical balance in the bridge, according to equation (2), are not fulfilled. This is due to the efforts made to reduce the terminating set loss, which must be the same in both directions of transmission, to between 0.9 and 1.05 N. It is not possible to obtain this figure when there is absolute electrical balance in the direction F_{arr} to F_2. For reasons of economy, therefore, a slight unbalance has been created to achieve a reduction in loss. Provided the outputs F_2 and z_{dep} are terminated by resistances of 600 ohms, the loss F_{arr} to F_{dep} reaches 6 N and above at the terminals F_{dep} when the voltage level is zero N. When the voltage level is -2 N, this figure only drops to 5.5 N. If the end F_2 is not terminated at all, there will be minimum figures of 3.5 N or 3.0 N. In the latter case, the loss F_{arr} to F_{dep} is 3.5 times greater than the loss F_{arr} to F_2. With the conventional terminating sets with differential transformers, we get a factor of only 2.0. A figure of about 1.0 N is not by any means extraordinarily high for the loss F_{arr} to F_2 of a terminating set, if we bear in mind that terminating sets of the conventional type have figures of 0.8 N, including the pad which is always provided. In this fashion, the excess loss of the type Gr terminating set amounts to only 0.2 N, and allowance can readily be made for it in the circuit arrangement of the final amplifiers.

The diodes used are of the OA5 type. These call for fairly high operating voltages. Hence a ratio of 1 : 10 was chosen for the windings of the three transformers, so as to obtain impedance ratios of 1 : 100. The low useful powers, of the order of 10^{-3} watts mean that the operating currents amount to about 10^{-6} A, so that the passband resistances of the diodes are 10 000 ohms and more. For all this, the distortion does not exceed 2% for the least favourable direction of transmission, from F_2 to F_{dep}, and a level of 0 to -3 N, and is no more than 1% in the direction $F_{arr}-F_2$. It would be desirable to have diodes with a better dynamic characteristic, but they are not available. For low input levels, the balance return loss of the bridge arrangement in the direction F_{arr} to F_{dep} falls off slowly. For an input level of -2.0 N, we get the figure of 3 N, already quoted. The four-wire circuit still does not sing, because, with low levels, the tendency of the singing frequencies to reach higher amplitudes is restrained by a simultaneous increase in the balance return loss.

A few words will suffice to demonstrate the great advantages of the new arrangement and the stability it affords, as far as transmission and switching are concerned.

Four-wire circuits terminated at both ends by Gr terminating sets can be adjusted to an equivalent of 0 N. With open inputs and outputs of the two-wire circuits, the stability will attain the

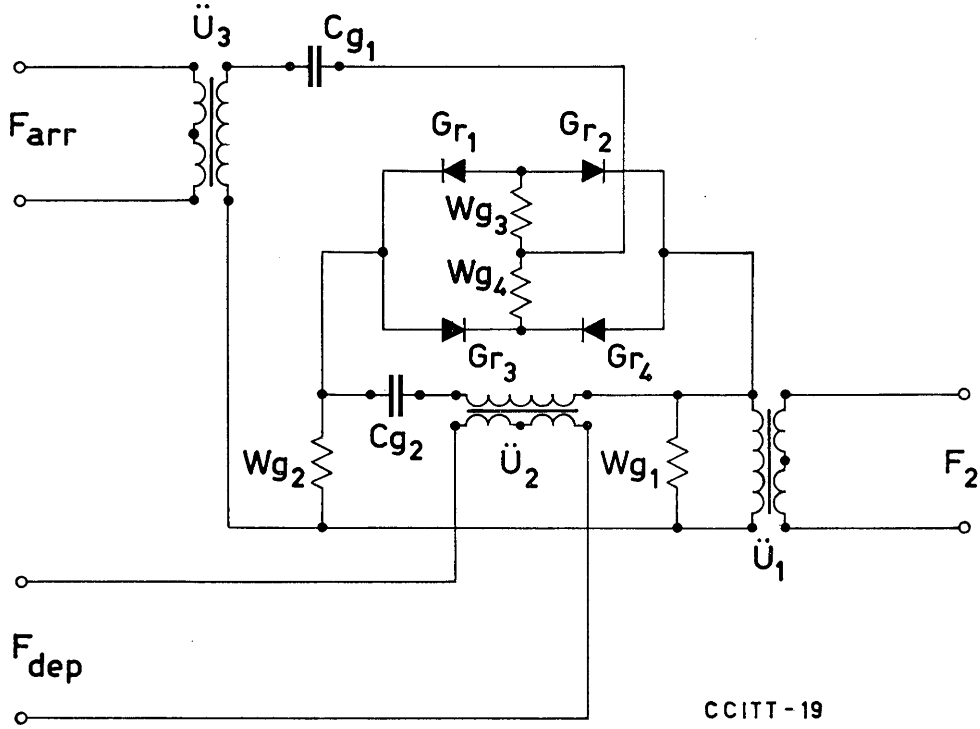


FIGURE 4. — **Circuit diagram of the terminating set giving improved balance return loss (type Gr)**

considerable figure of $3.5 - 2 \times 1.05 = 1.40$ N. This figure will suffice to absorb the reduction in loss occurring with some likelihood in a chain of 11 or 12 four-wire circuits, each of which is independently maintained.

The reduction of the normal figure of 0 N for the equivalent of the four-wire circuits makes for economy in laying underground cables linking group centres having four-wire terminating sets, to their terminal exchanges. Two-wire circuits on 0.9-mm pairs are always used exclusively between these two kinds of exchange. The maximum attenuation (0.4 N) provided for such circuits will be reached for a distance of 17 km. In this fashion, longer circuits must be equipped with intermediate repeaters, for which transistorized amplifiers for loaded circuits will be used. One such amplifier per circuit will be enough no matter how long the line may be. The outlay for such an amplifier is less than the additional cost for a pair of 1.4-mm wires.

The resistance Wg_2 maintains the electrical balance for the possible load impedances (z_2). This is why circuits with different loads can be incorporated in the same route, and the terminating sets definitely associated with the outputs of the four-wire circuits. This permits two-wire switching between four-wire circuits and two-wire ones. Thus it is that we can drop the procedure hitherto followed, i.e., that a special group of terminating sets should be reserved for each electrically uniform group, in order to make it easier to obtain perfect balance networks.

Of 4900 terminal equipments, 39% are of the Gr type and have already proved their worth. They need hardly any supervision and greatly facilitate the work of supervision and maintenance, as the balance networks do not need to be checked.

If the terminating set be traversed by currents of the two directions of transmission, there will be trouble, and mutilated signals. The direction of the current in the diode bridge is no longer clear: it depends on the momentary distribution of voltage peaks in the arrangement. This peculiarity is no handicap for circuits designed for semi-duplex working, such as telephone circuits. However, real duplex transmissions, such as data transmissions, comprising backward signals transmitted from the receiver to the transmitter, cannot be routed over four-wire circuits with terminating sets of the Gr type.

5. *Conclusion*

The circuit arrangement we use in the Austrian Administration does not, by any means, exhaust all the possibilities to be derived from use of non-linear components in four-wire terminating sets. There is the necessity to take cost into account and this inevitably leads to solutions involving varying degrees of compromise.

However, the new arrangement rigorously observes the physical conditions to be rigorously respected in the case of terminating sets providing improved balance return loss. For the direction incoming four-wire circuit → outgoing two-wire circuit, the electrical balance of the bridge arrangement, referring to the outgoing side of the four-wire circuit, must be rendered as high as possible. This balance must be destroyed for transmission of currents in the other direction by use of non-linear speech-operated components, in such a fashion as to enable the currents to flow in the required direction. Such arrangements cannot be used for duplex working. However, it is urged that they be used on a considerable scale in national telephone service, especially with fully automatic operation. Lastly, it may be affirmed that a full study of this kind of arrangement, with a view to its improvement, is desirable, and even entirely justified.

ANNEX 7

(to Question 19/XII)

Functioning of a terminating sewt ith improved attenuation characteristics

(Contribution by the Netherlands Administration)

1. *Introduction*

The most striking difference with the four-wire terminations used until now is the application of non-linear elements (diodes) in this termination. The object aimed at by using these elements was to obtain a large overflow attenuation, independent of the two-wire impedance.

In this annex the working of the Austrian four-wire termination described in Annex 6 will be discussed. It will appear that under special circumstances this termination has favourable properties regarding the solution of the stability problem on amplified connections. If signals are transmitted both on the four-wire and on the two-wire sides, the termination begins to show undesirable effects: the transmission two-wire side → four-wire side will be strongly distorted. This is the reason why this termination is altogether unsuitable for use in networks in which, e.g., MFC signalling is or will be applied.

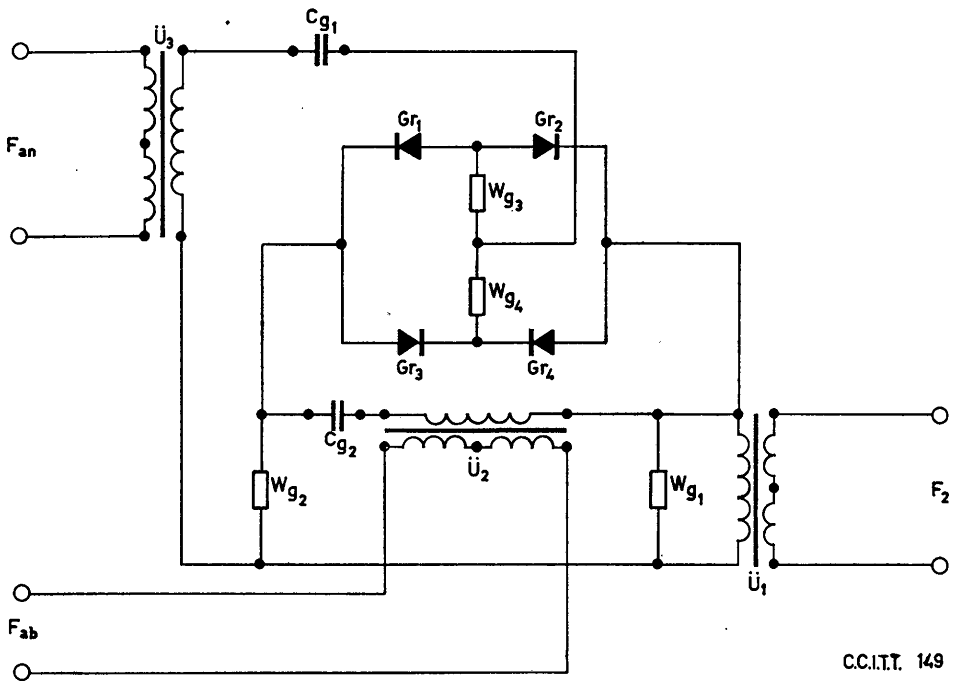


FIGURE 1. — **Switching diagram of a terminating set using diodes**

Wg_1 = 390 kohms
Wg_2 = 47 kohms
Wg_3 = 33 kohms
Wg_4 = 33 kohms
Cg_1 = 47 nF
Cg_2 = 47 nF
Gr_1 = Gr_2 = Gr_3 = Gr_4 = diode OA5

Transformer: turn ratios

U_1	1000 windings, side F_2	
	9500 windings, side Wg_1	
U_2	1000 windings, side F_{ab}	
	9000 windings, side Cg_2	
U_3	1000 windings, side F_{an}	
	10500 windings, side Cg_1	

Insertion loss

$F_{an} \to F_2$ and $F_2 \to F_{ab}$
for 800 c/s ... 0.9 ... 1.05 N
Tolerable deviation for 300 and 3400 c/s: ... ± 0.05 N
Input level 0 N to −2 N

Mismatch based on 600 ohms

Measuring frequency	F_{an}	F_2	F_{ab}
300 c/s	⩾ 2.3 N	⩾ 1.6 N	⩾ 2.3 N
800 c/s	⩾ 2.3 N	⩾ 2.3 N	⩾ 2.3 N
3400 c/s	⩾ 2.3 N	⩾ 2.3 N	⩾ 2.2 N

Return loss

$F_{an} \to F_{ab}$, F_2 terminated with 600 ohms:
for input level 0 N ⩾ 6 N
for input level −2 N ⩾ 5.5 N
F_2 not terminated:
for input level 0 N ⩾ 3.5 N
for input level −2 N ⩾ 3.0 N

Distortion

$F_{an} \to F_2$ ⩽ 1% test level 0, −1, −2 N
$F_2 \to F_{ab}$ ⩽ 2% test level 0, −1, −2, −3 N

The Netherlands Administration has had some of these four-wire terminations constructed according to data received from the Austrian P.T.T. (see Fig. 1). The measuring data have been mentioned and worked up in this annex.

2. *The termination*

Figure 1 shows the diagram of the termination with some characteristic quantities. F_{an} is the four-wire receiving side, F_{ab} the four-wire transmission side, and F_2 the two-wire side. The three transformers U_1, U_2, and U_3 effect the adjustment of the line impedances (600 ohms) to the termination. The resistors Wg_3 and Wg_4 have the same values. The type of diode used is the OA5 one.

3. *Measurements*

The measured results of the insertion loss and the overflow attenuation obtained by means of the laboratory model are shown in Figures 2, 3, and 4. Figure 2 shows the curve of the insertion loss $F_{an} \rightarrow F_2$ as a function of the frequency with the signal level as parameter. It appears that in this case the loss is practically independent of the level. Figure 3 shows a similar curve, now for the transmission direction $F_2 \rightarrow F_{ab}$. The dependence of the level comes more strongly to the fore here. This is the result of the fact that the resistance values of Wg_3 and Wg_4 are smaller than those of Wg_2 and W_2, which play a part in the transmission $F_{an} \rightarrow F_2$ (Fig. 5); so the resistance variation of the diodes is expressed relatively more in the case $F_2 \rightarrow F_{ab}$.

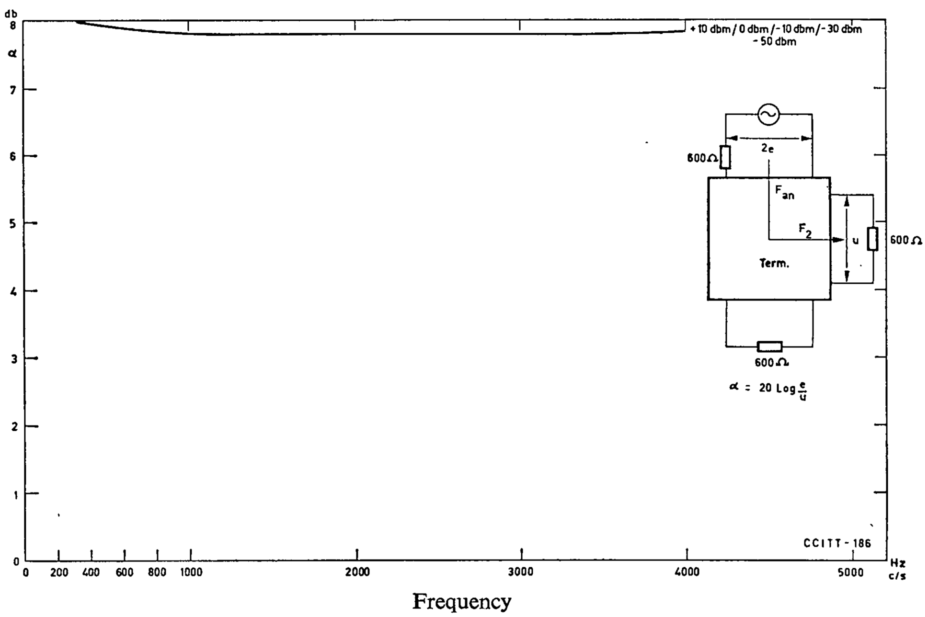
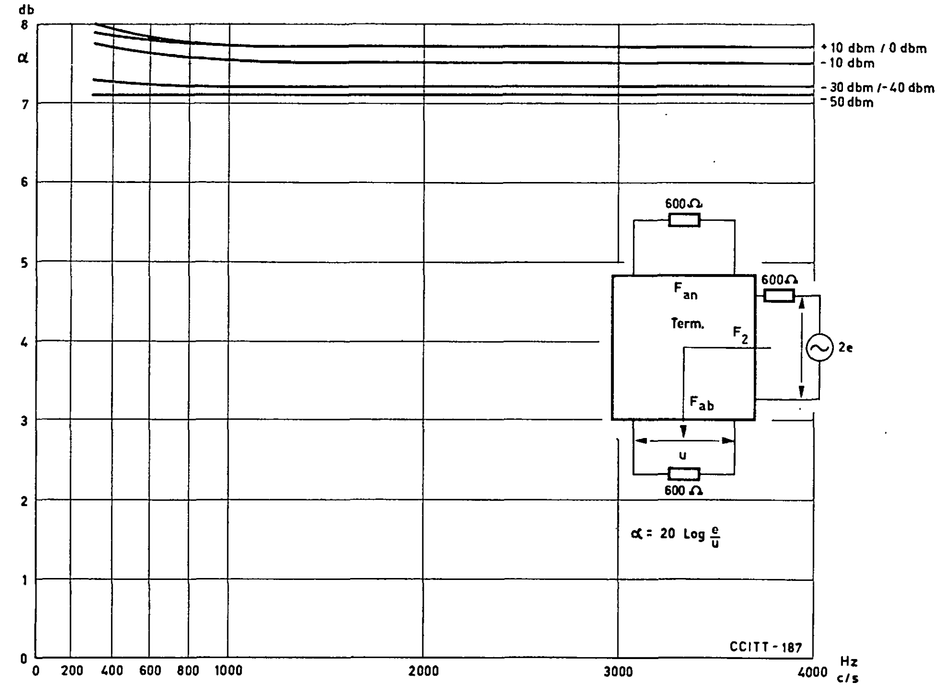


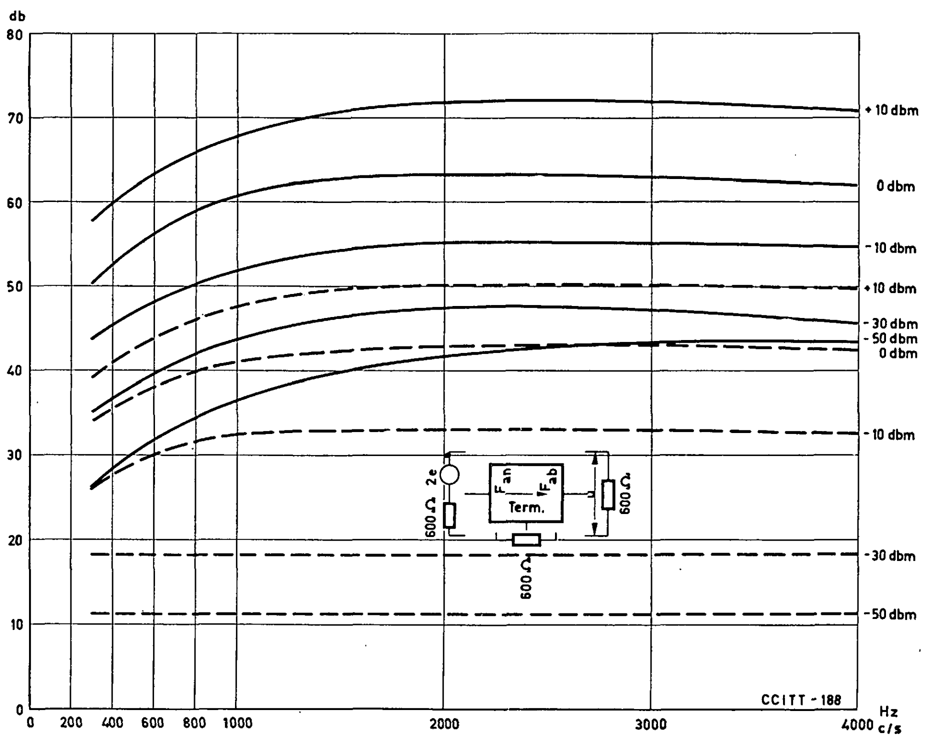
FIGURE 2. — **Insertion loss as a function of level (F_{an}) and frequency**

The overflow attenuation (Fig. 4) has been measured selectively. As a matter of fact, it appeared that the voltage shape measured on F_{ab} was peaked (Fig. 6). These voltage peaks result from the not simultaneously blocking of the diodes. During the transition from the positive half to the negative half of the period, the diodes Gr_1 and Gr_2 close and the diodes Gr_3 and Gr_4 open. If e.g. diode Gr_1 now closes a little earlier than Gr_2, the bridge equilibrium will be disturbed for a very short time, the result of this being a peak in the voltage over W_{an} (see Fig. 5).



Frequency

FIGURE 3. — **Insertion loss as a function of level (F_2) and frequency**



Frequency

———— F_2 with 600 Ω — — — — F_2 without 600 Ω

FIGURE 4. — **Return loss $F_{an} \rightarrow F_{ab}$ (Fundamental frequency)**

The time difference is caused by the inequality of the RC times of the diodes in combination with Wg_2 and W_2 (Fig. 5). If F_2 is open, $W_2 = Wg_1 \neq Wg_2$.

In Table 1 some characteristic values have been so arranged as to provide an easy survey. The values given hold for a frequency of 1 kc/s.

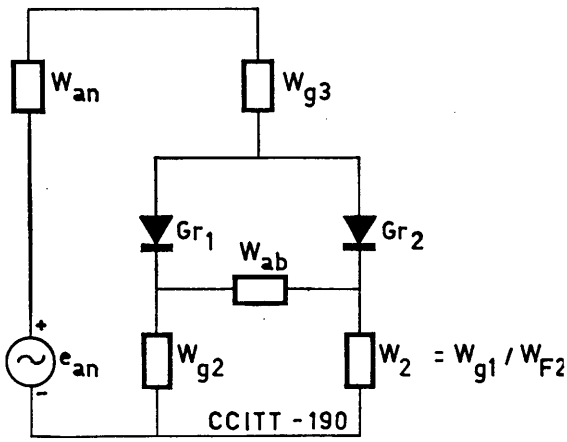


FIGURE 5

TABLE 1

Level	Insertion loss $Fan \rightarrow F_2$	Insertion loss $F_2 \rightarrow Fab$	Overflow attenuation in db	
dbm	db	db	600 ohms termination	open F_2
+10	7.8	7.1	68	47
−50	7.8	7.7	36	11

4. *Comparison of linear four-wire termination and the Austrian four-wire termination with respect to transmission, amplification, and stability of the four-wire circuit*

In Figure 7 a four-wire circuit is represented diagrammatically. As it is considered desirable that in the Netherlands telephone network the difference of level between the two-wire sides F_2 should be 0 db, and taking into account the values for the insertion loss α_1 and α_2, the amplification (G) of the amplifiers has been fixed as:

$$G = \alpha_1 + \alpha_2 \tag{1}$$

For the linear four-wire termination we have: $\alpha_1 = \alpha_2 = 3.5$ db, so that $G = 7$ db.

With the Austrian four-wire termination these values are more or less dependent on the level. From the measurements it appears that:

sending −50 dbm	*sending +10 dbm*
$\alpha_1 =$ 7.7 db	$\alpha_1 =$ 7.1 db
$\alpha_2 =$ 7.8 db	$\alpha_2 =$ 7.8 db
$\alpha_1 + \alpha_2 =$ 15.5 db	$\alpha_1 + \alpha_2 =$ 14.9 db

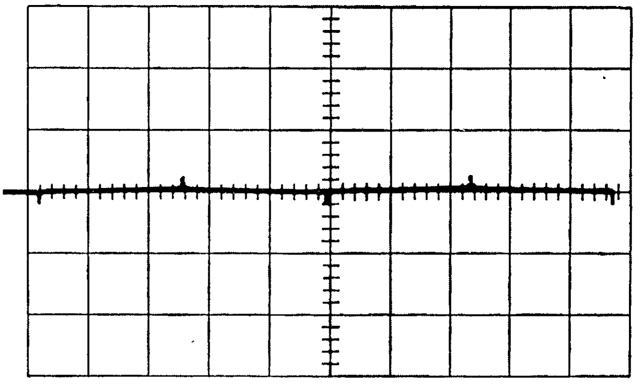


Photo 1. — F_2 terminated with 600 ohms

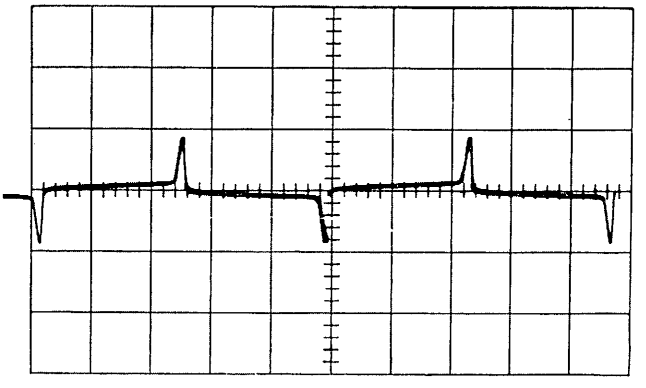


Photo 2. — F_2 not terminated

Vertical scale: 1 cm = 0.05 V Frequency = 1000 c/s

FIGURE 6. — **Return signal $F_{an} \rightarrow F_{ab}$. Level F_{an}: −15 dbm**

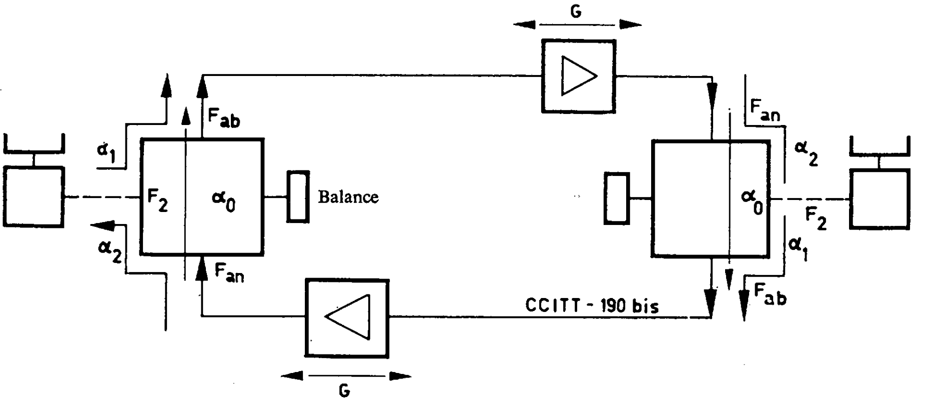


FIGURE 7

In this case the amplification will have to be about 15 db. Therefore a greater amplification will be required if the Austrian four-wire termination is used. Moreover, theoretically the level difference of 0 db between the two-wire sides cannot be obtained for all transmission levels. Because termination overflow ($F_{an} \rightarrow F_{ab}$) causes stability problems in a four-wire circuit, it is possible that oscillations will be started. Theoretically oscillations will be started if:

$$2\,G - 2\,\alpha_0 > 0 \tag{2}$$

In the most unfavourable case $\alpha_0 = 6$ db with the linear four-wire termination; oscillations will then be started if no special provisions are made.

With the Austrian four-wire termination the situation is somewhat different. Here α_0 is not 0 db in the most unfavourable case. From the measurements it has appeared that with a level of -50 dbm and open F_2 the overflow attenuation is 11 db. It is true that oscillations will occur, but these oscillations will stay at a relatively low level (low with respect to the oscillation level if linear four-wire terminations are used). As a matter of fact the overflow attenuation increases as the level rises.

To get an impression of the oscillation level a measurement has been made in which this level is determined as a function of the amplification (two-wire → two-wire). Figure 8 shows this relation.

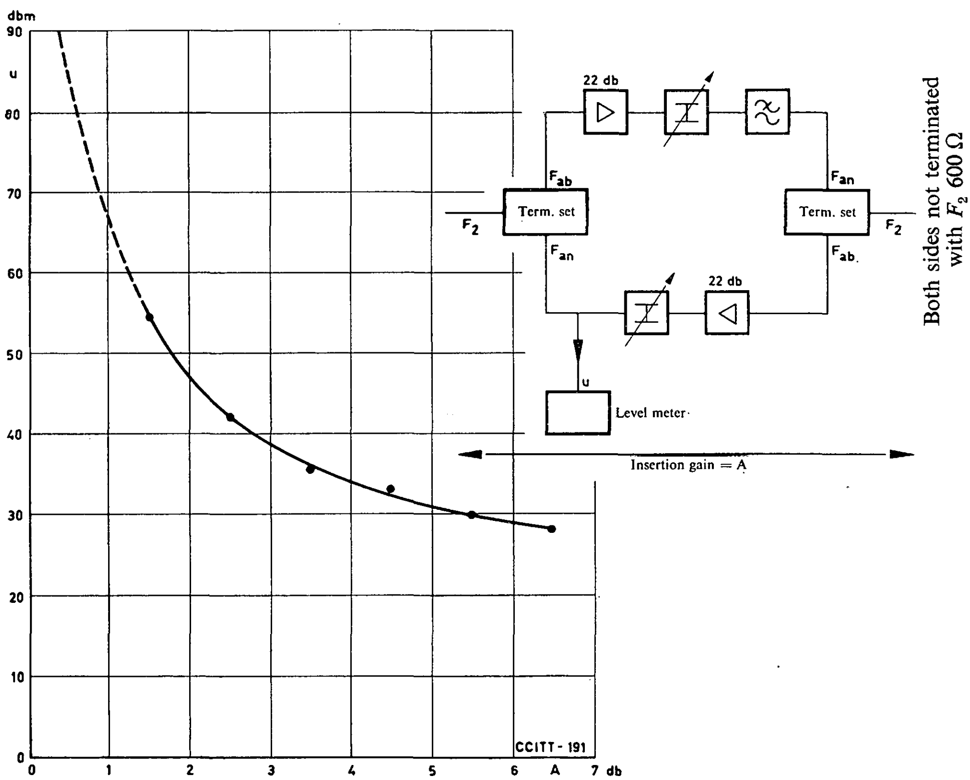
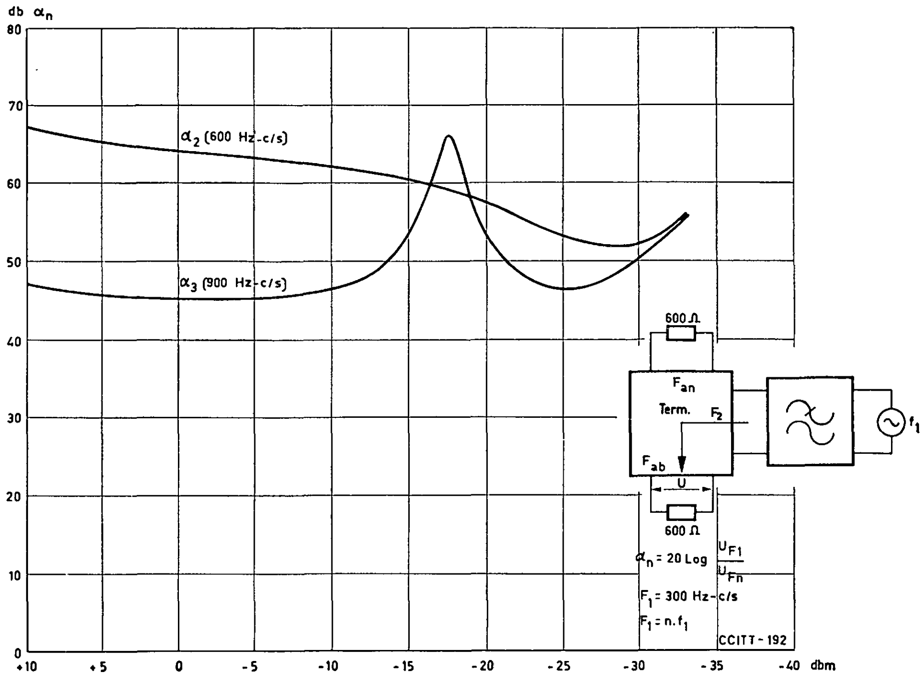


FIGURE 8. — **Singing level as a function of the insertion gain between the two-wire points F_2**

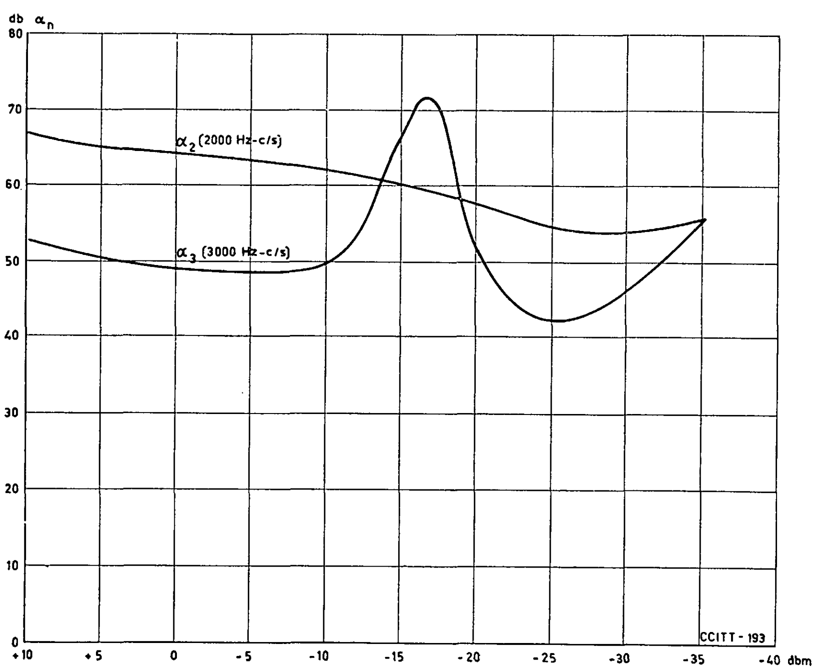
5. *Distortion and intermodulation*

Graphs 9 and 10 show the curves of the 2nd and 3rd harmonics as a function of the transmission level (transmission direction $F_2 \rightarrow F_{ab}$) for the frequencies 300 and 1000 c/s respectively. Similar measurements have been made for the transmission direction $F_{an} \rightarrow F_2$ (Figures 11 and 12).



Level F_2

FIGURE 9. — **Distortion $F_2 \rightarrow F_{ab}$ (300 c/s)**



See Fig. 9

Level F_2

FIGURE 10. — **Distortion $F_2 \rightarrow F_{ab}$ (1000 c/s)**

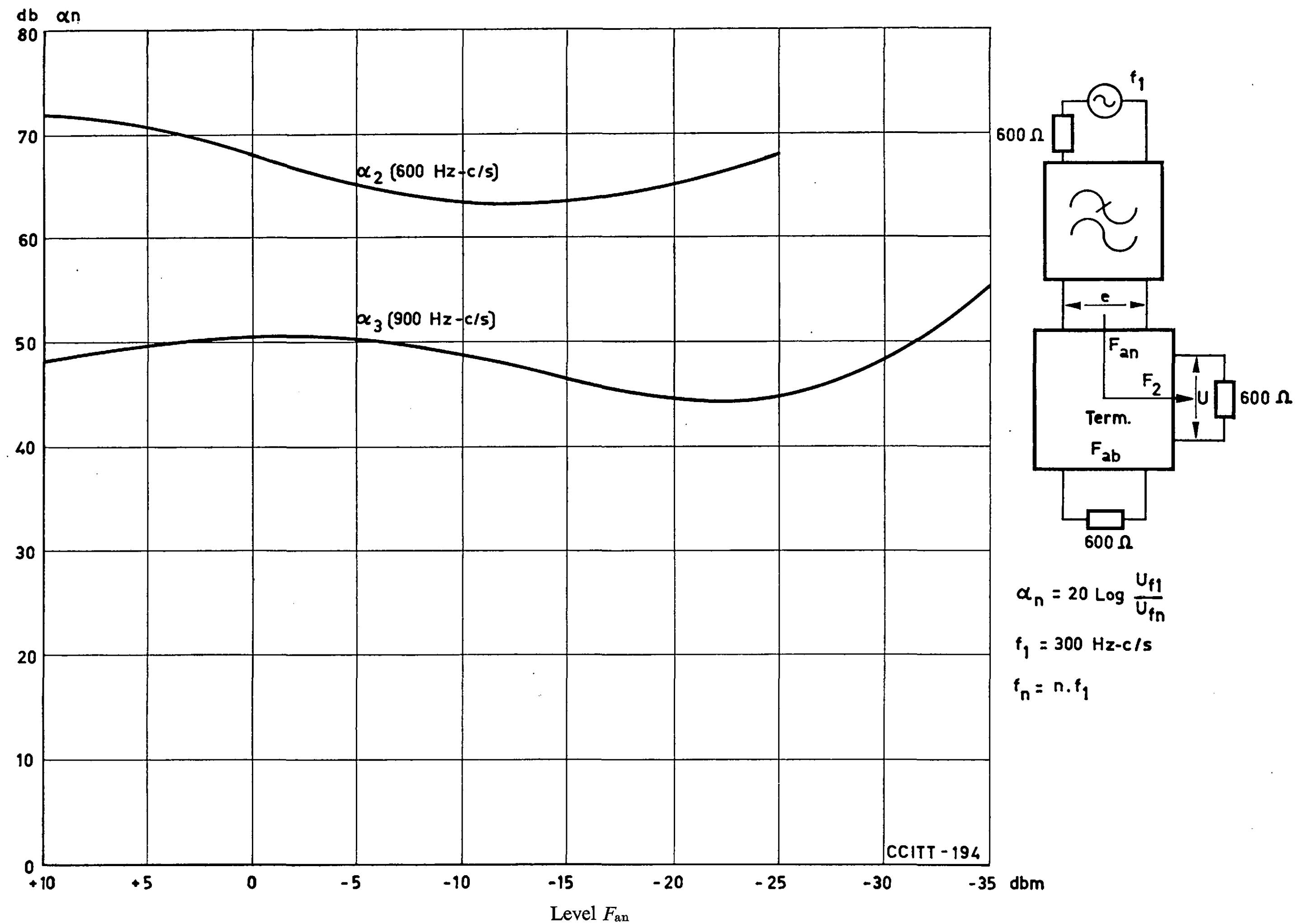
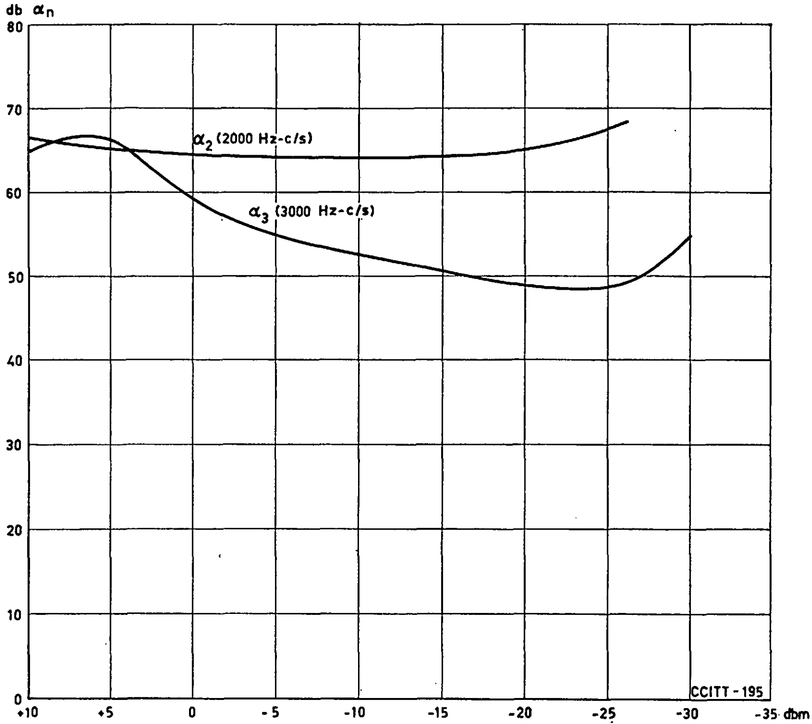


FIGURE 11. — **Distortion $F_{an} \rightarrow F_2$ (300 c/s)**

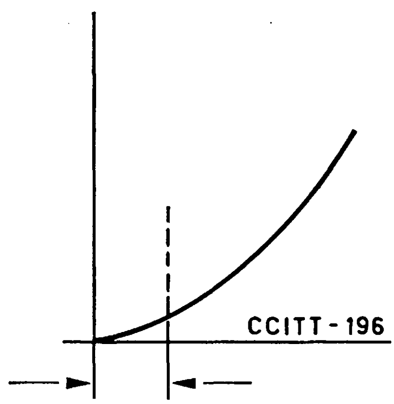


Level F_{an}

See Fig. 11

FIGURE 12. — **Distortion $F_{an} \rightarrow F_2$ (1000 c/s)**

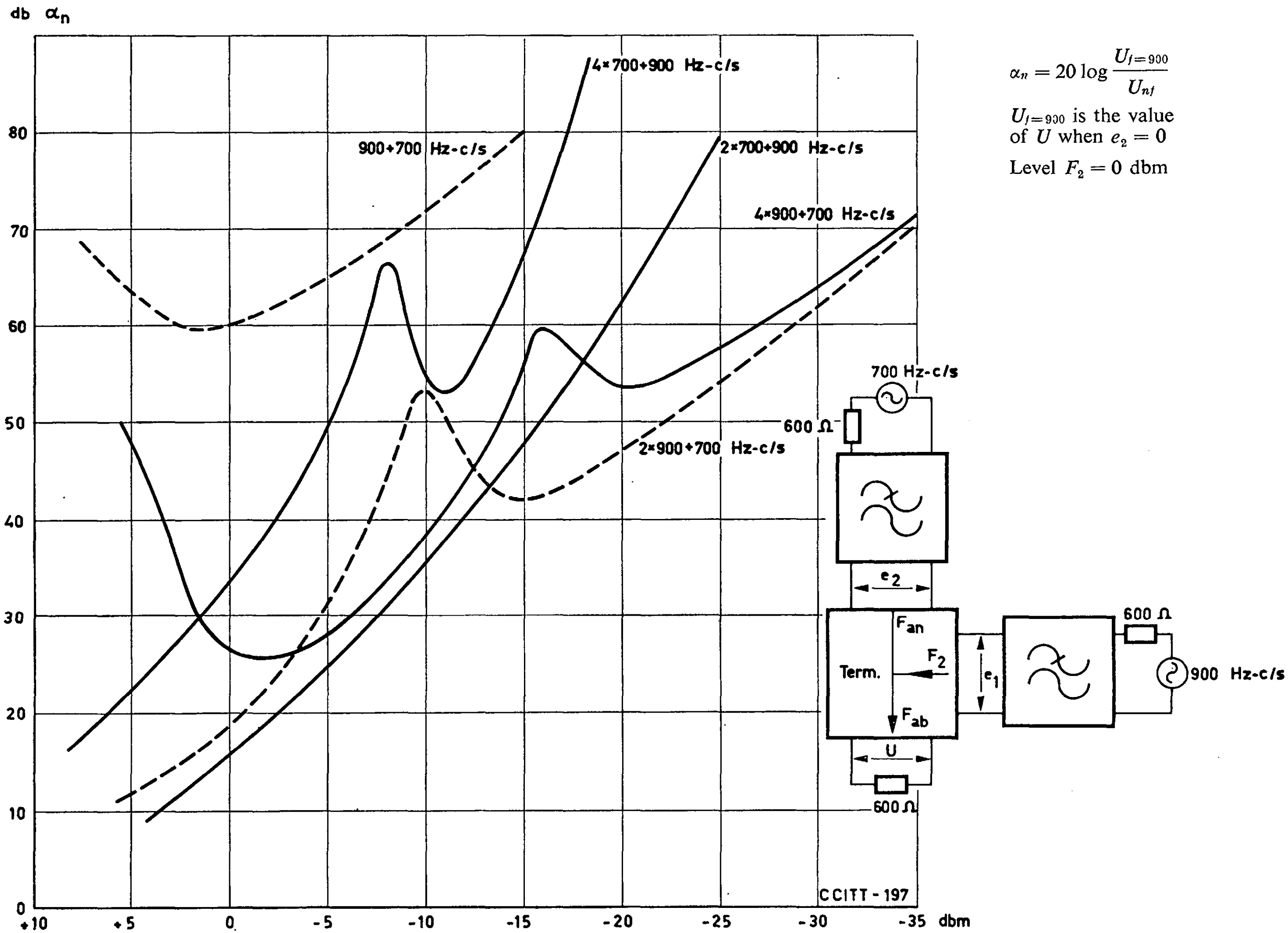
The symmetry in the diode arrangement explains the fact that the odd harmonics (3rd harmonic) are more in evidence. For lower levels the 2nd and 3rd harmonics become smaller and smaller; however, the curves rise again with levels lower than -30 dbm. This is the result of the fact that with low levels the measurements are made in a practically linear region of the diode characteristic (Fig. 13).



Linear region

FIGURE 13

The results of the intermodulation measurement are shown in Figure 14.

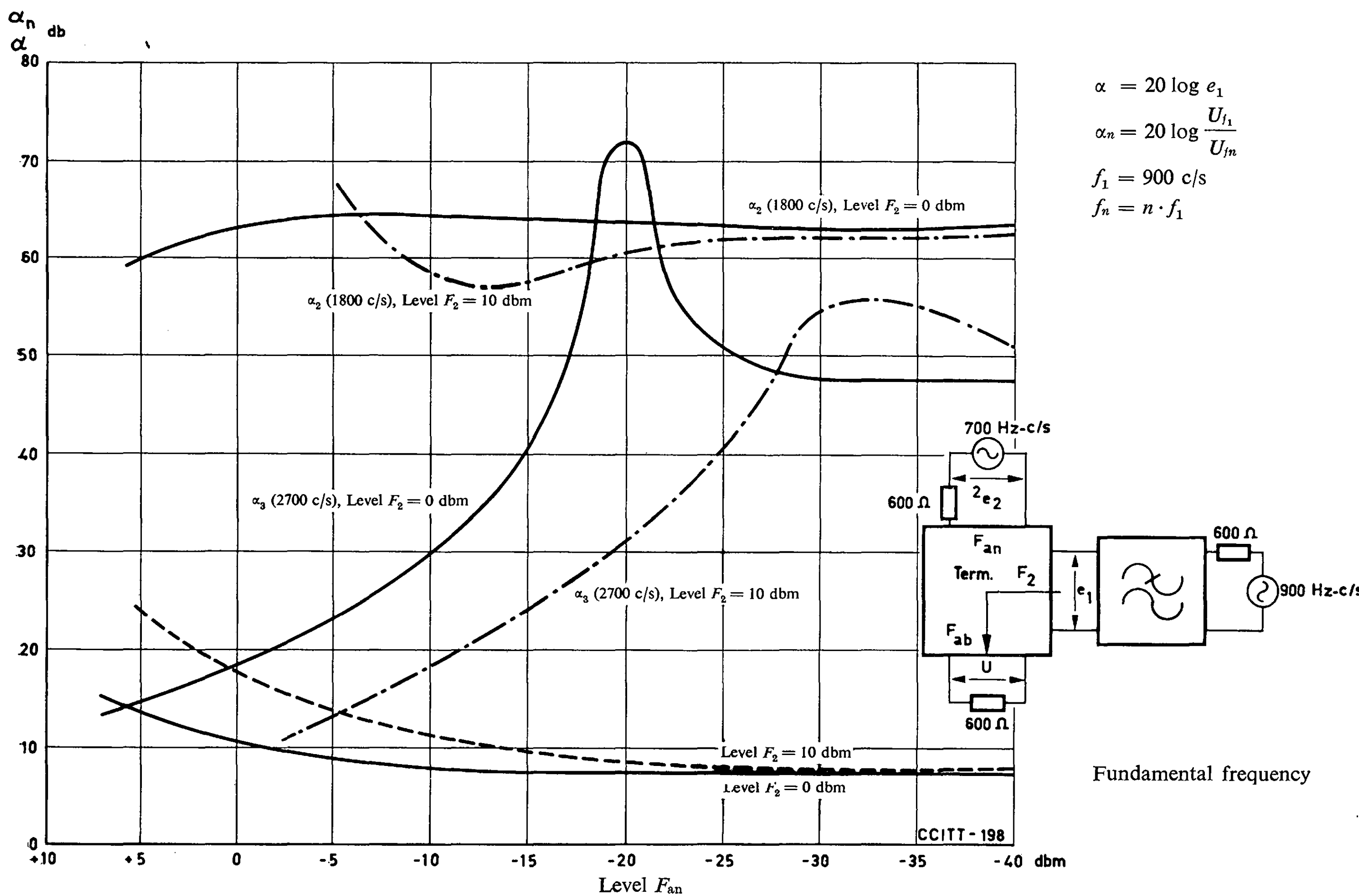


$$\alpha_n = 20 \log \frac{U_{f=900}}{U_{nf}}$$

$U_{f=900}$ is the value of U when $e_2 = 0$

Level $F_2 = 0$ dbm

FIGURE 14. — **Intermodulation (F_2 = 900 c/s, F_{an} = 700 c/s)**



$$\alpha = 20 \log e_1$$

$$\alpha_n = 20 \log \frac{U_{f_1}}{U_{fn}}$$

$$f_1 = 900 \text{ c/s}$$

$$f_n = n \cdot f_1$$

FIGURE 15. — **Insertion loss and distortion $F_2 \rightarrow F_{ab}$ as a function of the F_{an} level**

6. *Properties and working of the Austrian four-wire termination if signals are applied on two sides*

It is clear that because of the presence of non-linear elements in the termination the superposition principle does not apply. Therefore the output signal at the four-wire side cannot be obtained by adding up the voltages that are generated if transmission is effected at the two-wire side or four-wire side respectively.

This fact is confirmed by the photos of Figure 6 and the graphs above. From them it appears clearly that this four-wire termination is unsuitable if MFC signalling is used or when operating on communication channels having a poor signal-to-noise ratio (long-distance circuits, radio circuits, room noise etc.). The signalling coming from F_2 is affected by the signalling at F_{an}. In certain cases the output signal at F_{ab} is entirely distorted. This distortion is inherent in the arrangement with diodes and does not occur at all with linear four-wire terminations.

A quantitative impression of the phenomenon can to some extent be obtained from Figure 15.

Question 20/XII — Synthetic speech and frequency compression systems

(continuation of Question 20 of Study Group XII, 1961-1964)

a) What quality can at present be obtained from synthetic speech?

Note. — Synthetic speech is defined as a group of currents similar to voice currents produced by a system of generators, themselves controlled by signals transmitted according to a suitable code.

b) What is the quality of speech transmitted over a channel whose frequency band is compressed?

Note. — Administrations are asked to give a description of the operating principles of the system whose quality has been studied.

c) Under what circumstances could systems of the types studied under a) and b) be used in the international telephone network in the near future?

Note. — Attention is drawn to the following article in particular: E. E. David, Jr., H. R. Schroeder, B. F. Logan, and A. J. Prestigiacomo: Voice-excited vocoders for practical speech bandwidth reduction, published in *I.R.E. Transactions on Information Theory*, Vol. IT-8, pp. 101-105, Sept. 1962.

Question 21/XII — Systems using pulses or alternating currents modulated by the microphone

(continuation of Question 21 of Study Group XII, 1961-1964)

What quality of speech transmission may be obtained at present by systems using pulses or alternating currents modulated by the microphone?

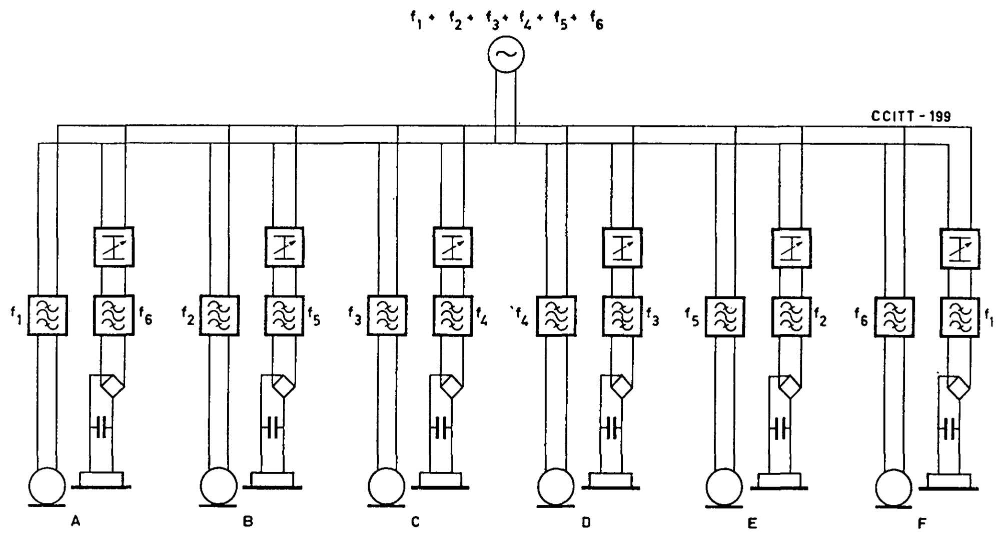
Note. — Preliminary experimental results submitted by the Italian Administration during the Geneva meeting of Study Group XII (June 1962) would seem to indicate that the differential resistance of a carbon microphone, which is determined for direct current, can be used for calculating the effects of modulation of alternating current by means of the same microphone. All these tests are described in the Annex below.

ANNEX

(to Question 21/XII)

Contribution by the Italian Telecommunication Administration

A test was made with a telephone system using carrier current instead of direct current to supply the carbon microphone. The circuit used is reproduced in Figure 1.



$f_1 = 21$ kc/s $f_2 = 35$ kc/s $f_3 = 49$ kc/s $f_4 = 63$ kc/s $f_5 = 77$ kc/s $f_6 = 91$ kc/s

FIGURE 1

This diagram shows six telephone sets, each of which has a receiving and a transmitting frequency. With this arrangement, three simultaneous connections are possible, namely, the circuits between the equipments A and F, B and E, and C and D.

There would, of course, be no difficulty in making a call between any two sets, provided each could use for transmission all the frequencies appropriate to the other sets at the receiving end.

Each sending set consists of a carbon microphone with a band filter and each receiving set of a bandpass filter, a rectifier, a condenser (acting as a high-frequency filter) and a telephone receiver. On the line connecting the six sets, a generator provides the frequencies required for supplying the microphones.

To vary the reference equivalent of each circuit, an adjustable attenuator controlling the receiving sensitivity of each set was provided; this attenuator had no effect on the value of the carrier current supplying the microphones. The tests were conducted with two different voltages for the carrier currents, namely 3 and 0.5 effective line volts.

In double-band transmission, a bandwidth of 6800 c/s should be used for each circuit. In these tests, the filters were of the kind shown in Figure 2, and the carrier currents were so selected that, for every out-of-band frequency, the attenuation was at least 30 db. Thus the values 21, 35, 49, 63, 77 and 91 kc/s, i.e. odd harmonics of 7 kc/s, were obtained for these frequencies.

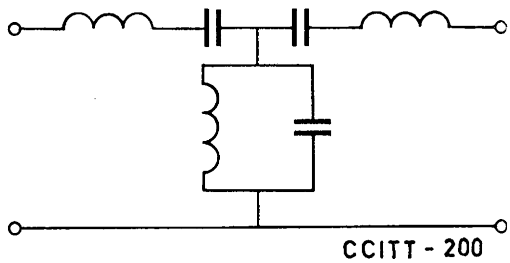


FIGURE 2

Transmission tests were carried out by objective measurements on the three circuits shown in Figure 1 and by comparative evaluations with the same equipments supplied with direct current and connected to an adjustable attenuator. Two values were used for the direct current, one corresponding to normal conditions with zero line and the other at half that value.

With the apparatus used in these tests, the reference equivalent of each four-wire circuit when the two sets are fed normally with direct current is 0 db and with direct current at half that value, 5 db. When the carrier current is 3 and 0.5 volts, the reference equivalents are 8 and 22 db, respectively.

When both calls are simulated with an artificial mouth, fed with a white noise such as to give a line level of −10 dbm for each call, a crosstalk noise at a level of −50 dbm is obtained in the receiver of the third circuit. Distortions of all kinds have identical values, whether the power is direct current or carrier current.

For the same values of reference equivalent on the circuits, the opinion tests gave the same results, for both types of feeding current.

Question 22/XII — Intelligibility of crosstalk in programme transmission

(continuation of Question 22 of Study Group XII, 1961-1964)

To what extent is the intelligibility of crosstalk between a telephone circuit and a programme circuit, or between two programme circuits, affected by the presence on the disturbed circuit of white noise and by the introduction into the programme circuit of emphasis networks, as recommended by the C.C.I.T.T. (see Recommendation J.21, Volume III of the *Red Book*)? Measurements should be made at various levels of disturbing speech and noise.

Note 1. — The C.C.I.T.T. recommendations for intelligible crosstalk are based on tests with the human voice. Designers of carrier systems require objective specifications and normally aim at providing the required crosstalk ratio at all frequencies within the relevant frequency band. Where the crosstalk path has a pronounced frequency characteristic (e.g. where emphasis is used in the disturbed or disturbing channel) or where the disturbed circuit is subject to a substantial level of noise, either of which conditions may affect the intelligibility of speech crosstalk, some modification in the recommendations seems desirable.

Note 2. — The following test conditions are suggested:

1. The disturbing telephone signal should be obtained from a typical telephone instrument with local exchange connections; the disturbing programme signal should be high-quality speech.

2. — The crosstalk signal on the disturbed programme circuit should be observed on high-quality monitoring equipment in quiet surroundings, set up first for normal listening levels and then having the sensitivity increased to simulate what happens when a broadcast receiver is switched on during a silent period.

3. At a point on a programme circuit at zero relative level the peak voltage of the programme signal is equal to that of a sinusoidal signal of 2.2 volts r.m.s. Values of crosstalk, speech volume and noise power on the programme circuit could conveniently be referred to such a point.

Note 3. — Only orders of magnitude are required from S.G. XII to begin with. On a long-term basis, Study Group XII might wish to consider the feasibility of studying a wider question on the possibility of devising a method of making an objective measurement of the intelligibility of crosstalk which will be significant for any likely frequency characteristic of the crosstalk path, for example, by means of an appropriate weighting network, as suggested in the text of Question 23/XII.

Note 4. — Annex 1, below, gives some results of subjective measurements made by the French Administration to estimate the masking effect of white noise on intelligible crosstalk. Annex 2 describes an experimental method which will be used by the British Administration.

ANNEX 1

(to Question 22/XII)

Tests carried out by the French Administration to assess the intelligibility of crosstalk in programme transmissions

Telephonometric measurements to assess how far white noise could mask intelligible crosstalk have been made by the French Administration.

Figure 1 shows the circuit arrangements used. Intelligible crosstalk was simulated by a succession of logatoms or sentences pronounced by operators, who spoke into a high-quality transmission system followed by a passband filter 300-3400 c/s, an attenuation line Z_1 (adjustable), and a fixed attenuation line (2 N).

A Gaussian white noise, adjustable in level by means of the attenuation line Z_2, could be included in the measurement chain.

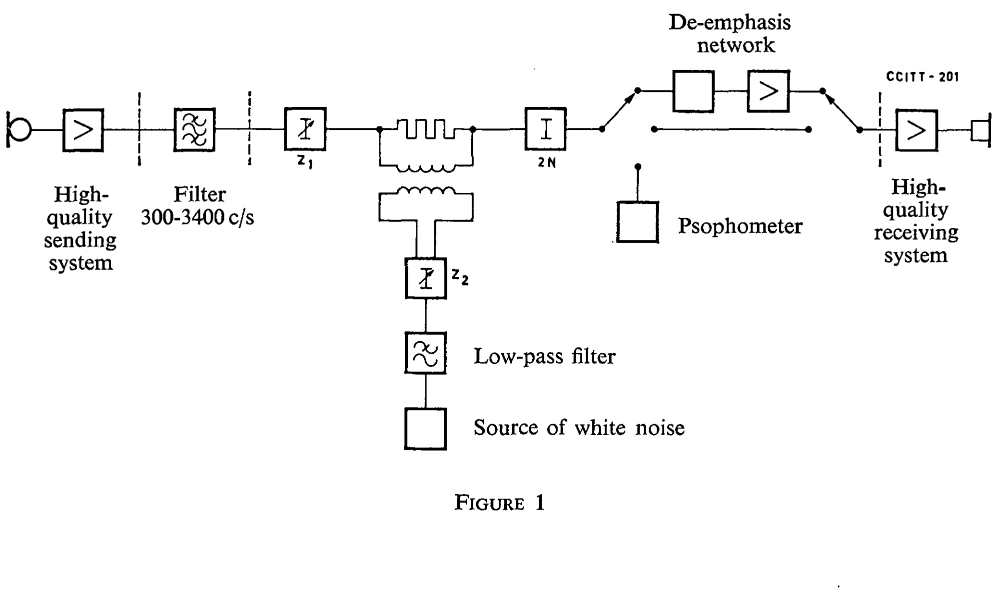


FIGURE 1

Two series of measurements were made, one with and one without the de-emphasis network recommended by the C.C.I.T.T. for programme circuits included in the chain between the fixed attenuator and the high-quality receiver.

This latter was in design and performance closely similar to the S.F.E.R.T. receiving system except that its reference equivalent had been adjusted to be 3.5 dN worse.

The results of these measurements are shown in Tables I and II. The crosstalk and noise levels shown therein are psophometric power levels with respect to the milliwatt, expressed in nepers (Nmp). When the de-emphasis network is not included in the chain (Table I), the measurement is made directly at the receiving system input. When the network is working (Table II) levels are measured at the input of the network, but have been increased by 1.5 dN to compensate for the gain introduced by the de-emphasis network and its associated amplifier.

Figures 2 and 3 show these results in diagram form. The curves show the variation of logatom or sentence articulation in relation to the noise-to-crosstalk ratio, the crosstalk level at the receiving system input being taken as parameter.

It seems that the desired masking effect becomes considerable when the difference between the level of white noise and that of crosstalk is 1 N at least. On the other hand, it seems that the distortion introduced by the de-emphasis network has no discernible effect on logatom articulation.

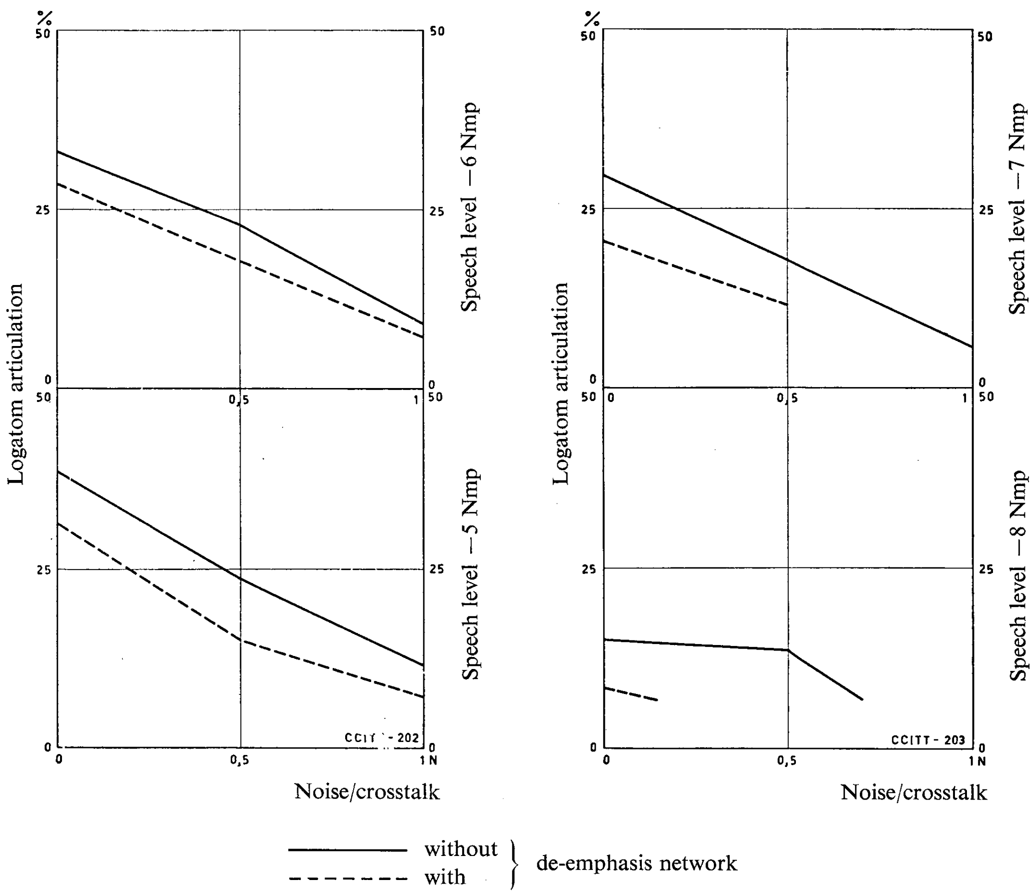


FIGURE 2

TABLE I

Intelligible crosstalk masked by white noise (de-emphasis network not inserted)

Articulation for logatoms and sounds

Crosstalk level in Nmp	Noise level in Nmp	Articulation in %	
		Logatoms	Sounds
−5	Without noise	93.2	97.7
	−5	38.6	73.9
	−4.5	23.7	61.8
	−4	11.5	47.4
	−3.9	2.8	39.9
−6	−6	33	70.4
	−5.5	23.2	63.3
	−5	9	48.1
−7	−7	29.5	67.5
	−6.5	18	59.5
	−6	5.7	41.8
−8	−8.5	19.2	59.2
	−8	15	53.2
	−7.5	13.7	48.3
	−7.3	6.3	42.2
−9	Without noise	3.8	34.2
	−9.5	1.5	24.2
	−9	1.5	21.4

Sentence articulation

Crosstalk level in Nmp	Noise level in Nmp	Articulation in %				
		A	B	C	D	E
−8	−7.5	30	2.9	4.6	7.9	54.6
	−7.3	11.2	2.1	2.9	3.7	80.1
	−7	8.3	0.8	0.8	5.8	84.3
−9	Without noise	12.1	4.2	2.5	4.1	77.1
	−9.5	6.4	0.7	0.7	4.8	87.4

A Sentences correctly received
B Sentences of which the general trend has been understood but with errors of detail
C Sentences in which one or more important words have been misunderstood
D Sentences of which the general sense has not been understood
E Received signal unintelligible

TABLE II

Intelligible crosstalk masked by white noise (de-emphasis network inserted)

Articulation for logatoms and sounds

Crosstalk level in Nmp	Noise level in Nmp	Articulation in %	
		Logatoms	Sounds
−5	−5	30.8	69.9
	−4.5	14.7	54.6
	−4	6.8	42.5
−6	−6	28.7	67.2
	−5.5	18.2	57.1
	−5	7.2	41.1
−7	−7.5	23.2	64.5
	−7	20.3	58.7
	−6.5	11.5	50.3
−8	Without noise	17.3	58.6
	−8.5	10.8	50.7
	−8	8.8	49.7
−9	Without noise	6.2	34.1
	−8.5	0	0

Sentence articulation

Crosstalk level in Nmp	Noise level in Nmp	Articulation in %				
		A	B	C	D	E
−7	−6.5	46.7	7.5	8.3	13.3	24.2
	−6	3.3	3.3	2.5	14.2	76.7
−8	−8.5	20.8	2.5	0.9	5	70.8
	−8	17.5	8.3	6.7	14.2	53.3
	−7.5	0	0	0	0	100
−9	Without noise	10	5.8	3.4	10	70.8
	−8.5	0	0	0	0	100

For the meaning of the letters A, B, C, D, and E, see Table I.

As far as sentence articulation is concerned, the results to hand are incomplete, since these measurements were made only where crosstalk levels were low. Figure 3 seems to show that the de-emphasis network interferes with sentences more than it does logatoms.

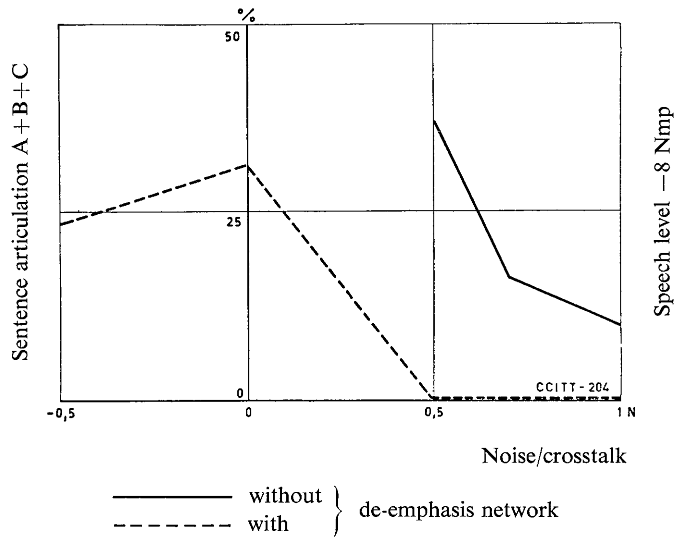


FIGURE 3

The use of these results for programme transmissions

To use these results to specify the crosstalk ratio on programme circuits would seem somewhat risky.

The following is proposed to Study Group XII as a working hypothesis.

Consider first of all the case when the disturbing circuit is a telephone circuit. Let x be the crosstalk ratio between the disturbing circuit and the disturbed programme circuit. The ratio between the signal and the intelligible crosstalk on this programme circuit will be of the order of $x + 2$, bearing in mind the difference in absolute levels at points of the same relative level in the disturbing and disturbed circuits (-1 Nm0 on the telephone circuit, $+1$ Nm0 for the peaks of the programme signal).

Reference to item 2 in Note 2 appended to Question 22/XII shows that the sensitivity of the broadcasting receiver can be increased during a silent period. Let y represent this additional gain. The intelligible crosstalk situated at $x + 2$ below the programme signal will become $x - y + 2$ below the " normal listening conditions " during the silent period.

For the high quality receiving system used in the subjective measurements, " normal listening conditions" represented an adjustment of $Z_1 = 0$, i.e., an absolute level of -2 Nmp at the receiving system input. This being so, the crosstalk level D to be taken into consideration in Tables I and II will be given by the equation:

$$-2 - D = x - y + 2$$

or

$$D = -4 - x + y \qquad (1)$$

When the disturbing circuit is itself a programme circuit, the absolute levels at points of the same relative level in the disturbing circuit and disturbed circuit are identical, and the signal-to-intelligible crosstalk ratio on the disturbed circuit is equal to the crosstalk ratio.

Equation (1) becomes:

$$D = -2 - x + y$$

ANNEX 2

(to Question 22/XII)

Method planned by the United Kingdom Administration for estimating the intelligibility of crosstalk on broadcast transmissions

Although no test results are yet available the following procedure has been planned.

Results showing the extent to which telephone conversations can be overheard intelligibly on radio receivers will be presented in the form illustrated in Figure 1.

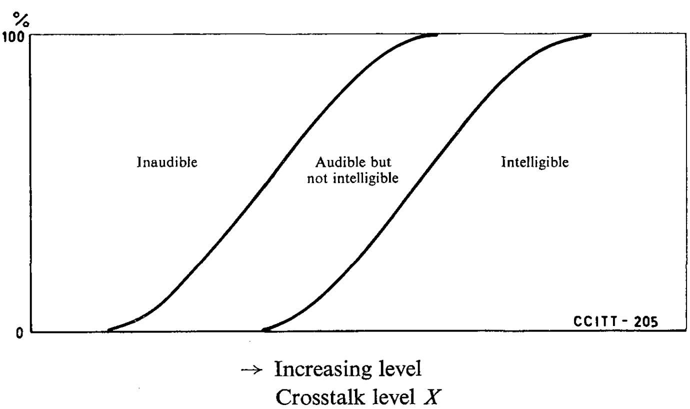


FIGURE 1. — **Presentation of results from crosstalk intelligibility tests**

The circuit shown in Figure 2 will be set up and the loudspeaker amplifier gain adjusted to provide the preferred listening level when the normal programme level is established at the zero level point. The frequency characteristic of the crosstalk path will be simulated by a network providing attenuation distortion which is in accordance with the upper sideband suppression of the appropriate filters.

Subjects will be presented with groups of five sentences at each of ten settings of attenuator X. The presentation of each group will be in the absence of any programme and will be signalled to the subject by a lamp. The subject will be asked to express his opinion concerning each group on the following scale:

(0) Inaudible;

(1) Audible but not intelligible;

(2) Intelligible.

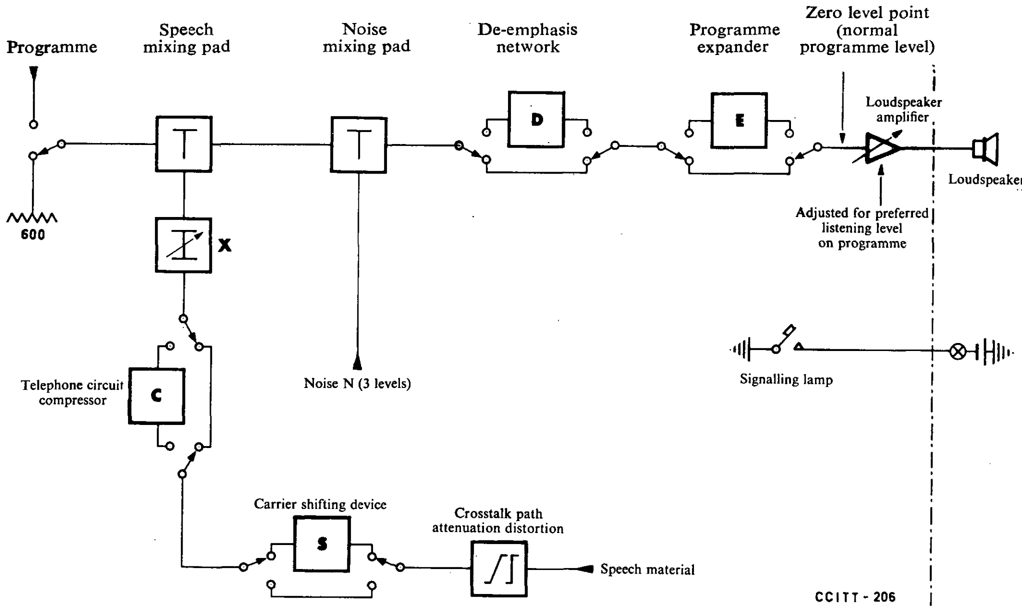


FIGURE 2. — **Circuit arrangement for crosstalk intelligibility tests**

The following factors require study:

N, Noise level on programme channel;

D, Absence or presence of programme de-emphasis;

S, Absence or presence of carrier shift in the crosstalk path;

C, Absence or presence of telephone compressor in the crosstalk path;

E, Absence or presence of programme expander in the programme path.

A systematic examination of all possible combinations of these would be much too long and so it is proposed to conduct two experiments as follows:

The first will serve to examine the effect of carrier shift upon intelligibility of crosstalk. Two levels of N will be used with each of the following carrier shifts, 0, 500, 1000 and 1500 c/s. Programme de-emphasis will be present but the telephone compressor and programme expander will be absent. Thus eight conditions will be covered. The experiment will require about four weeks work including preparation of recordings for each value of carrier shift.

In the second experiment will be included all combinations of N (three levels), D (without and with) and S (none and 500-c/s shift). Such an experiment (12 conditions) will require about four weeks work. The effect of a telephone circuit compressor will probably not be large if allowance is made for the increase in mean speech power. The effect of a programme expander can be estimated from the results with low values of N because, at the low levels involved and in the absence of programme transmission, the effect of the expander will merely be to introduce a constant loss.

Question 23/XII — Objective measurement of the intelligibility of crosstalk

(continuation of Question 23 of Study Group XII, 1961-1964)

a) Results of subjective tests on crosstalk from a telephone circuit to a programme circuit using pre-emphasis indicate that neither the loudness of the crosstalk nor its disturbing effect (such as will be measured for unintelligible crosstalk by the procedure described on page 63 of volume III of the *Red Book*) give a proper measure of its intelligibility.

Is it possible to make an objective measurement of the intelligibility of crosstalk which will be substantially independent of the frequency characteristic of the crosstalk path, for example, by using an appropriate weighting network?

b) If so, would it not be desirable to introduce this method of crosstalk measurement generally?

c) Given that circuit noise may tend to mask the intelligibility of crosstalk, would it be justifiable to make an allowance for, say, the lowest value of noise likely to appear on the circuit to be measured, and if so, how may this be done?

Note 1. — As a signal source, the use of the standardized speech spectrum shown on page 51 of Volume III of the *Red Book* (Figure 14) might be convenient.

Note 2. — Since the main purpose of the question is to define some means for the objective measurement of crosstalk intelligibility, this matter cannot usefully be considered before a reply, at least in part, has been given to Question 22/XII.

Question 24/XII — Extension of the bandwidth transmitted

(to be studied jointly with Question 1/XV)

(new question)

a) How much improvement in quality of transmission can be obtained by reducing the attenuation-frequency distortion in the lower-frequency range of the band of frequencies transmitted by international and national trunk circuits?

b) If such reduction in attenuation-frequency distortion were achieved, how much would the susceptibility to interference by low-frequency noise be increased, particularly so far as power-frequency harmonics are concerned?

c) How much additional speech power would need to be transmitted by carrier systems and what are the statistical properties of the additional low-frequency components?

Note. — It is the peaks of interfering power that are particularly important in considering disturbance to out-of-band signalling systems.

d) With such reduced attenuation-frequency distortion, what would be the effect of the phase-frequency distortion, bearing in mind the sharp low-frequency cut-off needed to reduce interference when out-of-band signalling systems are used?

e) What recommendations should be made to ensure balance return loss adequate for the new transmission plan, in the lower part of the frequency band transmitted?

Question 25/XII — Maintenance of subscribers' sets

(Question Africa H from Plan Sub-Committee for Africa)

(new question)

a) How should a satisfactory service be organized for the maintenance of telephone equipment and lines, particularly as regards the qualities covered by voice-ear tests: types of apparatus to be used—possibility of making simple voice-ear tests, particularly the sensitivity of sending systems (including the microphones), measurements of line noise, etc.?

b) Definition of measurement methods.

Note. — On this subject the C.C.I.T.T. has already issued Recommendation P.81 (*Red Book*, Volume V, pages 159-161). The texts cited in that recommendation should be completed or brought up to date by the following, which appear in the present volume: Annex H (part II) and Annex 1 to Question 15/XII.

Study Group XII will indicate whether important changes have been introduced by telecommunication Administrations in the procedures mentioned in Volume V or in the present volume V *bis* of the *Red Book*.

Question 26/XII — Transmission performance of carrier circuits for very short distances

(new question)

What general recommendation might be made by the C.C.I.T.T. concerning the transmission performance of circuits liable to form part of international connections set up on very short carrier systems (10 km or so long)?

The effect of syllabic or instantaneous compandors should be studied in particular.

Note 1. — Recommendation G.125 (*Blue Book*, Volume III) already makes certain recommendations applicable to such circuits, and Special Joint Study Group C, as part of its Question 9/C, is considering ways of supplementing these recommendations. As part of this question, the defects of transmission inherent in systems using coded-pulse modulation will be studied.

Note 2. — To study the effect of instantaneous compandors, information must be available about the filters and other parts of the systems. Study Group XV will be asked to supply this. Study Group XV, as part of Question 32/XV, is to assemble literature concerning the construction characteristics of these systems.

Question 27/XII — Transmission performance of pulse-code modulation systems

(new question)

a) What recommendation might be made by the C.C.I.T.T. about the standard of transmission performance assessment that ought to be achieved for a single audio-audio link provided by a practical engineered P.C.M. system bearing in mind the conditions under which such a link may form part of an international connection, as already indicated in Annex 1?

Note. — An attempt should be made to specify the method for expressing in quantitative form this standard of performance; this could take the form of an assessment score derived from a given assessment method or (perhaps more conveniently) as a setting of a given "reference device" which Study Group XII could specify.

In any case account must be taken of the wide range of input speech volumes and, possibly, separate figures would be necessary for low, medium and high input volumes.

b) So that acceptable values of fundamental parameters for an economical design may be recommended, the effects of the various factors contributing to the quantization distortion of an ideal P.C.M. system, e.g. peak clipping and centre clipping, should be assessed in the same units as used for a).

c) Since any economically engineered P.C.M. system will be liable to other forms of degradation due, for example, to transmission errors, quantization inaccuracies, synchronization difficulties, jitter, etc., how should such systems be measured in ordinary working conditions, to ensure that they live up to the requirements set forth under a) above?

Note 1. — It is essential that the method (or methods) be sufficiently simple to be carried out conveniently and without special precautions; it would also be very desirable, if possible, for this to be an instrumental measurement that could be made with readily available test gear.

Note 2. — Annexes 2 to 8 [1] give examples of the transmission performance which may be realized by practical P.C.M. systems.

ANNEX 1

(to Question 27/XII)

Considerations regarding the study of the question

1. Initially, attention might be concentrated on the following:

The P.C.M. system of transmission is operated only over the shorter national links; a maximum of up to four (two in each national network) unsynchronized systems may be assumed in an international connection. The characteristics to be assumed for the initial study are as follows (the comments of Study Group XII on the values of these parameters would be valuable):

Sampling rate: 8000 per second

128 sampling levels,

Companding law logarithmic, with a value of $\mu = 100$ (see *Bell System Technical Journal*, May 1957, page 653).

It may be assumed initially that the transmission equivalent of a link operated by P.C.M. is defined by the same mean value and standard deviation as for conventional transmission systems.

2. The study might be made taking as examples of international connections those studied by Study Group XII for Special Committee C, concerning the effects of circuit noise on performance.

3. The performance could usefully be expressed in terms of percentages of unsatisfactory calls.

[1] Annexes 3 to 6 are identical with Appendices 3 to 6 to Annex 2 to Question 33/XV.

ANNEX 2

(to Question 27/XII)

Effects of various transmission defects on the articulation of a pulse code modulation system

(Contribution by the Chile Telephone Company)

A series of determinations have been made of the impairment to sound articulation caused by a) centre clipping, b) time sampling and c) amplitude quantizing.

Tests were made over an orthotelephonic system with bandwidth limited to 300-2400 c/s (the S.E.T.E.D. " working standard "). The speech material consisted of lists of 80 logatoms each containing 200 sounds, phonetically balanced for the English language. A background of 50 db SL 'A' Hoth noise was present in the listening cabinets.

a) *Centre clipping*

A centre clipping circuit was placed in the trunk of S.E.T.E.D. The circuit was capable of being set at a particular level such that no speech was transmitted below this level. Figure 1 shows how the change in the degree of clipping effected the sound articulation. Each of the experimental

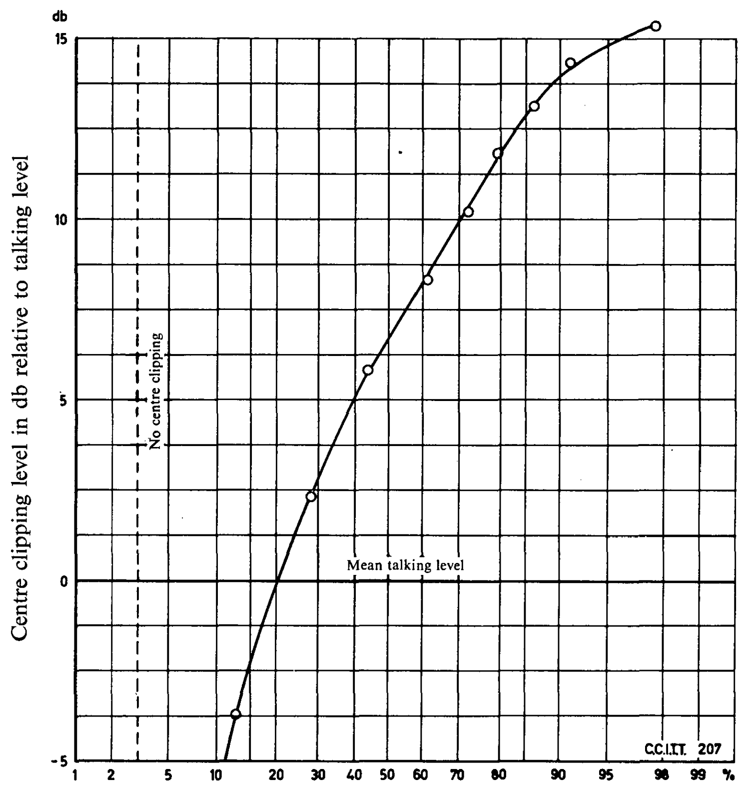


FIGURE 1. — **Centre clipper — Articulation error distribution**

points is based on an average of 4000 sounds heard over the circuit. The ordinate scale is the level at which the clipper was set in db relative to the logatom calling level as read on a vu meter. This was derived by an independent observer taking a series of vu meter readings in the actual logatoms rather than on the first part of the carrier phrase.

The study of centre clipping is important in ensuring that the threshold of the P.C.M. system is set sufficiently low so that weak limiting calls are adequately quantized.

b) *Time sampling*

A frequency controlled gating circuit followed by an RC circuit was inserted in the trunk of S.E.T.E.D. and the repetition rate varied from 1 to 6.8 kc/s. The results of articulation tests are as shown in Figure 2, together with a repeat test made over a commercial telephone circuit.

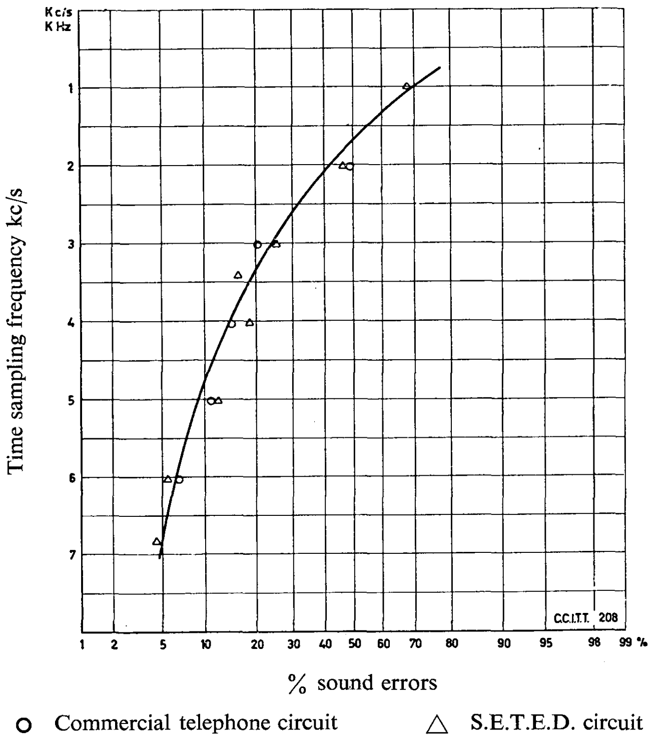


FIGURE 2. — **Time sampler — Articulation error distribution**

c) *Amplitude quantizing*

An amplitude quantizer with 32 linearly spaced levels (16 positive and 16 negative) was inserted in the trunk of S.E.T.E.D. and the input adjusted until the extreme levels were just operating. By attenuating the signal before the quantizer and by providing a compensating amplification after the quantizer it is possible to vary the number of operating levels. Further, by amplifying the signal before the quantizer and attenuating afterwards, it is possible to study the effect of all levels operating together with a controlled amount of peak clipping. The results obtained by articulation tests are as shown in Figure 3.

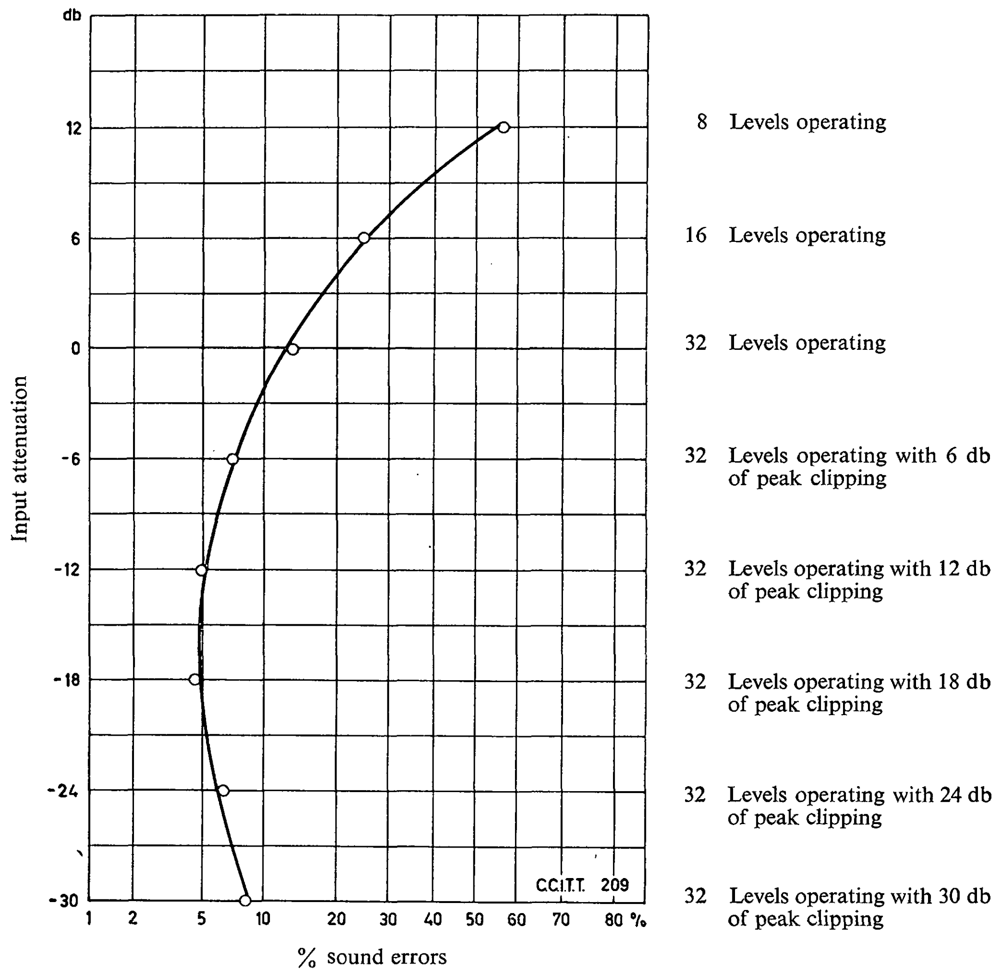


FIGURE 3. — **Amplitude quantizer — Articulation error distribution**

ANNEX 3

(to Question 27/XII)

Factors affecting speech impairment

(Note by the American Telephone and Telegraph Company)

As noted in the report of the Working Party on Question 33/XV, the impairment to an individual voice channel resulting from pulse code modulation of that channel is quite different from steady noise. Also, the impairments are introduced primarily by the P.C.M. terminals rather than by the digital transmission lines. Therefore, it is appropriate to study the over-all transmission performance of P.C.M. systems, and the terms in which that performance can be specified.

In considering applications of pulse code modulation it appears that the first applications will be for the provision of short-transmission links which may form portions of international connections. Standardization of equipment may not be required for this application because many of these links will not cross national boundaries. However, the formulation of recommendations with respect to performance is desirable to ensure satisfactory over-all connections. In these connections there will be only a few P.C.M. links in tandem and the major portion of the connection will generally consist of analog channels.

Second generation P.C.M. equipment will probably be used for longer distances and involve more P.C.M. links in tandem. Therefore, recommendations for the performance of P.C.M. links that may form a major part of international connections are important. The performance characteristics for individual channels must be established before system characteristics can be recommended.

The performance parameters should be stated in terms of parameters that are objectively measurable. The most definite measure now available is the noise and distortion power introduced as a function of input signal power. Any combination of number of digits, full load capacity, and companding characteristics that produces objectively measured noise and distortion powers equal to or better than a desired curve for noise and distortion will have equal or better subjective performance. Therefore, it appears appropriate to request Study Group XII to recommend a curve of maximum allowable distortion per link as a function of signal level.

In addressing Study Group XII, it will be beneficial to direct its considerations to distortion and noise curves representative of realizable P.C.M. systems. The following comments are offered to help Study Group XII appreciate the significance of this curve.

In order to achieve subjective advantage from companding, the noise and distortion power in the presence of speech will be higher than in the absence of speech. Thus the general tendency of any practical P.C.M. distortion curve is for increasing distortion with increasing signal. Since in the limit the quantization steps approach equal size at the origin, the distortion power becomes independent of signal power for low signal powers. At normal signal power it seems reasonable to desire a constant signal-to-(noise and distortion) ratio ($S/N + D$ ratio). Above the clipping level, the distortion increases very rapidly. Therefore, a reasonable noise and distortion curve for the formulation of performance recommendations is of the general shape shown in Figure 1.

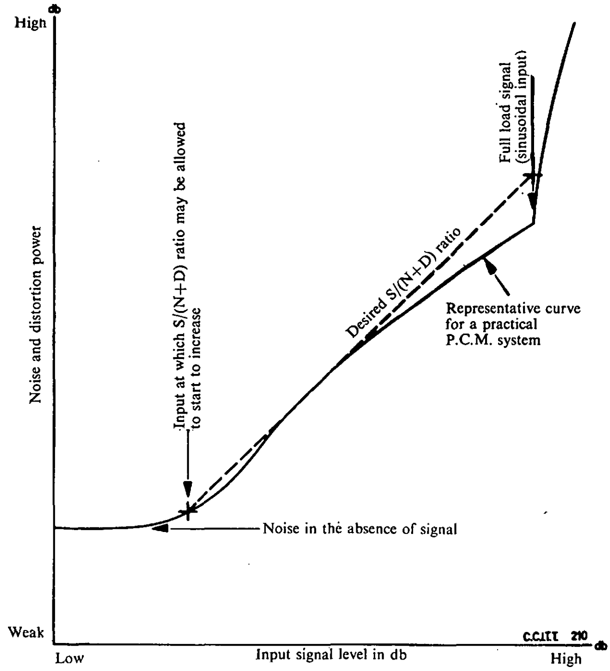


FIGURE 1

The desired distortion curve can be specified by a statement of the allowable noise power in the absence of signal, the signal-to-(noise and distortion) ratio that should be achieved for moderately high input powers, and the range of input powers over which this $S/N + D$ ratio should be preserved.

Except for the effects of clipping of the speech wave, the computed noise and distortion curves for sinusoidal inputs, assuming instantaneous logarithmic compandors, differ from computed curves for speech inputs by only a fraction of a db. Therefore, it is reasonable to use these curves interchangeably to express the results of objective measurements with sinusoidal inputs and to define the noise and distortion in the presence of speech. Consequently, they represent a good transition between subjective and objective measurement.

It should be noted that the distortion curve defines the average noise and distortion power over the speech burst and does not mean that the performance of the P.C.M. channel is subjectively equivalent to a circuit with constant noise of the same power. The noise on the circuit in the absence of signal is very important in the formulation of the subjective reaction of the listener; therefore, the equivalent constant noise is a function of both the noise and distortion in the presence of speech and the noise in the absence of speech.

ANNEX 4

(to Question 27/XII)

Companding law used in a P.C.M. system and its effect on transmission performance

(Note by Standard Elektrik Lorenz)

The following explanation refers only to companding with single-channel coding.

By the quantizing, which causes an unnatural reproduction of the input signal, a so-called quantizing noise arises which appears only during the actual duration of the conversation. This noise can be compared to that of a harmonic distortion noise; however, it results in a different subjective impression. In contrast to the harmonic distortion noise, the signal-to-quantizing noise ratio will decrease when the signal power decreases. Therefore, high levels give a better signal-to-noise ratio which, however, decreases substantially due to the clipping effect when the levels are still more increased.

To achieve a satisfactory transmission quality, a minimum signal-to-quantizing noise ratio must be provided in the volume range encountered in the operation (influenced by such factors as length of subscriber's line, quality of microphone capsule, attenuation plan etc.). If these requirements have to be achieved without companding, a large number of quantizing levels must be provided. By the companding, the signal-to-noise ratio will be improved, especially in the range of low signal power.

Nearly all contributions and publications have referred to a companding law conforming to the function

$$y = \frac{\log(1 + \mu x)}{\log(1 + \mu)} \tag{1}$$

where

y = output value normalized to the highest transmissible value

x = input value normalized to the highest transmissible value

μ = compression factor

The signal-to-noise ratio (ratio of signal to quantizing noise) ascertained with this companding law is shown as function of the dynamic ratio α in Curve 1 of Figure 1. This dynamic ratio α means the ratio of the r.m.s. signal level to the clipping level. If a definite volume range is specified in which a given signal-to-noise ratio is to be maintained, then it can be seen from Curve 1 that, in a part of this volume range, the signal-to-noise ratio is considerably better.

By choosing another companding law, the curve of the signal-to-noise ratio can be altered so that, in a large volume range, a given signal-to-noise ratio will be maintained. An example of this is shown in Curve 2 of Figure 1. To calculate this curve, a companding law of the following function is employed in the volume range with $\alpha \geqq -40$ db,

$$y = \frac{1 + \log \mu x}{1 + \log \mu} \tag{2}$$

The signs y, x and μ have the same meaning as in equation (1).

In the volume range $\alpha \leqslant -40$ db, a linear function for the companding law has been applied. The quantizing steps in this range are all of the same size. In summarizing, it will be found that the

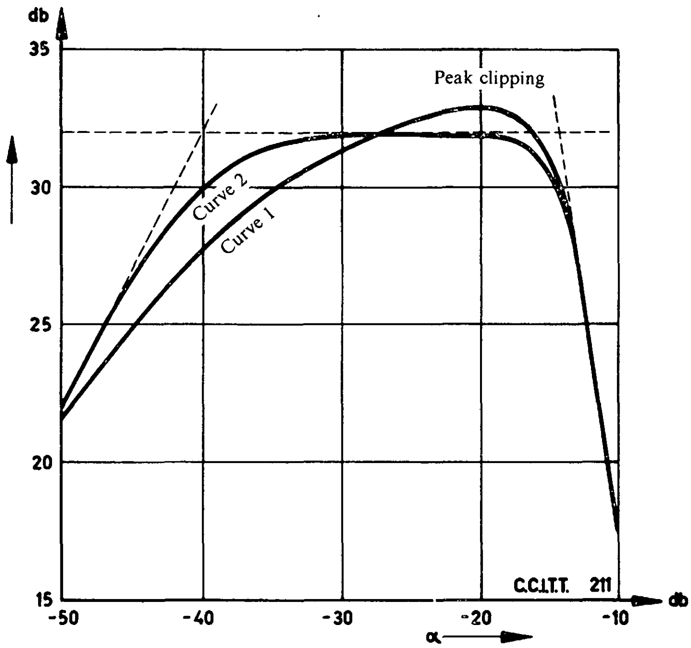


FIGURE 1. — **Signal-to-noise ratio *a* (logarithmic ratio of the signal to the quantizing and clipping noise) as a function of dynamic ratio α (ratio of the r.m.s. signal level to the clipping level)**

companding law of Curve 2, when compared with Curve 1, results in an advantageous increase of the volume range for a given signal-to-noise ratio.

On principle it can be said that the fixation of a companding law will still allow for technical solutions in very wide limits.

ANNEX 5

(to Question 27/XII)

Transmission quality in a P.M.C. system

(Note from Telefunken A.G., Backnang)

The definition of the transmission quality is of such an importance that this has to be clarified prior to the standardization of P.C.M. systems. Study Group XII has been asked for clarification and should also conduct experimental examinations.

As an aid for Study Group XII we would like to submit some contributions concerning the idle channel noise, the quantization distortion, and the companding problem. These contributions are based on the *Red Book* Volume III, and on the results of some examinations made by various Administrations.

A comparison with the noise recommendations for the usual carrier telephone terminal equipments and line amplifiers is difficult since, as is well known, the noise in P.C.M. systems is of a different nature and is moreover mainly produced in the terminal equipments and not on the transmission path.

Therefore, the usual reference to the circuit length of a transmission system does not appear to be very useful any more. It should be considered whether another definition would not be suitable, e.g. a determination of the idle noise and the modulation noise permitted in the over-all P.C.M. system from V.F. input to V.F. output.

The idle channel noise which mainly results from the performance of the terminal equipments being inevitably not ideally good due to random changes in the lowest amplitude stages caused by variations in the operating voltages or changes in the response thresholds of the switching stages etc., would have to be limited to a value acceptable in the noise planning for an international link. This problem is also closely related with Question 32/XV. Since the permissible noise power in short-haul systems is 1000 picowatts, this value could also be accepted for the idle channel noise permitted in P.C.M. terminal equipments.

Its realization also depends to a great extent on the number of amplitude stages in the P.C.M. system and on the companding law.

The same is true of the quantization noise and the peak clipping noise produced when a signal is transmitted over a channel.

According to the experience gained so far, a signal-to-noise ratio of 26-30 db will be required in case of speech transmission. The data published in the C.C.I.T.T. documents and the measurements of the speech power occurring at the input of an exchange show that the mean speech power of a talker is $N_m = 88\ \mu W$ or -10.6 dbm, as can be determined from Recommendation G.222, and that it can be expected to vary over a range of 20 db. These values would have to be used as a basis for the modulation and the companding in a P.C.M. system.

A good example for a companding law for better utilization of all quantization steps and for overcoming the level differences between the various speech channels with a tolerable distortion factor is deemed to be the logarithmic amplitude compression according to the relation

$$s_2 = A_1 \frac{\ln (1 + \mu \cdot s_1/A_1)}{\ln (1 + \mu)}$$

(C.C.I.T.T. 1961-1964, Contribution COM XV—No. 63, Appendix 1). The symbols used here and on the next pages have the following meanings:

A_1 = maximum amplitude that can be transmitted without limitation

s_1 = input amplitude

s_2 = output amplitude

μ = compression factor

s_{eff} = value of signal

α = A_1/s_{eff} = modulation ratio

a_k = signal-to-distortion noise ratio

q = number of quantization steps

The example in Figure 1 for $q = 128$ (i.e. 128 quantization steps) shows that the required minimum compression factor μ depends on the range of levels at which conversation shall be transmitted with a predetermined minimum signal-to-distortion noise ratio being maintained in the worst case.

In this figure, the total signal-to-distortion noise ratio (peak-clipping noise + quantization noise) is plotted as a function of ratio α, i.e. the ratio of the maximum amplitude that can be trans-

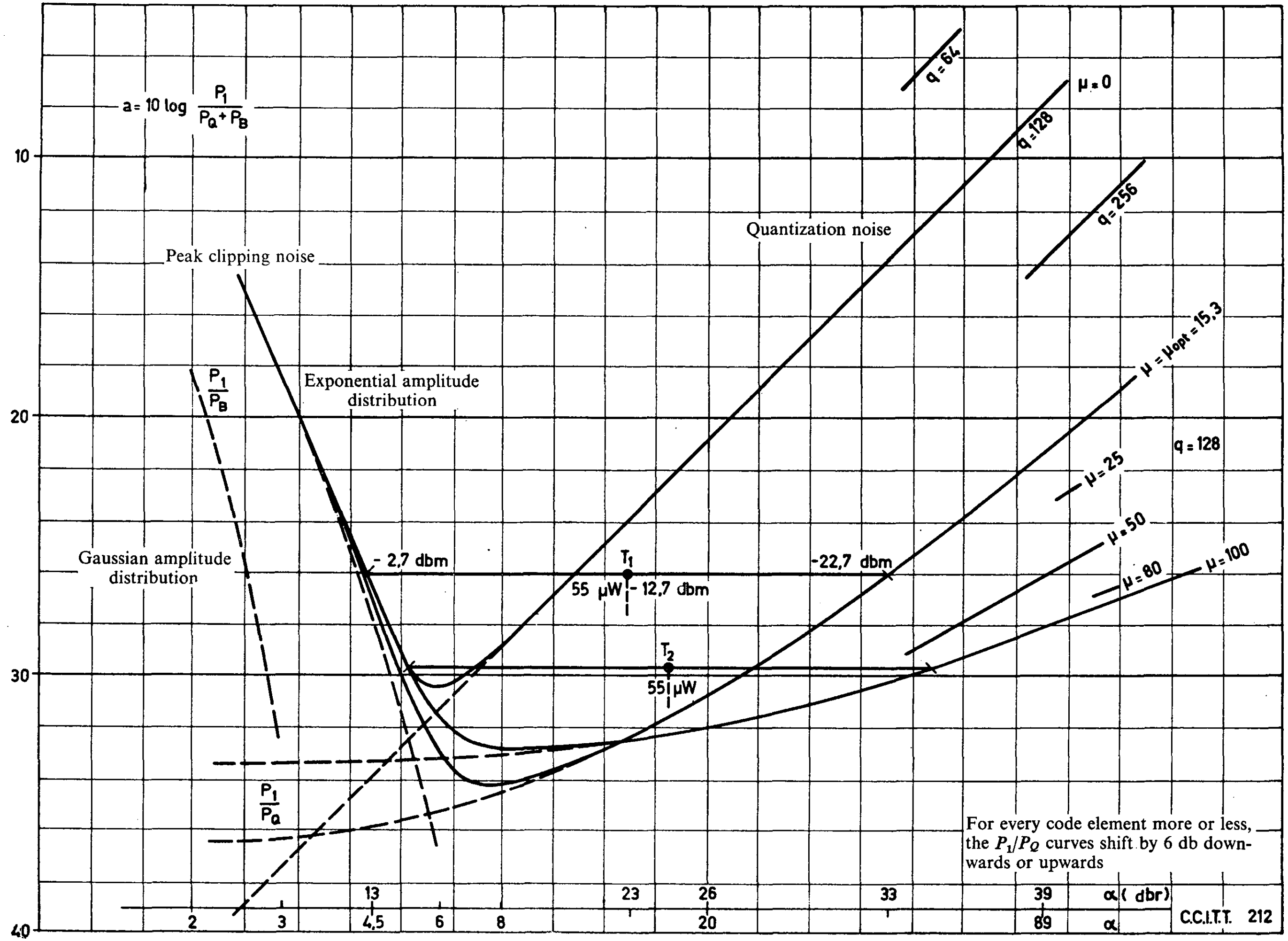


FIGURE 1. — **Total signal-to-distortion noise ratio in P.C.M. as a function of amplitude range**

mitted, to the r.m.s. value of the signal. If it is possible to take into account only one predetermined value for α (constant speech level), a signal-to-distortion noise ratio of $a_k = 30.4$ db can be achieved without compression ($\mu = 0$); when $\mu = \mu_{\text{opt}} = 15.3$, the maximum obtainable signal-to-distortion noise ratio in the case considered here (128 steps, exponential amplitude distribution in the speech signal, as generally assumed for single channels) is $a_k = 34.2$ db, and in the case $\mu = 100$, the signal-to-distortion noise ratio is $a_k = 32.8$ db. Hence, a compression factor of about $\mu = 15.3$ would be here the most appropriate one. However, since the mean speech powers vary considerably due to variations in talking power and differences in the losses of the subscriber lines, it will be necessary to deviate from this rating and provide for a greater compression in order to cover also very low levels.

For the determination of the mean speech power and its standard deviation only those times will be important during which speech is actually transmitted in one direction. The measurements carried out by most Administrations—among these also the Administration of the Federal Republic of Germany—are therefore only approximate clues, above all as far as the obtained standard deviations are concerned. Based on these measurements and on the measurements of the French Administration which appeared to be particularly well suited for these considerations, a value of 4.3 db has been assumed for the standard deviation of the mean speech power (to be distinguished from the mean conversation power which would also include the silent pauses).

Consequently, the median speech power N_{med} is, according to *Red Book*, Volume III, Annex 5, page 343:

$$N_{\text{med}} = N_m - 0.115\,\sigma^2 = -10.6 - 2.1 = -12.7 \text{ dbm} = 55\ \mu\text{W}$$

Assuming a Gaussian distribution of the speech powers of all channels, 98% of all conversations will vary within a range of $\pm$ 10 db with reference to the mean speech power. This total range of 20 db can be covered by 128 quantizing steps. Depending on the compression factor, the following signal-to-distortion noise ratios can be achieved: $\mu = \mu_{\text{opt}} = 15.3$, $a_k \geqslant 26$ db; $\mu = 50$, $a_k \geqslant 29$ db; $\mu = 100$, $a_k \geqslant 29.6$ db. The remaining 2% of the conversations will also be readily transmitted though with a somewhat higher non-linear distortion.

Therefore, compression factors between 50 and 100 will be suitable and it is to be noted that an increase of μ from 50 to 100 for the assumed volume range will bring, for very low speech powers, an improvement of 0.6 db only in the signal-to-distortion noise ratio whereas this ratio will be somewhat deteriorated for the majority of the conversations.

For the exact determination of the compression factor thus additional measurements and data would be valuable to indicate the extent of the deviation range around the mean speech power which should be taken into account (this is, in our opinion, the task of Study Group XII).

In case of 128-step quantization and a suitable companding following a logarithmic compression curve, the modulation noise for a rather wide amplitude range will be about 30 db lower than the effectively transmitted signal. This corresponds to a disturbance noise of about -40.5 dbm $= 90\,000$ picowatts. Changing the step number by the factor 2 will change the noise level by about $\pm$ 6 db and consequently the noise power by the factor 4 approximately.

The value of the permissible noise power is at present being considered by Study Group XII, and we should wait for the results of these studies. The signal-to-noise ratio of 30 db as obtainable for a rather wide amplitude range with 128 steps corresponds to a distortion factor of about 3%. This value could be used provisionally as a basis as long as more results expected from Study Group XII with regard to the disturbing effect of the quantization noise are not available.

ANNEX 6

(to Question 27/XII)

Effect of noise on transmission performance in a pulse code modulation system

(Note by the United Kingdom Administration)

1. *Introduction*

The purpose of this contribution is to describe a method whereby the acceptability in service of junction P.C.M. systems may be estimated and conclusions arrived at concerning the necessary number of amplitude quantization steps and the desirable companding law.

The method consists first of calculating a mean effective value of the ratio of speech power to quantization distortion power for each of a number of values of speech input level. This ratio, expressed in db, is converted to the ratio of speech power to constant random noise power that would, in the absence of the quantization distortion, give the same degree of acceptability. This enables the large body of information, already available, relating to various combinations of over-all reference equivalent and circuit noise level to be used to calculate the percentage of unsatisfactory calls likely to occur in various selected classes of telephone connection.

Some of the assumptions made and the numerical values of certain parameters used need checking experimentally but the method described is believed suitable for the problem in hand.

2. *Symbols and definitions*

B = number of possible output states (voltages) for positive values of instantaneous input voltage. In addition the same number of states are also available for negative-going voltages. Systems will be distinguished by referring to the number of output states rather than the minimum number of binary digits into which they could be coded; thus a 64+64-state system represents one often referred to elsewhere as a seven-digit system;

i = a number to identify a particular output state. The states are numbered positively from 1, nearest to zero volt, up to B and negatively from -1 to $-B$;

v = instantaneous voltage of the input signal;

$p(v)$ = probability density function describing the statistical distribution of v;

V = maximum value of v for a fully-loading sinusoidal test signal;

$T_{\max}$ = mean power level (dbm) of a fully-loading sinusoidal test signal, measured at the input of the system;

v_i = instantaneous value of input voltage at which the decision is taken whether, in the case of positive voltages, to code as the $(i+1)$-th or the i-th output state. For negative voltages read $(i-1)$ for $(i+1)$;

v_0 = centre decision voltage ideally coinciding with zero input voltage;

w_i = voltage of the output signal corresponding to the i-th state;

h_i = voltage interval between adjacent output states;

S = long-term mean power of a given talker (dbm), measured at the input of the system;

$i/B = f(v_i/V)$ is a mathematical function that describes the companding law and therefore enables h_i to be expressed as a function of v_i. When B is large, h will be treated as continuous and denoted by $h(v)$;

$\overline{R}$ = average value of the ratio of signal power to quantization distortion power. $\overline{R}$ will be expressed in db. The averaging can be performed in various ways which distribute differently the relative weights accorded to high- and low-level components of the speech. Two forms of averaging are mentioned below, distinguished by the symbols $\overline{R}'$ and $\overline{R}''$;

N' = level of constant random noise occupying the same frequency bandwidth as that of the speech channel transmitted by the P.C.M. system, that would give the same degradation to speech if present in place of the quantization distortion. N' is expressed as unweighted power (dbm), its level being referred to the input of the system.

3. *Determination of* $\overline{R}$

The output of a P.C.M. system consists of a waveform, quantized in time and in amplitude, which is filtered to restrict the bandwidth to that of the input signal, namely that of a normal telephone channel. The time quantization (sampling rate) is likely to be of the order 8000 per second. The amplitude quantization, however, is a subject for discussion; a likely choice for B being between 60 and 130.

The speech signal from a given talker in a conversation covers a very wide range of power levels; the peak instantaneous power observed during a few seconds of talking may reach 18 db above the long-term mean power and significant syllabic power is present down to 20 db below the mean power and even lower.

The waveforms of signals will be assumed to be, on average, symmetrical about zero and so, for convenience, positive and negative ranges of input and output voltage can be treated without regard to sign. This convention much simplifies the treatment of companding.

The output states of a P.C.M. system intended to transmit speech effectively must correspond to values of w_i that are not equally displaced from their neighbours; h_i must increase as i increases.

The ratio of signal power to quantization distortion power in a given P.C.M. system thus varies from instant to instant depending upon: a) the long-term mean speech power from a given talker; b) the value of B; and c) the companding law. This ratio must be averaged over the range of speech power levels that occur within utterances and this can be performed in several different ways.

One form of averaging (see, for example, reference [1]) treats all distortion equally, irrespective of the level of the signal with which it is for the moment associated. The operation can be defined by the formula:

$$\overline{R}' = 10 \log_{10} \left[\frac{\int_0^\infty v^2 \cdot p(v)\, \mathrm{d}v}{\int_0^\infty (1/12)\, h^2(v) \cdot p(v)\, \mathrm{d}v} \right] \tag{1}$$

Figure 1 shows the quantity $\overline{R}'$ as a function of $S - T_{\max}$ for the following companding law when the parameter $A = 100$.

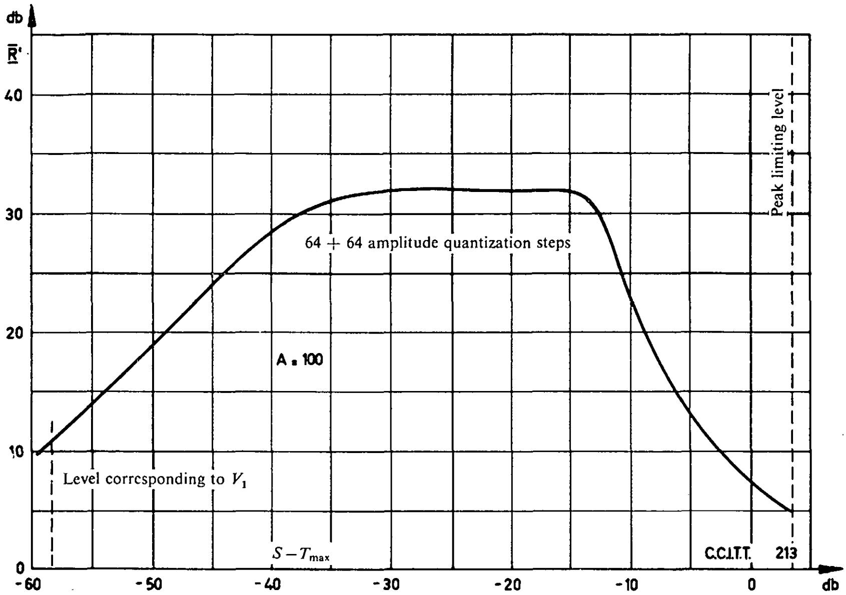


FIGURE 1. — $\overline{R}'$ as a function of $S - T_{max}$

$$i/B = \frac{A\,(v/V)}{1 + \log A} \text{ when } 0 \leqslant (v/V) \leqslant 1/A \tag{2}$$

$$\text{and } i/B = \frac{1 + \log A\,(v/V)}{1 + \log A} \text{ when } 1/A \leqslant (v/V) \leqslant 1 \tag{3}$$

The instantaneous speech voltage distribution used in reference [2] has also been used here, namely

$$p(v) = \frac{0.828}{v_{\text{r.m.s.}}} \exp\,(-1.183\; v/v_{\text{r.m.s.}}) \tag{4}$$

It can be shown (see reference [3] that choice of a companding law to maximize $\overline{R}'$ does not yield satisfactory results because too much importance is attributed to distortion products that are associated with high-power speech sounds and insufficient to those that accompany weak sounds. An interesting suggestion to overcome this effect is made by the same author.

Preliminary tests by the British Administration have shown that it is necessary to use functions more like those shown in Figure 2 if good agreement is to be secured with subjective opinions. Figure 2 has been calculated, for the range $0 \leqslant v \leqslant V$, from the formula below; m has been made equal to 4.

$$\overline{R}'' = m \cdot 10 \log_{10} \int_0^V \left[\frac{v^2}{(1/12)\, h^2(v)}\right]^{1/m} p(v)\, \mathrm{d}v \tag{5}$$

In the region $v > V$ where peak clipping occurs an arbitrary allowance has been made, as will be seen by comparing Figure 1.

The best form of averaging is currently being studied and the curves of Figure 2 must be regarded merely as provisional.

They will nevertheless be used to illustrate what follows.

4. *Relationship between $(S - N')$ and $\overline{R}$*

$S - N'$ is the ratio, expressed in db, of speech power to unweighted random noise power that is equivalent, as regards impairment, to $\overline{R}$. It will be assumed here, until the subjective tests now in progress are completed, that these two quantities are related in the form

$$(S - N') = \overline{R}'' + k \tag{6}$$

where k is a constant assumed independent of S and N'. Preliminary experience suggests that $k = 5$ db would not be an unreasonable guess at the present stage.

The right-hand scale of Figure 2 shows, on these assumptions, $(S - N')$ as a function of $(S - T_{max})$.

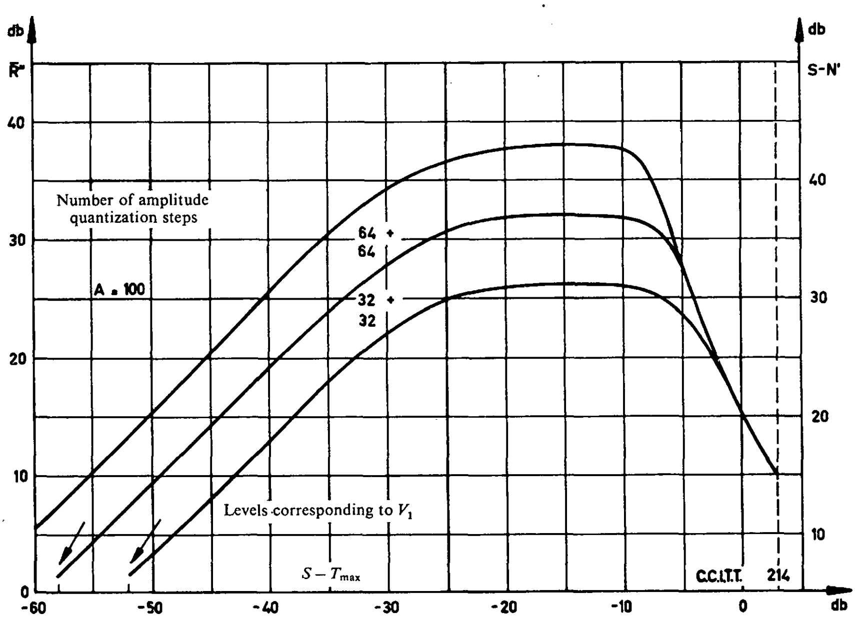


FIGURE 2. — $\overline{R}''$ as a function of $S - T_{max}$

5. *Performance of a P.C.M. system in a telephone network*

The P.C.M. systems to be considered will be assumed to have the following characteristics:

Sampling rate, 8000 per second.

Companding law according to equations (2) and (3) with $A = 100$.

The cases of $B = 32$, 64 and 128 will be examined.

The mean power of a fully-loading sinusoid, T_{max}, will be taken as + 6 dbm referred to the input of the P.C.M. system.

The transmission loss of the link containing the system will be taken as 3 db.

The class of telephone calls taken as an example is defined by Figure 3 and Table 1. The distribution of over-all reference equivalent was derived from the following components:

1. Mean reference equivalent of local ends and junctions for sending: 10.8 db
2. Standard deviation of (1): 3.3 db
3. Mean reference equivalent of local ends and junctions for receiving: 3.4 db
4. Standard deviation of (3): 2.7 db.

TABLE 1

Particulars assumed for various values of over-all reference equivalent (see Figure 3)

Over-all reference equivalent	Relative frequency of occurrence	E_S	$J_1 = J_4$	$J_2 = J_3$	E_r
db	%	db	db	db	db
40	0.02	12	0	6	3
36	0.3	11	0	5	2
32	2.7	10	0	4	1
28	12.3	9	0	3	0
24	27.9	8	0	2	−1
20	31.9	7	0	1	−2
16	18.6	6	0	0	−3
12 and lower	6.3	—	—	—	—

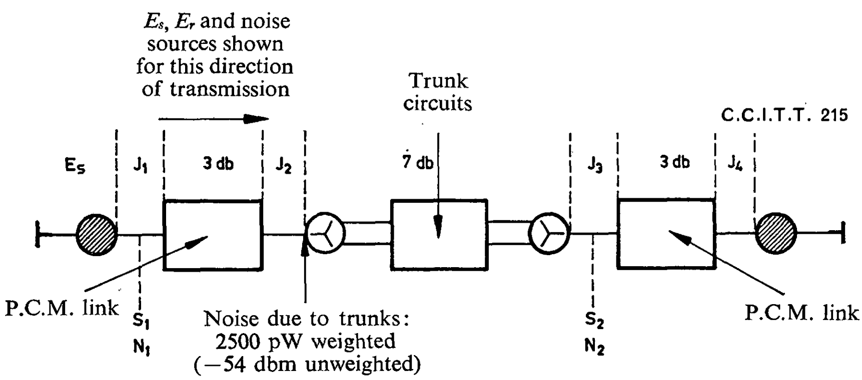


FIGURE 3. — **Telephone connection containing P.C.M. junction systems**

The trunk circuit is assumed to be composed of three links each with standard deviation 1 db: the over-all reference equivalent therefore has a mean of 21.2 db and a standard deviation of 4.6 db. These particulars are approximately in accordance with the results of statistical surveys taking calls as the basis (rather than possible connections).

Transmission performance calculations are carried out as follows. The median speech levels S_1 and S_2 at the points indicated in Figure 3 are estimated from the results of conversation tests and the values used here are given in Table 2 below. The values of $(S_1 - N'_1)$ and $(S_2 - N'_2)$ are read from the right-hand scale of Figure 2 for the appropriate values of $(S - T_{max})$. N'_1 and N'_2 can then be calculated, the values obtained being shown in Table 2. N'_1 and N'_2 are then referred to the input of a hypothetical receiving end having a reference equivalent of 0 db by subtracting $(13 + J_2 + J_3 + J_4 + E_r)$ and $(3 + J_4 + E_r)$ db respectively; these noise levels are denoted by N_1 and N_2. Noise arising in the trunk circuit is assumed equivalent to 2500 pW weighted power referred to the two-wire input; when expressed in unweighted form the level is — 54 dbm at the same point, and $(-54) - (10 + J_3 + J_4 + E_r)$ dbm at the hypothetical point referred to above; the latter level is denoted by N_3.

The total equivalent unweighted noise level at this point is given by power addition of N_1, N_2 and N_3 and denoted by N.

TABLE 2

Calculation of the total unweighted circuit noise level equivalent to the quantization distortion and other noise for various values of over-all reference equivalent

64 + 64 step system

Over-all reference equivalent	S_1	$S_1 - N'_1$	N'_1	S_2	$S_2 - N'_2$	N'_2	N_1	N_2	N_3	N
db	dbm	db	dbm	dbm	db	dbm	dbm	dbm	dbm	dbm
40	−18	36	−54	−40	18	−58	−82	−64	−73	−63
36	−18	36	−54	−38	20	−58	−79	−63	−71	−62
32	−17	36	−53	−35	23	−58	−75	−62	−69	−61
28	−16	36	−52	−32	26	−58	−71	−61	−67	−60
24	−16	36	−52	−30	28	−58	−68	−60	−65	−58
20	−15	36	−51	−27	31	−58	−64	−59	−63	−57
16	−15	36	−51	−25	32	−57	−61	−57	−61	−54

The percentage of " poor " or " bad " opinions in conversation over connections having the combinations of over-all reference equivalent and circuit noise level given in the first and last columns of Table 2 can be read from tables available elsewhere. These results are summarized in Table 3 for the following cases:

— No P.C.M. systems, junctions by loaded cable

— P.C.M. systems with $B = 128$

— P.C.M. systems with $B = 64$

— P.C.M. systems with $B = 32$

TABLE 3

Percentage of unsatisfactory ("poor" or "bad") calls

Over-all reference equivalent	No P.C.M. system	Value of B for P.C.M. system			Junction carrier system
		B=128	B=64	B=32	
db	%	%	%	%	%
40	33.6	36.3	40.6	48.3	43.4
36	18.9	20.6	23.9	30.4	25.8
32	9.7	10.5	12.1	16.1	13.5
28	4.2	4.5	5.4	7.5	6.5
24	1.76	1.94	2.3	3.3	2.6
20	0.67	0.71	0.84	1.17	0.94
16	0.28	0.30	0.35	0.46	0.35

In addition, for comparison, the two P.C.M. systems have been replaced by F.D.M. junction carrier systems having the following characteristics:

— Attenuation/frequency distortion: negligible in comparison with that caused by the trunk circuits;

— Compandors: none;

— Circuit noise level: equivalent to 2000 pW weighted noise power referred to the input terminals of the junction carrier systems.

In all five cases considered below the noise levels present on the trunk circuits as shown in Figure 3 remained unchanged. Room noise at a level of 50 db was also assumed in all cases to be present at both subscribers' stations.

The results in Table 3 can be reduced to a single figure for each type of system by weighting the percentages according to the relative frequencies of occurrence as quoted in Table 1. The results are as follows:

— No P.C.M. system: 1.60%

— P.C.M. systems with $B = 128$: 1.73%

— P.C.M. systems with $B = 64$: 2.04%

— P.C.M. systems with $B = 32$: 2.84%

— Junction carrier systems: 2.34%

6. *Conclusions*

Section 5 above describes a method whereby Point 1 of the Communication from Study Group XV to Study Group XII (part *a* of Question 27/XII) could be studied. If the validity of the assumptions made is accepted, a P.C.M. system that shall introduce no more degradation than a junction carrier system having the suggested limit of circuit noise of 2000 pW, the P.C.M. system must be equivalent to one having tpe companding law assumed and at least 64 + 64 amplitude quantization steps. The subjective performance of such a system could be specified in terms of the levels of constant channel-limited, random noise that, for each of a set of mean speech powers covering a given range of speech volume, incur respectively the same impairments to transmission performance. Alternatively an adjustable reference device can be used, constructed so that it introduces random noise of which the power (averaged over a short interval of time) is always proportional to that of the short-term mean (syllable) power of the speech of a given talker. The setting of this device is defined by the ratio of speech signal power to that of the noise. The device is used for comparison purposes, a setting being determined by subjective tests to give the same impairment, at a given speech level, as the P.C.M. system.

Point 2 of the Communication (Part b of Question 27/XII) can be studied by some method similar to that described in Section 3 and Section 4 above. The validity must, however, be verified and the correct numerical values of the parameters established.

An instrumental method for checking the performance of a P.C.M. system (Point 3 of the Communication, Part c of Question 27/XII) is unlikely to be simple if it is to provide a reliable check of all possible forms of error, inaccuracy or other trouble. An exhaustive study of these would take an excessively long time and so it will probably be necessary to adopt instrumental tests that will check only the commonest troubles and then to rely on a subjective test as a final check.

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ANNEX 7

(to Question 27/XII)

Quantization noise in a P.C.M. system and its influence on the quality of the transmission

By I. BARDUCCI, F. BIANCHI and L. GIACOMELLI

(Contribution by the Italian Administration)

1. *Introduction*

To transmit information contained in a signal with a complex spectrum limited at a maximum frequency f_0, it is sufficient to transmit a discrete number of the values assumed by the signal in interval instances which are more than $T_0 = \frac{1}{2f_0}$; in other words it is sufficient to transmit a series of sample signals taken with a rate that is not inferior to double the maximum frequency.

Because of the impulsive character of such a representation of the original signal, the possibility of multiplying the messages that can be transmitted on the same link with a time division system is evident.

The signal samples taken in an analogical form can be converted for the transmission into a numerical form with a proper code. The numerical representation of an ordinary physical quantity can only be discontinuous, because of the limited number of the different combinations possible in a practical code. The analogico-numerical conversion is therefore necessarily linked to a subdivision of the whole range of the possible values in partial intervals, to each of which a single representative value is given, and which is called the quantization level.

A transmitting system based on those principles takes the name of pulse code modulation system and includes the following basic operations: sampling, multiplexing, quantization, coding, and, when received, decoding and separation into the various channels.

The advantages offered by this transmitting system are the possibilities of multiplying messages and the independence, within vast limits, of the transmission quality from line disturbances. Its disadvantages are its need of occupying a much greater frequency band than occupied by the original signal and in the formation of a new type of disturbance, which is linked to the quantization process. This disturbance stems from the alterations effected by the substitution of discrete levels to the continuous levels of the signal. Finally, the quantization process creates a disturbance, called quantization noise or distortion, whose characteristics are those of a background noise with complex spectrum which is added to the signal and is present only when the signal is present.

Various theoretical studies have been made on this quantization noise to try to explain its origin and to determine its strength as well as its spectral distribution in relationship to the number of binary digits used for representing the signal. The principal results of these studies, in the case of uniform quantization and for complex signals, can be summed up as follows:

— the quantization noise has a uniform spectrum (white noise) in all of the signal frequency band;

— the noise level is independent of the amplitude of the signal and depends solely upon the quantum size, i.e. upon the value of the quantization step;

— the ratio of the r.m.s. value of the maximum sinusoidal signal which can be transmitted to the r.m.s. value of the quantization on noise voltage can be expressed in the case of n binary digits by the equation:

$$S/N = 6\,n + 1.8 \text{ db} \qquad (1)$$

With an increase in the number of the binary digits (called bits) there is a decrease in the noise intensity; however, there is an increase in the complexity of the equipment and in the frequency band necessary for transmission.

It is to be noted, however, that as the noise voltage is independent of the signal intensity equation (1) does not express, in reality, the signal-to-noise ratio, but only the ratio between the maximum r.m.s. sinusoidal voltage which can be transmitted and the r.m.s. value of the quantization noise voltage re. to given zero level, which is characteristic of the equipment. The maximum level of r.m.s. sinusoidal voltage which can be transmitted may be chosen as reference level. Thus equation (1) becomes:

$$N_q = -[(6\,n + 1.8)] \text{db} \qquad (2)$$

where n is the number of bits of the equipment, no matter how many quanta are actually used for reproducing a given signal. This second form for the quantization noise level makes it evident that we may speak of a quantization noise level for a certain number of bits, whatever be the signal amplitude.

2. *Research purpose and experimental equipment*

For practical purposes, in a P.C.M. system the coding of the signals is necessarily made with a limited number of bits. This means that sampled signals can be effectively transmitted only if their amplitude does not surpass the coder's saturation voltage, given in the case of uniform quantization by the sum of the number of quanta times the voltage assigned to the quantum in the particular device.

This limit value can be usefully characterized by means of r.m.s. voltage of a sinusoidal signal having the above-mentioned amplitude. This r.m.s. value is called " full sinusoidal load voltage ".

Because of quantization, the output signal will always be affected by a noise whose intensity is proportional to the quantum size associated with a characteristic " granular effect ", i.e. with a sensation of intermittence in the circuit continuity, which is not to be found in the conventional systems.

This study was undertaken in order to obtain experimental data on the disturbances caused by quantization to the transmission of speech signals and the improvement of the quality attainable with instantaneous dynamic compression-expansion devices.

For this purpose the following determinations were made:

1) Determination of the white noise level equivalent in loudness to the quantization noise level in a P.C.M. system as a function of the bit number, with or without instantaneous compandor in the case of the speech transmission.

2) Determination of the transmission quality obtainable with a P.C.M. unit with or without compandor as a function of the bit number.

To test the experimental equipment objective measurements of the quantization noise with sinusoidal signals and determinations of the noise spectrum both with sinusoidal signals and with a third-octave band of white noise were made.

The experimental equipment used consisted of a P.C.M. monocanal system with uniform quantization, using a binary code, manually set from 2 to 8 bits, and with a 8000-c/s sampling frequency. An instantaneous compandor could be inserted at the terminals of the equipment as indicated in the block diagram in Figure 1.

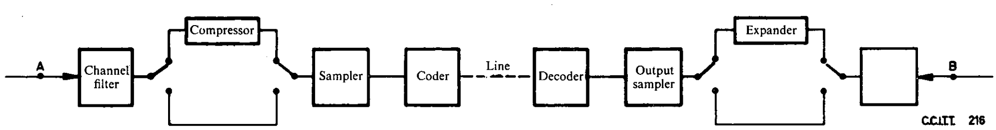

FIGURE 1. — **Block diagram of the P.C.M. link, with and without compandor**

3. *Choice of the equipment working level and the testing reference signal*

Because the P.C.M. equipment cannot distinctly transmit signals whose amplitude is greater than the coder saturation voltage, it is necessary to regulate the signal level so that the peak-clipping distortions are contained within acceptable limits.

To ensure these conditions it is necessary to establish a signal reference level at a given point at the input of the equipment and to establish the maximum admissible peak clipping. As reference level the full load sinusoidal voltage in the above-mentioned measuring point was assumed.

This value at the input of the tested equipment (point A in the block diagram in Fig. 1) was 100 mV across 600 ohms, or -18 dbm.

The measurements were made with a sinusoidal signal at 800 c/s. The equipment was so set as to have a zero equivalent between A and B points; in these conditions full load sinusoidal voltage at the output terminals was 100 mV as at the input.

The zero equivalent and the 100 mV full load sinusoidal conditions were re-established in the A point even when the compandor was inserted. The voltage at the earphone terminals (S.T.C. type 4026A) was regulated to 15 mV for the full sinusoidal load. The acoustic pressure generated by the earphone in the artificial ear recommended by C.C.I.F. (*Green Book*, Volume IV, p. 118) was 70 db (re. 2×10^{-4} dyne/cm^2), when at the earphone terminals a 800-c/s and 15-mV signal was applied.

The working level assumed was the maximum signal level which caused inaudible peak clipping even with the maximum bit number.

In this way the peak-clipping noise did not affect the determination of the white noise which was equally as loud as the quantization noise. It must be pointed out, however, that the best transmission quality is not normally obtained in this condition, because the quality improvement related to the signal average level increasing initially overcomes the coder saturation negative effects chiefly for the lowest bit number.

The speech volume at the input of the first channel filter (point A on the diagram in Fig. 1) was read on a normalized vu meter.

The maximum working speech level stated above was initially determined in this way: a speaker read a passage with a natural and constant voice: a listener varied the signal level by an attenuator inserted before the A point, until the peak-clipping distortions were hardly detected. In those conditions an oscilloscope monitoring showed a slight saturation of the coder in short intervals corresponding to the emission of accented syllables for several words which were pronounced with greater emphasis; the measurements effected with the vu meter indicated a level about 3 db below the full sinusoidal load level at the same point.

These determinations were often repeated with various operators, but the level always remained about 3 db below the full sinusoidal load. Therefore the emission was so adjusted as to have a signal volume of that value.

4. *Subjective methods used for evaluating quantization noise and transmission quality*

The essential feature of quantization noise is that of being present only when the signal is present; furthermore, the intensity and the spectrum of noise do not depend on the signal intensity, but only on its waveform. The direct experimental determination of the quantization noise intensity in case of speech signals must therefore be made with the signal present.

Because it is impossible to separate the speech signal from the complex spectrum noise, such as quantization noise, it was necessary to use subjective methods of evaluation, namely:

a) Determination of the noise loudness by direct comparison with another noise of known intensity; this method, hereafter called *balancing method*, is based on the discriminating power of hearing, i.e., on the capacity of the ear to detect the noise and to evaluate its loudness even in the presence of a louder signal.

b) Determination of the threshold of intelligibility based on the masking effect of noise on the signal.

c) The *users' opinion test method* was used to determine the transmission quality.

5. *Testing the experimental equipment*

5.1 *Distortions introduced by the samplers*

In Figure 2, the results of samplers distortion measurements vs. frequency are shown in db referred to full sinusoidal load; such measurements were effected by means of an oscillator having a 0.2 % harmonic distortion (−53 db).

As can be noted in the figure, the distortion introduced by the samplers is so low as to be negligible in comparison with quantization noise even with the highest number of bits.

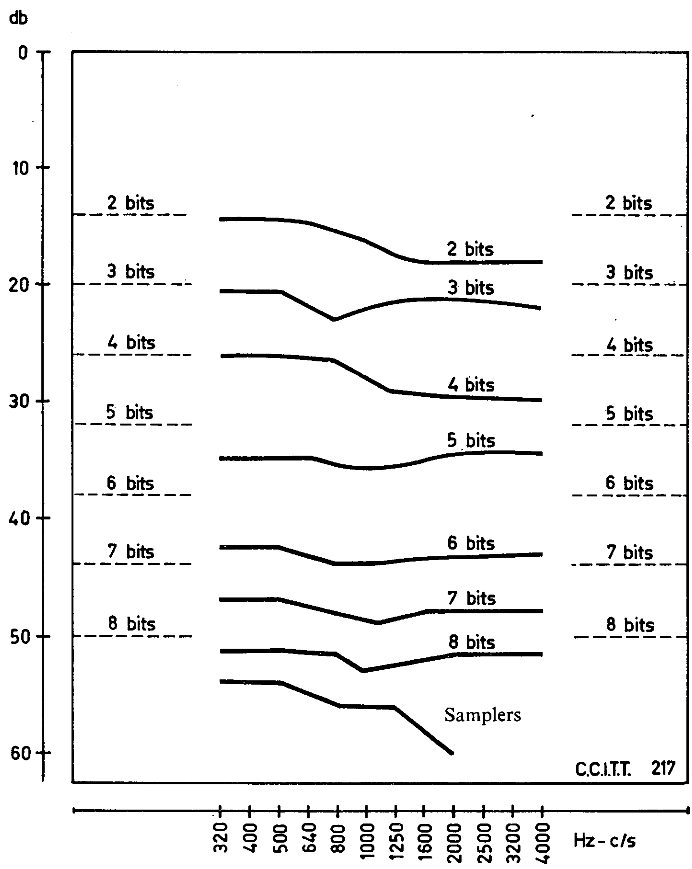

FIGURE 2. — **Over-all level of spurious components at the output of the P.C.M. equipment, for various numbers of bits and of the samplers, as a function of the frequency of the sine wave applied**

5.2 *Quantization noise in case of sinusoidal signals*

The measurements were taken by feeding, at the input of the P.C.M. system without compandor, a full load sinusoidal signal and measuring the S/N ratio by a distortiometer inserted at the output of the receiver channel filter.

The noise levels encountered for the different numbers of bits, referred to the full load sinusoidal voltage, are shown in Figure 2 as a function of frequency. On both sides of the figure the theoretical values of quantization noise are also indicated in dotted lines according to equation (2), which is valid for signals having a constant distribution of probability density of the instantaneous levels. From these measurements it appears that for sinusoidal signals the quantization noise power is always less than the value calculated for complex spectrum signals and depends slightly upon the signal frequency.

Furthermore the average discrepancies between the values given in equation (2) and the measured values vary with the number of bits.

These discrepancies are not to be attributed to an anomalous behaviour of the equipment, but to the fact that the quantization noise level and spectrum depend, although slightly, upon the waveform of the signal. In fact, the theoretical calculations of the quantization noise in case of sinusoidal signals lead to noise levels represented by the following equation:

$$N_q = -(6n + 1.8 + K) \text{ db} \tag{3}$$

where K is a corrective term which increases as the number of bits decreases; furthermore the noise spectrum is not continuous but is a line spectrum. The number of lines increases with the number of bits while the amplitude of each line increases as the number of bits decreases.

The dependence of the noise level on the frequency can be explained by the fact that, as the frequency band is limited within 300 c/s and 3400 c/s, some lines of the noise spectrum may fall either inside or outside the band, according to the signal frequency. This explanation seems to be supported by the greater variation observed for the noise level as a function of signal frequency at the lower numbers of bits.

5.3 *Quantization noise spectra with sinusoidal signals and with complex spectrum signals*

To prove the statements made in the preceding paragraph about the dependence of quantization noise on the frequency in case of sinusoidal signals, the spectra of the noise obtained by feeding either a pure tone or a third-octave band of a complex and uniform spectrum signal at the equipment input were determined.

The separation of the noise from the signal was carried out by third-octave filters. The results of these determinations are shown in Figures 3, 4, 5.

It is easy to see that, for complex signals, the spectrum level of the quantization noise is fairly uniform, if we bear in mind that the slope of the input filter characteristic outside the band is not infinite. On the contrary, in the case of sinusoidal signals the spectrum level of the noise is greatly variable, with energy concentration in some frequency bands, which are more pronounced the lower the number of bits; furthermore the frequency intervals where such concentrations take place are greatly dependent on the signal frequency.

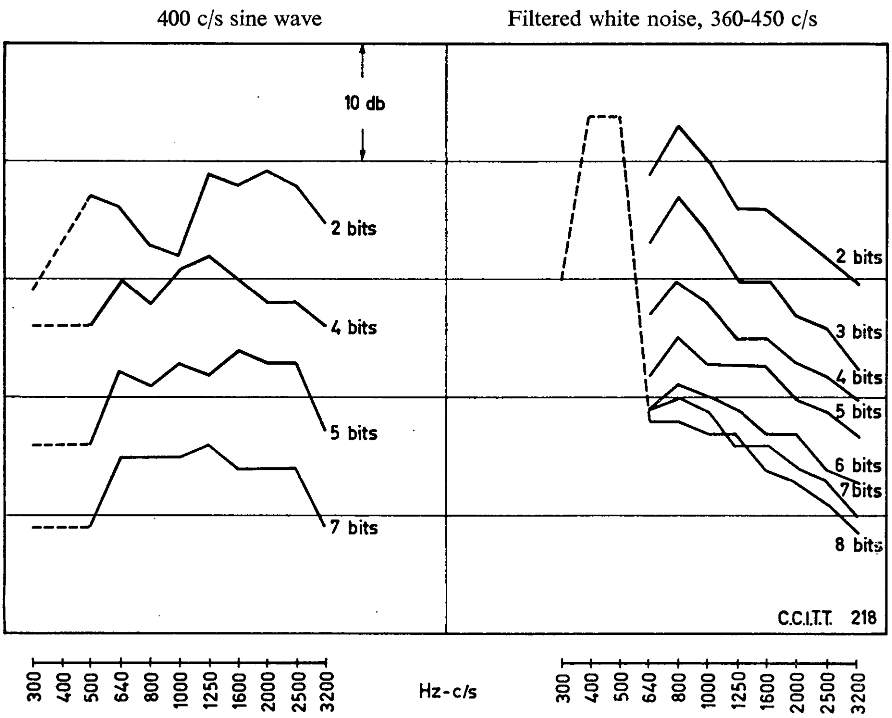

FIGURE 3. — **Spectra of quantizing noise for sinusoidal signals and for uniform continuous spectrum signals in a band of third-octave**

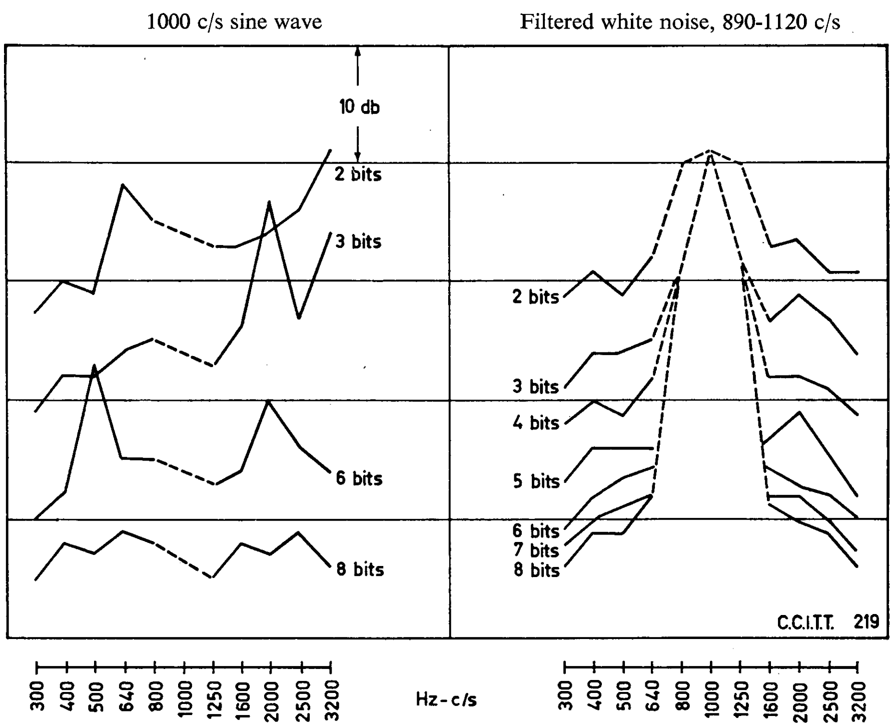

FIGURE 4. — **Spectra of quantizing noise for sinusoidal signals and for uniform continuous spectrum signals in a band of third-octave**

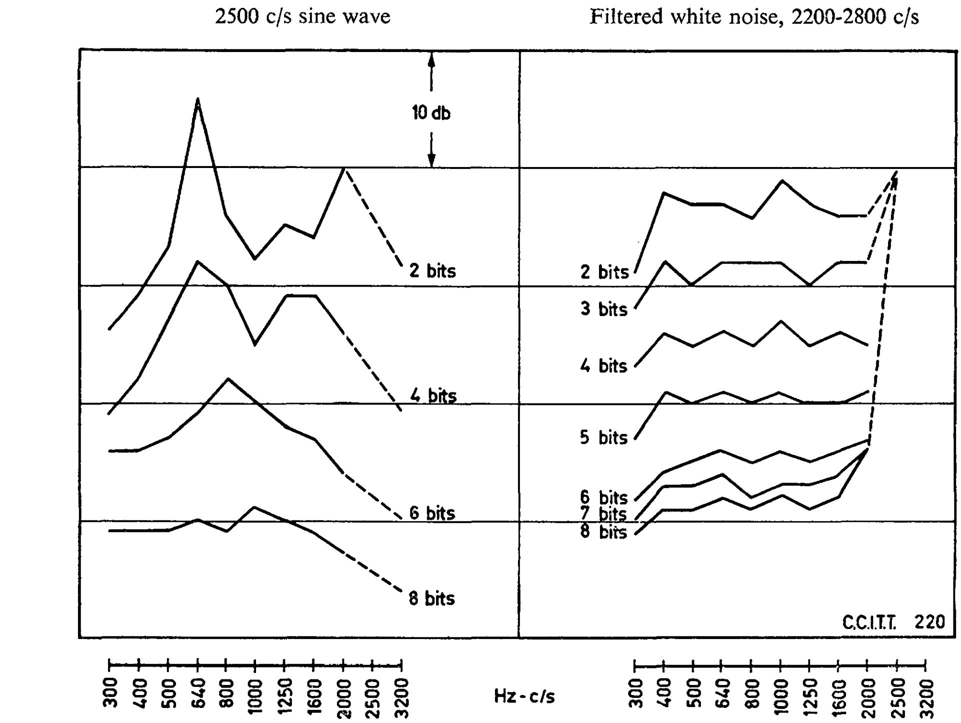

FIGURE 5. — **Spectra of quantizing noise for sinusoidal signals and for uniform continuous spectrum signals in a band of third-octave**

5.4 *Notes on the operation of the instantaneous compandor; transfer characteristic and non-linear distortions of the system used*

An instantaneous dynamic compressor is an amplifier whose gain is a function of the instantaneous signal level. The instantaneous expander, on the other hand, can be considered as an attenuator whose attenuation depends on the signal level, with the same law as the compressor gain.

The transfer characteristic of a compressor must have the maximum slope round the origin and the minimum slope at the highest signal levels. As a consequence the gain will be maximum at minimum levels of the input voltage and will gradually decrease as the latter rises. The transfer characteristic of the expander must be reciprocal to that of the compressor.

The application of a compandor to the terminals of a zero equivalent telephone line can cause the following modifications: if there is no line noise the instantaneous input voltage of the expander will be identical to the corresponding output voltage of the compressor; thus the expander output voltage will be identical to the compressor input voltage. In these conditions the signal undergoes no other modification (in the ideal case where the characteristic of the expander is perfectly reciprocal to that of the compressor).

If, on the contrary, the line noise voltage is not zero and at a certain instant assumes the value ε, the input voltage will be different from that at the signal of the compressor by the quantity ε; consequently the instantaneous expander output voltage will differ from the corresponding compressor input voltage by ε' which is given with a certain approximation by the equation $\varepsilon' = \frac{\varepsilon}{\text{tg}\,\alpha}$, where tg α is the slope of the compressor characteristic at the working point.

The above-mentioned equation can easily be verified by graphical means. In Figure 6 the transfer characteristic of an instantaneous compressor is traced. The horizontal and vertical axes represent the ratio of the instantaneous input and output voltages respectively to their corresponding maximum values. The same curve also represents the expander characteristic, provided the two axes are interchanged.

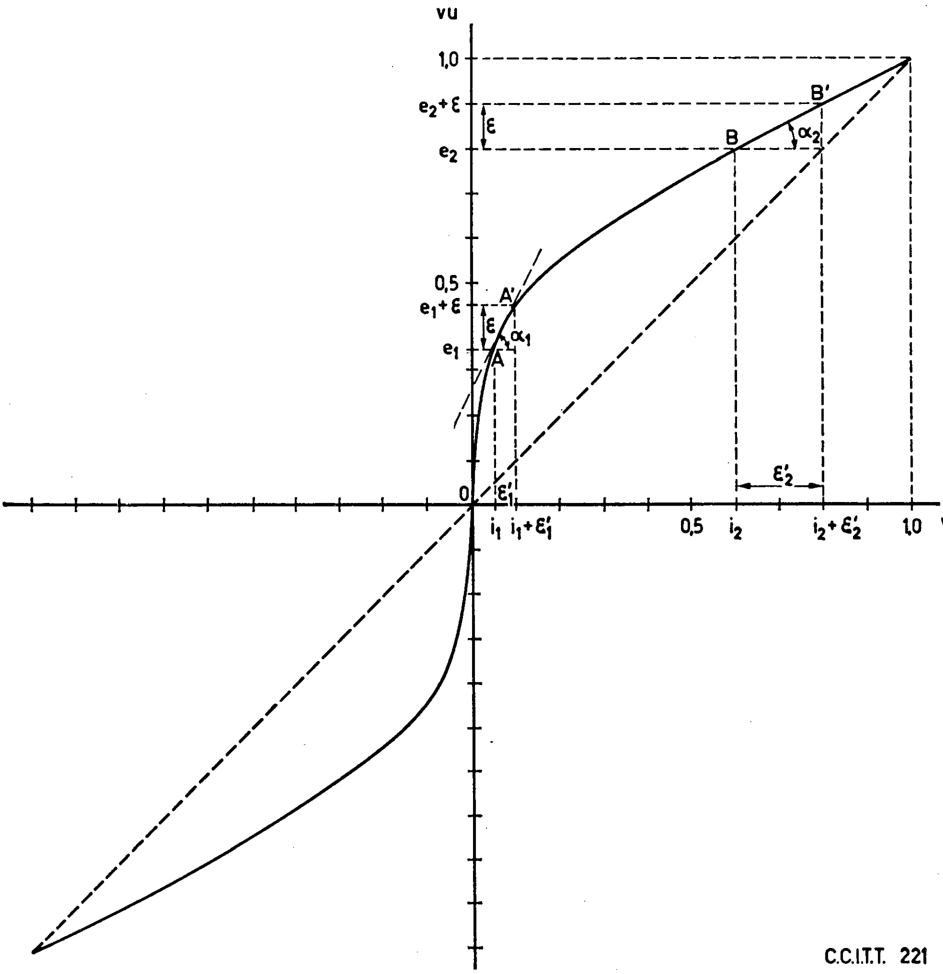

FIGURE 6. — **Graphical computation of the variation of the instantaneous noise voltage as a function of the instantaneous level of the signal in a dynamic compandor system**

Let i_1 be the compressor input voltage at a certain instant, e_1 the corresponding output voltage, A the operating point on the characteristic and tan α the slope at the same point. If the line noise voltage at the same instant is ε_1, the expander input voltage will be $e_1 + \varepsilon$; the operating point on the characteristic will shift to A' and the output voltage will be $i_1 + \varepsilon'$. If ε is so small that the tangent in A can be replaced by the straight line AA' there results $\varepsilon' = \dfrac{\varepsilon}{\text{tg}\,\alpha}$.

The value of tan α is greater than 1 near the origin and less than 1 at the highest voltage levels; therefore the instantaneous noise voltage associated with the expander output signal can be greater or smaller than the corresponding instantaneous voltage of line noise if the portion of characteristic involved relates to high or low signal levels respectively.

The use of the instantaneous compandor therefore results in an improvement of the local signal-to-noise ratio when the signal is weak, i.e. when the signal-to-noise ratio is minimum, while it results in an impairing of the signal-to-noise ratio for strong signals [1]. Consequently it may be concluded that the r.m.s. noise voltage at the expander output can result greater or smaller than the r.m.s. line noise voltage, according to the statistical distribution of the instantaneous signal levels and also independence on the compression curve.

The over-all line noise reduction obtained by a compandor for a given signal, even though this quantity is not the most convenient to represent the improvement of the transmission quality, is useful information which enables us to give a judgment on the efficacy of the system for any type and level of signal and of noise.

The transfer characteristic of the instantaneous compressor used in our tests is shown in Figure 7, while its derivative, i.e., the value of tan α as a function of the instantaneous input voltage, expressed in db referred to the maximum value, is shown in Figure 8.

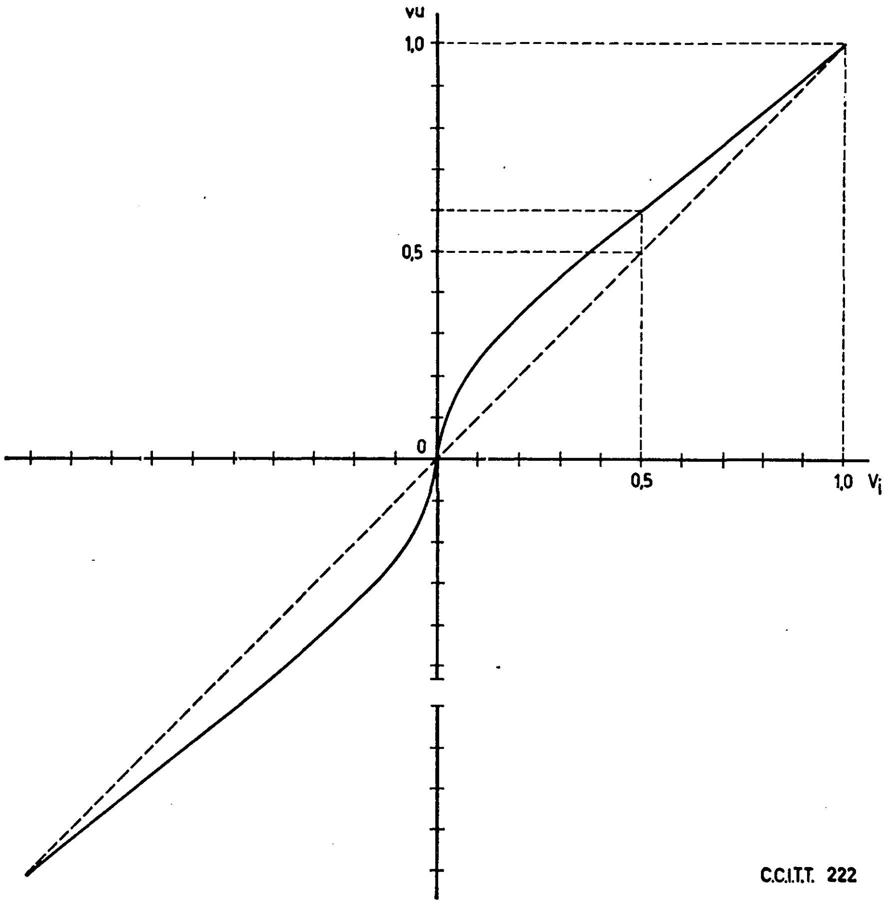

FIGURE 7. — **Input-output characteristic of the compandor**

[1] The variation of the local signal-to-noise ratio at the lowest signal levels is more important in this respect.

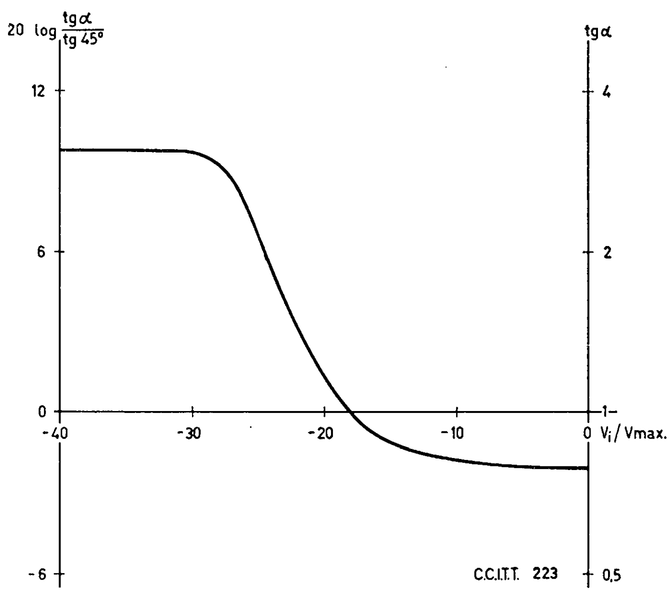

FIGURE 8. — **Variation of the derivation of the characteristic curve of the compressor, i.e. the slope of this curve, as a function of the input level**

The non-linear distortion of the compandor caused by the imperfect reciprocity between the compressor and expander characteristic was 3 % (−30 db).

This prevented us from making objective measurements of the quantization noise with the compandor. For the subjective determinations, on the contrary, there was no excessive disturbance.

6. *Determination of the level of white noise equivalent in loudness to quantization noise*

6.1 *Determination, by balancing method of the white noise equivalent in loudness to quantization noise in a P.M.C. equipment without compandor*

The signal used for this determination was made up of tape recordings of news items and technical papers read by various speakers with a natural and constant volume in an anochois room. To obtain the same speech volume in the various tests, a 800-c/s signal having a level 3 db above the speech level, read on the vu meter, was recorded on the same tape before each news item. The level at the equipment input was adjusted according to the criteria stated in section 3.

With the above-mentioned level, determinations of the noise equivalent to quantization noise with 2 to 6 bits were made. With 7 and 8 bits, the quantization noise level was too low and the operators did not succeed in evaluating it. For those two determinations, therefore, it was necessary to attenuate the signal at the input by 20 db. This can be done, because (as already stated in section 1) the quantization noise amplitude is not influenced by the signal level, at least until the signal amplitude is such as to cover a certain number of quantization stops.

The circuit used for the balancing tests is shown in block diagram in Figure 9.

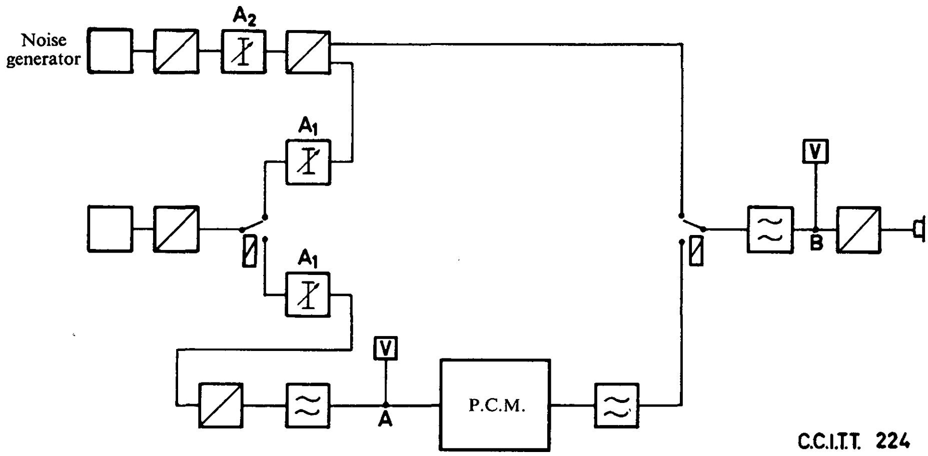

FIGURE 9. — **Circuit arrangement used in the balancing tests and in the tests on intelligibility threshold**

The results obtained are shown in Table I.

TABLE I

Levels of white noise equivalent to the quantization noise vs. number of bits, referred to full sinusoidal load

Number of bits	2	3	4	5	6	7	8
Level (db)	−14	−20	−26	−32	−38	−44	−50

These data show that the two circuits were judged equivalent, as far as noise loudness is concerned, when the level of the white noise fed into the reference line was equal to the level of the quantization noise as given by equation (2).

The value of the noise level which gave the equivalence can be reliably evaluated by the analysis of the operators' answers. As an example, Figure 10 shows the percentages of preference in favour of the reference line, with respect to the line including the P.C.M. equipment operated at 5 bits, plotted vs. the level of the injected white noise.

The difference between the white noise level which gave 50% of the preferences " both for the reference and the P.C.M. lines " and the theoretical value of the corresponding quantization noise level was, for all determinations, less than ±0.5 db.

The operators, in giving their preference, concentrated their attention exclusively on the noise loudness, therefore the equivalence of the two circuits means only that the noise loudness is equal in the two circuits. Since the quantization noise spectrum is quite similar to the spectrum of the white noise used for the comparison and since the comparison itself was effected by using the same earphone throughout, it can be stated that, when the two circuits were judged equivalent, they were affected by noises of equal level.

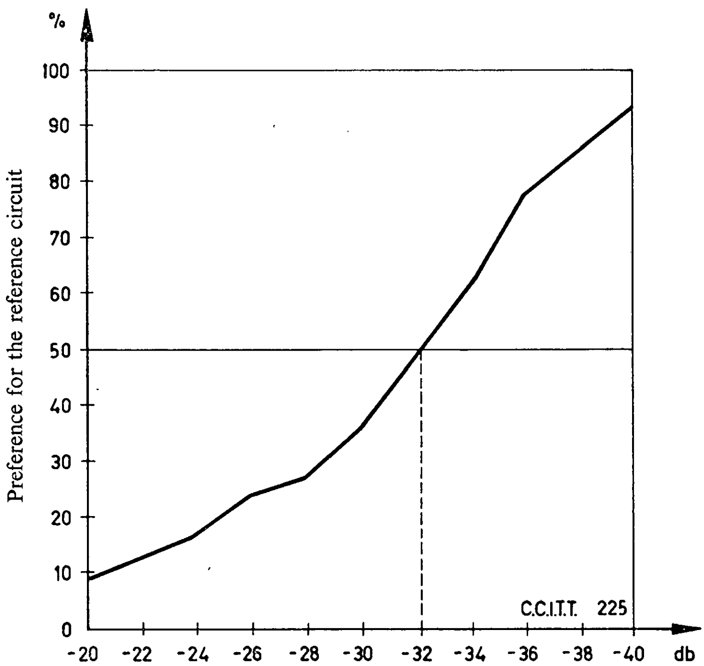

FIGURE 10. — **Preference in % for the reference circuit, as a function of injected white noise, compared with a P.C.M. system operating at 5 bits**

It is therefore confirmed that, in the case of speech signals, the quantization noise power in a P.C.M. system making use of a n-digit binary code and uniform quantization, can be expressed in db, in reference to the full sinusoidal load, by equation (2).

The results of the balancing tests, which at first sight could seem obvious, have, on the contrary, an important significance because they point out that the very particular characteristic of the quantization noise of being present only when the signal is present has no influence on the subjective evaluation of the noise loudness.

6.2 *Determination by the balancing method, of the white noise equivalent to the quantization noise in a P.C.M. equipment making use of an instantaneous compandor*

These determinations were also made by the balancing method, using the same circuit and the same recordings used in the determinations of the quantization noise without the compandor. The level setting was effected before the tests according to procedures established in section 3.

The results are listed in Table II. In the same table, the results obtained without the compandor are also indicated.

From the data in Table II we note that the use of the compandor having a compression characteristic as indicated in Figure 7 had, for the bit numbers 2 to 6 a decrease of 6 db in the quantization noise power in comparison to the values obtained without compression. For 7 and 8 bits the equivalent noise level found was -44 db, that is, identical to that found for 6 bits. This fact can

TABLE II

Levels of white noise equivalent to the quantization noise vs. the number of bits and referred to full sinusoidal load, with and without compandor

Number of bits	2	3	4	5	6	7	8
With compandor	−20	−26	−32	−38	−44	−44	−44
Without compandor	−14	−20	−26	−32	−38	−44	−50

be attributed to the distortions caused by imperfections in the compandor whose effect in the case of a time-division system is more severe than in case of harmonic distortion of equal value. In fact, as the signal is on the pulse carrier at 8000 c/s all of the distortion components whose frequency is higher than 4000 c/s are in the band, with a spectral distribution which depends on the frequencies associated with the pulse carrier and on the harmonics of the signal, i.e. they lose the harmonic correlation with the signal and appear as a true noise associated with the signal.

Determinations of equivalent noise with a signal level attenuated by 20 db at the compandor input, effected with the P.C.M. equipment operated with 5 and 8 bits, gave the following results: for 5 bits a noise level of −40 db instead of the −38 db found with the normal level of signal; for 8 bits a level of −48 db instead of −44 db.

This reduction of noise encountered by lowering the signal amplitude at the input to the compandor can be attributed both to the smaller value of the distortion components, and to the greater rate of utilization of the high-slope portion of the compression characteristic.

6.3 *Determination, by the threshold of intelligibility, of the masking effect due to the quantization noise*

The application of the threshold method to a circuit which includes a P.C.M. equipment is based on the independence of the quantization noise from the signal amplitude. The method consists in determining the lowest signal level that still permits an acceptable sentence intelligibility: this signal level is called the " threshold of intelligibility in the presence of noise ". The threshold of intelligibility in the presence of noise is a function solely of the difference between the spectral levels of the signal and the noise, when the noise levels are always higher than the levels of the threshold of detectability in the whole frequency band considered. In this case there exists a relationship of simple proportionality between threshold of intelligibility and noise level.

The circuit used in this test is illustrated in block diagram in Figure 9. It has permitted the determination of the threshold of intelligibility both on the P.C.M. line and on a normal line affected by white noise of equal loudness to the quantization noise. The equivalent between points A and B was equal to zero in both circuits. The setting of the levels was achieved during the calibration procedure according to what has been stated in section 3. The adjusting of the signal level to reach the threshold level was achieved by calibrated attenuators A_1. The adjusting of the noise fed into the normal line was achieved by attenuator A_2.

Before presenting the results of the determination of the threshold it is necessary to clarify some points in the functioning of the P.C.M. coder. The coder can accept signals of only one polarity; therefore the alternating signal at the input must be added to a d.c. bias voltage E_p, which brings the signal's zero level into the quantization interval located at the centre of the dynamic which can be reproduced by the coder itself. Therefore the value of the bias voltage must be such that:

$$q\,(2^{n-1} - 1) < E_p < q \cdot 2^{n-1}$$

i.e. it must be equal to a number of quantization steps given by $E_p/q = 2^{n-1} - \varepsilon/q$, where q is the amplitude of the quantization step, n is the number of the bits and ε is a number between zero and q. For $\varepsilon = 0$ the level of the bias voltage is situated at the upper extreme of the central quantization step while for $\varepsilon \cong q$ this level is closer to the lower extreme.

We shall see in the course of this section the great importance of the bias voltage setting on the transmission quality of the quantized signals. A first series of determinations of the threshold of intelligibility was carried out by setting the d.c. bias to a value very near to the upper extreme of the central quantization step ($\varepsilon = 0$). The results of these measurements have shown that:

a) the level of the threshold of intelligibility was the same both with the P.C.M. line and with the normal line affected by a white noise having a level equal to the quantization noise level associated with the number of bits used on the other line;

b) a 1-bit decrease corresponded to a 6-db increase in the level of threshold of intelligibility;

c) the " granular effect " on the P.C.M. line was barely noticeable and independent of the number of bits.

A second series of determinations was made by setting the level of the bias voltage in the middle of the central quantization step ($\varepsilon = q/2$). These measurements were made using the P.C.M. equipment set for functioning only with 7, 5 and 3 bits; the following results were obtained:

a) the levels of the threshold of intelligibility determined on the P.C.M. line, set for functioning with 7 and 5 bits were 2 db above those obtained with the same number of bits when the bias voltage was set near the upper extreme of the central quantization step; with the equipment set for 3 bits, the increase in the threshold level was 6 db.

b) the " granular effect " was fairly evident, and it appeared the more pronounced the lower the number of bits.

A possible explanation for the difference in the results of the two series of determinations could be the following: it is well known that the speech signal peaks themselves contribute in an irrelevant manner to the intelligibility. In order that the content of information of a signal received is such as to ensure the threshold of intelligibility, it is necessary for the system to be capable of distinctly transmitting signal levels situated at least 10-12 db below the level of the signal peak itself.

In the case of a P.C.M. equipment, the sensitivity of the coder, i.e. the minimum amplitude of the signal that can be transmitted, is given by the difference between the value of the bias voltage and that of the voltage corresponding to the nearest of the two levels determining the central quantization step; therefore ε represents the working threshold of the equipment and can have all of the values between zero and $q/2$. When ε is lower than the value $q/2 \cdot \sqrt{3}$ of the quantization noise r.m.s. voltage, it is the noise which fixes the lower limit to the signal dynamic. Therefore to obtain the threshold of intelligibility it is necessary for the signal peaks to be greater by 10-12 db than the quantization noise voltage, i.e., the peak voltages of the order of a quantum are necessary. When, on the other hand, ε is greater than $q/2 \cdot \sqrt{3}$ it is the working threshold voltage that fixes the dynamic and therefore the intelligibility. To obtain the threshold of intelligibility with $\varepsilon = q/2$, it is necessary for the signal peak voltages to be less than two quanta that is, a signal level of 5-6 db more than what is needed in the case of ε less than $q/2 \cdot \sqrt{3}$ is necessary.

In these determinations the stability and precision of the d.c. voltage setting was within an 8-bit step; consequently the condition $\varepsilon = q/2$ was reached with fair approximation only att he lower numbers of bits. This explains the increase of only 2 db encountered in the threshold of intelligibility with 7 and 5 bits instead of the calculated 5 or 6 db.

The " granular effect " is a result of the fact that, whenever the amplitude of the input signal is smaller than ε, no signal is received. This gives the impression of a temporary interruption in the transmission. The duration and the frequency of these interruptions in the case of a speech signal depend upon the ratio of the signal peak value to ε. The " granular effect " will therefore be greater the smaller the signal and the larger ε.

On the basis of these results, the following conclusions may be drawn:

1) The " granular effect " is due to the existence of a lower working threshold of the coder. In the case of speech signals this effect will be more detectable the smaller the ratio of the signal peak values to the value of the working threshold.
2) When the value of the working threshold is very small, the quantization noise intensity can still be evaluated, in the case of speech signals, with equation (2), even when the signal level is so low as to cover a limited number of quantization steps (2 or 3 steps).
3) When the value of the working threshold surpasses the r.m.s. value of the noise voltage, the signal masking due to the centre clipping brought about by the working threshold prevails upon the masking due to noise. The shifting of the threshold of intelligibility in the case of a uniform binary code cannot exceed 5-6 db.

7. *Determinations by the users' opinion test method of the transmission quality in a link consisting of a P.C.M. equipment and orthophonic transducers*

The users' opinion test consists in having a group of non-professional persons talk, by means of the link under study, and asking them separately for their judgment on the transmission quality. For our determinations two groups of 6 persons in all the possible combinations of two were used separately. Thus we were able to have 60 judgments on each listening post for every circuit condition. The scale of judgments was in 5 steps: Bad (B), Poor (P), Fair (F), Good (G), Excellent (E). The transmission quality was determined by the mean opinion score calculated by attributing weight zero to judgment B, weight 1 to judgment P, weight 2 to judgment F, weight 3 to judgment G and weight 4 to judgment E. The circuit used for the determinations is illustrated in block diagrams (Fig. 11).

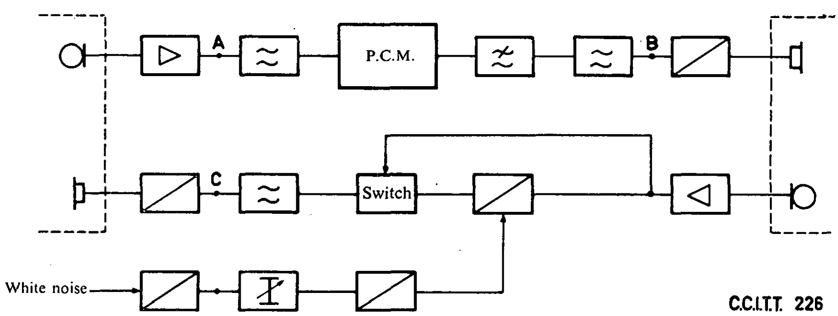

FIGURE 11. — **Circuit arrangement used in determining transmission performance**

It was not possible to make an identical connection in both directions (according to common practice) because the experimental equipment available was of unidirectional type. To make up for this difficulty the link in the opposite direction was made by means of a conventional line affected by a white noise and equipped with a voice-controlled electronic switch, which interrupted the current whenever the signal level was less than an assigned limit, in order to simulate the " granular effect " of the P.C.M. line. The switch consisted of a gate circuit, turned on when the input signal exceeded a given value. The microphones used, of a cardioid type, were placed about 40 cm. from the speaker's mouth. In these conditions, the voltage observed at their terminals was nearly 3 mV in a reasonably reverberating room.

The earphones were two S.T.C. type 4026 A. The frequency response of the links was fairly uniform within the telephonic band.

The levels were fixed, according to what has been established in section 3, by feeding at the microphone amplifier input a 3 mV 800-c/s voltage, and adjusting the amplifier gains to have: 100 mV at points A, B, C of the two circuits, the full sinusoidal load at the coder input and 15 mV at the terminals of the two earphones. The d.c. bias voltage in the P.C.M. equipment was adjusted each time so that it would fall, as far as possible, in the middle of the central quantization step corresponding to each number of bits. The minimum signal level capable of operating the electronic switch was adjusted, by means of a direct comparison, to give the signal the same " granular effect " as on the P.C.M. line.

The white noise fed into the conventional line was adjusted in such a manner as to have a loudness equal to that of the P.C.M. line. The equivalence of the noise on the two directions was obtained when the white noise fed into the line equipped with the electronic switch was 6-8 db lower than the quantization noise on the opposite direction. This fact can easily be explained if we think that the cutting-off of the minimum signal levels operated by the switch is equivalent to a noise added to the signal. The disturbances present on the two channels of the link were judged almost identical, so that the mean opinion scores for the two directions fairly coincided. The conversation posts were set up in two sound-insulated rooms of about 50 m^3 with reverberation time of 0.4-0.5 s.

The average speech level in these tests was about 3 db higher than the full sinusoidal load.

The determinations of the transmission quality were made in various test shifts for each value of the number of bits from 2 to 8, both with the compandor and without; the same determinations on a link made up of two identical conventional lines affected by white noise and using the same transducers were effected.

Each test shift also included a quality measurement on a reference circuit with constant characteristics, with the purpose of making comparable the results obtained with different circuit conditions and in different shifts.

In Figure 12 we find the curves showing the mean opinion scores obtained for the three types of links, that is: curves A and B indicate the average judgment vs. the numbers of bits relative to the links equipped with P.C.M. without and with compandor, respectively; curve C represents the mean opinion score vs. the white noise level in the conventional link.

To complete the evaluation of the transmission quality that can be obtained with a P.C.M. system we deem it useful to add that the " granular effect " in the link without compandor was still evident with the equipment operated at 8 bits and it increased as the number of bits decreased; in the link with the compandor on the contrary the " granular effect " was observed only for

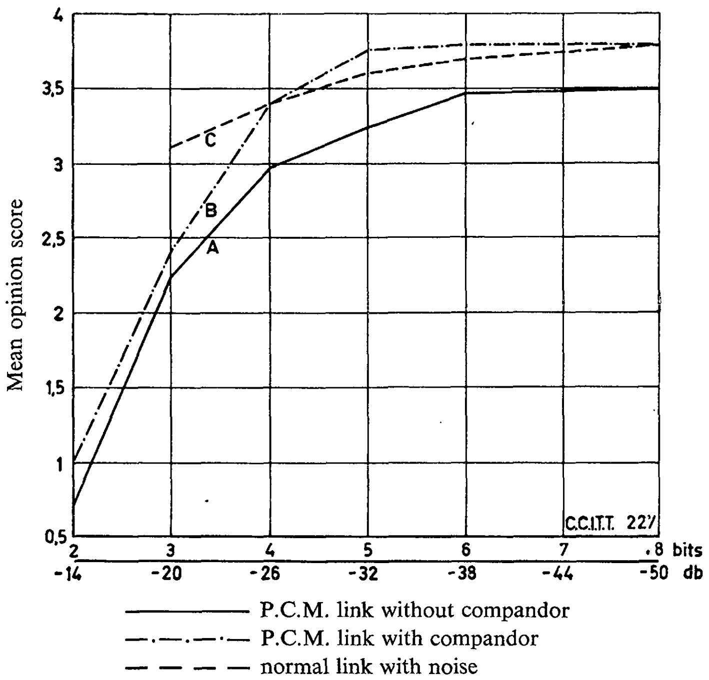

FIGURE 12. — **Results of determination of performance by opinion tests on links made up of P.C.M. lines, with and without compandors, and on a normal line affected with a white noise**

numbers of bits ranging from 2 to 4 and it could not be observed for a higher number of bits. The lower mean opinion score attributed to the link without compandor with respect to the one with the compandor was justified by the operators with the greater disturbance caused by the " granular effect ".

8. *Conclusions*

The determinations effected with the balancing method have confirmed that formula (2) is suitable for evaluating the quantization noise level in the case of speech signals. The determinations made with the method of threshold of intelligibility have emphasized the importance of the centre clipping caused by the existence of a minimum working threshold in the device; such threshold is linked to the quantum amplitude and to the d.c. bias voltage setting within the central quantization step. In fact, when the bias was adjusted so as to have a very small working threshold the results showed that the impairing of intelligibility due to quantization was equal to that due to the quantization noise level.

When the bias was adjusted at the centre of the quantization step, the intelligibility was worse than that depending on the noise level and appeared instead to depend upon the level of the working threshold.

From the results of the determination of transmission quality effected by the users' opinion test method it seems appropriate to conclude that the working threshold is probably the factor affecting the transmission quality more than any others both on account of the influence on the intelligibility and because of the " granular effect " it gives the signal.

The use of compandor systems improves the transmission quality because it reduces the noise level, lowers the working threshold level and greatly attenuates the disturbance due to the " granular effect " even with very high numbers of bits. From all of the tests carried out it may be concluded that a P.C.M. system functioning at 5 bits and equipped with an instantaneous logarithmic compandor, whose variable amplification ranges from zero db at the maximum signal levels up to 12 db at the minimum signal level transmitted, is capable of ensuring a good transmission quality when the speech signal volume is near the full sinusoidal load level. As the average volume of the speech signal can undergo a variation of level of 25-30 db, the compandor should be capable of maintaining the signal-to-noise ratio constant within that field. In fact the compandor used gave a signal-to-noise ratio fairly constant for a level variation of an input signal of 15 to 20 db. To ensure a good transmission quality to the great majority of users with the above-mentioned compression rate, a seven-bit code is necessary. With these characteristics, the transmission quality will be good for most of the users even if along the link two or three P.C.M. systems in cascade could be present. This obviously implies that a slight decrease in quality is to be tolerated for a minimum percentage of the users, subject to particularly severe transmission loss.

ANNEX 8

(to Question 27/XII)

Tests on a pulse code modulation system

(Contribution by the French Administration)

For these tests, the French Administration used a prototype pulse code modulation system in which the telephone channels presented the following characteristics:

— sample rate: 8000 per second,

— 6-unit code (1 unit representing the sign and the 5 others the absolute value of the amplitude of the sample, in pure binary code),

— logarithmic compression according to the Smith law[1] with two possible values for the parameter μ:

$$\mu = 36, \quad \text{or} \quad \mu = 100$$

Intelligibility and opinion tests were carried out with this equipment.

1. *Intelligibility tests*

The pulse code modulation (P.C.M.) system was locally looped, transmission to reception. The telephone circuit established for the tests was one-way as shown in Figure 1.

[1] Cf. *Bell System Technical Journal*, volume 36, May 1957, page 667.

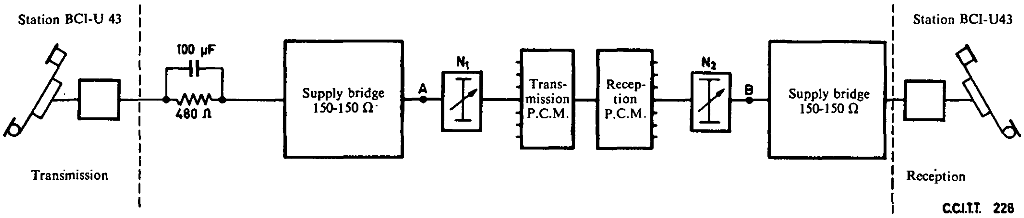

FIGURE 1. — **One-way telephone circuit for intelligibility tests**

The reference equivalent of the subscriber system was equal to:

+0.90 N for sending

−0.22 N for receiving

Two adjustable attenuation lines were inserted in the chain, one before and the other after the P.C.M. system. The sum of both attenuations thus introduced was constant and equal to 4 N. Given the gain of 1.9 N caused by the P.C.M. system, the reference equivalent of the circuit A B was 2.1 N.

The reference equivalent of the call was thus 2.78 N.

The intelligibility tests were made by silent listening, without the addition of Hoth noise, at various cut-off levels (manipulating the attenuations N_1 and N_2) and both values of the parameter μ.

For comparison purposes, a test was carried out in which the P.C.M. system was replaced by a 300-3400 c/s bandpass filter and the equivalent of the call was adjusted to 2.78 N.

Table I summarizes the results of these tests. To interpret them correctly, it should be noted that the root mean square voltage of the speech signals measured across 600 ohms at the output of the subscriber system was between 70 and 100 mV and that the peak-to-peak voltage at the same point varied from 1 to 1.7 V.

TABLE I

Intelligibility tests

Circuit	Cut-off level (in Nm) at the output of the subscriber system	Articulation (%)	
		Logatoms	Sounds
U43 — U43 300-3400 c/s filter	—	93	97.5
U43 — U43	−2.1	86	95
	−1.4	91.5	97
P.C.M.	0	93	97.5
	+1.4	89.5	96
μ = 36	+2.1	85	95
U43 — U43	−1.4	93	97.5
	−0.7	91.5	97
	0	93	97.5
P.C.M.	+0.7	93	97.5
	+1.4	92.5	97.5
	+2,1	88	96
μ = 100	+2.8	83.5	94

It may be deduced from the results in Table I that the peak-clipping of the signals becomes troublesome, so far as intelligibility is concerned, when the cut-off level at the system output falls below −1.4 Nm, the average level of the signals at this point being approximately −2 Nm.

The quantizing noise becomes an appreciable nuisance when the cut-off level at the same point exceeds 0 Nm for $\mu = 36$ or 1.4 Nm for $\mu = 100$.

2. *Opinion tests*

The two-way telephone circuit used for these tests is shown in Figure 2. The equivalent of the circuit AB could be adjusted to 2.1 or 3.5 N, giving a reference equivalent for the call to 2.78 N or 4.18 N.

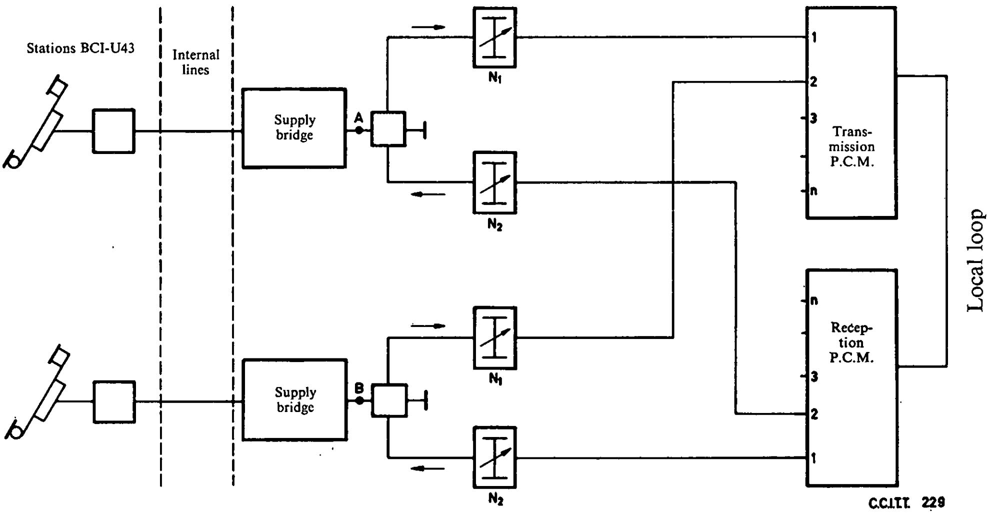

FIGURE 2. — **Two-way telephone circuit for opinion tests**

In all the opinion tests, the logarithmic compressor of the P.C.M. system followed the Smith law in which $\mu = 100$. By manipulating the attenuation lines N_1 and N_2, it was possible to obtain 5 different cut-off levels at the output of the subscriber system. As in the intelligibility tests, a sixth condition was provided for comparison purposes, which consisted in replacing the P.C.M. system by a 300-3400 c/s bandpass filter and adjusting the reference equivalent to 2.78 or 4.18 N.

One hundred people took part in the tests and made the complete series of 12 (2 values of the equivalent; 6 circuit conditions). Each test lasted about one minute during which two speakers conversed freely. At the end of that time, each participant was asked to give his opinion on the quality of the circuit in accordance with the usual scale of marking:

excellent circuit	4
good circuit	3
fairly good circuit	2
mediocre circuit	1
bad circuit	0

The 12 tests took place in variable and random order.

These results are summarized in Table II and shown graphically in Figure 3.

TABLE II

Opinion tests

Circuit	Cut-off level (in Nm) at the output of the subscriber system	Assessment of quality	
		Average	Standard deviation
U43 — U43 300-3400 c/s filter E = 2.78 N		3.51	0.64
U43 — U43 P.C.M. μ = 100 E = 2.78 N	−1.4 −0.7 +0.7 +2.1 +2.8	2.60 3.28 3.16 2.47 1.89	0.79 0.65 0.71 0.96 1.03
U43 — U43 300-3400 c/s filter E = 4.18 N		2.29	0.86
U43 — U43 P.C.M. μ = 100 E = 4.18 N	−1.4 −0.7 +0.7 +2.1 +2.8	1.27 1.84 2.08 1.85 1.51	0.81 0.85 0.87 0.83 0.99

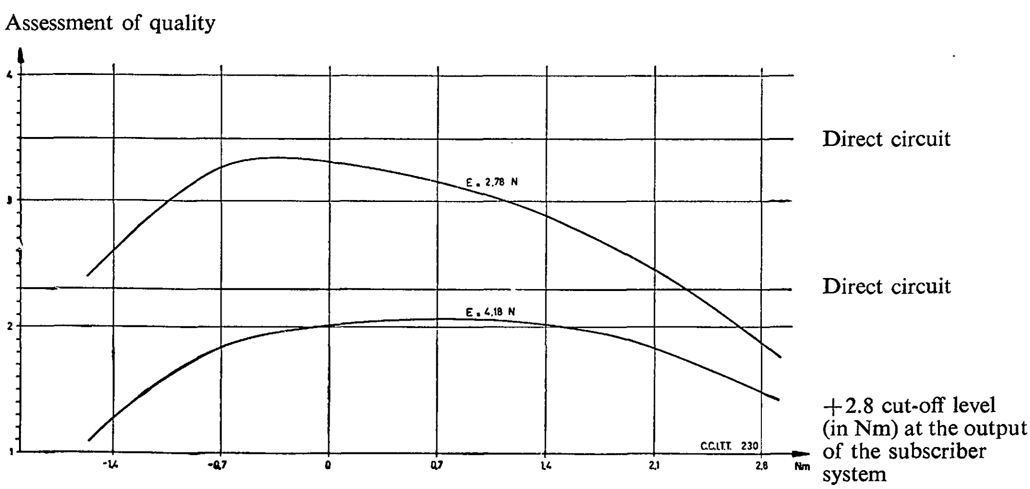

FIGURE 3

It will be noted that, in this case, the deterioriation in quality caused by peak-clipping of the speech signals appears earlier (cut-off level approximately −0.7 Nm at the output of the subscriber system). This phenomenon may be connected with the fact that the speech level is distinctly higher in free conversation than during intelligibility tests.

Quantizing noise, on the other hand, has a less noticeable effect in calls with a high reference equivalent than in calls whose equivalent is distinctly below the limit.

Question 28/XII — A statistical study of the implications of Spanish phonetics for telecommunication systems

(Question Latin America No. 5 from the Plan Sub-Committee for Latin America)

(new question)

A statistical study of the implications of Spanish phonetics for telecommunication systems.

Summary of Questions allocated to Study Group XII in 1964-1968

Question No.	Short title	Remarks
1/XII	National system reference equivalents in the new transmission plan	Of interest to S.G. XVI
2/XII	Measurement and limits of reference equivalent for sidetone	
3/XII	Measurement of the disturbing effect of clicks	
4/XII	Effect of circuit noise on transmission performance	
5/XII	Accuracy of reference equivalents by subjective methods	
6/XII	Users' tolerance of echo and propagation time	
7/XII	Determination of transmission quality by objective measurement	
8/XII	Measuring the efficiency of a microphone or a receiver	
9/XII	Limits applied in national trunk and local networks	
10/XII	Increase in the efficiency of local systems	
11/XII	Statistical methods of checking subjective tests	
12/XII	Artificial voices, mouths and ears	
13/XII	Non-linear distortion of telephone apparatus	
14/XII	Premises and equipment for the C.C.I.T.T. Laboratory	
15/XII	Measuring indices based on sound power	
16/XII	Effect of compandors on transmission quality	Reply to be sent to S.G. XVI
17/XII	Loudspeaker telephones	
18/XII	Calculating the reference equivalent of a subscriber's line	
19/XII	Impedance variations in subscribers' lines and telephone sets	
20/XII	Synthetic speech and frequency compression systems	
21/XII	Systems using pulses or alternating currents modulated by microphone	
22/XII	Intelligibility of crosstalk in programme transmission	
23/XII	Objective measurements of the intelligibility of crosstalk	
24/XII	Extension of the bandwidth transmitted	Linked with Question 1/XV
25/XII	Maintenance of subscribers' sets	Question Africa H
26/XII	Transmission of performance carrier for circuits on very short distances	Linked with Question 32/XV
27/XII	Transmission performance of pulse code modulation systems	Linked with Question 33/XV
28/XII	A statistical study of the implications of Spanish phonetics for telecommunication systems	Question Latin America 5

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