JOURNAL OF

The British Institution of Radio Engineers

(FOUNDED IN 1925—INCORPORATED IN 1932)

"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

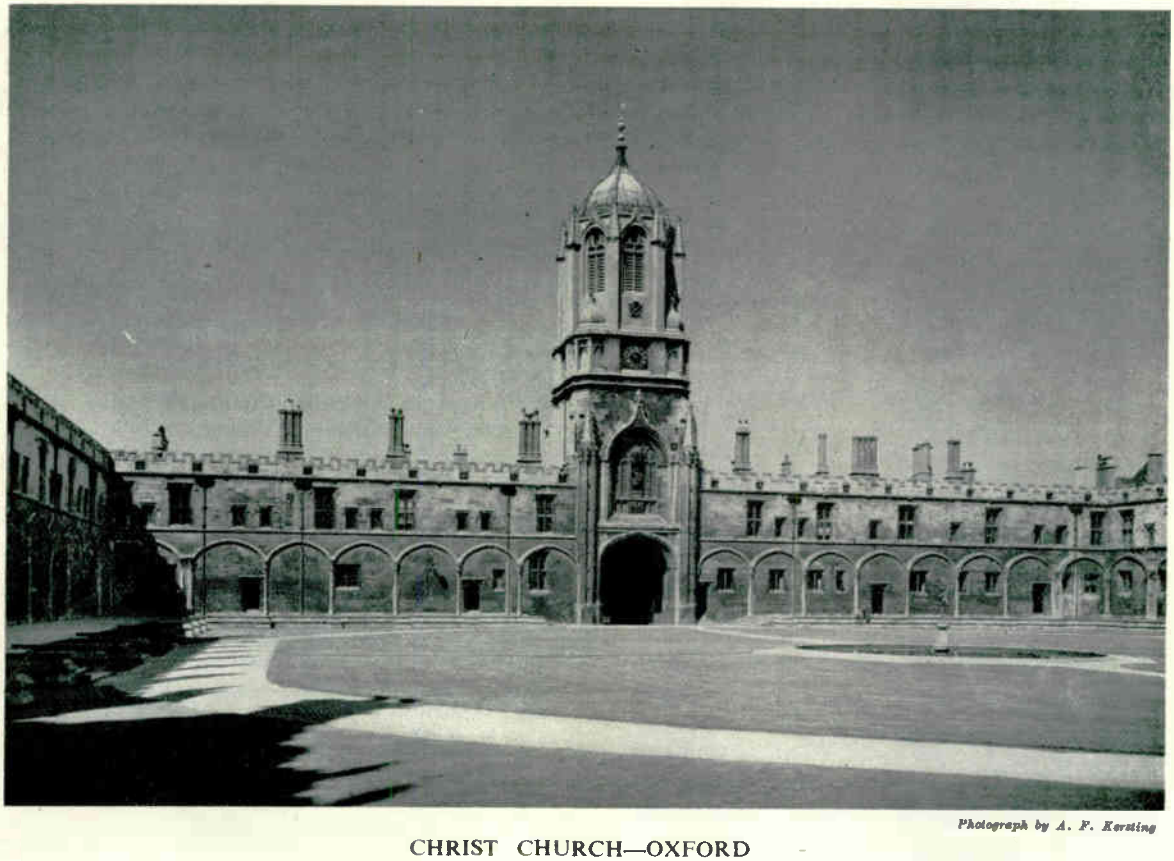
Vol. XIV. No. 4 APRIL 1954

THE 1954 CONVENTION VENUE

Christ Church was founded by King Henry VIII in 1532, although the buildings actually date from 1525 when Cardinal Wolsey began to build a new college at Oxford. The picture below shows the Great Quadrangle, completed in 1681, and "Tom Tower," named from the bell "Great Tom" which now hangs in it.

Students of Christ Church have given to the world many famous books, perhaps one of the best known of them being "Alice in Wonderland," written by Lewis Carroll whilst he was in residence.

Accommodation will be available in Christ Church for Brit.I.R.E. delegates, and the lectures will be delivered in the Clarendon Laboratory Theatre and the Electrical Laboratory Theatre.



Photograph by A. F. Kersting

CHRIST CHURCH—OXFORD

INDUSTRIAL ELECTRONICS CONVENTION

Selection is now being made of about 30 papers for presentation at the 1954 Convention. It is intended that all papers selected will be issued to delegates in preprint form and the final arrangements for each session will be published in the May and June Journals and circulated to all who are attending the Convention.

The Clerk Maxwell Memorial Lecture will be given by Sir John Cockcroft, K.C.B., C.B.E., F.R.S., on Thursday, July 8th, at 8 p.m. This will be an open meeting and the Lecture will be delivered in the Clarendon Laboratory Lecture Theatre, Parks Road, Oxford.

Details are given below of some of the papers *provisionally* selected for presentation during the five formal sessions.

Thursday, July 8th (afternoon)

SESSION 1—INDUSTRIAL APPLICATIONS OF ELECTRONIC COMPUTORS

Chairman: L. H. Bedford, O.B.E., M.A., B.Sc. (Past President)

1 "Statistical Computors as Applied to Industrial Control"
by P. Huggins, A.M.Brit.I.R.E.

2 "Electronic Digital Computors for Commerce and Industry"
by R. L. Michaelson, F.I.A.

3 "The Function of Leo"
by T. R. Thompson and J. M. M. Pinkerton.

(*Note—A number of other papers on analogue and digital computors are still under consideration.*)

Friday, July 9th (morning)

SESSION 2—INDUSTRIAL APPLICATIONS OF X-RAYS AND SONICS

Chairman: H. G. Foster, M.Sc., M.Brit.I.R.E.

1 "Some Typical Circuits for Industrial X-ray Apparatus"
by J. Jeremy Bliss, B.Sc.(Eng.).

2 "An X-ray Thickness Gauge for the Measurement of Hot-Rolled Strip Steel"
by F. H. Gottfeld and D. Tidbury.

3 "Industrial X-ray Fluoroscopy with Particular Reference to Ultra-fine Focus X-ray Equipment"
by E. W. Kowol.

4 "Electronics in Automatic Direct Reading Spectrometers"
by F. Holmes.

5 "A Method of Ultrasonic Gauging"
by F. M. Savage.

6 "The Analysis of Binary Gas Mixtures by a Sonic Method"
by E. W. Pulsford, B.Sc., A.M.Brit.I.R.E.

Friday, July 9th (afternoon)

SESSION 3—NUCLEONIC INSTRUMENTATION AND APPLICATION

Chairman: N. C. Robertson, C.M.G., M.B.E., M.Brit.I.R.E.

1 "The Alpha Gauge"
by E. N. Shaw, B.Sc.

2 "The Scintillation Counter in Industry"
by J. S. Eppstein, B.Sc.(Eng.), A.C.G.I.

3 "Nucleonic Instruments in Industry"
by Denis Taylor, M.Sc., Ph.D.

4 "Multi-electrode Counting Tubes"
by K. Kandiah.

5 "A Combined Beta and Di-Electric Gauge"
by R. Y. Parry.

Saturday, July 10th (morning)

SESSION 4—ELECTRONIC SENSING DEVICES (TRANSDUCERS)

Chairman: Professor E. E. Zepler, Ph.D. (Vice-President).

1 "Optical Transducers and Some Applications in Industrial Production"
by John A. Sargrove, M.Brit.I.R.E.

2 "Photometric Test Techniques using Barrier-Layer Cells"
by R. W. J. Cockram.

3 "Piezo-electric Vibration Pick-ups"
by S. Kelly.

5 "The Measurement of Pressure"
by J. L. Thompson (Vice-President).

Saturday, July 10th (afternoon)

SESSION 5—PROCESS CONTROL

Chairman: J. L. Thompson (Vice-President).

1 "The Automatic Indication and Recording of Minute Concentrations of Organic Gases in Air"
by H. A. Thomas, D.Sc.

2 "Electronic Welding Controls"
by C. R. Bates, A.M.Brit.I.R.E.

3 "Electronics in the Tin-plate Industry"
by M. N. Lapper.

THE HISTORY OF THE HOMODYNE AND SYNCHRODYNE*

by

D. G. Tucker, D.Sc., Ph.D. (*Member*)

SUMMARY

The Homodyne and Synchrodyne are systems of demodulation for amplitude-modulated signals; they use a local oscillator, synchronized in frequency to the carrier of the wanted signal, to modulate the received wave. Thus the wanted signal is obtained immediately as an audio output without further detection, and unwanted signals are left on carriers of the difference-frequency between their original carrier and that of the wanted signal. There is no detection of the unwanted signals since the whole signal path can be kept free of non-linearity. Unwanted signals can thus be completely rejected by a low-pass filter in the audio output. The receiver consequently has the advantages of high quality combined with high selectivity, and is inherently a demodulator of precision. The principles can be extended to permit separation of signals whose sidebands overlap, and they also lead to improved detection of pulse signals in the "coherent detector."

The history of the systems over the last 30 years is outlined in this paper. Although they are of great interest, they appear never to have had any commercial exploitation until recent application in colour television.

1. Introduction

In 1947 the author published a short article[1] on the "Synchrodyne" radio receiver. As this application was only a by-product of other work, no attempt was made to elaborate the article before publication. However, in view of the interest it evoked, further articles[3, 4, 5] were published, and interest spread to many other countries, numerous articles[21–35] being published by British and foreign authors, some based entirely on the original ones, but some containing new work. For the most part the Synchrodyne was attributed to the present author, although it can be seen from the published items of correspondence[11–20] that there was some criticism and several claims of anticipation.

When the first articles were published, the author was aware of the Homodyne system only from the one paper published by Colebrook[36] in 1924, and Colebrook's system was certainly not the same as the synchrodyne. The author therefore felt fully justified in coining the new name "Synchrodyne" and applying for a patent. Even now, after reasonably full searches have been made, only a few other articles[42, 47, 53] have come to light. But these articles, together with many patent specifications cited by the Patent Office and by several correspondents, make it perfectly clear that there had been a development of ideas under the "generic" title of Homodyne which had culminated in a system absolutely the same as that which was published under the name Synchrodyne. The synchrodyne, therefore, cannot be regarded as the invention of the present author.

In view of the facts set out above, it seems very desirable that the full history of the homodyne should now be published. The account which follows should be reasonably complete, although it cannot be claimed with certainty that *all* published and patent information has been obtained.

2. The Original Homodyne

The circuit arrangement of the original homodyne, as published by Colebrook[36] in 1924, is shown in Fig. 1. This is nearly as old as the superheterodyne, which appears to date from 1918. It is an oscillating detector, the anode and grid coils being tightly coupled so that the valve circuit oscillates gently in the absence of input signal. When the input is applied, the tuning is adjusted until the harsh beat-note disappears; this is the condition of synchronization, and the modulation signal is heard clearly. Colebrook points out the benefit which is obtained from the fact that this system gives effectively linear (i.e., distortionless) rectification. However, it is clear that the frequency response of the highly regenerative tuned circuit (so highly regenerative that it

* Manuscript received January 4th, 1954. (Paper No. 259.)
U.D.C. No. 621.396.59.

oscillates) is imposed on the received r.f. signal, and it is inevitable, therefore, that there will be frequency-response distortion of the audio output, since most of the higher frequency sidebands will have been attenuated. Colebrook states that "the quality of the reception was very full and rich in tone", but it is likely that this was due to the relatively enhanced bass response, which may have been advantageous with the headphones and loudspeakers used in those early days. He points out also that the amount of input signal relative to the local oscillation amplitude is critical, and that it may be necessary to reduce the latter, or to provide a r.f. amplifying stage to increase the former.

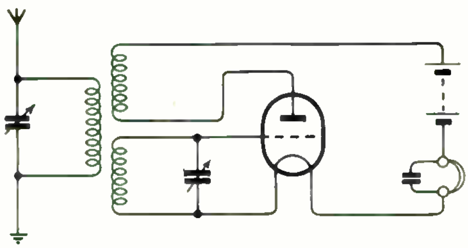


Fig. 1.—Colebrook's homodyne receiver.

Following this article, a letter[37] was published in the same journal under the initials F.G.G.D., and this referred to Appleton's earlier paper[54] on the synchronization of oscillators and to the fact that this method of reception had been in use for some time at Cambridge. This letter also pointed out that the demodulation was not equally effective over the whole frequency band of the synchronized condition, but that the output was zero at the edges and a maximum in the middle of this band. Colebrook had rather inferred that the output was constant over the band, but this may not have been intentional.

The homodyne in this form had the same circuit arrangement as the "autodyne", which was an "automatic heterodyne" receiver of c.w. signals, having its oscillating detector tuned to a frequency different from the incoming carrier by an amount sufficient to give a good heterodyne note. It is clear that the relation between the homodyne and the autodyne did not extend beyond their circuit diagrams.

One cannot now avoid the suspicion that in those days of reaction coils and tuning whistles, homodyne reception was almost certainly used far more often by accident than by design!

3. Carrier-Reinforcement Systems

In that the homodyne is a kind of carrier-reinforcement system, it is important to notice that the advantages of carrier reinforcement had been appreciated rather earlier than the first announcement of the homodyne. In March 1922, Robinson had applied for a patent[38] for a system in which the carrier of the incoming signal is filtered out in a path separate from the main signal path, and amplified in a high-Q regenerative tuned amplifier before being re-combined with the original input signal, which has been amplified if necessary by a low-Q tuned amplifier. The combined signal is then detected in the usual way. Fig. 2 shows the basis of the system.*

The importance of this patent—and of another[39] by the same inventor, applied for in August 1930—is that it shows an appreciation of how important it is to reinforce the carrier in such a way that highly tuned stages are not used in the main signal path. In this way, the attenuation of the higher modulation frequencies, which was a weakness of the homodyne, can be avoided.

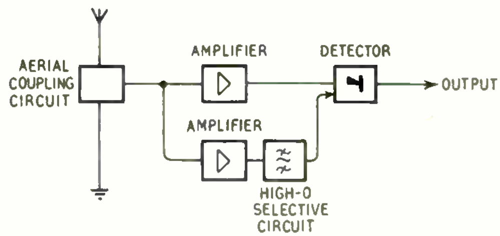


Fig. 2.—Robinson's carrier-reinforcement system.

The second patent referred to above, while not basically different from the first, is interesting in that in Claim 6 it is stated that the carrier-wave tuned stage may even feebly oscillate; no advantages are claimed for this condition, but

* Diagrams given in this paper to illustrate patents are not, in general, copied from the patent specification, but are modified by the author to show his interpretation of the system. The method of operation is often not clear in the patent.

it is clearly identical with the improved homodyne or synchrodyne to be discussed later. No appreciation of the significance of the synchronization requirement is evident.

A much more recent treatment of carrier reinforcement is given by Crosby,[60] in 1945.

4. Improved Homodyne

A system was patented by a Frenchman named de Bellescize,[40] with a convention date of November 1930 (U.K. application date November 1931), and also published in a very full and thorough paper[42,43] in 1932, in which the advantages of carrier reinforcement combined with the homodyne scheme of obtaining the enhanced carrier from a synchronized local oscillator seem to be fully recognized. The synchronization in this system is, however, not obtained by the injection of the input signal into the oscillator as described by Appleton and used by Colebrook, but by a more complicated method using a separate valve as a control impedance. The local oscillator, as shown in Fig. 3, is applied with a large oscillation amplitude to one grid of a two-grid valve, to the other grid of which the input signal is applied. The difference-frequency between the local and carrier frequencies (which is ideally zero, i.e., d.c.) is applied to the grid of the control valve in such a way that it varies the impedance produced by the control valve across the tuned circuit of the oscillator in a manner which leads to the attainment and maintenance of exact synchronization. This method is nowadays quite well known.

A second patent application[41] in June 1932, by the same inventor, merely clarifies some of the points of the first, and was not accepted as a complete specification by the British Patent Office.

Another system which seems to be a sort of homodyne is described by Walton[44] in a patent applied for in December 1930. This patent specification is a rather difficult one; the method of operation is not made very clear. However, the system uses a local oscillator synchronized to the incoming carrier to produce an oscillation which fully loads a balanced valve circuit: the balance is upset by the application of the incoming signal, which must be of opposite phase to the local oscillation. This seems very much like the use of a balanced modulator in place of the usual "detector" valve in the previous homodynes. The r.f. circuits have very flat tuning. The advantages of this system are not fully set out, and the inventor does not appear to be aware of the homodyne as previously published, or alternatively considers his scheme to be quite different.

Another form of homodyne receiver was described by Reimann in 1932[45], but the author has been unable to obtain details of this.

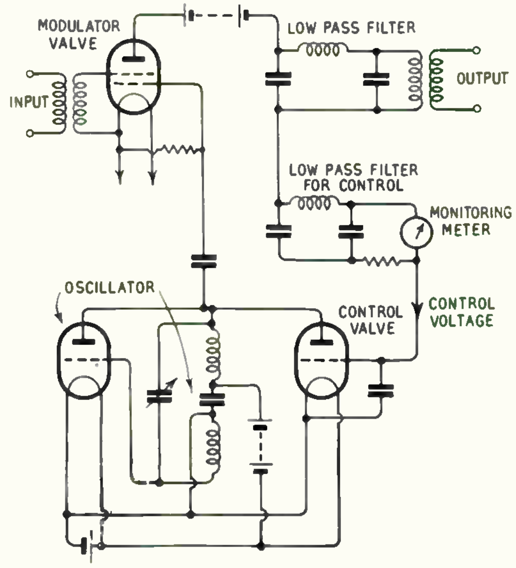


Fig. 3.—Bellescize's homodyne arrangement.

A system which was in all essential respects identical to the "Synchrodyne" described by the present author in 1947 was patented by a German named Urtel[46] in German Patent No. 670,585, applied for in December 1932. A British application for a patent based on the same specification was made in February 1934 but this did not mention the inventor's name. A switched linear modulator and synchronized local oscillator are specified, although the method of synchronization is not absolutely clear—the local oscillator is "triggered" by the carrier wave. However, this can hardly mean anything other than synchronization by direct injection, on Appleton's basis.

Some further analysis of the signal discrimination in a homodyne system was published by Groszkowski[47] in 1933.

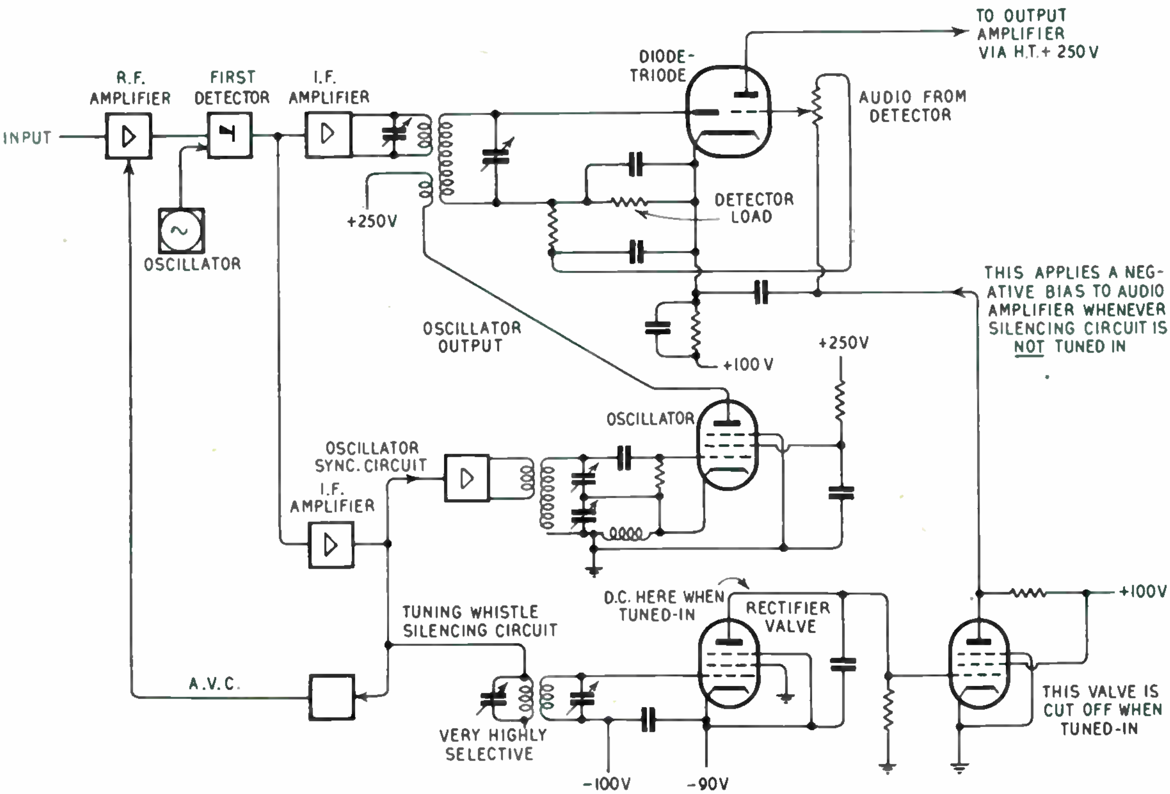


Fig. 4.—Jarvis's homodyne circuit.

A very detailed homodyne design is described by Jarvis[48] in U.S. Patent No. 2,166,298, applied for in November 1933. Synchronization is by plain injection, and the local oscillation is added to the incoming signal at a very high level. The homodyne is used as an i.f. stage in this design, and this appears to be the first record of its use in this way. Automatic volume control is provided in the usual way—i.e., the output of the i.f. amplifier is rectified to provide the control bias. For the first time, too, an effort is made to remove the objectionable tuning whistle which occurs when the local oscillator is not properly synchronized. However, this is accomplished at the expense of using a highly tuned stage preset at the wanted carrier, the rectified output of this being used as a muting bias. This arrangement, if it is really supposed to work this way, seems hardly an advance, since if this stage can be so accurately preset, why not the whole receiver? Fig. 4 shows one of Jarvis's circuit arrangements.

An interesting version of the Homodyne is a single-valve circuit described by Starnecki[49] in Polish Patent No. 28655, applied for in August 1934. The one valve serves as local oscillator, r.f. amplifier, and modulator. The circuit arrangement is shown in Fig. 5. It is doubtful if this circuit gives enough control of factors such as synchronization to be satisfactory over any range of conditions.

Although normally an effort is made to obtain linear action in the receiving circuit of a homodyne, Oltze[51], in a German patent of 1938, claims additional selectivity by the use of suitably non-linear demodulation. This has not been investigated further.

One further homodyne scheme was patented by Curtis,[52] an American, with a convention date (U.S.A.) of March 1939 and a British application date of February 1940. This arrangement has not only the desirable features of a separate path for the synchronized oscillator and a balanced modulator for the signal demodulation, but also the features, later described for the synchrodyne by Garlick,[5] of using both injection and a reactance valve method for synchronization—i.e., using both Appleton's and de Bellescize's principles. Although the advantages of this combination are not described in the patent, it is, in fact, a very effective combination, as shown by Garlick.[59] Fig. 6 shows the circuit arrangement, and it will be seen that a.v.c. is provided by using the rectified input signal—not a very suitable method if strong unwanted signals are present. An important feature of the patent is that it reports the production of non-linear distortion due to the phase-modulation of the synchronized oscillation by sidebands of the input signal; this effect was analysed by the present author and a colleague[7,8] in 1949-50.

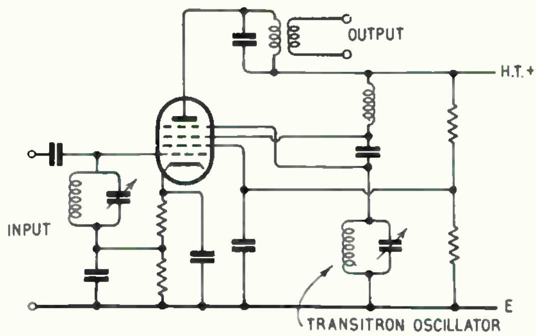


Fig. 5.—Starnecki's single-valve homodyne.

Finally there was the important, but anonymous, article[53] in 1942 that gave a very lucid and well-reasoned discussion of the position and prospects of the homodyne, although it gave no references to previous work.

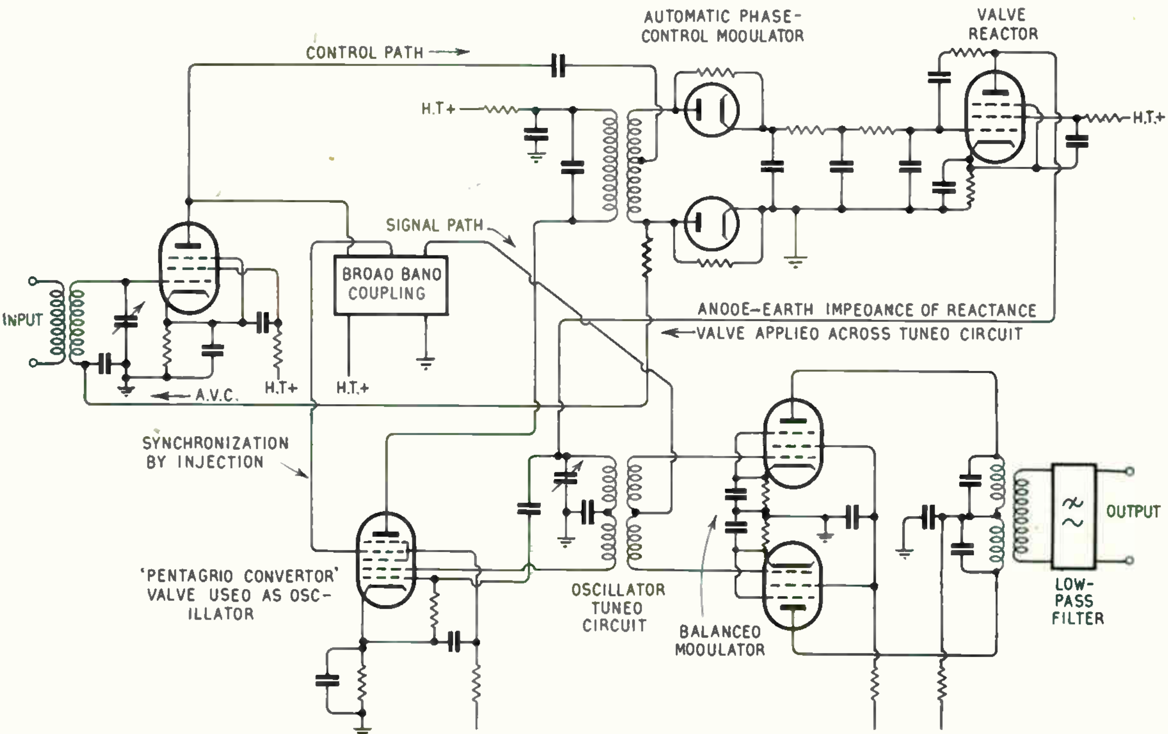


Fig. 6.—Curtis's homodyne receiver.

5. The Synchrodyne

The simple synchrodyne arrangement published by the present author in his first 1947 article[1] was as shown in block schematic form in Fig. 7. A typical circuit arrangement published later is shown in Fig. 8. This is clearly quite different from Colebrook's original homodyne, and the name "Synchrodyne" was chosen for it as being very descriptive of its method of operation. However, it is quite clear from the account of the history of the homodyne given above that the original homodyne had been gradually improved until almost all the desirable features mentioned by the present author had been incorporated; the synchrodyne was, in fact, only an up-to-date homodyne. The true position may well be as nicely summarized by a French journal,[35] which reported (literal translation) as follows:—"It is possible that the principle of the synchrodyne is not new, but it is, to our knowledge, the first time that this principle has been applied in practice and that a report of practical trials has been given".

However, these questions of originality interested only the very small minority, and did not affect the enthusiasm which was shown by the great majority of interested readers, both amateur and professional. But although this enthusiasm gave the author a great deal of work (and some satisfaction), it did not lead to any important developments in the technique of synchrodyne or homodyne reception. The letters and articles which were published by various people in several countries did not disclose any previously unrecorded principles, and there is therefore no point in examining them here; they are, however, listed in the Bibliography.[11-35]

6. Tuning-Whistle Suppression

One new feature which the author incorporated in one of his designs[4] was an effective tuning-whistle suppressor. When the synchrodyne is being tuned-in and the oscillator is not quite synchronized, a very harsh beat-note is produced which approaches zero frequency as synchronism is reached. It is desirable, for comfort in operating the receiver, that this "tuning-whistle" should be considerably attenuated, although not entirely eliminated, since it is a great help in guiding the operator to the synchronizing band. This result was achieved by utilizing the fact that, when the oscillator is synchronized, a d.c. output is obtained from the carrier component of the signal; but when synchronism fails, this output disappears. A network consisting of series and shunt rectifiers is inserted in the audio output, as shown in Fig. 8, and the standing bias is such that in the absence of synchronism the series rectifier presents a high resistance and the shunt rectifier a low resistance. Thus a considerable loss is inserted in the output circuit, 30 db being easily achieved with suitable rectifiers. When synchronism is attained and a d.c. is produced, this d.c. opposes the standing bias of the rectifiers, and causes the series rectifier to become a low resistance and the shunt rectifier to become a high resistance; thus no loss is produced in the output circuit while synchronism is maintained. The only drawback to this device is that a small amount of nonlinearity is introduced into the audio output by the rectifier network unless the depth of modulation of the r.f. signal is very small; it is therefore advisable to have a switch to cut out the network when the required signal is satisfactorily tuned in.

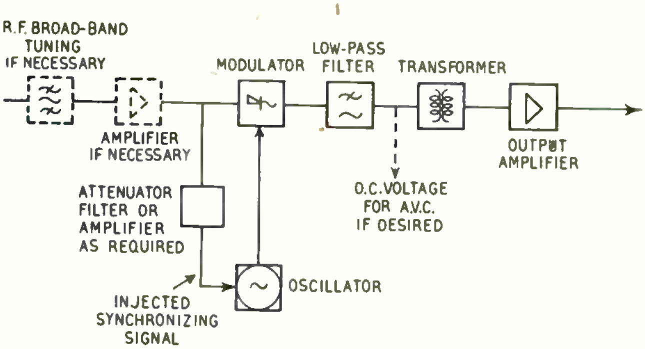


Fig. 7.—Schematic diagram of "Synchrodyne" system of reception.

It should be noted that this whistle-suppressing circuit will operate on any signal which may be selected provided the amplitude is sufficient to produce a d.c. large enough to overcome the bias on the rectifiers.

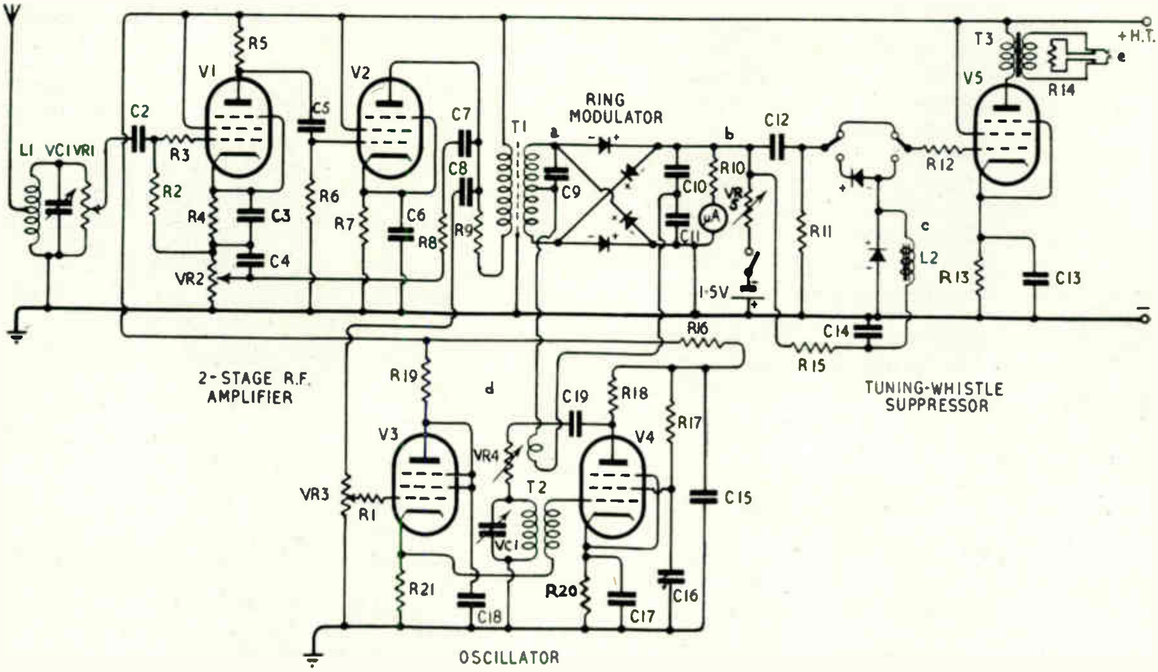


Fig. 8.—Typical synchrodyne circuit including tuning-whistle suppressor.

R1 100Ω	R12 100k	VR2 500Ω	C7 100p	C18 0·1μ
R2 100k	R13 150Ω	VR3 50k	C8 100p	C19 0·01μ
R3 82Ω	R14 1k	VR4 250k	C9 300p	L1 Input coil
R4 150Ω	R15 4·7k	VR5 50k	C10 0·01μ	L2 L.F. choke
R5 10k	R16 4·7k	VC1 500p+500p (ganged)	C11 0·01μ	T1 Modulator transformer
R6 100k	R17 22k	C2 0·01μ	C12 1μ	T2 Oscillator transformer
R7 150Ω	R18 10k	C3 0·01μ	C13 50μ	T3 Output transformer (~4:1 turns ratio)
R8 22k	R19 4·7k	C4 0·01μ	C14 0·05μ	V1–V5 5P41
R9 4·7k	R20 390Ω	C5 100p	C15 0·1μ	Rectifiers Germanium
R10 3k	R21 200Ω	C6 0·01μ	C16 0·01μ	
R11 4·7k	VR1 20k		C17 0·01μ	

Notes on letters in diagram

(*a*) The value of C9 should be adjusted if possible to give the minimum leakage of the oscillator output back into the r.f. amplifier.

(*b*) This meter serves as a tuning indicator, and to check the operation of the network C.

(*c*) Tuning-whistle suppressing network. The choke L2 should have upwards of 100 henrys inductance and will probably need a magnetic screen to avoid 50-c/s pick-up.

(*d*) Cathode-follower in synchronizing path to make oscillator frequency independent of setting of VR3.

(*e*) Output to earphones—or to power stage for loud-speaker direct from anode.

7. The Synchrodyne as a Precision Demodulator

As has been previously stated, the author's interest in the synchrodyne was not primarily in its application to radio reception; it was, in fact, more specifically in its application as a very accurate demodulator for highly-selective transmission-measuring equipment.[2, 6, 10] This accuracy arises largely due to the facts that the centre of the frequency band selected is determined almost entirely by the carrier to which the local oscillator synchronizes, and that the shape of the band is determined by the low-frequency filter in the output. In ordinary circuits, where the filtration is done at a relatively high i.f., neither the shape nor the centre of the band selected is easily made exactly what is desired. To enhance this inherent accuracy, the author and his colleagues investigated the cause and cure of distortion due to the synchronizing process when the natural frequency of the local oscillator* and the injected carrier

* The natural frequency is that frequency at which the oscillator oscillates when the synchronizing control is removed.

frequency are not exactly equal; this distortion arises as a result of phase-modulation in the oscillator. A reasonably full theoretical study of this effect has been published,[7,8] and it has been shown that the main result is harmonic distortion on the output of modulation frequency and an error in the d.c. output if this is retained. Since the phase modulation in the oscillator is due to the amplitude modulation of the carrier, there is no such effect if the synchrodyne is used merely to select a tone from a spectrum of unwanted signals, the wanted output being then a d.c. voltage. It was subsequently discovered that this distortion effect had been reported by Curtis[52] in a patent specification previously discussed, but he gave no analysis of it.

To avoid the distortion discussed above it is evidently necessary to prevent the natural frequency of the oscillator from drifting very far from the frequency to which it is made to synchronize; this can be done by what is really an automatic phase-control arrangement.[5] A reactance valve is used to readjust the natural frequency of the oscillator according to a d.c. voltage obtained by modulating the oscillator output with the injected carrier. The direction of the control is such that the d.c., which is approximately proportional to the phase difference between the oscillator and the carrier, is brought towards zero; and this condition means that the natural frequency of the oscillator is made very nearly equal to the actual frequency at which it is synchronized. This method of control was also described by Curtis in his patent.

8. Separating Overlapping Amplitude-Modulated Signals

Although the ordinary conception of separating signals in a spectrum by means of filters leads to the idea that when the upper sideband of one modulated signal overlaps the lower sideband of another signal, then interference between the signals is inevitable, this is, in fact, far from true. In the simplest case, for instance, when the overlap is small, interference can be avoided by sacrificing that part of the modulation frequency band corresponding to the overlap—i.e., the higher-frequency part. Even better, if r.f. filters are used which accept each signal except the overlapping portion, then on detection the whole modulation frequency band of each signal is obtained separately, the only distortion* being a falling-off of up to 6 db over that portion of the band where only one sideband has been accepted. This can easily be corrected if desired, but will often be unimportant.

However, when the overlap is considerable, and especially when the sideband of one signal overlaps the carrier of the other, r.f. filtration is impracticable or useless. In such cases, the synchrodyne technique can be used to obtain theoretically perfect separation of the two signals. Two methods have been published, one by Gabrilovitch[50] as a patent specification with a convention date of September 1936 and

* This statement assumes that the non-linear distortion which can occur in the detection of unsymmetrical sidebands when the depth of modulation is large, is negligible.

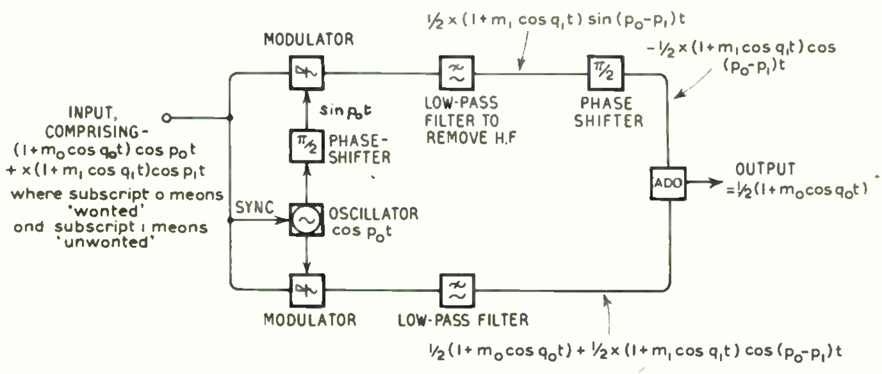


Fig. 9.—Separating overlapping sidebands: method based on Gabrilovitch.

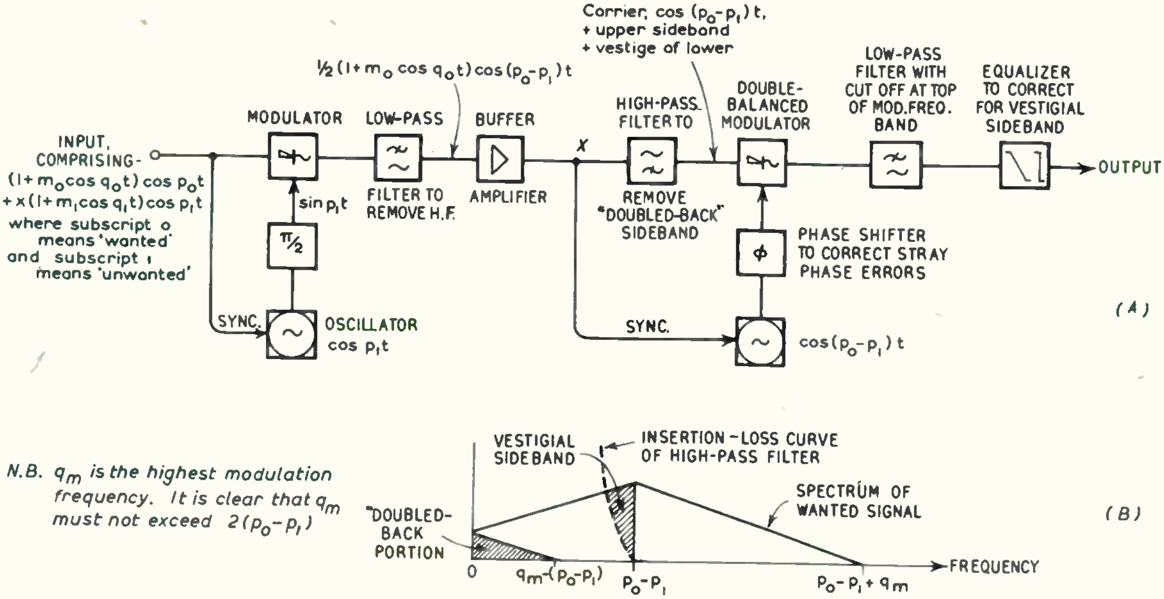


Fig. 10.—Separating overlapping sidebands (Author's method)
(a) Block schematic. (b) Spectrum at point X.

a date of acceptance of April 1939, and the other by the present author,[5] in 1948. Both methods use the synchrodyne principle in the first stage as a rejector instead of the usual acceptor. Gabrilovitch's system rejects, in this stage, the signal which is ultimately wanted; the author's system rejects the unwanted signal. Unfortunately, Gabrilovitch's patent is very difficult to understand in detail, so the account given here is really the author's personal interpretation of the main principles, and the suggested schematic arrangement does, in fact, differ considerably from that given by Gabrilovitch; the latter, in the author's opinion, will not work. This may account for the fact that Gabrilovitch claims that his system will separate signals whose sidebands overlap the adjacent carrier by a large amount, whereas the author thinks that Gabrilovitch's system, even as re-arranged, will not work unless the overlap of the sidebands is restricted so that they do not overlap the adjacent carrier. On the other hand, in the author's own scheme, the spacing between the carriers of two adjacent signals need not exceed one-half of the highest modulation-frequency, although this system has its own limitation in that it can deal with only one overlapping signal at a time.

The author's version of Gabrilovitch's system is shown in block schematic form in Fig. 9, which should be largely self-explanatory. The first synchrodyne path acts as a rejector because the local oscillation is in quadrature with the incoming signal which is to be rejected. The second path accepts all incoming signals. One of the paths is then given a phase shift of 90 deg, and on combining the two paths suitably, all signals cancel out except for the wanted signal which was contained in only one path. This signal is undistorted and free of the signals which overlapped it. It is clear that the wanted signal must not overlap the carrier of any adjacent signal, because, if it did, the spectrum of the lower sideband of the output of the low-pass filter in the first path would be doubled back on itself, and the 90 deg phase shift would have the wrong sign for the doubled-back portion.

In the author's own scheme, shown in outline in Fig. 10, the signal rejected by the first stage is actually the unwanted signal whose spectrum overlaps the wanted signal. The wanted signal then remains on a carrier equal to the difference between the carriers of the wanted and unwanted signals. If its modulation-frequency band is greater than this difference, then the lower sideband is doubled-back on itself. This signal is demodulated by first filtering out the doubled-back band with a high-pass filter, and then applying the remaining signal (which is now a vestigial-sideband signal) to a synchrodyne stage, arranged as an acceptor, with the oscillator synchronized, in phase, to the new carrier frequency. The output of this stage is the wanted modulation-frequency band, complete, but with a small distortion of frequency response due to the inclusion of the contributions from two sidebands over the very low-frequency part of the band. This can be equalized if desired.

Synchrodyne principles can also be used to enable a system to be made in which two different channels of information occupy the same frequency band and use the same carrier frequencies. The channels are separated by virtue of the fact that their carriers, although of the same frequency, are in phase quadrature. If the carriers are transmitted, then at the receiver a local oscillator can be synchronized to their resultant, and from this, two quadrature carriers can be obtained by means of phase-shifters for use in demodulating the two channels into separate output circuits. The basic principles of this "carrier-phase duplex" or "two-phase" system were set out (except for the synchrodyne application) by Nyquist[57] in 1928, and a fuller discussion of it was given by the present author[58] in 1948. It is used in the American N.T.S.C. colour television system[61] on the synchrodyne basis.[62]

9. The Coherent Detector: Detection of Pulse Signals

In all the published work by other people on the homodyne receiver, one particularly important idealization is made, namely, it is always assumed that the local oscillator gives a pure oscillation even though it is synchronized to a component of a complex input spectrum. It has been pointed out by the present author in several papers[55, 56] that the local oscillation will not, in general, be a pure tone of the frequency of the wanted carrier. Many other, unwanted, components of the input spectrum will be present, although, if the circuit is correctly designed and adjusted, these will be largely discriminated against, and the local oscillation can usually be made substantially pure while retaining adequate synchronization. Some analysis of the mechanism of the discriminating properties of a synchronized oscillator has been published,[56] and all that need be given here is a reminder that the difficulties of synchronizing are the ultimate limitations of the performance of the synchrodyne system.

However, sometimes it can be arranged that a pure tone of the wanted carrier frequency is available, and then this tone can be used (in the correct phase relationship) to operate the frequency-changer of what is otherwise a synchrodyne receiver. Moreover, since the local oscillation is no longer dependent on the received carrier for synchronization, the carrier can be intermittent—i.e., the receiver can be used for pulse signals. This arrangement has become known as the "coherent detector", and has been fully discussed in published papers;[32, 9] it has many important properties, those of most practical use being associated with the reception of pulse signals against a high level of noise background. If a comparison is made with the detection of pulse signals by an ordinary so-called linear detector (the usual diode detector with a large signal amplitude approximates to this), then it is found that as the input signal/noise ratio is decreased from about 3 to small fractional values, the improvement in detection given by the coherent detector increases rapidly. If the detection is measured by the ratio

$$\frac{\text{increase in d.c. output when signal is applied}}{\text{r.m.s. noise voltage when signal is absent}}$$

then, calling the input signal/noise ratio R_1, it is found that for the coherent detector the detection is proportional to R_1, but for the linear detector it is proportional to R_1^2, provided R_1 is small compared to unity.

These properties of the coherent detector appear to have been noticed only comparatively recently.

10. Synchrodyne or Homodyne?

In view of the fact that the homodyne developed through the years into a more refined system which was identical with the synchrodyne, there seems to exist a difficulty as to what name the system should be given. It is clearly unsatisfactory that one system should have two different names.

In discussing the coherent detector, Smith[32] appears to identify it with the homodyne. This suggests a very nice way out of the difficulty. Since the prefix "homo" suggests "same", while the prefix "synchro" suggests "synchronized", the logical distinction of meaning is the following: "homodyne" should signify a system where the local oscillation is, in fact, the *same* as the wanted carrier, being obtained by some direct process (e.g., by direct transmission from the transmitter), while "synchrodyne" should signify a system in which the local oscillation is obtained by *synchronizing* an otherwise free oscillator to the incoming signal. This nomenclature has the big advantage of making obsolete the somewhat vague name "coherent detector"; the word "coherent" seems to lack an adequate technical definition. Whether it is now too late to make these changes remains to be seen.

11. Conclusions

It is hoped that this review of the somewhat obscure history of the homodyne, and its subsequent development as the synchrodyne, will have cleared up many previously doubtful and mistaken matters. The subject is one which concerns the author rather personally, but his position in relation to it should now be unambiguous.

Although many interesting and important properties and principles have emerged from the work described here, nevertheless no commercial application of them in the radio field appears to have been made until the advent of colour television.

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MEMBERS OF STANDING COMMITTEES

Bernard Joseph Burridge was born in 1901 in Portsmouth, where he received his general and technical education. In December, 1916, he joined the W.T. Branch of the Royal Navy, with which he served for seven years. In 1921, he was attached to the W.T. Experimental Establishment at Portsmouth.

Mr. Burridge joined the Telephone Manufacturing Company in 1923 as a Telecommunications Engineer and has remained with this company ever since. He went to Australia in 1924 in connection with radio broadcasting and line telephony schemes, and on his return in 1925 was attached to the Development Laboratories.

Whilst with T.M.C., he has been successively Development Engineer, Consultant Engineer to the Works, Engineer-in-Charge of the "Physical" Laboratory and Development Laboratories, and Acting Assistant Chief Engineer, and was appointed to the position of Assistant Chief Engineer in 1939. He is now responsible, under the Chief Engineer, for the general administration of the Engineering Division and engineering activities at the company's St. Mary Cray works.

Elected an Associate Member of the Institution in 1935, Mr. Burridge was transferred to Full Membership in 1950. Since January, 1952, he has served on the Membership Committee.

William McMenemy was born in Newcastle upon Tyne in 1906 and was educated at Newmills, Fife. In 1925 he joined the Royal Air Force as an apprentice and received his technical training at the Electrical and Wireless School, Flowerdown.

After five years service on radio ground and air duties, Mr. McMenemy joined the British Broadcasting Corporation as a Maintenance Engineer, and just prior to the war was appointed Senior Maintenance Engineer of the Westerglen Station.

During the war Mr. McMenemy served in the R.A.F., being mainly engaged on special duties in connection with radio equipment and he was demobilized in 1946 as a Wing Commander. After returning to the B.B.C. for a short time, he obtained his present appointment as Signals Officer with the Ministry of Civil Aviation in 1947.

Elected to Full Membership of the Institution in 1945, Mr. McMenemy has served on the Membership Committee for the past three years.

Professor Emrys Williams was born in Liverpool in 1910, and graduated with the degree of Bachelor of Engineering of Liverpool University in 1932. After obtaining the Ph.D. degree for a thesis on research in electro-acoustics, he was appointed to the scientific staff of the Research Laboratories of the General Electric Co.

In 1937, Dr. Williams became lecturer in Electrical Engineering at King's College, Newcastle upon Tyne, and since 1947 he has occupied the Chair of Electrical Engineering at the University College of North Wales, Bangor. Professor Williams has recently been appointed the first professor of Electrical Engineering of the University College of South Wales and Monmouthshire, Cardiff, and will take up his duties in October, 1954.

Professor Williams was elected a Full Member of the Institution in 1944, and served on the North-Eastern Section Committee while in Newcastle. He has been a member of the Library Committee for several years and was appointed its Chairman three years ago; he is also Chairman of the Ministry of Labour Negotiating Committee. Professor Williams was appointed to the General Council in 1951, and was re-elected for a further period at the last annual general meeting.

In addition to being the author of papers in various technical journals, Professor Williams is the author of the book "Thermionic Valve Circuits," and has given papers to several local sections of the Institution.

NOTICES

Obituaries

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

Group Captain George Norman Hancock, C.B.E., R.A.F., was killed on April 1st when flying a Meteor aircraft on a duty flight at Farnborough. He was aged 40 years and was unmarried.

Group Captain Hancock's tragic death ended a distinguished Air Force career. Educated at Nottingham High School, he entered the R.A.F. College, Cranwell, in 1931 and passed out two years later as "Prize Cadet." After a period on general duties, he was Experimental Signals Officer at the Aeroplane and Armament Experimental Establishment, Boscombe Down, from 1938 until 1942 when he was appointed Staff Officer to the British Air Commission in Washington. Between 1943 and 1946 he held staff appointments with Coastal Command, latterly as Chief Signals Officer. From 1946 to 1951 he was Group Captain (Signals) at Air Traffic Headquarters in the United Kingdom and in Germany and he was subsequently attached to the Ministry of Supply on liaison duties.

In the Birthday Honours List for 1944 Group Captain Hancock was Mentioned in Despatches and in the New Year Honours List for 1946 he was appointed a Commander of the Most Excellent Order of the British Empire.

Elected to Associate Membership of the Institution in 1941, Group Captain Hancock was transferred to Full Membership in 1947. He had served on the Technical Committee since April, 1953.

* * *

Frank Edwin Matthews (Associate), who was born at Belmont, Surrey, in 1907, was during the war a Civilian Instructor in radio subjects with the Royal Air Force. He subsequently received an appointment with the Admiralty Signal and Radar Establishment, Cosham, where he remained until the time of his death on February 1st last. He is survived by his widow.

Mr. Matthews was elected an Associate of the Institution in 1944.

Production Exhibition and Conference

The Institution of Production Engineers is sponsoring an Exhibition and Conference in London in July, 1954.

Concurrently with the Exhibition, which is being held at Olympia, London, from July 7th to the 14th, a Conference will be held at which papers relating to the improvement of productivity will be presented.

This new venture by the Institution of Production Engineers will be of interest to members attending the Brit.I.R.E. Convention on Industrial Electronics since the President of the I.Prod.E., Sir Walter Puckey, will be taking part in the discussion on "How electronics can increase production" at Oxford on July 11th.

Details of the Exhibition and Conference may be obtained from the Secretary of the I. Prod. E., 36 Portman Square, London, W.1.

European Exhibition on Productivity

The problems of increasing productivity are receiving wide attention not only in the British Isles but also in Europe. An Exhibition is to be held in Strasbourg, from May 14th to June 7th, under the aegis of the Organization for European Economic Co-operation.

Almost every aspect of industrial activity will be dealt with in the Exhibition, which is being organized by an International Committee.

Information may be obtained from the Secretary-General, L'Exposition Européenne de la Productivité, 101 rue de Prony, Paris 17e.

South Devon Television Station

Reference was made in the December 1953 issue of the *Journal* (page 589) to the Public Enquiry into the application made by the B.B.C. for planning permission to establish a television station on North Hessary Tor in the Dartmoor National Park area.

The Minister of Housing and Local Government, Mr. Harold Macmillan, has now announced in the House of Commons that the B.B.C. will be permitted to build the station subject to certain conditions, among which are the requirements that materials used in the buildings and approach road should be satisfactory to the Local Planning Authority and that all cables serving the station should be placed underground.

The B.B.C. has stated that the transmitter will operate on the same channel in Band I as Holme Moss (Channel 2) with vertical polarization, and the vision transmitter will have an output of 5 kW.

DECAY OF EMISSION FROM AN OXIDE-COATED CATHODE DUE TO ADSORPTION OF MATTER LIBERATED FROM THE ANODE*

by

S. Deb, M.Sc.†

SUMMARY

The paper examines critically the process of decay of emission current of an oxide-coated cathode when poisoning agents liberated from the anode by electron bombardment are adsorbed on the emitting surface. Two types of adsorption process, mobile and immobile, are considered. The final equations for the decay process which appear in integral form in the two types of adsorption are solved numerically by estimating the probable values of the constants involved. An approximate analytical solution of the integral is also given for the mobile type of adsorption. It is found that the decay resulting from mobile adsorption is generally of short duration and is likely to be spontaneously recoverable, but that resulting from immobile adsorption is of long duration and is probably permanent. For the former case the time constant of decay is about 10^{-3} sec, provided that the poisoning agent is oxygen at a pressure of the order of 10^{-7} mm Hg and temperature of the order of 1,000°K, and that the adsorption process is confined to a fraction (about 0·1) of the actual cathode area. The hypothesis that a decay process resulting from mobile adsorption is responsible for the differences between the pulsed and the d.c. behaviour of an oxide-coated cathode is examined in general terms and is found to deserve consideration.

1. Introduction

It has long been known that gases like oxygen, carbon monoxide, chlorine and carbon dioxide have very adverse influence on electron emission from an oxide-coated cathode.[1] Although powerful vacuum and gettering techniques have been developed in recent years to eliminate such "poisoning" effects, valves using oxide-coated cathode are still known to exhibit peculiar characteristics which point to the presence of a residual effect of this nature. In particular, the following peculiar behaviour of valves with oxide-coated cathode may be noted:

(i) Very often immediately on application of the anode voltage the current begins to decay. The decay continues ordinarily for a minute or so.[2] On switching off the anode voltage, a time of the same order is required to regain the initial electron emission from the cathode. The phenomenon has not yet been satisfactorily explained. One of the possibilities suggested is that the effect might be due to liberation of gases from the anode.

(ii) It has been observed that with pulsed voltages of the order of a micro-second duration it is possible to obtain current values which are 10–100 times larger than that obtainable with steady d.c. voltage. This difference in current yield is ascribed to a very sharp and recoverable decay of current which follows the application of anode voltage. The time constant of this decay is much smaller than of that described in (i) above, being of the order 10^{-4}–10^{-2} sec. A significant observation was made in this connection by Wright[3] and by Feaster.[4] Both of these authors found that the decay of current occurs only when the anode surface is contaminated with a poisoning agent. Wright further observed that the decay process may be prevented by replacing the anode used during the "activation process" of the cathode by a fresh one after the completion of the same. This suggests that during the activation process the anode is probably contaminated with some kind of poisonous matter and, when a current is drawn, the contaminating material is ejected from the anode and poisons the cathode surface temporarily.

(iii) Permanent decay of emission from an oxide-coated cathode is also often found to occur. There are evidences that this arises from the effect of gases released from the anode under electron bombardment.[2,5]

* Manuscript received January 19th, 1954. Based on a paper published in the Indian Journal of Physics, Vol. 25, 1951, p. 197. (Paper No. 260.)

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U.D.C. No. 621.385.

Some attempts have been made to interpret theoretically the observed decay of current.[6-9] However, all such attempts are based on results of earlier experiments and are concerned solely with the possible processes that may occur only in the cathode. It was, therefore, thought desirable to make a fresh study of the problem taking into account the recent evidences that condition of anode surface might also play a decisive role in the whole process. This has been done in the present paper by examining closely the hypothesis that adsorption—both mobile and immobile—by the cathode surface, of matter released from the anode by the impact of the oncoming electronic stream is ultimately responsible for the observed current decay. The short-period decay, as described under (ii), and the main differences between the properties of the cathode under pulsed and steady state conditions of operations have also been discussed in general terms, keeping in view the possibility that these might originate from the mobile adsorption process referred to above.

2. Analysis of the Basic Process of the Current Decay

2.1. *Setting up of the mathematical equation*

The basic process which is sought to be analysed is as follows: it is assumed that electrons emitted from the cathode and impinging on the anode surface by being driven by the applied field, eject matter particles from the same. These travel towards the cathode surface and are adsorbed. As this adsorption process progresses, an opposite process, namely evaporation (or better, desorption), begins to develop. If the instantaneous number of adsorbed particles be denoted by N then,

$$\frac{dN}{dt} = \text{Time rate of adsorption—Time rate of desorption}$$

In order to estimate the two rate processes it is convenient to consider separately two distinct cases:

(*a*) When the adsorbed particles remain mobile on the surface.

(*b*) When these are immobile.

2.1.1. Case (*a*). Mobile adsorption

The reaction rate for such adsorption as given by the theory of absolute reaction rates[10] is given by

$$N_g \frac{kT}{h} \frac{f_0}{f_g} \exp\left[-\frac{E}{kT}\right],$$

where N_g = Concentration of gaseous particles in front of the surface

k = Boltzmann's constant

h = Planck's constant

T = Absolute temperature

f_0 = Complete partition function of particles in the activated state

f_g = Complete partition function in the gaseous state

E = Energy of activation for adsorption.

The rate of desorption as given by the theory of absolute reaction rates may be written as

$$kT\beta N \exp\left[-\frac{E'}{kT}\right],$$

where β is a constant involving partition functions and E' is the energy of activation for desorption.

Now N_g is related to the anode current density i. If the reasonable assumption is made that Ng is directly proportional to i, then

$$N_g = \alpha' i,$$

where α' is a constant dependent on anode voltage, material of the anode, nature of particles contaminating the anode and the electrode geometry.

From the foregoing one readily obtains

$$\frac{dN}{dt} = \alpha' i \frac{kT}{h} \frac{f_0}{f_g} \exp\left[-\frac{E}{kT}\right] - kT\beta N \exp\left[-\frac{E'}{kT}\right] \quad \ldots\ldots\ldots(1)$$

It is to be noted that in deriving eq. (1) it has been tacitly assumed that the inter-electrode transit times for the electrons and for the poisoning particles are negligible in comparison with the duration of the whole decay process. Where this assumption is not valid the time constant of decay as derived from eq. (1) will be a little higher than the actual value.

We now introduce into eq. (1) the *fractional coverage* (θ) of the cathode surface. We put

$$N = N_0 \theta \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(2)$$

where N_0 is the maximum number of mobile particles which can settle down per unit area of the cathode surface, without mutual overlapping.

Substituting (2) in (1)

$$N_0 \frac{d\theta}{dt} = \alpha' i \frac{kT}{h} \frac{f_0}{f_g} \exp\left[-\frac{E}{kT}\right] - N_0 kT\beta\theta \exp\left[-\frac{E'}{kT}\right] \quad \ldots\ldots(3)$$

$$= \alpha i - A N_0 \theta \quad \ldots\ldots(4)$$

where $\alpha = \alpha' \frac{kT}{h} \frac{f_0}{f_g} \exp\left[-\frac{E}{kT}\right]$ ......(5)

$$A = kT\beta \exp\left[-\frac{E'}{kT}\right] \quad \ldots\ldots(6)$$

It remains to express θ in terms of the current i in (4). In order to do this one might follow the procedure adopted by Wagener[9] and assume the following relation based on the experimental results of Becker,[1]

$$\ln \frac{i}{i_0} = b(\theta_0 - \theta) \quad \ldots\ldots(7)$$

where i_0 = the limiting value of the current for large values of t

θ_0 = values of θ corresponding to i_0

b = a slowly varying function of θ, which for changes of current by one order may be taken to be a constant.

We thus have from (7) and (4)

$$\frac{1}{i}\frac{di}{dt} = -A \ln \frac{i}{i_0} - \frac{b\alpha i}{N_0} + Ab\theta_0 \quad \ldots(8)$$

$$= -A \ln \frac{i}{i_0} - Bi + C \quad \ldots\ldots(9)$$

where $\left.\begin{aligned} B &= \frac{b\alpha}{N_0} \\ C &= Ab\theta_0 \end{aligned}\right\}$ ......(10)

whence

$$t = \int \frac{di}{i\left[-A \ln\left(\frac{i}{i_0}\right) - Bi + C\right]} \quad \ldots\ldots(11)$$

It is difficult to solve the integral equation (11) analytically. A solution may, however, be obtained by graphical method, for which the values of the constants A, B and C have to be estimated.

2.1.2. Case (*b*). Immobile adsorption

When the adsorbed particles are immobile the following time rates are obtained from the theory of absolute reaction rates:

Rate of adsorption

$$= N_g N_s \frac{kT}{h} \frac{f_0}{f_g f_s} \exp\left[-\frac{E}{kT}\right]$$

$$= \alpha' i N_s \frac{kT}{h} \frac{f_0}{f_g f_s} \exp\left[-\frac{E}{kT}\right],$$

where

N_s = Number of vacant places available for adsorption/cm^2

f_s = Complete partition function of particles in the adsorbed state.

$f_0, f_g, h, k, T, E, \alpha'$ and i have the same meanings as in mobile adsorption.

Rate of desorption

$$= kT\beta N \exp\left[-\frac{E'}{kT}\right]$$

β and N having the same meanings as in the case of mobile adsorption. Hence

$$\frac{dN}{dt} = \alpha' i N_s \frac{kT}{h} \frac{f_0}{f_g f_s} \exp\left[-\frac{E}{kT}\right] - kT\beta N \exp\left[-\frac{E'}{kT}\right]$$

$$= \alpha i N_s - AN \quad \ldots\ldots(12)$$

where

$$\left.\begin{aligned} \alpha &= \alpha' \frac{kT}{h} \frac{f_0}{f_g f_s} \exp\left[-\frac{E}{kT}\right] \\ A &= kT\beta \exp\left[-\frac{E'}{kT}\right] \end{aligned}\right\} \quad \ldots\ldots(13)$$

Assuming

$$N = N_0'\theta \text{ and } N_s = N_0'(1-\theta),$$

and expressing (12) in terms of θ

$$\frac{d\theta}{dt} = \alpha i(1-\theta) - A\theta = \alpha i - \theta(A + \alpha i) \quad (14)$$

Eliminating θ in terms of i with the help of (7)

$$-\frac{1}{i}\frac{di}{dt} = \alpha i + \frac{\ln \frac{i}{i_0} - b\theta_0}{b}(\alpha i + A) \quad \ldots\ldots(15)$$

This can be rearranged and rewritten as

$$\frac{1}{i}\frac{di}{dt} = \alpha(b\theta - b + \ln i_0)\, i - \alpha i \ln i - A \ln i + A(b\theta_0 + \ln i_0)$$

$$= Bi - (\alpha i + A)\ln i + C \quad \ldots\ldots(16)$$

where

$$\left.\begin{aligned} B &= \alpha(b\theta_0 - b + \ln i_0) \\ C &= A(b\theta_0 + \ln i_0) \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots(17)$$

Integrating (16)

$$t = \int \frac{\mathrm{d}i}{i[Bi - (\alpha i + A)\ln i + C]} \quad \ldots(18)$$

To evaluate the integral numerically, we must estimate the values of A, B and C.

2.2. *Determination of Constants: Evaluation of the Integrals* (11) *and* (18)

2.2.1. Case (*a*). Mobile adsorption

In order to evaluate the constants for the case of mobile adsorption, use may first be made of the boundary condition

$$\frac{\mathrm{d}i}{\mathrm{d}t} = 0, \text{ for } i = i_0.$$

From this and eq. (9) one obtains

$$C = Bi_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(19)$$

Integral (11) then takes the form

$$t = \int \frac{\mathrm{d}i}{i(-A\ln i - Bi + C')} \quad \ldots\ldots\ldots(20)$$

where

$$C' = Bi_0 + A\ln i_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots(21)$$

Eqs. (10), (19) and (21) show that A and C' may be calculated if B is known. The problem is, therefore, reduced to the evaluation of the constant B. This involves a knowledge of the values of f_0/f_g, kT/h, b, α', N_0 and E.

Value of f_0/f_g.—Value of f_0/f_g may be inferred if the various degrees of freedom associated with the particles in the free and in the activated state are known. For mobile adsorption particles in both of these states will have sensibly identical degrees of freedom for rotation and vibration. Coming to the consideration of degrees of translational freedom it may be noted that there are three degrees for the gaseous and two for the activated state. Thus,

$$\frac{f_0}{f_g} \simeq \frac{h}{\sqrt{2\pi mkT}}$$

m being the mass of the particles. With $T \simeq$ 1,000°K, which is the usual operating temperature of the oxide cathode, and m = 10–100 atomic mass

$$\frac{f_0}{f_g} \simeq 10^{-9}$$

Value of kT/h.—For $T \simeq$ 1,000°K, this has a value 2×10^{13}.

Value of b.—For obtaining an idea of the value of b use may first be made of the boundary condition

$$i = i_m \text{ at } t = 0,$$

which, together with (7), gives

$$\frac{i_m}{i_0} = \exp(b\theta_0).$$

The usually observed values of i_m/i_0 lie between 10 and 100. Taking the two boundary values of this range

$$b\theta_0 = 4{\cdot}6 \text{ for } \frac{i_m}{i_0} = 100,$$

$$b\theta_0 = 2{\cdot}3 \text{ for } \frac{i_m}{i_0} = 10.$$

No information is available about θ_0. However, considering the pronounced drop in current that is usually observed, θ_0 may be expected to lie in the range 0·1–1·0. Uncertainty in the value of B that may result from any arbitrary choice of θ_0 within this range cannot be more than an order. This is trivial compared to the uncertainties resulting from our choice of the other comparatively more obscure factors like α' and N_0. Putting $\theta_0 = 0{\cdot}5$, one obtains

$$b = 9{\cdot}2 \text{ for } \frac{i_m}{i_0} = 100$$

$$b = 4{\cdot}6 \text{ for } \frac{i_m}{i_0} = 10.$$

Value of α'.—In order to estimate a probable value of α' one must have some knowledge of the partial pressure of the poisonous gas within the valve and of the emission current. Under favourable conditions the d.c. emission is of the order of 1 A and, for realizing this, the total pressure inside the valve envelope must be reduced at least to 10^{-6} mm of mercury. Partial pressure of the poisoning agent for a newly made valve of this type would be considerably lower—probably 10^{-8} mm Hg or less. When fresh matter is released from the anode this latter pressure may attain the highest limit of 10^{-6} mm Hg. Taking the intermediate value 10^{-7} mm Hg one obtains $\alpha' \simeq 2 \times 10^9$/A.

Value of N_0.—For a rigidly bound complete mono-molecular layer of gas to be formed on any homogeneous surface N_0 will have a value

10^{14}–10^{15}/cm^2. For the case of a mobile layer to be formed on an inhomogeneous surface like that of an oxide-coated cathode, the situation is a little different and the value of N_0 is likely to be less. The reasons are twofold. Firstly, mobility of the particles deposited on the surface requires the existence of appreciable intermolecular distances. This necessitates a less dense packing of the particles deposited. Secondly, the deposited particles may cluster round preferred regions of the inhomogeneous surface; in particular, in the case of an oxide cathode the clustering may be round those regions of the surface where thermionic activity varies sharply.

In view of the above, N_0 for mobile adsorption is likely to be 10^{14} or less and this may be expressed by saying that the effective area for such adsorption is a fraction of the geometrical area of the cathode surface. Computations will, therefore, be carried out with three selected values of N_0, namely, 10^{14}, 10^{13} and 10^{12}, i.e., for cases when the effective area for mobile adsorption is 1, ·1 and ·01 times the geometrical cathode area.

Value of E.—The energy of activation for adsorption (E) is known to be very small for a rapid process and also for the case when the adsorbed particles are atomic. The short-period decay that is observed may be classed as a rapid process. Further, such decay always shows a slowing down with lowering of temperature. These indicate that the energy of activation must be small but finite. Assuming that the decay process is connected with diffusion of matter inside the body of the oxide layer, Blewett[11] had found the energy of activation for diffusion to be 0·7 V. If, however, the decay process is ascribed to adsorption rather than diffusion, then remembering that both the diffusion and the adsorption rates are expressible in the same general form, namely, $R \exp[-E/kT]$, E for the former should be of the order of 0·7 V. It is to be noted in this connection that Blewett had confined his studies to long period decay processes lasting for minutes and as such his value may be taken as an upper limit. It is also to be noted that the poisoned cathode regains its activity quickly when the applied anode voltage is withdrawn, indicating that the height of the surface potential barrier, which determines E, cannot be much larger than the thermal energy of the particles at the cathode temperature, which is of the order of 0·1 V. In view of the above E may be estimated to lie in the range 0–0·2 V. Computation will, therefore, be made for two values of E, 0 and 0·2 V.

From the values of the various factors deduced above those of the constants A, B and C may be readily obtained. These are given in Table 1.

Graphical solution.—Graphical solution of eq. (11) obtained by utilizing the values of the constants recorded in Table 1 are shown in Figs. 1–3.

TABLE 1

Estimated values of constants for the case of mobile adsorption

	i_m (amps)	i_0 (amps)	N_0	E (volts)	A	B	C	C'
1	10	·1	10^{12}	0	8	$3·68 \times 10^{2}$	$3·68 \times 10^{1}$	$1·84 \times 10^{1}$
2	10	·1	10^{12}	·2	8×10^{-1}	$3·68 \times 10^{1}$	3·68	1·84
3	10	·1	10^{13}	0	8×10^{-1}	$3·68 \times 10^{1}$	3·68	1·84
4	10	·1	10^{13}	·2	8×10^{-2}	3·68	$3·68 \times 10^{-1}$	$1·84 \times 10^{-1}$
5	10	·1	10^{14}	0	8×10^{-2}	3·68	$3·68 \times 10^{-1}$	$1·84 \times 10^{-1}$
6	10	·1	10^{14}	·2	8×10^{-3}	$3·68 \times 10^{-1}$	$3·68 \times 10^{-2}$	$1·84 \times 10^{-2}$
7	10	1	10^{12}	0	8×10^{-1}	$1·84 \times 10^{2}$	$1·84 \times 10^{2}$	$1·84 \times 10^{2}$
8	10	1	10^{12}	·2	8	$1·84 \times 10^{1}$	$1·84 \times 10^{1}$	$1·84 \times 10^{1}$
9	10	1	10^{13}	0	8	$1·84 \times 10^{1}$	$1·84 \times 10^{1}$	$1·84 \times 10^{1}$
10	10	1	10^{13}	·2	8×10^{-1}	1·84	1·84	1·84
11	10	1	10^{14}	0	8×10^{-1}	1·84	1·84	1·84
12	10	1	10^{14}	·2	8×10^{-2}	$1·84 \times 10^{-1}$	$1·84 \times 10^{-1}$	$1·84 \times 10^{-1}$
13	1	·1	10^{12}	0	8	$1·84 \times 10^{2}$	$1·84 \times 10^{1}$	0
14	1	·1	10^{12}	·2	8×10^{-1}	$1·84 \times 10^{1}$	1·84	0
15	1	·1	10^{13}	0	8×10^{-1}	$1·84 \times 10^{1}$	1·84	0
16	1	·1	10^{13}	·2	8×10^{-2}	1·84	$1·84 \times 10^{-1}$	0
17	1	·1	10^{14}	0	8×10^{-2}	1·84	$1·84 \times 10^{-1}$	0
18	1	·1	10^{14}	·2	8×10^{-3}	1·84	$1·84 \times 10^{-2}$	0

We can now evaluate approximately the integral in eq. (20). We note from the data given in Table 1 that A is always less than B by more than an order. The term involving A may, therefore, be neglected in the denominator of eq. (20). The integral equation then becomes readily soluble, the final form of the solution being

$$i = \frac{i_0}{1 - D\exp[-Bi_0 t]} \quad \ldots\ldots\ldots(22)$$

where $D = \dfrac{i_m - i_0}{i_m}$ ....(23)

The time required for the current to fall to $1/e$th part of the initial value i_m is given by

$$\tau = -\frac{1}{Bi_0}\ln\frac{i_m - ei_0}{i_m - i_0} \quad \ldots\ldots\ldots\ldots(24)$$

and this may be defined as the "time constant" of the decay process. For the set of data recorded in column 3 of Table 1, i.e. for the case when energy of activation is negligible and the effective area for mobile adsorption is 0·1 times the geometrical area $\tau \simeq 10^{-3}$ sec, if the current intensity decreases from a value 1·0 to 0·1. This agrees with the value observed for the short-period decay.

2.2.2. Case (*b*). Immobile adsorption

In order to determine the values of the constants for the case of immobile adsorption, use may first be made of the condition

$$\frac{di}{dt} = 0, \text{ for } i = i_0.$$

This, together with eq. (4), gives

$$\left.\begin{aligned} A &= \frac{\alpha}{\theta_0}(1-\theta_0)i_0 \\ C &= \frac{\alpha i_0}{\theta_0}(1-\theta_0)(b\theta_0 + \ln i_0) \end{aligned}\right\} \quad \ldots.(25)$$

Values of b, θ_0, i_0 and α' would be of the same order of magnitude as in Case (*a*) of mobile adsorption. Value of E, the energy of activation, may be higher in this case. Nevertheless, computation may be carried out with the values assumed for mobile adsorption, it being remembered that the rate of decay thus obtained would represent rather a lower limit of the same. It remains only to determine the value of $f_0/f_g f_s$ which may be inferred as follows.

When the adsorbed particles are immobile—as presumably are also those in the activated state—f_0 and f_s will not have any contribution due to translational degrees of freedom. Contributions due to rotation and vibration would be sensibly the same for f_0 and f_s. f_g will have contributions from three degrees of translational freedom and a smaller one from rotation and vibration, the latter being altogether absent if the particles are atomic. Thus, one obtains

$$\frac{f_0}{f_g f_s} \simeq \frac{h^3}{(2\pi mkT)^{3/2}} \simeq 10^{-27} \text{ at } T = 1{,}000°\text{K}.$$

Graphical solutions.—The values of the constants A, B and C as obtained from the above considerations are given in Table 2 and a typical graphical solution of eq. (18) is given in Fig. 4.

3. Results and Discussion

The values of the different parameters for which graphical solutions of eqs. (11) and (18) are given in Figs. 1–4, are listed below. (Figs. 1–3 are for mobile and Fig. 4 for immobile adsorption).

For Fig. 1, Curve I, $i_m = 10$, $i_0 = 0·1$ and $E = 0$.

For Fig. 1, Curve II, $i_m = 10$, $i_0 = 1·0$ and $E = 0$.

For Fig. 2 (Curves I and II) the parameters are the same as for Fig. 1, excepting for $E = 0·2$ V.

For Fig. 3 $i_m = 1·0$, $i_0 = 0·1$ and $E = 0$.

TABLE 2

Estimated values of constants for the case of immobile adsorption

	i_m (amps)	i_0 (amps)	E (volts)	α	A	B	C
1	10	·1	0	4×10^{-5}	4×10^{-6}	$-2·76 \times 10^{-4}$	$9·2 \times 10^{-6}$
2	10	·1	·2	4×10^{-6}	4×10^{-7}	$-2·76 \times 10^{-5}$	$9·2 \times 10^{-7}$
3	10	1	0	4×10^{-5}	4×10^{-5}	$-9·2 \times 10^{-5}$	$9·2 \times 10^{-5}$
4	10	1	·2	4×10^{-6}	4×10^{-6}	$-9·2 \times 10^{-6}$	$9·2 \times 10^{-6}$
5	1	·1	0	4×10^{-5}	4×10^{-6}	$-1·84 \times 10^{-4}$	0
6	1	·1	·2	4×10^{-6}	4×10^{-7}	$-1·84 \times 10^{-5}$	0

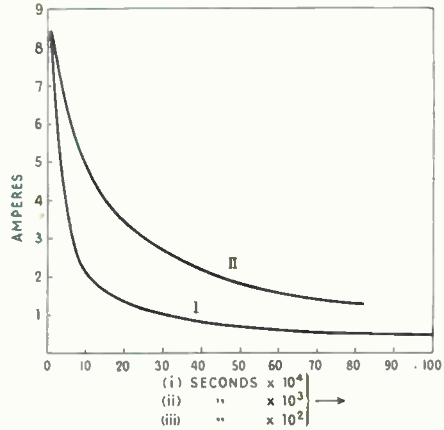

Fig. 1

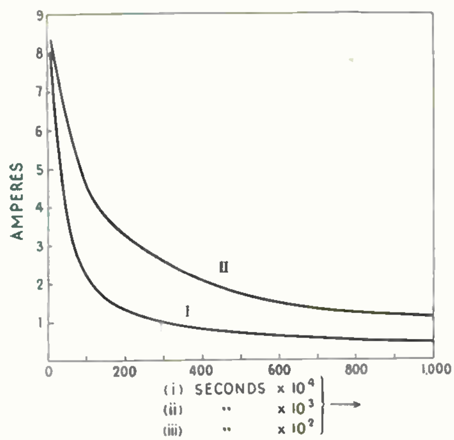

Fig. 2

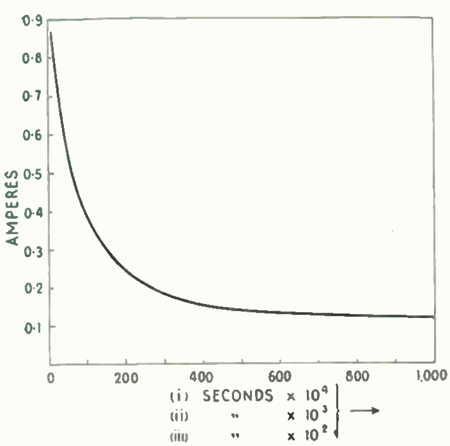

Fig. 3

Fig. 1 (top).—Typical curves showing the decay of current for mobile adsorption when the energy of activation, E, is zero. Curve I is for $i_m = 10$ A, $i_0 = 1$ A and Curve II for $i_m = 10$ A, $i_0 = 0{\cdot}1$ A. Time scales (i), (ii) and (iii) are for $N_0 = 10^{12}$, 10^{13} and 10^{14} respectively.

Fig. 2 (centre).—Curves showing decay of current for the same case as in Fig. 1 excepting that $E = 0{\cdot}2$ V.

Fig. 3 (bottom).—Curve showing decay of current for the same case as in Fig. 1 excepting that $i_m = 1$ A and $i_0 = 0{\cdot}1$ A.

For all the above figures the time scale marked (i) is for $N_0 = 10^{12}$, (ii) for $N_0 = 10^{13}$, and (iii) for $N_0 = 10^{14}$.

Fig. 4 is for the case of immobile adsorption with $i_m = 10$, $i_0 = 1{\cdot}0$ and $E = 0$.

Considering the curves in Figs. 1-3, for mobile adsorption we note that these represent a quick process—the major part of the decay taking place within 10^{-4} to 10^{-2} sec and the whole process is complete within 0·1 to a few seconds. This period is found to be almost of the same order as that obtained from the approximate relation (22). The relation may, therefore, be taken to yield sufficiently accurate results for all practical purposes and, in all subsequent discussions, interest will be confined to this formula only.

The result stated above clearly indicates that for a time constant of decay within the range 10^{-2}—10^{-4} sec., N_0 must be within the range 10^{12}—10^{13} and E within 0—0·2 volt. In other words, the effective area for mobile adsorption must be sensibly less than the geometrical area. This is not surprising because studies have shown[9, 12] that the surface of an oxide cathode is usually rich in defects arising out of conspicuous loss of oxygen atoms. The adsorption process may, therefore, have a tendency to be confined to the defective regions resulting in a reduction of the effective area for the adsorption process.

The curve in Fig. 4 for immobile adsorption requires little consideration. It is characterized by a large time constant of decay—of the order of minutes—although it is for the case of zero energy of activation. If the energy of activation is 0·2 volt—which is quite a modest estimate—the decay process would last for about an hour. Such long time for decay is usually observed when the cathode is de-activated permanently.

It has not yet been observed for the spontaneously recoverable type of de-activation. The result obtained, therefore appears to suggest that the immobile type of adsorption is probably connected with permanent decay of activity of the cathode and that during a time interval of the order of millisecond such a process cannot bring about any significant change in the emission current.

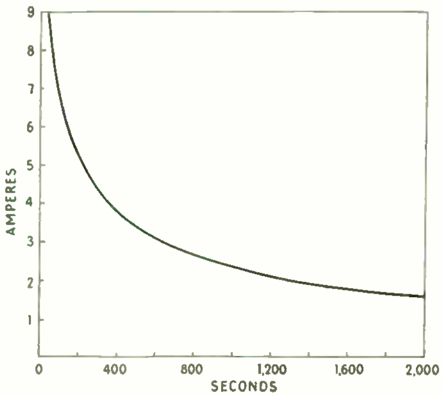

Fig. 4.—A typical curve showing the decay of current due to immobile adsorption when E the energy of activation is zero. Here $i_m = 10$ A and $i_0 = 1$ A.

We will conclude this section with a discussion regarding the possible nature of the adsorbed particles. We recall, in this connection, that the partial pressure of the adsorbed gas should vary within the range $10^{-6}-10^{-8}$ mm Hg during the whole process of decay and that, at the same time, the nature of the gas should be such as to reduce the emission current to a marked extent. We also accept the experimental evidence that the particles causing the decay are deposited on the anode during the process of activation.[3] And, since the substances that are likely to be liberated from an oxide cathode during activation are CO, Ba, BaO and O_2, we discuss the possible roles of these in the observed phenomenon.

CO is produced by the reducing action of carbon which often exists in the core of the cathode as an impurity. CO, however, readily combines with oxides of Ba and Sr producing carbonate and is not likely to give rise to any adsorption phenomenon.

Ba is set free by the action of reducing agents which reside in the core. One very efficient reducing agent is silicon which reacts as follows:—

$$2\,BaO + Si \rightleftharpoons SiO_2 + 2\,Ba.$$

It is also interesting to note that Sproull has observed decay with a cathode whose core contained silicon as impurity. Equilibrium pressure of Ba for this reaction at 1,000°K is about 10^{-5} mm[13] which is above the required range of pressure ($10^{-6}-10^{-8}$ mm Hg). For pressure within this latter range Ba will have a tendency to escape through evaporation rather than to settle on the cathode by adsorption. Further, the required pressure range being too near the equilibrium pressure value, any change brought about by a change in pressure of an order one or two cannot be a very rapid process. Above all, Ba is not known to have any adverse effect on emission—rather it is believed to be the active agent responsible for emission.[14, 15] As such, the possibility of Ba being connected with the observed decay process may be ruled out.

BaO is liberated from the surface by evaporation. Its vapour pressure is 10^{-7} mm Hg at 1,275°K and 10^{-12} mm Hg at 1,000°K. This is quite favourable for a rapid process taking place through the deposition of BaO on the cathode surface at pressures $10^{-6}-10^{-8}$ mm Hg. BaO is however, not known to have a detrimental effect on emission. Again, evaporation of BaO, although a little more pronounced during the activation processes, goes on all the while. Further, there is the possibility that any BaO contaminating the anode would breakdown into Ba and O under the impact of the oncoming electron beam.

Even when no reducing agent is present oxygen may be liberated by dissociation of BaO thus,

$$2\,BaO \rightleftharpoons 2\,Ba + O_2.$$

Equilibrium pressure of oxygen for this reaction is 10^{-15} mm Hg at 1,000°K and 10^{-11} mm at 1,275°K.[13] During the activation process the cathode temperature is often raised above 1,275°K—perhaps to 1,500°K. This increases somewhat the liberation of oxygen throughout the body of the oxide layer. Again, free oxygen liberated through the electrolytic dissociation process, which comes into play when the activating potential is applied, also travels on to the surface of the cathode. This outward rush will give rise to an enormous increase in the rate of loss of oxygen from the cathode to the vacuum at the oxide/vacuum interface. Contamination of the anode by oxygen during activation is, therefore, likely to be very heavy and the subsequent release of the same from the anode under electron bombardment may lead to a sudden increase in its partial pressure to a value 10^{-6}—10^{-7} mm Hg. This being far above the equilibrium value for oxygen in the cathode, rapid deposition on the cathode becomes possible. Again, since oxygen is known to have a pronounced poisoning effect on the cathode, the process of deposition would easily bring about a decrease in current by one order or more.

To make this discussion complete mention may be made of two other possible poisoning agents, CO_2 and Cl_2. CO_2 is liberated when carbonate coating is broken down for obtaining the oxide. During this process the pressure inside the valve may increase to a value as high as 10^{-4} mm Hg. If reducing agents are absent this rise would be almost entirely due to CO_2. It is however, difficult to associate CO_2 with any adsorption process, because, it would combine readily with excess Ba to form carbonate rather than remain adsorbed.

Regarding Cl_2 it may be liberated by oxygen if the glass that is used in the valve is contaminated with HCl.

$$4\,HCl + O_2 \rightleftharpoons 2\,H_2O + 2\,Cl_2.$$

The water formed may be carried away by pump and the chlorine deposited on anode either as such or as chloride. As is, however, evident, liberation of chlorine is incidental and not an inevitable accompaniment of the activation, process.

The above considerations show that the substance ejected from the anode due to electron bombardment, the adsorption of which on the cathode surface causes current decay in a valve, is most likely oxygen. Occasionally, and to a lesser extent, it may also be chlorine or BaO.

4. Decay due to Adsorption and Difference between Pulsed and D.C. Behaviour of an Oxide Cathode

The discussions and the results obtained in Secs. 2 and 3 show that a decay of current lasting over a millisecond or so may arise out of mobile adsorption on the surface of an oxide cathode of oxygen which has been ejected from the anode by electron bombardment. As pointed out in the introduction a decay of this type is known to be responsible for the large difference between the currents obtainable from such a cathode with pulsed anode voltage of micro-second duration and with d.c. voltage. The hypothesis may, therefore, be advanced that the difference between the pulsed and the d.c. behaviour of an oxide cathode is due to temporary poisoning of the same by adsorption of oxygen brought to its surface from the anode by electron bombardment. The hypothesis is necessarily tentative as the possible mechanism of emission from an oxide cathode is still uncertain. Nevertheless, it may be of interest to examine whether the main characteristics of pulsed emission and the associated decay phenomenon may be explained —at least qualitatively—from the standpoint of such an hypothesis. These characteristics are as follows:—

(i) The rate of decay is increased with increasing anode current and at times anode voltage.
(ii) The rate of decay is lowered with lowering of temperature.
(iii) The short period decay is most pronounced in aged cathode.
(iv) The rate of decay is independent of thickness of the coating.
(v) The decay of current is accompanied by a somewhat similar decay of coating conductivity.
(vi) Good pulsed emitters are also good d.c. emitters but not always the reverse.
(vii) Deviation from the law governing the so-called Schottky effect is larger for pulsed emission.
(viii) Work function under the pulsed condition of operation is equal to and sometimes a little higher than that under d.c. condition.
(ix) A square-topped pulse of current is obtained with voltage of micro-second duration.

(i) The picture of the decay process as developed in Secs. 2-3 also leads to an increased decay rate with higher voltage and current. If the current remains constant then an increase in anode voltage would increase the energy of the impinging electrons. This may often favour a larger value of α' and hence a larger value of B in eq. (10). di/dt would thereby increase which would quicken the adsorption process and hence also the decay. (*Note.* In the calculations carried out in Sec. 3, α' was obtained from the value of the partial pressure which is the quantity directly observable. Actually, of course, α' determines the pressure.) It is also evident from eqs. (9) and (24) that di/dt would increase with increasing value of i (or i_0) indicating a quicker rate for larger current. A comparison of curves II in Fig. 1 and Fig. 3 would also lead to the same conclusion.

(ii) According to the decay process suggested a slower decay rate at lower temperature would be observed if a small but finite energy of activation is assumed. A value of about 0·1 volt is sufficient at the usual temperature of operation of the cathode, viz., 1,000°K. The effect of this would be to lower the value of α' and hence of B leading to a higher time constant of the decay represented by (24). It may be pointed out that the non-exponential factor of eq. (5) also decreases with decrease in temperature thus lowering the value of B. But the lowering effect on B of the exponential factor will always play the dominant role.

(iii) The phenomenon of short lived decay being best observed in aged cathodes, cannot be easily associated with the adsorption of matter coming from the anode. It may be thought that N_0 will decrease and hence B will increase with ageing.[12] This, however, is also accompanied by a lowering of i_0 in (24) and thus opposes somewhat the effect due to B. There is, however, a likelihood of the effect arising out of the lowering of efficiency of the getter with ageing. Initially, the getter is quite active and may remove a substantial number of particles released from the anode thereby reducing the number of particles which are available for adsorption. This would effectively reduce the value of α' and hence of B leading to a slowing down of the decay process. As the getter ages, it fails increasingly to remove any significant amount of matter appearing from the anode and hence permits the adsorption and the associated decay process to proceed with full speed. The quicker decay rate with ageing may, therefore, be associated more with the ageing of the getter than with that of the cathode.

(iv) Non-dependence of decay on the thickness of coating is readily associated with adsorption, as this latter process depends on the state of the surface and not on the conditions in the interior.

(v) It is not possible to associate the phenomenon of close correlation between decay of emission current on the one hand and of coating conductivity on the other, with any simple type of adsorption phenomenon without a close examination of the mechanism of emission from an oxide cathode—an involved subject still in a somewhat fluid state. It is, however, known from experimental observations that an oxide cathode gains conductivity when oxygen is driven off and an excess of the metal content is left behind. The process of deposition of oxygen on the surface (even though in a mobile state) may, therefore, bring about a reduction in conductivity by disturbing the proportion of excess metal to free oxygen which is necessary for maintaining the so-called stoichiometric equilibrium.

(vi) That good pulsed emission ensures good d.c. emission but not the reverse is easily shown to be a result of the decay phenomenon due to adsorption. Eq. (22) suggests that the final current i_0 should be proportional to the initial current i_m where adsorption process is operative. This ensures good d.c. emission from a cathode for which the pulsed emission is high. If, however, the electrode system is clean enough, decay resulting from adsorption would be insignificant. For such a valve the cathode would not yield greatly different currents under d.c. and pulsed condition of operation. This would be the case of a good d.c. emitter which does not give a proportionately large pulsed emission.

(vii) Greater deviation from the law governing the so-called Schottky effect under pulsed condition is also to be anticipated from the effect of poisoning arising out of adsorption. As already pointed out the oxide surface is initially highly inhomogeneous due to oxygen deficiency, and the adsorbed oxygen would have a tendency to remove this. Thus, the oxide surface will be less inhomogeneous under the state of equilibrium. The law of Schottky effect, which is valid for a homogeneous surface, is therefore expected to be more at variance with the facts observed initially, i.e., under pulsed condition.

(viii) Lack of any appreciable difference between the work function under pulsed and d.c. condition does not follow easily from the adsorption phenomenon that has been analysed. Any conclusive evidence in this regard again requires a consideration of emission mechanism. Qualitative reasonings of course suggest that a mobile layer on the surface can hardly effect the work function which is determined chiefly by the arrangement of matter in bulk.

(ix) Appearance of square-topped current pulses with voltages of micro-second duration is easily associated with the decay resulting from adsorption, if account is taken of the transit time of the electrons and of the poisoning agent. It is clear that the poisoning agent will arrive on the cathode only after a finite interval of time—an interval determined by the finite time required for the electrons to reach the anode from the cathode plus that required by the poisoning agent to reach the cathode from the anode. If the total interval is of the order of micro-seconds a square-topped current pulse will be obtained for drainage lasting over a period of the same order.

The above discussion shows that the main features of the difference between the pulsed and the d.c. behaviour of an oxide-coated cathode are also in accord with the hypothesis that the decay of current is due to mobile adsorption on the cathode surface of oxygen ejected from the anode by electron bombardment.

5. Concluding Remarks

The analysis carried out shows that both short and long period decay of current may occur in an oxide-coated cathode due to adsorption of oxygen originating from the anode through electron bombardment. Short period decay will result from mobile adsorption and long period ones from immobile adsorption—a deduction which is supported by results of recent experiments.[2, 16] Differences which are usually observed between the performances of such a cathode under pulsed and under d.c. condition of operation may also be explained—at least qualitatively—as being due to mobile adsorption of oxygen released from the anode in sufficient quantity to raise its partial pressure about 10^{-6} mm Hg.

It will be noted that owing to lack of sufficient experimental data, the values that have been assigned to the various parameters in the working equations are only those which appeared most probable. Improved results will undoubtedly be obtained when more experimental data are available. However, any major change in the main trend of the results obtained appears unlikely.

6. Acknowledgments

It is a pleasure to thank Professor S. K. Mitra for his constant encouragement and help in course of the work. Thanks are also due to Professor M. N. Saha and Dr. H. Rakshit for going through the original manuscript and to Messrs. B. K. Banerjee and U. C. Guha for helpful discussion.

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PLASTICS IN THE TROPICS

Behaviour of polythene (polyethylene) under tropical conditions is the subject of a report* resulting from the latest of the tests being carried out at the instance of the Ministry of Supply—British Plastics Federation sub-panel of the Inter-Services Plastics panel. Evaluations of the effect of climatic conditions in the tropics on some properties of plastics are continually being carried out as a long-term policy and a report on tests on expanded plastics has already been published.

The tests into the effects of climatic conditions are essentially long-term and will no doubt be continued at intervals as new and improved materials come into use. A large number of plastics materials are already undergoing tests.

The tests on polythene are regarded as particularly important, as the material is specified for a variety of high frequency electrical applications because of its low power factor and permittivity, and is also suitable for such purposes as packaging and waterpiping because it is ductile and inert.

The report shows that the inclusion of as small an amount as 0·1 per cent. of carbon black in polythene provides a considerable measure of protection against deterioration of properties due to light—usually 2 per cent. of well-dispersed carbon black is incorporated.† At the Ministry of Supply's Tropical Testing Establishment in Nigeria, two grades of polythene, each sample of which contained antioxidant (0·1 per cent. of "Nonoxol" D.C.P.) but which were presented both with and without the inclusion of carbon black (0·1 per cent. "Kosmos" B), were exposed for periods of three, six or 12 months to four sets of conditions—jungle undergrowth, jungle clearing, semi-desert and tropical surf beach.

Conclusions of the report are that polythene is resistant to the moisture of tropical climate as represented mainly by the jungle exposure sites, but is susceptible to the actinic effects of the tropical sun, preponderant on the beach and desert exposure sites. These effects, however, are greatly reduced by carbon black.

The main results of the continued sun exposure were: an increase in "brittleness," exemplified by loss of flexibility and extensibility, a rise in cold bend temperature and the appearance of surface cracks; and a thirty-fold rise of power factor in the absence of carbon black or tenfold rise in the presence of carbon black, under the worst conditions of the trials described in the report.

As ductility and low power factor were the foremost properties leading to the selection of polythene in preference to any other material, the report concludes that it would be necessary either to be satisfied with a limited service life in direct sunlight, or to provide additional protection for the material.

There was no difference in the behaviour of the two grades on trial: "Alkathene" grade 7 is a medium extrusion grade of polythene, and grade 20 softer and of lower average molecular weight.

Among the detailed results are the following:—

Biological examination: There was no indication that fungal growths (of which there were only traces at all sites except the jungle clearing) had attacked the polythene itself. Growths seemed to be confined to extraneous matter on the surface.

Changes in dimensions: A general progressive shrinkage in both directions at all sites (greatest at the semi-desert site, least in jungle undergrowth).

Changes in weight: All specimens showed a progressive loss in weight.

Tensile properties: The tensile strength remained largely unaffected by exposure on all sites, but exposure on the semi-desert and beach sites ultimately lead to a considerable reduction in percentage elongation at break particularly in the absence of carbon black.

Dielectric properties: Permittivity and power factor properties remained practically unaffected in all batches in the jungle undergrowth. Carbon black protected batches in the jungle clearing, where, however, considerable deterioration took place in the absence of carbon black. The protection afforded by carbon black was no longer sufficient for specimens exposed for more than six months at the semi-desert or surf beach sites.

Low-temperature flexibility: The small concentration of carbon black used afforded partial protection against embrittlement, which, however, is nearly exhausted after one year on the sunny sites. Unprotected material has a useful life as regards brittleness of six months or less in the desert, less than a year on the surf beach and about one year in the jungle clearing.

* "Reports on Plastics in the Tropics—2. Polythene" (H.M. Stationery Office, 1953, price 2s).

† Reference to this effect and its remedy was made in two 1951 Convention papers: L. W. Meyers, "Plastic-insulated land communication cables," *J.Brit.I.R.E.*, 11, December 1951, p. 559; A. Cross and F. R. Yardley, "Wireless broadcasting and rediffusion systems for colonial territories," *J.Brit.I.R.E.*, 12, Feb. 1952, p. 98.

CODES OF PRACTICE

In a written answer to a question raised in the House of Commons on March 9th, the Minister of Works replied: "In agreement with the Professional Institutions the British Standards Institution are establishing a Council for Codes of Practice to be responsible for all work on codes of practice, including building construction and civil, mechanical and electrical engineering codes. The Chairman-elect is Mr. Allan Stephen Quartermaine, C.B.E., M.C., Past President of the Institution of Civil Engineers.

"After March 31st the services carried out by my Department for the present Council for Codes of Practice for Buildings will be undertaken by the British Standards Institution."

In the past 12 years the Council for Codes of Practice for Buildings under the Ministry of Works has prepared more than 80 codes, whilst others affecting engineering techniques outside the scope of the Council have been produced independently by the relevant professional bodies. Although the British Standards Institution has thus not been considered the responsible authority for the preparation of the codes, it has always been one of the consultative groups and actually published the majority of the codes.

The need for the integration of the Codes of Practice in all spheres was contained in the Cunliffe Committee report which investigated the organization of B.S.I. in 1950.

There has been therefore set up a Council for Codes of Practice within the B.S.I. which will be concerned with building and a wide range of interests in engineering. Representatives will be appointed from B.S.I. and 15 founder organizations, and invitations to appoint representatives are to be extended to other interested organizations as may be deemed appropriate from time to time.

Proposals for preparing a code may originate from a professional organization, from within the Codes of Practice Committee, or from outside. Whatever the origin, the proposal will be referred by the Council to the appropriate Codes of Practice Committee for an opinion on the action to be taken. If it is decided to proceed, this Codes of Practice Committee will, with the approval of the Council, either invite an appropriate professional society to set up a committee to carry out the Code-drafting work or otherwise arrange for a committee to be set up for this purpose.

At the outset four Codes of Practice Committees will be appointed as follows:—

Codes of Practice Civil Engineering Committee.

Codes of Practice for Building, Construction and Engineering Services Committee.

Codes of Practice Mechanical Engineering Committee.

Codes of Practice Electrical Engineering Committee.

In the radio engineering field only five Codes of Practice have been prepared so far, two under the Codes of Practice for electrical equipment in buildings and three engineering Codes of Practice.

CP 327.201 : 1951. Broadcast reception—sound and television by radio.

CP 327.300 : 1952. Sound distribution systems.

CP 1001 : 1947. Abatement of radio interference caused by motor vehicles and internal-combustion engines.

CP 1002 : 1947. Abatement of radio interference from electro-medical and industrial radio-frequency equipment.

CP 1005 : 1954. The use of electronic valves (Parts 1 and 2). See below.

CP 1005 : 1954. The Use of Electronic Valves (Parts 1 & 2).

The first two parts of the British Standard Code of Practice on the use of electronic valves have now been published.

Part 1 gives general information and recommendations for all types of electronic valves, whilst Part 2 sets out additional recommendations on the use of domestic receiving valves, cathode-ray tubes and rectifiers. Further parts giving recommendations for other types of valves are now being prepared and will be published in due course.

The Code of Practice, which is intended to give guidance to designers of equipment using electronic valves so that optimum performance and life may be achieved, will be revised and extended from time to time to keep it abreast of developments.

APPLICANTS FOR MEMBERSHIP

New proposals were considered by the Membership Committee at a meeting held on March 29th, 1954, as follows: 26 proposals for direct election to Graduateship or higher grade of membership and 22 proposals for transfer to Graduateship or higher grade of membership. In addition, 63 applications for Studentship registration were considered. This list also contains the names of two 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.

Direct Election to Full Member

DUNN, Wg. Cdr. Walter Edward. *Carshalton Beeches, Surrey.*

Direct Election to Associate Member

APTE, Bhagwan Madhav. *Burbank, California, U.S.A.*
COCKER, Sqn. Ldr. Francis Stanislaus. *Abu-Sueir.*
DEB RAY, Flt. Lt. Mihir, M.Sc. *Jalahalli, Bangalore.*
DE SA, Joseph Francis Brian. *Madras.*
EVANS, William Francis. *London, N.W.9.*
FIELDING, Henry George. *Qatar.**
GOOLRY, Lt.-Col. Kirpal Singh, B.A. *Bombay.*
RIDLEY, Gp. Capt. Leslie Ralph, O.B.E. *London W.C.1.*
SHAW, Lieut. Com. (L) Simeon Basil. *Barnstaple, Devon.*
SZEKELY, Georges Gavriel. *Tel-Aviv.*
WEBSTER, Bernard Raymond, B.Sc.(Eng.). *Southend-on-Sea, Essex.*

Transfer from Associate to Associate Member

BRAGANZA, Lt.-Col. Joseph Vincent Paul. *Poona.*
CARR, Lieut. (L) Alistair Gordon, B.Sc. *Halifax, Nova Scotia, Canada.*
DAVIES, Christopher Sylvester. *Freetown, Sierra Leone.*
FRYER, Frank Hardman. *Lichfield, Staffordshire.*
NEELY, Terence Joseph Prescott. *Grahamstown, South Africa.*
TOMS, David Morgan Storer. *Sidcup, Kent.*
WYMAN, Lieut. Com. (L) Kenneth Henry. *Corsham, Wiltshire.*

Transfer from Graduate to Associate Member

JENKINS, Ieuan Huw, B.Sc.(Hons.). *Sunbury-on-Thames, Middlesex.*
KRISHNAMURTHI, Namipudi, B.Sc.(Hons.), M.Sc. *Secunderabad.*
NUTTALL, Lieut. (L) Leslie Frederick. *London, W.10.*

Direct Election to Associate

BAILEY, George Joseph. *Ilford, Essex.*
BONAGE, William Frederick. *Lagos, Nigeria.*
BROWN, Lieut. (L) Kenneth Stanley. *Lossiemouth, Morayshire.*
GILBERT, John. *London, W.5.*
KILROY, James. *Liverpool.*
MAY, Harold. *Feltham, Middlesex.*
ROBERTSON, Robert Gordon. *Wau, Sudan.*

Transfer from Student to Associate

RAINE, George. *Newcastle-upon-Tyne.*
SHERWIN, Ronald. *Chelmsford, Essex.*

Direct Election to Graduate

ASOKENDU MOZUMDER, B.Sc., M.Sc. *Calcutta.*
BRIGGS, Richard Richardson. *Larbert, Stirlingshire.*
BURROWS, Keith, B.Sc.(Hons.). *Yorkshire.*
FITCH, Peter Mark, B.Sc. *Ruislip, Middlesex.*
SANSOM, John Stuart. *London, W.13.*
STOCKS, Edward John, B.Sc.(Eng.). *Ilford, Essex.*
WILLIAMS, Raymond Henry. *London, E.11.*

Transfer from Student to Graduate

ANDERSON, Charles William Michael. *Waterford, Eire.*
CAMERON, Archibald. *Glasgow.*
CURRY, Eric. *Parkstone, Dorset.*
NEWMAN, Robert Hanmer. *Haywards Heath, Sussex.*
PENFOLD, Reginald Charles. *Malvern, Worcestershire.*
PHILLIPS, James Hugh. *London, W.13.*
RAMANATHAN, P. V. N., B.Sc.(Hons.). *Allahabad.*
YAM YAU BAN. *Hong Kong.*

Studentship Registrations

ACHOW, Carl Ruthven. *London, N.19.*
BAIGENT, John Andrew Mortimer. *Surbiton, Surrey.*
BANNISTER, Frederick James. *London, S.E.3.*
BEARDALL, James Howitt. *London, N.W.3.*
BENERAGAMA, Don Kingsley Wimalasiri. *Kalutara South, Ceylon.*
BEZUIDENHOUT, Christoffel Jacobus. *Pietersburg, South Africa.*
BHARUCHA, Jehangir. *Bexleyheath, Kent.*
BHATTACHARYA, J. N., B.Sc. *Naihati, West Bengal,*
BHATTI, Mohammed Aslam. *Rawalpindi.*
BLUE, Johnston Richmond. *Glasgow.*
BOLAR, Jayanth Purshotam, B.Sc. *Bangalore.*
BOWN, Geoffrey Charles Stanley. *Penarth, Glamorgan.*
CAMPBELL, Malcolm, B.Sc. *South Croydon, Surrey.*
CHARANJIT SINGH. *Agra.*
CHEW BAK KHOOK. *Singapore.*
CRIDLAND, William Wyndham. *Liverpool.*
CURRY, Thomas William. *Pietersburg, South Africa.*
DOCKERTY, William Harold. *Henley-on-Thames, Oxfordshire.*
ELLIS, Brian Norman. *Cambridge.*
FARNWORTH, Geoffrey. *Manchester.*
FERLIE, Thomas Darling. *Coventry.*
GERRITSEN, Johannes Theodosus Aloysius. *Apeldoorn, Netherlands.*
GURUSWAMY, K. R. *Madras.*
HELPS, Robert Lewis, B.Sc. *Stoke Poges, Buckinghamshire.*
HORSFALL, Frank Alan. *Pietersburg, South Africa.*
JARVIS, John Walter. *Sidcup, Kent.*
JEFFERY, Falga Terence. *Nairobi, Kenya.*
KALYANASUNDARAM, G. *Kallakurchi, Madras.*
KANAAR, John Christopher Norris. *London, W.13.*
KEEBLE, Derek Leslie. *Harlow, Essex.*
KELLY, Joseph. *Formby, Lancashire.*
KHAN, Maqbool Husain. *Karachi.*
KING, Gordon Henry. *Gympie, Queensland.*
KRISHNAN, S. *Mylapore, Madras.*
KULKARNI, Dattatray Yeshwant. *Bombay.*
MCLAREN, Robert Ian. *Edinburgh.*
MAHER, Stephen. *Anderton, Lancashire.*
MAHAJAN, Prabhakar Shankar. *Dadar, Bombay.*
MAMO, Denis Ronald. *Malta, G.C.*
MATHUR, Sushil Kumar. *Jaipur.*
MUKHOPADHYAY, 2nd Lt. P. M., B.E. *Bangalore, India.*
MURPHY, Dermot Michael. *London, W.12.*
NAIDU, M. K. R. *Madras.*
O'DONNELL, John Gerald. *Donnycarney, Dublin.*
POLISHKO, W. *Adelaide, South Australia.*
RAMACHANDRA RAO, Udipi, M.Sc. *Ahmednagar.*
ROBERTS, William Edward. *Plymouth.*
ROMANOWSKI, Richard. *London, N.W.6.*
SABARSOEDIMAN. *Bandung, Indonesia.*
SEKHAR, Yadavalli Sivarama Chandra. *Calcutta.*
SESHA, Venkatarama Adi. *Delhi.*
SIDHU, Gulzar Singh. *New Delhi.*
SMITH, David Trevor, B.Sc. *Fleet, Hants.*
STEPHEN, Sidney George. *Llandaff, Cardiff.*
STOCK, Peter Anthony. *New Malden, Surrey.*
STRIDE, Regent. *London, S.W.19.*
TURNER, Brian Robert. *Heybrook Bay, Devon.*
VAN DER WALT, Leopold. *Durban, South Africa.*
VAN WIERINGEN, George Alfred. *Singapore.*
VAN ZYL, Nicolaas. *Welgedacht, South Africa.*
VERSTER, Ryno. *Bloemfontein, South Africa.*
WOO CHI KONG. *Hong Kong.*
WORDSWORTH, Geoffrey. *Stevenage, Hertfordshire.*

* Reinstatement.

EXPERIMENTAL RADIO BEARER EQUIPMENT FOR CARRIER TELEPHONE SYSTEMS*

by

W. S. McGuire† and A. G. Bird†

SUMMARY

Two types of frequency-modulated radio bearer circuits are described, one operating in the frequency range of 420 to 470 Mc/s, and the other in the frequency range 860 to 960 Mc/s. Both systems are crystal controlled with a frequency deviation of ± 180 kc/s and each capable of carrying 12 to 17 telephone circuits.

The description includes a radio repeater operating on the heterodyne principle which permits long distances to be traversed with a minimum of accumulated noise and distortion. The performance of both systems meets C.C.I.F. transmission specifications when used with the appropriate carrier equipment. Various aspects of supervisory and order-wire facilities, together with alarm and automatic change-over equipment are discussed.

A typical installation, operating in the 420-470 Mc/s band and consisting of a transmitter, repeater and receiver, is described and its performance over a 100-mile circuit is given.

1. Introduction

The demand for trunk telephone circuits has increased considerably during the past few years, thus overloading the available lines to remote centres. Over these years, v.h.f. and u.h.f. multi-channel radio technique has developed into an efficient and reliable method of telecommunication, which has proved to be particularly suitable for rapid extension of line networks.

The radio bearer circuit, when used in conjunction with appropriate frequency-division carrier equipment, also offers many advantages for remote-controlled transmitting and receiving centres. In such circumstances it is common practice to use a one-way circuit from the receiving station to the control centre and a one-way circuit from the control centre to the transmitting centre. By using appropriate voice-frequency telegraph channel equipment, one or more voice channels may be further subdivided to provide a number of control functions over the radio bearer circuit.

The two radio bearer systems to be described employ frequency modulation with a frequency deviation of ± 180 kc/s and operate in the frequency ranges of 420 to 470 Mc/s and 860 to 960 Mc/s; both systems are directly crystal controlled. The base-band frequency response is constant within ± 1 db from 250 c/s to 4 kc/s and within ± 1 db from 4 kc/s to 70 kc/s. The former section of the base-band is normally used for order-wire and supervisory circuits and the latter section is used for frequency division carrier equipment. A description is also given of a radio repeater system using the heterodyne principle in the above frequency range. Facilities are described for adding or dropping carrier channels at each repeater, together with a party line telephone circuit for supervisory purposes.

2. System Planning

Whilst it is beyond the scope of this paper to deal in detail with the all-important subject of system planning a few of the outstanding problems will be briefly discussed.

2.1. *Operating Frequencies*

In the past, most multi-channel systems were restricted to the frequency range 160 to 250 Mc/s and the microwave range (3,000 to 4,000 Mc/s). This was due mainly to limitations of available valves, particularly the output valve for the terminal and repeater transmitter. With the recent introduction of new low-power transmitting valves, such as the 4X150A, 4X150G,[4] DET. 24[9] and QQE06/40,[11] considerable interest is now being taken in the frequency range from 250 Mc/s to 900 Mc/s.

* Reprinted from *The Proceedings of the Institution of Radio Engineers, Australia*, Vol. 14, No. 6, June, 1953. (Paper No. 261.)

† Amalgamated Wireless (A/sia) Ltd., Ashfield, Sydney, N.S.W.

U.D.C. No. 621.396.6 : 621.395.44.

2.2. *Number of Channels*

The number of telephone channels a radio bearer circuit is capable of carrying depends upon a number of factors, some of which are:

(*a*) base-band frequency response of radio bearer circuit.

(*b*) linearity of the radio equipment.

(*c*) type of carrier equipment and bandwidth of each channel.

(*d*) signal-to-noise ratio permissible in each channel.

(*e*) the distance between terminal and repeater stations and the nature of the intervening terrain.

The degree of linearity and signal-to-noise ratio required will be governed by the type of service for which the equipment is intended.

When the radio bearer circuit is to be included as part of the telephone trunk network the circuit must at all times meet the minimum C.C.I.F. requirement, i.e. the total open-circuit psophometric voltage due to noise and cross-talk in a 600-ohm cable circuit shall not exceed 2 mV when measured at a relative level of —6 db. Radio bearer circuits used for conveying information to and from telecommunication centres may have less stringent requirements in regard to signal-to-noise ratio and linearity. For this type of service it is often possible to increase the number of channels and/or increase the separation between terminal and repeater stations and still provide an adequate circuit.

2.3. *Power Supply Problems*

When the system consists of terminal equipment only, and is situated at main communication centres, there is usually adequate power available from the mains supply, and from elaborate emergency supplies. The main problem is usually at a repeater station where the power supply may fail due to zoning or other causes. The degree of reliability required from a telephone communication system is often such that an alternative supply must be provided and arrangements made for automatic change-over.

The choice of an alternative power supply will be governed by the expected duration of power failure and power consumption of the equipment. In its simplest form the emergency supply may be a rotary converter operated from batteries which are charged by the mains. For larger power supplies, petrol or diesel-driven alternators may be used; either system requires batteries.

It would appear from the above that in any case batteries are required, and it may be more economical to operate the equipment from batteries, both h.t. and heaters, at all times and float the batteries across a mains-operated battery charger. This would allow uninterrupted operation of the equipment during power failures with minor changeover facilities.

Wind-driven generating equipment is another possibility. It is interesting to note that both the radio repeater and the tower for the wind-driven generating equipment require an elevated site. In such circumstances a common tower could be used to mount the aerials and the wind-driven equipment.

2.4. *Choice of Sites*

The choice of suitable sites for the terminal and repeater equipment involves a number of considerations and, in general, cannot be satisfactorily determined until some field tests have been carried out. These tests should be carried out at or near the final operating frequency and for as long a period of time as is economically possible. (Whilst it is possible, in the 450 and 900 Mc/s bands to use the same frequency allocation at alternate repeater stations, it is advisable to choose the path so that the sites are not in a direct line. In general the physical separation between sites should be restricted to between 40 or 50 miles and should be in direct line of sight. At 450 Mc/s satisfactory results have been obtained on paths other than line of sight, but such paths should be restricted to 15 miles or less and backed up by field tests.) Other factors such as accessibility for maintenance and availability of a public electricity supply will also be important in determining the route over which circuits are most conveniently operated.

2.5. *Modulator*

In general there are two types of modulators in common use as exciters for v.h.f. and u.h.f. link transmitters. The first type is the reactance-controlled oscillator which produces direct frequency modulation and is capable of excellent linearity providing the percentage deviation is reasonably small. The main disadvantage is that it requires a more-or-less elaborate automatic-frequency-control system of which there are two common types. The first comprises variations of the Crosby ring arrangement in which a control voltage is obtained from a discriminator via a reference crystal and i.f. amplifier; the control voltage is applied to the reactance valve grid in such a manner as to correct the outgoing frequency. The second type of automatic frequency control is derived from a chain of frequency dividers and a low-frequency reference crystal, the differential voltage being applied to a motor which rotates a small capacitor associated with the LC oscillator; this corrects the outgoing frequency as referred to the low-frequency crystal. Both arrangements depend upon a number of valves and circuit components outside the main circuit for the frequency control of the transmitter.

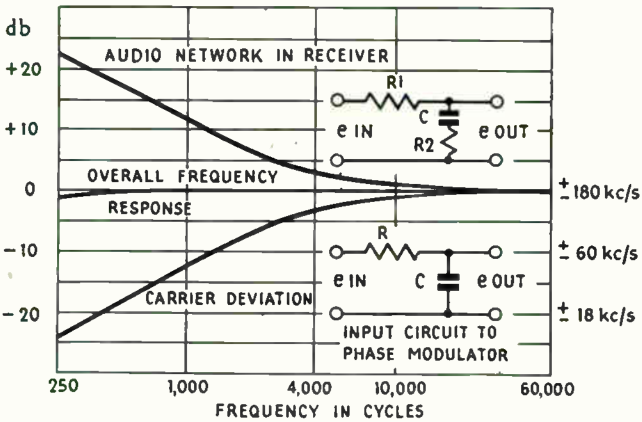


Fig. 1.—Curves showing the carrier frequency deviation with modulating frequency and the correction network response in receiver audio circuit.

The second type of exciter in common use is the phase modulator with associated network to produce frequency modulation. The phase modulator has the fundamental advantage that the centre frequency of the emitted wave is directly crystal controlled, thus eliminating one of the major causes of erratic operation. The difficulty associated with most phase modulators is the large degree of frequency multiplication necessary to obtain the final carrier deviation, as compared with the direct f.m. type of modulator.

The modulator used in the equipment to be described is a phase modulator of the delay line type, similar in principle to that previously described[5] and used in an early type of 160-Mc/s radio relay system.[10] The aim was to develop a modulator to act as an exciter unit for a link transmitter, operating in the 450 and 900-Mc/s bands and having the following characteristics:

Modulation frequency range: 250 c/s to about 70 kc/s.

Type of modulation: f.m.

Carrier frequency: 450 and 900 Mc/s bands.

Carrier deviation: ± 180 kc/s.

Minimum modulation index: 3.

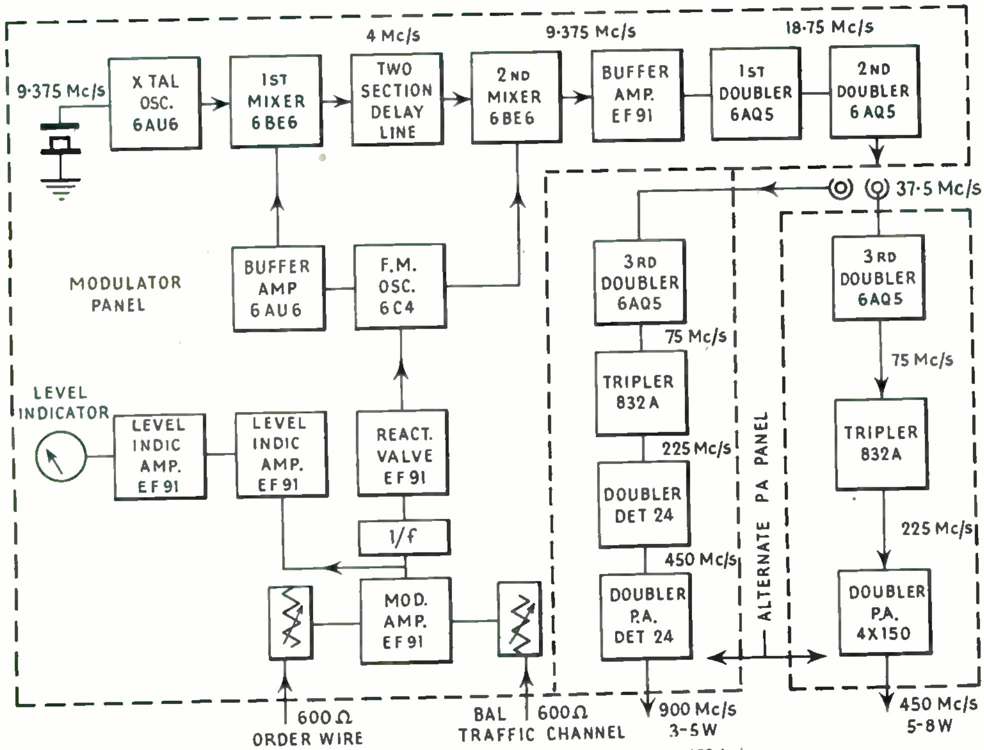
Noise and distortion: at least 60 db down in a 4-kc/s channel.

It can be seen that if pure f.m. is used throughout the base-band, at 240 c/s the modulation index would be 180/0·24 = 750 and would require a large degree of frequency multiplication. As the carrier equipment used in conjunction with this system does not use modulating frequencies below 4 kc/s, it would appear that such frequencies could be considered separately from those above 4 kc/s. Frequencies below 4 kc/s can only carry one voice channel and this band is normally used as an order-wire circuit for supervisory purposes.

It is thus possible to use a system in which constant frequency deviation (f.m.) is used for frequencies above 4 kc/s and phase modulation below 4 kc/s. By using the usual inverse-frequency network in the modulator input, with the 3 db point placed at 4 kc/s, the modulation is pure phase at 250 c/s with a transition to pure frequency modulation at 60 kc/s. Table 1 illustrates the type of response. It can be seen that the modulation index at 250 c/s is now reduced from 750 to 45, thus reducing the frequency multiplication in the transmitter by 16 times.

TABLE 1

Mod. Frequency	Mod. Index	Final Deviation
250 c/s	45·0	11·25 kc/s
1 kc/s	43·4	43·4 ,,
2 ,,	39·85	79·7 ,,
4 ,,	31·8	127·2 ,,
10 ,,	16·75	167·5 ,,
60 ,,	3·0	180·0 ,,



Total multiplication: 48; Modulation index: 3; Deviation: ±180 kc/s.
Audio frequency network:—250 c/s–4 kc/s: phase modulation; 4 kc/s 60 kc/s: frequency modulation.

Fig. 2.—Block diagram of 450-Mc/s and 900-Mc/s terminal transmitter.

The maximum angular deviation required at the final frequency is now ± 45 radians; with the delay line type modulator this represents a frequency multiplication of 48 times for a deviation of ± 180 kc/s at 450 Mc/s as compared with a multiplication of 768 times if pure f.m. were used down to 250 c/s. Fig. 1 shows the carrier frequency deviation with modulating frequency and the correction network response in the receiver audio circuit.

3. Terminal Transmitter

The block circuit diagram of Fig. 2 shows the general arrangement of the modulator and frequency multipliers, together with the 4-Mc/s delay line panel and power amplifier stages. Fig. 3 illustrates the rear view of the 450-Mc/s terminal transmitter and Fig. 4 illustrates the power amplifier stages used on the 900-Mc/s equipment; the top unit shown in Fig. 3 is replaced by the panel shown in Fig 4 for the 900-Mc/s system. No specific description is given of the power supply units for operation of the equipment as these may be of any convenient form; the maximum high-tension voltage does not exceed 300 to 350 volts.

Briefly, the modulator depends for its operation upon the time delay which occurs when an f.m. signal is passed through a multisection band-pass filter. A signal from a fixed-frequency source is mixed with an f.m. signal from a reactance-valve modulated oscillator; the f.m. difference-frequency component is selected, and after passage through a time-delay network is once more mixed with the original f.m. signal. The difference-frequency component in the output of the second mixing stage has the same centre frequency as the fixed source and is phase modulated; frequency modulation is obtained when the base-band frequency is passed through an inverse-frequency network as described previously.

The fixed-frequency source is obtained from a crystal, operating in the vicinity of 10 Mc/s, thus

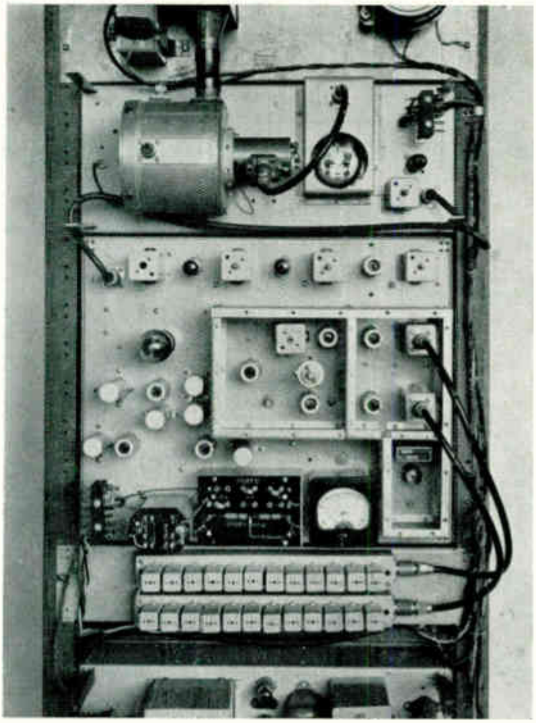


Fig. 3.—Rear-of-rack view of the 450-Mc/s terminal transmitter.

providing a transmitter frequency accuracy of 0·003 per cent. over a wide range of temperature and mains variation.

3.1. *Baseband Input Circuits*

In normal use, two independent pairs of modulation-input terminals are required—(*a*) to handle the multiplex signal from the carrier equipment, and (*b*) to handle an order-wire and/or alarm circuits. It is desirable that these two input circuits be so isolated that significant power is not directly transferred from either signal source to the other.

There are numerous ways in which the desired isolation may be achieved. One method is to isolate each signal source by a separate attenuator and connect a bridging amplifier across the junction of the two attenuators; sufficient gain is then provided in the bridging amplifier to offset the loss in the attenuators. This arrangement has the advantage that feed-back may be applied to the amplifier without destroying the isolation as would be the case with an amplifier with separate grids and a common anode load.

The multiplex signal is connected via a 600-ohm balanced transformer and variable attenuator to the grid of the bridging amplifier, which in turn feeds the inverse-frequency network in the input to the reactance-controlled oscillator.

3.2. *Frequency Multipliers*

The modulator panel houses an amplifier and the two early frequency multiplier stages, each stage being transformer-coupled to provide a band-pass circuit. Whilst the frequency deviation in the early stages is only ± 3·8 kc/s, the coupling circuits must have a band-pass of at least twice the highest modulating frequency (72 kc/s) and still provide adequate attenuation for harmonics other than the desired one. This constitutes one serious disadvantage of phase modulators using low-frequency crystals of the order of 500 kc/s, as there is a great danger of producing a number of spurious signals around the final frequency, spaced at intervals equal to the crystal frequency.

The output from the modulator is link-coupled to the input of the power amplifier panel which contains a 6AQ5, doubling to 75 Mc/s, followed by an 832A push-pull tripler to 225 Mc/s.

3.3. *Final Amplifier Stage*

The final amplifier in the 450-Mc/s terminal transmitter employs a 4X150A tetrode, operating as a frequency doubler to 450 Mc/s.

The grid connection on the 4X150A is made to the centre pin of an 8-pin loctal socket. By modifying the socket, direct connection may be made between the grid pin and inner conductor of a quarter-wave coaxial grid circuit. The grid line is considerably shorter than a quarter-wave, due to the valve input capacity. Additional capacitance is lumped across part of the line in the form of a self-supporting grid tuning capacitance. The opposite end of the line is earthed to r.f. by a disc-type mica capacitor and bias is fed to the grid via the grid line. Four cathode pins are provided, each pin being earthed by a short copper strap to the main cavity.

The anode circuit also uses a capacitance shortened quarter-wave cavity. The inner conductor is a copper tube the same diameter as the anode and contact is made with the anode by spring fingers; the circuit is tuned by a small variable capacitance tapped across part of the line.

The screen-grid is earthed to r.f. by a special mica capacitor attached to the inner wall of the anode cavity. The input and output circuits are both inductively coupled to their respective lines using small loops of copper strip.

A small blower is used to cool the valve; air is forced into the anode cavity where it circulates and leaves the cavity via the valve anode. A number of small holes in the wall separating the grid-anode cavity bypass a small amount of air for cooling the valve base connections. The valve manufacturers specify an air flow around the valve base, irrespective of the anode dissipation.

By totally enclosing the input and output circuits, the loss due to radiation is extremely low and this, in turn, produces high tank circuit efficiency and a high Q for the output circuit.

Whilst the 4X150A has an anode dissipation considerably in excess of that required for the link, it is always advisable to operate the output valves, where possible, well below their rated dissipation. The 4X150A is operated under the following conditions when doubling from 225 Mc/s to 450 Mc/s:

$$\left.\begin{array}{l} E_a = 350\text{ V} \\ I_a = 56\text{ mA} \end{array}\right\} \text{anode input} = 19{\cdot}5 \text{ watts,}$$

$$E_{sg} = 120\text{ V}$$

$$\left.\begin{array}{l} I_g = 2{\cdot}1\text{ mA} \\ \text{Bias} = -45\text{ V} \end{array}\right\} \text{grid resistor 30,000 ohms,}$$

Power output = 9·5 watts into 50 ohm load,

Efficiency = 49 per cent.

3.4. 900-*Mc/s Terminal Transmitter*

For the 900-Mc/s terminal transmitter an additional doubler stage is added to the system. Initially a 4X150A was used as the frequency doubler to 900 Mc/s; this arrangement produced an output of 3 to 5 watts but was rather inefficient. The main limitation to the use of a 4X150A for the 900-Mc/s system was the inability of the valve to function as a power mixer at these frequencies for the repeater transmitter. The repeater transmitter for the 450-Mc/s and 900-Mc/s system is described in detail in a later section and as it was considered most desirable to use the same type of valve in the terminal and repeater transmitter, the 4X150A was discarded in the 900-Mc/s system.

The 900-Mc/s terminal transmitter uses the same modulator and frequency multiplier chain up to the 225-Mc/s stage as the 450-Mc/s system. This is followed by two doubler stages using grounded-grid disc-seal triode valves (type DET.24) which are arranged for common-grid, earthed-anode coaxial circuits.

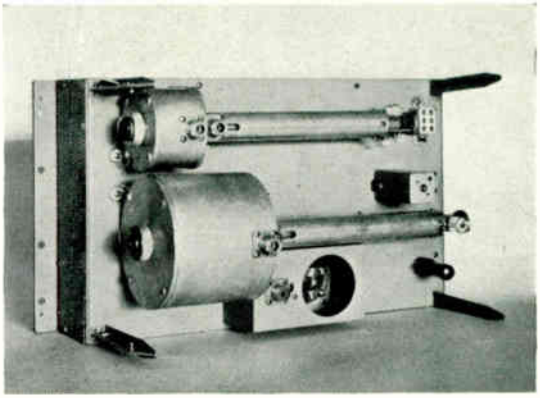


Fig. 4.—Rear-of-rack view of 900-Mc/s final multiplier unit.

The 225-Mc/s signal is fed to the grid-cathode line which is operated in the quarter-wave mode and is tuned by adjusting the line length by means of a non-shorting plunger. The grid-anode cavity is tuned to 450 Mc/s by means of plugs at the high and low impedance end of the cavity. The output from the 450-Mc/s stage is fed into the grid-cathode line of the second doubler stage; the grid cathode and anode-grid lines are the same as the 450-Mc/s stage except that both circuits operate at twice the frequency. Each stage is separately biased by means of an adjustable resistor in the junction of the cathode-line and earth. A special mica capacitor earths the anode for r.f. potentials but allows d.c. to be fed to the valve.

The valve is metered separately in the cathode and anode circuit and the difference in reading between the two meters is then equal to the grid current.

The DET.24 is operated under the following conditions when doubling from 225 Mc/s to 450 Mc/s, followed by a second doubler from 450 Mc/s to 900 Mc/s.

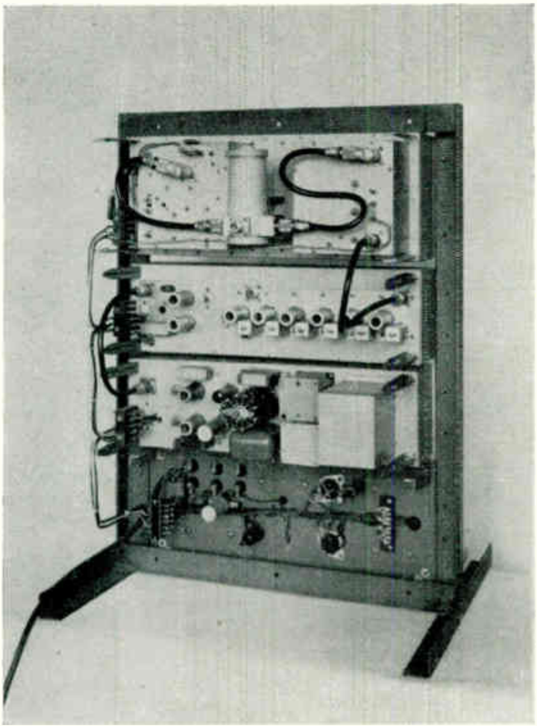


Fig. 5.—Rear view of 450-Mc/s receiver.

First stage doubling from 225 Mc/s to 450 Mc/s:

Condition (1)	*Condition* (2)
E_a = 315 V	E_a = 350 V
I_a = 26 mA	I_a = 30 mA
I_k = 35 mA	I_k = 40 mA

Second stage doubling from 450 Mc/s to 900 Mc/s:

Condition (1)	*Condition* (2)
E_a = 315 V	E_a = 350 V
I_a = 35 mA	I_a = 40 mA
I_k = 44 mA	I_k = 50 mA
Bias = 44 V	Bias = 50 V
Anode input = 9·6 W	Anode input = 15 W
Power output = 5 W	Power output = 6·5 W

4. Terminal and Repeater Receiver

This section describes receiving equipment operating in the frequency bands of 420 to 470 Mc/s and 860 to 960 Mc/s. Equipment operating in the 420 to 470-Mc/s band has been chosen as the example, with a view to adapting it to the higher frequency assignment. The circuit arrangement of the 860 to 960-Mc/s unit is the same except where modifications are required on account of the signal frequency; the only modification is an additional frequency multiplier stage in the local oscillator circuit and a mixer cavity capable of tuning to the assigned band. Several design problems are discussed and a description of the various units is given.

A block diagram of the receiver appears in Fig. 7 and an illustration in Fig. 5. The three units, excluding the power supply, are contained in the rack and consist of the following equipment:—

The top unit is the r.f. head comprising the crystal oscillator, multiplier stages, mixer cavity, and low-noise i.f. amplifier input unit; the unit immediately below is the main i.f. amplifier channel with detector and first audio amplifier stage; the remaining unit is the audio amplifier for "traffic" and "supervisory" circuits.

The choice of r.f. amplifiers and mixers becomes a problem at these frequencies, and two separate design trends are now followed. A block circuit diagram illustrating the use of a thermionic valve radio-frequency amplifier and mixer is given in Fig. 6 and is the first basic design. Several miniature valves are suitable for operation at 500 Mc/s, such as the 6J4 grounded grid triode and the 6J6 twin triode, but with the higher frequency assignment in view it was decided that if valves were to be used it would have to be the disc-seal type, which include the 2C40, DET.22, DET.23, DET.24 type valves.

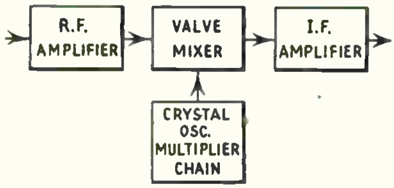


Fig. 6.—Block diagram illustrating the use of a thermionic valve mixer with r.f. amplifier.

An important characteristic of a u.h.f. receiver is its ability to amplify weak signals to a usable level, which means the apparatus must introduce as little noise as possible; thus the noise factor must be made low. With such valves as the 2C40 r.f. amplifier and mixer, noise factors of

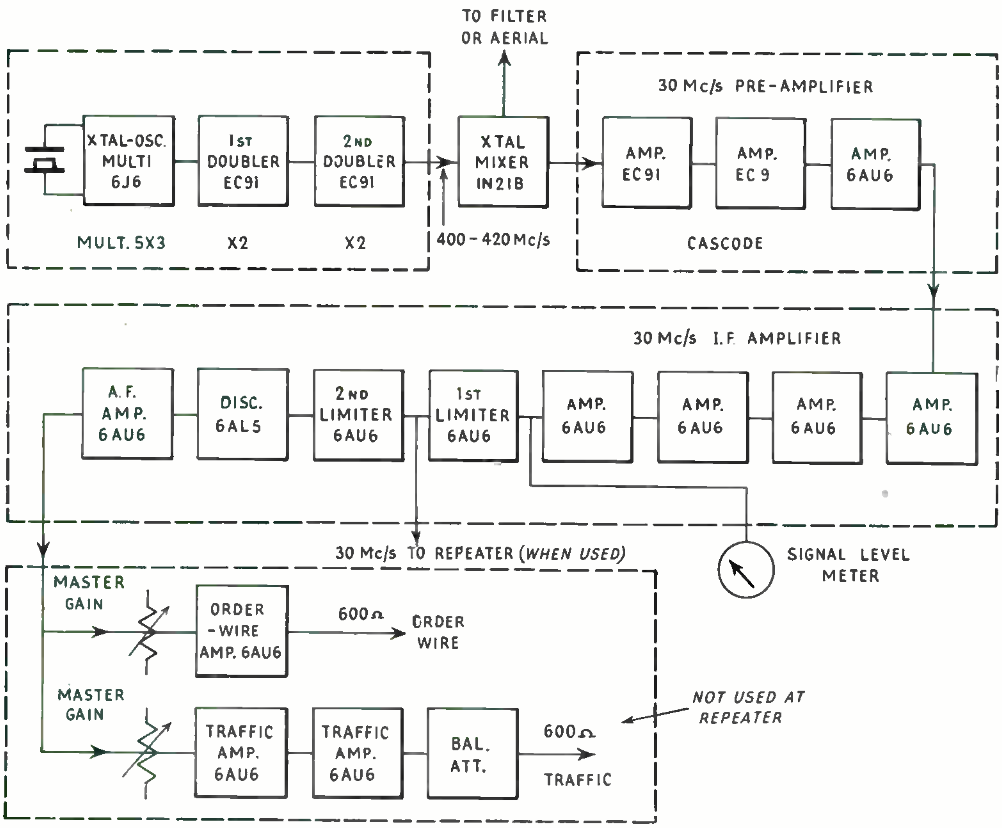


Fig. 7.—Block diagram of 450-Mc/s terminal or repeater receiver.

10 db can be achieved as the following calculation will indicate.

In the 420 to 470-Mc/s range the noise factor of a 2C40 r.f. amplifier and mixer will be 7·5 db and 15·5 db respectively, and a gain of 10 db can be expected from the amplifier;[8,12] thus the noise factor of a receiver can be calculated.

Fig. 8.—Cascade arrangement of networks.

The noise factor for networks in cascade,[3] illustrated in Fig. 8, is given by

$$F_{12} = F_1 + \frac{F_2 - 1}{G_1}$$

where F_1 = noise factor of network 1 expressed as a power ratio,

F_2 = noise factor of network 2 expressed as a power ratio,

G_1 = gain of network 1 expressed as a power ratio,

F_{12} = noise factor of networks 1 and 2 in cascade.

Thus, for the 2C40 r.f. amplifier and mixer, $F_1 = 5{\cdot}6$, $F_2 = 35{\cdot}5$, $G_1 = 10$ and

$$F_{12} = 5{\cdot}6 + \frac{35{\cdot}5 - 1}{10}$$

$$= 9{\cdot}05 \text{ (or } 9{\cdot}56 \text{ db)}$$

Similarly the noise factor at 900 Mc/s would be 13 to 14 db.

Thus, assuming the noise factor to be the main consideration, the 2C40 disc seal triode would be acceptable as the r.f. amplifier and mixer for either frequency band. However, the cost of this type of valve added to that of the mechanical design and construction of tunable cavities makes them uneconomical, so that it was decided to check the performance of a crystal mixer followed by a low-noise i.f. amplifier. The above case using the 2C40 does not take into account the i.f. channel noise factor as the mixer noise is considerably higher.

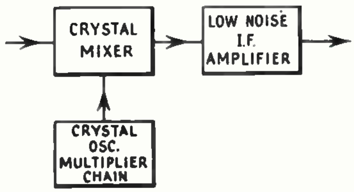


Fig. 9.—Block diagram showing an alternative r.f. head design using a crystal mixer.

The block diagram of Fig. 9 shows an alternative r.f. head design. The r.f. amplifier and mixer are replaced by a crystal mixer, which is followed by a low noise "cascode" i.f. amplifier. The noise factor of this arrangement is dependent primarily on the intermediate frequency as the average 1N21B crystal mixer has a constant loss of approximately 6 db up to 1,000 Mc/s.

The formula for noise factor can be used for a receiver with a crystal mixer and low-noise i.f. amplifier, if we let

F_1 = noise factor of crystal mixer,

F_2 = noise factor of i.f. amplifier,

and G_1 = gain of crystal mixer.

Thus, for a 1N21B crystal mixer and an i.f. amplifier with a 3-db noise factor, we can write

$$F_1 = 8,\ F_2 = 2,\ G_1 = \tfrac{1}{4}$$

$$\text{and } F_{12} = 8 + \frac{2-1}{\tfrac{1}{4}}$$

$$= 12 \text{ (or 10·8 db)}$$

The noise factor in both cases does not include noise developed in the local oscillator, but as it is crystal controlled, noise should be quite low. A figure of 10·8 db for a crystal mixer and low noise amplifier will be constant to approximately 1,000 Mc/s but the same general performance can be expected from a crystal mixer followed by a low-noise i.f. amplifier, as from a thermionic r.f. amplifier and mixer in this frequency range. Fig. 10 illustrates the noise factor of a typical arrangement.

Finally, the crystal mixer is generally preferred on account of the lower cost and simpler circuitry.

4.1. *Crystal Mixer Unit*

The top unit of the receiver illustrated in Fig. 5 shows the crystal mixer cavity centrally mounted. It is a $\lambda/4$ cavity, capacitance loaded at the high impedance end so as to provide a tuning adjustment. The crystal is mounted inside the cavity, near the short-circuited end and is readily accessible in case of replacement. An insertion loss of less than $\frac{1}{2}$ db is achieved by making the ratio of unloaded Q to loaded Q greater than 17, and additional selectivity is obtained by means of coupled filter units in the antenna circuit[2] as described in a later section. Connections are made to the cavity by means of coaxial cables.

4.2. *Low-Noise I.F. Amplifier*

As previously stated the input stage of the i.f. amplifier must be designed to have a low noise factor. Several types of amplifiers can meet this requirement and the arrangement known as the "cascode" amplifier[6] is used in this case. It consists of two triode amplifier stages, the first

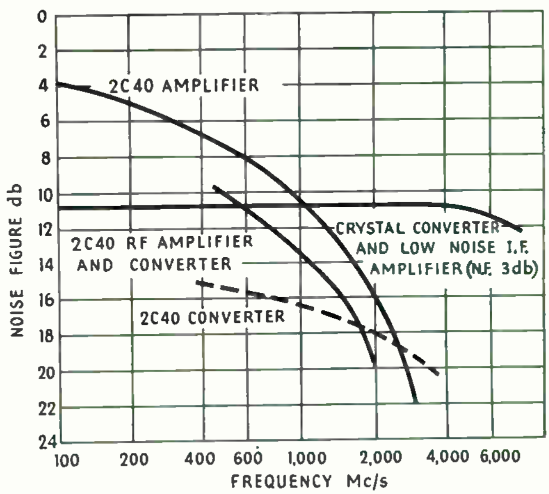


Fig. 10.—Graph showing noise figure of 2C40 lighthouse valve r.f. amplifier and mixer compared with crystal mixer.

being a neutralized grounded-cathode stage followed by a grounded-grid stage. The neutralized grounded-cathode stage can be operated without neutralization because the plate load of that stage is the input impedance of a grounded-grid stage—hence its gain is quite low. However, if the best noise factor is to be achieved, the stage must be neutralized. This combination is non-critical in its operation and combines the low noise factor of triode amplifiers with a gain that is characteristic of a pentode.

The block circuit diagram of the receiver is shown in Fig. 7 and the "cascode" arrangement shown as the 30-Mc/s pre-amplifier. Two type EC91 valves are used in the cascode circuit followed by a single 6AU6 pentode amplifier. The noise factor of the combination is 2½ to 3 db at 30 Mc/s with a gain of 35 db. The pre-amplifier unit is enclosed in a shield box (top unit Fig. 5), as the gains involved in the i.f. channel are quite high.

4.3. *Crystal Oscillator and Multiplier*

The crystal oscillator operates at a frequency higher than that of the i.f. amplifier, otherwise harmonics of the oscillator could fall in the i.f. passband and cause spurious beats. The crystal is operated on the third harmonic in the region of 35-40 Mc/s. This frequency is multiplied 12 times to give the final frequency at 30 Mc/s above or below the signal frequency as required. Inductively coupled circuits are used throughout the multiplier chain in order to minimize any feed-through of a side-band of the crystal frequency after the first multiplier valve and thus to reduce spurious responses in the receiver.

The output of this unit is sufficient to drive the 1N21B crystal to a current of 0·5–1 mA; the normal current is 0·5 mA. A block diagram of the unit is shown in Fig. 7. The whole unit is enclosed in a shielded box similar to that of the i.f. pre-amplifier.

4.4. *Intermediate Frequency Amplifier*

Two types of i.f. amplifier were considered, namely (*a*) synchronously tuned and (*b*) stagger tuned.

Synchronously tuned amplifiers are simple and easily aligned, but have the disadvantage of being inefficient owing to the fact that the overall bandwidth shrinks rapidly when stages are cascaded.[3] For *n* stages, the overall bandwidth is equal to the bandwidth of a single stage multiplied by the factor $\sqrt{(2^{1/n}-1)}$. The gain-megacycle bandwidth product is given by $g_m/2\pi C$ for a single amplifier stage so that in the case of *n* stages it becomes

$$\frac{g_m}{2\pi C}\sqrt{2^{1/n}-1}$$

For example, the bandwidth of six stages is only 0·35 times that of a single stage, which means, for a given overall bandwidth, the stage bandwidth must be made wider, hence the gain per stage is reduced and the number of stages increased to give the required gain.

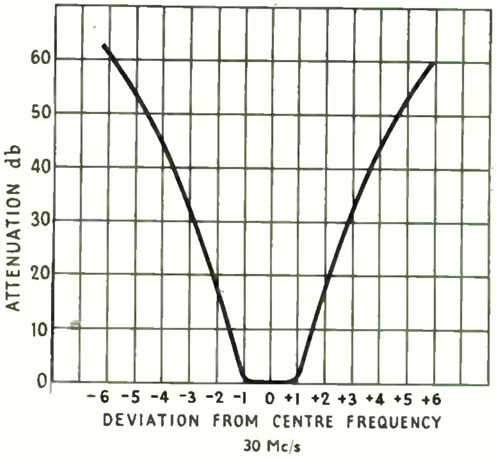


Fig. 11.—Selectivity curve for stagger-tuned i.f. amplifier.

In the stagger-tuned amplifier, the individual stages are resonant at different frequencies and have different bandwidths; the effect of this procedure is to reduce the overall bandwidth shrinkage when stages are cascaded. For these stages the gain-bandwidth product is the same for one stage or "*n*" stages; hence, the number of stages required to give a certain bandwidth is less than the number required to give the same bandwidth with synchronous tuning. Stagger tuning has disadvantages but it was felt that they were relatively unimportant.

The i.f. amplifier consists of four stages of amplification, followed by two limiters and a discriminator. The four stages are stagger tuned about a mean frequency, whereas the two limiter circuits are tuned to the mean frequency of 30 Mc/s.

The i.f. channel bandwidth is ± 1·2 Mc/s at the 3-db points (typical curve Fig. 11) and total gain approximately 100 db in the four stages, the gain-bandwidth product for a single stage being 22·7. A staggered quadruple is used in the amplifier, the single tuned circuits being adjusted to the following resonant frequencies[7]:—

$$f_1 = f_0 + 0{\cdot}46\ B,$$
$$f_2 = f_0 - 0{\cdot}46\ B,$$
$$f_3 = f_0 + 0{\cdot}92\ B,$$
$$f_4 = f_0 - 0{\cdot}92\ B,$$

where f_0 is the centre frequency, B is the overall bandwidth and $B/f_0 < 0{\cdot}3$.

The bandwidths of the above circuits are given by the following equations:—

$$B_1 = 0{\cdot}38\ (B/f_0) f_1,$$
$$B_2 = 0{\cdot}38\ (B/f_0) f_2,$$
$$B_3 = 0{\cdot}19\ (B/f_0) f_3,$$
$$B_4 = 0{\cdot}19\ (B/f_0) f_4.$$

From these equations, the frequencies and bandwidths can be calculated and the approximate values of damping resistances can be found from the following formula[3]:—

$$R_d = \frac{10^6}{\omega_0 C D},$$

where $\omega_0 = 2\pi f_0$
C = tuning capacitance (including strays),
and D = dissipation factor,
$= 1/Q$

Although the resistance calculated above will not be the exact value required, due to the input conductance of the valve, it will be a guide and will be lower than the actual value.

The main i.f. amplifier consists of four 6AU6 valves operated at 105 volts, regulated on both anode and screen, with the cathode connected directly to earth. An a.g.c. voltage is fed back from the first limiter grid to the grid of each stage via a filter network and, in the no-signal state, a bias due to the rectified noise voltage is applied in order to keep the tube within its ratings. A meter connected in the grid return of the first limiter grid indicates the received signal strength for signal levels between 10 microvolts and 10,000 microvolts.

A second limiter stage drives the Foster-Seeley discriminator[1] which has a peak separation of 2 Mc/s and a linear region of ± 600 kc/s; a typical response curve is shown in Fig. 12. Trouble was experienced in designing a discriminator to operate at 30 Mc/s owing to the effect of capacitance coupling between windings; hence the discriminator coil is not placed in a screen can as in the case of the i.f. coils since leads would then have to be connected to the base, giving rise to capacitance coupling. The winding is placed in a shielded compartment of the i.f. chassis enabling the leads to go direct to the valve sockets; also the primary was specially wound so as to preserve the symmetry of the primary and secondary windings.

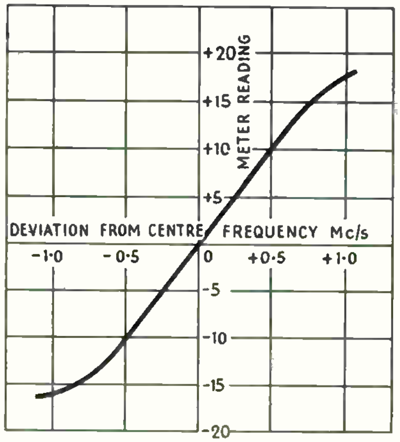


Fig. 12.—Discriminator characteristic of 30 Mc/s i.f. channel.

Included on the i.f. chassis is the first audio amplifier stage using a 6AU6 pentode. Connected between the discriminator output and first audio stage is the frequency-correcting network; a typical curve is shown in Fig. 1. A high-impedance shielded line connects the audio output from the i.f. chassis to the main audio amplifier.

4.5. *Audio Amplifier Unit*

This consists of two amplifiers, (*a*) the "traffic" amplifier and (*b*) the "supervisory" amplifier.

(*a*) The "traffic" amplifier is a high-quality amplifier capable of handling frequencies in the range of 3 kc/s to 70 kc/s. The r.m.s. distortion is less than 0·1 per cent. and the maximum power output is 100 mW into a 600-ohm line. A potentiometer in the input circuit is used as the coarse level control and a balanced 600-ohm attenuator calibrated in 1-db steps in the output line as a fine level control.

This amplifier has two stages of amplification, a 6AU6 pentode driving a 6AQ5 pentode, with a combination of negative voltage and current feedback over both stages. The proportion of current and voltage feedback is adjusted so that the "return loss" of the amplifier is better than 20 db over the frequency range.

(*b*) The supervisory or "order wire" amplifier has a single stage, using a 6AU6 pentode. A low-pass filter, with a cut-off at 3 kc/s is incorporated in the 600-ohm output line to attenuate frequencies in the range 3 kc/s to 70 kc/s and pass frequencies in the range 250 c/s to 3 kc/s. The power output is several milliwatts.

5. Relay Station

The function of the relay station is much the same as that of a landline repeater, i.e. to amplify the signal prior to repeating. The radio repeater must not only introduce gain, it must also shift the radio frequency to prevent the amplifier, which has a very high gain, from singing, since the receiving aerial cannot be screened from the transmitting aerial to the necessary degree.

The repeater station to be described uses the heterodyne principle in that the traffic circuit is received at one frequency and, after amplification, is heterodyned to a new frequency prior to re-radiating. This process is carried out without demodulating and remodulating the carrier, thus eliminating the accumulated distortion and cross-talk associated with this process.

The repeater receiver is identical with the terminal receiver and includes a demodulator and audio amplifier. A tap is taken from the 30-Mc/s i.f. amplifier prior to the demodulator and is fed to the repeater transmitter where it is heterodyned to the new carrier frequency. The demodulator is used to drop out the order-wire or supervisory tone at each repeater.

A simple phase modulator is included in the repeater transmitter and is used for introducing supervisory tones and order-wire facilities at each repeater.

It must be clearly understood that any modulation introduced at the terminal transmitter or repeater transmitter cannot be prevented from going through the system and at the remote receiver all modulating frequencies introduced throughout the chain will be present.

When the repeater is part of a trunk telephone system it must function in both directions. This necessitates having two receivers and two transmitters, one in each direction. As the repeater station is usually unattended it is often necessary

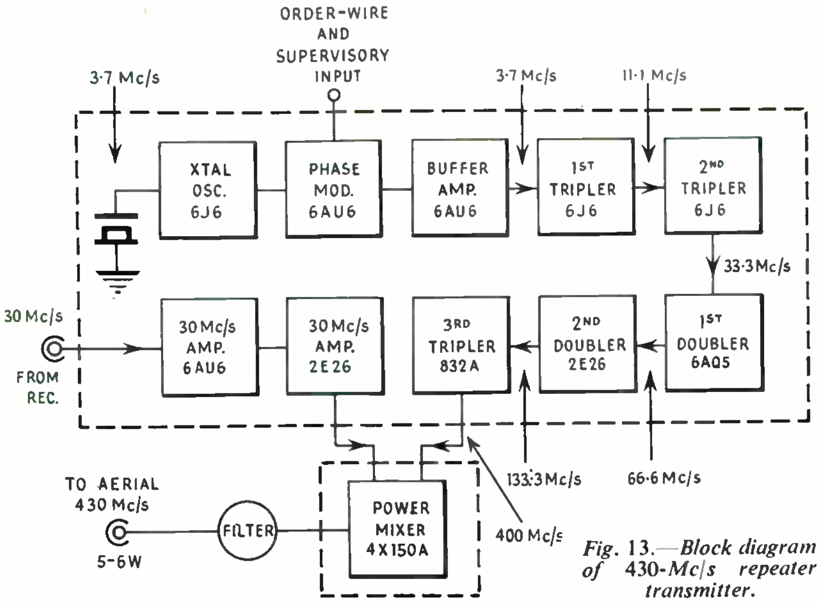


Fig. 13.—Block diagram of 430-Mc/s repeater transmitter.

to duplicate the equipment in both directions and to provide some form of automatic change-over.

5.1. 430-*Mc/s Repeater*

The block diagram of Fig. 13 shows the general arrangement of a one-way repeater for the 430-Mc/s system. Fig. 14 illustrates the rear view of the repeater transmitter together with its modulator for order-wire and supervisory purposes. The repeater modulator contains a

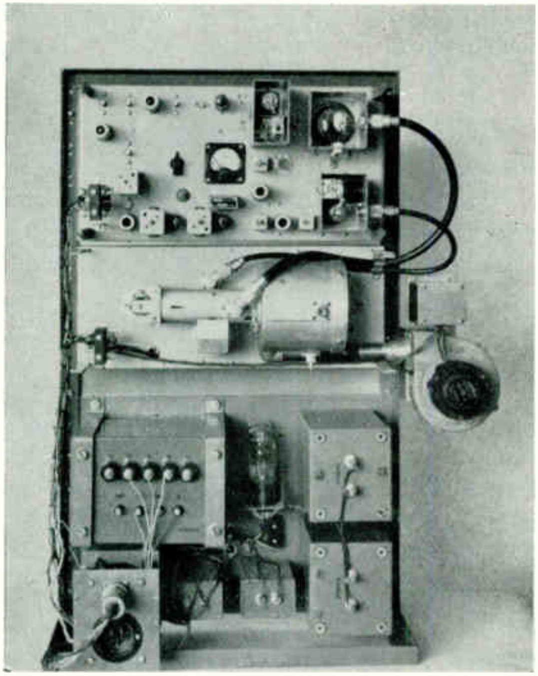


Fig. 14.—*Rear-of-rack view of* 430-*Mc/s repeater transmitter.*

crystal oscillator and modified Van Roberts phase modulator together with a chain of frequency multipliers to produce an output frequency in the vicinity of 400 Mc/s with an output of approximately 3 W. All circuits throughout the multiplier chain employ band-pass transformers; push-pull triplers are used in the early stages to eliminate spurious signals. A tap is taken from the i.f. stage of the receiver and is fed via a 70-ohm line to a two-stage limiting amplifier where the level is raised to about 3 W.

The two signals, one at 30 Mc/s bearing the original frequency-modulated signal and the other at 400 Mc/s bearing the modulation introduced at the repeater, are fed via coaxial cables to the power mixer panel.

5.2. *Frequency Converter Stage*

A 4X150A tetrode is used as a power mixer to shift the outgoing carrier to a new frequency, slightly different from that of the incoming carrier. The new carrier is radiated at a power of approximately 5 W.

The grid circuit is a half-wave coaxial line, tuned to resonance by means of a small butterfly-type capacitor at one end of the line; the other end of the line is terminated by the grid pin of the valve. Bias is fed to the grid line via a small r.f. choke in the form of a compression spring which contacts the inner conductor near the quarter-wave junction. The other end of the choke is by-passed to earth by a small mica capacitor.

The 30-Mc/s signal is inductively coupled to a variable inductance, connected to the junction of the r.f. choke and r.f. by-pass capacitor. At 30 Mc/s, the valve input capacitance, grid-line capacitance and r.f. by-pass are all effectively lumped together and form a circuit with the inductance which is resonant at 30 Mc/s.

The 400-Mc/s signal is fed to the grid circuit of the mixer by means of a small loop inductively coupled to the first quarter-wave section of the line.

The mixer valve is driven into grid current by both the 30-Mc/s and 400-Mc/s signal source. Owing to the non-linear action of the grid, a number of frequencies are produced in the plate circuit, the predominating frequencies being 370 Mc/s, 400 Mc/s and 430 Mc/s.

The anode circuit of the mixer is identical with that in the terminal transmitter and in this arrangement is tuned to the sum frequency of the 30-Mc/s and 400-Mc/s signals; an output of approximately 6 W is obtained at 430 Mc/s. Considerable attenuation of the unwanted frequencies is provided by the anode circuit of the mixer and further attenuation is provided by a cavity filter between the anode circuit and aerial; the total attenuation of unwanted frequencies is better than 60 db.

The 4X150A is operated as a power mixer under the following conditions:

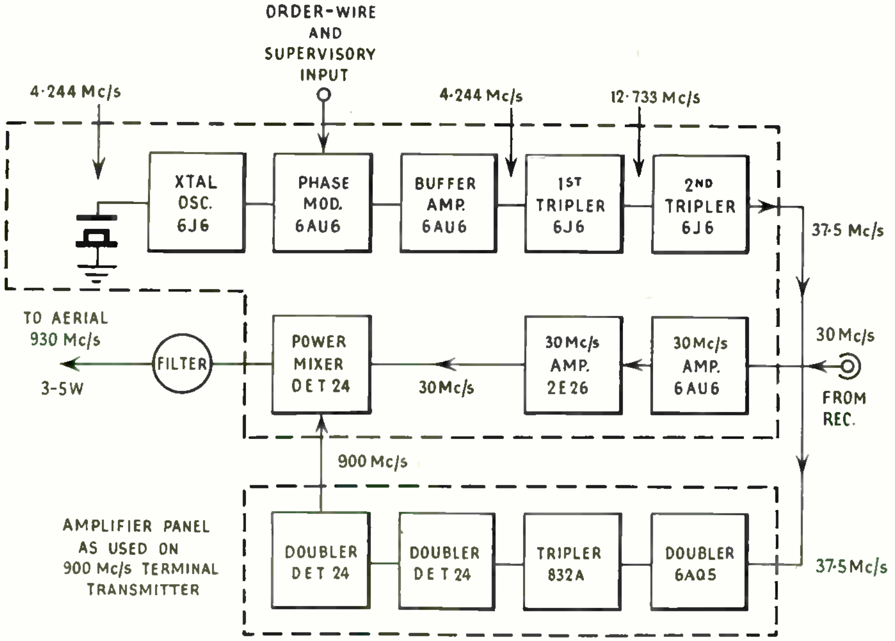


Fig. 15.—Block diagram of 930-Mc/s repeater transmitter.

E_a = 300 V } anode input = 16·8 W
I_a = 56 mA }
E_{sg} = 225 V
I_{sg} = 4·5 mA
I_g = 7 mA
Bias = −45 V
Output = 5·6 W
Efficiency = 33 per cent.

5.3. 930-*Mc/s Repeater*

As can be seen from the block circuit diagram, Fig. 15, the 930-Mc/s repeater makes use of the heterodyne principle as used in the 430-Mc/s repeater. Fig. 16 illustrates the rear view of the power mixer and modulator panel.

The output from the modulator chain following the order-wire modulator is fed to a separate multiplier panel where the signal is multiplied a further 24 times to produce a signal source at or near 900 Mc/s. This panel is the same as for the multiplier chain and output circuit used in the 900-Mc/s terminal transmitter. The output from this panel is fed back to the repeater transmitter panel where it is combined with the output of the two-stage 30-Mc/s amplifier to produce a new frequency of 930 Mc/s.

The 930-Mc/s repeater transmitter employs a DET.24 grounded-grid triode as the frequency converter. The construction of grid-cathode line and grid-anode cavity is identical with that of the 450–900-Mc/s stage in the terminal transmitter and is tuned in a similar manner. As the grid-cathode line is now tuned to 900 Mc/s, the non-shorting plunger is adjusted to the three-quarter wave-length position as the first quarter-wave-length position is too close to the valve for convenient coupling.

The 30-Mc/s source from the receiver i.f. is fed to the mixer input via a bi-filar winding in the heater-cathode leads. The double winding is joined together for r.f. by a capacitor at each end to form an inductor which resonates at 30 Mc/s with the valve and line capacitance; the 30-Mc/s source is link coupled to the inductor.

In this way the two signals are independently fed to the grid-cathode line where they are mixed and the sum or difference frequency is selected in the grid-anode cavity.

The DET.24 is operated as a power mixer at 870–930 Mc/s under the following conditions.

E_a	= 300 V
I_a	= 48 mA
I_k	= 60 mA
Input	= 14·5 W
Output	= 5·2 W
Efficiency	= 37 per cent.

6. Aerial Line Filters

The amount of frequency pre-selection required in the receiver input circuit will be governed by the location of the receiver, and the number of systems and general interference in the area.

The only pre-selection in the receiver is that provided by the mixer cavity. Additional selectivity is readily obtained by using a one-, two- or three-section pre-tuned cavity filter prior to the mixer.

Similarly in the terminal and repeater transmitter output circuit, spurious and harmonic suppression is obtained by using high Q cavities in the output anode circuit. Additional suppression is again obtained by using a one-, two- or three-section pre-tuned cavity filter in the transmitter output circuit.

By using a high Q cavity filter with a ratio of unloaded to loaded Q greater than 17, the insertion loss per filter will be less than 0·5 db. When greater attenuation than that provided by a single cavity is required, it is possible to couple two or more filters together, using quarter-wave couplings (Fano-Lawson type) to make a band-pass filter.

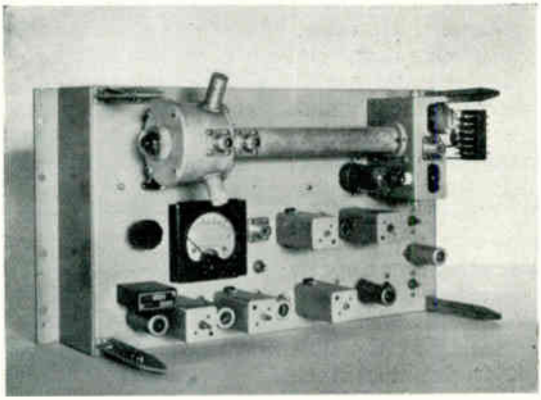


Fig. 16.—Rear-of-rack view of power mixer and modulator panel.

A two-section quarter-wave coupled filter for 450 Mc/s is illustrated in Fig. 17, together with a single-cavity filter for 900 Mc/s; a 4X150A valve is also illustrated.

Fig. 17.—Two-section 450-Mc/s quarter-wave coupled filter; on the right-hand side is a single cavity filter for 900 Mc/s, in the foreground a 4X150A valve.

The two-section cavity filter, tuned to 450 Mc/s, has a bandwidth of ± 1 Mc/s at the 3-db points and is approximately ± 10 Mc/s at 40 db down. The loaded Q is 160 per cavity as measured, with a calculated unloaded Q of approximately 11,000 and the insertion loss for the two cavities is approximately 0·25 db.

A three-section quarter-wave coupled filter with a loaded Q of 160 was computed to have an attenuation of 60 db at ± 8·5 Mc/s from 450 Mc/s with a bandwidth of ± 1 Mc/s at the 3-db points.

7. Order-Wire, Pilot Tone and Fault-locating Equipment

7.1. *Order-Wire System*

The object of the order-wire facility is to provide a telephone channel between all stations for use by maintenance and operational personnel. The channel operates on the party line basis in the range 250 c/s to 2·8 kc/s with a two-wire extension to a remote magneto-type phone. Provision is made for transmission and reception in both directions at repeater stations.

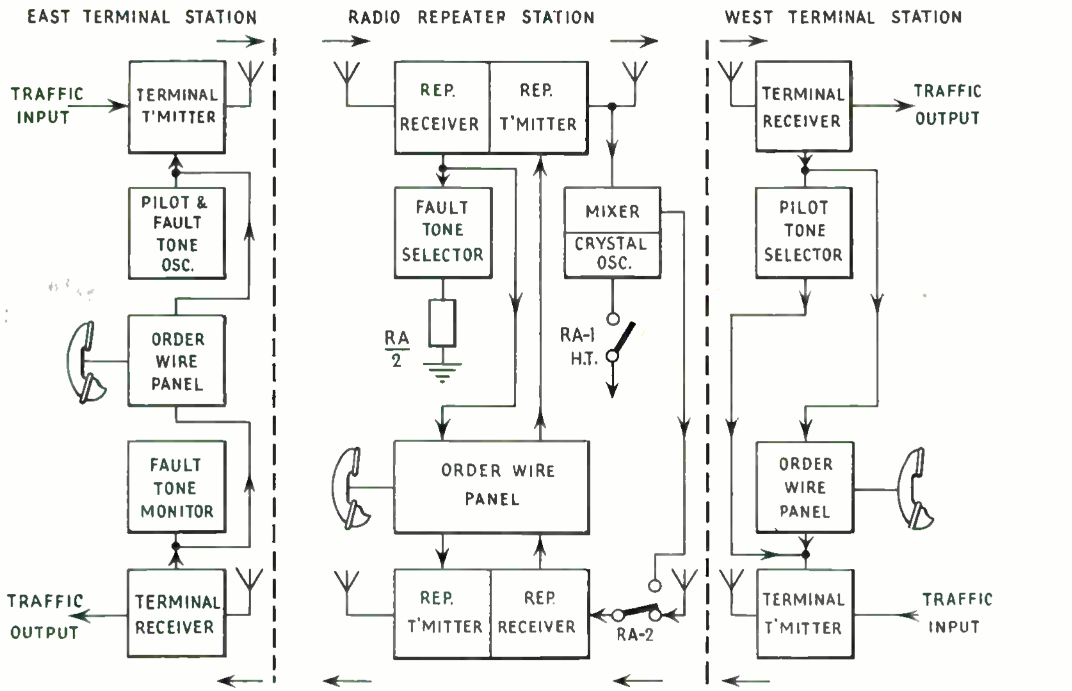


Fig. 18.—Block diagram showing the order-wire and supervisory facilities at a typical terminal and repeater station.

Ringing oscillators are installed at each station for calling purposes when used with the local phone.

7.2. *Pilot Tone and Alarm Circuits*

The block diagram of Fig. 18 shows the order-wire and supervisory facilities employed at typical terminal and repeater stations. The method adopted of indicating and locating a failure of any section of the equipment throughout the chain will now be described. By means of a circulating pilot tone in the frequency range below 4 kc/s the continuity of all traffic circuits within the link system may be continuously checked. The pilot tone is fed into the terminal transmitter at one end of the system. At the opposite end of the link the tone is selected and fed into its associated transmitter. The tone is finally selected at the end of the system which originated the tone, then applied to an alarm panel via a rectifier. A failure in any part of the system will sound an alarm. When the system is duplicated throughout the chain, a failure in any section of the system could cause a changeover of the traffic connections to the stand-by equipment with a minimum interruption to the service.

7.3. *Fault-locating Equipment*

When the system contains only one repeater in the chain, it is relatively simple to determine which part of the system has failed, and it is normally unnecessary to add fault-locating equipment. However, when there are a number of repeaters in the chain it is necessary to know which repeater has failed, assuming the failure is not at the terminal. By using a fault-tone generator, capable of generating up to, say, 5 tones in the frequency range below the pilot tone, means may be provided for locating a fault in the system. Each repeater station may have a unit responsive to one of these tones, and when this tone appears in the receiver output it is selected by this unit which causes a relay to close. This starts an auxiliary crystal oscillator, operating at the difference frequency between the transmitter in the E-W repeater and the receiver in the W-E repeater. The output from the oscillator together with some output from the E-W transmitter is fed to a crystal mixer in a cavity tuned to the frequency of the W-E receiver. At the same time a relay changes the W-E receiver input from its antenna to the cavity output; this completes a loop and returns the fault tone to the control terminal, where it is identified in the fault tone monitor. Thus, signals may be looped through successive repeater stations until the faulty station is located, by sending each of the fault tones in turn. When a long haul is envisaged, it is advisable, for maintenance reasons, to break the system up into groups of say five repeaters per section, and to use a terminal repeater at these junctions; each terminal repeater would then have its own fault tone generating equipment.

8. Performance using 12-Channel Carrier Equipment

The terminal transmitter, operating at 450 Mc/s, was set up in the Grace Building, at the corner of York Street and King Street, Sydney. The radio repeater, receiving on 450 Mc/s and transmitting on 430 Mc/s, was installed at Broadcasting Station 2 KA Wentworth Falls, with both receiving and transmitting aerials directed towards Sydney. The terminal receiver was installed at Grace Building alongside the terminal transmitter. This arrangement provided an overall test loop of the three major units of the system over a total path length of 100 miles. The experimental aerial used in conjunction with the equipment was a relatively simple type of directive array with a gain of approximately 15 db.

The received signal at Wentworth Falls was approximately 300 mV and the received signal at Grace Building varied between 200 to 300 mV over a period of three weeks.

Prior to these tests, the equipment was set up at the A.W.A. Laboratory, Ashfield, with the repeater at Wentworth Falls. Over a period of several months the signal variation was less than 6 db.

8.1. *Harmonic Distortion and Noise Tests*

Before attaching the radio bearer circuit to the carrier equipment, harmonic distortion and noise tests were carried out by using three band-pass filters, centred at 4, 8 and 12 kc/s, each with a band-pass of 3 kc/s. The transmitter was fully modulated by a 4-kc/s tone, fed from a low-distortion oscillator.

The receiver output was fed to the three filters with their inputs in parallel and the second and third harmonic selected by the 8- and 12-kc/s filters. The results obtained are as follows:—

1. Second harmonic: better than —60 db.

   Third harmonic: better than —60 db.

   Noise only (no input) in each channel about —65 db unweighted.

2. With the signal input to the transmitter reduced to —12 db on 100 per cent. modulation and the received output adjusted to a level of —1 dbm, the following results were obtained:—

   Noise only: —53 db unweighted.

   Distortion plus noise: —53 db unweighted.

   This indicated that at a line-up level of —12 db per channel, as would be the case for a 12-channel carrier system, the unweighted noise is greater than the harmonic distortion.

8.2. *Power-handling Capacity*

When a radio-bearer circuit is used in conjunction with multichannel carrier equipment, the line-up level per channel, in relation to full deviation, will depend upon the number of channels in use.

The line-up level per channel also has considerable bearing on cross-talk and signal-to-noise ratio on the receiver side of the carrier equipment. If the line-up level is based on the transmission of continuous sinusoidal signals simultaneously on all channels, i.e. based on total peak deviation, divided by the square root of the number of channels, the line-up level shown in Table 2 would be used.

TABLE 2

Channel Loading

2	channels	3 db	down	on	single	channel
4	,,	6 db	,,	,,	,,	,,
8	,,	9 db	,,	,,	,,	,,
12	,,	11 db	,,	,,	,,	,,
16	,,	12 db	,,	,,	,,	,,
24	,,	14 db	,,	,,	,,	,,

With the above line-up level as a basis it is customary to insist that each channel is checked at 8 db above the line-up level to allow for an overload.

Additional tests over a path other than line-of-sight have been conducted over a period of nine months with satisfactory results. The terminal transmitter was located at Liverpool with the repeater at Sutherland and the terminal receiver at Mascot; the circuit has been carrying traffic continually over this period.

This path between Liverpool and Sutherland has four obstacles in the form of hills, each of at least 50 ft above direct air line; the path length is about 10 miles over this section of the link.

Signal fading in the order of 6-8 db has been observed at Sutherland, but owing to the action of the limiters in the receiver, the variation in the output from each channel is consistently less than 1 db.

9. Brief Performance Specification

Using 12-Channel Standard Carrier Systems

Line-up level per channel: −12 db on full deviation.

Frequency response: within 1 db from 4 kc/s to 60 kc/s.

Power-handling capacity: linear from −12 db to +10 db on full deviation.

Return loss input circuit: better than 20 db.

Return loss output circuit: better than 20 db.

Psophometric noise: 1 mV or less on all channels.

Crosstalk plus noise: 2 mV or less on all channels.

Noise unweighted: −53 db, as referred to switchboard level of −6 dbm.

Gain stability: variation in channel level less than 1 db recorded over a period of 72 hours.

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