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LOCAL SECTIONS AND THE CONVENTION

In any international or even national organization there are few opportunities for bringing together the majority of members. Even Annual General Meetings are usually only attended by representatives of Sections; this is understandable since the business involved does not normally include technical issues which constitute the main reason for membership of a professional institution.

It has become generally acknowledged, therefore, that Conventions, if properly timed, are a most useful adjunct to local meetings. Members who have attended previous Brit.I.R.E. Conventions know that one of the most valuable aspects of such gatherings is the opportunity to exchange experience and ideas on a wider scale than is possible at local Section meetings.

The subject of this year's Convention—Industrial Electronics—will attract members from widely different parts of the country and from overseas, since it will deal with the applications of the electronic art to many other industries.

The interests of many of the local Sections in industrial electronics, as distinct from communication engineering, may be seen from their programmes of meetings over the past few years. Whilst the general trend has been to meet the main interests of local members, certain Sections have already found that papers on industrial electronics, which cater particularly for the interests of members employed in other industries, have also a very wide appeal to members engaged in the communication field. In particular, the West Midland Section held meetings of this nature which attracted considerable interest among the heavy industries in the "Black Country" area.

The success of such meetings throughout the Sections was one of the factors which determined the choice of theme for the Institution's third post-war Convention. It also initiates the formation of the first specialized group within the Institution.*

Such a policy implements the views expressed by Mr. W. E. Miller in his Presidential Address on the importance of considering the particular requirements of members in relation to their local Section activities. Indeed, Mr. Miller made special emphasis of the importance of papers on the techniques and application of electronics. The majority of papers now being considered by the Convention Committee for inclusion in the programme fall into this category rather than that of fundamental research.

Response to publication of the arrangement of sessions of the Convention has been most favourable. Final selection should result in a minimum of 30 papers being presented; it is not intended, however, that the time of the Convention shall be taken up wholly by presentation of the papers. Stress will be laid on providing adequate time and facilities for discussion and demonstration. It is believed that in this way all attending will derive the maximum benefit from a Convention of this character.

During the course of the Convention a special meeting of local Section representatives will be arranged in order to give the fullest opportunity for an open discussion on the Institution's work and problems, with particular reference to the development of the Sections, both at home and overseas.

Local Committees are invited to submit items for the agenda of this meeting. The Council is anxious to frame an agenda suitable for free debate and of common interest to *all* the Sections. This new addition to the programme of Conventions will, it is believed, subsequently be of great advantage to the entire membership.

G. D. C.

* "Annual Report 1952-3." *J.Brit.I.R.E.*, **13**, September 1953, p. 422.

NOTICES

Obituary

The Council of the Institution has expressed its sympathy with the relatives of the following member:

John Auspitz, advice of whose death on August 24th last has only just been received, was elected a Companion of the Institution in 1950. Mr. Auspitz qualified in Law from Temple University Law School, Philadelphia, and was executive director and joint owner of the Radio Electronics Institute of Philadelphia. He was 48 years of age.

Radio Component Exhibitions

This year's Radio Component Show, arranged by the Radio and Electronic Component Manufacturers' Federation, will take place from Tuesday, April 6th to Thursday, April 8th, inclusive. As in previous years, the show will be at Grosvenor House, Park Lane, London, W.1, and the hours of opening will be: Tuesday 10-6, Wednesday 10-9, Thursday 10-5. Over 130 exhibitors are taking part and the show will be officially opened by Sir Robert Renwick, Bt., K.B.E. (Member.)

The 1954 Salon National de la Piece Detachee Radio will be held at the Parc des Expositions, Porte de Versailles, Paris 15, from March 12th-16th. Some 200 exhibitors are participating in the exhibition, which normally attracts over 50,000 visitors. During the period of the exhibition a series of technical meetings will be held at which papers will be presented on new techniques and trends in the French component industry. Further information may be obtained from S.N.I.R., 23 Rue de Lubeck, Paris 16.

Royal Commission on Awards to Inventors

At a public hearing of the Commission recently the Chairman of the Royal Commission on Awards to Inventors, Lord Cohen, announced that the Commission anticipated completing the hearing of claims by the end of the current year.

Unanswered Questions

The work of the Department of Scientific and Industrial Research in providing information for industry was referred to in the July issue of the *Journal*. Most of this service is provided by the Technical Information and Documents Unit who are able to answer the majority of questions from industry.

In "Unanswered Questions," the bulletin published by T.I.D.U., specific questions are given to which no answer can be found from usual sources, and among recent questions which are relevant to the field of interest of members of the Institution are the following:

(1) When using electro-acoustic instruments for locating schools of fish it is very important to be able to adjust the trawl to the correct depth. A method is required for measuring the depth of trawls and the following characteristics are desirable:

(*a*) Accuracy must be better than 1 per cent., i.e. the depth must be measured to 1 fathom over the range 0-150.

(*b*) The measuring device must be insensitive to temperature variations and to the length of trawl warp.

(*c*) The meter or recorder must be ship-proof. It must withstand tilting in any direction to 30 deg with a period of 5 sec, and must be adequately gimballed. In addition it must withstand acceleration of 0·3 *g* in any direction without appreciably affecting the reading.

(*d*) The whole apparatus must be small in size and operated by one unskilled man.

(*e*) The device should not interfere with normal fishing procedure.

(2) A suitable cement is required for fixing a strain gauge to a tensile test piece at 500°C.

Members able to supply any information at all in answer to these problems are invited to write to the Chairman of the Technical Committee of the Institution, giving, if possible, the source of the information provided.

Back Copies of *Electronics*

The Library files of the American journal *Electronics* for 1952 and 1953 are not complete. Members and others who have the following issues for disposal are invited to write to the Librarian: August, 1952; September, 1952; October, 1952; November, 1952; December, 1952; January, 1953.

Correction

In the report of the Discussion Meeting, "Education in the Radio and Electronics Industry," which was published in the January *Journal*, there is a reference at the top of page 47 (col. 2, line 7) to "the Final Technological Certificate level." This should, of course, read "the Full Technological Certificate level."

SOME FACTORS IN THE ENGINEERING DESIGN OF V.H.F. MULTICHANNEL TELEPHONE EQUIPMENT*

by

W. T. Brown†

Read before the Institution in London on January 6th, 1954. *Chairman: Mr. P. Adorian (Past-President).*

SUMMARY

The principal clauses of a specification for a v.h.f. multichannel telephone equipment are discussed, and methods of approach to these problems are suggested. Particular attention is given to ancillary equipment such as automatic change-over systems, remote control, and fault warning and location facilities. The general method of construction is described together with a brief description of power supply problems and suitable aerial arrangements.

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1. Introduction

Experience over a period of years indicates that the development of a "standard" production multichannel system is indeed a very formidable task. For one thing, the job such a system has to do depends very largely on where it is situated in the world. If, for example, the system was located in Britain, then the probability of having a great distance to cover would be fairly remote. On the other hand, countries like Africa, where great distances are commonplace, may require circuits covering 300–400 miles. The engineering of a system, therefore, will depend on where it is most likely to be used. Further, the facilities required often depend to a considerable extent on the conditions pertaining in the actual locality in which the equipment will be used. The degree of reliability required of any particular circuit is often governed by political and/or financial consideration, defence circuits demanding as little breakdown time as possible, whilst equipment used for civilian purposes may not require such a high degree of serviceability.

In short, from the sales point of view, it is very difficult to estimate just what will be required for a world market. There are, however, certain requirements which are common to all multichannel radio circuits, and it is proposed to deal with these in the order which they are considered to be important.

Throughout this paper only one type of modulation will be discussed, namely, frequency

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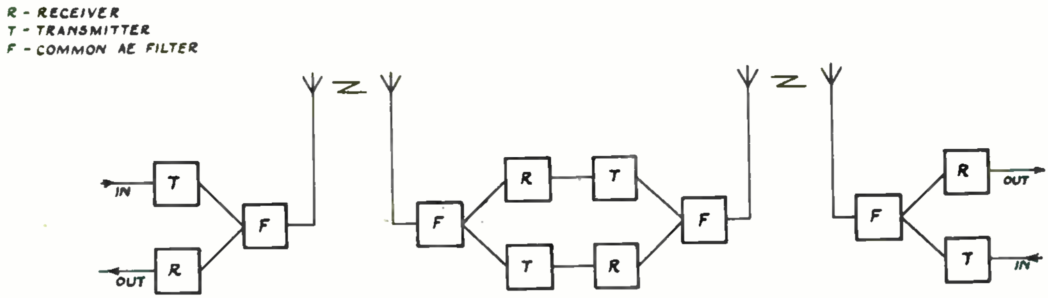
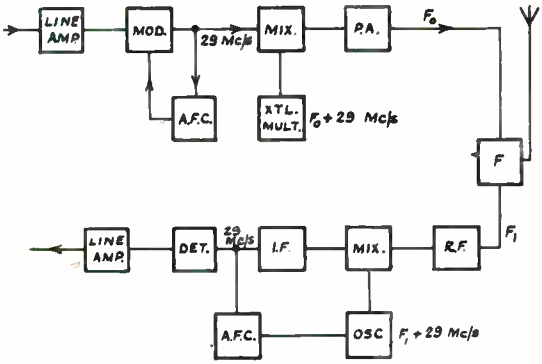
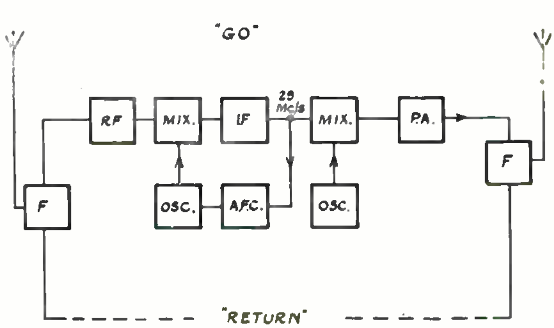
U.D.C. No. 621.396.41.029.62.



(*a*) *General arrangement.*



(*b*) *Block schematic—Terminal station.*



(*c*) *Block schematic—Repeater station.*

Fig. 1.—V.h.f. multichannel telephone equipment terminal and repeater stations.

modulation. Many types of modulation are available to the designer of a multichannel system, but the pros and cons of the selection of any particular method is outside the scope of this paper.

In a discussion of this length, it would be impossible to deal with all the ramifications, permutations and combinations of all the parameters associated with multichannel radio telephony. It is the purpose of this paper, therefore, to give only the general approach to the problems involved, quoting where necessary idealized cases as an illustration of the application of the fundamental principles.

2. Signal/Noise Ratio

The C.C.I.F.* lays down certain recommendations regarding international transmission systems and a v.h.f. multichannel system normally comes under these recommendations since the circuit may form part of a national network which in turn may be part of an international network.

By its very nature, a radio path is subject to continual change in attenuation, which may or may not be rapid over a short period or which may be seasonally gradual. It would obviously be uneconomic to build a system which would give a high signal/noise ratio under all circumstances of fading. A compromise has been reached inasmuch as it is agreed that when a figure ratio is quoted it implies the signal/noise ratio will be equal to or better than this for all but a small percentage of the operating time. The recommendation of the 3émé Commission d'études du C.C.I.F. is that in a band 300–3,400 c/s at a zero level point the weighted noise power should not exceed 10,000 micro-microwatts (pico-watts) for more than 1 per cent. of the busy

* International Telephone Consultative Committee.

hour (this being for a 1,600-mile typical circuit).

The assumption is made that 20 per cent. of the total noise power is generated at the terminals and 8,000 pico-watts of noise can be introduced in the transmission path. Since the noise power is proportional to the length of path, the maximum noise (weighted) which can be tolerated in any section of the circuit may be estimated as follows:

Assuming a total length of radio path (terminal to terminal) = 180 miles

Noise power "allocation" to this path

$$= \frac{180 \times 8{,}000}{1{,}600} = 900 \text{ pw.}$$

$$= 60{\cdot}4 \text{ db (weighted).}$$

In a v.h.f. system the only significant "noise power" implies noise from two sources:

(*a*) Thermal noise of the valves and circuits of the first stage of the receiver.

(*b*) Spurious signals caused by intermodulation products due to the modulation of the carrier and caused by non-linearity of the transmission equipment.

No mention has been made of "man-made" noise because this is of an entirely random nature and depends on many factors, such as location of the site, height and direction of aerials, etc. The degree to which any system rejects impulsive type of interference also depends on factors such as bandwidth, limiter type, discriminator type, etc.

It is normally assumed that noise from thermal sources should be balanced against the noise from non-linearity sources. This immediately sets the signal to noise (thermal) and signal to noise (intermodulation) at 63 db for each. The noise degradation against numbers of repeaters is shown in Table 1 below. This is for the ideal case where the path attenuation is the same for all cases. For any given link of "n" repeaters, there are $n + 1$ receivers, the noise power adds arithmetically and is equal to $10 \log (n + 1)$.

Table 1

No. of Repeaters	0	1	2	3	4	5	6	7	8	9
Degradation in db	0	3·0	4·8	6·0	7·0	7·8	8·4	9·0	9·5	10

In the case under consideration, a 5-repeater station circuit, the signal to thermal noise ratio per individual repeater section would require to be $63 + 7{\cdot}8 = 70{\cdot}8$ db.

It can be shown (see Appendix) that the signal to noise ratio weighted of a single channel in a multichannel f.m. system is

$$168 + 20 \log M - (A + K + F) + (P + G) \qquad \text{.............(1)}$$

where A = path attenuation (db)

P = transmitter power (db above 1 W)

F = noise factor of receiver (db)

G = total aerial gain db

K = other losses, feeders, filters, etc. (in db)

M = Mod. index $\frac{\Delta f}{f'}$, (Δf = channel deviation, f' = channel mid-frequency). (Channel deviation is that produced by a (1 mW) test tone injected at a point of zero level.)

(Below frequencies of f' = 12 kc/s the formula tends to become inaccurate.)

From the above expressions it is therefore possible to estimate the magnitude of the parameters under design control. As an example, it is assumed:—

$F = 7$ db; $K = 2$ db $+ 2$ db; $A = 124$ db

$G = 8 + 8$ db (aerial at each end 8 db)

$\Delta f = 30$ kc/s; $f' = 30$ kc/s

Hence

$$71 = 168 + 20 \log 1 - (124 + 4 + 7) + (P + 16)$$

$$\therefore P = +22 \text{ dbw} = 158 \text{ watts.}$$

The figures shown against the symbols are average values which might very well be found in practice.

Since the noise voltage in an f.m. system is triangulated, i.e. the thermal noise voltage is proportional to the channel frequency, the noise in the top channel (usually 156 kc/s for a 36-channel system) will be worse than the 30 kc/s channel assumed above by 20 log 156/30 db = 14 db. The transmitter power will therefore require to be 22 + 14, i.e. + 36 dbw = 4,000 W to give a signal/noise ratio of 70 db in the 156 kc/s channel.

It would obviously be unrealistic to attempt to make a transmitter of this power, particularly for unattended repeater station operation where the only power supplies available are diesel-electric generators. In actual practice a power of about 50 W (+ 17 dbw) appears to be a reasonable compromise between available valves, mains power consumption and capital cost of the equipment. This represents a 19 db (36 db − 17 db) drop in power, and therefore an equivalent fall in signal/noise ratio. This may be compensated by

(*a*) Improving the aerial gains by 8 db at each end, giving a total of 16 db per aerial.

(*b*) By increasing the channel deviation by 3 db, i.e. to 42 kc/s.

It should be borne in mind that an increase of aerial gain is valueless unless the bandwidth of the aerial is maintained, and in order to achieve this condition the cost of the aerial will increase more rapidly than if only pure gain is considered.

In the case of (*a*) the economic limit appears to be about 16 db per aerial, and it is therefore a better proposition to settle for 16 db aerial gain and increase the deviation by 3 db, rather than try to achieve the total amount of gain required by increasing the aerial height and/or transmitter size. Owing to the increase in deviation the distortion or cross-talk will increase, and if this is intolerable the phase linearity must be improved accordingly.

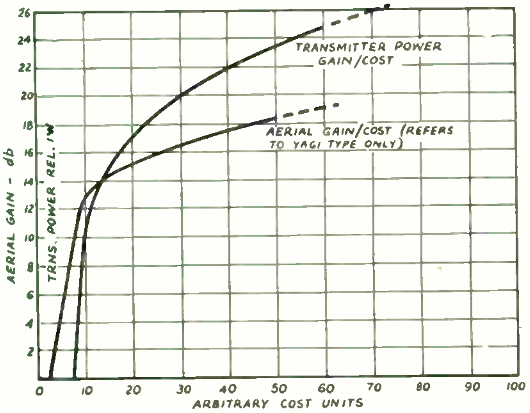


Fig. 2.—Cost comparison between aerial gain and transmitter power.

In Fig. 2 are given smooth curves showing aerial gain *versus* cost and transmitter power gain *versus* cost. It should be noted that about 16 db aerial gain represents the optimum economic value. Further increases in required gain are therefore better obtained by increasing the transmitter power. It should be pointed out that these graphs assume the following:

(*a*) That no increase in the size of the mast is required as the aerial gain is increased.

(*b*) Frequency of operation 175 Mc/s approximately.

(*c*) That the cost of valves for the transmitter is proportional to the power output taking the Mullard QQVO6-40 and QY3-125 valves as reference levels.

Throughout the discussion, it has been assumed that the first stage of the receiver contributes the bulk of the noise power measured at the output. The signal/noise ratio formula above assumes that the limiting action is perfect and it makes a useful guide to the efficiency of a receiver to measure the noise levels in different carrier channels and plot the result against r.f. input to the receiver.

In the same way the estimated values are plotted and comparison of the curves arrived at provides an overall check of the receiver, from noise factor to limiting efficiency. In a properly designed receiver the theoretical and practical results should be within 1 to 2 db.

Figure 3 shows typical curves illustrating the method of display. The solid lines are the calculated values whilst the broken curves are the measured values. On the equipment from which these curves were derived, considerable attention has been given to the limiter stages. It should be noted that at low level r.f. input, the limiting action breaks quite suddenly and the noise output therefore increases very rapidly with small changes of input. On the other hand, at high r.f. input the noise output of the measuring equipment and post-detector circuits tends to be of the same order as the noise from the apparatus under test, the curves tend therefore to converge to a point of no further change for increase of signal input.

When the limiters in an f.m. receiver are fully operated the level of the modulated output is independent of change in r.f. input within wide limits. The term "fully limited at x μV" is a vague but common inclusion in specifications for f.m. systems. It is therefore advisable to be certain what is actually meant. A convenient method is to specify the r.f. input at which a test tone changes its level by say ± 0·5 db.

Figure 4 is a curve of test tone output level plotted against r.f. input. Point A indicates the

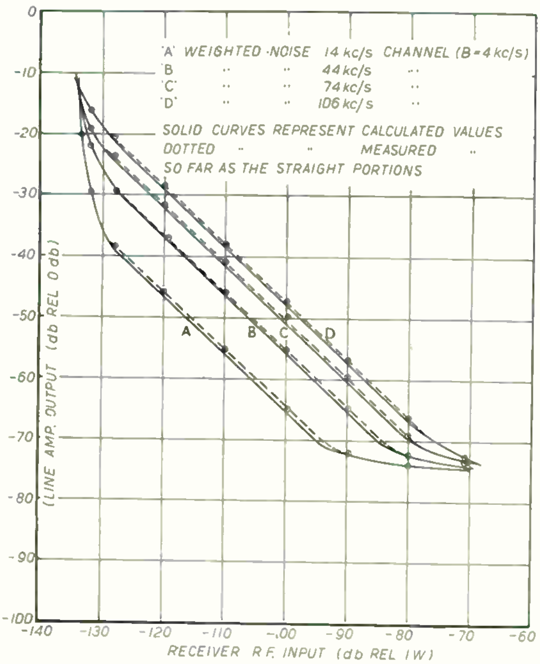


Fig. 3.—Thermal noise levels in f.m. receiver.

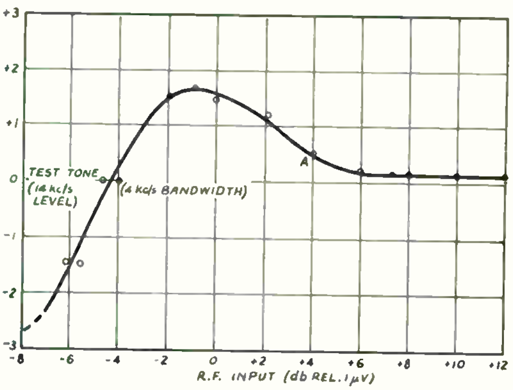


Fig. 4.—Measured limiter characteristics at low r.f. input.

+ 0·5 db point which represents a signal input of about 1·5 µV. It should be noted in passing that this curve represents the characteristic of one particular type of limiter, and is not necessarily representative of limiters in general. A knowledge of this part of the receiver characteristic is of value in assessing the performance of the system under poor signal conditions.

3. Intermodulation Noise (Crosstalk)

In the type of transmission at present being discussed, it was stated earlier that one source of noise could be caused by the intermodulation products or crosstalk. This phenomenon arises as a result of non-linearity in the system. There are three possible sources for this,

(i) the modulator } variation of amplitude
(ii) the demodulator } frequency response.
(iii) variation of group delay with frequency (non-linearity of the r.f. and i.f. phase characteristics, including reflections due to the aerial and feeder.

Non-linearity at any of the three sources above manifests itself as a production of harmonics of the modulating tone, the most important of these being the second and third order, and if two voltages of frequencies f_1 and f_2 are injected into the same non-linear element, combinations such as $f_1 - f_2$ and $2f_1 - f_2$ appear, these being the 2nd and 3rd orders. The crosstalk noise is a summation of these and higher order products.

3.1. *Distortion Measurement*

The measurement of intermodulation products is a complicated technique, and at the best, only gives a partial answer. The direct measurement of second and third harmonics is limited to the lower end of the frequency spectrum. This is because the third harmonic of a test tone at the top end of the modulation spectrum would normally fall outside the modulation spectrum and be therefore unobtainable. The usual method of measurement is to apply two equ l sinusoidal tones at a known level such that the sum voltage of the amplitudes of $f_1 + f_2$ results in peak deviation. The level of the frequency $(f_1 - f_2)$ or $(f_1 + f_2)$ can then be taken to represent the second harmonic figure and $(2f_1 - f_2)$ the third harmonic content. It is normally convenient to measure the peak voltage applied to the modulator and the harmonic ratio is then specified as relative to the peak deviation. The actual channel crosstalk may be obtained theoretically from the knowledge of the 2nd and 3rd harmonic powers.

Another, and possibly better way of measuring crosstalk is to apply a spectrum of noise and to band stop various channels. The noise voltage produced in these particular channels may then be considered directly as crosstalk power. If, say, a 24-channel equipment is available, then it is only necessary to apply a noise voltage at a suitable level to 23 channels and measure the noise power from the 24th channel. This method assumes, of course, that the noise generated by the channelling equipment itself is of a very low order. This latter method has the advantage that it is a direct measurement whereas the harmonic method has to be converted to crosstalk and this is open to conflicting arguments as to the validity of the result. The noise method appears to be the most satisfactory way of measurement, in spite of the filtering cost and equipment difficulties, but is more awkward to specify in terms of the catalogue performance.

The two-tone measurement on the other hand lends itself to easy specification, but its practical application is open to argument.

When speech is applied to a system the deviation is obviously going to change continuously, and only when the speech voltages add up for short periods on a random basis will excessive deviations occur. This condition is covered by the C.C.I.F. recommendations for noise power in terms of time percentages of the busy hour, e.g., -63 db for 1 per cent. busy hour. In the case of harmonic distortion due to modulator/demodulator characteristics, the resulting crosstalk noise is distributed fairly uniformly over the modulating spectrum. However, in the case of distortion due to the r.f. and i.f. phase response the crosstalk noise is distributed, in general, in a triangular way similar to the thermal noise distribution. One of the chief problems is the separation of the main sources of distortion from the measurement point of view. The modulator and demodulator produce amplitude distortion of the modulating signal, whilst the r.f. and i.f. circuits produce phase distortion of the modulated signal. The combination of the two may add on a power basis or depending on the type of circuits in use, may tend to cancel.

It is good practice when designing a multichannel radio system to make the modulator and demodulator as good as possible, even if extra expense may be incurred at this point. If the distortion from the modulator and demodulator is comparable with distortion accruing from the r.f. circuits considerable confusion may exist in eliminating the type of distortion and its source. Further, the modulator and demodulator are, or should be, stable, and not subject to the vagaries of tuning and attention by the maintenance personnel. On the other hand, if the modulator and demodulator are such that their respective distortions are low by comparison with that from the r.f. and i.f. stages, comparatively simple tests enable the crosstalk to be determined. In the following part of this discussion it is assumed that the great proportion of the intermodulation products occur in the r.f. and i.f. sections of the equipment. In fact, in the v.h.f. band using several repeater stations this must necessarily be so.

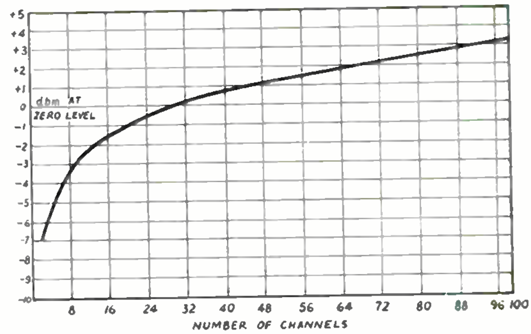


*Fig. 5.—Speech power exceeded for 1 per cent. of the time.**

It should be borne in mind that an exceptionally good signal thermal noise ratio tends to be incompatible with a low crosstalk figure, and that the channel deviation is the crucial parameter. This being so, it is obviously necessary to work back to the channel deviation from some starting point, and the obvious starting point is the required specification. It is assumed that the figure for intermodulation noise should be -63 db, and it can be shown that, in general, the intermodulation noise due to delay or phase distortion in an f.m. system may be expressed in the form

$$N_2 = H_2 + 2P + A_2 \text{ for 2nd order distortion} \quad \ldots\ldots\ldots\ldots(2)$$

$$N_3 = H_3 + 3P + A_3 \text{ for 3rd order distortion} \quad \ldots\ldots\ldots\ldots(3)$$

(This arises from the fact that the intermodulation noise power is proportional to the 2nd harmonic ratio and also proportional to the

* Figures 5 and 6 are due to B. D. Holbrook and J. T. Dixon ("Load rating theory for multichannel amplifiers." *Bell Syst. Tech. J.*, 18, October, 1939, pp. 624-644).

square of the signal power in the case of the 2nd harmonic. In the case of the 3rd harmonic, the intermodulation noise power is proportional to the 3rd harmonic ratio and also proportional to the cube of the signal power. This statement expressed in decibels takes the form of equations 2 and 3.)

where N_2 = 2nd order intermodulation noise unweighted in dbm at zero level point.

N_3 = 3rd order intermodulation noise unweighted in dbm at zero level point.

P = total speech power in multichannel systems exceeded for a given percentage of the time (say 1 per cent.). Expressed in dbm at zero level points.

A_2 and A_3 = constants dependent on the number of channels, channel frequencies, etc. (see Table 2).*

H_2 = level of the 2nd order products $f_1 - f_2$ or $f_1 + f_2$ and is measured in db relative to the level produced by peak deviation.
(The sum of the amplitudes of $f_1 + f_2$ being equal to the peak deviation.)

H_3 = level of the 3rd order products $2f_1 - f_2$ and is measured in db relative to the level produced by peak deviation.
(The sum of the amplitudes of $f_1 + f_2$ being equal to the peak deviation.)

* The constants A_2 and A_3 are derived from an unpublished work by J. L. Slow of B.T.R. Ltd.

For the purpose of the argument it is at first assumed that the measured 2nd and 3rd harmonic intermodulation powers are equal and would therefore be 66 db each in order to achieve a total of 63 db.

By rearrangement of (2)

$$H_2 = N_2 - 2P - A_2$$

From Fig. 5, P (for, say, 24 channels) = −0·5 db
Table 2, A_2 (for, say, 24 channels) = −18·2
N_2 (from specification) = −66 db

H_2 is therefore equal to −66 + 1 + 18·2 = 46·8 db.

The practical method would be to carry out a two-tone test on the equipment, increasing the deviation until the amplitude of the sum

Table 2

No. of Channels	Freq. of Channels kc/s	Peak Factor $= 20 \log_{10} \left(\frac{\text{Pk. Devn.}}{\text{Ch. Devn.}}\right)$ db	Test Tone Intermod. Product Freq. f_c ($f_1 + f_2$ or $f_1 - f_2$ for 2nd order; $2f_1 - f_2$ or $2f_1 + f_2$ for 3rd order)	A_2 db	A_3 db
			kc/s		
12	12–60	13·3	50	−15·1	−17·5
12	60–108	13·3	90	No 2nd order cross-talk	−17·7
24	12–108	14·2	90	−18·2	−21·8
36	12–156	14·7	90	−17·2	−21·4
60	60–300	15·2	200	−21·8	−26·8

Note 1.—The two tones which combine to form the product f_c should be chosen such that their average value $(f_1 + f_2)/2$ falls approximately in the middle of the band occupied by the channels.

Note 2.—The values of A_2 and A_3 are for the particular intermod. frequencies f_c quoted. If it is more convenient to use some other value of intermod. product f_c then A_2 and A_3 must be corrected by *adding* a factor $-20 \log_{10} f_c/f_{c'}$.

$(f_1 + f_2)$ is increased to give this figure. For example, it is assumed that the measured deviation is found to be, say, 200 kc/s.

Similarly by rearrangement of (3):

$$H_3 = N_3 - 3P - A_3$$

The values for N_3, P and A_3 are determined as before, and hence H_3 calculated, but if in practice the deviation caused by $(f_1 + f_2)$ to give a calculated figure of 42 db is now, say, 300 kc/s, then obviously the 3rd harmonic distortion is lower than the 2nd and has little effect on the total crosstalk noise. This means that the deviation so far as 2nd is concerned could be increased. This procedure is continued until a reasonable compromise figure for deviation is obtained which will give the total crosstalk power required. If it is assumed that the final figure arrived at is, say, 230 kc/s, it only remains to determine the channel deviation. This can be obtained from Fig. 6 in terms of the ratio between peak deviation and channel

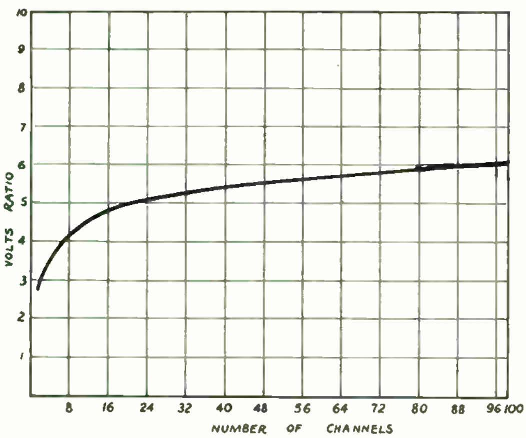


Fig. 6.—Multichannel peak voltage/single channel test tone peak voltage equivalent to peak deviation/single channel in f.m. system.

deviation. In the example above this value is 5·1 times, and therefore the channel deviation is 230/5·1 = 45 kc/s.

This figure may now be looked at from the signal/noise aspect and any modifications to the overall system carried out accordingly. Depending on which parameter is more or less important for any particular application so the other parameters may be adjusted accordingly.

It is pointed out that the whole subject of distortion in multichannel systems is long and complicated and that the above is merely intended to give guidance and to arrive at the order of the operating conditions, and is in no way intended as a rigorous treatment. Experience, however, has shown that it is reasonable to expect an accuracy of about $\pm 2 - 3$ db by the method described.

4. Performance Compromises

Figure 1 illustrates in block schematic form a typical v.h.f. radio link with one repeater station. The actual distance between either terminal and repeater, or repeater to repeater, may vary from 15 to 70 miles, depending on many factors, such as terrain covered, transmitter power, aerial gain, etc.

The number of repeater stations which it is possible to use in tandem is almost entirely dependent on the signal/noise ratio and the permissible non-linear distortion. In general, it may be said that in the v.h.f. band about six repeater stations is about the limit before fairly extensive measures have to be taken to improve the signal to thermal noise ratio and the crosstalk.

Like most aspects of radio engineering, the final result is largely governed by compromise and/or economics. For example, the signal/noise ratio may be improved by:—

(i) Increasing the transmitter power.

(ii) Increasing the aerial gain.

(iii) Improving the noise factor of the receiver.

(iv) Increasing the channel deviation.

(v) Reduction of transmission losses.

(ii) above increases the capital cost of the system only, whilst (i) increases not only the capital cost but the running cost as well. The latter may be recovered by increasing the toll charges over a long period or may be paid for by capital investment. However the financial aspect is considered, a higher-grade circuit in terms of signal to thermal noise so far as (i) and (ii) are concerned costs money.

On the other hand, development costs on the receiver to improve the noise factor performance are mitigated by the fact that the manufacturer of the equipment will spread this cost over his whole production and the customer only pays a fraction on each receiver he buys. However, there is a theoretical limit to the noise factor in the present state of the art, and when this is approached improvement must be sought in other directions.

By increasing the channel deviation, (iv) above, an improved signal/thermal noise ratio will be obtained at theoretically no monetary cost. The price is, however, paid in terms of increased crosstalk, and it is here that the designer of a multichannel system must make his most careful compromise.

If, however, the increase in crosstalk caused by an increase in deviation is intolerable, it may be reduced by

(i) Improving the frequency/amplitude characteristics of the system.

(ii) Improving the deviation/amplitude characteristics of the system.

(iii) Increasing the overall effective bandwidth.

(iv) Reducing the non-uniformity of the group delay/frequency characteristics of the r.f., i.f. and aerial feeder sections.

Factors (i) and (ii) fall largely in the same category as receiver noise factor, i.e., one of design technique and development costs.

Factors (iii) and (iv) may not only require a considerable amount of development to compromise with other conflicting factors, but are liable to cost the customer money as well, in equipment, phase correcting networks, and wide band sharp cut-off i.f. amplifiers, etc., which might not have been necessary if the deviation had not been increased to improve the signal/noise ratio.

In short, a very precise and realistic specification of the performance of a multichannel system is a vital requisite if the system is to be designed for an optimum performance to cost ratio.

5. Bandwidth

Experience has shown that under present conditions one of the most important factors is bandwidth, and a subtle compromise must be made in this direction. The i.f. bandwidth is considered to be of first importance because of its ability to reject adjacent channel and local site interference, and should be therefore no wider than is absolutely necessary; if this concept is accepted then the i.f. phase characteristic should tend to be the determining factor so far as intermodulation is concerned.

It is theoretically possible to design an i.f. amplifier in such a way as to achieve a good compromise between very selective circuits and good phase characteristics. The practical fulfilment is unlikely, owing to the necessity for maintaining stable circuits and the requirement of field maintenance.

The nominal bandwidth of an f.m. radio system may be shown to be given approximately by

$$B = 2(\Delta f_0 + f)$$

where Δf_0 = peak deviation.

f = highest modulation frequency.

Owing to various errors, such as the alignment of i.f. amplifiers, r.f. stages, frequency instability of tuned circuits, etc., it is necessary to allow a margin of safety which in general may be taken as 1·6 times.

The nominal bandwidth B is increased therefore to:—

$$3{\cdot}2(\Delta f_0 + f) \qquad (4)$$

The bandwidth given by (4) may be taken as a starting point. Distortion, adjacent channel selectivity, spurious interference and stability tests will show, however, under the particular conditions the system is required to operate, whether the bandwidth will require to be modified.

As a result of modern conditions, the bandwidths of equipment operating in the region 30–200 Mc/s should be kept to the absolute minimum.

6. Spurious Interference

The amount of interference from spurious sources and its cause and remedy depends very largely on the system and the circuits used in each case. It would therefore be impossible to cover all the aspects of interference but a few brief notes on one or two possible sources may be of interest.

6.1. *R.F. Intermodulation*

In one particular case the aerial filter characteristic produced amplitude modulation of the local transmitter signal arriving at the grid of the r.f. stage. A deviation of ±200 kc/s produced about 10 per cent. amplitude modulation. This unwanted signal after detection by the r.f. stage caused the input capacitance of the valve to be modulated at the transmitter modulating frequency. The varying capacity in turn caused phase modulation of the required signal. The net effect manifested itself as an increase in channel noise in the channel of frequency equal to the modulating frequency.

6.2. *R.F. Interference*

In a particular type of equipment incorporating a 2nd i.f. of 1 Mc/s spurious beats were noticeable with a send/receive frequency separation of 5 Mc/s. Difference frequencies of 4–6 Mc/s were produced and beat with harmonics of a counter type demodulator. The resultant frequency should theoretically be zero but owing to slight discrepancies in alignment occasionally appeared in the multichannel spectrum.

6.3. *R.F. Amplifier Blocking*

High level unwanted signals from, say, the local transmitter, at the grid of the r.f. amplifier may tend to block the grid of the first r.f. stage. If the grid circuit resistance is high, then the current produced will bias the valve back and so reduce the gain offered to the required signal.

6.4. *"Looping" Interference at Repeater Stations*

A more serious form of interference may arise at a "non-demodulating" type of repeater station, and it may be of interest to discuss this in some detail. The figures below refer to equipment in the early stages of development.

Figure 7 shows diagrammatically the path of wanted and unwanted signals at a repeater station having common aerials at each end and a frequency spacing of $5\frac{1}{2}$ Mc/s. If f_1 is the frequency of the transmitter T1, the wanted signal received by R1 is $5\frac{1}{2}$ Mc/s away, say at $f_1 + 5\frac{1}{2}$ Mc/s. The effective rejection of f_1 by the filter FR1 is 60 db so that for a transmitter power of +17 dbw, the level of f_1 at the input of R1 is −43 dbw. Assuming the level of the wanted signal to be −90 dbw, the relative levels of the interfering and wanted signal is 47 db.

If the rejection of the interfering signal by R1 is 90 db the input level to the transmitter, T2 at $34\frac{1}{2}$ Mc/s (29 Mc/s receiver i.f.) will be $90 - 47 = 43$ db below the required signal at 29 Mc/s. The measured rejection at $34\frac{1}{2}$ Mc/s of the transmitter is 62 db so that if the frequency of T2 is f_2, a spurious signal $f_2 - 5\frac{1}{2}$ Mc/s will be radiated at 105 db below the level of the transmitter output. Further, the amplitude distortion of the class C stages of the transmitter results in a second spurious radiation at $f_2 + 5\frac{1}{2}$ Mc/s. If the transmitter power is +17 dbw the level of the transmitted spurious signals will be −88 dbw. One of these spurious signals is on the same frequency as the wanted signal received by R2 (in Fig. 7 it is assumed to be the signal at $f_2 + 5\frac{1}{2}$ Mc/s). The transmitter filter will reduce the level of the spurious signal by 36 db, the level of the input to R2 is therefore −124 dbw.

On a particular type of equipment the maximum interfering signal which could be

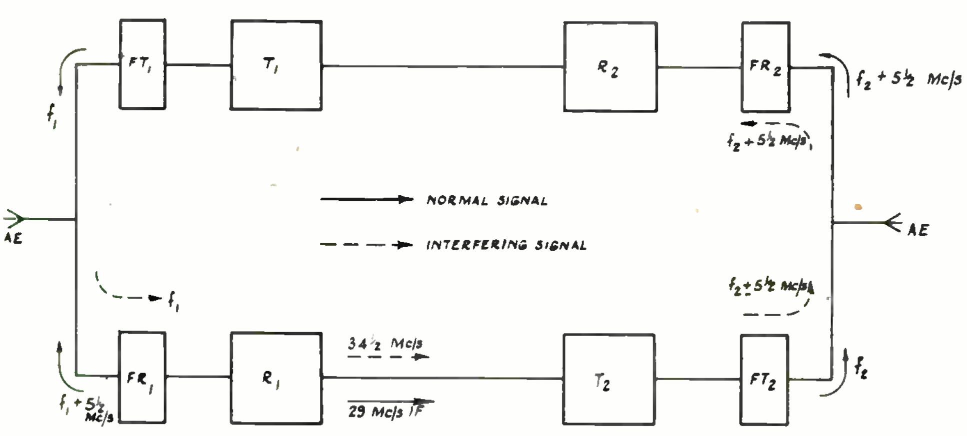


Fig. 7.—Repeater station "loop" type interference.

tolerated was -159 dbw. It was therefore necessary to provide further rejection of the order of about 40 db at $\pm 5\frac{1}{2}$ Mc/s. A reduction in the level of the interfering signal in the above case could be achieved by making the spacing between R1–T1 $5\frac{1}{2}$ Mc/s and between R2–T2 $7\frac{1}{2}$ Mc/s. The resulting interference signal at R2 would then be 2 Mc/s away from the wanted signal instead of on the same frequency. Particular requirements of frequency planning will require to be considered for each case before the latter or similar scheme may be adopted.

In general it may be said that the best method of reviewing a system for freedom of spurious noise is by trial and error. The signal/noise ratio for a particular channel should be calculated from formula (1) to give the required input signal level. A suitably modulated signal of the correct level is then injected into the equipment and the signal/noise ratio measured. An interfering signal of the expected level may then be injected and the channel signal/noise ratio again measured in various channels. It is advisable to move the interfering signal a few kc/s whilst doing this in order to make sure that the beat frequencies (if any) are actually falling in the channels under test. A panoramic receiver to review the modulation spectrum is also of great assistance in tracing down spurious signals.

7. Automatic and Auxiliary Equipment

One of the less obvious requirements in a communications system is reliability. Many factors contribute to this feature but even in the best equipment failures do occur and the "goodness" of the circuit will in the long run be measured in terms of availability for service.

With this in view it has now become common practice to provide automatic change-over facilities which come into operation in the event of a failure of the working equipment. In general, there are two ways of doing this:

(*a*) Automatic change-over at each station.

(*b*) Duplicate path operation with automatic change-over at the terminals.

Figure 8 shows diagrammatically the various arrangements. In each case, however, the principle and result is the same, the only difference being the degree of reliability achieved.

Basically the principle is that a pilot signal at a convenient frequency is injected into the transmitter and monitored at the following points:—

(1) the output of the transmitter.

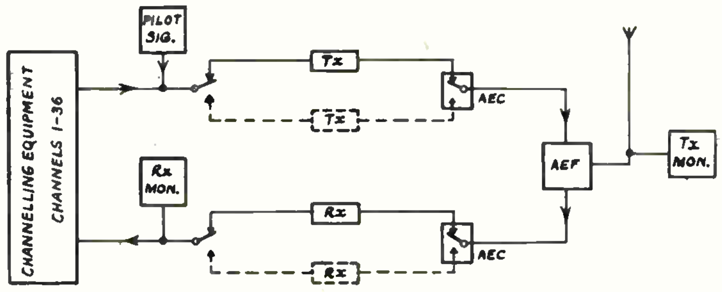
(2) the output of the receiver at a repeater station.

(3) the output of the transmitter at a repeater station.

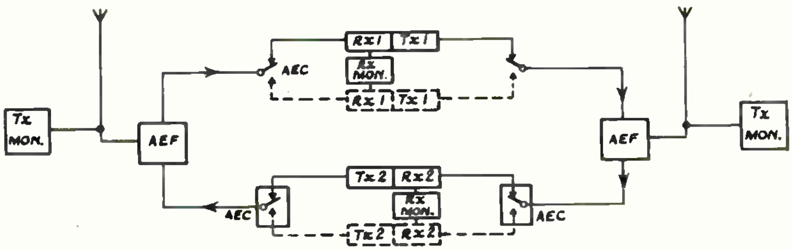
(4) the output of the receiver at the terminal station.

Disappearance of the signal at any of these points initiates the automatic equipment. The fundamental difference between systems (*a*) and (*b*) above is that no repeater station change-over is required at (*b*), and the modulation input is changed over at the terminals only. The reason behind this difference in technique is that in case (*a*) there is no guarantee that a faulty working equipment will change over to good standby equipment since the standby equipment may have developed a fault in the standby condition. If the arrangements are such that the standby equipment is available either in a completely cold condition or with heaters on and reduced h.t., it is not possible to monitor it, and therefore a fault may develop unknown to the operating staff. The provision of separate monitoring equipment for the standby sets would increase the complication considerably and raise the cost of the ancillary equipment beyond a reasonable limit. Even with the above minimum of monitoring and change-over facilities the cost of such equipment tends to approach that of the radio equipment itself.

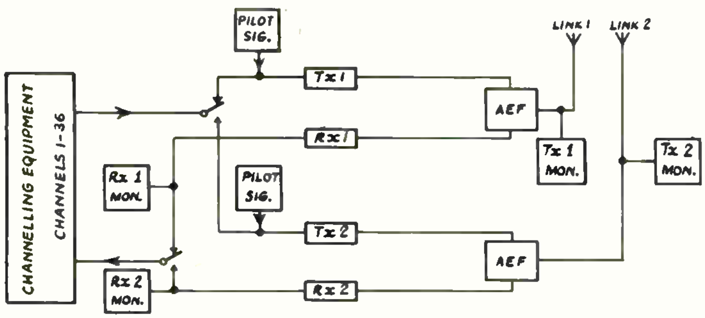
On the other hand, an arrangement such as Fig. 8c permits both the working and standby equipments to be monitored, continuously, and therefore suitable alarms given in the event of failure of either link. Apart from the main advantage described above, the other benefit of this system is that the standby link may be brought into operation during peak hours in order to ease traffic congestion. The main drawback, however, is that twice the number of radio frequencies are required and twice the number of aerials. Further, since both equipments are in full operation the power consumption is doubled. In isolated sites where the power is supplied by diesel generators, the supply and maintenance of fuel may be a considerable problem.



(a) *Terminal station with automatic change-over to standby equipment.*



(b) *Repeater station with automatic change-over to standby equipment.*



(c) *Terminal station with automatic change-over facilities. Duplicate path operation.*

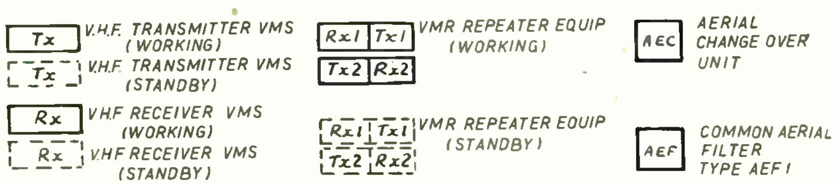


Fig. 8.—Block diagrams of various automatic change-over arrangements.

8. Monitoring Arrangements

The choice of a suitable pilot signal depends largely on the modulation spectrum of the equipment involved, but as a guide the following points may be considered.

(i) If possible the frequency of the pilot should be above the highest channel frequency. This avoids harmonics of the pilot becoming troublesome. Further, it avoids the requirement of band-stopping the pilot from the in and out lines, and the necessity of squeezing it in between channels.

(ii) If the equipment is not going to change over due to a severe fade the signal/noise ratio of the pilot channel must be good. As mentioned earlier, the signal/noise ratio is inversely proportional to the frequency of the channel, and unless a high deviation is used and/or narrow band filters, it will be found difficult to achieve a good signal/noise ratio. However, the deviation used must be considered in conjunction with the system loading in order that crosstalk problems may not arise.

(iii) The determination of the worst pilot channel signal/noise ratio tolerable for any system depends largely on the design of the receiver monitor. It is advisable therefore to determine the limits of this part of the equipment first and then choose the pilot signal frequency in conjunction with the other considerations.

The receiver monitors must fulfil the following requirements:

(i) respond only to the frequency of the pilot signal.

(ii) be incapable of operation on noise.

(iii) be able to operate at a critical signal/noise ratio without dither.

(iv) have the highest possible reliability.

The requirement of (iii) above is often made variable so the change-over takes place at a predetermined signal/noise ratio. This is open to the objection that it is difficult to determine a figure for this. Occasional deep fades may cause false operation, and only operational experience can determine a final compromise value. It is possible and even desirable to fit a thermal delay relay which will only operate after say two minutes. The assumption is made that no single deep fade will last longer than this time, and therefore if the signal drops out for more than two minutes it must be due to a faulty link. In order to avoid indecision on the part of the control relay it is advisable to provide some form of flip-flop circuit. If this is not done the exact signal/noise ratio at which a change occurs depends to a certain extent on the contact tensions of the relay. This factor is a highly variable one and provision of some form of trigger circuit avoids this marginal condition.

The requirements of the transmitter monitor are as follows:—

(i) must be tuneable over the transmitter working range.

(ii) if an f.m. system is used an f.m. detector must be provided and this must be free from variations of output caused by amplitude variations of input.

(iii) no appreciable mismatch should be introduced into the aerial feeder cable by the presence of the monitor.

(iv) must be capable of decisive operation for a predetermined drop in input power.

Automatic change-over at repeater stations is complicated by the fact that an indication of failure of the transmitter monitor may be caused by:—

(*a*) the transmitter itself failing.

(*b*) failure of the driving receiver.

(*c*) failure of the incoming signal due to the failure of a previous station.

There is no simple way of distinguishing (*b*) and (*c*) from each other, and the most convenient method is to provide some form of time delay on the receiver monitor. The transmitters are allowed to change over as fast as they are able, the following receiver being delayed sufficiently long to enable the previous transmitter to become fully operational. One of the dangers that arise in this connection is that the lack of a signal to a receiver results in that receiver driving its transmitter with noise. This condition is obviously unacceptable since the transmitted spectrum is very wide and might cause considerable interference to adjacent channels. In order to avoid this state, it is necessary to provide some form of "dummy drive" to the transmitter. This is brought into operation immediately the receiver monitor shows a no pilot signal condition and before the transmitter

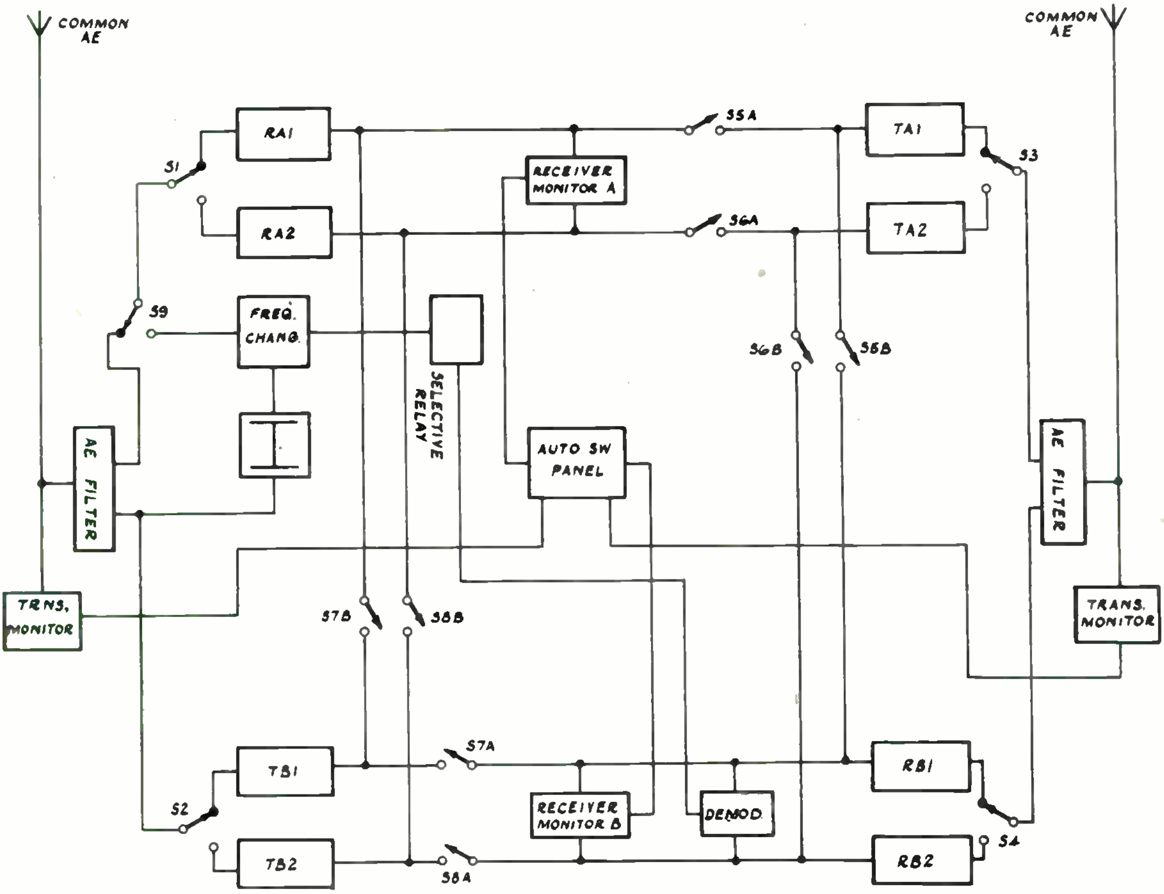


Fig. 9.—Block schematic of a repeater station showing dummy drive and loop switching.

monitor has time to operate. The transmitter monitor, for this reason, has a short time delay of just sufficient length to mask a receiver failure as indicated by the monitor.

Figure 9 is a block schematic of a repeater station fitted with automatic change-over illustrating the principles described above. RA1, TA1 is the left-right path for the signal and RB1, TB1 is the right-left path. RA2, TA2 and RB2, TB2 are the appropriate standby equipments. If it is assumed that the station is operating as shown, a failure of say RB1 will cause the following sequence.

(i) No signal will be received by receiver monitor 'B'. Immediately a thermal delay comes into operation which will not permit a change-over for, say, 1 minute.

(ii) Electronic switches S7A and S7B operate. S7A opening to remove the receiver RB1 and S7B closing to put some of the RA1-TA1 drive on to TB1 and thereby hold in the left-hand transmitter monitor. (Switches S8A, S8B, S5A, S5B, S6A and S6B operate in the same manner depending on which receiver either normal or standby shows a fault.)

(iii) If the fault was actually on the previous transmitter to RB1, then on the restoration of the signal, S7A and S7B would be put back to normal and the thermal relay would drop out. However, if at the end of 1 minute no signal is received at monitor 'B' then a change-over of RA1–RA2, TA1–TA2, RB1–RB2 and TB1–TB2 takes place, and a suitable warning signal injected into the engineering channel.

It should be noted that S7A and S7B operate faster than the transmitter monitor relay.

If, however, TB1 is at fault then the relay operated from the transmitter monitor causes an immediate change-over of all equipments.

9. Location of Faulty Repeater Stations

In a long circuit employing several repeater stations, it is obviously advantageous to locate a faulty unit before a service party is sent out. This may be done in one of two ways.

(*a*) By permitting the station that has changed over to standby to send a discrete signal which is then displayed on a repeater fault indicator at the terminals.

(*b*) To provide means at the terminal for fault location.

System (*a*) is open to the disadvantage that a failure of any common equipment such as power supplies renders the station completely incapable of sending signals. In general, however, the equipment required to operate (*a*) is less than required for (*b*). System (*b*) removes this disadvantage and therefore reduces the overall fault liability. The most straightforward method in (*b*) is to provide some form of selective signalling from the terminals so that each repeater station will respond to its own signal. On receipt of this signal (coming from right-left by, say, RB1 in Fig. 9) the selective relay operates. Switch S9 changes over and the frequency changer unit is switched on. A small portion of TB1 output is then applied via attenuator A to the frequency changer. Frequency conversion takes place so that TB1 frequency is now equal to RA1 frequency. The signal sent from the terminal is now returned there and therefore that particular station is proved. This process is continued down the line until it is no longer possible to loop, when it may be assumed that the fault lies in that particular station.

In the type of system shown in Fig. 9 the difficulty lies in the fact that when a fault occurs the system changes over and if all is well the circuit is restored. Therefore by looping no fault will be disclosed. However, if the standby equipment is faulty then the looping type of fault finding will prove indispensible. If the circuit consists of two radio circuits in parallel as in Fig. 8c when repeater stations are not on automatic and the change-over is at the terminals, the loop switching technique is all that is required. However, in the automatic system of repeater change-over it is therefore advisable to have both methods (*a*) and (*b*) above available to meet with all contingencies.

A further advantage of having some loop switching arrangement is that it enables the maintenance staff to locate a deterioration of the circuit as distinct from a complete failure. When the circuit is operating normally the signal/noise and distortion figures are tabulated for each looped condition. When a partial fault occurs it is then only necessary to loop each station and check the measured values against those of the existing tables. It should be noted in Fig. 9 that looping is shown only on the left side, there is no reason of course why it cannot be provided at both sides other than the fact that twice the number of selective signals will be required. A typical monitoring and auto change panel is shown in Fig. 16.

10. Signalling Systems

When automatic equipment of the kind described above is developed it becomes increasingly difficult to find enough modulation spectrum into which the selective signalling tones can be put; the danger is that more and more of the useful modulation frequency range of the equipment will be utilized for signalling and not for its rightful purpose. To overcome this danger it is necessary therefore to have the signalling frequencies spaced very close together. This in turn involves narrow-band filters and in consequence stable oscillators. Very quickly therefore the cost of the ancillary equipment rockets up and gets out of proportion to the function it fulfils. An interesting adaptation of a well-known device largely overcomes this danger and it may be of interest to describe it briefly.

The Automatic Telephone & Electric Co. Ltd. has for some time past been marketing a system called "Rythmatic Control." This system was developed initially for the switching of street lights, and in view of the difficulties described above, lends itself remarkably well to the radio application.

The system comprises a generator of square pulses with a p.r.f. in the range 7–12 per second approximately. The pulses are then used to modulate a tone of 800 c/s. This pulsed tone is transmitted on the engineering channel and may be dropped off at the required point and utilized to operate a mechanical relay. It is possible to have 12 different p.r.f.'s or "rhythms" in the range 7–12 and therefore if the receiving mechanical relay is adjusted to be tone and p.r.f. resonant at the required rhythm, a switching operation can take place. The mechanical relay or galvo (Fig. 10) consists of a length of tensioned wire which supports in the middle a small permanent magnet at right angles to the wire. Surrounding this assembly is a coil into which is fed rectified output derived from the 800-c/s pulsed modulation. If a particular galvo is tuned to the correct received p.r.f. then a movement takes place torsionally of increasing amplitude until such time as a contact mounted on the wire makes contact with a fixed member. This

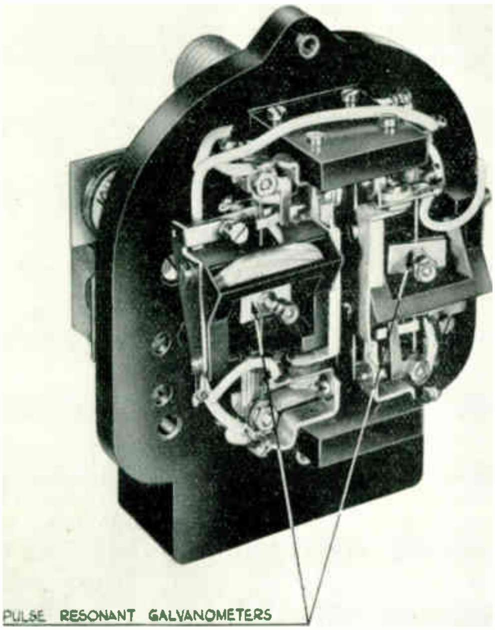


Fig. 10.—"Rythmatic" relay type 46.VS.

contact then energizes a latching relay which in turn carries out the required switching function. The relay is unlatched by means of a different p.r.f. and a different receiving galvo tuned to the off rhythm.

In the system which has been developed for loop switching the repeater stations, each "end" of each repeater station is allocated its own switching rhythm. A common "off" rhythm is used throughout. The above switching functions (up to 12) are carried out on one tone, but if it is desired to increase the number of operations then other tones using the same rhythms as before may be used. The change in galvos is one of frequency discrimination only. This frequency discrimination is carried out by means of a simple tuned circuit.

Apart from the actual switching of repeater stations it is obvious that many facilities may be provided by means of selective rhythm working, for example, information may be transmitted from the repeater station to the terminal and there used to operate alarm signals. Examples of this are power failure and restoration of standby power supplies, illegal entry, fire, indication of repeater change-over, on/off switching of terminal stations from a remote point, etc. In order to increase the number of switching operations it is only necessary to change the tone frequency to, say, 1,200 c/s when all the original rhythms may be used over again. The many applications and ramifications of this type of signalling are outside the scope of this paper but are fairly obvious.

11. Practical Layout of a Multichannel Equipment

The following is given as a brief example of a multichannel equipment designed to comply with the foregoing discussion. The large part of the radio conforms to normal practice and therefore is outside the scope of the paper.

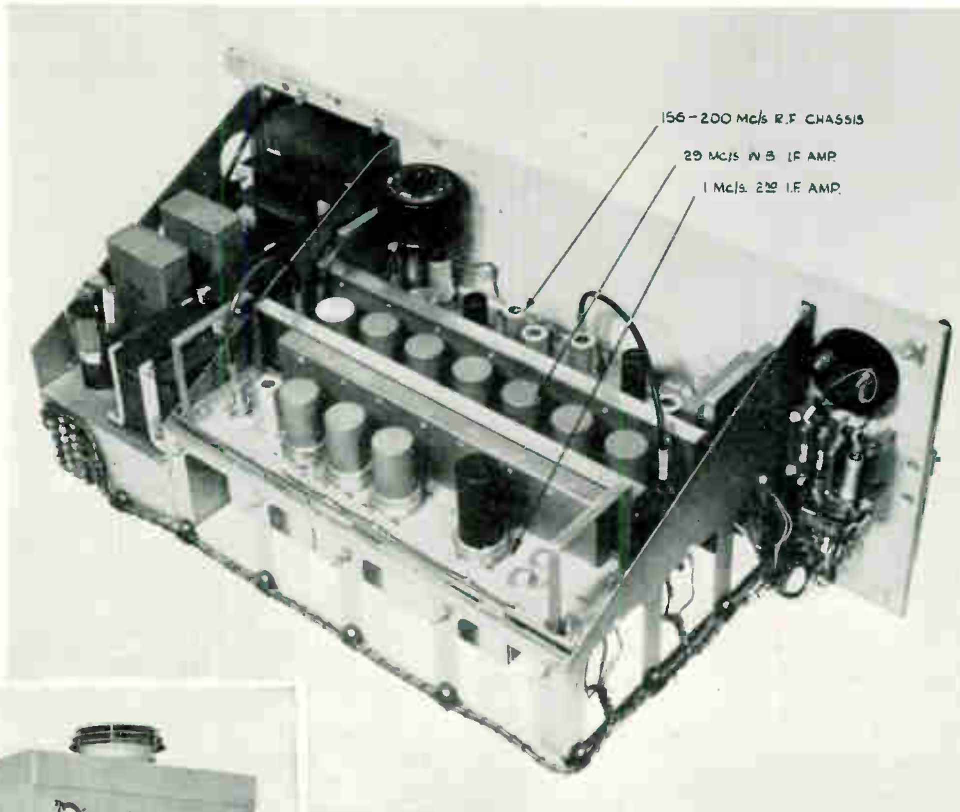
The incoming multichannel carriers modulate the frequency of a 14·5 Mc/s oscillator with a very high degree of linearity and this 14·5-Mc/s signal, after frequency doubling, is mixed with a suitable crystal-controlled frequency to provide the transmitting frequency in the ranges 40–60 Mc/s, 60–90 Mc/s and 156–200 Mc/s.

In the receiver, automatic tuning ensures that the signal is always brought to the middle of the i.f. band. The multichannel carriers are finally recovered by an electronic frequency integrator (demodulator).

Figures 11, 12 and 14 illustrate the general mechanical layout of a terminal equipment, while Fig. 13 is a block diagram of the circuitry.

11.1. *Transmitter Cabinet*

Referring to Fig. 13, a 29 Mc/s signal from the modulator is first amplified by a two-stage wideband amplifier which has an overall bandwidth of 3 Mc/s and then mixed with the output of a crystal harmonic amplifier to give the transmitting frequency.



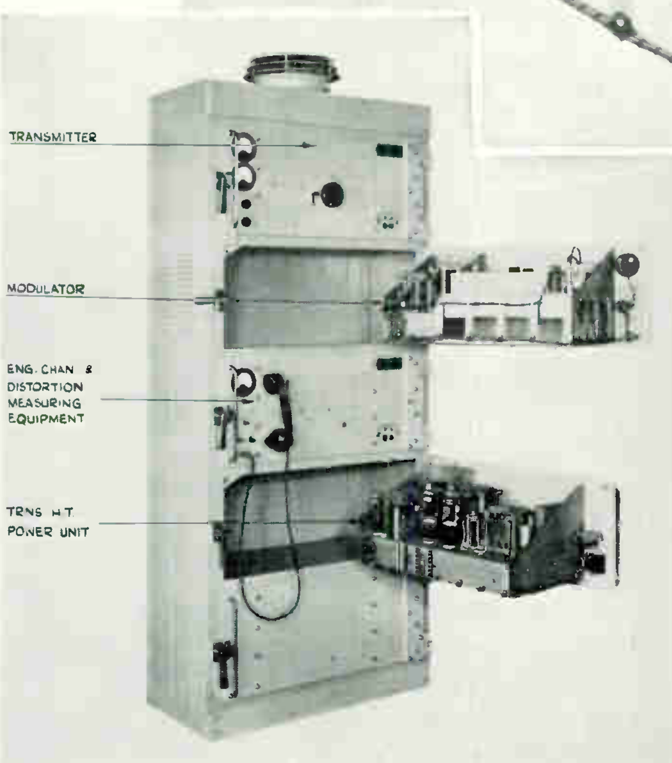


Fig. 11 (above).—V.H.F. Radio Receiver, V.R.2.

Fig. 12 (left).—Terminal Station Transmitter V.M.S.

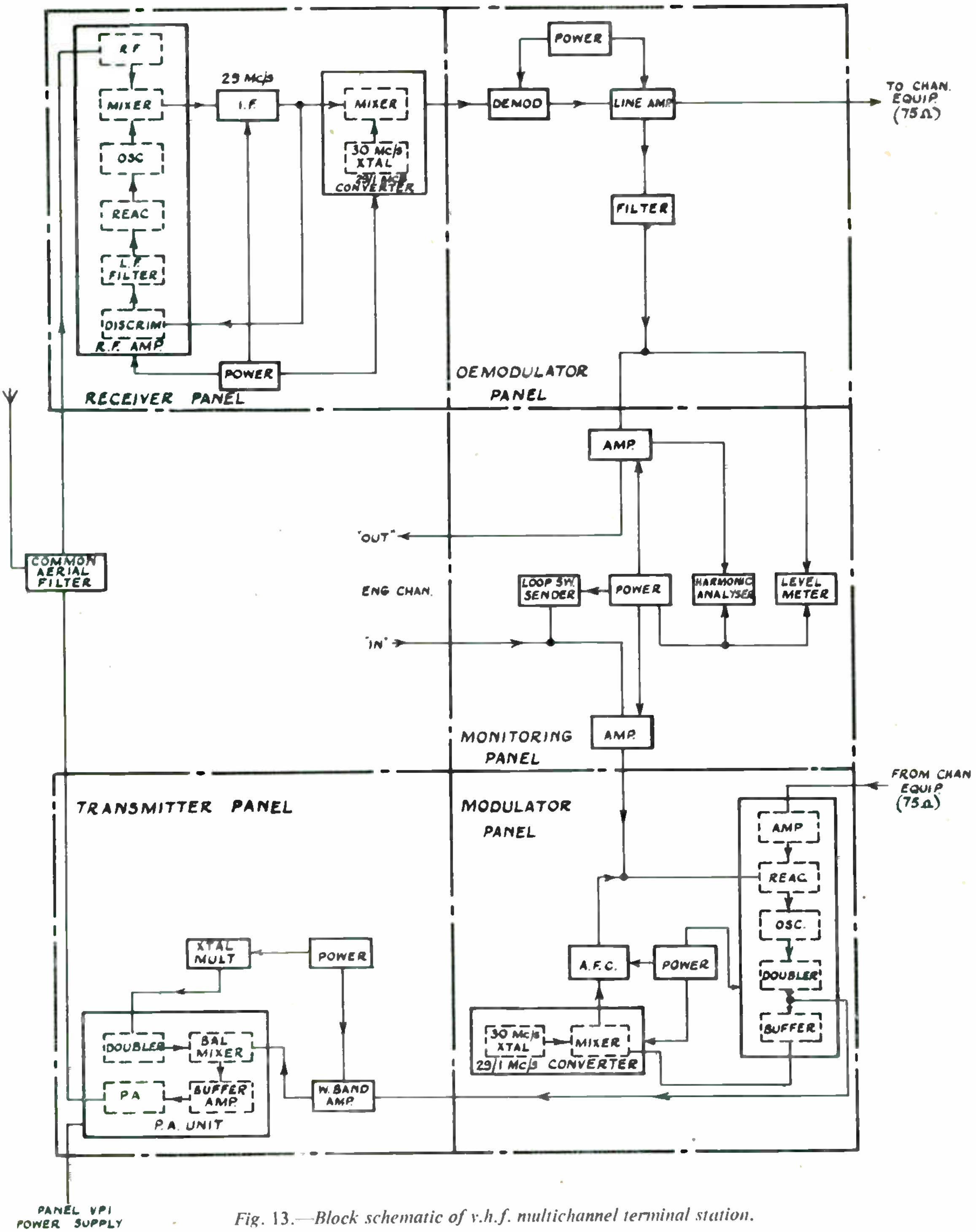


Fig. 13.—Block schematic of v.h.f. multichannel terminal station.

The harmonic amplifier generator consists of crystal oscillator and multiplier stages giving an output frequency 12 times that of the crystal frequency at a level of approximately 1 W.

The 20 Mc/s signal from the wideband amplifier and the 12th or 24th harmonic, as appropriate, of the crystal frequency are mixed in the first stage of the power amplifier. In the 40–90 Mc/s equipment this first stage consists of two pentodes in push-pull, but the 156–200 Mc/s equipment has a frequency-doubler, a mixer and a buffer amplifier to drive the final power output stage.

The r.f. output of the 40–90 Mc/s equipment is about 45 W at 600 Ω unbalanced whilst that of the 156–200 M/cs equipment is about 50W with a 75 Ω unbalanced output.

11.2. *Modulator Panel*

Linear frequency modulation is produced in the modulator which consists of four sub-units.

Wide-band modulator; 29/1 Mc/s convertor;

A.F.C. stabilizer; Power unit.

In the wide-band modulator, incoming multi-channel carrier signals are applied to a line amplifier via a 75 Ω coaxial cable, or a matching transformer, and a variable attenuator network. The attenuator has an attenuation variable between 0 and 12 db to compensate for variations of input level. The input level required is between −27 and −37 db relative to 1 mW per channel.

11.3. *Monitor Panel*

The servicing gear required for routine testing of the complete equipment is contained in the monitor unit. This servicing equipment comprises engineering channel amplifiers with ringing device, level meter for setting up the

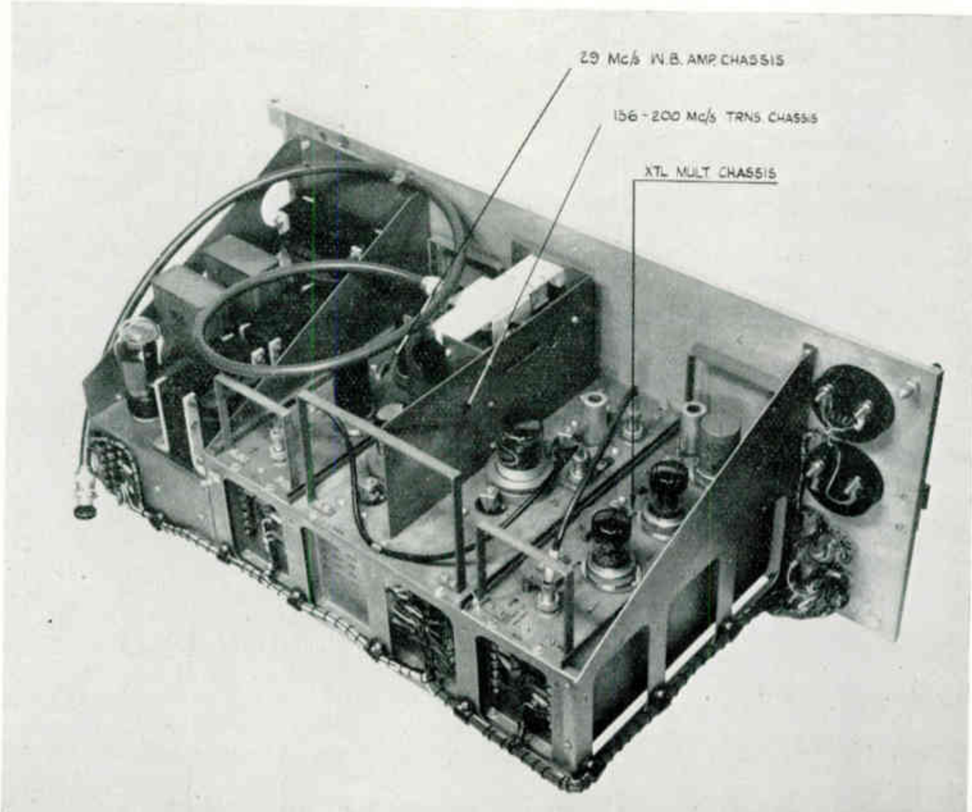


Fig. 14.—V.H.F. Radio Transmitter V.T.2.

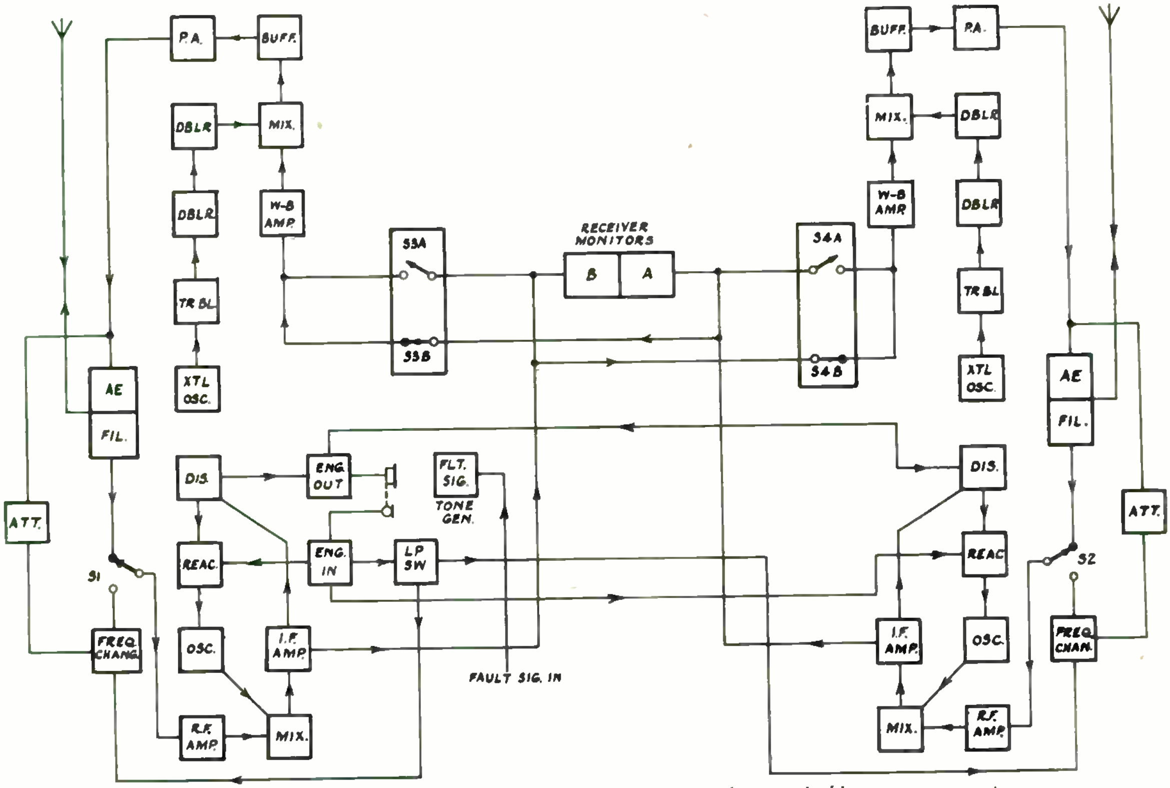


Fig. 15.—Block schematic of repeater station showing loop switching arrangements.

modulation level, a harmonic analyser for checking the overall performance of the link and a d.c. meter for connections brought out to a jack.

The engineering channel provides a speech circuit and ringing facilities for communication between all the terminal and repeater stations of a chain without interfering with the normal multichannel traffic. The speech circuit operates within 300–3,000 c/s and the ringing frequency at 3,000 c/s. Both visible and audible indication of an incoming call is given.

11.4. *H.T. Power Panel Transmitter*

This unit provides the high tension for the output stage of the transmitter panel VT.1. Self-contained power units are fitted to all other panels of a transmitter cabinet and operate directly from the mains.

11.5. *Receiver Cabinet*

The r.f. and i.f. panel (Fig. 13) receives input signals from the aerial which are amplified by an r.f. amplifier and mixed with the output of a local oscillator to provide a 29 Mc/s i.f. Automatic tuning of the local oscillator is used to ensure that the signal is always in the middle of the i.f. band, thus compensating for any transmitted frequency drift and avoiding the tendency towards distortion experienced in broad-band systems employing crystal-controlled receivers, when differential transmitter and receiver frequency drift occurs. Five stages of i.f. amplification with a.g.c. follow the mixer to drive a 29/1 Mc/s convertor which gives a 1 Mc/s output to feed the demodulator.

The overall bandwidth of the r.f. and i.f. stages is $\pm$750 kc/s to the 3 db points and amply provides for low distortion amplification. Automatic frequency control of the local oscillator is performed in the same way as in the transmitter. In the 40–90 Mc/s equipment, a portion of the 29 Mc/s i.f. signal is mixed with the output of a 30 Mc/s crystal oscillator to give a 1 Mc/s signal which is applied to a 1 Mc/s centre-frequency discriminator. In the 156–200-Mc/s equipment the 29 Mc/s i.f. is applied directly to a 29 Mc/s centre-frequency discriminator. Any change in i.f. from 29 Mc/s gives a bias voltage from the discriminator and this bias is used to control the local oscillator frequency by means of a reactor valve to maintain the i.f. stable within 0·01 per cent.

11.6. *Demodulator Panel*

In the demodulator the 1 Mc/s second i.f. is demodulated in order to recover the multi-channel carrier groups. The demodulator panel contains three sub-units:

- Demodulator;
- Output line amplifier;
- Power unit.

The 1 Mc/s signal from the 29 Mc/s convertor sub-unit is applied to the demodulator via two stages of video amplification and a limiter. Output signals from the limiter are level within ± 1 db for input variations up to ± 25 db from a normal level of -55 db relative to 1 mW. The demodulator used is of the frequency integrator or counter type and is particularly free from distortion. It does not depend upon the accurate setting-up of the tuned circuits for its operation, and is thus reliable and simple to maintain. A low-pass filter and an amplifier stage follow the demodulator and feed the output line amplifier.

The line amplifier sub-unit follows conventional practice and has three stages of amplification with variable negative feedback to give a 6 db gain control. The standard output transformer has a 75-Ω unbalanced output but alternative transformers can be supplied to match 600 Ω or 140 Ω balanced lines. The normal output level is -8 db relative to 1 mW per channel.

11.7. *Repeater Stations* (*Figs.* 15 *and* 17)

The mechanical construction of the repeater station is similar to that of the terminal station, inasmuch as sub-unit construction is used throughout. Furthermore, the transmitter receiver and h.t. power units and panels are interchangeable with those at a terminal. The main feature, however, is that the multichannel carrier signal is not subject to demodulation and modulation. The drive for the transmitter is taken directly from the i.f. amplifier of the receiver (29 Mc/s). The a.f.c. circuit in the receiver pays a further dividend at repeater stations in allowing the engineering channel facilities without a complexity of equipment. The a.f.c. discriminator provides the engineering channel detector, whilst the oscillator reactor valve is available for modulation of the i.f.

In Fig. 15, S3A, S3B, S4A and S4B are electronic switches which provide for dummy drive arrangement when required.

S3B is the normal switched condition for left-right operation and S3B for right-left operation. If no signal is received by say, receiver monitor "A", S3B opens and S3A closes. This has the effect of breaking the right-left drive and replacing it with some of the left-right drive.

The engineering channel is cross connected so that talk and listen takes place in both directions; calling is therefore on a party line basis.

12. Power Supplies

Fully automatic dual diesel/alternator sets are usually made available to supply the station load, where no public mains supplies exist, such as at unattended repeater station sites. The diesel sets each comprise two engines directly coupled to alternators with a common control cubicle. They are normally supplied with fuel, oil and water for four weeks' operation. Each engine is arranged to run for a period of 96 hours at the termination of which the second set is automatically started and takes over the load for a further 96 hours. The cycle is then repeated.

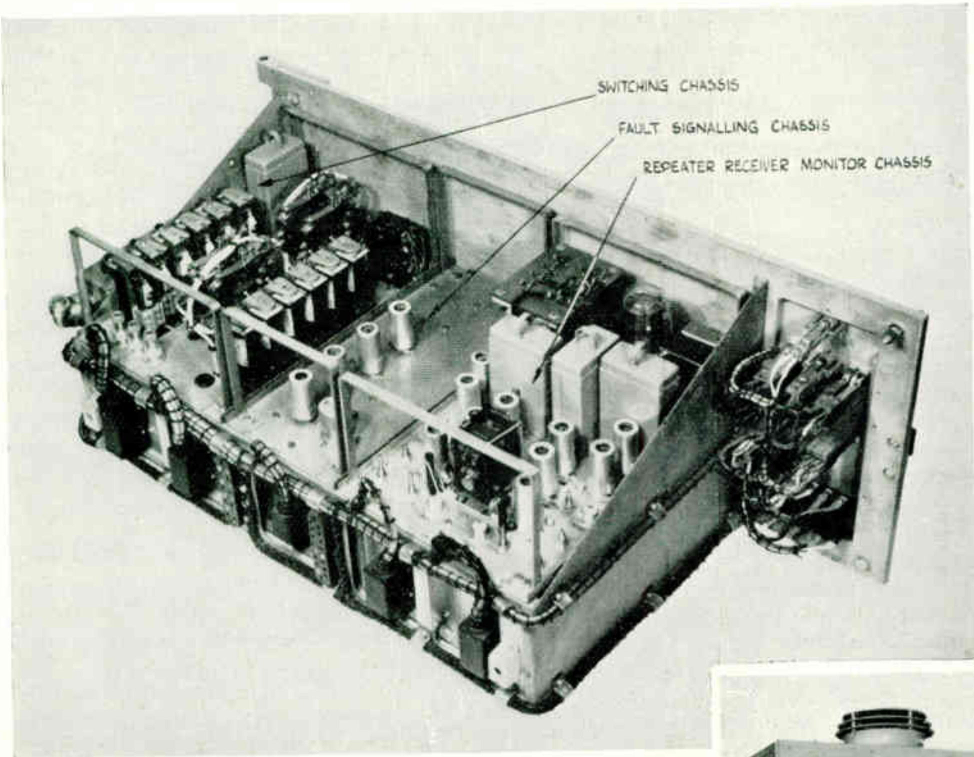
In the event of a failure of the working set, the standby set is automatically brought into service and an alarm condition extended over the engineering channel to the terminal stations.

The importance of ensuring that only the highest grade power plant is employed at radio stations is evident, since a power failure will immobilize all the circuits and invalidate all the automatic systems.

13. Aerials

As was discussed earlier in the paper, the choice of an aerial depends on many factors, and it is not proposed to discuss the design of aerials suitable for wide band transmission, but merely to illustrate two types of aerials in use with the equipment described earlier.

Figure 18a is a Yagi type aerial having a $\lambda/2$ dipole, four directors and a reflecting screen.



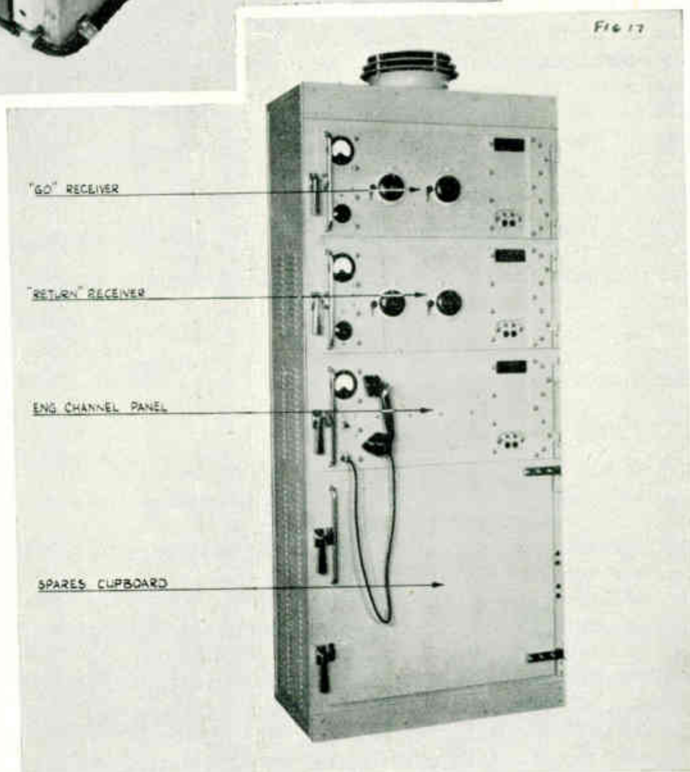
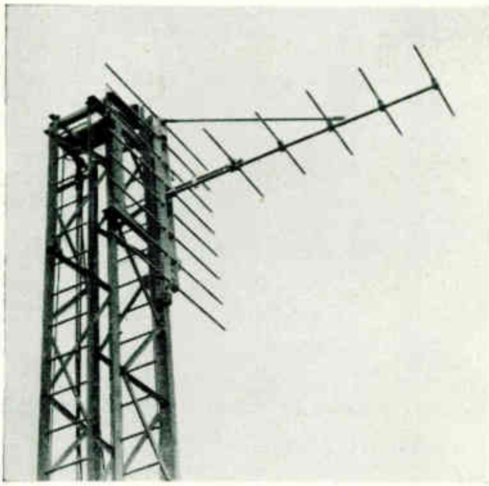
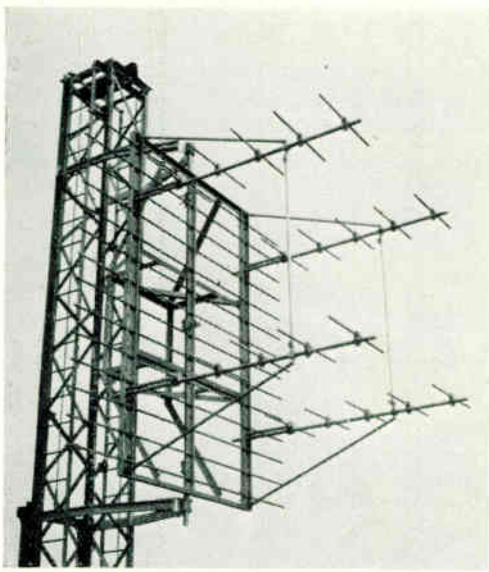


Fig. 16 (above).—V.H.F. Radio Change-over Unit V.M.C.2.

Fig. 17 (left).—Repeater Station Receiver V.M.R.





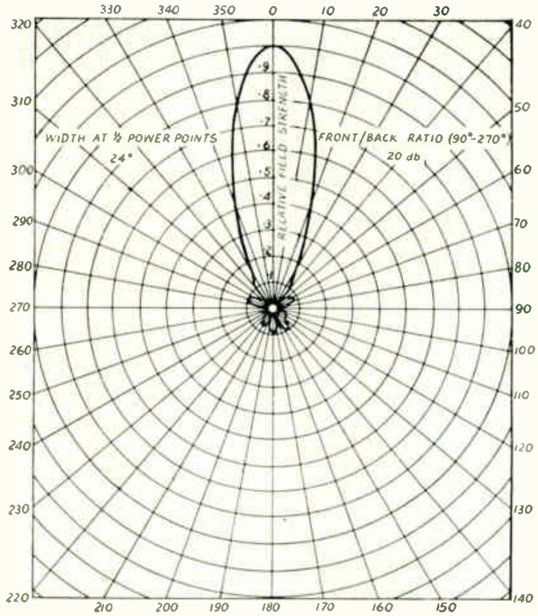


Fig. 18.—(*top left*) (*a*) 10 *db gain Yagi aerial.* (*bottom left*). (*b*) 16 *db gain Yagi aerial, type AER*9. (*top right*). (*c*) *Typical A.E.R.* 9 *polar diagram in plane of elements.*

The gain of this type is of the order of 10 db relative to a $\lambda/2$ dipole.

In all cases it is necessary to keep the impedance match between the aerial and the equipment as good as possible. This is necessary because a mismatch occurring at this point may introduce group-delay distortion which might cause crosstalk. It has been found that for all practical purposes a v.s.w.r. of 1·1/1 is adequate. This figure takes into account all equipment associated with the aerial, e.g. filters, change-over switches, etc.

Figure 18b shows a development of the type shown in Fig. 18c. This aerial gives a gain of 16 db and probably represents the greatest economical gain achievable in the present state of the art. As can be seen from the illustration it comprises four 4-director Yagis with reflecting screen. Data on the specification of such an aerial may be of interest and a few of the more important parameters are listed below.

Operating frequency: 150–200 Mc/s with a transmit/receive frequency separation of 5·5 Mc/s.

Working bandwidth: ±750 kc/s.

Gain: 16 db relative to a $\lambda/2$ dipole in free space. (See Fig. 20.)

Impedance match: The aerial is required to give a v.s.w.r. of 1·18 or better over the working bandwidth.

Front-Back ratio: The radiation between 90° and 270° relative to the direction of maximum radiation is less than 16 db down on the maximum radiation. (Fig. 18c.)

Wind loading: The aerial is capable of withstanding with an adequate safety factor wind loadings of 30 lb/ft^2 acting on its projected area.

14. Acknowledgments

The author wishes to thank Mr. F. O. Morrell, Director of Research of British Telecommunications Research Ltd., for permission to use information acquired during the course of the company's development programme, particularly in respect of work carried out by Mr. J. L. Slow and Dr. W. Hackett.

An article entitled "Basrah Auto-Radio System," in the *A.T.E. Journal*, Vol. 10, No. 1, January 1952, describes a complete system installed by the Basrah Petroleum Company, using equipment referred to in the paper.

15. Appendix

Calculation of Signal-to-Noise Ratio

An unpublished report by Dr. W. Hackett, late of B.T.R., gives the expression for the signal/noise ratio in a high frequency channel of a multichannel group, after transmission over an f.m. radio link as

$$\frac{Sq}{Nq} \simeq \frac{c}{n} \cdot \frac{fdq}{fq}\left[\frac{1}{2fq_2 - fq_1}\right]^{\frac{1}{2}} \quad \ldots\ldots\ldots(1)$$

where Sq = r.m.s. signal voltage in channel.

Nq = r.m.s. output noise voltage in channel.

c = carrier amplitude in r.m.s. volts.

n = amplitude of single r.f. noise component (i.e. noise in unit bandwidth).

fdq = frequency deviation due to signal in channel.

fq = frequency of signal in channel.

fq_1 = lower cut-off frequency of channel.

fq_2 = Higher cut-off frequency of channel.

The approximation is assumed that over the comparatively small bandwidth of the single high-frequency channel, the noise may be regarded as constant.

Substituting,

$fq_2 - fq_1 = B$ (i.e. channel bandwidth)

$\frac{fdq}{fq} = M$ (i.e. deviation ratio)

and expressing formula (1) in power ratios such that

Sw = output power of single channel in watts.

Nw = output noise power of single channel in watts.

Cw = r.f. carrier input power in watts.

nw = r.f. input power in watts for unit bandwidth,

then $$\frac{Sw}{Nw} = \frac{Cw \cdot M^2}{2nw \cdot B} \quad \ldots\ldots\ldots\ldots\ldots\ldots(2)$$

The thermal noise agitation developed across any resistance is $4kT$ watts per unit bandwidth.

When $k = 1{\cdot}374 \times 10^{-23}$ joules/° Kelvin

$T = 300°$ Kelvin

$4kT = 1{\cdot}64 \times 10^{-20}$

Since the noise power is divided equally between the aerial and the receiver, the noise power delivered to the receiver is $0{\cdot}82 \times 10^{-20}$.

For a receiver noise factor F the noise power is increased to

$$\frac{0{\cdot}82 \times 10^{-20} \times F}{2} = 0{\cdot}41 \times 10^{-20}$$

Then from equation (2),

$$\frac{Sw}{Nw} = \frac{Cw \cdot M^2}{0{\cdot}82 \times 10^{-20} \cdot Fw \cdot B} = \frac{1{\cdot}21\ Cw \cdot M^2 \times 10^{20}}{Fw \cdot B}$$

Assuming a channel bandwidth = 4,000 c/s and substituting for B

$$\frac{Sw}{Nw} = \frac{3{\cdot}03 \cdot Cw \cdot M^2 \times 10^{16}}{Fw}$$

Expressed in decibels, $R = 164{\cdot}8 + 20 \log M + C - F$.

where R = signal/noise ratio in any channel of multichannel group (in db)

M = deviation ratio due to single channel

C = r.f. carrier input power expressed in dbw.

For weighted noise measurements the signal/noise ratio is improved by 3·2 db approximately, i.e. $R = 168 + 20 \log M + C - F$.

DISCUSSION

E. G. Hamer (Associate Member): I notice that common aerial working is used in the system described by the author. From the illustrations, the aerials appear to consist of an array of stacked Yagis with a gain of 16 db, and a standing-wave ratio of less than 1·1/1 over a transmit-receive bandwidth of 5·5 Mc/s at 175 Mc/s. Yagi aerials are noted for their narrow bandwidth, and I would be pleased if Mr. Brown could give us more details of the aerials and special matching networks, which enables this high-gain and low standing-wave ratio to be obtained over such a wide bandwidth.

In the case of unattended repeater stations with diesel-electric power supplies, I would welcome the author's views on the telemetering of information about the state of the generator sets. Would he advocate sending information on, say, engine defects, such as low oil pressure, excessive engine- or bearing-temperatures, etc., or would he simply wait until the engine failed and then indicate that a change to standby had occurred?

Another question where Mr. Brown's experience would be of value, is the operational procedure for running the diesel sets. Is it better to run one set continuously for long periods of several weeks or months, or to change over from one set to another, say every 24 hours?

N. Fedida: When designing multichannel radio links, complying with the recommendations of the C.C.I.F. for intercontinental cable circuits, it is usual to allow a total of 10,000 pW for noise of all sources, measured at a point of zero reference level in a channel 300-3,400 c/s wide.

A quarter of this amount is usually allocated to the channelling terminals, while the remainder, i.e. 3 pW/km is allocated to the cable route, including repeaters.

In the case of radio links, which may or may not be part of cable circuits, the same allowance of 3 pW/km is made for the route, and it enables the designer to lay down specifications for the repeaters.

The noise arising in the radio terminals on account of thermal agitation and intermodulation caused by non-uniform group delay, is, in general, equal to, or greater than, the noise arising in a repeater from the same causes.

However, because of the additional modulating and demodulating functions performed by the radio terminals, these must of necessity contribute an additional amount of noise. What special allocation, if any, is made for this additional amount of noise, arising at the radio terminals, over and above the allocation made for repeaters on a route-mile basis, i.e. 3 pW/km or any other similar figure?

At what deviation and modulation frequencies are the figures of −65 and 70 db for the second order and third order harmonic margins, of the counter-type discriminator described, obtained?

Dr. T. G. Hammerton: I would like to refer to the question on aerial characteristics raised by Mr. Hamer, which implied that because the transmitter and receiver frequencies at a given station, working to and from the same distant station, may be operating on carrier frequencies separated by, say, 5·5 Mc/s, then their common aerial had a bandwidth of 5·5 Mc/s. This is not so, as it is usual to arrange such a common aerial to be a "two-frequency" type. By suitable disposition of the Yagi elements—length and spacing—it is possible to make the aerial impedance purely resistive at two frequencies, which, by suitable arrangement, can be varied up to about 8 Mc/s apart in the 156 to 200 Mc/s band. Around each working frequency there is a frequency band, as Mr. Brown stated, of width ±750 kc/s over which the impedance match can be maintained with a voltage standing-wave ratio of 1 to 1·1 or better. Providing the spurious emission and receiver bandwidths are of an adequate standard, then the aerial impedance characteristic is between as well as beyond the two working bands, and can be outside that required to give the 1·1 match.

E. J. Schindler (Associate): Do field surveys in tropical countries provide sufficient data for the establishment of an u.h.f. link without access being available to accurate records of climatic conditions over a number of years?

B. B. Migawski (Student): The v.h.f. multichannel telephone equipment discussed, has a reactance valve as a means for producing f.m. Would Mr. Brown explain why the reactance valve has been chosen and whether other means of producing f.m. have been considered?

One of the factors for better signal-to-noise ratio is a linear function between modulating signal and frequency deviation. Can this be fully exploited by using the reactance valve?

A. G. Williamson: Mr. Brown has given a warning about the expense of elaborate monitoring systems in terms of capital and bandwidth.

If it is considered necessary to have elaborate monitoring systems engineered as part of the radio equipment, could he indicate what facilities could be provided over a system if the cost of the monitoring equipment is allowed to approach half that of the working and standby radio equipment in a link employing six repeaters?

P. P. Eckersley (Member): Mr. Brown referred to the measurement of cross-talk (caused by intermodulation products) by means of a noise spectrum applied to the system. Is this noise weighted to simulate speech conditions or is it "white" noise?

AUTHOR'S REPLY

The aerial characteristics together with a polar diagram, are given in the paper. However, a further comment might be made that the four Yagi elements are, in fact, in parallel, and the impedance value corrected by suitable matching transformers. The resistive component is corrected by means of $\lambda/4$ matching transformers, whilst the reactive component is corrected by adjusting the length of the driven element.

In reply to Mr. Hamer's other questions, there does not seem to be a great advantage to be gained in telemetering the various functions of the diesel generator. What is important, however, is that a distinctive signal at the control centre indicates whether a fault on the circuit is due to power equipment or radio equipment. This information permits the servicing department to send a suitable engineer to the site. The faulty repeater station may well be many hours travelling away from the control centre, and the knowledge that, say, the diesel is overheating might well be valueless because the servicing party could not get there in time to prevent a failure in any case. In the author's opinion, the diesel should be fitted with its own alarm circuit, which would switch it off in the event of something extraordinary taking place, and then a distinct "diesel failure" alarm sent to the control centre.

Opinions differ on the best operational procedure for diesel sets, but in general it is better for the engine to be run fairly frequently in order to avoid excessive corrosion taking place, which might occur were the engine allowed to stand idle for long periods. The concensus of engineering opinion on the subject appears to be that about 96 hours on/off periods is a reasonable time. However, fuel oil and water supplies, should be adequate for at least a month's continuous running.

I am in agreement with the first three statements made by Mr. Fedida, but cannot entirely agree with the fourth statement. So far as thermal noise and intermodulation noise caused by non-uniform group delay is concerned, the terminals of a radio link, i.e. a transmitter and a receiver are just the same as a repeater station. Therefore, for a circuit of n repeaters there are $n + 1$ equipments, and each of these equipments should contribute the same amount of thermal noise and noise from group delay sources (assuming, of course, that the path attenuations are the same for all sections of the link). This being so, the only additional noise source is that of the modulator and demodulator at the terminals. Since there are many tuned circuits between terminals, the possibility of intermodulation noise arising at these points is fairly high and should, therefore, considerably outweigh the intermodulation noise arising from the modulator and demodulator circuits. Even in a terminal to terminal link, it is good practice to have this state of affairs.

The noise allocation of 25 per cent. to the terminals of a 1,600-mile circuit is, in fact, for the channelling equipment only. So that if 10,000 pW is the total noise power allowable, 7,500 pW is the amount allocated to the path itself, be this line, radio or a combination of both. As was stated above, there is no difference between a radio repeater and a radio terminal so far as noise generation is concerned, and therefore the amount of noise allowable in any one part of a circuit is directly proportioned to the length of that part of the circuit. For example, a four-repeater station circuit has five links, and if this covers 1,600 miles (which is very improbable), then the noise power permissible per link would be 1,500 pW.*

The particular figures of -65 db and -70 db for 2nd and 3rd order harmonic powers, referred to a centre frequency of 1 Mc/s, a deviation of ± 200 kc/s and a modulation tone of 3 kc/s.

The degree of reliability that can be placed on a field survey, is only as good as the thoroughness with which that survey has been carried out. If the

* See C.C.I.F. 3émé Commission d'études—Geneva, October, 1953—Reply to questions A, B and C.

survey has been carried out over a period sufficiently long to embrace all the vagaries of the local climate, then a very good estimate of the expected path attentuation can be made. Looking at it from the other point of view, it would be of rather dubious value to estimate the path attenuation purely from a knowledge of the climatic conditions only. This is because it would be almost impossible to have all the relevant meteorological data available for all parts of the link for all of the time. It should, however, be emphasized that an extensive survey together with a theoretical estimation of the path will, together, give a very close estimation of what can be expected under operational conditions.

The reactance valve circuit as a means of producing f.m. has many commendable features, provided always one can achieve the required degree of linearity. For example, the fundamental frequency may be made sufficiently high to permit of the minimum amount of frequency multiplication. It is also capable of wide deviations so that if only a small portion of the characteristic is utilized, that portion closely approaches a truly linear state. Other means of producing f.m. are, of course, available such as phase modulation of a crystal. The ultimate choice depends on many factors, amongst which the degree of rejection given by the rest of the equipment to spurious frequencies from the modulator is important.

In the second part of Mr. Migawski's question, it is presumed that by signal-to-noise ratio is meant signal/intermodulation noise only. This point is covered largely in the answer to Mr. Fedida's question, viz.: that the design of a radio system should be such that the intermodulation noise accruing from the modulator and demodulator should be such that it is small compared to that arising from sources causing non-uniformity of the group-day characteristic. If this is not so, then the intermodulation noise, for a single link, would be out of proportion to the length of the link.

It is, in fact, possible by good design and careful choice of the parameters, to make a reactance valve modulator adequate for all practical purposes. However, since an oscillator controlled by a reactance valve must of necessity be inherently unstable, some form of centre frequency stabilization must be employed. This may well take the standard form of a.f.c. circuit using a conventional discriminator, which, in turn, must have a stability better than the required overall stability. An overall frequency stability of better than 0·01 per cent. may be achieved by this means.

As Mr. Williamson suggests the monitoring and alarm facilities may reach alarming proportions in respect of the radio equipment cost. To detail just what facilities could be provided for a cost of half that of the radio would be very difficult without knowing more about the particular application. For example, if the circuit is a duplicated radio path, then no monitoring will be required at repeater stations. However, the facilities might include the following: remote control and monitoring of terminal stations; local control and monitoring of terminal stations, including all the appropriate relays and switching; transmitter and receiver monitors; repeater fault location apparatus; repeater signalling to terminal equipment.

The cost of any of the above equipment depends, in turn, on the number of particular facilities required by the customer and on the lengths to which the designer is prepared to go to make the system reliable and foolproof.

To measure the fully loaded characteristics of a system from the input of the modulator to the output of the demodulator a spectrum of white noise of the required baseband is applied, and the input spectrum is bandstopped at the frequency one is interested in. The use of channelling equipment for this purpose is merely because it represents a convenient assembly of filters. The most crucial point is the level at which this noise is injected, and is based on the statistical analysis of numbers of channels and average speech power (a curve for this is given in Fig. 5).

If a complete system which includes channelling equipment is used then, since the input to each channel is at l.f., application of a noise voltage to all channels in parallel would represent an unreal condition. This is because the noise "waveform" would be the same for all channels and therefore the peak voltages would add up in an unrealistic manner, i.e. unrealistic so far as an equivalent number of speech channels is concerned.

In order to achieve something approaching a random condition, a satisfactory method is to take a large spectrum of noise and select by means of simple-tuned circuits the required number of noise spectrums. These noise spectrums are then rectified and applied to the channel inputs at the correct level. By this means purely random voltages appear in each channel, and bear no relation to any other channel. Any voltage additions which then takes place will represent a reasonably true condition.

GRADUATESHIP EXAMINATION NOVEMBER, 1953

FIRST PASS LIST

This list contains the results of all candidates in the British Isles and those of the overseas candidates which were available on January 14th.

Eligible for Transfer or Election to Graduateship or Higher Grade of Membership

(These candidates have satisfied the Examiners in all parts of the examination)

BETTERIDGE, John E. (S) *Stoneleigh, Surrey.*

CHRISTMAS, Bernard Harrison. (S) *London, E.11.*
CURRY, Eric. (S) *Parkstone, Dorset.*

DIX, Dennis Lee. (S) *Ruislip, Middlesex.*
DOREY, Cecil Frank. (S) *Ilford, Essex.*
DOWSETT-MARSH, Julian Caryl. (S) *Rayleigh, Essex.*

ELLIS, Alfred Brian Edwin. (S) *Eglinton, Northern Ireland.*

HAMBLETON, James. *Consett, Co. Durham.*

KULAHIAN, Kevork. (S) *Beirut, Lebanon.*

LEE, Kenneth. (S) *Leeds.*

MILLBURN, John Richard. *Aylesbury, Bucks.*

NEELY, Terence Joseph Prescott. *Grahamstown, South Africa.*

ORMISTON, Peter Thomas. *Morden, Surrey.*

PORAT, Dan Israel. (S) *Rehovoth, Israel.*

ROBERTS, Stanley John. (S) *London, S.E.18.*

SCHLOSS, Ralf. (S) *Tel-Aviv.*
SHAHZAD, Iqbal Hasan. (S) *London, W.4.*

WEGER, Meir. (S) *Tel-Aviv.*

The following candidates were successful only in the Part or Parts indicated

ANWAR, Ahmad. (S) *Fareham, Hants.* (IIIa).
ARDITTI, Joseph. (S) *Tel-Aviv.* (IIIa).

BARRETT, Brendan. (S) *Dublin.* (I, II).
BASCH, Ernest Frederic. (S) *Haifa, Israel.* (I).
BEAMISH, Desmond Arthur. *Malvern, Worcestershire.* (II, IIIa).
BERLOWITZ, Julius Gad. (S) *Gedera, Israel.* (II).
BLEZARD, Kenneth Brian. (S) *Winchester, Hants.* (II, IIIa).
BRAUN, Simon. (S). *Haifa, Israel.* (I, IIIa).
BRONSTEIN, David. (S) *Tel-Aviv.* (I)

CARR, Eric. (S) *Southampton.* (I).
CHACKAL, Joseph Farouk. (S) *London, W.2.* (II, IIIa).
CLEARY, Alan. (S) *Warrington, Lancashire.* (II).
COCKS, Frank Dennis. (S) *B.A.O.R.* (II).
CRANE, Patrick Herbert. *Send, Surrey.* (IIIa).

DAVIES, Mervyn William. (S) *Johore Bahru, Malaya.* (IIIb).
DESAI, Ramesh Chandra. (S) *London, W.12.* (II).
DICKIN, Frank Douglas. (S) *Bawdsey, Suffolk.* (I, II, IIIa).

GAY, Lionel Stanley. (S) *Durban, South Africa.* (IIIa).
GILVARY, Davis Francis. (S) *Bray, Co. Wicklow.* (II).
GITTINS, Leonard. (S) *Wolverton, Bucks.* (II).
GRIFFIN, David John (S) *Leighton Buzzard, Bedfordshire.* (I).

HABIBULLAH, Mohd. *London, S.W.1.* (IV).
HARUNI, Daniel. (S) *London, N.16.* (I).
HOWETT, Victor Frederick. (S) *Harrow, Middlesex.* (I).
HUNTER, John Hynd. (S) *Kirkcaldy, Fife.* (II).

JASTRZEMBSKI, Jerzy Andrzej. (S) *London, N.7.* (I).

KAPLAN, Zeev. (S) *Givaatayim, Israel.* (II).
KEYMER, Clifford Wilfred. (S) *Cheltenham, Gloucestershire.* (IIIa).
KNITER, Edmund. (S) *London, N.10.* (I).

LEVENE, Arnold Ian. (S) *London, E.1.* (II).

MAHLAB, Ezra Salim. (S) *Jerusalem.* (IIIa).
MASTER, Minoo Ratansha. (S) *London, W.3.* (IIIa).
MEHTA, Mahendra Kumar. (S) *London, W.12.* (I).
MILES, Ronald Boyce. (S) *Lydd, Kent.* (I, II).
MITTELMANN, Mordechai. (S) *Shickun Misrach, Israel.* (II).
MORAWSKI, Wojciech. (S) *London, S.W.4.* (I).
MORDEKHAI, Ephraim Hay. (S) *Ramat Gan, Israel.* (I, II, IIIa).

NELSON-JONES, Laurence. (S) *London, E.17.* (II, IV).

PHAN-CAO-HAN, Louis Jean. (S) *London, W.2.* (II).
PHILLIPS, James Hugh. (S) *London, W.13.* (II, IIIb, IV).

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MICROWAVE SHOT NOISE IN ELECTRON BEAMS AND THE MINIMUM NOISE FACTOR OF TRAVELLING WAVE TUBES AND KLYSTRONS*

by

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SUMMARY

A theorem recently proposed by J. R. Pierce concerning the intrinsic noise of electron streams is verified for electron beams originating in thermionic diodes under arbitrary conditions. The theorem is applied to the calculation of the minimum noise factor of klystrons and travelling wave tubes which is found to be 6 db. The treatment of partition noise is made possible by a generalization of the theorem.

LIST OF SYMBOLS

A, B	Amplitudes of uncorrelated noise waves
B	Bandwidth
C	T.W.T. Gain parameter
$E^*F^*H^*I^*$	Llewellyn Peterson coefficients
E_i	T.W.T. field components
F	Noise factor
I, I_0	Beam current
K	Llewellyn-Peterson parameter (Section 2)
K	Klystron gain parameter (Section 3)
P	Signal power
Q_i	Generalized mean square amplitude of a noise standing wave
T	Ambient temperature 293°K
T_c	Cathode temperature
V, V_0	Beam voltage
V_A, V_t	Induced and thermal noise voltage at klystron buncher (Section 3)
Z_1, Z_2	Buncher and catcher impedances
Z	Helix impedance
e	Electron charge
k	Boltzmann's constant
q	Alternating convection current
m	Electron mass
u	Steady beam velocity
v	Alternating beam velocity
z	Distance along beam
α	Fraction of beam intercepted
β	Propagation constant of helix
γ	Propagation constant of a wave in an electron beam
δ	T.W.T. incremental propagation constant
ϵ	Angle of convergence in an electron beam
ζ	Llewellyn and Peterson's space charge parameter
θ	Transit angle
ρ, ρ_0	Space charge density
χ	Transit angle; phase difference between uncorrelated noise waves in the beam
ω	Angular frequency
ω_p	Angular plasma frequency
Γ_i	T.W.T. propagation constant
Γ^2	Space charge smoothing factor

1. Introduction

It has been known for some time that a small disturbance of frequency $\omega/2\pi$ is propagated in an electron beam by two waves, one travelling slightly slower than the mean velocity of the beam, the other slightly faster.[1] The propagation constants are

$$\gamma = \frac{\omega}{u_0} \pm \frac{\omega_p}{u_0} \qquad \text{(1)}$$

where u_0 is the beam velocity and ω_p is the plasma frequency in the beam. For beams of large transverse dimensions, ω_p is related to the space charge density ρ by

$$\omega_p^{\,2} = \frac{e\rho}{\epsilon m} \qquad \text{(2)}$$

It is thus possible for standing waves of convection current q and velocity modulation v to exist of the general form

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$$q = q \sin \frac{\omega_p z}{u_0} \exp\left(\frac{j\omega t - j\omega z}{u_0}\right) \quad \ldots.(3)$$

In particular such waves may be excited by shot noise. Theoretical treatments of these noise waves have been given by numerous authors and they have been observed experimentally by Cutler and Quate.[2]

Recently Pierce[3] has proposed the following general theorem:—

"If measurements are made in bandwidth B about a frequency ω of the minima and maxima of the noise current standing wave in a beam, carrying current I at a potential V with respect to the cathode at temperature T_c from which the beam originates, then

$$q_{min}\, q_{max} \geqslant (4 - \pi)^{\frac{1}{2}} \frac{\omega}{\omega_p} \frac{kT_c}{eV} eIB \quad \ldots.(4)$$

Pierce has verified that equation (4) is satisfied with the equality sign by an electron beam, rapidly accelerated from a temperature limited cathode, and then allowed to drift at constant potential. He has also proved that if (4) holds for any part of a beam it will continue to do so no matter to what system of accelerating and decelerating fields the beam is subsequently subjected. Thus the method of space charge amplification and reduction considered by Field, Tien and Watkins[4] only results in a change in the noise standing wave ratio. It cannot affect the beam's intrinsic noisiness given by (4).

We shall show in Section 2 that equation (4) is satisfied by an electron beam originating in a planar diode gun of any transit angle and with any degree of space charge. We shall also demonstrate that Pierce's theorem allows us to set a lower limit to the noise factors of travelling wave tubes and klystrons. This will form the subject of Sections 3 and 4. In Section 5 we shall generalise the theorem to include the effects of partition noise considered by Robinson and Kompfner.[5]

2. Pierce's Theorem for the General Diode Gun

The noise current in the beam emerging from a space charge limited diode gun has been considered by Pierce[6] on the assumption that there is full space charge and that the Maxwellian distribution of initial electron velocities can be neglected, so that a single valued analysis such as that of Llewellyn and Peterson[7] can be used.

Robinson has shown that the initial velocity distribution results in a spread of electron transit times from the potential minimum to the anode which is not negligible, and that the use of the single valued analysis is not justified.[8] However, by noting that the major part of this spread arises in a narrow region between the minimum and a plane at a potential of the order of kT_c/e with respect to it, and by assuming that up to this plane the flow is entirely random and that thereafter the single valued analysis is valid, he obtains results very similar to those of Pierce. This method of calculation also demonstrates that at these frequencies the existence of electron beams exhibiting less than the full shot noise is independent of the existence of a potential minimum in the gun.

We shall find that equation (4) is verified whether we adopt Pierce's or Robinson's analysis.

Llewellyn and Peterson[7] have obtained equations:

$$\begin{aligned} q_b &= E^* q_a + F^* v_a \\ v_b &= H^* q_a + I^* v_a \quad \ldots\ldots\ldots\ldots(5) \end{aligned}$$

relating the first order alternating current q_b and velocity v_b at a plane "b" in the diode to their values at an earlier plane "a".

$$\begin{aligned} E^* &= u_b^{-1} [u_b - \zeta (u_a + u_b)] \\ F^* &= K \zeta u_b^{-1} [u_a + u_b] \\ H^* &= -K^{-1} (1 - \zeta) [u_a + u_b]\, u_b^{-1} \\ I^* &= u_b^{-1} [u_a - \zeta (u_a + u_b)] \quad \ldots\ldots\ldots(6) \end{aligned}$$

where u_a and u_b are the steady state velocities at "a" and "b". ζ is the space charge parameter. $\zeta = 0$ for zero space charge, $\zeta = 1$ for space charge limited flow. K is a parameter depending on the transit angle and the frequency.

We may identify the "b" plane with the anode of the diode and express the current and velocity fluctuations there in terms of their values at the "a" plane, which, following Pierce[6] we may identify with the potential minimum or, following Robinson,[8] with a plane near to it. In either case we have

$$\overline{q_a^2} = 2eI_0 B \quad \ldots\ldots\ldots\ldots(7)$$

$$\overline{v_a^2} = \frac{e}{m} \frac{4kT_cB}{I_0} \left(1 - \frac{\pi}{4}\right) \frac{kT_c}{\frac{1}{2}mu_a^2} \quad \ldots.(8)$$

Equation (8) follows from a generalization of Rack's formula[9] given by Robinson, for the mean square velocity fluctuation observed in bandwidth B.

If, at the anode of the diode, i.e. the beginning of the drift region, the convection current is q and the velocity v then[1] the standing wave excited is

$$q = \left\{q^2 + \left(\frac{j\omega\rho_0 v}{\omega_p}\right)^2\right\}^{\frac{1}{2}} \sin\left(\frac{\omega_p z}{u_0} - \chi\right) \quad \ldots\ldots(9)$$

$$\tan\chi = -\frac{\omega_p q}{j\omega\rho_0 v} \quad \ldots\ldots(10)$$

Now v_a and q_a are uncorrelated, we must therefore evaluate the noise standing waves due to each of them separately, and then add their squared moduli in order to find the total noise current. We find for the standing wave due to q_a

$$\overline{q'^2} = \overline{q^2_a}\left[E^{*2} + \left(\frac{j\omega\rho_0 H^*}{\omega_p}\right)^2\right]\sin^2(\theta - \chi') \quad \ldots\ldots(11)$$

$$\tan\chi' = -\frac{\omega_p E^*}{j\omega\rho_0 H^*} \quad \ldots\ldots(12)$$

where we have written

$$\theta = \frac{\omega_p z}{u_0}$$

Similarly the standing wave arising from v_a is given by

$$\overline{q''^2} = \overline{v_a^2}\left[F^{*2} + \left(\frac{j\omega\rho_0 I^*}{\omega_p}\right)^2\right]\sin^2(\theta - \chi'') \quad \ldots\ldots(13)$$

$$\tan\chi'' = -\frac{\omega_p F^*}{j\omega\rho_0 I^*} \quad \ldots\ldots(14)$$

Now if we have a noise standing wave consisting of two components which we may write as

$$\overline{q^2} = A^2\sin^2(\theta - \chi') + B^2\sin^2(\theta - \chi'')$$

we find that

$$\overline{q^2}_{max}\,\overline{q^2}_{min} = A^2 B^2 \sin^2(\chi' - \chi'')$$

Applying this result to the sum of $|\overline{q''^2}|$ and $|\overline{q'^2}|$ and expressing $\sin^2(\chi' - \chi'')$ in terms of $\tan\chi'$ and $\tan\chi''$ we find

$$\overline{q^2}_{max}\,\overline{q^2}_{min} = \left(\frac{\omega\rho_0}{\omega_p}\right)^2 \{E^*I^* - F^*H^*\}^2\,\overline{q_a^2}\,\overline{v_a^2}$$

Noting that

$$\rho_0^2\,\overline{v_a^2} = 2eI_0 B\frac{kT_c}{eV}\left(1 - \frac{\pi}{4}\right)\frac{kT_c}{\frac{1}{2}mu_a^2}$$

and that by equations (6)

$$E^*I^* - F^*H^* = \frac{u_a}{u_b}$$

we find

$$q_{max}\,q_{min} = (4 - \pi)^{\frac{1}{2}}\frac{\omega}{\omega_p}\frac{kT_c}{eV}\,e\,I_0 B$$

Thus Pierce's theorem is verified for the diode gun under arbitrary conditions of space charge and transit angle. Since u_a and hence the position of the "a" plane do not enter into the final result the theorem is also independent of whether we neglect, or allow for, the effect of Maxwellian velocity distribution on the electron flow in the gun. It therefore appears likely that the theorem will prove to be of more general validity than any particular model of the space charge smoothing process.

3. Minimum Noise Factors of Klystrons

We consider a two resonator klystron in which a beam (current I, voltage V) passes through a buncher resonator of impedance Z_1, to which the signal is applied, drifts for a time and then traverses a catcher resonator of impedance Z_2.

We define the impedance Z_1 in such a way that signal power P produces a r.m.s. voltage modulation $V = (PZ_1)^{\frac{1}{2}}$ of the beam. We suppose that the beam has a noise standing wave consisting of two components:

$$\overline{q_a^2} = (4 - \pi)^{\frac{1}{2}}\frac{\omega}{\omega_p}\frac{kT_c}{eV}eI_0 B$$
$$\{A^2\sin^2\theta + B^2\sin^2(\theta + \chi)\} \quad \ldots(15)$$

where $\theta = \dfrac{\omega_p Z}{u_0}$ is the plasma transit angle.

Then Pierce's theorem tells us that:

$$A^2B^2\sin^2\chi = 1 \quad \ldots\ldots(16)$$

Let us consider the noise current at the catcher due to the component A. If θ_1 is the transit angle at the buncher and θ_2 at the catcher, then there will first of all be the direct component $A\sin\theta_2$. There will also be a r.m.s. voltage

$$V_A = -Z_1 A\sin\theta_1$$

induced at the buncher which will cause a velocity modulation

$$v_a = \frac{e}{m}\frac{V_a}{u}$$

where u is the beam velocity. This will set up a further standing wave

$$q_a = j\frac{\omega}{\omega_p}\rho_0 v_a \sin(\theta - \theta_1)$$

and so the mean square current due to A at the catcher will be

$$\overline{q^2_A} = (4 - \pi)^{\frac{1}{2}} \frac{\omega}{\omega_p} \frac{kT_c}{eV} eIB \left[A^2 \sin^2 \theta_2 + K^2 A^2 \sin^2 (\theta_2 - \theta_1) \sin^2 \theta_1\right] \quad \ldots(17)$$

where we have written

$$K = \frac{\omega}{\omega_p} \rho_0 \frac{eZ_1}{mu} = \frac{1}{2} \frac{\omega}{\omega_p} \frac{I}{V} Z_1 \quad \ldots(18)$$

The component B will result in a similar current and so the total noise current at the catcher will be

$$\overline{q^2_n} = (4 - \pi)^{\frac{1}{2}} \frac{\omega}{\omega_p} \frac{kT_c}{eV} eIB \left[A^2 \sin^2 \theta_2 + B^2 \sin^2 (\theta_2 + \chi) + K^2 \{A^2 \sin^2 \theta_1 + B^2 \sin^2 (\theta_1 + \chi)\} \sin^2 (\theta_2 - \theta_1)\right] \quad \ldots(19)$$

Thermal noise in the buncher resonator will result in a voltage

$$V_t^2 = 4kTBZ_1$$

which will produce a current

$$\overline{q_t^2} = 4\left(\frac{\omega}{\omega_p}\right)^2 \rho_0^2 \left(\frac{e}{mu}\right)^2 Z_1 kTB \sin^2 (\theta_2 - \theta_1)$$
$$= 4kTBZ_1^{-1} K^2 \sin^2 (\theta_2 - \theta_1) \quad \ldots(20)$$

Defining the noise factor F in the conventional way, we have

$$F = \frac{\overline{q_n^2} + \overline{q_t^2}}{\overline{q_t^2}}$$

The gain of the klystron will be, from equation (20),

$$G = \frac{Z_2}{Z_1} K^2 \sin^2 (\theta_2 - \theta_1) \quad \ldots(27)$$

We can optimize this expression by putting $\theta_2 - \theta_1 = \pi/2$ and we then can give a simple interpretation to equation (25) for we find $\theta_1 = -\phi$. That is to say, the buncher must be at a noise current minimum. We also find that from equation (22) we must have

$$K^2 = \frac{A^2 + B^2 + \{(A^2 + B^2)^2 - 4\}^{\frac{1}{2}}}{A^2 + B^2 - \{(A^2 + B^2)^2 - 4\}^{\frac{1}{2}}} \quad \ldots(28)$$

This has a minimum value of unity when there is no standing wave on the beam. In order to obtain a large value of the gain we desire as large a value of K as may be achieved within practical limitations. For large values of either A or B, K is approximately given by:

$$K = A^2 + B^2 \quad \ldots(29)$$

Thus, in theory at least, by having a beam with the correct noise standing wave ratio it should be possible to make a klystron having a useful gain and the minimum noise figure which for a cathode temperature of 1,000°K is approximately 6 db.

We may also note that if the noisiness of the beam is in excess of the minimum value allowed

$$F - 1 = \tfrac{1}{2}(4 - \pi)^{\frac{1}{2}} \frac{T_c}{T} K^{-1} \left[\sin^2 (\theta_2 - \theta_1) \{A^2 \sin^2 \theta_2 + B^2 \sin^2 (\theta_2 + \chi)\} + K^2 \{A^2 \sin^2 \theta_1 + B^2 \sin^2 (\theta_1 + \chi)\}\right] \quad \ldots(21)$$

We now assume that this is minimized by making

$$K^2 = \{A^2 \sin^2 \theta_2 + B^2 \sin^2 (\theta_2 + \chi)\} \left[\sin^2 (\theta_2 - \theta_1) \{A^2 \sin^2 \theta_1 + B^2 \sin^2 (\theta_1 + \chi)\}\right]^{-1} \quad \ldots(22)$$

Giving

$$F - 1 = (4 - \pi)^{\frac{1}{2}} \frac{T_c}{T} \left[\frac{\{A^2 \sin^2 \theta_2 + B^2 \sin^2 (\theta_2 + \chi)\} \{A^2 \sin^2 \theta_1 + B^2 \sin^2 (\theta_1 + \chi)\}}{\sin^2 (\theta_2 - \theta_1)}\right]^{\frac{1}{2}} \quad \ldots(23)$$

We find using Pierce's theorem that this has a minimum value

$$F - 1 = (4 - \pi)^{\frac{1}{2}} \frac{T_c}{T} \quad \ldots(24)$$

when $(A^2 + B^2) \cos (\theta_2 - \theta_1) = \{(A^2 + B^2)^2 - 4\}^{\frac{1}{2}} \cos (\theta_1 + \theta_2 + 2\phi) \quad \ldots(25)$

where

$$\tan 2\phi = \frac{B^2 \sin 2\chi}{A^2 + B^2 \cos 2\chi} \quad \ldots(26)$$

by Pierce's theorem the minimum noise factor will be given by

$$F - 1 = AB \sin \chi \, (4 - \pi)^{\frac{1}{2}} \frac{T_c}{T} \quad \ldots(30)$$

which will be achieved when, from equation (25),

$$(A^2 + B^2) \cos (\theta_2 - \theta_1) = \{(A^2 + B^2)^2 - 4A^2B^2 \sin^2 \chi\}^{\frac{1}{2}} \cos (\theta_1 + \theta_2 + 2\phi) \quad \ldots(31)$$

It will be observed that the noise behaviour of the klystron depends on four parameters A, B, χ and the zero from which we measure θ. Measurements such as those of Cutler and Quate[2] in which the noise is measured at different points along the beam with a resonator yield three parameters:

$$A^2 + B^2 = (q^2{}_{max} + q^2{}_{min}) \left\{(4-\pi)^{\frac{1}{2}} \frac{\omega}{\omega_p} \frac{kT_c}{eV} eI_0B\right\}^{-1} \quad \ldots\ldots(32)$$

$$A\,B \sin\chi = q_{max}\, q_{min} \left\{(4-\pi)^{\frac{1}{2}} \frac{\omega}{\omega_p} \frac{kT_c}{eV} eI_0B\right\}^{-1} \quad \ldots\ldots(33)$$

and $-\phi$ the phase angle at a current minimum. Only in the special case when $\theta_2 - \theta_1 = \pi/2$ are these parameters sufficient to specify the behaviour of the klystron.

The reason for this is that in a klystron, and also in a travelling wave tube, we effectively make simultaneous measurements of the noise at more than one point in the beam. The phase relations between the coherent components of the noise at these points are therefore of importance and these phase relations are not measured in experiments which only sample the noise at one point at a time. It is for this reason that it is not permissible to represent the sum of two standing waves by one standing wave together with a constant component.

The special case when the catcher and buncher are $\pi/2$ apart in transit angle arises, therefore, because measurement made at two points of this separation are insensitive to the particular decomposition of the noise standing wave into its components.

4. The Minimum Noise Factor of a Travelling Wave Tube

Throughout this section we shall follow the treatment and notation of Pierce[11] and we shall assume the reader to be familiar with that paper.

The operation of the T.W.T. is described in terms of three normal modes. Associated with each mode is a field strength E_i and a propagation constant Γ_i. The power flow in each mode is given by

$$P_i = \beta^{-2} E_i E_i^* Z^{-1} \quad \ldots\ldots\ldots\ldots\ldots(34)$$

which serves as a definition of the helix impedance Z. β is the propagation constant of the helix in the absence of the electrons and the three propagation constants are given by

$$\Gamma_i = j\beta - \delta_i \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots(35)$$

where the three incremental propagation constants for the waves in the presence of the electrons are given, for the special case of a lossless helix and equal beam and wave velocities, by

$$\delta_1 = \tfrac{1}{2}\{\sqrt{3} - j\}\beta C$$

$$\delta_2 = \tfrac{1}{2}\{-\sqrt{3} - j\}\beta C$$

$$\delta_3 = j\beta C \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots(36)$$

C the gain parameter is defined by

$$C^3 = \frac{I_0 Z}{8V_0} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots(37)$$

E_1 and δ_1 correspond to the growing wave.

At the beginning of the helix the boundary conditions are

$$\sum E_i = E$$

$$\sum \frac{E_i}{\delta_i} = -\frac{u_0 v m}{e}$$

$$\sum \frac{E_i}{\delta_i^2} = \frac{2V_0}{j\beta I_0} q \quad \ldots\ldots\ldots\ldots\ldots\ldots(38)$$

where E is the impressed signal, v the velocity modulation of the beam, q the convection current modulation and u_0 the beam velocity.

Putting $E^2 = \beta^2 ZkTB$ and $v = q = 0$, we find for the amplitude of the growing wave excited by thermal noise in the input circuit

$$E_t^2 = \frac{8V_0\beta^2 C^3 kTB}{I_0 \left|\left(1 - \frac{\delta_2}{\delta_1}\right)\left(1 - \frac{\delta_3}{\delta_1}\right)\right|^2} \quad ..(39)$$

We now consider the growing wave excited by a noise wave in the beam of the form

$$q = q \sin\theta$$

$$v = -j\frac{\omega_p q}{\omega\, \rho_0} \cos\theta \quad \ldots\ldots\ldots\ldots(40)$$

and we find

$$E_1{}^2 = \frac{2C^2 V\omega_p{}^2(\cos^2\theta + \sqrt{3}X\cos\theta\sin\theta + X^2\sin^2\theta)}{I_0{}^2 \frac{e}{m}\left|\left(1 - \frac{\delta_2}{\delta_1}\right)\left(1 - \frac{\delta_3}{\delta_1}\right)\right|^2} q^2 \quad \ldots\ldots\ldots\ldots(41)$$

where we have written

$$X = \frac{\omega}{\omega_p} C \quad \ldots\ldots(42)$$

There will also be a second uncorrelated component of the noise standing wave in the beam and we will again write the total noise in the normalized form

$$\overline{q_n^2} = (4-\pi)^{\frac{1}{2}} \frac{kT_c}{eV} \frac{\omega}{\omega_p} eI_0 B \{A^2 \sin^2\theta + B^2 \sin^2(\theta+\chi)\} \quad \ldots(43)$$

Since E_1^2 is a mean square field while E_t^2 is a peak field, the noise factor is given by

$$F - 1 = 2\frac{E_1^2}{E_t^2}$$

This yields

$$F - 1 = (4-\pi)^{\frac{1}{2}} \frac{T_c}{T} X^{-1}[A^2\{\cos^2\theta + \sqrt{3}X\cos\theta\sin\theta + X^2\sin^2\theta\} + B^2\{\cos^2(\theta+\chi) + \sqrt{3}X\cos(\theta+\chi)\sin(\theta+\chi) + X^2\sin^2(\theta+\chi)\}] \quad \ldots(44)$$

We now use Pierce's theorem and put $B^2 = (A^2 \sin^2\chi)^{-1}$. We then minimize F with respect to A, obtaining

$$F - 1 = 2(4-\pi)^{\frac{1}{2}} \frac{T_c}{T} X^{-1}(\sin\chi)^{-1}[\{X^2\sin^2(\theta+\chi) + \sqrt{3}X\sin(\theta+\chi)\cos(\theta+\chi) + \cos^2(\theta+\chi)\}\{X^2\sin^2\theta + \sqrt{3}X\sin\theta\cos\theta + \cos^2\theta\}]^{\frac{1}{2}} \quad \ldots(45)$$

for $$A^2 = (\sin\chi)^{-1}\left\{\frac{X^2\sin^2(\theta+\chi) + X\sqrt{3}\sin(\theta+\chi)\cos(\theta+\chi) + \cos^2(\theta+\chi)}{X^2\sin^2\theta + X\sqrt{3}\sin\theta\cos\theta + \cos^2\theta}\right\}^{\frac{1}{2}} \quad \ldots(46)$$

After some very tedious algebra we find that (45) has a minimum value with respect to θ when

$$\sin(2\theta + 2\psi + \chi) = -\frac{X^2+1}{(X^4+X^2+1)^{\frac{1}{2}}}\cos\chi \quad \ldots(47)$$

where $$\tan 2\psi = \frac{1-X^2}{X\sqrt{3}} \quad \ldots(48)$$

This minimum value is then

$$F - 1 = \frac{T_c}{T}(4-\pi)^{\frac{1}{2}} \quad \ldots(49)$$

We may note that from (47) it will only be possible to achieve this minimum if

$$(X^4+X^2+1)^{\frac{1}{2}} \geqslant (X^2+1)\,|\cos\chi|$$

When $X = 1$, which is a value readily obtainable in practical tubes, the equations become rather simpler and we find that (46) reduces to

$$\frac{A^2}{B^2} = \frac{1-(1-\frac{1}{4}A^2B^2)^{\frac{1}{2}}}{1+(1-\frac{1}{4}A^2B^2)^{\frac{1}{2}}} \quad \ldots(50)$$

and that this equation implies that a minimal noise factor may only be attained if

$$A^2B^2 < 4,\ 1 > \frac{A^2}{B^2} > 0{\cdot}072 \text{ and } \frac{\pi}{2} > \chi > \frac{\pi}{6}$$

It is interesting to see how a large departure from optimum conditions will affect the noise factor. Suppose that $X = 1$, $A^2 = 20$, $B^2 = 1$, $\sin^2\chi = 0{\cdot}05$, then we find that if θ is optimized

$$F = 1 + 2{\cdot}8(4-\pi)^{\frac{1}{2}}\frac{T_c}{T}$$

corresponding for $T_c = 1{,}000^\circ$K to a noise figure of 10 db instead of the minimum value of 6 db.

Since noise figures for travelling wave tubes of 8·5 db have been quoted in the literature,[12] it would appear that there is now little room for improvement.

We conclude this section by noting that if the noisiness in the beam is in excess of the minimum predicted by Pierce's theorem then the minimum noise factor will be increased in the same ratio (see equation (3)). We have already shown that A, B and χ cannot be determined from the results of experiments which only measure the noise at one point in the beam at a time, and, except in the trivial case when $A = B = \sin\chi = 1$ so that there is no standing wave, there will be no special case, such as there is for the klystron, when the results of such experiments will be sufficient to predict the behaviour of a T.W.T. exactly.

5. Partition Noise and a Generalization of Pierce's Theorem

Robinson and Kompfner[5] discuss three main types of partition noise which are of importance in microwave tubes involving electron beams:

(i) Induced partition noise due to non-uniform coupling of the beam to the external circuit.

(ii) Direct partition noise due to interception of part of the beam current by electrodes.

(iii) Velocity partition noise due to a spread of velocities, in excess of that due to the thermionic origin of the beam, produced by electron optical defects of the gun.

Their treatment is based on that of Thomson, North and Harris which is only valid at low frequencies when space charge smoothing is a function of the action of the potential minimum near the cathode as an electron gate. However, Kompfner[14] has shown that whatever the mechanism of smoothing, similar results will be obtained. His argument, which we now reproduce, is as follows.

Electrons are not only emitted with a distribution of velocity in the direction of the flow, but also with a distribution of transverse velocities. Thus, after the electrons have travelled a sufficient distance, which in nearly all practical cases is negligible, an electron in any one part of the beam may initially have come from any other part of the beam. As a result of this the number of electrons in any sufficiently small part of the beam will be quite random, although the beam as a whole may show a considerable degree of order. This order will be due to correlations between the fluctuations in the several parts of the beam. Now this is exactly the state of affairs which subsists at lower frequencies. We may therefore take over the Thomson, North and Harris theory of partition noise and its extension by Robinson and Kompfner, even though our model of space charge smoothing is quite different.

We shall not here be concerned with induced partition noise as this does not affect the intrinsic noisiness of the beam but merely produces additional noise in the external circuit. Robinson and Kompfner[5] show that direct interception of a fraction α of a beam at a point where the noise current is

$$q^2 = \Gamma^2 \, 2eI_0B \quad \text{(51)}$$

introduces an additional noise current

$$q_p^2 = \alpha(1 - \Gamma^2)\, 2eI_0B \quad \text{(52)}$$

For small Γ^2 this reduces to

$$q_p^2 = \alpha \, 2eI_0B \quad \text{(53)}$$

and this will excite a noise standing wave

$$q_p^2 = \alpha \, 2eI_0B \sin^2(\theta - \chi_3) \quad \text{(54)}$$

where $\chi_3 + \pi/2$ is the plasma transit angle co-ordinate of the point where interception takes place. q_p^2 will be uncorrelated with any other noise existing in the beam.

Velocity partition noise arises when, due to electron optical defects, electrons at different points in the cross-section of the beam have different velocities in the direction of the flow. Such a spread may, for example, arise when an electron beam is caused to converge in a uniform magnetic field by an electrostatic lens. In this case if ϵ is the angle of convergence of electrons on the edge of the beam the resulting spread in velocities is expressed by

$$\overline{\Delta u^2} = \overline{u^2} - (\bar{u})^2 = (\bar{u})^2 \tfrac{1}{48}\epsilon^4 \quad \text{(55)}$$

We may show by a slight modification of the argument given by Robinson[8] that the existence of a velocity spread $\overline{\Delta u^2}$ in an electron beam, in which the number of electrons with any particular velocity is random, results, whether the beam as a whole is random or not, in an equivalent velocity modulation in bandwidth B

$$\overline{v^2} = \frac{2eB}{I_0}\overline{\Delta u^2} \quad \text{(56)}$$

Using equation (9) this will excite a noise standing wave

$$\overline{q_v^2} = \left(\frac{\omega}{\omega_p}\right)^2 \rho_0^2 \bar{v}^2 \sin^2(\theta - \chi_4)$$

$$= 2eI_0B\left(\frac{\omega}{\omega_p}\right)^2 \frac{\overline{\Delta u^2}}{u^2}\sin^2(\theta - \chi_4) \quad \text{(57)}$$

where χ_4 is the plasma transit angle co-ordinate of the point at which the velocity spread first occurs. This current standing wave, also, is uncorrelated with the noise due to other causes.

We see, therefore, that each of these kinds of partition noise will add an additional component standing wave to the total noise.

We now generalize Pierce's theorem to apply to a beam in which the noise is due to an arbitrary number n of standing waves.

$$\overline{q_n^2} = \sum_{i=1}^{n} Q_i \sin^2(\theta - \chi_i) \quad \text{(58)}$$

We can easily show that

$$q^2_{max} + q^2_{min} = \sum Q_i \quad \text{(59)}$$

and

$$q^2_{max}\, q^2_{min} = \sum_{i>} Q_i Q_j \sin^2(\chi_i - \chi_j) \quad \text{(60)}$$

From (60) we may deduce that

$$q^2_{max}\, q^2_{min} \geqslant Q_1 Q_2 \sin^2(\chi_1 - \chi_2) \left\{1 + \frac{1}{Q_1 + Q_2} \sum_{j>2} Q_j\right\} \quad \ldots .(61)$$

If we now identify Q_1 and Q_2 with the intrinsic noise components due to the thermionic origin of the beam we obtain our generalization of Pierce's theorem.

To illustrate the application of these results to partition noise let us suppose that we have a beam in which the thermal noise components are

$$Q_1 = 2(4-\pi)^{\frac{1}{2}} \frac{kT_c}{eV} \frac{\omega}{\omega p} e I_0 B = 2N$$

$$Q_2 = 10\,N$$

$$\sin^2(\chi_1 - \chi_2) = 0{\cdot}05$$

$$\omega/\omega p = 20 \qquad kT_c/eV = 10^{-4}$$

and we have direct partition noise due to interception of a fraction α of the beam at the current minimum.

We find that the value of q^2_{min} is $N/12$ in the absence of partition noise and that with it it is

$$\frac{N}{12} + 2\alpha \frac{\omega p}{\omega} \frac{eV}{kT} N = \frac{N}{12} + 10^3 N \alpha$$

If α is as little as 0·01 this results in an increase of over a hundredfold in the current minimum and, therefore, the intrinsic noisiness of the beam is increased at least tenfold. Similarly, the introduction of a small amount of velocity partition noise at a current maximum will have equally serious effects.

6. Conclusion

The validity of Pierce's theorem for an electron beam originating in a planar diode of arbitrary transit angle, under any degree of space charge and independently of the particular model of the space charge smoothing process has been demonstrated. As a result of the theorem it has been shown that the minimum noise factor of any klystron or T.W.T. is not less than 6 db. By a generalization of the theorem a treatment of partition noise by the same method has been outlined.

7. Acknowledgments

The author is indebted to Dr. J. R. Pierce of the Bell Telephone Laboratories for the advance information about his new theorem which led to the writing of this paper, and for his kindness in letting him see the manuscript of a further paper in which the minimum noise factor of travelling wave tubes is derived.

This work was completed during the tenure of a Nuffield Research Fellowship.

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# of current interest . . . .

Exports of Radio Equipment

The report of the Radio Communication and Electronic Engineering Association for 1953 states that exports of broadcast transmission, radio communication and navigational aid equipment exceeded all previous figures, the annual rate in 1953, according to Customs and Excise figures, being just short of £11 million for direct exports, the value of indirect exports being estimated at £4 million. The comparative figures for 1952 were £8 million and £3 million respectively.

Transatlantic Telephone Cable

It was recently announced by the Postmaster-General that an agreement had been signed by the American Telephone and Telegraph Company, the Canadian Overseas Telecommunication Corporation, the Eastern Telephone and Telegraph Company, and H.M. Post Office for the establishment of a transatlantic submarine telephone cable. The proposed cable would be laid between Oban in Scotland and Newfoundland and thence to Nova Scotia.

The most important engineering feature about the cable, of course, is the fact that it will have to incorporate a considerable number of submarine repeaters, and this will be the first occasion on which deep-sea repeaters have been used on such an extensive scale. The cable will be designed to carry 29 telephone circuits to the United States, six to Canada, as well as a number of telegraph circuits.

The deep-sea repeaters are to be supplied by the American company, while the repeaters between Newfoundland and Nova Scotia are to be of a shallow-water type designed by the British Post Office. The cable will be laid by H.M.T.S. *Monarch*, which is the largest cable-laying ship in the world. It is understood that the last link in the chain to the United States border will be a micro-wave radio relay system, and, at some point on this link, a connection will be made to the Canadian telephone network.

The agreement between the participating organizations extends for 25 years, and it is estimated that the total cost will be £12½ million.

Some of the problems associated with the design of submarine telephone repeaters were briefly described in a 1951 Convention paper by Dr. E. W. Smith (published in the *Journal* for October 1951, pages 463-468).

American Research

A recent publication, "Science in U.S.A., 1953,"* draws attention to principal features of scientific effort in the United States. The report has been written by the British Commonwealth Scientific Office in Washington, which makes an annual review of American scientific progress.

Federal expenditure on research during 1951-52 was $1,800 million, and during 1952-53 it was expected to be $2,200 million; 15 per cent. of this amount pays for work at non-profit institutions, 60 per cent. for work at profit-making organizations and only the remaining 25 per cent. for research in government stations.

Since 1900 the proportion of the population of the United States occupied in scientific and professional work has doubled; the total number is now about 5 millions and over half a million are engineers. There are about 155,000 workers in the natural sciences, about 15,000 of them engaged on fundamental research. American authorities consider, nevertheless, that there still is a shortage of some kinds of scientific manpower; for example, 80,000 more engineers could be used.

Shortages of men in individual branches of science have been difficult to estimate, and are as much in quality as in quantity. Frequently there is a lack of a few individuals of the highest calibre. Wasteful employment of professional manpower also seems to exist, two engineering societies having drawn attention to waste in their particular field. The complaints were that too many engineers were doing work that did not exercise their engineering talents to the full, and that others who were doing the work for which they were trained were being hampered by routine duties.

A new and significant trend in the industrial pattern of the U.S. is the increasing industrialization of the South. Plans to spend more than $700 million on new plants in Texas, Louisiana, Oklahoma and Arkansas were announced during the year. The reasons for this change are the abundance of raw materials in the South, together with plentiful petroleum and natural gas and low-cost water transportation to markets in the East.

Other subjects dealt with in the report include radio astronomy, v.h.f. television broadcasting, mechanical engineering, metallurgy, building, etc.

* "Science in U.S.A., 1953," published by H.M.S.O. for the D.S.I.R., price 2s. (50 cents U.S.A.), by post 2s. 1½d.

APPLICANTS FOR MEMBERSHIP

New proposals were considered by the Membership Committee at a meeting held on January 12th, 1954, as follows: 10 proposals for direct election to Graduateship or higher grade of membership and 18 proposals for transfer to Graduateship or higher grade of membership. In addition 31 applications for Studentship registration were considered. This list also contains the names of three applicants who have subsequently agreed to accept lower grades than that for which they originally applied.

The following are the names of those who have been properly proposed and appear qualified. In accordance with a resolution of Council and in the absence of any objections being lodged, these elections will be confirmed 14 days from the circulation of this list. Any objections received will be submitted to the next meeting of the Council, with whom the final decision rests.

Transfer from Associate Member to Full Member

PARSONS, David Reginald. *Newcastle-upon-Tyne.*

Direct Election to Associate Member

BROWN, Arthur, B.Sc. *Durham.*

GOLDSMITH, Michael Norton Grenier, M.A. *Newport Pagnell, Bucks.*

SUBRAMANIAN, Chinnaswamy Rajam, B.Sc. *Dehra Dun, India.*

Transfer from Associate to Associate Member

BIRAM, John Graham Swinton. *Abingdon, Berkshire.*

Transfer from Graduate to Associate Member

JAMES, Antony Bracher. *Welwyn Garden City, Hertfordshire.*

Transfer from Student to Associate Member

BORUM, Per Anton. *Windhoek, South-West Africa.*

LOVELL, Frank Ward. *Zomba, Nyasaland.*

RAMABHADRAN, Srinivasa, M.A. *Poona, India.*

Direct Election to Associate

NICHOLLS, Leslie Robert. *Shrewsbury.*

ROGERS, Capt. Donald Thomas Chaplin. *Arborfield, Berkshire.*

Transfer from Student to Associate

HEDLEY, Thomas William. *Pinner, Middlesex.*

Direct Election to Graduate

BOND, Martin Edward. *Worcester Park, Surrey.*
HATTON, Terence John Peter. *Kingston-upon-Thames.*

Transfer from Student to Graduate

BISPING, Piotr. *London, N.22.*
BRISTOW, Hubert Ronald. *London, N.W.3.*

CLARK, Thomas Gordon. *East Molesey, Surrey.*

DUNLOP, John Alexander. *Blantyre, Lanarkshire.*

EVANS, Hugh Maitland. *Newbridge, Monmouthshire.*

JAGANNATHA RAO, Baji Ramachandra, B.Sc.(Hons.), *Madras.*

MACMULL, Jacob Isaac, B.Sc. *Byculla, Bombay.*

RAMAGOPALA RAO, Pemmaraju, B.Sc., M.Sc. *Bharkatpura, Hyderabad.*

RICKETTS Peter William. *Loughborough, Leicestershire.*

TEMBE, Sakharam Bhaskar, B.Sc. *Bombay.*
THREADINGHAM, Roy Lawrence. *East Southsea, Hants.*

Studentship Registrations

AMIR CHAND GAUR. *Delhi.*
ARMITAGE, Major Neville John. *London, S.W.7.*

BAKISCH, Hanna. *Jaffa, Israel.*
BALWANT SINGH, Sangha. *Jullundur Cantt., India.*
BHANAGE, Prabhakar Gangadhar. *Poona, India.*
BHURANI, Dindayal Harismal. *Deolali, Bombay.*

CHARANJIT SINGH, Chhabra. *Delhi.*

ELIAHOU, Emile. *London, W.11.*

INDER, James Haviland. *Waimauku, New Zealand.*
IQBAL AHMAD, Syed. *Karachi.*

JAIN, Naim Chand, B.A.(Hons.), *New Delhi.*
JAIN, Om Prakash. *Borivali, Bombay.*
JANGRA, Ram Narayan. *Ranchi, Bihar, India.*
JEZIERSKI, Eugeniusz Czeslaw Jan. *London, N.4.*
JHA, Chandra Shekhar, B.Sc.(Hons.). *Edinburgh.*

KANWAR, Sri Inder Singh. *Jaipur, India.*

MALHOTRA, Chaman Lal. *Poona, India.*
MILNER, Geoffrey. *Sheringham, Norfolk.*
MITCHELL, Roy Herbert. *Woodford Green, Essex.*

PREM NATH. *New Delhi.*

RAGHAVACHARI, Colathur Pathangi, M.A. *Madras.*
REES, Heath Watkin, B.Sc. *Coventry, Warwickshire.*
RILEY, J. Maxwell. *Glasgow.*

SATYANARAYANA MURTY, Challa, B.Sc. *Chromepet, Madras.*
SEN, Birendra Mohan, B.Sc. *Calcutta.*
SHARMA, D. N., B.A. *Agra, India.*
SUBRAMANIAN, T. S. *Bombay.*

THAMPI, Sridhar Ramavarma Srikumaran, B.A. *New Delhi.*

UPASANI, Capt. Bhaskar Purushottam. *Poona, India.*

VARGHESE CHERIAN. *Thiruvalla, India.*
VENKATESWARAN, Vadakanthara Venkatachalan. *Julahalli, Bangalore.*
VENKATESWAR RAO, Gundimada. *Madras.*

SOME MEASUREMENTS OF FADING AT A WAVELENGTH OF 8 MM OVER A VERY SHORT SEA PATH*

by

D. G. Kiely, M.Sc.†

SUMMARY

The results of a pilot experiment designed to investigate the magnitude of atmospheric refraction effects in the 8-mm wavelength band are described. Records of signal fading on a 1-mile over-sea path with a low transmitter and a high receiver indicate that these effects are very large.

1. Introduction

That the effects of vertical gradients of refractive index in the lower atmosphere on the propagation of microwaves will become more marked as the wavelength decreases has been appreciated in a general manner for some years. It has been shown by Booker and Walkinshaw[1] that ray bending and partial and complete duct trapping may be expected to become increasingly pronounced as the wavelength varies from the centimetre band, which has been well examined experimentally in this respect, to the millimetre band which is at present relatively unexplored. Booker and Walkinshaw say that a state of semi-permanent super-refraction for wavelengths of the order of 1 cm is expected to exist over the sea due to a semi-permanent duct of the order of 10 to 20 ft in depth which was shown by Wust[2] in 1920 to exist over Northern European waters and which is associated with evaporation. This prediction was fulfilled by later experimental work and, indeed, the evaporation duct is several times deeper in other sea areas than northern European waters, giving similar effects in the 10-cm wavelength band.

Several isolated observations by workers in the 8-mm wavelength band indicated the presence of strong refraction effects, and, during some rain clutter measurements[3] on the west coast of Scotland, the present author obtained certain recordings which could be explained in terms of fairly complete duct trapping. In this case a separate T. and R. c.w. radar set operating in the 8-mm wavelength band was situated on the coast some 15 ft above sea level with its aerials inclined upwards at an angle of some 5 deg to enable the main beams of the radiation patterns to clear hills on the opposite side of Loch Goil approximately a mile distant. On several occasions when the weather was calm and warm in the afternoons a strong signal 83 db below the transmitter level was obtained when there was no rain nor visible source of reflection. This signal

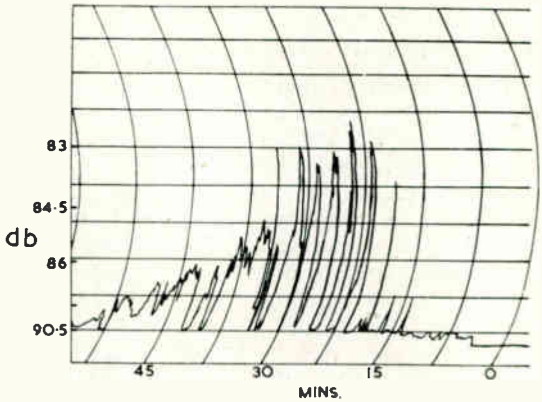


Fig. 1.—Chart record of an anomalous echo obtained with separate T and R radar equipment.

usually lasted for about an hour, rising from noise level to a maximum and falling away again. A slow and very deep fluctuation of period two to three minutes was superposed on the signal. An example is shown in Fig. 1. A possible explanation of this effect is that the radiation was being trapped fairly completely in a low duct over the sea and the energy was being directed at and not over the hills which were causing the

* Manuscript received December 14th, 1953. (Paper No. 252.)

† Royal Naval Scientific Service.

U.D.C. No. 621.392.2.029.65.

reflection. Such a mechanism could account for the very strong received signal.

It was decided to establish a short one-way link across Loch Goil and record the signal fluctuations to see if any evidence of duct effects was apparent which would manifest itself as fading of the received signal. To obtain this effect it was considered that a low transmitter site about 15 ft above mean sea level and a high receiver site about 100 ft above mean sea level would serve, as there was a good chance that the transmitter would be within any low ducts which formed while the receiver would be above duct level, and so trapping would be shown as fading on the received signal. The variation of sea level due to the tide was approximately 5 ft during the period of these measurements.

For reasons of convenience of supervision this investigation was carried out in the vicinity of the rain clutter measurements at Loch Goil. It was conducted for only five weeks in October and November, 1952, and was regarded purely as a pilot experiment which did, however, provide some highly indicative and interesting observations.

2. Description of Equipment and Site

The equipment used was very simple. The receiver was a crystal feeding a tuned amplifier operating at 3·2 kc/s and the transmitter was a klystron modulated at the same frequency. A crystal and amplifier, similar to the receiver, were used as a transmitter monitor and ink recorders were operated from the amplifiers of the receiver and monitor. Calibration of the receiver was carried out using a calibrated r.f. attenuator. The aerials in both cases were 2 ft-diameter lenses fed by small conical horns giving 3 db beam widths of 1 deg, and these were protected from the weather by Onazote windows which were, when dry, transparent to 8-mm wavelength radiation.

The equipment stability with the klystron working from a stabilized power supply was, in general, very good. The transmitter monitor provided a record of variations in the valve power which were noted on the receiver chart; for periods of several hours the klystron output varied by only a small fraction of a decibel.

A site was found at Loch Goil in which the receiver was approximately 100 ft and the transmitter 15 ft above mean sea level. Loch Goil is a narrow land-locked loch approximately ½ mile wide with steeply sloping hills coming down to the water's edge. The maximum height of these hills is some 1,800 ft. and there are many lower saddles. The link erected was across this loch between terminals which were not directly opposite each other. Unfortunately this site was only 1 mile long which meant that only the strongest super-refraction effects would be detected. The angle of elevation of the transmitter, aligned on the receiver, was a little greater than 1 deg, hence with 3-db aerial beam widths of 1 deg it may be safely assumed that the sea was not illuminated except by the sidelobes and that the signal arriving at the receiver was effectively independent of sea reflection.

3. Experimental Observations

During the five weeks that this link was maintained the principal general characteristics of the received signal were as follows:

(*a*) Signal characteristics were markedly dependent on weather conditions and changed considerably and frequently with weather conditions. During stormy weather the signal was very constant but frequently when the wind dropped or died the signal immediately began to fluctuate by ± 1 db to ± 2 db.

(*b*) Signal scintillation occurred on several occasions and was associated with periods of dry weather with little or no wind. The extent of the scintillation was less than 1 db as would be expected on a short path of this kind.

(*c*) Fading occurred very frequently and varied in depth from 2 db to 16 db and in duration from 15 minutes to several days. The deep fades were always associated with calm, warmish weather conditions and the shallow fades occurred in rather more varied weather conditions which included light and moderate winds.

(*d*) During rain the signal dropped due to attenuation by the rain in the ray path and due to attenuation by the rain forming a water film on the Onazote windows. The latter gave sometimes very considerable attenuation of some 20 db which was produced unequally by the two windows depending on the direction of the wind and the amount of water on the Onazote sheets.

It was, however, noticed on many occasions that the signal fall anticipated the rainfall by 10 or 15 minutes; before the rain actually fell the signal dropped by one or two decibels as though an associated temperature or humidity effect preceding the rain was affecting refraction conditions in the ray path. Such a change of the vertical distribution of humidity before a rain shower is indeed likely to occur. Whether the considerable falls in signal level during rain were due entirely to water attenuation in the ray path and on the Onazote or partly to refraction fading as a continuation of the initial drop in signal level is difficult to say.

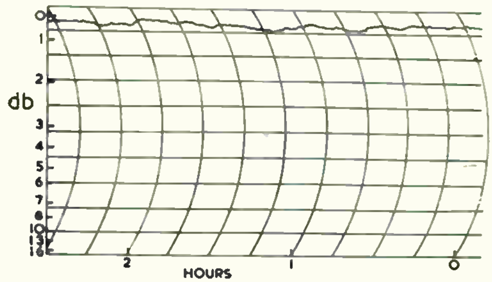
The significant aspect of these observations of signal fluctuations, particularly the deep fades, is that they were recorded on such a short path. On a longer path of several miles one would expect on this basis considerably greater fluctuations when the ray path in the stratified medium is larger.

TABLE 1

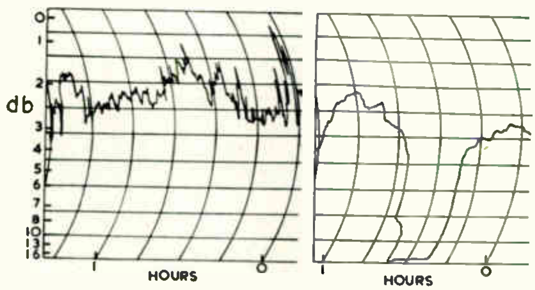
Depth of Fade, db	Duration	Remarks
2	1 hr	Weather calm and dull, overcast
2	8 hr	Weather cold with a light wind
4½	3 hr	Weather cold with a light wind
4	½ hr	Weather calm, sunny, clear blue skies with a light breeze
4	½ hr	Weather calm, sunny, clear blue skies with a light breeze
13½	15 min	Weather calm, warmish and moist with occasional very light breeze
10	15 min	Weather calm, warmish and moist with occasional very light breeze
15	30 min	Weather calm, warmish and moist with occasional very light breeze
16	1 hr	Weather calm, warmish and moist with occasional very light breeze
3	1 hr	Bright sunny weather, with scattered showers, cool, light wind
3	1 hr	Bright sunny weather, with scattered showers, cool, light wind
5½	15 min	Weather calm, dull and overcast
2	30 min	Weather calm, dull and overcast
7	6 days	Uniform flat calm frosty weather for this time
1½	6 hr	No wind, calm, cool night
1½	3½ hr	Dull, overcast, no wind

The principal characteristics of the signal fades not caused by rain are given in Table 1.

The contribution to attenuation by water vapour in the atmosphere, at even saturation density, is less than 0·1 db for this path; thus water vapour attenuation can have no significant effect on the fading figures.



(*a*)



(*b*) (*c*)

Fig. 2.—Typical chart records showing :

(*a*) *Normal steady signal*

(*b*) *Scintillating signal*

(*c*) *Deep fade recorded on a one-way link.*

Examples of typical signal characteristics are shown in Fig. 2.

(*a*) shows a steady constant signal associated with wind conditions producing a well-mixed atmosphere,

(*b*) shows signal scintillation, and

(*c*) gives an example of a deep fade associated with calm weather.

It is suggested that the most probable mechanism causing these fades is a surface or elevated duct which captures the transmitted signal with varying degrees of completeness. The 1-deg angle of elevation of the beam and the fact that practically all the energy is concentrated in a narrow beam of 1 deg are favourable conditions for trapping. The situation would be probably quite different if the two terminals of the link were the same height, either low or high, as the ray bending would not affect the signal in quite the same manner.

4. Conclusion

The principal importance of these results is that they show the major part played by atmospheric effects on the propagation of millimetre waves. Although this was expected in general terms from the results of previous theoretical work the order of magnitude of the effects was largely obscure.

5. Acknowledgments

This paper is published by permission of the Admiralty and of the Controller, H.M. Stationery Office.

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