ELECTRONIC ENGINEERING

VOL. 30 No. 363

MAY 1958

Microwave Communications

THE wideband multi-channel microwave radio link is a branch of communications that has expanded enormously of late years and it owes much of its development to the radar techniques originated during the war.

It has now become one of the accepted methods of carrying from point to point a large number of separate telephone channels or, alternatively, a wide band television signal.

The multi-channel radio link itself is no new thing for much development work was carried out before the war in the v.h.f. band and valuable contributions to the techniques were made on both sides of the Atlantic. Interest in the band of frequencies above 300Mc/s for commercial purposes began to grow in 1930 onwards and one of the first microwave links was set up on an experimental basis in 1931. This was the microwave link across the English channel between Dover and Calais. Operating at what was then regarded as an extremely high frequency of 1 700Mc/s with a radiated power of about half a watt it was regarded as an enormous advance on the techniques of the day and it demonstrated that the little used band of frequencies from 300 to 3 000Mc/s the Ultra High Frequency Band was ripe for exploitation.

There followed in 1934 what was probably the world's first microwave link operating on a regular commercial basis between the airports of Lympne in Kent and St. Inglevert in France, a distance of some 35 miles.

A halt to further progress was caused by the outbreak of war but the many radar techniques of the war period and the advent of special types of valves such as the klystron and the travelling wave tube, which were developed for radar purposes, were to make a great contribution when hostilities ceased.

One of the first microwave links to take advantage of the new techniques was installed in 1949 between London and Birmingham. This link operated at a frequency of 900Mc/s and was designed to carry a single channel television link from London to the BBC's new television transmitter at Sutton Coldfield. It was followed in 1952 by another microwave link on 4 000Mc/s from Manchester to Edinburgh to convey the BBC's television signals to Kirk o'Shotts.

Since then, similar systems have been installed in Europe and it is interesting to recall that in 1954 when the Coronation ceremonies were televised, an extensive network of microwave systems had been brought into being to link the countries of Europe for this occasion.

Investigations had also shown that the microwave link offered distinct possibilities for use in multi-channel telephony and in 1946 the British Post Office set up its own experimental link between London and Castleton, near Cardiff. It was used in 1949 to convey the BBC's television signal to Wenvoe and, although operating at a frequency of 200Mc/s, it was one of the first links to use frequency modulation for the transmission of television. Much valuable information was gained on the problems involved in the use of this form of modulation and the Post Office followed this installation by a single hop frequency modulated microwave link across the Moray Firth, carrying 240 separate telephone channels and designed to meet C.C.I.T.T. standards as far as conditions would permit.

The foregoing is a very brief account of the history and development of post war microwave communications but however incomplete this may be, reference should be made to enhanced possibilities that are suggested by tropospheric propagation.

It has been evident for a number of years that the propagation of microwaves could extend for distances considerably beyond the horizon and Marconi himself was one of the early pioneers in this respect.

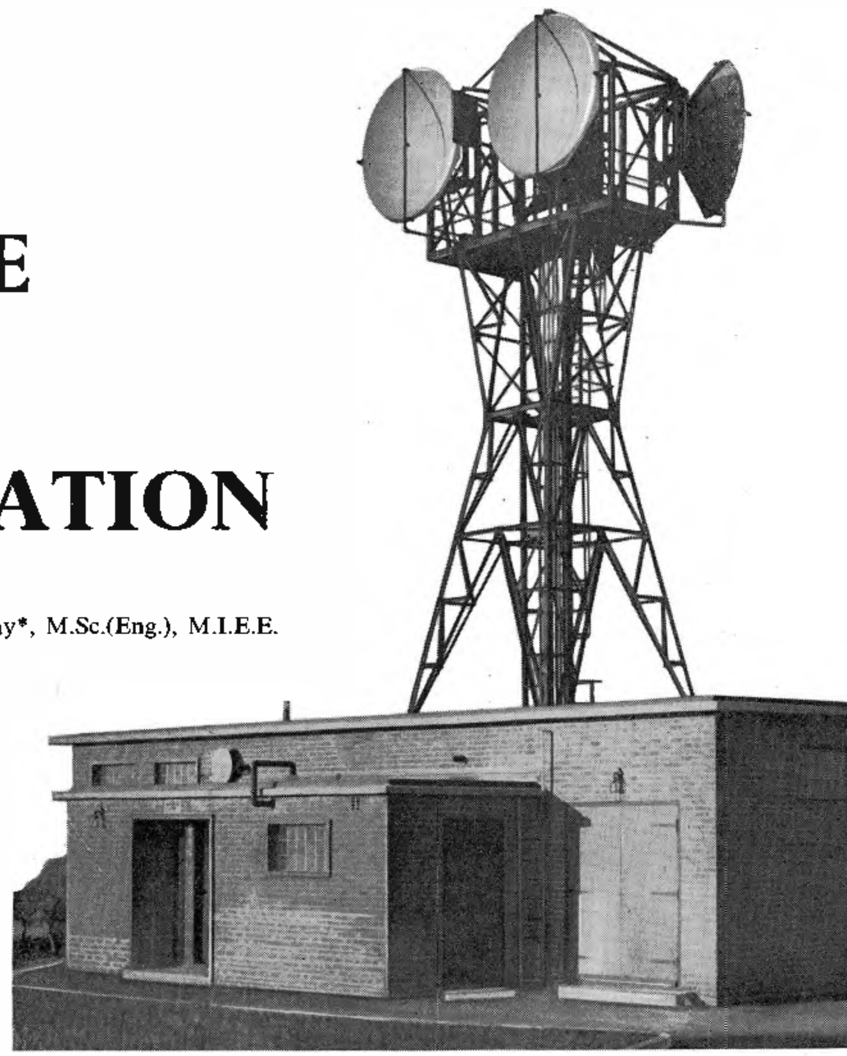
Several theories have been put forward to account for this new phenomenon of tropospheric scatter propagation but the mechanism is not yet fully understood and much investigation remains to be carried out. It does, however, appear that long distance microwave propagation, including the bridging of the Atlantic, is within the bounds of possibility.

The world wide interest in microwave transmission and tropospheric propagation has grown to the extent that it was felt that the time is appropriate to present a survey of current British practice in this field and this month's issue has been given over to this purpose.

It deals principally with propagation, both line of sight and tropospheric, microwave systems and their essential components, such as valves. Radar and navigational devices have been excluded.

The amount of reading matter included in this issue is probably more than can normally be digested in a month, but the subject to be covered is an extensive one and care has been taken to make the survey as complete and as representative of British practice as is possible.

A Survey of MICROWAVE RADIO COMMUNICATION



By W. J. Bray*, M.Sc.(Eng.), M.I.E.E.

This article is a broad survey of line-of-sight and tropospheric forward-scatter radio communication systems operating in the frequency range from about 1 000 to 10 000Mc/s, for the transmission of multi-channel telephony and television signals. The applications of such systems, the transmission requirements, the factors determining the choice of frequency, and the basic equipment techniques are described. Possible future developments are also considered.

THIS article is a broad survey of microwave radio communication systems existing in the United Kingdom and elsewhere, and of current developments in this field. Most of the systems discussed are intended to operate on frequencies in the range 1 000 to 10 000Mc/s. They may be divided into two main classes:

(a) Low-power systems with line-of-sight paths between adjacent stations, the paths being some 10 to 50 miles in length[1-4].

(b) High-power systems with paths extending well beyond the visible horizon up to 200 miles or more long; these systems include 'tropospheric-scatter' systems[5-7].

Microwave communication systems with line-of-sight paths are rapidly assuming a position of considerable importance in the field of telecommunications and in many respects such systems are competitors to wire-line and coaxial-cable systems. A wide range of applications is possible, extending for example from systems providing a small number of telephone circuits to systems providing several thousands of telephone circuits or several television channels over distances of a thousand miles or more with the high standards of performance and reliability essential for national and international trunk circuits.

The increasing use of 'line-of-sight' microwave systems may be attributed to the following advantages:

(1) In some applications the capital cost and annual charges are less than for line or coaxial-cable systems of comparable capacity and performance.

(2) Microwave systems can often be installed more rapidly than line systems.

(3) Additional capacity can be provided quickly once a chain of relay stations has been established.

(4) Microwave systems are well adapted to the provision of communications in difficult terrain, particularly in sparsely populated, mountainous or densely wooded areas, where line systems may be impracticable or unduly costly.

There are however certain disadvantages which must be recognized and taken into account in the planning of such systems, viz.:

(1) The siting requirements of the stations are often critical and there may be difficulties in obtaining suitable sites due, for example, to objections from local planning authorities or unwillingness of land owners to sell.

(2) Mains power supplies or access roads may not be available, or available only with difficulty, at some sites.

(3) Adverse propagation conditions may at times cause a deterioration of performance due to fading; while this deterioration can be minimized by suitable equipment design and choice of paths, it cannot be completely eliminated.

The above discussion applies primarily to low-power microwave systems with line-of-sight paths; high power tropospheric-scatter systems are a comparatively recent development and have as yet been used only on a relatively limited scale; these systems are discussed later in the present article (see also Millington and Grisdale page 248 and 275 this issue).

* *Post Office Engineering Department.*

Applications of Microwave Systems

The main fields of application for microwave communication systems are as follows:

(a) "Integrated" fixed systems for multi-channel telephony or television, forming part of national or international networks.

(b) "Non-integrated" fixed systems for multi-channel telephony or television, not normally, or only occasionally, interconnected with the national or international networks.

(c) "Transportable" systems, generally for outside broadcast television purposes, which may be interconnected at times with the national or international networks.

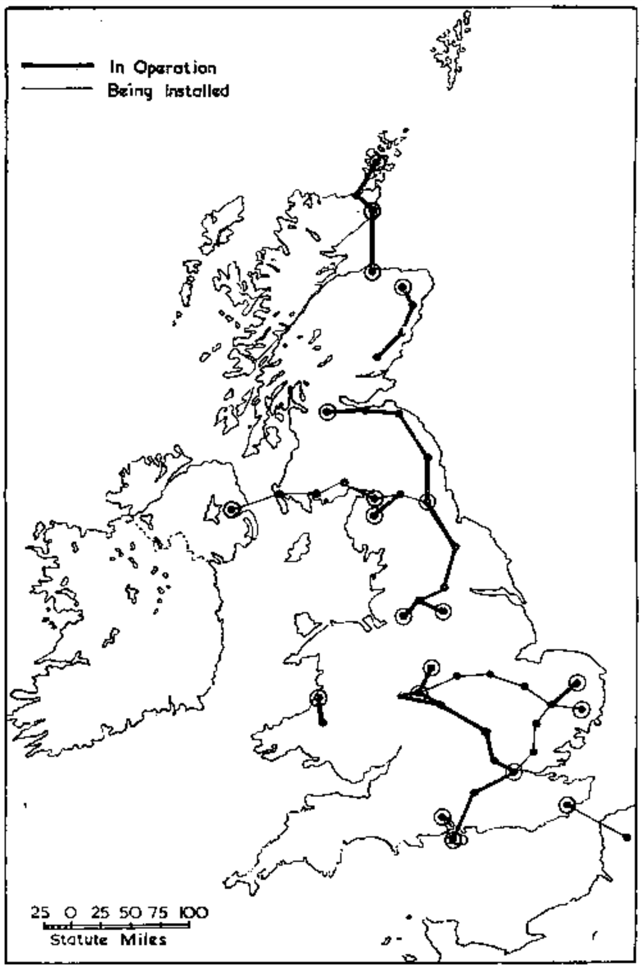


Fig. 1. Post Office microwave links in operation and being installed in the United Kingdom (Feb. 1958)

(a) Integrated Microwave Systems

Typical "integrated" microwave systems in operation or being installed in the United Kingdom are shown in Fig. 1; the buildings and aerials of a typical repeater station are shown on page 226 and the equipment is shown in Fig. 2.

In the United States of America, some three quarters of the total mileage of long-distance television channels and about a quarter of the total trunk telephone circuit-mileage owned and operated by the American Telephone and Telegraph Co. are now provided by microwave systems, and the proportion of trunk telephone circuit mileage provided by microwave is expected to approach a half the total in ten years' time. These systems are mainly of the 'low-power line-of-sight' type, although some tropospheric-scatter systems are beginning to make their appearance.

In addition to the strictly 'microwave' systems using frequencies above about 1 000Mc/s, use is also made in the United Kingdom and elsewhere of systems operating at frequencies around 150 and 450Mc/s for the provision of blocks of up to 24 or 48 telephone channels. The main field of application of the latter systems in the United Kingdom has been for the provision of communications between the mainland and off-shore islands; in overseas countries extensive inland networks have also been set up with this type of system.

(b) Non-integrated Fixed Microwave Systems

Non-integrated microwave systems are used in some countries for such purposes as the communications networks of the fire, police, forestry and highway maintenance services, oil pipe line and electricity power line communications, railway and highway communications and for the broadcast auxiliary services (e.g. studio to transmitter links).

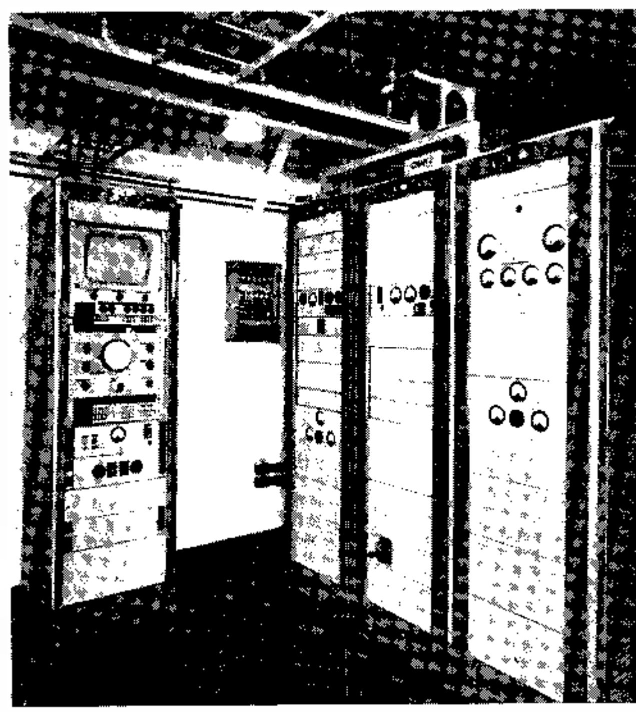


Fig. 2. Typical microwave radio-relay station equipment

(c) Transportable Microwave Systems

The transportable microwave systems referred to here are essentially point-to-point systems set up and operated for a few hours or a few days in one location and then moved to a new location. Typical examples are the microwave equipments used by the British Broadcasting Corporation and the Programme Contractors of the Independent Television Authority in this country for relaying television and in some cases sound signals from the scene of an outside broadcast to a point on the Post Office network for onward transmission to the television control centre of the broadcasting organization[8,9] (see Bridgewater, page 291, this issue).

Choice of Frequency: The Frequency Bands For Microwave Systems

Microwave systems for television or some hundreds of telephony channels occupy several megacycles per second of frequency space for each direction of transmission, and if several broadband channels are to be provided on the same route for both directions of transmission, hundreds of megacycles per second of frequency space are required. It follows that sufficient space can only be found if frequencies above about 1 000Mc/s are used.

Electromagnetic waves with frequencies above about 1 000Mc/s are propagated more or less freely over line-of-sight paths but above about 10 000Mc/s. appreciable absorption is introduced at times by rain, fog and snow; at still higher frequencies, above 20 000Mc/s, absorption by water vapour and oxygen in the atmosphere becomes significant[10].

Low-power microwave systems must use sensibly line-of-sight paths between the transmitting and distant receiving aerials in order to avoid excessive loss of signal strength. Obstructions such as the surface of the earth, buildings, trees etc. should preferably be removed from the direct transmission path by at least 'first Fresnel zone' clearance even under the most unfavourable condition of atmospheric refraction; this implies that the path from one aerial to the other via any obstruction should be at least one half-wavelength longer than that via the direct path. There is an additional requirement that appreciable reflected components should not be received from outside the Fresnel zone since the resulting multi-path transmission may impair transmission quality.

As the frequency is increased both the gain and the directivity of microwave aerials of a given size increase, the beamwidth correspondingly decreases; furthermore, the width of the Fresnel zone also decreases so that the necessary obstacle clearance is more readily obtained at the higher frequencies. These advantages are, however, offset to some extent by the tendency for fading to be greater at the higher frequencies.

Summing up, it may be said that the range from about 1 000 to 10 000Mc/s is particularly suitable for microwave line-of-sight systems, but frequencies up to 15 000Mc/s can be used effectively under suitable conditions.

The frequency bands allocated in the United Kingdom for microwave systems conform with the agreement reached by the member countries of the International Telecommunications Union at Atlantic City in 1947. For example, frequencies around 2 000, 4 000 and 6 500Mc/s are available for integrated systems.

As already mentioned frequencies around 150 and 450Mc/s are also used for the smaller capacity systems, sometimes on a shared basis with mobile services where adequate geographical separation is possible. These lower frequencies have the advantages that longer paths, and some obstruction of the paths, can be tolerated and the equipment is often cheaper.

Although the less favourable propagation characteristics of frequencies above 10 000Mc/s restrict the range to a certain degree, there are compensatory advantages, for example the equipment is more compact and there is at present less demand on the available frequency space.

Transmission Requirements

In general the telephony or television channels provided by integrated microwave systems may, at times, form part of international connexions several thousands of miles long. It follows that the transmission characteristics in respect of noise, gain stability, bandwidth and amplitude linearity, and in the case of television channels, waveform distortion, must be such as to permit satisfactory transmission over these long distances.

Two international bodies, both forming part of the International Telecommunication Union (I.T.U.), are responsible for the preparation of recommendations facilitating the setting up of long international circuits, viz. the International Radio Consultative Committee (C.C.I.R.), and the International Telegraph and Telephone Consultative Committee (C.C.I.T.T.). The C.C.I.R. through its Study Group IX, is responsible for the preparation of recommendations relating to microwave systems[11]; the C.C.I.T.T., through its Study Group III, is concerned with the introduction of such systems into the general line telecommunication network. The C.C.I.R. and the C.C.I.T.T. are thus both concerned in the formulation of the transmission requirements of international microwave systems.

The recommendations of the C.C.I.R. in respect of the transmission requirements of microwave systems are in close conformity with the corresponding recommendations of the C.C.I.T.T. for line systems, but take into account the special characteristics of microwave systems, e.g. the higher noise levels that may occur for a very small percentage of the time due to fading.

Noise Requirements for Telephony and Television

In a microwave system the noise requirements are of particular interest since these determine the transmitter power, aerial size and other parameters of the system. One important source of noise is that due to valves and mixers in the early stages of receivers and repeaters, since external radio noise is generally negligible; this noise is a maximum when the fading is deepest. In a telephony system additional noise may arise due to intermodulation between the signals in the various telephony channels; this noise is a maximum when most channels are active, i.e. during the busy hours. In general the noise in a microwave system is fairly steady at a low value for most of the time but there are occasional surges of higher level due to fades; by suitable system design and planning these surges can be confined to a negligibly small percentage of the time.

The permissible noise levels are defined internationally in terms of 'hypothetical reference circuits' 2 500km (1 600 miles) long with a specified number of modulation and demodulation processes (Rec. No. 200, 201, 202 and 203 Ref. (11) and Doc. 12, Ref. (12)); if these levels are expressed in terms of a typical microwave system 250km (160 miles) long the resulting signal-to-noise ratios are as follows:

Telephony channel (300c/s to 3·4kc/s); the ratio of the level of a test-tone (1mW) at a point of zero relative level to the mean value of psophometric noise in any hour should exceed 63dB. (The higher levels of noise permissible for very short periods of time have not yet been defined.)

Television channel (450 line system, 50c/s to 3Mc/s or 625 line system, 50c/s to 5Mc/s): the ratio of the peak-to-peak signal to r.m.s. noise should not be less than 40dB for more than 0·1 per cent of the time (linear noise distribution, i.e. rising with frequency by 6dB per octave, in the video band).

The foregoing relates to the transmission requirements in respect of noise that are desirable for integrated microwave systems forming part of long international connexions; in the case of other systems, not integrated with the general network, appreciably higher levels of noise are permissible and still enable satisfactory service to be provided.

Waveform Transmission Requirements for Television

The waveform transmission requirements are of great importance for television signals and it is now customary to specify the requirements in terms of a rating factor K, defining the permissible distortion of a number of test signals including the following:

(1) A pulse of sine-squared envelope and half-amplitude duration 0·17μsec (405 line system) or 0·10μsec (625 line system).

(2) A 40μsec bar test signal.

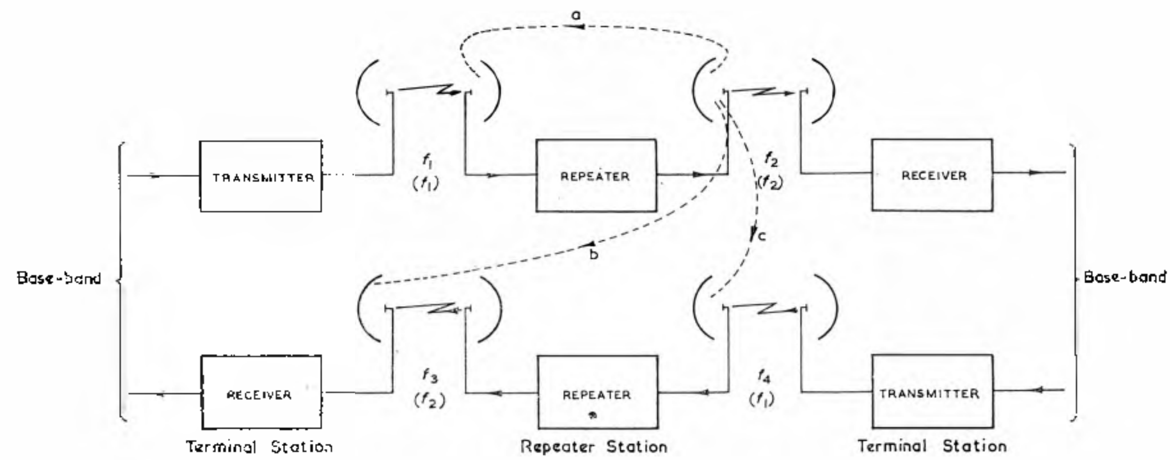


Fig. 3. Principles of radio-relay system

(3) A 50c/s square-wave test signal.

The detailed determination of the K-rating from the received signal waveform is described elsewhere[12,13]; as a guide it may be said that a K-rating factor of 1 per cent, which is considered permissible for a 250km (160 mile) link, would correspond to a long-delay echo of 1 per cent observed with test signal (1), a tilt of 2 per cent on test signal (2) or a tilt of 4 per cent on test signal (3).

Space does not permit a complete statement of all the transmission requirements for telephony and television; reference should be made elsewhere[11,12] for more detailed information.

Line-of-Sight Microwave Systems

The basic elements of a multi-section radio-relay system using line-of-sight paths are shown in Fig. 3; such a system comprises two terminal stations and a chain of repeater stations spaced at intervals of some 20 to 50 miles. At the terminal stations a microwave carrier is generated, modulated by the multi-channel telephony or television signal to be transmitted, and after amplification to a power of a few watts, is applied to a directional aerial for radiation to the first intermediate or repeater station. At the repeater station the microwave signal is received on a directional aerial, is amplified and shifted in frequency to avoid interference between the incoming and outgoing signals and is then re-radiated to the second repeater station and so on. The microwave signal at the distant terminal station is, after amplification, demodulated and the multi-channel telephony signal recovered.

In the system shown in Fig. 3, four frequencies f_1, f_2, f_3 and f_4 are used to avoid crosstalk over the dotted paths a, b and c. With adequate aerial directivity however it is possible to reduce the number of frequencies required to two, i.e. to f_1 and f_2, as shown in brackets in the Figure.

Multiplexing Techniques for Telephony

With a few exceptions general practice in integrated microwave systems is to use frequency-division multiplexing (f.d.m.) when a number of telephony channels are to be transmitted on the same carrier. The f.d.m. signal comprises a number of telephony channels arranged side-by-side in the base-band* as shown in Fig. 4; each telephony channel transmits audio frequencies from 300c/s to 3·4kc/s, the base-band signals being in the form of single-sideband signals with suppressed carriers at intervals of 4kc/s. The

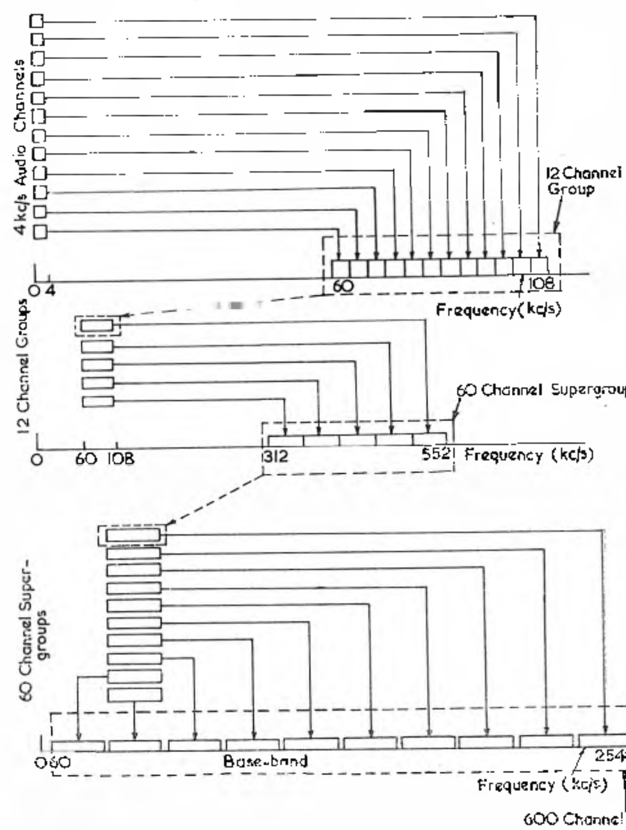


Fig. 4. Frequency-division multiplex multi-channel telephony signals.

lower and upper frequency limits required for typical systems with various numbers of telephony channels are shown in Table 1. The frequencies given are those recommended by the C.C.I.R. for radio systems (Rec. No. 191, Ref. (11)), and are compatible with those recommended by the C.C.I.T.T. for line systems.

TABLE 1.

Base-band Frequency Limits for Various Numbers of Telephony Channels.

NUMBER OF CHANNELS	BASE-BAND FREQUENCY LIMITS (KC/S)
12 (1 group)	12-60 or 60-108
24 (2 groups)	12-108
60 (1 supergroup)	12-252 or 60-300
120 (2 supergroups)	12-552 or 60-552
240 (4 supergroups)	60-1052
600 (10 supergroups)	60-2540

* *The term 'base-band' designates the band of frequencies transmitted by the microwave system between its input and output terminals.*

Consideration is now being given to microwave systems with even larger capacity, e.g. 1 800 or more telephony channels, in which case the base-band might extend from 60 to 8 500kc/s; a microwave system for up to 2 220 telephone channels has been developed by the Bell Telephone Laboratories[6].

The use of f.d.m. techniques is of course an important advantage when line and microwave systems are to be interconnected, since little or no additional equipment is required at the point of interconnexion; it also permits the use of standardized multiplex terminal equipment throughout a mixed line and microwave radio network; and the use of common techniques for dropping and inserting blocks of channels.

Time-division multiplex (t.d.m.) techniques have also been used in systems providing up to 24 or 48 telephone channels[14]. In t.d.m. systems pulses with a half-amplitude duration of a few microseconds are modulated in amplitude, width or position by the audio signals to be transmitted; each pulse in a train of pulses corresponds to a different telephony channel and the train of pulses is repeated at a frequency at least twice the highest audio frequency to be transmitted, e.g. 8kc/s. The main advantages of t.d.m. techniques are that the terminal equipment is cheap and compact, amplitude linearity of the microwave links is unimportant, and individual telephony channels can readily be dropped and inserted; however, t.d.m. systems cannot be interconnected with f.d.m. systems except on an audio-to-audio basis. T.D.M. systems also require considerably greater frequency space than f.d.m. systems of similar capacity and are not suitable for large numbers of telephone channels because of the very narrow pulses and large bandwidths that would be required.

Transmission of Television Signals

Microwave systems are well adapted to the transmission of television signals since the video signal may be transmitted directly in the base-band, i.e. without modulation on to a sub-carrier as is necessary in a long coaxial-cable system in order to avoid low-frequency interference and delay distortion. Furthermore, little or no equalization is required in a microwave system whereas in a long coaxial-cable system the equalization required to correct the variation of the cable attenuation over the frequency band of the television signal may amount to hundreds of decibels.

In current practice, and for monochrome television, the video signal is transmitted over radio-relay systems without modification; however, it would appear that there are significant advantages to be obtained by the use of pre-emphasis, i.e. the attenuation of the lower video components by some 15dB relative to the high frequency components before transmission, the appropriate de-emphasis correction being applied at the output of the terminal receiver. Pre-emphasis enables the same modulators and demodulators to be used either for television or for multi-channel telephony, thus facilitating the interchangeability of broad-band channels and the use of a common standby channel in a multi-r.f. channel system transmitting both television and multi-channel telephony. It would also appear, from experience in the U.S.A., that pre-emphasis facilitates the transmission of colour television signals over microwave links by minimizing the differential delay and amplitude distortion of the chrominance sub-carrier signal.

Microwave systems of very large capacity, e.g. 1 800 or more telephone channels, may be used alternatively to provide a television channel together with some 600 or 900 telephone channels on a single microwave carrier; such systems would offer marked economic advantages on routes where traffic requirements are heavy compared with systems using a number of carriers to provide the same capacity.

Modulation Methods

Frequency modulation is in general use for f.d.m. multi-channel telephony with 12 to 600 or more channels and for television transmission over microwave systems; both frequency modulation and amplitude (pulse) modulation are used for t.d.m. multi-channel telephony with up to 48 channels.

The reasons for this general preference for frequency modulation as compared with amplitude modulation are as follows:

(1) The transmission of f.d.m. multi-channel telephony signals requires a highly linear relation between the input and output amplitudes of the base-band signal if crosstalk between the telephony channels is to be avoided; at the present state of the art this is more readily achieved with f.m. than with a.m.

(2) By using f.m. with deviation ratios greater than unity a substantial improvement in signal-to-noise ratio can be obtained in both multi-channel telephony and television systems. To obtain an equal improvement in an a.m. system would require appreciably larger aerials or increased transmitter power.

(3) The use of f.m. for television results in a distribution of noise in which most of the noise is concentrated at the higher video frequencies, where it is less visible than at low video frequencies; furthermore and for a similar reason, echo signals are less visible with f.m. than with a.m.

However, the use of f.m. for the transmission of f.d.m. multi-channel telephony places stringent requirements on the uniformity of the group-delay/frequency characteristics of the intermediate frequency and microwave components of the system if crosstalk between telephony channels is to be avoided. For the same reason the levels of echo signals, e.g. arising from impedance mismatch in feeders or in the radio path, must not be excessive[15].

Aerials, Feeders, Branching Filters and Isolators

The chief requirements of the aerials for a microwave system are: high gain to reduce the transmitter power required for a given signal-to-noise ratio, good directivity to minimize interference to and from other systems and to reject echoes due to reflections from buildings, hillsides etc., in the radio path, good impedance matching between the aerial and its feeder to minimize reflections which would otherwise cause echoes in television systems or crosstalk in multi-channel telephony systems.

Perhaps the most commonly used aerials in microwave systems are those employing parabolic reflectors or mirrors with a waveguide feed or feed horn at the focus, Fig. 5(a). The sizes of parabolic reflectors used range from about 3 to 6ft diameter for mobile equipment, to 12 or 15ft in diameter for fixed systems; the power gain is proportional to the area of the aperture and is some 40dB relative to a non-directional or isotropic aerial in the case of a 4 000Mc/s aerial with a diameter of 12ft, the corresponding beam-width at half power being about 3°. In some cases arrangements are made for the aerial to be energized with two planes of polarization simultaneously, e.g. to enable a single aerial to be used simultaneously for transmission and reception on nearby frequencies. The maximum bandwidth available on parabolic reflector aerials is usually of the order of 10 to 20 per cent of the mid-band frequency, this being adequate for systems with several radio frequency channels on the same aerial.

In the case of heavily loaded main-route microwave systems it may be advantageous to accommodate three frequency bands centred on say 2 000, 4 000 and 6 500Mc/s on the same aerial; in such cases the horn-reflector aerial shown in Fig. 5(b) can be used. This type of aerial has excellent impedance matching properties over a frequency range of four to one or more, together with very low levels of radiation in unwanted directions.

Although air-spaced coaxial feeders are used up to about 2 000Mc/s, the losses and impedance irregularities of such feeders tend to be prohibitive at higher frequencies and rectangular waveguides are generally used at 4 000Mc/s

modulated, e.g. by means of a reactor valve in u.h.f. systems, or directly using a klystron oscillator in s.h.f. systems, Fig. 6(a).

(2) In which an intermediate-frequency carrier is frequency modulated and this i.f. carrier is combined with a radio-frequency carrier in a frequency-changer, thus producing the required output frequency, Fig. 6(b).

Arrangement (2) is more complex but has certain advantages, e.g. the frequency modulator is a standard unit working on the same frequency irrespective of the radio frequency used. Furthermore the functions of frequency stabilization and modulation of the carrier are carried out separately and generally more effectively.

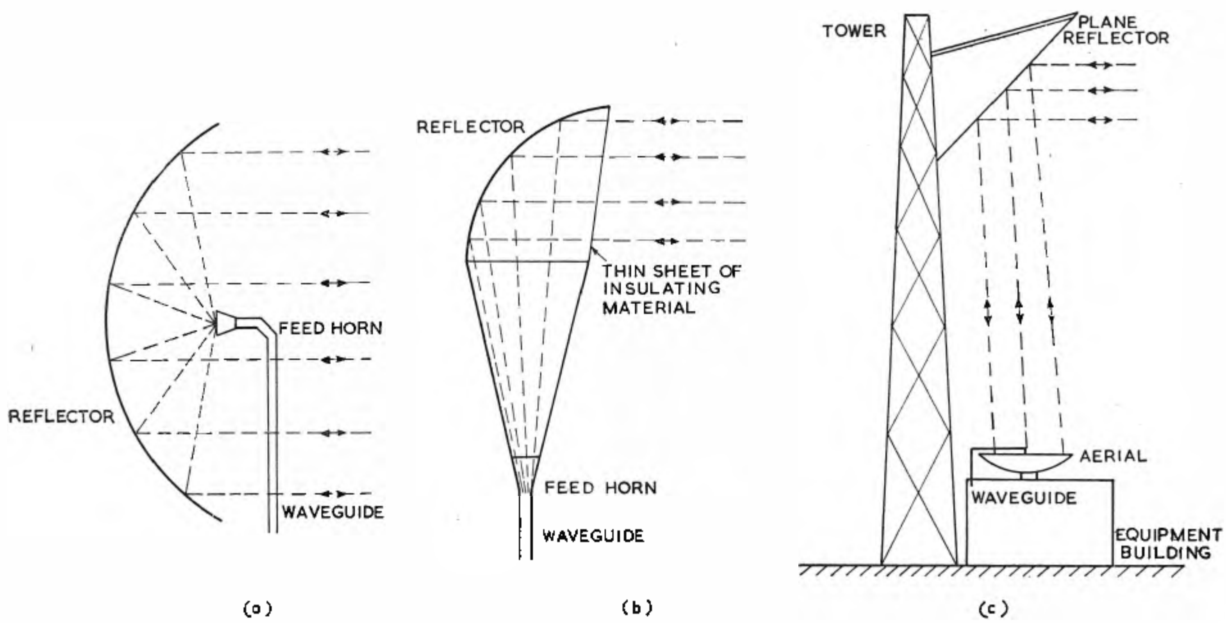


Fig. 5. Microwave aerial systems
(a) Parabolic reflector
(b) Horn reflector
(c) Plane-reflector

upwards; the loss of a typical 2·4in by 1·1in copper waveguide at 4 000Mc/s is about 1dB/100ft.

Plane reflectors are sometimes used, as shown in Fig. 5(c); such reflectors enable the aerials to be kept near ground level, thus facilitating maintenance and avoiding the need for, and cost of, long feeders.

Where transmit and receive radio-frequency channels are to be combined for application to a common aerial, or when several transit (or receive) channels are to be combined for this purpose, branching filters using coaxial-line techniques up to 2 000Mc/s or waveguides techniques at 4 000Mc/s upwards are available.

The non-reciprocal properties of ferrite-loaded coaxial or waveguide components can be used to minimize the effects of impedance mismatch in long feeders; these 'isolators' are particularly useful when transmitters with frequency-modulated oscillators are to be operated directly into long feeders without a buffer amplifier[16].

Terminal Transmitters and Receivers

The basic arrangements of typical frequency-modulation terminal transmitters are shown in Figs. 6(a) and 6(b); they are of two main types:

(1) In which a radio-frequency oscillator is frequency-

In the case of multi-channel telephony microwave systems the mean frequency of the carrier is stabilized; in most of the television microwave systems in current use the carrier frequency corresponding to the tips of the synchronizing pulses is stabilized. However, as noted earlier, there is now a tendency in television microwave systems to use 'pre-emphasis'; this removes most of the asymmetry of the video waveform and enables mean-frequency stabilization of the carrier to be employed, it also enables the same modulators and demodulators to be used for television and multi-channel telephony.

In the case of microwave equipment for fixed links providing large numbers of telephone channels or a television channel with the high standards of performance necessary for long national or international connexions it is usual to interpose an amplifier, e.g. a travelling-wave tube, between the frequency-modulated oscillator and the aerial in order not only to provide a higher power output but also to isolate the oscillator from feeder reflections. This avoids the non-linearity due to frequency-pulling that would otherwise occur; in the case of mobile television link equipment, or multi-channel telephony equipment providing a relatively small number of telephone channels, the frequency-modulated oscillator feeds the transmitting aerial directly, or if a long-feeder is used, via a ferrite isolator[16].

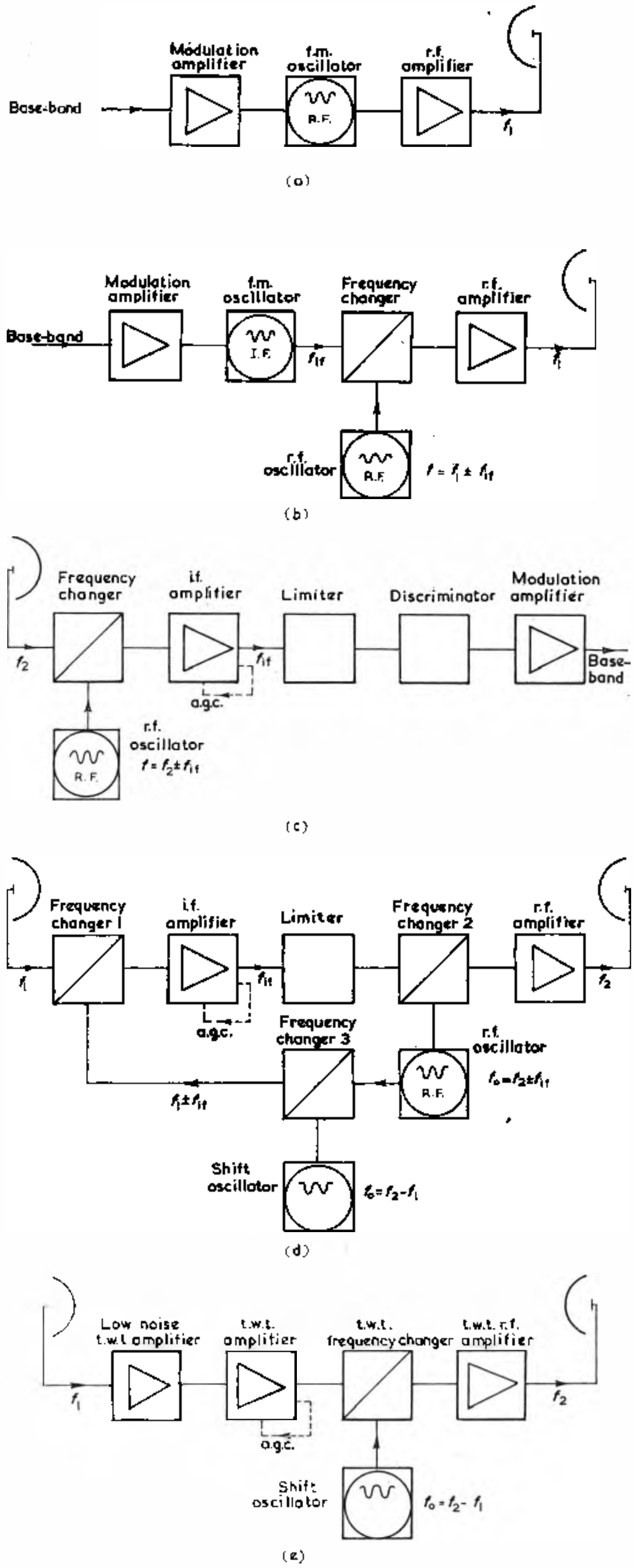


Fig. 6. Basic arrangements of terminal transmitters, terminal receivers and repeaters

(a) Terminal transmitter with frequency-modulated r.f. oscillator
(b) Terminal transmitter with frequency-modulated i.f. oscillator
(c) Terminal receiver
(d) Repeater with i.f. amplification
(e) Repeater with all r.f. amplification

Terminal receivers are almost invariably of the superheterodyne type with a frequency-stabilized local oscillator, as in Fig. 6(c); the superheterodyne principle is used because it enables the desired gain to be readily achieved and facilitates automatic gain control, limiting and demodulation, which are at present difficult to achieve at radio-frequency.

The C.C.I.R. has indicated preferred values for the base-band signal levels and impedances, and for the modulation, intermediate-frequency and radio-frequency characteristics of microwave radio-relay systems used for international connexions (Rec. Nos. 184 to 195, Ref. (11)).

Repeaters

The repeaters generally employed are of the non-demodulating type to avoid an increase in distortion and noise, and the greater level instability that would be entailed by the use of repeaters in which the signal appears at the base-band and is then modulated on to a carrier for onward transmission.

A typical non-demodulating repeater is shown in Fig. 6(d); such a repeater incorporates two frequency changes in the main signal path, the first change is from the incoming radio frequency f_1 to an intermediate frequency f_{if} at which most of the required gain is achieved, the second frequency change is from the intermediate frequency to the outgoing frequency f_2. Automatic gain control and some limiting are provided at intermediate frequency in order to stabilize the output level. The output from the second frequency changer is amplified to a power level suitable for radiation to the next repeater station.

To avoid cumulative frequency errors provision is made for deriving the local oscillator feeds for the first and second frequency changers from a common radio-frequency source; in the example shown in Fig. 6(d) the radio-frequency oscillator feeds the second frequency-changer at a frequency $f = f_2 \pm f_{if}$ and the feed for the first frequency-changer is obtained from it by modulating a portion of the radio-frequency oscillator output by the shift frequency $f_o = f_2 - f_1$ in a third frequency-changer, the sideband frequency $f_1 + f_{if}$ being selected and used.

It is not essential to translate the incoming signal to an intermediate frequency in a non-demodulating repeater, all the necessary amplification can be achieved at radio frequency. For example, a repeater may consist of a two-stage low-noise travelling-wave tube amplifier with fast-operating automatic gain control, followed by a travelling-wave tube used as a frequency-changer[17] and finally a travelling-wave tube power amplifier, Fig. 6(e). In such a repeater the shift-frequency oscillator f_o is applied to the travelling-wave tube frequency changer and generates sidebands $f_1 + f_o$, one of which is selected and amplified.

Multi-R.F. Channel Systems

In the case of heavily loaded routes it may be necessary to provide several independent broad-band channels over the same route. The wide bandwidths that can be accommodated on microwave aerials makes it possible to use the same aerial for a number of r.f. channels, simultaneously for transmission and reception if required. Such an arrangement reduces the cost of a link, not only by reducing the number of aerials and feeders but also by enabling lighter and the therefore cheaper towers or masts to be used to support the aerials.

A repeater for three r.f. channels is shown schematically in Fig. 7; in this case each aerial transmits and receives three r.f. channels simultaneously. A bi-polar feed can be used to provide some discrimination between transmitted and received signals on the same aerial. Additional discrimination is provided by the selectivity of the repeaters and the associated band-pass filters (the latter are not shown in Fig. 7). Branching filters are used to separate the individual r.f. channels without significant loss.

The frequency plan for multi-r.f. channel systems must be chosen with great care in order to minimize the risk of interference from (a) the various oscillators in the system, (b) the high-level transmitted signals and (c) from any intermodulation products of the latter. Account must also be taken of the 'image' channels of the transmitter and receiver, i.e. channels spaced from the wanted channel by twice the intermediate frequency.

The frequency plan recommended by the C.C.I.R. (Rec. No. 194, Ref. (11)) for 600-channel telephony or television systems is shown in Fig. 8. The plan provides for up to six r.f. channels in each direction and is intended for use with an intermediate frequency of 70Mc/s; the spacing between the centres of adjacent channels is 29Mc/s and

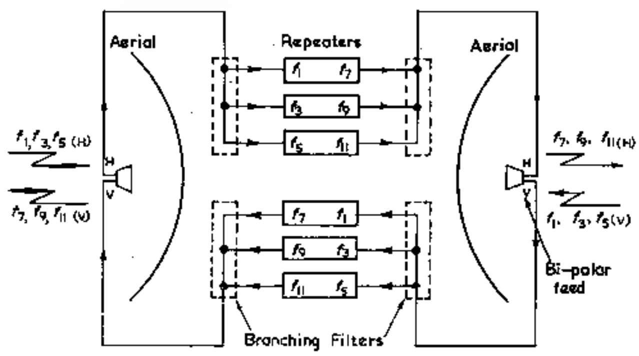


Fig. 7. Three r.f. channel repeater

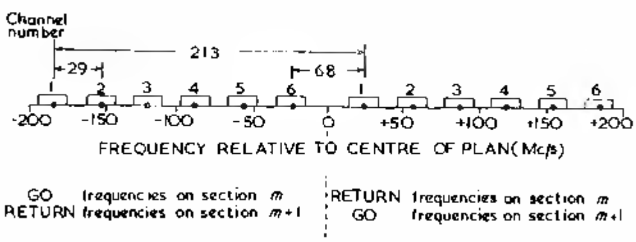


Fig. 8. Frequency plan recommended by the C.C.I.R. for six r.f. channel systems for multi-channel telephony (600-channels) or television

the 'shift frequency', i.e. the frequency interval between corresponding go and return channels, is 213Mc/s. Adjacent r.f. channels operate with different polarization in order to facilitate the separation of channels. Channels on frequencies interleaved with those on the main plan can also be used if desired, e.g. for spur or crossing routes.

If six r.f. channels are required on one route, two 3-r.f channel systems such as that shown in Fig. 7, can be provided; in such a case one system would use the odd number channels of the frequency plan, and the other the even channels.

Protection or Stand-by Channels; Stand-by Power Supplies; Supervisory and Control Systems

A microwave system may be required to achieve a reliability of 99·9 per cent or more; in order to achieve a reliability of this order it is necessary to provide a 'protection' or stand-by channel which can be switched into use manually or automatically when a working channel fails. In systems with up to five working channels on the same route only one protection channel may be needed since the latter is required for a very small percentage of the time and simultaneous faults on two or more working channels are extremely unlikely. The use of a stand-by channel on a separate radio frequency is generally preferred to the use of stand-by equipment operating on the same frequency as the working equipment and switched at individual stations, since the former arrangement enables the stand-by channel to be monitored continuously or used as a temporary additional working channel when required, e.g. for television outside broadcasts.

A pilot tone is sometimes transmitted in the base-band of both working and stand-by channels, above the range of frequencies occupied by the multi-channel telephony or television signal; the pilot tone enables the continuity of each channel to be established, and by measuring the noise level in a narrow band adjacent to the pilot, provides a basis for switching automatically from a working to a stand-by channel should the signal-to-noise ratio in the working channel become inferior to that in the stand-by channel. The C.C.I.R. has recommended preferred values of the pilot frequencies, i.e. 3·2Mc/s for 600-channel systems and 7 or 8·5Mc/s for television systems (Rec. No. 198 and 199, Ref. (11)). In some systems the stand-by channel is also used to provide frequency-diversity operation in order to minimize the effects of fading[16].

Since the normal mains supply cannot always be relied on to the degree of reliability required, it is usual to provide a stand-by generator driven by a petrol or oil engine. In some systems, e.g. those transmitting multi-channel telegraphy, it may be essential to avoid interruptions and in such cases a 'continuity set' is employed. These sets comprise a motor driven from the mains, a generator and a heavy flywheel on the same shaft, and an engine. In the event of a mains failure the engine is started up automatically and is coupled to the motor-generator via a magnetic clutch, the flywheel providing energy to maintain the electrical power output until the engine comes into operation.

Since the repeater stations are normally unattended it is necessary to provide a supervisory and control system for identifying at staffed control stations associated with terminal stations, the repeater station at which a fault has occurred, in order that a maintenance man may visit the station and correct the fault. It is also desirable to indicate broadly the nature of the fault, e.g. urgent or non-urgent, and to indicate to the control station when a mains failure has occurred, so that steps may be taken to replenish the fuel supply of the stand-by generator set. This information is usually transmitted by keyed audio-frequency tones, similar to voice-frequency telegraph signals, sent either over a line circuit, or in an additional 3kc/s wide channel or channels in the base-band of the radio-relay system, or in channels provided by an auxiliary low-power radio-relay system. A service channel for speech transmission is also provided between all the repeater stations and the staffed control stations to facilitate maintenance operations and control of the system.

Tropospheric Forward-Scatter Systems

The fact that radio waves in the frequency range from about 100 to 10 000Mc/s are consistently propagated well beyond the visible horizon to distances of up to a few hundred miles has now been firmly established, mainly as a result of experiments carried out since 1950. However, it should be noted that Marconi, as far back as 1932, demonstrated that signals on about 500Mc/s could be received at a distance of 168 miles, but because of the low power used in his experiments the signals were not reliable.

In the United Kingdom, the pioneering work of the late Dr. E. C. S. Megaw of the Admiralty Signal and Radar Establishment demonstrated, in 1950, that signals on about 3 000Mc/s could be consistently received on an oversea path well beyond the visible horizon. It was shown that these signals could not be accounted for by ray bending due to refraction in the normal undisturbed atmosphere, nor could they be accounted for by propagation in ducts over the sea surface; the signals were in fact attributed to radio-wave scattering caused by atmospheric turbulence.

Since 1950, a great deal of experimental and theoretical work on tropospheric forward-scatter propagation has been carried out[5,6,7,19]; there is now available a considerable volume of data on which to base the design of systems.

Tropospheric forward-scatter radio systems have now emerged from the experimental and development stage and a number of such systems are in operation; for example, extensive tropospheric forward-scatter radio links have been installed in Canada and Alaska to provide communication facilities between radar stations, and for civil communications; these links provide 18, 36 or in some cases up to 132 telephone circuits over distances of 1 000 miles or more in hops of 100 to 200 miles. A tropospheric forward-scatter link is now in operation on a 184-mile oversea path between Florida and Cuba, providing up to 240 telephone circuits (up to 120 circuits on each of two carriers) or a television channel. In Europe a tropospheric forward-scatter link is in operation between Sardinia and Minorca, a distance of 240 miles, and other systems are planned. However, extensive applications of tropospheric forward-scatter systems for civil communications in countries such as the United Kingdom with extensive line, cable and line-of-sight microwave systems seem unlikely.

The prime advantage of tropospheric forward-scatter systems, compared with line-of-sight microwave systems or coaxial-cable systems, is that they provide reliable communication over distances of up to 200 miles or more without repeater stations or interconnecting cables. This feature is of considerable value when the intervening terrain is difficult of access or when sea crossings are involved.

On the other hand the larger range tropospheric forward-scatter systems require very large aerials and very high-power transmitters which not only involve heavy capital expenditure but also, in the case of the transmitters, considerable annual charges.

It is by no means obvious that a tropospheric forward-scatter system is cheaper, either in capital costs or in annual charges, than a line-of-sight microwave system or a cable system; each must be considered on its merits in relation to the job to be done. Furthermore, line-of-sight microwave and cable systems can readily provide for inserting or dropping blocks of channels at intermediate points; this is of considerable value and is often required in a complex inland network. By its nature, such a facility is not available in a tropospheric forward-scatter system.

Tropospheric forward-scatter systems are subject to certain technical limitations. For example, it would appear that the frequencies used cannot readily be shared in the same geographical area with conventional low-power line-of-sight microwave systems since the high powers used by the scatter systems, and the ease with which scatter signals are propagated over hundreds of miles, introduces a considerable risk of interference.

At their present stage of development tropospheric forward-scatter systems require much more frequency space than line-of-sight systems of comparable capacity, a limitation which arises from the large distances over which scatter signals are propagated and the need to use different frequencies for the various sections of a multi-section route in order to avoid mutual interference.

Finally, it is not practicable at the present time to engineer the longer tropospheric forward-scatter links fully to meet the high standards recommended by the C.C.I.T.T. and C.C.I.R. for international connexions. For a small fraction of the time the noise level on a long tropospheric forward-scatter link may be higher than is allowable on international circuits; nevertheless the performance may be acceptable for many purposes.

Characteristics of Tropospheric Forward-scatter Propagation

Propagation by forward-scattering is essentially one involving a large transmission loss, thus it is necessary to use high-gain and therefore narrow beam aerials both for transmitting and receiving if the loss is not to be excessive. In practice the aerial beams are arranged as shown in Fig. 9 to intersect in a 'common volume' in which scattering occurs.

The scattering angle θ between the transmitting and receiving beams is of great importance and for effective transmission must be kept as small as possible. This implies

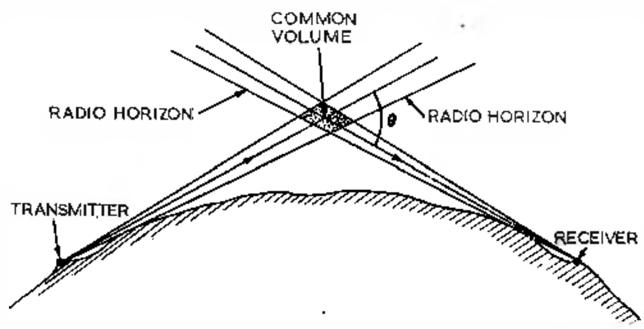


Fig. 9. Geometry of tropospheric forward-scatter radio link

that the transmitting and receiving sites must be chosen so that they have an unobstructed view to the natural horizon.

The maximum range of signals scattered from the troposphere is determined by the greatest height at which scattering can occur in that region; the troposphere extends up to about six miles and the corresponding maximum range is about 400 miles. However, recent investigations[19] suggest that signals scattered from the stratosphere, i.e. the region between the troposphere and the ionosphere, may extend the maximum range to some 700 miles.

The more important characteristics of tropospheric forward-scatter propagation are as follows:

(1) The instantaneous received signal strength fluctuates continuously since the signal itself is the resultant of a large number of components of random and varying phases.

(2) The mean signal level varies from hour-to-hour, day-to-day and month-to-month by some 10 to 20dB or more.

(3) The received signal levels are relatively weak, being from 50 to 100dB or more below the free-space level.

(4) Usable signals can be received consistently even up to some 400 miles or more provided that very large aerials, very high-power transmitters and sensitive receivers are employed.

(5) Frequencies from 100 to 10 000Mc/s can be used; however, frequencies below about 1 000Mc/s are generally more effective for the longer paths.

(6) Bandwidths of three or four megacycles per second, i.e. sufficient for a television channel or some hundreds of telephone channels, may be transmitted over distances up to about 200 miles provided very large, narrow-beam aerials and high transmitter powers are employed; however, the useful bandwidth decreases rapidly as the distance is increased. There is as yet insufficient experimental data on which to assess the quality of transmission from the aspect of television signals and more data of this type is needed.

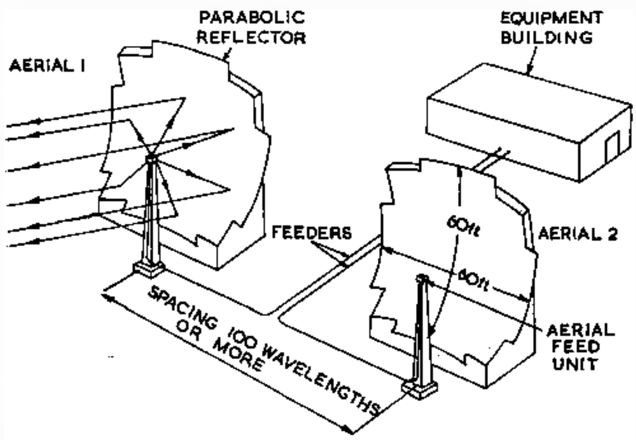


Fig. 10. Arrangement of aerials: tropospheric forward-scatter radio link

Techniques Employed in Tropospheric Forward-scatter Systems

Tropospheric forward-scatter systems are characterized by large aerials and high-power transmitters; the transmitter powers currently employed range from 1 to 20kW and the aerial sizes from 12 to 60ft diameter respectively for systems providing from 12 to 120 telephone circuits over paths from 100 to 250 miles long. Still larger aerials, up to 120ft diameter, and powers up to 50kW are planned for systems providing 100 telephone circuits over paths up to 400 miles long[20].

The aerials used are generally of the parabolic reflector type (Fig. 10), mounted some twenty wavelengths above the local ground; elevated sites are advantageous since they enable the scattering angle θ, and thus the scatter transmission loss, to be reduced.

Aerials with gains of 30dB or more may not realize the nominal plane-wave gain due to lack of coherence of the wave-front, thus an aerial with a nominal plane-wave gain of 50dB may yield under 40dB for 10 per cent of the time. Nevertheless high-gain, narrow-beam, aerials may be necessary in order to achieve the desired bandwidth by reducing the time-delay differences between the components of the scatter signal.

It is usual because of the large size and high cost of the aerials to employ the same aerials for transmission and reception, diplexers being used to combine the transmitter output and the receiver input, Fig. 11. R.F. filters are provided at the transmitter output to minimize noise in the received frequency band, and at the receiver input to reject the high-level signal at the transmitted frequency.

Klystron amplifiers are usually employed in the high-power output stages of transmitters because of the large power gains and high efficiency obtainable; gains of 50dB and output powers up to 10kW at 2 500Mc/s, 2kW at 6 000Mc/s are available[21,22].

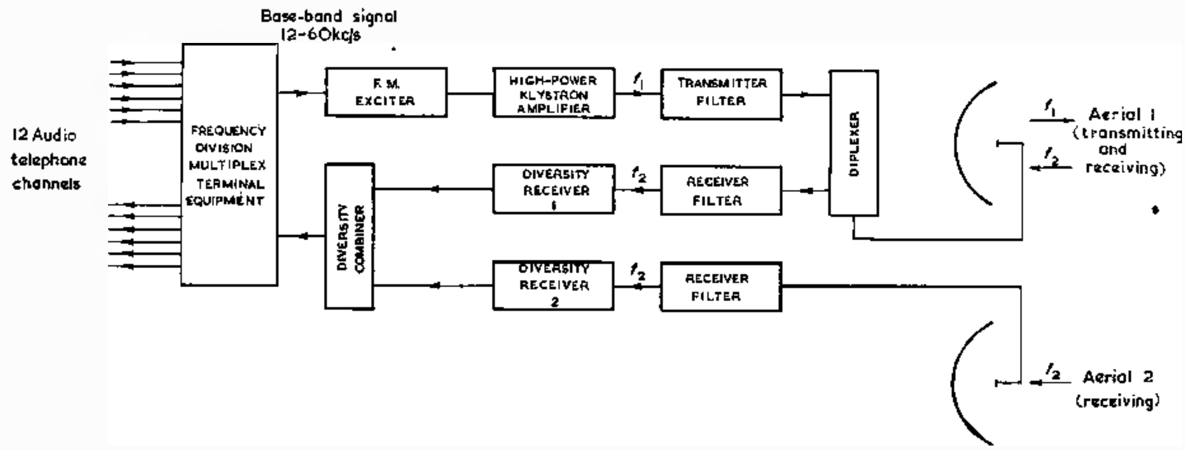


Fig. 11. Tropospheric forward-scatter link: equipment schematic

In most systems the high-power transmitter amplifier is driven by a low-power frequency-modulation exciter, with frequency-division multiplexing of the telephony channels. Most of the advantages of frequency-modulation already noted in connexion with line-of-sight microwave systems also apply to tropospheric forward-scatter systems; however, it would appear that single-sideband modulation has certain characteristics which may make it attractive for tropospheric forward-scatter systems, these include the ability to operate satisfactorily at low levels of received signal due to the absence of a threshold effect, a saving of up to 10 to 1 in the frequency space required, freedom from intermodulation due to multipath signals, reduced interference to other systems and economy in high-tension power consumption. On the other hand a high degree of amplitude linearity is essential in a single-sideband system[23].

Because of the random fading which is characteristic of scatter propagation it is usual to employ spaced-aerial or other types of diversity operation; the simpler systems employ selection of the stronger of two or more signals by switching, more advanced systems employ all the diversity signals in combination, the signal levels being automatically and continuously adjusted for optimum diversity gain before combination. A double-diversity system of the latter type yields an improvement compared with a non-diversity system of about 4dB for 50 per cent of the time and 17dB for 0·1 per cent of the time; corresponding improvements for a quadruple-diversity system are 7dB and

27dB respectively. Since the large improvements relate to the small percentages of time when the signal-to-noise ratio is poor, the advantage from diversity is specially valuable.

The use of frequency diversity is deprecated because of the additional frequency allocations required and the need to conserve the use of the spectrum.

Fig. 12 shows a quadruple-diversity system in which spaced aerials and crossed polarizations can be employed to achieve in effect four separate paths in space; it is notable that only two frequencies are employed in this system[24].

Future Developments

The future developments of line-of-sight microwave systems for main routes carrying heavy traffic is likely to be in the direction of systems transmitting perhaps 1 800

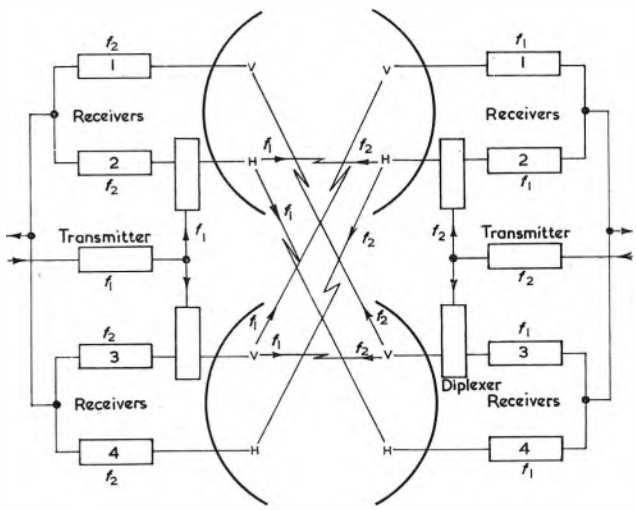


Fig. 12. Quadruple-diversity tropospheric forward-scatter radio link with spaced aerials and crossed polarization

telephone circuits, or a television channel and some hundreds of telephone circuits, on a single radio carrier; such systems could lead to substantial cost savings as compared with systems requiring several radio carriers to provide the same total capacity.

At the other end of the scale there would appear to be scope in some countries for low-capacity microwave systems accommodating up to 48 telephone circuits. In this field there is scope for printed 'micro-strip' circuits and for intermediate-frequency amplifiers and other components including time-division multiplex terminal equipment, using printed-circuit techniques and transistors where appropriate, to reduce the size, weight, power consumption and cost[25,26].

It is interesting to speculate on the advantages which low-noise solid-state microwave amplifiers, 'MASERS'[27], might bring to microwave systems; if these amplifiers enabled the noise factor of the receivers to be reduced by say 10dB and if transistors were used for the intermediate frequency and other circuits, these measures would reduce the power supply requirements for a typical main-route repeater station from several kilowatts to a few hundred watts. Standby power supplies could be provided from batteries and large engine-generator sets would be unnecessary and the volume of the radio equipment and the size of the buildings could be considerably reduced, with large cost savings.

Tropospheric forward-scatter radio systems are likely to find applications where difficulties of terrain make it impracticable or too costly to use conventional line-of-sight microwave systems or coaxial-cable systems. The disadvantages of tropospheric-scatter systems, notably their lack of spectrum economy and interference potentiality, make it unlikely that they will find wide application for normal civil communications in countries with extensive networks of conventional systems.

The future development of tropospheric forward-scatter systems is likely to include improved modulation techniques, e.g. single-sideband modulation, aiming at spectrum economy and improved performance under difficult propagation conditions. Additional spectrum economy could be achieved by the development of improved designs of aerials with smaller minor lobes, since this would reduce the number of frequency allocations needed to avoid 'over-reach' interference via the minor-lobe responses. The use of single-sideband modulation and the development of very high-power transmitters, e.g. with peak powers of up to 100kW, and exceptionally large aerials, e.g. 120ft in diameter, may be expected to lead to systems capable of providing 100 telephone circuits over distances of some 400 miles without intermediate repeater stations.

Acknowledgments

The thanks of the author are due to the Engineer-in-Chief of the General Post Office for permission to publish this article.

REFERENCES

1. Dawson, G., Hall, L. L., Hodgson, K. G., Meers, R. A., Merriman, J. H. H. The Manchester-Kirk o'Shotts Television Radio-relay System. *Proc. Instn. Elect. Engrs. 101*, Pt. 1, 93 (1954).
2. Starr, A. T., Walker, T. H. Microwave Radio Links. *Proc. Instn. Elect. Engrs. 99*, Pt. 3, 241 (1952).
3. Roetkin, A. A., Smith, K. D., Friis, R. W. The TD-2 Microwave Radio-relay System. *Bell Syst. Tech. J., 30*, 1041 (1951).
4. McDavitt, M. B. 6000 Mc/s Radio-relay System for Broadband, Long-Haul Service in the Bell System. *Alta Frequenza. 26*, No. 5. (Oct. 1957).
5. The Scatter Propagation Issue. *Proc. Inst. Radio Engrs. 43*, (Oct. 1955). (This issue contains several papers on tropospheric forward-scatter systems). See also the papers of the Symposium on Long-Distance Propagation above 30 Mc/s, held by the Institution of Electrical Engineers in Jan. 1958. (To be published as a supplement to Pt. B of the Proc. Instn. Elect. Engrs.).
6. Symposium on Communications by Scatter Techniques. I.R.E. Transactions on Communications Systems. CS-4, No. 1. (March, 1956).
7. Telford, M. Communications Potentialities of Tropospheric Scatter. *Point-to-Point Telecommunications. 1.* No. 2. (Feb. 1957).
8. Rendall, A. R. A., Anderson, W. N. Temporary Linkages for Outside Broadcast Purposes. *Proc. Instn. Elect. Engrs. 99*, Pt. 3A, 323 (1952).
9. Dawson, G. Portable Equipment for a Microwave Television Link. *Proc. Instn. Elect. Engrs. 99*, Pt. 3A, 379 (1952).
10. Bullington, K. Radio Propagation Fundamentals. *Bell Syst. Tech. J. 36*, 593 (1957).
11. International Radio Consultative Committee (C.C.I.R.), Documents of the 8th Plenary Assembly, Warsaw, Vol. 1. (1956).
12. Joint CCIR/CCITT Committee for Television Transmissions (CMTT), Paris, Document 12 (1957).
13. Lewis, N. W. Waveform Responses of Television Links. *Proc. Instn. Elect. Engrs. 101*, Pt. 3, 258 (1954).
14. Speed, R. F. B. A 24-Channel Pulse-Time Modulation System. *Proc. Instn. Elect. Engrs. 102B*, 375 (1955).
15. Albersheim, W. J., Schafer, J. P. Echo Distortion in the Transmission of Frequency-Division Multiplex. *Proc. Inst. Radio Engrs. 40*, 316 (1952).
16. Clarricoats, J. B. et al. A Survey of the Theory and Applications of Ferrites at Microwave Frequencies. I.E.E. Convention on Ferrites. (Oct, Nov. 1957). (Published as supplement to Pt. B of the *Proc. Instn Elect. Engrs.*).
17. Bray, W. J. The Travelling-Wave Valve as a Microwave Phase-Modulator and Frequency-Shifter. *Proc. Instn. Elect. Engrs. 99*, Pt. 3, 15 (1952).
18. Bellows, B. C. Automatic Switching for TD-2 Radio. *Bell Syst. Tech. J. 35*, 98 (1956).
19. Booker, H. G., Gordon, W. E. The Role of Stratospheric Scattering in Radio Communication. *Proc. Inst. Radio Engrs. 45*, 1223 (1957).
20. Mellen, G. L. UHF Long-Range Communication System. *Proc. Inst. Radio Engrs. 43*, 1269 (1955).
21. Moreno, T. Transmitting Tubes for Scatter Communications. I.R.E. Transactions on Communications Systems. CS-4, No. 1, 64 (1956).
22. Speaks, F. A. Power Amplifier Klystron for UHF Transmission. I.R.E. Transactions on Communications Systems. CS-4, No. 1. 69 (1956).
23. Morrow, W. E., Mack, C. L., Nichols, B. E., Leonhard, J. Single-Sideband Techniques in UHF Long-Range Communications. *Proc. Inst. Radio Engrs. 44*, 1854 (1956).
24. Altman, F. J., Sichak, W. A Simplified Diversity Communication System for Beyond-the-Horizon Links. I.R.E. Transactions on Communication Systems. CS-4 No. 1. 50 (March, 1956).
25. Dukes, J. M. C. The Application of Printed Circuits to the Design of Microwave Components. (To be published in *Proc. Instn. Elect. Engrs. 105*, Pt. B. 1958).).
26. Engelmann, H. et al. Portable Multi-channel 11 000Mc Radio Link. I.R.E National Convention Record. Pt. 8. 150 (1957).
27. Wittke, J. P. Molecular Amplification and Generation of Microwaves. *Proc. Inst. Radio Engrs. 45*, 291 (1957).

Microwave Line-of-Sight Propagation

By M. W. Gough*, M.A., A.M.I.E.E.

After sketching the historical and physical background of the subject, this article surveys some fundamental investigations into microwave line-of-sight propagation made, mainly since 1939, in Europe and the U.S.A. Theoretical work, based principally on classical optical concepts, is reviewed in conjunction with related experimental investigations. Statistical processes important in the research and engineering aspects of the subject are also discussed.

At the turn of the century when Marconi was carrying out his experiments on long distance transmission on low frequencies, the theoretical study of terrestrial radio propagation was based on an idealized model of a smooth spherical earth surrounded by free space. The great discrepancy between the experimental results and the theoretical predictions led to the conclusion that a physical mechanism not included in the mathematical formulation, namely the reflection of the waves by an ionized region in the upper atmosphere, must be exerting a major influence on the propagation.

As far as the effect of the ground and of the lower atmosphere was concerned, however, the model was reasonably good. Only in very mountainous country would the ground irregularities be serious, while the effect of refraction would be small on such low frequencies where the wavelength is large compared with the scale of the atmospheric variations.

With the extension of ground wave theory to ultra high frequencies where the effect of the ionosphere is negligible, it becomes possible under most practical conditions to treat the propagation within the horizon by geometrical-optics, but the ground irregularities become of major importance, while the ray paths are curved by refraction in the non-homogeneous atmosphere. On the assumption that there is a standard atmosphere with a constant negative gradient of refractive index in the troposphere immediately above the surface of the earth, it has been shown that the propagation is equivalent to transmission through an homogeneous atmosphere, or in effect free space, over an earth with its radius increased by a geometrical transformation that removes the curvature of the ray paths.

It was not at first appreciated that variations from the assumed standard atmosphere can produce profound modifications to the nature of tropospheric propagation, though fluctuations of signal-strength, or fades, were in general attributed to temporal changes in the structure of the troposphere. With the development of radar techniques prior to and during the Second World War, evidence accumulated of long distance tropospheric propagation under conditions of extreme super-refraction and during the presence of elevated inversion layers. This phenomenon has been explained by an appropriate extension of the mathematical theory.

Further experiments after the war showed that even under normal conditions when the field-strength beyond the horizon drops away exponentially a distance is reached beyond which the average signal decreases much more slowly and persists out to great distances. These signals have been interpreted in terms of scattering by atmospheric turbulences and have become the subject of intensive study in connexion with the establishment of long distance scatter links.

The whole subject is now referred to as tropospheric propagation and includes reception within the horizon and ground-wave propagation in the presence of a standard atmosphere. It can be conveniently divided into two parts, one dealing with line-of-sight propagation and the other with propagation beyond the horizon with special reference to tropospheric scattering. The present article is concerned with the former, while the companion article by G. Millington reviews the latter with regard to the present position of research being carried out in the United Kingdom.

There is of course an intermediate situation in which the transmission distances are quite small but the propagation is not strictly line-of-sight, i.e., where local obstructions rather than the curve of the earth prevent an optical path between the transmitter and the receiver. Such conditions are of prime importance in considering the service areas of television stations, where the diffractive losses over hills have to be assessed by Fresnel 'knife-edge' theory and the variations in the meteorological factors are of secondary importance except at the fringes of the service area.

This is a specialized aspect of ground-wave propagation that will not be considered in detail here, where the emphasis will be on point-to-point links between terminals with an unobstructed path between them, but account has to be taken of the ground irregularities in considering rays reflected from the surface of the earth, and of the effect of the troposphere on the ray paths through it.

In this article microwaves will be deemed to comprise wavelengths shorter than 1m (frequencies above 300Mc/s), where, as shown in the first part of this review, many features and applications of propagation theory are borrowed from the concepts of classical optics. The second part of this survey gives examples of the statistical approach to propagation problems, necessitated by the temporal and spatial variations in the atmosphere. Space prevents discussion of much pioneering work, and this review is therefore restricted mainly to war-time and post-war investigations, when the development and operation of radar, followed by rapid advances in wide-band multichannel radio systems, gave a great impetus to research on microwave propagation.

Optical Aspects

Ground Clearance

Reliable microwave communication necessitates a clear line-of-sight between the terminals, with the further require-

* *Marconi's Wireless Telegraph Co. Ltd.*

ment of first Fresnel zone clearance of all points on the intervening ground. Under these conditions the signal strength will be near the 'free-space' value, except for possible departures due to ground reflections and atmospheric effects discussed later. Simple formulae for the attenuation between various types of aerial in free space have been given by Friis[1], and the basic principles of microwave propagation have been briefly reviewed by Megaw[2], Bullington[3], Thompson[4] and others.

Failing first Fresnel zone clearance and disregarding secondary influences, signal strengths will fall below the 'free-space' level. Adapting classical edge-diffraction theory, Megaw[2], Bullington[3], Thompson[4] and others give charts showing the 'shadow loss' due to a specified obstructing screen at various wavelengths, while curved earth diffraction has been rigorously investigated by Eckersley[5], Millington[6], van der Pol and Bremmer[7], Burrows and Gray[8] and others. While ideally, microwave paths should maintain first Fresnel zone clearance under all atmospheric conditions, some in practice default during periods of sub-refraction and suffer consequent long-period fading. The influence of inadequate ground clearance on the statistical behaviour of such paths has been investigated experimentally by Durkee[9], Wheeler and Mathwich[10], and Thompson[4].

Ground Reflections

Ground reflections may adversely affect line-of-sight propagation because, due to siting exigencies and changing atmospheric conditions, phase opposition between direct and ground-reflected waves cannot always be avoided. In this situation near-equality in their amplitudes gives a weak signal and therefore microwave paths should where possible be chosen with small ground reflection coefficients—less than 0·5 for example. Over nearly 50 established 4000Mc/s. paths in the U.S.A., Bullington[11] has measured reflection coefficients seldom exceeding this value, and averaging about 0·28.

Much theoretical and practical study has been devoted to reflection coefficients. McPetrie[12], Kerr[13], Ohman[14] and others have published theoretical charts giving the amplitude and phase of the reflection coefficient of radio waves at a plane surface under various conditions, and conformatory measurements have been made by Megaw[2], Ford and Oliver[15] and others. For smooth ground in the frequency range under review, the reflection coefficient is virtually unaffected by conductivity and depends only on the wave polarization, the dielectric constant and the reflection angle θ as defined in Fig. 1(a). For vertical polarization the Brewster effect, whereby the reflection coefficient becomes zero at a particular angle, always results in a lower reflection coefficient with vertical than with horizontal polarization for a specific path. Nevertheless at the shallow reflection angles characteristic of ground-to-ground propagation, which seldom exceed 2°, both polarizations give reflection coefficients near unity, and thus for land paths the controlling factor is the roughness of the ground rather than its dielectric properties.

For sea, surface roughness cannot be relied on for reducing the reflection coefficient to an innocuous level, but this disadvantage is partly offset by the fact that, for sea reflection, a pseudo-Brewster effect with vertical polarization causes a rapid initial reduction in reflection coefficient with increasing reflection angle. Thus compared with horizontal polarization, which entails coefficients very near unity in practical cases, the use of vertical polarization should reduce fading attributable to direct and sea-reflected wave interference.

Fig. 1(b), after Kerr[13], shows theoretical relationships between reflection coefficient and reflection angle for smooth sea for wavelengths between 1m and 1cm, using the best available data for dielectric constants and conductivities, which for sea water are somewhat frequency-dependent. For vertical polarization the pseudo-Brewster minimum approaches zero at very short wavelengths due to the subordinate influence of conductivity in that case. However, there is a progressive reduction in the initial slope of the curves with decreasing wavelength (Fig. 1(b)), which reduces the benefit from vertical polarization on the

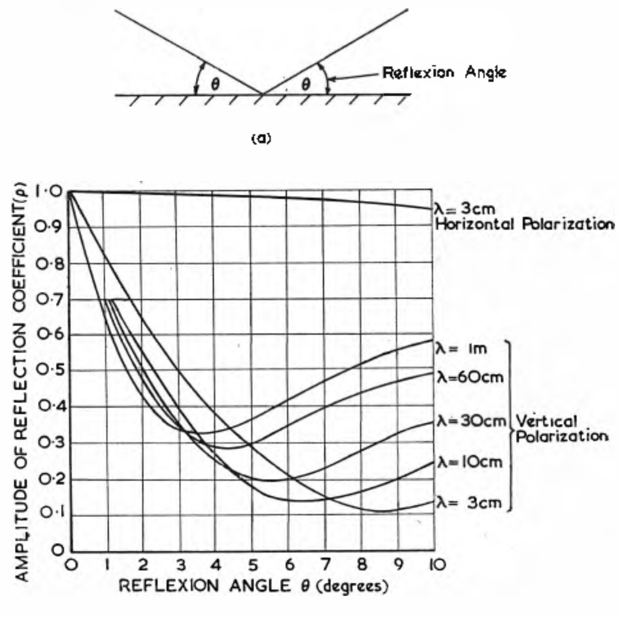


Fig. 1. Reflection coefficient versus reflection angle for sea (after Kerr)

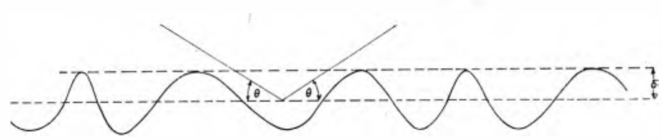


Fig. 2. Reflection at a rough surface

shorter wavelengths. Bray and Corke[16] in fact found that vertical polarization showed no benefit over horizontal on a 3 930Mc/s optical sea path, but the present author anticipates a significant advantage on the longer microwaves.

Several workers have considered ground roughness from the statistical aspect of Rayleigh's roughness criterion, which states that reflection from a rough surface becomes diffuse when wavelets scattered from irregularities are disturbed in phase by more than a specified amount from the mean value. For a reflection angle θ, and irregularities of height δ above and below the mean level (Fig. 2) the maximum phase deviation ϕ_m from the mean value is $4\pi\delta \sin\theta/\lambda$, where λ is the wavelength. The Rayleigh criterion is variously quoted as $\phi_m > \pi/4$ or $> \pi/2$ radians. The crucial feature is the association between the reduction in reflection coefficient due to ground roughness and the quantity $\delta \sin\theta/\lambda$. Discussing vertical polar patterns of radar aerials, Norton and Omberg[17] imply that for a well developed lobe the quantity $\delta \sin\theta/\lambda$ must not exceed $1/16$ throughout the first Fresnel zone for ground reflection.

Day and Trolese[18], Bullington[11] and others have deduced reflection coefficients of rough ground from interference patterns revealed by height-gain measurements. Fig. 3 shows height-gain curves obtained on 3 microwave frequencies over the Arizona Desert, using an aircraft ranging in height between 600 and 7 700ft. The transmitting aerial remained at 190ft and the aircraft distance was about 55 miles. The depth of any signal minimum (Fig. 3) is a measure of the nearness to unity of the ground reflection coefficient for the test condition indicated. The minima are labelled with the coefficients inferred.

From 4 000Mc/s height-gain tests on 200ft towers, Bullington[11] has deduced values of the reflection coefficient, ρ, for land paths in the U.S.A. already mentioned, and has correlated these with corresponding values of ϕ_m estimated from the path irregularities. ρ ranged between 0·1 and 0·7, and a plot of ρ versus ϕ_m shows scattered points whose trend for high values of ϕ_m lies well above a theoretical curve based on random ground irregularities, but conforms better to theoretical curves based respectively on 'saw-tooth' and 'sine-wave' surfaces. Bullington

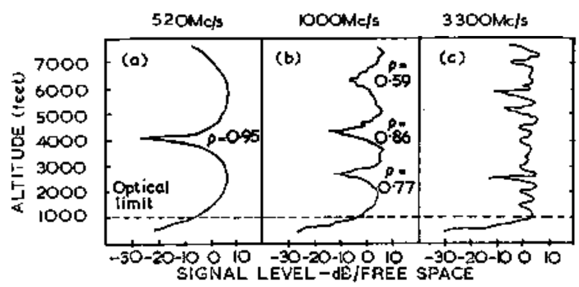


Fig. 3. Aircraft height-gain curves (after Day and Trolese)

remarks that departures from true randomness in the actual path profile are likely to increase ρ at high values of ϕ_m, in conformity with the observed trend. A discussion by Norton and Omberg[17] on Fresnel zones for a reflecting surface prompts the suggestion that Bullington's sporadic high values of ρ may have resulted from a specular reflection from surfaces comparable in size with the first Fresnel zone, which on the test frequency can be quite small. The erratic height-gain curve of Fig. 3(c), after Day and Trolese, shows a similar effect.

Ford and Oliver[15] have made extensive measurements on 9cm of the reflection coefficient of a small area of prepared ground, using a direct method needing a very short radio path (50ft), narrow beam aerials and steep reflection angles. For each polarization the reflection coefficient of smoothed sandy loam was measured in various states of wetness. Each condition was measured at 3 or 4 reflection angles ranging between 12° and 46·5°. By finding the value of dielectric constant that fitted each set of measurements best (conductivity having negligible influence at the wavelength used), Ford and Oliver obtained values of 2 and 15 for the dielectric constant of very dry and wet ground respectively. Measurements were then repeated with the ground raked into regular ridges, whose heights between crests and troughs were varied in successive experiments between 5 and 16cm. In general the results showed progressively reduced values of ρ below the 'smooth' value as θ and δ (Fig. 2) were increased. An exception to this trend sometimes occurred with vertical polarization when θ was near the Brewster angle, which varied between about 15° and 25° according to the wetness of the ground. In this critical range, roughening the ground disturbed the Brewster effect, thereby sometimes *increasing* in reflection coefficient.

Following Bullington's technique, the author has computed values of ϕ_m associated with 33 of Ford and Oliver's measurements of ρ for prepared rough ground, omitting the above-mentioned anomalous results with vertical polarization. Fig. 4 shows ϕ_m plotted against ρ/ρ_s, where ρ is the measured reflection coefficient of rough ground giving a phase deviation ϕ_m, and ρ_s is the measured reflection coefficient of the same ground when smooth. Further points on Fig. 4 are derived from height-gain tests and ground roughness estimates made by Day and Trolese[18] on 520Mc/s and 1 000Mc/s, and by Gough[19] on 504Mc/s and 9·3cm. The author has added a further point similarly obtained on 200Mc/s. On Fig. 4 a scale of $\delta \sin \theta/\lambda$, (which is simply related to θ_m) has been included because is it the fundamental parameter governing ρ/ρ_s.

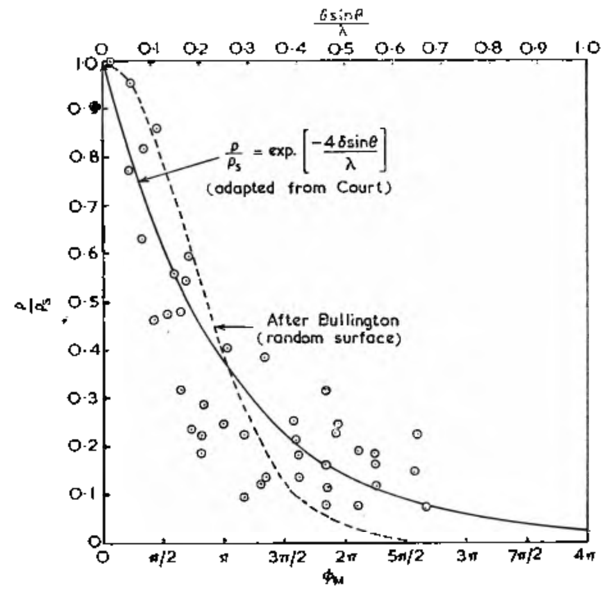


Fig. 4. Effect of ground roughness on reflection coefficient

The points on Fig. 4 scatter less than Bullington's, perhaps because they are derived mostly from artificially regular ground undulations, whereas Bullington's are taken from natural paths. The latter's theoretical curve for a random surface, which is superimposed on Fig. 4 (dotted line) lies below the experimental trend when ϕ_m exceeds about π radians.

Optical measurements made by Court[20] of diffuse reflection of monochromatic light from a matt surface have shown, for fairly shallow reflection angles, an exponential relationship between the reflection coefficient and the reflection angle. This suggests by analogy that the radio reflection coefficient ρ for a rough surface might follow the empirical law:

$$\rho/\rho_s = \exp[-c\,\delta \sin \theta/\lambda]$$

where δ and θ are defined in Fig. 2, ρ_s is the reflection coefficient of a smooth surface of the same material and c is a constant.

$c = 4$ best fits the experimental points of Fig. 4, and the resulting empirical relation

$$\rho/\rho_s = \exp[-4\,\delta \sin \theta/\lambda]$$

gives a better fit than Bullington's curve at large values of ϕ_m. Taking $\rho_s = 1$, which is justified for conventional ground-to-ground paths, the important condition that $\rho < 0{\cdot}5$ requires

$$\delta \sin \theta/\lambda > 0{\cdot}17$$

and $\phi_m > 2{\cdot}14$ radians.

From the aspect of ensuring that ground reflection from a radio path is adequately diffuse, this condition is more stringent than the Rayleigh criteria already quoted.

SEA-REFLECTION FADING

Even with vertical polarization the reflection coefficient of sea-water at shallow incidence approaches unity at very short wavelengths (Fig. 1(b)), and sea paths may thus present special problems of mutual cancellation between direct and sea-reflected waves. Changes in the refractive index gradient of the atmosphere are equivalent from the propagation standpoint to changes in the radius of a fictitious earth at which reflection effectively takes place.

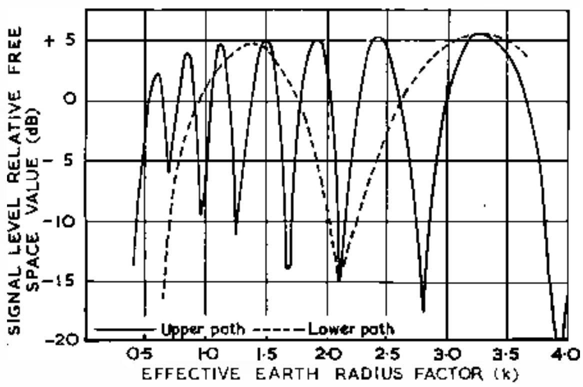


Fig. 5. Theoretical relations between signal level and K on 3930Mc/s for specific oversea paths (after Bray and Corke)

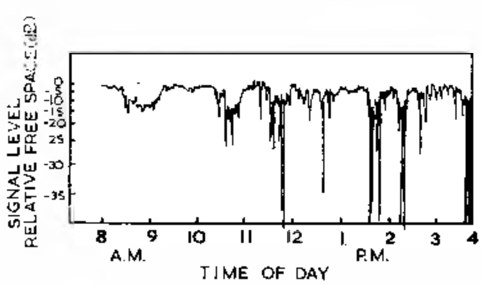


Fig. 6. Water-reflection fading on 9·2cm (after Bogle)

Thus diurnal or weather-related atmospheric changes can cause large phase shifts on microwaves between direct and water-reflected rays, leading to deep fades under conditions of phase opposition. Fig. 5, after Bray and Corke[16], shows theoretical curves relating the effective earth radius factor K to the signal level (relative to free space) expected over two 40-mile paths across the Irish Sea tested on 3 930Mc/s. The paths had a common high terminal in Ireland and respectively high and low terminals in Scotland. A fortnight's recording of October signal-levels showed that the higher path, which was seldom fade-free, faded more than 15dB for 1 per cent of the time, whereas the lower path faded to this extent for only 0·06 per cent of the time. As is evident from Fig. 5, the superiority of the lower path results from its reduced susceptibility to departures of K from the standard value of 4/3.

Several workers comment that the observed fading attributable to this mechanism is often deeper than theoretically expected after allowing for ray divergence at the curved sea surface. Bray and Corke consider that atmospheric inhomogeneities or the state of the sea surface might provide the small change in relative ray amplitudes needed to explain the observed fading. Discussing the same problem, Bogle[21] shows a theoretical expectation of deep fades when K is near infinity, but argues nevertheless that the behaviour of his 9·2cm and 3·26cm oversea records is inconsistent with the occurrence of this K-value. He concludes that direct and sea-reflected waves must be occasionally equalized by some agency which Megaw (writing privately to him), attributed to momentary defocusing of the direct ray by larger-scale refractive index fluctuations caused by turbulence.

Fig. 6 shows a record of fading of this type obtained by Bogle on 9·2cm. Smith-Rose and Stickland[22] have published similar records from line-of-sight transmissions across Cardigan Bay, showing rapid fading of about 3dB on 3cm and slow fading of about 15dB on 9·2cm. Cabessa[23] has observed severe fading on 1 200 and 2 000Mc/s over sea paths in Greece, while Gudmandsen

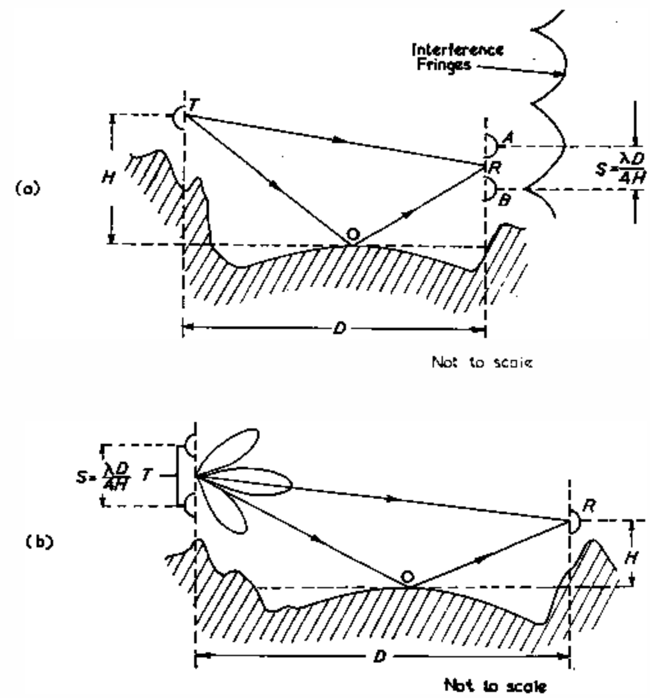


Fig. 7. Complementary diversity system (after Bateman)

and Larsen[24] have experienced similar effects over sea paths in Denmark on 17 and 6·4cm. The two last-named investigations are reviewed in some detail later.

Two-path interference by sea reflection causes an approximately sinusoidal variation of field strength with height, giving vertical interference fringes as illustrated on Fig. 7(a). Near R, the height interval S, between a point of virtually zero signal and the nearest point of maximum signal, is about $\lambda D/4H$, where λ, D, and H are respectively the wavelength, path length and the transmitter height above the tangent to the reflection point O. Bateman[25] has devised a 'complementary diversity' system whereby two receiving aerials A and B (Fig. 7(a)) are vertically spaced by the specified amount, S. When the effective earth radius factor, K, alters, the interference fringes move bodily past the aerial system with little change in the fringe height, S, for paths well within visual range. Thus zero signal from one aerial is always accompanied by a large signal from the other. To exploit this effect fully, Bateman suggests an attractive alternative to the conventional use of separate diversity receivers on each aerial, namely linking the two aerials directly together so that, used for transmission for example, they direct a null towards the reflecting sea at O, and a lobe towards the receiver at R as in Fig. 7(b), thus virtually eliminating the sea reflection. The requisite spacing for this condition is, as might be expected, identical with that already defined for complementary diversity reception. In fact the linked aerial system functions equally as a receiving array, thus abolishing the necessity of a second diversity receiver in this special case, where a coherent amplitude and phase pattern allows suitably spaced aerials to be paralleled to give a nearly constant joint signal while K varies.

Another application of spaced aerials propounded by Bateman is the suppression of injurious multi-path signals propagated by lateral reflections from hills, etc., flanking the radio path. In this case the aerials are spaced an appropriate distance horizontally.

Fresnel Zones

By analogy with classical edge-diffraction theory, Norton and Omberg[17] give general formulae for the dimensions and position of the N^{th} Fresnel zone for reflection at an unbounded flat surface. The reflected wave is the resultant of contributions from an infinite number of half-period (Fresnel) zones on the plane surface, but near-cancellation between adjacent zones (which comprise elliptical rings), results in most of the reflected energy coming from the first zone, which is thus of fundamental importance in evaluating the reflecting ability of surfaces of limited area.

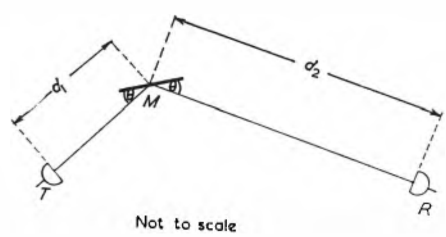


Fig. 8. Use of a flat mirror as a passive reflector

To reflect as well as a limitless plane area, a flat surface must compare in size with the first Fresnel zone. This zone is an ellipse, such that the length of a diffracted ray path via the edge of the zone is half a wavelength longer than the geometrically reflected ray path. For ground-to-ground propagation the ellipse is always elongated in the direction of transmission and its size, which depends on the path geometry, always decreases with the wavelength. Thus the shorter the wavelength the smaller the surface needed for specular reflection.

Passive Reflectors

When intervening obstructions prevent direct microwave communication, a link is sometimes achievable by erecting a suitably angled flat radio mirror on a prominent site visible from each terminal. Andrieux[26] outlines the principle for a flat mirror (Fig. 8) which is far enough from the terminals T and R for the contributions from all points on the mirror to be virtually in phase at the receiver, i.e., the mirror spans only a small portion of the first Fresnel zone.

Andrieux defines the efficiency of the system by the ratio:

$$\frac{\text{Power received via the mirror}}{\text{Power received over a direct path of equal total length in free space}}$$

which amounts to $A_p^2(d_1 + d_2)^2/\lambda^2 d_1^2 d_2^2$,

where A_p is the mirror area projected on the aperture plane of the aerial T or R, d_1 and d_2 are specified on Fig. 8 and λ is the wavelength. For a practicable mirror size, good efficiency requires d_1 or d_2 to be small and the wavelength to be short. Moreover the reflection angle θ must be large enough to avoid an unduly small projected area A_p. This last limitation can be overcome by a 'periscope' system (included in Fig. 9) whereby two mirrors M_1 and M_2 spaced a few feet apart ensure steep reflection angles θ_1 and θ_2 even when the incoming and outgoing rays are parallel.

Chaux and Dascotte[27] describe an elaborate installation in Morocco, shown diagrammatically in Fig. 9, for providing 9·5cm communication without active repeaters between two low-lying points T and R. A single mirror M_3 was erected on a hill which is optical to R but not to T, thus necessitating a second mirror system M_1M_2. The latter was of the 'periscope' type because T, M_1 M_2 and M_3 are nearly collinear. The mirrors, all of 10m² area, were made of flat aluminium sheet. The system is inefficient according to Andrieux's definition, due to unsuitable topography and the rather small mirrors used, but performance was within 6dB of the theoretical figure. Autumn and winter tests showed little fading despite sandstorms, fog and snow.

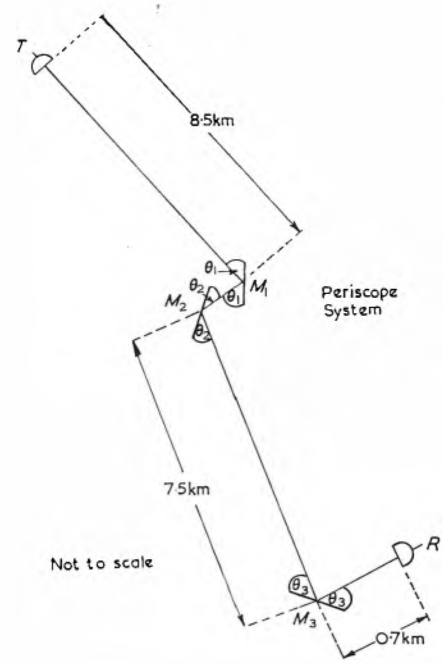


Fig. 9. Passive reflector chain in Morocco (after Chaux and Dascotte)

Magnuski and Koch[28] describe a passive reflector for a 6 700Mc/s link for a hydro-electric scheme in the U.S.A. The single mirror, of area 480ft², was 1¼ miles from the nearer radio terminal. The predicted system performance, closely approached on test, was only 7dB below the free space figure. The mirror required a flatness tolerance of $\pm\frac{1}{8}$in to satisfy Rayleigh's roughness criterion, and an angular tolerance of about $\pm 0{\cdot}13°$ to keep the reflected beam aligned on the receiver.

Zonal Screens

Bussey[29] has devised a means of suppressing microwave ground reflections by erecting, broadside to the direction of propagation, a suitable screen at the geometrical reflection point for flat ground. It can be shown that the reflected field contributed by half the first Fresnel zone (e.g. the unshaded part of the ellipse on Fig. 10) exactly cancels the contribution from the remaining Fresnel zones. Thus if a screen blocks contributions from the other half of the first Fresnel zone (shown shaded on Fig. 10) the reflected wave is sensibly zero. This requirement is met, among other possible shapes of screen suggested by Bussey, by an opaque quadrant of radius equal to the half-width of the first Fresnel zone, located as illustrated in Fig. 10. Experiments showed a very useful reduction in ground reflection, and the system is stated to be tolerant of small siting errors and variations in the effective earth radius factor. For 20 to 30 mile paths on 4 000Mc/s Bussey specifies, without complete dimensional details, a rectangular screen mounted on 40ft poles at mid-path.

Fading Mechanisms

The atmosphere causes intermittent fading of varying degrees over most microwave paths. While absorption due to fog, snow, rain etc., which has been discussed by Ryde[30] and others, is very small below 5 000Mc/s, the redistribution of transmitted energy due to small spatial variations in the atmospheric refractive index can produce large changes in received signal strength.

Several investigators have distinguished between the various situations giving rise to fading illustrated on Fig. 11. Fig. 11(a) relates to sea-reflection fading already discussed, in an atmosphere whose refractive index gradient varies with time. While most prevalent over water paths, this type of fading can also occur over flat land. Cabessa[23]

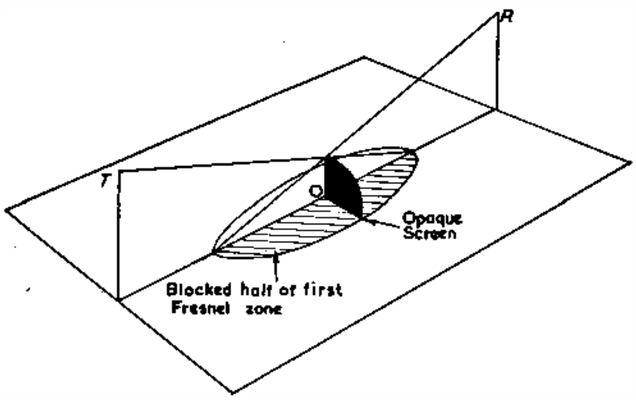


Fig. 10. Suppression of ground reflection by zonal screen (after Bussey)

and Thompson[4] mention the possibility of fading from interfering reflections at an elevated inversion layer (Fig. 11(b)), involving a sharp discontinuity in the vertical refractive index gradient. Millington[31] has shown that such a layer has unity reflection coefficient when the grazing angle is less than a critical value depending only on the total change of refractive index across the layer. Small changes in its height or structure can cause large phase shifts between direct and layer-reflected waves, which can promote deep fading under the conditions described. Above the critical angle, which seldom exceeds $\frac{1}{2}°$, layer reflections and the associated fading decline very sharply.

When an elevated inversion layer separates terminals of widely differing altitudes (Fig. 11(c)), total or near-total reflection at the layer's lower surface may prevent transmission through it to the receiver. The formation and disintegration of this layer, or changes in its height or structure, may promote and later destroy this critical condition, thus again causing fading. Crawford and Jakes[32] emphasize that this fading, not being an interference effect, is not subject to diversity improvement. Rivet[33] points out that it can be long-lasting. A similar effect, associated with trapping within a duct, may occur when the lower terminal lies within the layer. In this case the fading is explainable, with certain reservations, in terms of total internal reflection at the layer's upper boundary.

With shortening wavelength, propagation is increasingly influenced by random irregularities in the atmospheric refractive index, whose existence has been directly revealed in fine detail by microwave refractometer measurements made by Essen[34] in the laboratory and by Bussey and Birnbaum[35] in aircraft. By using 4 000Mc/s pulses of 3mμsec duration over a 28-mile path devoid of ground reflections, de Lange[36] has revealed the simultaneous presence on summer nights of two or three distinct radio paths through the atmosphere (Fig. 11(d)), resulting from the irregularities just discussed. Pulses travelling by these paths had a maximum time separation of 7mμsec, corresponding to a 7ft path difference. It was inferred from fading observations that path differences of only 6in were sometimes present. In this situation a c.w. transmission would be subject to interference fading between the multiple paths.

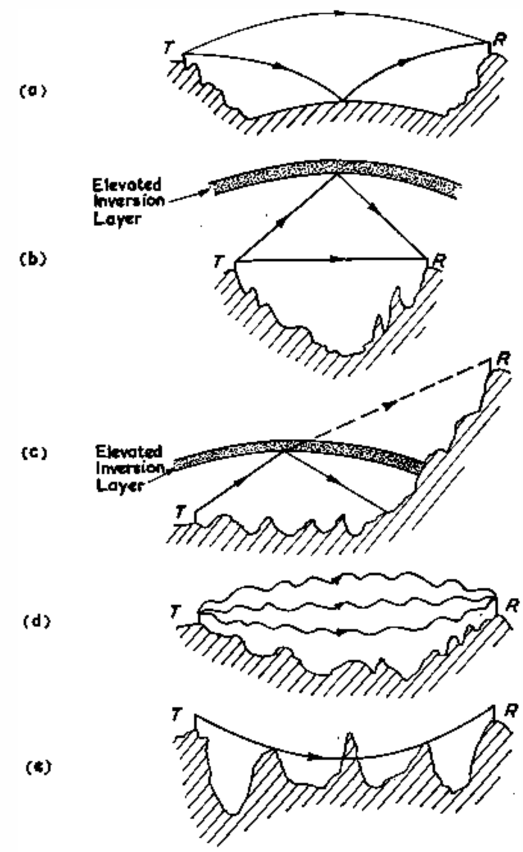


Fig. 11. Situations promoting microwave fading

Several other workers have strikingly illustrated this effect with oscillograph displays of changing microwave signal levels as the transmitted frequency was quickly and linearly varied through 400Mc/s in the 4 000Mc/s band. Fig. 12 shows samples of many traces obtained by Kaylor[37] over a path in Iowa having negligible ground reflections. During summer nights the traces often changed significantly within a few seconds. Mathematical synthesis of the observed waveforms revealed the amplitudes and delay times of the significant signal components, of which the following table is a sample:

Signal amplitude relative to that of first pulse received	1	0·45	0·26	0·10	0·115	0·025	0·02
Delay time, millimicroseconds, or path difference, feet	0	0·122	0·370	2·86	3·14	3·9	12·1

Kaylor draws the following conclusions:

(1) All deep fades are frequency-selective, and associated with a general loss over the band of about 10dB relative to the daytime level, due probably to short-delay signals of substantial amplitude.

(2) All deep fades need 4 to 6 components for adequate synthesis.

(3) Although weaker than short-delay components, long-delay components critically influence the shape of deep fades.

Crawford and Jakes[32] describe similar observations over another path in conjunction with conventional recording of an undeviated 4 195Mc/s transmission. The authors observed level frequency-sweep patterns during daytime when the conventional recording showed a 'free-space' level, and also when signals occasionally exceeded the free-space value by 10 to 15dB (a condition tentatively attributed to trapping or focusing within a surface duct). At night behaviour resembled that found by Kaylor[37], exhibiting several signal components with delays up to 10 mμsec.

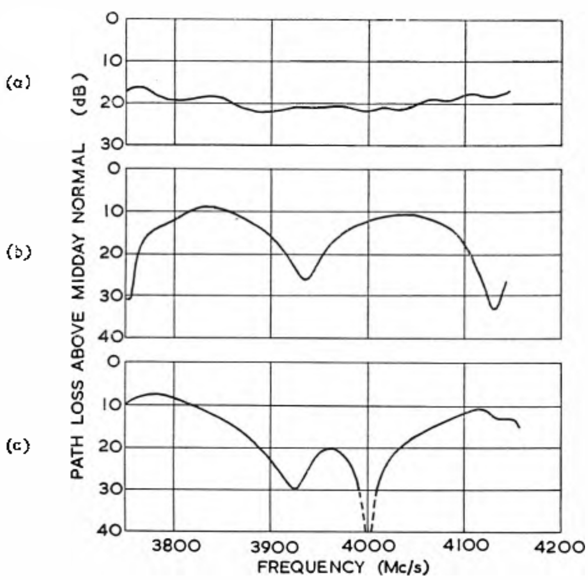


Fig. 12. Typical path loss-frequency curves (after Kaylor)

The authors supplemented these observations with vertical angle-of-arrival measurements on 24 000Mc/s, made by nodding an aerial of 0·12° vertical beamwidth through an angle of 2°. The system could detect changes of 0·05° in vertical arrival angle. During daytime a single ray was usually indicated, while at times of exceptionally strong signal this widened into a solid wedge of rays, (plausibly attributable to focusing) covering a sector of 0·8°. At night three individual rays separated by about 0·25° were sometimes seen, or two rays about 0·5° apart appeared.

The three techniques just described convincingly demonstrate that, under temperate summer conditions, microwave paths may carry three or more signal components with maximum delays of about 10mμsec. De Lange[36] notes that this dispersion limits the ultimate performance of microwave multiplex systems.

Paths with small ground clearance may fade due to a spell of atmospheric sub-refraction causing significant diffraction loss through reversed ray bending (Fig. 11(e)). Over a 37½ mile near-grazing path, Thompson[3] has observed 30dB fades on 4 000Mc/s clearly attributable to this mechanism. He calculates that under these conditions the reversed ray radius of curvature was 5·5 times the earth's radius, equivalent to an effective earth radius factor of only 0·85.

Diurnal and Seasonal Variations

Microwave propagation is profoundly affected by refractive index variations in the lower atmosphere amounting to only a few parts in a million, and thus diurnal, seasonal and weather changes are reflected in propagation behaviour. Many workers have noted the disturbing effects of nocturnal temperature inversions near the ground, caused by cooling of land by radiation on clear windless nights.

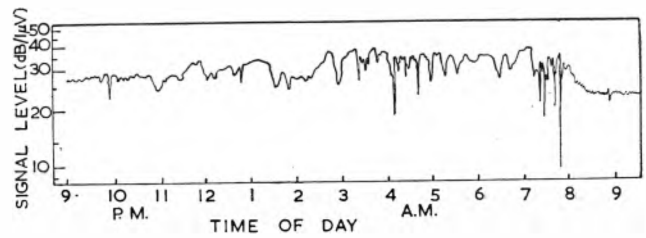


Fig. 13. 9·2cm fading on a summer radiation night (after Megaw)

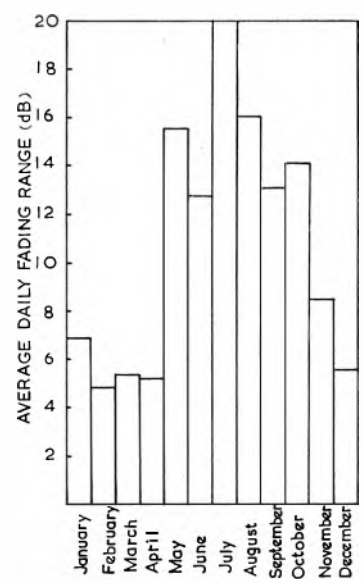


Fig. 14. Seasonal fading variations on 6·5cm (after Durkee)

Particularly during summer, an associated humidity lapse assists the inversion in forming a ground-based or slightly elevated super-refracting layer which can enhance signal strength by reducing diffraction losses, or cause fades from layer reflections at grazing incidence. The latter effect is illustrated on Fig. 13, after Megaw[2], which shows pronounced nocturnal fading on 9·2cm over a 38 mile land path near London. Discussing prolonged tests over this path, Megaw[2] and Smith-Rose and Stickland[22] note the tendency to abnormally high signal strengths on cloudless summer nights, while Bray and Corke[16] mention early morning fading in similar conditions over a neighbouring 28 mile path tested on 3 930Mc/s. Bean[38], discussing 1 046Mc/s tests between a mountain site in Colorado and a low site 70 miles away, notes maximum fading at night between May and September, always associated with low-level irregularities in vertical refractive index gradient measured near the path.

Durkee[9], and subsequently Millar and Byam[39], have made long-term signal recordings over a 42 mile land path near New York, using wavelengths ranging from 42 to 1·25cm. Diurnal variations, pronounced in summer and weak in winter, showed maximum fading in early morning and a stable period round noon, with fading generally increasing in severity with decreasing wavelength. Fig. 14, after Durkee, illustrates the enhanced fading in summer characteristic of all wavelengths and most land paths.

Vecchiacchi[40] and Carassa[41] have noted similar effects during 500 and 1 000Mc/s tests over 'high-low' land paths in Italy, ranging in length between 83·5 and 196km. Fading, which was generally worst in August and least in March, was deeper and more rapid on the higher than on the lower frequency. Fig. 15 from the above authors, which covers three separate years, shows median values of maximum and minimum signal strength for each hour of the day for the whole of August, and for two paths. The charts show the increased variability characteristic of the higher frequency and the longer path. The reduced fading between 1 200 and 1 800 hours results from the minimizing of atmospheric stratification during this period by convection above heated ground. It is noteworthy that 1 000Mc/s variations over the longer path altered very little in two successive years.

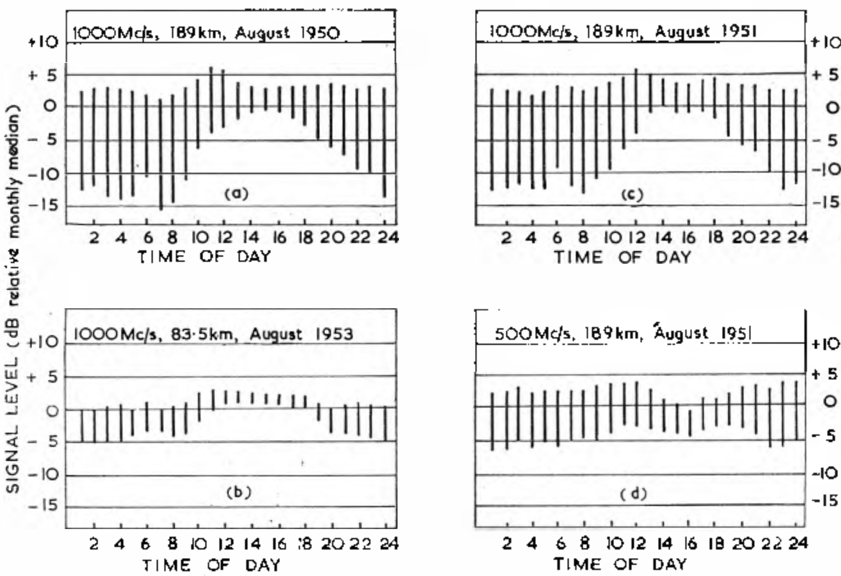


Fig. 15. Summer fading characteristics (after Vecchiacchi and Carassa)

Day and Trolese[18] have investigated diurnal changes over a 27 mile path in the Arizona Desert, using 5 microwave transmissions and numerous aerial heights up to 190ft. Simultaneous meteorological soundings showed the regular recurrence between 1 800 and 0 900 hours of a ground-based super-refracting layer 100 to 200ft thick, which in early morning intensified into a weak surface duct. During the rest of the day refraction was generally slightly sub-standard. Hourly height-gain tests reflected the diurnal atmospheric changes. Between 0 900 and 1 800 hours, regular interference fringes due to ground reflection were evident on 3 300 and 9 375Mc/s, but after 1 500 hours the fringes progressively contracted due to increasing surface refraction. By 1 900 they had become completely incoherent. Random patterns persisted until about 0 900 next morning, when conventional fringes reappeared with the return of normal atmospheric conditions. It is significant that, apart from some signal strength enhancement near the ground, height-gain curves on 1 000 and 520Mc/s were hardly influenced by the nocturnal duct.

For sea paths, particularly when they are well clear of coast lines, diurnal variations may be intermittent or totally absent, for reasons connected with the sea's thermal properties. Gudmandsen and Larsen[24] have occasionally observed oversea diurnal effects on 17 and 6·4cm, with maximum variations in late afternoon. They also observed seasonal variations with maximum variability in spring. A 9·2cm path across Cardigan Bay, discussed by Megaw[2] and Smith-Rose and Stickland[22], exhibited no diurnal effects, and judged by variations in mean daily signal strength appeared free of seasonal variations also.

Weather Influences

Broad generalizations can be made on the effects of weather on microwave propagation. Calm cloudless weather promotes nocturnal signal variations over land paths of the type discussed by Megaw[2], Smith-Rose and Stickland[22], Bray and Corke[16], and Bean[38]. Vecchiacchi[40] and Carassa[41] have observed long periods of stable signal in rainy weather. Millar and Byam[39] noted low signal strength on 9 550Mc/s, connected with prolonged thick fog, while lower frequencies were unaffected. Smith-Rose and Stickland have noted both low and high signal strengths associated with fog on separate occasions. Disturbances from frontal conditions have been observed by the above and by Mismé[42], who also treats the effect theoretically. Tests described by Klein[43] over a 106km path in the Swiss Alps illustrates the remarkable absence of fading peculiar to high altitude paths. Signal variations in June on 15cm amounted to only ± 6dB, a fact attributed by Klein to the prevention of atmospheric stratification by vertical air currents in Alpine valleys, and to the low humidity.

A theoretical investigation by Ryde[30] has shown that while the attenuation due to all kinds of precipitation and fog increases rapidly with frequency, it is negligible below 5 000Mc/s. At 10 000Mc/s, rain covering a 30 mile path might cause an attenuation of about 5dB.

Statistical Aspects

Random processes inherent in microwave propagation through an inhomogeneous atmosphere necessitate a statistical approach to many problems. Many useful methods of statistical analysis, which cannot all be discussed here, have been employed by Gudmandsen and Larsen in an exhaustive investigation of oversea microwave propagation in Denmark, and by other authors already mentioned. It must suffice here to discuss a few points of particular statistical interest and importance.

Statistical Properties of Fades

Investigation of the statistical character of fading is becoming increasingly important, not only in connexion with signal reliability and its improvement by diversity systems, but also in relation to the speed of fading, which influences the phase stability of the received signal. With the rapid development of f.m. multiplex systems carrying progressively more channels, crosstalk and noise introduced by fluctuations in the phase of the received signal may become significant.

Gudmandsen and Larsen[24], Wheeler and Mathwich[10] and others have observed that the durations of microwave fades below a given level have an approximately log-normal distribution, so long as the path has sufficient clearance to avoid diffraction losses during subrefraction periods. The average duration of fades, which is of considerable interest, decreases with the reference level. Fig. 16, after Gudmandsen and Larsen, shows observed relations between the reference level and the average fade duration below this level for oversea transmissions on 17cm and 6·4cm during a 24 hour period. The curve clearly indicates the more rapid fading on the shorter wavelength for all but very high levels.

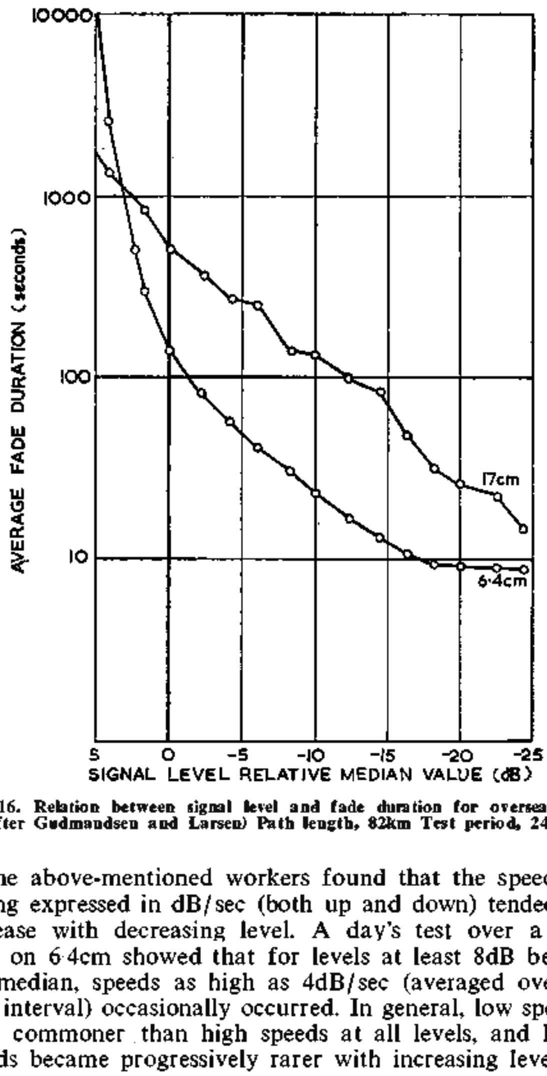


Fig. 16. Relation between signal level and fade duration for oversea path (after Gudmandsen and Larsen) Path length, 82km Test period, 24hr

The above-mentioned workers found that the speed of fading expressed in dB/sec (both up and down) tended to increase with decreasing level. A day's test over a sea path on 6·4cm showed that for levels at least 8dB below the median, speeds as high as 4dB/sec (averaged over a 5sec interval) occasionally occurred. In general, low speeds were commoner than high speeds at all levels, and high speeds became progressively rarer with increasing level.

The Inference of Propagation Mechanisms from Fading Statistics

Fig. 17 is a cumulative distribution derived from a fortnight's measurements by Bray and Corke[16] of a 3 930Mc/s sea path already discussed. Assuming propagation confined to direct and water-reflected paths, the expected behaviour in terms of the effective earth radius factor, K, is described by the full line (upper path) on Fig. 5. Because of the multiple nulls covered by the probable range of K, the phase difference between ray paths is virtually random when K varies. Changes in K alter the divergence factor at the sea surface, and thus the received signal is the resultant of two components in virtually random phase whose relative amplitudes are related to K in calculable way. Assuming a certain probability distribution of K-values, a theoretical signal strength distribution can be calculated for comparison with the experimental curve. The dotted line on Fig. 17 shows such a curve, based on statistics of K-values derived by Gough[44] from oversea v.h.f. tests in the mediterranean, where it appeared that K was approximately Gaussianly distributed. Despite the possible inappropriateness of this distribution to the present path, agreement is good for levels exceeded for less than 95 per cent of the time. The progressive divergence above 95 per cent is associated with the unexpectedly deep fading noted by various workers. The discrepancy might be explained by a third small randomly phased component (itself possibly the resultant of many others) caused for example by turbulence or weak reflection from an inversion layer.

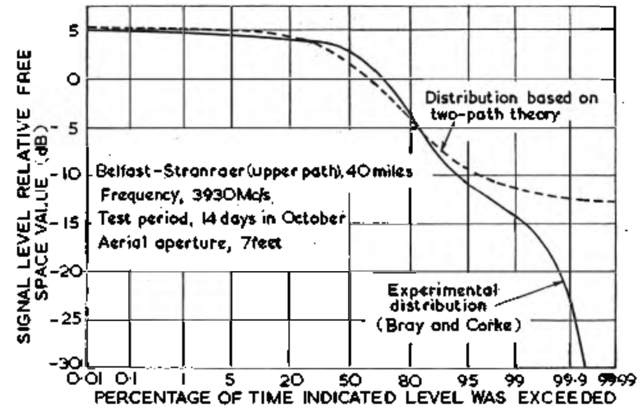


Fig. 17. Cumulative distribution for oversea path (after Bray and Corke)

Signals comprising many components having various amplitudes in random phase should show a Rayleigh distribution when the test period avoids long-term influences such as sub-refraction. Gudmandsen and Larsen[24], Cabessa[23]. and Kaylor[37] give many distributions for periods of an hour and upwards which approximate well to Rayleigh's law. Long-term measurements of sea paths made by the first three investigators for periods up to five months show fading close to the Rayleigh value and median levels often significantly below free space. The latter effect may derive from reduction in aerial gain due to a non-uniform field across the aerial aperture in the presence of multiple ray paths. The 'single aerial' curve on Fig. 18, after Cabessa, shows a distribution derived from a 5 months' oversea test on 2 000Mc/s, with Rayleigh's distribution (curve (a)) added for comparison. The path in question has sufficient clearance to preclude the possibility of diffraction losses during sub-refraction periods, which may partly explain its long-term conformity with Rayleigh's law.

It is noteworthy that while the long-term distributions of Cabessa and Gudmandsen and Larsen indicate the general presence of multiple ray paths, Bray and Corke's oversea test (Fig. 17) points to dominantly two-path propagation, for reasons possibly connected with the persistence of a well-mixed atmosphere in the latter case.

The statistical investigation of fading mechanisms is complicated by field variations across the aerial due to multi-path effects. Increasing the aerial aperture increases the chance, at any instant, of a usable field somewhere in the aperture, with consequent reduction in fading. Cabessa, using small aerials over a sea path, measured a fading range of 37dB on 1 200Mc/s, while simultaneous measurements on 2 000Mc/s over the same path with 10ft paraboloid aerials showed fades of only 25dB.

Table 1 lists representative observed fading characteristics in temperate regions under the conditions specified. The fading depth, expressed relative to the median level, is here defined as the depth exceeded for 0·1 per cent of the time. Rayleigh fading on this basis amounts to 28dB.

Diversity Effects

As is well known, fades due to multi-path transmission are not always synchronized when separately received at two points spaced a sufficient distance apart. A like effect occurs when two transmissions over a common radio path differ sufficiently in frequency. In the frequency band under discussion, space and frequency diversity systems exploiting this principle can provide a useful increase in circuit reliability. The subject has been treated theoretically by Jelonek, Fitch and Chalk[45], with illustrative statistics derived from microwave measurements made in Wales.

For maximum diversity advantage, fading on one channel must always be accompanied by high signal strength on the other, such that the correlation between signals is -1. This condition, termed by Bateman[25] 'Complementary Diversity', can only exist with coherent two-path propagation, e.g., the situation approached in Bray and Corke's oversea test (Fig. 17). For two randomly fading signals (for example of the commonly experienced Rayleigh-distributed type), correlation can be zero but not negative. Experimental observation of negative correlation indicates either the existence of two coherent waves or the presence of sampling errors due to finite measuring time.

Only the important case of two independently fading signals (between which the correlation is zero) can be considered here. If each signal falls below the usable threshold level for a fraction p of the time, they are both unusable together for only p^2 of the time, which may represent a large improvement in reliability. Even when, as often occurs, some correlation between channels remains, a valuable diversity gain may still be obtainable.

The dotted curves on Fig. 18, given both by Cabessa[33] and Gudmandsen and Larsen[34], show (a) the conventional Rayleigh distribution and (b) the diversity Rayleigh distribution based on the selection of the strongest of two independently fading Rayleigh signals. The full-line curves, after Cabessa, show distributions for a 90km sea path obtained respectively with single-aerial reception (already discussed) and with diversity reception using 5·75m (vertical) aerial spacing. This diversity performance approximates well to theoretical expectation, indicating satisfactory independence between the two signals. Cabessa considers that for best results over short paths, the diversity spacing should correspond to the distance between 'bright' and 'dark' interference fringes for reflection at a flat earth. In the present case his spacing somewhat exceeds this figure, but this is of small consequence because Rayleigh-type rather than two-path propagation is dominant, so that any spacing exceeding a certain minimum will suffice.

For a 120-km sea path, which Cabessa considers to be the limiting distance for a high quality microwave link, fading with appropriate space diversity was 2dB worse, on a 0·1 per cent basis, than the diversity Rayleigh value.

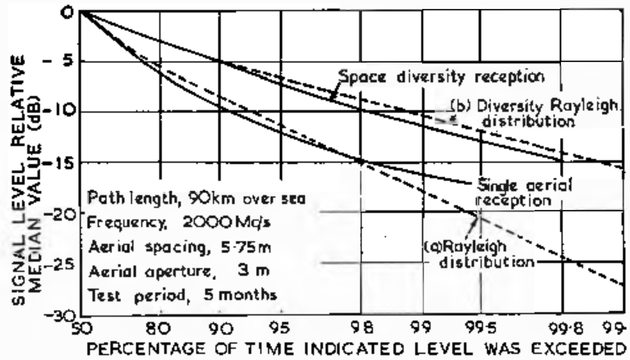


Fig. 18. Effect of space diversity on oversea path (after Cabessa)

TABLE 1

Representative Microwave Fading Depths

INVESTIGATOR	PATH LENGTH (Miles)	(Km.)	DESCRIPTION	WAVELENGTH OR FREQUENCY	TEST PERIOD	FADING DEPTH EXCEEDED FOR 0·1% OF THE TIME (dB)
Bray and Corke[16]	28	45	Overland, near London	3930Mc/s	0200-0300 hours throughout May*	36
				3930Mc/s	(All hours in May except 0100-0400)	13
Kaylor[37]	31	50	Overland, Iowa, U.S.	4190Mc/s	July and August	27
Millar and Byam[39]	42	68	Overland, near New York	16·2cm	Jan.-March inclusive	6
					June-Oct. inclusive†	17
				7·2cm	Jan.-March inclusive	10
					June-Oct. inclusive†	18
				4·7cm	Jan.-March inclusive	12
					June-Oct. inclusive†	21
				3·1cm	Jan.-March inclusive	22
					June-Oct. inclusive†	24
Vecchiacchi[40]	117	189	Overland, N. Italy	1000Mc/s	August†	27 (extrapolated)
Carassa[41]	117	189	Overland, N. Italy	500Mc/s	August†	20 (extrapolated)
Bray and Corke[16]	40	65	Upper oversea path, Irish Sea	3930Mc/s	Single hour of worst fading in May‡	40
				3930Mc/s	Fortnight in October‡	25
Cabessa[33]	56	90	Oversea, Greece	2000Mc/s	Jan.-June inclusive	24 (extrapolated)
Gudmandsen and Larsen[34]	51	82	Oversea, Denmark	17cm	August-Sept. inclusive	24
	51	82		6·4cm	August-Sept. inclusive	20
	33½	54		17cm	April-July inclusive	26
	33½	54		6·4cm	10 days in May	29

* *Diurnal period of worst fading.* † *Season of worst fading.* ‡ *No diurnal variations.*

Fig. 19 shows a 4 hour simultaneous record of diversity signals over this path. The resultant fading is defined by the upper boundary of the chart. While major fades on the two channels are mostly out of step, there is occasional synchronization, as for example at *S*. This is of course to be expected on statistical grounds when Rayleigh's law operates. With coherent two-path propagation diversity reception could theoretically be made virtually fade-free, and any lapse would have to be explained by some agency which attenuated each ray path about equally, e.g., fading of types (c) and (e) on Fig. 11.

Gudmandsen and Larsen have made oversea diversity tests on 17 and 6·4cm, using aerials vertically spaced at such a distance as to avoid synchronized fading on a two-path basis with the range of atmospheric refraction likely to occur. Like Cabessa, they found good agreement with the diversity Rayleigh law for test periods of about a month. For single days, however, diversity gave sometimes less and sometimes more improvement than expected from this law. The occasional betterment of expectation suggests the spasmodic occurrence of coherent two-path propagation, as opposed to the apparently predominating multi-path propagation.

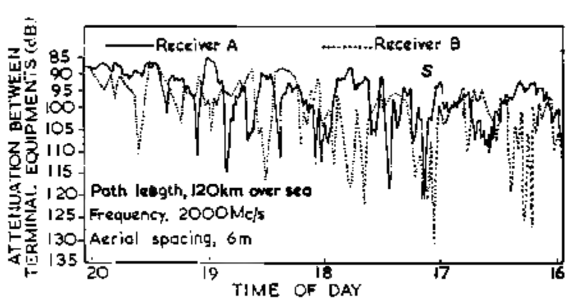


Fig. 19. Space diversity reception over a sea path (after Cabessa)

With the aim of mitigating fading by frequency diversity, Kaylor[37] has measured the frequency separation between two microwave transmissions over a common path required to avoid correlation between fades on the respective channels. He found that the requisite frequency separation increased with the depth of fading encountered. For example, in the 4 000Mc/s band, fading of 10dB required a separation of 40Mc/s to avoid correlation, while fading of 20dB required a separation of 160Mc/s.

Compared with long established h.f. ionospheric techniques, and with the scatter systems now rapidly evolving, the exploitation of diversity principles on conventional microwave links is still in its infancy. Under the stimulus of demands for increased information-carrying capacity with high circuit quality, coupled with the economic advantage of extended ranges, it is likely that this field will see much development in the near future.

Acknowledgments

The author wishes to thank the Engineer-in-Chief, Marconi's Wireless Telegraph Company Limited, for permission to publish this article. He is further indebted to Mr. G. Millington for the Introduction to the article, and to the last named and Mr. G. A. Isted for helpful suggestions. The assistance of Mr. F. Immirzi with statistical computations is also acknowledged.

The author acknowledges the use of illustrations from the following publications: The Journal and the Proceedings of the Institution of Electrical Engineers; The Proceedings of the Institute of Radio Engineers; The Bell System Technical Journal; L'Onde Electrique; Annales de Radioélectricité; Alta Frequenza; Acta Polytechnica; Electronics; Tele-Tech.

REFERENCES

1. Friis, H. T. A Note on a Simple Transmission Formula. *Proc. Inst. Radio Engrs. 34*, 254 (1946).
2. Megaw, E. C. S. Experimental Studies of the Propagation of Very Short Radio Waves. *J. Instn. Elect. Engrs. 93*, 79 Pt. 3A (1946).
3. Bullington, K. Radio Propagation at Frequencies Above 30Mc/s. *Proc. Inst. Radio Engrs. 35*, 1122 (1947).
4. Thompson, L. E. Microwave Propagation Experiments. *Proc. Inst. Radio Engrs. 36*, 671 (1948).
5. Eckersley, T. L. Ultra-Short-Wave Refraction and Diffraction. *J. Instn. Elect. Engrs. 80*, 286 (1937).
6. Millington, G. The Diffraction of Wireless Waves Round the Earth. *Phil. Mag. 27*, 517 (1939).
7. Van der Pol, B., Bremmer, H. The Propagation of Radio Waves Over a Finitely Conducting Spherical Earth. *Phil. Mag. 25*, 817 (1938).
8. Burrows, C. R., Gray, M. C. The Effect of the Earth's Curvature on Ground-Wave Propagation. *Proc. Inst. Radio Engrs. 29*, 16 (1941).
9. Durkee, A. L. Results of Microwave Propagation Tests on a 40-mile Overland Path. *Proc. Inst. Radio Engrs. 36*, 197 (1948).
10. Wheeler, B. F., Mathwich, H. R. Use of Distribution Curves in Evaluating Microwave Path Clearance. *Proc. Inst. Radio Engrs. Trans. on Instrumentation*, 31 (Oct. 1955).
11. Bullington, K. Reflection Coefficients of Irregular Terrain. *Proc. Inst. Radio Engrs. 42*, 1258 (1954).
12. McPetrie, J. S. The Reflection Coefficients of the Earth's Surface for Radio Waves. *J. Instn. Elect. Engrs. 82*, 214 (1938).
13. Kerr, D. E. Propagation of Very Short Waves—Pt. 2. *Electronics 21*, 118 (Feb. 1948).
14. Ohman, G. P. Universal Curves for the Vertical Polarization Reflection Coefficient. *I.R.E. Trans. on Antennas and Propagation*, p. 140 (Jan. 1957).
15. Ford, L. H., Oliver, R. An Experimental Investigation of the Reflection and Absorption of Radiation of 9-cm. Wavelength. *Proc. Phys. Soc., Lond. 58*, 265 (1946).
16. Bray, W. J., Corke, R. L. A Technique for 4000-Mc/s Propagation Testing for Radio-Relay Systems. *Proc. Instn. Elect. Engrs. 99*, Pt. 3A, 281 (1952).
17. Norton, K. A., Omberg, A. C. The Maximum Range of a Radar Set. *Proc. Inst. Radio Engrs. 35*, 4 (1947).
18. Day, J. P., Trolese, L. G. Propagation of Short Radio Waves Over Desert Terrain. *Proc. Instn. Radio Engrs. 38*, 165 (1950).
19. Gough, M. W. V.H.F. and U.H.F. Propagation Within the Optical Range. *Marconi Rev. 12*, 121 (1949).
20. Court, G. W. G. Determination of the Reflection Coefficient of the Sea for Radar Coverage Calculation by an Optical Analogy Method. *Proc. Instn. Elect. Engrs. 102B*, 827 (1955).
21. Bogle, A. G. Some Aspects of Microwave Fading on an Optical Path Over Sea. *Proc. Instn. Elect. Engrs. 99*, Pt. 3, 236 (1952).
22. Smith-Rose, R. L., Stickland, A. C. An Experimental Study of the Effect of Meteorological Conditions Upon the Propagation of Centrimetric Radio Waves. *Phys. Soc. and Ryl. Met. Soc. Report of a Conference on Meteorological Factors in Radio-Wave Propagation*, p. 18 (1946).
23. Cabessa, R. Réalisation de Liaisons par Faisceaux Hertziens de Qualité sur les Trajets Maritimes en Grèce. *L'onde Electrique 35*, 714 (1955).
24. Gudmandsen, P., Larsen, B. F. Statistical Data for Microwave Propagation Measurements on Two Oversea Paths in Denmark. Acta Polytech. No. 213 (1957)
25. Bateman, R. Elimination of Interference Type Fading at Microwave Frequencies with Spaced Antennas. *Proc. Inst. Radio Engrs. 34*, 662 (1946).
26. Andrieux, G. Réflecteurs Passifs pour Faisceaux Hertziens. *L'Onde Electrique 36*, 57 (1956).
27. Chaux, R., Dascotte, J. Les Relais Passifs de la Liaison Afourer-Bin-el-Ouidane. *Ann. Radioélectricité 5*, 220 (1950).
28. Magnuski, H., Koch, T. F. Passive Repeater Bends Microwave Beam. *Electronics 26*, 134 (Feb. 1953).
29. Bussey, H. E. Suppressing Microwaves by Zonal Screens. *Tele-Tech. 10*, 45 (Dec. 1951).
30. Ryde, J. W. The Attenuation and Radar Echoes Produced at Centimetre Wave-Lengths by Various Meteorological Phenomena. *Phys. Soc. and Royal Met. Soc. Report of a Conference on Meteorological Factors in Radio-Wave Propagation*, p. 169 (1946).
31. Millington, G. The Reflection Coefficient of a Linearly Graded Layer. *Marconi Rev. 12*, 140 (1949).
32. Crawford, A. B., Jakes, W. C. Selective Fading of Microwaves. *Bell Syst. Tech. J. 31*, 68 (1952).
33. Rivet, P. Essais de Diversité et Etude de L'Effet de Focalisation sur les Liaisons Longues en Visibilité. *L'Onde Electrique 36*, 24 (1956).
34. Essen, L. A Highly Stable Microwave Oscillator and Its Application to the Measurement of the Spatial Variations of Refractive Index in the Atmosphere. *Proc. Instn. Elect. Engrs. 100*, Pt. 3, 19 (1953).
35. Bussey, H. E., Birnbaum, G. Measurements of Variations in Atmospheric Refractive Index with an Airborne Microwave Refractometer. *J. Res. Nat. Bur. Stand., Wash. 51*, 171 (1953).
36. De Lange, O. E. Propagation Studies at Microwave Frequencies by Means of Very Short Pulses. *Bell Syst. Tech. J. 31*, 91 (1952).
37. Kaylor, R. L. A Statistical Study of Selective Fading of Super-High-Frequency Radio Signals. *Bell Syst. Tech. J. 32*, 1187 (1953).
38. Bean, B. R. Prolonged Space-Wave Fadeouts at 1046 MC Observed in Cheyenne Mountain Propagation Program. *Proc. Inst. Radio Engrs. 42*, 848 (1954).
39. Millar, J. Z., Byam, L. A. A Microwave Propagation Test. *Proc Inst. Radio Engrs. 38*, 619 (1950).
40. Vecchiacchi, F. Prove di Propagazione con le Frequenza di 1000 MHz su Vari Percorsi. *Alta Frequenza 25*, 100 (1956).
41. Carassa, F. Prove di Propagazione con le Frequenze di 250, 500, 1000 MHz. *Alta Frequenza 25*, 378 (1956).
42. Misme, P. Fadings d'Intérferences Causés par des Discontinuités Frontales. *L'Onde Electrique 36*, 43 (1956).
43. Klein, W. Wave Propagation Studies in the Alps, over 15-cm, 90-cm, and 2-m Wave Radio Relays. *Brown Boveri Rev. 36*, 387 (1949).
44. Gough, M. W. Some Features of V.H.F. Tropospheric Propagation. *Proc. Instn. Elect. Engrs. 102B*, 43 (1955).
45. Jelonek, Z., Fitch, E., Chalk, J. H. H. Diversity Reception. *Wireless Engr. 24*, 54 (1947).

Tropospheric Scatter Propagation

By G. Millington*, M.A., B.Sc., M.I.E.E.

The physical mechanisms of long-distance propagation are reviewed with special reference to forward scattering. Recent research in the United Kingdom is described on the nature of the signal variations and their relation to meteorological conditions. The use of scatter links for wide-band systems is discussed, including the problems of aircraft reflections and mutual interference due to frequency sharing.

THIS article is a companion to the preceding one by M. W. Gough on 'Microwave line-of-sight propagation'. The opening section of that article forms a general introduction to propagation in the u.h.f. and higher frequency bands and reference should be made to it in connexion with the present article also, particularly with regard to the division of the subject between the two articles.

It may appear that the trans-horizon region in which there is a transition to the diffraction field has been neglected in passing from line-of-sight to scatter propagation, but it should be borne in mind that at the present time there is no general agreement about the physical mechanism of long distance tropospheric propagation. It is interesting to note that at a recent Symposium at the Institution of Electrical Engineers on 'Long-distance propagation above 30Mc/s'[1] the session on ionospheric propagation was entitled 'Ionospheric forward scatter propagation' whereas the sessions on tropospheric propagation were qualified by 'beyond the horizon' without specific reference to forward scatter.

The subject of long distance tropospheric propagation has been intensively studied in recent years from two different points of view, (1) as a disturbing factor limiting the amount of frequency sharing that can be achieved in relatively short-distance applications such as optical relay links and television broadcasting, and (2) as a method of establishing long-distance communication links of several hundred miles by using large powers and high-gain aerials. The literature is already extensive and only a few selected references to survey papers and conference proceedings will be given here[2,3,4,5] which themselves contain useful bibliographies.

No attempt will be made to give a detailed review of all this work. Instead, the aim will be to assess the present position with regard to the successful development of tropospheric scatter links for long-distance communication on a world-wide basis. In the U.S.A. work has gone ahead on the establishment of operational links, whereas in the United Kingdom attention has so far been largely concentrated on the better understanding of the propagation phenomena involved, as is shown from the papers contributed to the symposium already mentioned[1]. The ionospheric research described at that symposium has revealed that the claims originally made for the general use of forward scatter circuits need careful revision, and it is found that in the tropospheric case also much thought needs to be given to the factors which may control the use of long-distance tropospheric communications at any given time and place.

Factors in Long-distance Tropospheric Propagation

It is well known that the possibility of receiving signals far beyond the horizon on quasi-optical frequencies depends upon the existence of a non-homogeneous atmosphere. It is true that communication can be established over considerable distances by the mechanism of obstacle gain when there is a high mountain ridge between the transmitter and the receiver that is well optical from both terminals[6], but this remarkable phenomenon can only be exploited in certain cases and does not enter into the consideration of long-distance transmission in general. It has also been suggested by Bullington[7] that energy may be conveyed into the diffraction region by scattering from the rough surface of the ground, but though this may be a contributory factor it is probably not a major one.

If the non-homogeneity of the atmosphere had the form of a smooth variation of refractive index in the vertical direction such as a linear variation or one decreasing exponentially from the surface value to unity at great heights, the situation would lead to propagation over the surface of the earth characteristic of that through a homogeneous atmosphere above an earth of modified radius. Under the conditions of a standard atmosphere the effective radius would be the true radius modified by a factor of 4/3. This would give an increased horizon distance, but no major increase of range beyond the horizon.

When there is a large negative gradient of refractive index at the surface of the earth, super-refraction occurs and a radio duct can be formed leading to anomalous propagation in which strong signals can be obtained at great distances. This aspect of long-distance propagation will not be considered in detail here, though it is an important factor, especially in tropical regions, in the disturbance caused for certain small percentages of time to services working on a frequency sharing basis.

It has been maintained by Carroll and Ring[8] that partial reflections at all levels within a smoothly graded atmosphere are of themselves sufficient to explain propagation to long distances, but it appears from their analysis that this conclusion is a result of the model they have used, namely a bi-linear distribution of refractive index in which at a certain height there is a discontinuity in the gradient. At the shallow angles of propagation involved, such a discontinuity can give rise to appreciable reflection and thus profoundly modify the attenuation of the propagation modes in the diffraction region.

This mechanism does, however, draw attention to the general phenomenon of partial reflections from elevated layers in the troposphere where there is a rapid change of refractive index over a small range of height associated with changes in temperature and humidity. It is well known from meteorological soundings by radio-sonde and from refractometer measurements made in aircraft[9] that such layers often occur, and there is now a large body of evidence that the presence of such layers can give rise to long-distance propagation. Under many conditions, partial reflections are a major factor, while some authors[10] even claim that such reflections from high level layers, and in particular from the tropopause, are the sole cause of long-distance tropospheric propagation.

The scattering theory itself is based on the idea that there exists in the troposphere a series of scattering centres

* *Marconi's Wireless Telegraph Co. Ltd.*

or 'blobs' in which the refractive index differs from the mean value for the surroundings. These blobs can scatter energy falling on them in various directions and the polar diagram of the scattered energy depends on the scale size of the 'blob', the wave frequency, the scattering angle between the direction of the incident wave and the scatter direction under consideration, and upon the law of distribution of refractive index assumed.

At the Symposium[1] all these factors affecting long-distance tropospheric propagation came under review and a division of the papers was made into two categories: (1) those dealing with transmissions in which the mechanism was primarily one of reflection from elevated layers, and (2) those in which the propagation was thought to be by forward scattering from 'blobs' formed by turbulence in the atmosphere. The first dealt with the trans-horizon region and with reception at considerable distances for a small percentage of time of high field-strengths capable of causing interference between services sharing the same frequency in well separated areas. The second was concerned mainly with the relatively weak signals receivable continuously at long distances up to several hundred kilometres.

It is interesting to note in retrospect that the first category dealt entirely with transmissions on frequencies below 300Mc/s, i.e. in the v.h.f. range and the second with frequencies above 300Mc/s, i.e. in the u.h.f. and s.h.f. ranges. It is in these higher frequency bands that work is going ahead with the aid of large powers and high-gain aerials to use the weak but continuous signals for long-distance communication by tropospheric propagation. The remainder of this article will be concerned with this aspect of the subject and in particular with the work being carried out in the United Kingdom.

Long-Distance Forward Scatter Propagation

Evidence of the propagation of microwaves well beyond the horizon has been accumulating for more than a quarter of a century and it has been pointed out that Marconi was among the pioneers in studying this part of the radio spectrum[11,12]. It was not, however, until after the development and use of microwave radar that Pekeris[13] in 1947 first put forward the suggestion that a scattering process might be involved. In 1949, Megaw[14] carried out a pioneer experiment establishing beyond doubt that at a certain distance beyond the horizon the exponential decrease of field strength in the diffraction region gives place to a much slower decrease with distance, so that a detectable signal persists out to great distances.

The signals are not constant, in fact it was the nature of the fading to which they are subject that led to the scattering theory which was first worked out in detail by Booker and Gordon[15] and subsequently on a somewhat different basis by Villars and Weisskopf[16] and by Batchelor[17]. Megaw's own independent theoretical work has recently been published posthumously[18]. The subject of turbulence and the formation of eddies is complex and no attempt will be made here to compare the various scattering formulae that have been derived, especially as it is difficult to decide between them from the experimental results. They have, however, been reviewed by Johnson[19] who points out the danger of attributing differences between theory and observation to wrong assumptions about the scattering parameters when there may be the unsuspected presence of radio ducts or elevated inversion layers modifying the propagation process.

The explanation of the observed results in terms of scattering from atmospheric turbulences is based on the condition that the fading of the signals is random in character, though an equivalent effect could be produced by the addition of a number of coherent signal components in random relative phases arising, for example, from partial reflections from a number of discontinuities of limited extent. Actually the signal is often the resultant of a random component and an intermittently coherent component, even at considerable distances and when the field-strength is only of the order associated with the scattering process. Such a signal has been termed 'quasi-scatter[20,21] and is sometimes revealed in a statistical analysis of a fading record by the departure from a Rayleigh law typical of a random signal. Even at 10 000Mc/s there is evidence that partial reflections may play some part as well as scattering in the propagation mechanism[21]. In analysing records obtained from tropospheric scatter links to determine the characteristics of forward scattering by atmospheric turbulences, care has to be taken to exclude cases where propagation in a radio duct or by partial reflections from elevated layers is revealed by an unusually high signal level.

Recent Work in the United Kingdom

In reviewing the work in progress in the United Kingdom, attention will be mainly directed towards the general results that have been obtained. With the exception of the experiments that have been conducted by the Admiralty[21,22] in which measurements have been made at sea out to ranges as great as 700km, the tests have been made over fixed distances wholly or mainly overland. These distances lie in the range 150km to 400km, and the frequencies in the range 850Mc/s to 3 500Mc/s, apart from one of the Admiralty tests carried out on about 10 000Mc/s.

Although some tests have been made with speech modulation, the transmissions have been on continuous wave to obtain amplitude recordings for studying the various propagation characteristics upon which the successful use of communication circuits depends. These will be dealt with briefly in turn.

Attenuation Loss Relative to the Free-space Signal

In agreement with the scattering theory as applied to the kind of turbulence that is believed to occur in the troposphere, the increase of the attenuation or transmission loss with frequency between 300Mc/s and 3 000Mc/s is not great, though at 10 000Mc/s the effect of atmospheric absorption is becoming important. The frequency dependence is in fact much less severe than in the case of ionospheric forward scatter propagation. In the region beyond the horizon where the permanent scattered signal first predominates over the diffracted signal, it is of the order of 50dB below the free space value, but the attenuation with distance is only 0·1dB/km increasing to 0·2dB/km at 10 000Mc/s.

Even on 300Mc/s at the lower end of the u.h.f. band, the rate of attenuation of the diffracted field under standard atmospheric conditions is 0·65dB/km and the field falls below the level of the scattered signal at about 60km beyond the horizon. At 3 000Mc/s this attenuation rate is 1·4dB/km and the scattered signal takes over from the diffracted field within about 30km of the horizon.

Diurnal and Seasonal Variations

The above figures for the scattered field are of course approximate as they are subject to diurnal and seasonal variations. In broad terms the signals are greater by night than by day with an overall variation of the order of 10dB, while they are from 5 to 10dB weaker in the winter months than in the summer. These conclusions depend somewhat upon the method of averaging, according, for instance, to whether daily or hourly median values are taken, and upon the way in which allowance is made for non-standard conditions typified by ducting or the presence of inversion layers.

These conditions are more likely to occur by night than by day and may have an influence on the night-time estimates. There is some evidence that the signal is most likely to be a purely scattered one in the early afternoon with the suggestion that the seasonal variation should be based on measurements made between 1 200 to 1 400 hours[23].

Fading Characteristics

An indication that the signal is effectively a scattered one is obtained when the fading is rapid and the amplitude exhibits a Rayleigh distribution. Under these conditions a typical fading rate is one per second between limits of $\pm$10dB relative to the median value. This fast fading is sometimes superimposed on a more slowly varying signal with a rate of only a few per minute but with a fading depth of 20dB or more, implying the presence of another mechanism of propagation.

The Rayleigh distribution strictly refers to measurements made over a short interval of time. Variations in the troposphere which cause changes in the scattering coefficient associated with the turbulences make the distribution measured over longer intervals tend towards the log-normal type. A modified Rayleigh distribution has been found in those cases where a coherent mode of propagation is presumed to be present under quasi-scatter conditions[2,24].

Beam Setting Requirements

Tests have been carried out to find the optimum aerial arrangements with regard to beam gain and setting. On the basis of a presumed scattering mechanism, theory indicates that at the scattering source the angle between the direction from the transmitter and the direction to the receiver should be kept as small as possible. For an assumed height of the source this scattering angle is least for the mid-way point in the great-circle plane through the transmitter and receiver.

It is in fact found necessary to keep the angle to a minimum by directing the beams at the horizon. It is therefore important to choose a site with an unobstructed horizon and indeed it is beneficial to use high terminals where possible and to direct the beams downwards to the horizon. Measurements show that even on 860Mc/s at a distance of 160km the received signal drops by about 14dB if the transmitting and receiving beams are elevated together by as little as 3°.

In the horizontal plane over the same circuit the signal dropped by 15dB when the beams were moved together through 3° from the great-circle plane. In such a case the drop in signal may not be solely due to the consequent increase in the scattering angle, but also to the failure of the beams to intersect at a sufficiently low level in the troposphere. It appears in fact that with increasing height the intrinsic scattering coefficient decreases quite apart from the value of the scattering angle. There is thus a limited scattering volume which is effective in conveying the signal from a transmitter to a receiver at a given distance.

It is thus essential to use a beam that is as narrow as possible in the vertical plane and narrow enough in the horizontal plane to lie inside the relevant scattering volume. On the other hand the experiments show that it is desirable to illuminate the whole of the scattering volume, and that if the horizontal polar diagram is narrowed further than is necessary the extra gain of the aerial on a plane wave basis is not achieved.

Polarization Characteristics

Tests carried out[23] on about 860Mc/s at distances of 160 and 320km have shown that the polarization of the signals received by forward scatter propagation is preserved to a high degree independently of the initial state of polarization launched by the transmitting aerial. Records taken on receivers with the correct and crossed polarization consistently showed a difference of 30dB or more, the suppressed signal often being among noise.

It appears that the transmission characteristics do not depend upon the state of polarization which is substantially maintained over a scatter link. This fact precludes any use of polarization diversity such as can be used in long distance high-frequency communication by normal reflection from the ionosphere due to the anisotropic nature of the medium in that case[25].

Height-gain Relations

In ground-wave propagation under standard atmospheric conditions over smooth terrain there exists over the wavefront at the receiving point a coherent amplitude and phase relationship. As a result there is a definite height-gain function determined by the superposition of an incident and reflected wave system above the surface of the earth. With a signal consisting of a number of randomly related components arriving over a small cone of angles from the scattering volume, the interference pattern above the earth should be blurred and variable with time. However, depending upon the measure of incoherence across the wavefront, some vestige of a height-gain relationship may exist near to the ground.

Height-gain measurements were made on the 320km link on 860Mc/s already mentioned[23]. A well defined interference pattern was obtained up to a height of 10m, but at greater heights the maxima and minima decayed and showed little coherence between successive tests. The difference between the maxima and minima was only of the order of 3dB, and there was no sign of the large height-gains that would be exhibited by a purely diffracted signal, indicating the entirely different mechanism by which the energy reaches the receiver at large distances beyond the horizon.

The effect is thus of little practical interest especially as the dimensions of a high-gain aerial system such as a large parabolic dish are great compared with the scale of the interference pattern. Such tests, however, may give some indication of the scale of the incoherence across the incident wavefront and of the position of the scattering source.

On the other hand, as has already been pointed out, there may be some advantage in placing a terminal at a considerable height, e.g. on a convenient hill-top, where the horizon direction may lie appreciably below the horizontal plane through the aerial.

Diversity Measurements

It is well known that the troubles caused by fading in long-distance radio communication can be alleviated by the use of some form of diversity reception. It has already been seen that no use can be made of polarization diversity at a given point, but frequency and space diversity can be employed with advantage. So far no tests have been carried out in the United Kingdom on frequency diversity, but correlation measurements have been made on signals received on aerials spaced both vertically and horizontally.

Preliminary tests on 3 500Mc/s show that at times the vertical spacing has to be increased to 25 wavelengths before the correlation coefficient drops to zero, and that the figure is somewhat greater for horizontal spacing. On 860Mc/s respective figures of 35 and 55 wavelengths have been obtained for a correlation coefficient of 0·2.

The figures are very variable, depending upon the changing characteristics of the troposphere. They indicate, however, that it may be advisable to use a spacing as large as 100 wavelengths if the full benefit of this type of diversity reception is to be obtained.

Aircraft Effects

In the development of tropospheric scatter circuits for long distance communications, it has been realized from the start that reflections from aircraft, which are the basis of their detection by radar, would cause interference. Special observations have been made in the 860Mc/s and 3 500Mc/s tests of the aircraft effects which appear on a continuous wave transmission as a sinusoidal oscillation on the record caused by the characteristic Doppler beat between the wanted scatter signal and the signal changed in frequency by reflection from the moving aircraft.

In view of the large number of aircraft in flight over the United Kingdom, it is difficult to identify a given interference effect with a particular aircraft, except in some of the cases when the aircraft is in sight and its movement in relation to the path of the tropospheric link can be observed. The problem of the reflection received from an aircraft in relation to its position in the scattered field set up by the transmitter is complex, but it can be shown that the signal is strongest when the aircraft is near to either terminal.

On both frequencies this was found to be the case with respect to the receiving terminal and under such conditions the signal can be comparable with the wanted scatter signal. Although the phenomenon should be reciprocal, attempts to correlate interference effects at the receiver with the presence of aircraft in the vicinity of the transmitter on 860Mc/s had no success. It was, however, found during siting tests that when the receiver was unfavourably placed with the horizon obstructed by rising ground, the aircraft signals could overwhelm the wanted signal.

Meteorological Factors

As the mechanism of long-distance tropospheric propagation is essentially bound up with the physical properties of the lower atmosphere, it must obviously depend upon the meteorological situation typified by the weather existing along the transmission path. The diurnal and seasonal variations in the received signal must in some way be associated with changing atmospheric conditions. Attempts have therefore been made to correlate weather data obtained from the nearest meteorological stations with the signal variations.

It is, however, difficult to assess the true situation along the path from measurements taken intermittently at isolated points which are not actually on the path or near the mid-point, so that correlations with hourly median values are in general impracticable. The making of a continuous and detailed study of the refractive index all along the path is a formidable task that has not yet been attempted in the United Kingdom, and some of the difficulties involved may be appreciated by reference to an elaborate project carried out in New Zealand some years ago[26]. However, the development of a sensitive and quick-acting refractometer for use in an aircraft is a step towards this end.

The information obtainable from such a meteorological survey is, moreover, more likely to be concerned with the existence of refractive index gradients and sharp inversion layers that give rise to periods of propagation by ducting and by partial reflections when the signal is strong compared with the normal scattered signal under consideration. Even so, attempts to correlate such occasions with specific weather conditions have been largely inconclusive apart from general trends established on a long term basis[27].

The most that has been done in the United Kingdom tests is to show that there is a significant correlation between the mean refractive index gradient and the median signal strength[23,24], the presumed explanation being that with an increase in the gradient there is an increased curvature of the ray paths from the transmitter to the scattering volume and from the latter to the receiver, leading to a decrease in the associated scattering angle and hence to an increase in the received signal.

A significant correlation has also been claimed between wind velocity measurements and the movements of the scattering centres across the transmission path as deduced from time delays between the amplitude records made on two aerials spaced horizontally transverse to the path[23].

The experiments all agree that the signal is enhanced during fog or mist, though on the highest frequencies dense fog causes a reduction in signal due to atmospheric absorption. Heavy rain or snow can likewise depress the signal considerably over the whole frequency band, the attenuation being of the order of 0·06dB/km at a frequency of 3 500Mc/s. Under anticyclonic conditions the signal on 3 500Mc/s tends to be below average, while on lower frequencies the fading is often deeper and slower than during the well mixed atmospheric conditions associated with cyclonic weather.

It is not clear to what extent these observations imply a connexion between the weather and the nature of the turbulences responsible for the scattered signal. It appears that some degree of stratification may assist the scattering process, but such a conclusion is difficult to establish because of the effect of such stratification on the signal by the formation of ducts and partially reflecting layers.

Conclusions

The experimental results and the difficulty of relating them precisely to the meteorological conditions emphasize the complexity of the problem under consideration. In the Introduction it was stated that there is no general agreement about the physical mechanism of long distance tropospheric propagation and this is exemplified by the fact that for much of the time the process is not one of pure scattering typified by a Rayleigh distribution of signal amplitude. Changing weather conditions often imply the presence of some measure of layer or duct formation, and where the reflecting layers may not be continuous over a large area but broken up into a number of reflectors of limited extent, the signal may be the resultant of a number of partial reflections in random relationship and similar in characteristic to scattering from turbulences.

When such reflecting areas become comparable with the size of the scattering centres, there is experimentally no clear cut distinction between the two mechanisms. This is equivalent to saying that the fundamental physical process is one of scattering and that reflection from a plane surface of sufficient extent represents a limiting condition in which the scattering is completely coherent. While the theoretical study of the formation of turbulent eddies is of great importance and the statistical analysis of scattering on certain idealized assumptions has led to a knowledge of the general form of the laws of scattering, it is doubtful whether it is fruitful to pursue this line of approach in great detail in the attempt to explain the experimental results.

The tests have shown the need to establish the propagation characteristics by experimental observation, the theory acting as a general guide in the interpretation of the records. The United Kingdom results in general fall into line with those found in the U.S.A. and in northern Europe insofar as the average meteorological conditions are similar. There is plenty of evidence that in tropical regions, especially in the desert regions of the Middle East, the mechanisms of partial reflection and ducting can play a dominating part at some seasons of the year, giving rise to very strong signals for hours at a time.

As far as is known, no really long distance u.h.f. tropospheric link has been established in the United Kingdom on an operational basis. Steps are now being taken to engineer such a circuit that can be modulated for traffic and pulse transmissions to implement the findings of the experimental links. Preliminary tests with speech modulation over the experimental links have shown promising results, though it must be remembered that in the case of this type of intelligence the ear and mind are tolerant to momentary breaks due to fading that would be unacceptable with some other modulation systems.

On the other hand these tests were carried out without the use of diversity reception and with lower transmitted power than would be employed in a fully engineered circuit. Tests must be made with such a circuit making full use of diversity reception before it is possible to pronounce with complete confidence on the operational performance of tropospheric links for any projected service. In particular, such tests are needed to study with the aid of pulses the problem of time delays in relation to multi-channel working and television relaying.

These time delays are known to be much shorter with tropospheric than with ionospheric forward scatter propagation because of the much lower level at which the scattering occurs and of the small volume of the medium involved, but troubles may occur at times of abnormal propagation when the mechanism ceases to be solely one of scattering. The problem of available bandwidth is a complex one, but some discussion of it has been given in a recent paper by Saxton[28] who concludes that the prospects for television relaying are promising. Although some claims have already been made in this direction[29], it is not proven that the mechanism was truly that of tropospheric scattering, and as far as is known, no long term performance tests have yet been carried out to substantiate the possibility of, for example, bridging the Atlantic by a chain of long-distance tropospheric television links.

There are obvious reasons for employing this mode of propagation for military purposes, especially across inaccessible intervening terrain, even if the performance is not as perfect as would be demanded of a commercial installation. The problem of using such a link in preference to a chain of line-of-sight relays where the latter is practicable in an economic as well as a technical one, and as such is outside the scope of this article.

As in many other applications of radio, some measure of disturbance must be expected which has to be tolerated, provided it can be kept within acceptable limits. Mention has been made of aircraft effects. Although a large measure of protection will be obtained by the use of highly directive aerials, it obviously cannot be entirely eliminated in regions of high air-traffic density. Where the type of modulation permits, use can be made of error detection and correction methods, but these would presumably not apply for television where storage techniques are difficult to envisage.

Finally it may be pointed out that the use of this new mode of propagation raises further problems of frequency allocation and sharing. It is worth noting that in international circles this aspect of the problem is already coming under review and attention is being drawn to possible interference problems under conditions of abnormal propagation in view of the large powers that will be used. It is well known that serious difficulties have been met in the case of ionospheric scatter propagation which are requiring special measures to overcome them. It is important that in the case of tropospheric propagation such difficulties should be anticipated as far as possible in the preliminary research survey along the lines that have been discussed in this article.

Acknowledgments

The author wishes to thank the Engineer-in-Chief, Marconi's Wireless Telegraphy Company Limited for permission to publish this article.

REFERENCES

In the following list of references, the description 'Symposium Paper' refers to the Symposium in reference 1.

1. Symposium on 'Long-distance propagation above 30Mc/s' held at the Institution of Electrical Engineers on 28th January, 1958. To be published in Vol. 105B of the Proceedings.
2. SAXTON, J. A. The Propagation of Metre Radio Waves Beyond the Normal Horizon. *Proc. Inst. Elect. Engrs. 98*, Pt. 3, 360 (1951).
3. Colloque International sur la Propagation. (Paris-September, 1956). *L'Onde Electrique. 37*, 411 (1957).
4. Les Problemes d'actualite dans la Propagation des Ondes Radioelectriques. (Colloque International. Paris 17-21, September, 1956) *Ann. Telecommun. 12*, 140 (1957).
5. STARAS, H. Forward Scattering of Radio Waves by Anisotropic Turbulence. Scatter Propagation Issue. *Proc. Inst. Radio Engrs. 43*, 1174 (1955).
6. DICKSON, F. H., EGLI, J. J., HERBSTREIT, J. W., WICKIZER, G. S. Large Reductions of v.h.f. Transmission Loss and Fading by the Presence of a Mountain Obstacle in Beyond-Line-of-Sight Paths. *Proc. Inst. Radio Engrs. 41*, 967 (1953).
7. CARROLL, T. J. Normal Tropospheric Propagation Deep into the Earth's Shadow: the Present Status of Suggested Explanations. *Trans. Inst. Radio Engrs. PGAP-3*, 6 (1952).
8. CARROLL, T. J., RING, R. M. Propagation of Short Radio Waves in a Normally Stratified Troposphere. *Proc. Inst. Radio Engrs. 43*, 1384 (1955).
9. GRAIN, C. M. Survey of Airborne Microwave Refractometer Measurements. *Proc. Inst. Radio Engrs. 43*, 1405 (1955).
10. STARKEY, B. J., TURNER, W. R., BADCOE, S. R., KITCHEN, G. F. The Effects of Atmospheric Discontinuity Layers Up To and Including the Tropopause on Beyond-the-Horizon Propagation Phenomena. Symposium Paper No. 2486R.
11. CARROLL, T. G. Marconi's Last Paper, On the Propagation of Micro-Waves over Considerable Distances. *Proc. Inst. Radio Engrs. 44*, 1056 (1956).
12. ISTED, G. A. Guglielmo Marconi and Communication Beyond the Horizon: A Short Historical Note. Symposium Paper No. 2479R.
13. PEKERIS, C. L. Wave Theoretical Interpretation of Propagation of 10cm and 3cm Waves in Low-Level Ocean Ducts. *Proc. Inst. Radio Engrs. 35*, 453 (1947).
14. MEGAW, E. C. S. Scattering of Electromagnetic Waves by Atmospheric Turbulence. *Nature, Lond. 166*, 1100 (1950).
15. BOOKER, H. G., GORDAN, W. E. A Theory of Radio Scattering in the Troposphere. *Proc. Inst. Radio Engrs. 38*, 401 (1950).
16. VILLARS, F., WEISSKOPF, V. F. The Scattering of Electromagnetic Waves by Turbulent Atmospheric Fluctuations. *Phys. Rev. 94*, 232 (1954).
17. BATCHELOR, G. K. The Scattering of Radio Waves in the Atmosphere by Turbulent Fluctuations in Refractive Index. Research Report No. EE 262, School of Electrical Engineering, Cornell University (1955).
18. MEGAW, E. C. S. Fundamental Radio Scatter Propagation Theory. *Proc. Inst. Elect. Engrs. Monogr. 236R*, 15 (May, 1957).
19. JOHNSON, M. A. A Review of Tropospheric Scatter Propagation Theory and its Application to Experiment. Symposium Paper No. 2534R.
20. KITCHEN, F. A., RICHARDS, E. G., RICHMOND, I. J. Some Investigations of Metre-Wave Radio Propagation in the Trans-Horizon Region. Symposium Paper No. 2509R.
21. JOY, W. R. R. Radio Propagation Far Beyond the Horizon at about 3.2cm Wavelength. Symposium Paper No. 2538R.
22. JOY, W. R. R. The Long-Range Propagation of Radio Waves at 10cm Wavelength. Symposium Paper No. 2522R.
23. RIDER, G. C. Some Tropospheric Scatter Propagation Measurements and Tests of Aerial Siting Conditions at 858Mc/s. Symposium Paper No. 2528R.
24. ANGELL, B. C., FOOT, J. B. L., LUCAS, W. J., THOMPSON, G. T. Propagation Measurements at 3480Mc/s over a 173-Mile Path. Symposium Paper No. 2510R.
25. GRISDALE, G. L., MORRIS, J. G., PALMER, D. S. Fading of Long-Distance Radio Signals and a Comparison of Space and Polarization-Diversity Reception in the 6-18Mc/s Range. *Proc. Inst. Elect. Engrs. 104B*, 39 (1957).
26. MILNES, B., UNWIN, R. S. A Radio-Meteorological Investigation in the South Island of New Zealand. *Proc. Phys. Soc. Lond. 63B*, 595 (1950).
27. SPENCER, C. W. Radio-Wave Propagation at 89Mc/sec in Relation to Synoptic Conditions. *Met. Mag. 81*, 206 (1952).
28. SAXTON, J. A. Scatter Propagation and its Application to Television. *J. Televis. Soc. 8*, 1 (1957).
29. TIDD, W. H. Demonstration of Bandwidth Capabilities of Beyond-Horizon Tropospheric Radio Propagation. *Proc. Instn. Radio Engrs. 43*, 1297 (1955).

MICROWAVE LINK DEVELOPMENT in the Radio Laboratories of the Post Office Engineering Department

By C. F. Floyd*, M.A., A.M.I.E.E. and R. W. White*, B.Sc., F.Inst.P., A.M.I.E.E.

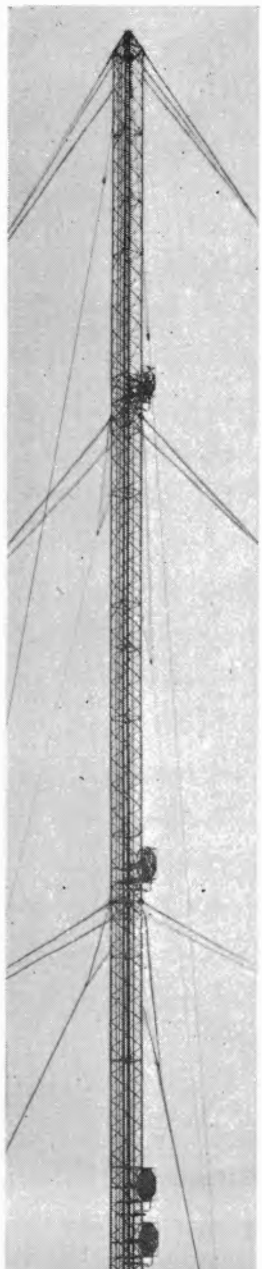
The contribution made by the Radio Branch Laboratories of the Post Office to broad-band microwave system development is surveyed. Their work includes the design and installation of complete point-to-point 4 000Mc/s television and multi-channel telephony systems and associated test equipment.

THE British Post Office has had a close interest in the development of point-to-point radio circuits since the earliest days of radio communication, and its Radio Development Laboratories at Dollis Hill and elsewhere are concerned with research and development on new systems and with the investigation of new measuring and testing techniques.

The development of modern broad-band links owes much to earlier work on v.h.f. links before and during the war[1-6]. The last two decades have seen a rapid growth of interest throughout the world in microwave radio systems for multi-channel telephony or point-to-point television distribution. Many laboratories have made important contributions to the art and the pioneering work of the Bell Laboratories in America has been of fundamental importance. Significant contributions have also come from the laboratories of several British firms. This article is, however, restricted to a survey of some of the work of the Post Office Radio Laboratories; limitations of space preclude much detail, and for further information reference should be made to the extensive bibliography.

Microwave Systems

In 1946 a one-way experimental television link operating at approximately 200Mc/s was established from London to Castleton, near Cardiff[7], and pictures were successfully transmitted over it in March. 1949. This experimental link was one of the first to use frequency modulation for the transmission of television signals and it provided much valuable information on the problems involved in the use of this form of modulation. Subsequently three hops of this link were converted to 4 000Mc/s operation and in 1952 the composite link shown in Fig. 1 was used to provide a temporary service to the BBC Wenvoe transmitter pending completion of the permanent coaxial cable link from London.



The photograph shows the buildings and aerials at the southern terminal of the Elgin-Wick link.

THE ELGIN-WICK MULTI-CHANNEL TELEPHONY SYSTEM

After the successful demonstration of broad-band transmission over the London-Castleton radio link, the development of a single-hop frequency-modulation telephony system on a 55-mile long oversea path across the Moray Firth was undertaken[8,9]. This link was required to carry 240 channels and, so far as it was possible on such a long hop, had to meet C.C.I. standards.

The likelihood of multi-path propagation and fading, made space diversity reception essential. Preliminary propagation tests also showed that frequency selective fading was to be expected, and this led to the provision of frequency diversity in addition to space diversity. The use of frequency diversity was acceptable in view of the isolated location of the link. However, in locations where pressure on available frequency space is acute, the use of frequency diversity may not be acceptable. The Elgin-Wick link is, in effect, a quadruple diversity system, with the general arrangement shown in Fig. 2. As complete transmitters and receivers are provided for each frequency, there is no need for standby equipment.

Of the three methods of diversity switching:

(a) At microwave frequency at the receiver input.

(b) At intermediate frequency, 60Mc/s.

(c) At base-band.

intermediate frequency switching (b) was the most attractive at that time.

It was chosen for the spaced aerial diversity, the aerial giving the higher output being

* *Radio Laboratories of the Post Office Engineering Research Station*

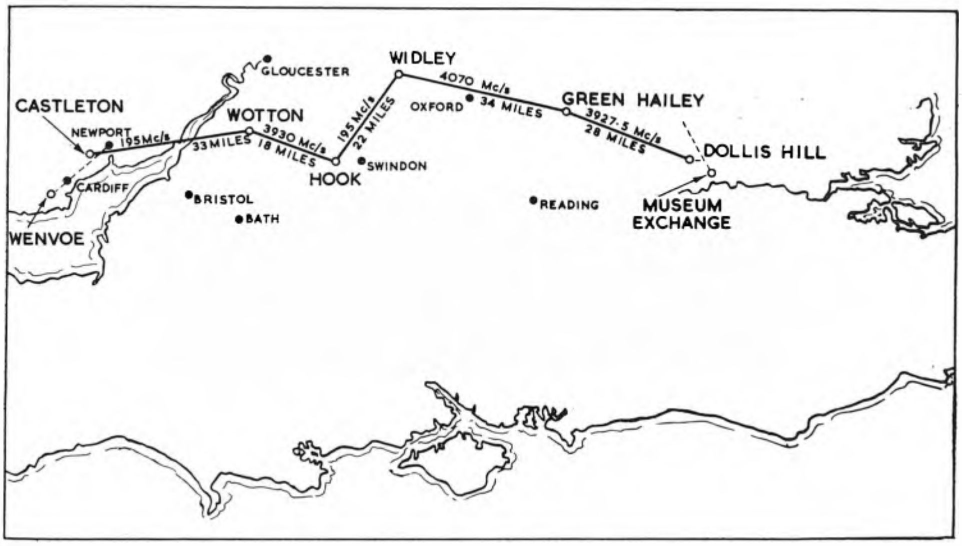


Fig. 1. The route of the Dollis Hill to Castleton broadband television radio link

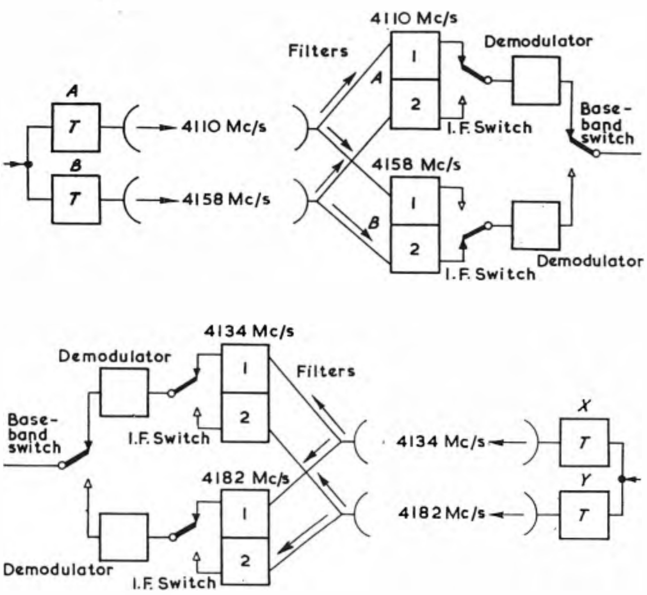
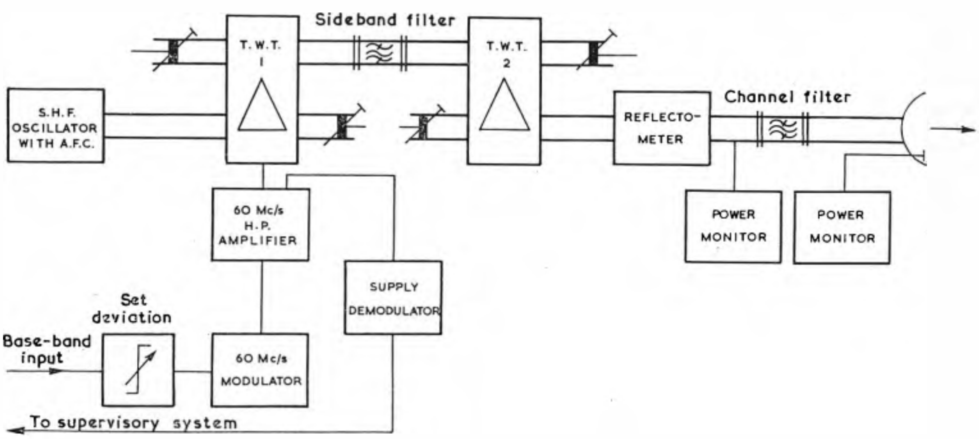


Fig. 2. Simplified block schematic of radio equipment

Fig. 3. Block schematic diagram of transmitter



selected by the diversity switch. The changeover time of this switch is less than 0·3μsec and the time required for the selecting circuit to decide that a changeover is needed is about 3msec. The switch is required to be most sensitive under deep fade conditions, i.e. when diversity improvement is most needed, and the equipment is adjusted so that a 2dB level difference causes changeover from one aerial circuit to the other. For normal reception levels, i.e. in the absence of deep fading, the margin becomes 5dB.

The two broad-band channels providing frequency diversity are independent as far as the final demodulator output (i.e. to base-band) and at this point, a fast-operating relay selects the output with the lower basic noise. The noise is continuously monitored in a narrow pilot channel selected near the top of the base-band; the diversity switch selects the radio channel with the better signal-to-noise ratio. The circuit interruption arising from a changeover is less than 5msec, which is short enough to prevent appreciable interference to traffic.

The optimum heights for the four dish aerials—two transmitting and two receiving at each end of the link—were determined by experiment. The mast at the southern terminal is shown on page 253; the two transmitting dishes are mounted low down and the space diversity receiving aerials are at widely different heights higher up the mast. It was found desirable to restrict the heights of the dishes to minimise the effects of reflected rays, so as to give the optimum space diversity action.

The transmitter, Fig. 3, gives an output power of 1W into the waveguide feeder. The base-band signals, comprising 240 telephone channels in the range of 60 to 1 052kc/s plus a sub-carrier band for the supervisory and speaker circuits, frequency modulate a 60Mc/s i.f. carrier: the resulting signal in the range 55 to 65Mc/s, phase modulates the 4 000Mc/s carrier in a travelling-wave amplifier[10]. One sideband is selected by a waveguide filter and amplified to 1W by a second travelling-wave amplifier before transmission to the aerial. The intermediate frequency of 60Mc/s was in use here before the C.C.I.R. had adopted the value of 70Mc/s as a standard.

The microwave receiver, Fig. 4, is a superheterodyne without any r.f. stage. The signal is selected by a waveguide filter and mixed in a crystal frequency changer with a local oscillator output to give a difference frequency centred on 60Mc/s; this is amplified, automatic gain control keeps the level constant, and, after the i.f. diversity switch, a demodulator extracts the

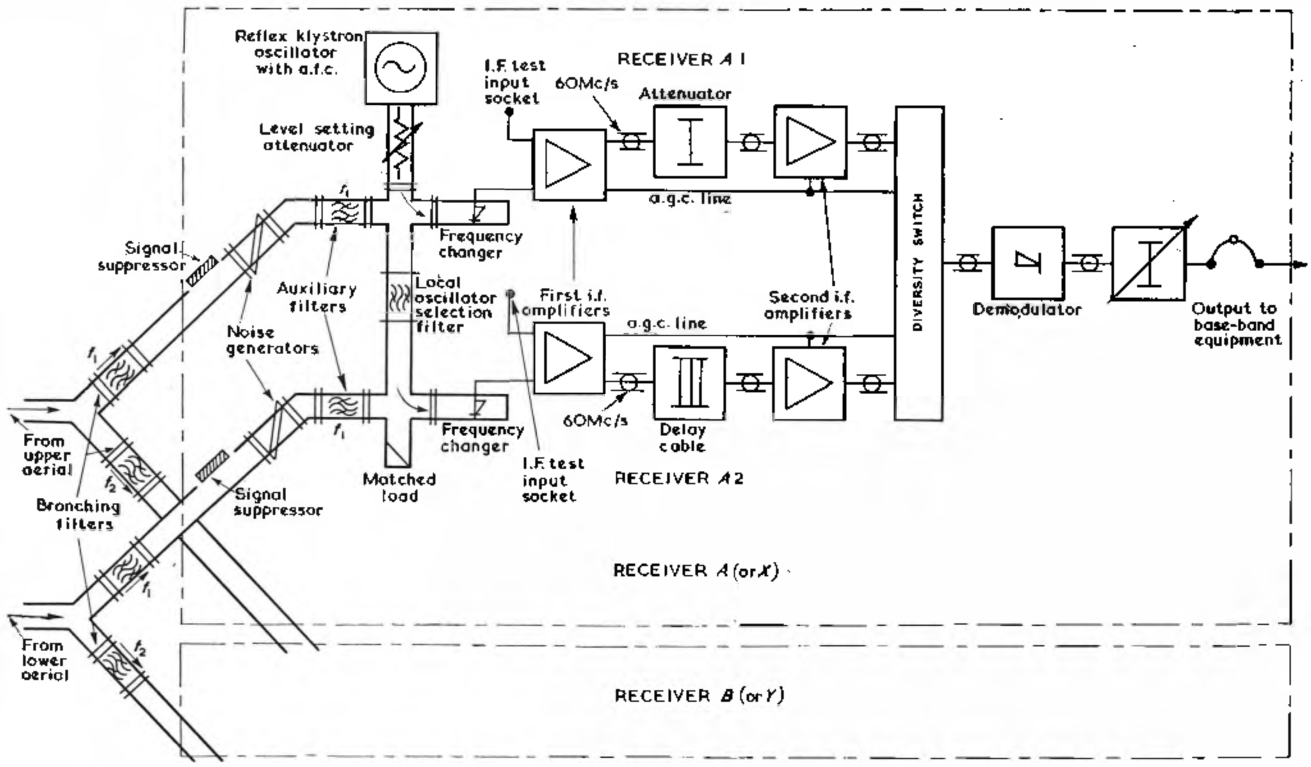


Fig. 4. Block schematic diagram of receiver

base-band signal. The sub-base-band, with supervisory channels and speaker circuits, is separated by filters, leaving the 240 channel band, 60 to 1 052kc/s.

However good the transmission characteristics of a radio link may be, its performance is assessed by reliability and ease of maintenance; the supervisory system is therefore of great importance. The supervisory system[9] uses where possible standard types of equipment as employed on co-axial cable systems. Separate audio frequency tones, drawn from the v.f. telegraph series with 120c/s separation, carry each of the various supervisory indications required. Two speaker circuits are also included. All these audio channels are combined and translated into the band 32 to 48kc/s which is then added as a sub-base-band to the 240-channel base-band, 60 to 1 052kc/s, making a total band of 32 to 1 052kc/s.

The link was opened to traffic in September, 1957. Typical average performance figures under non-fading conditions measured with the white-noise intermodulation test equipment, which will be described later in this article, are given in Table 1. These values are, of course, measured in this type of test as noise power ratios, but they have been converted in Table 1 to signal-to-noise ratios by adding the conversion factor, 18·7dB.

As an indication of the fading experienced on this Elgin-Wick path, fades exceeding 20dB below the steady propagation condition existed for 0·3 per cent of the total time when

TABLE 1

Signal-to-noise ratios on the Elgin-Wick radio link

	Signal/Noise Ratio with 240 Channel Loading			Basic Signal/ Noise Ratio at
Frequency ..	70kc/s	534kc/s	1 002kc/s	1 002kc/s
Elgin to Wick	78dB	69dB	68dB	72dB
Wick to Elgin	78dB	74dB	69dB	72dB

measured on one aerial. Using two aerials in switched space diversity, no fades exceeding 13dB were recorded, and 10dB fades lasted only 0·05 per cent of the total time. These values are representative of mid-winter conditions.

The London-Isle of Wight Television Transmission System

The London-Isle of Wight system is a two-hop link designed by the Laboratories for the transmission of television signals; it was opened for service in October 1954. Initially, it used a direct pick-up receiver at Alton and a single-hop, 40 mile, 4 000Mc/s link from Alton to Rowridge on the Isle of Wight[11], but later the direct pick-up receiver at Alton was replaced by a 4 000Mc/s hop from a transmitter on the roof of Museum Telephone Exchange. Diversity working is not used, but a standby channel is provided.

The video signal at the Museum terminal frequency-modulates a 60Mc/s i.f. carrier; the broad-band signal from the i.f. amplifier is then translated to the 4 000Mc/s band and this signal is radiated from a 10ft diameter paraboloid dish. A two-way branching filter is needed to couple the two transmitters that provide the alternative channels to the transmitting feeder and aerial. The subsequent installation of a third 'go' channel has entailed the use of an additional branching filter. A 'return' channel, which is also now operating, uses separate dish aerials. Under normal conditions the basic signal-to-noise ratio on the link is about 50dB which gives a useful, though not excessive, margin against fading. The signal-to-noise ratio is here defined as the ratio of the peak-to-peak picture signal (excluding synchronizing signals) to the r.m.s. noise.

Recent Development Work on Microwave Systems

Progress in the design of microwave systems continues; for example, travelling-wave tubes have improved since the completion of the Elgin-Wick system and low-noise and high-power (10W) output travelling-wave tubes are now available. Similarly, i.f. modulators, filters and equalizers for these wideband systems have improved. Such diverse items of new equipment must be tested in the field and as a system; a laboratory test cannot meet all these needs. An experimental 4 000Mc/s installation has been set up over a 33-mile path for the field trial of improved components and for the investigation of systems of greater capacity than 600 telephone channels.

Broadband Intermediate Frequency Equipment

Amplifiers

During 1950-1 three types of 60Mc/s amplifier were developed. One of their notable features was the extensive use of single-tuned circuits as interstage coupling networks, with negative feedback applied across alternate valves to provide a band-pass characteristic substantially flat from 55 to 65Mc/s. The three designs were, respectively, a low-noise pre-amplifier for direct association with the mixer unit of a receiver, a main amplifier providing the bulk of the gain and a high level amplifier used to feed a travelling wave tube operating as a convertor from i.f. to microwave frequency[10]. They were used in the London-Castleton experiments and in the London-Isle of Wight link.

The i.f. amplifiers for the Elgin-Wick link[8], developed during 1953-4, resembled the earlier units by being centred on 60Mc/s and by making considerable use of single-tuned circuits plus negative feedback; but they were of different mechanical design, used wired-in valves of the flying lead type and had a much improved electrical performance—particularly in group delay characteristic. They were designed to meet C.C.I.R. overall transmission standards for either 240-channel telephony or 405-line monochrome television.

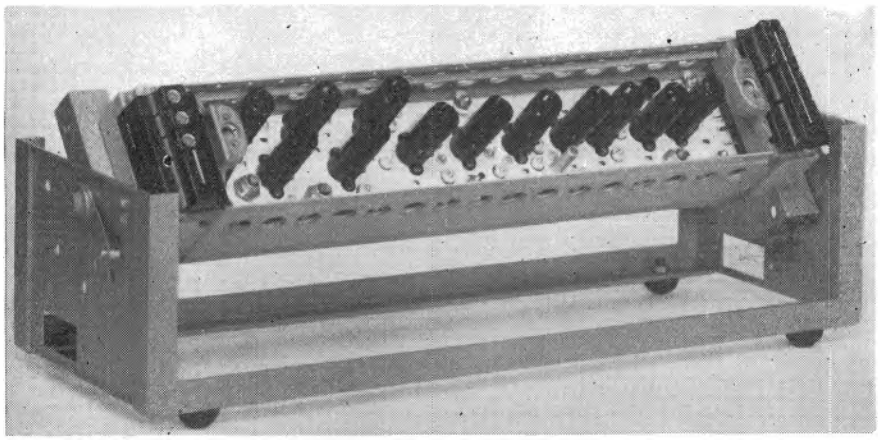


Fig. 5. A Post Office Modulator No. 9A, with front cover removed, shown in a bench mounting cradle of the type used during wiring and laboratory testing. *This modulator accepts either television (up to 625 lines) or multi-channel telephony (up to 960 channels) and provides a frequency modulated carrier centred on 70Mc/s*

All the 60Mc/s units were designed before C.C.I.R. agreement had been reached on the choice of i.f., but more recent designs of i.f. amplifier have been centred on 70Mc/s—the current C.C.I.R. standard. A typical five-stage amplifier recently described by Hamer and Gibbs[12] provides 50dB gain over an almost flat pass-band of approximately 50 to 90Mc/s, and over the central 12Mc/s (i.e. 70 ± 6Mc/s) the group delay is constant to within 1mμsec.

Frequency Modulators

In early v.h.f. work, several forms of reactor-valve frequency-modulated oscillator were used; but the maximum deviation available with adequate linearity, and without excessive amplitude modulation, was restricted to one or two per cent of the oscillator mean frequency.

Klystrons can be used as sources of microwave frequency modulated signals, but there are operational advantages in having a modulator unit whose output is at intermediate frequency rather than at the microwave carrier frequency. A method commonly used in U.S.A. is to beat together two klystrons with frequencies separated by the i.f., and to apply the modulating signal to one of the klystrons. This arrangement has also been used by the Post Office Laboratories for test equipment.

During 1950-1 a simple and convenient form of i.f. modulator was developed from the well-known ladder network type of resistance-capacitance phase-shift oscillator by using earthed-grid triode valves as the variable impedance elements[13,14]. This type of frequency modulator has proved suitable for 405 line monochrome television or for 240 telephone channels; but it is not good enough for 600 channel operation. During 1955 two important improvements were made[15,16], and the frequency modulator which is shown mounted in a laboratory testing cradle in Fig. 5 has a performance adequate to meet C.C.I. transmission standards when carrying either 625 line television or considerably more than 600 telephone channels. The circuit diagram of the frequency-modulated oscillator section of this unit is shown in Fig. 6. The mean frequency of the oscillator is 35Mc/s and it is frequency doubled to provide the required 70Mc/s output signal.

Demodulators

The demodulators used in broad-band systems have to handle high base-band frequencies and relatively high deviations, and at the same time non-linear distortion must be kept low. Discriminators of the Foster-Seeley or of the two-tuned-circuit type can be operated satisfactorily in systems carrying television or small numbers of telephony channels; but, when higher standards of performance are required, more complex networks have to be used in the discriminator, and attention has to be paid not only to the linearity of the frequency/voltage characteristic but also to the phase or group-delay characteristic. Some recent 70Mc/s demodulators have used an asymmetrical or unbalanced form of discriminator of high linearity; this has the advantage of simplicity relative to balanced designs of comparable linearity and stability, but the preceding limiter must be somewhat better.

The performance of a limiter should be assessed in terms of the efficiency with which it suppresses amplitude modulation up to at least twice the highest base-band frequency with which it is to be used, and its freedom from conversion of amplitude modulation into phase modulation. Many designs of limiter have been tried out and there has been useful progress during recent years, but much remains to be done in this field. Current designs of demodulator have limiter stages giving well over 40dB suppression of 10 per cent amplitude modulation at any frequency up to 8Mc/s.

The base-band amplifier which follows the discriminator should have the frequency response required by television considerations, yet must have low enough non-linear distortion for multi-chanel telephony, if either type of traffic is to be handled. A video amplifier with a considerable amount of negative feedback is therefore necessary.

Group-Delay Equalizers

A typical i.f. amplifier or microwave filter, which has been aligned for a transitionally flat amplitude/frequency characteristic, has a group-delay/frequency characteristic which is approximately quadratic in shape; i.e. it may be represented by a linear plus a square-law component, with the latter (parabolic curvature) generally predominant. The group-delay at the apex is normally a minimum. In modern systems fixed equalizers, generally of the all-pass bridged-T type, are used extensively to counteract group-delay distortion: they may be built into i.f. units or inserted as required in the interconnecting cables between units.

Even with the best available techniques, the cumulative delay distortion in a complete system may be larger than is permissible and may be subject to small variations with ageing of equipment or environmental changes. It is desirable, therefore, to have the facility for inserting a 'mopping-up' equalizer prior to the terminal demodulator of a system. A suitable variable group-delay equalizer has been developed[17,18] to provide independent and continuously variable controls for slope and parabolic curvature over a nominal bandwidth of ± 6Mc/s centred on 70Mc/s. The range of slope adjustment is from − 1 to + 1mμsec/Mc/s, while the parabolic control has a range of 0 to +10mμsec delay at midband relative to that at ± 6Mc/s. The principles involved are applicable to other bandwidths, and can perhaps be extended to provide variable correction for curvature corresponding to higher order terms of a power series, thus allowing more accurate equalization of the edges of the band in a long system.

Aerials and Feeders

One of the remarkable features of line-of-sight microwave point-to-point radio circuits is that a very low transmitter power gives an adequate performance; for example an output of less than 10W is normally satisfactory. One reason for this is the fact that almost all the radiated energy is concentrated in a narrow beam accurately steered towards the receiving station by a highly directive aerial. The principles of operation of paraboloid dish reflectors are now well documented[19], and the Laboratories have concentrated on the development of the associated waveguide feeders and launching units. Particular care must be taken with accuracy of impedance matching at joints, as a poorly constructed joint in a long waveguide feeder can cause enough reflection of energy to give a perceptible return wave. If this suffers a reflexion at a second mismatch the forward-travelling echo, which has been delayed relative to the main signal, will create intermodulation noise on the telephony channels or an echo on a television screen. For example, a voltage-standing-wave-ratio of 0·93 (i.e reflection coefficient of 3·5 per cent) at each end of a 125ft length of copper waveguide feeder can give rise to intermodulation noise of −75dB in the uppermost channels of a 600 channel telephony link and there may be many such points in a system. Brass flanges ½in thick with accurately ground faces, dowelled together at joints, are used to connect lengths of waveguide to build up long runs. Corrosion inside the waveguide is prevented by pressurizing[20] with a dry gas.

The performance of a dish aerial depends among other things on the design of its launching unit, that is, the device that radiates the energy carried by the feeder so as to illuminate the whole area of the paraboloid. The prevention of spill-over at the edges of the dish is important in reducing radiation in unwanted directions. For this, and other, reasons a 90° type of paraboloid (i.e. one which has its focus in the plane of the rim of the dish) is preferred by the Laboratories.

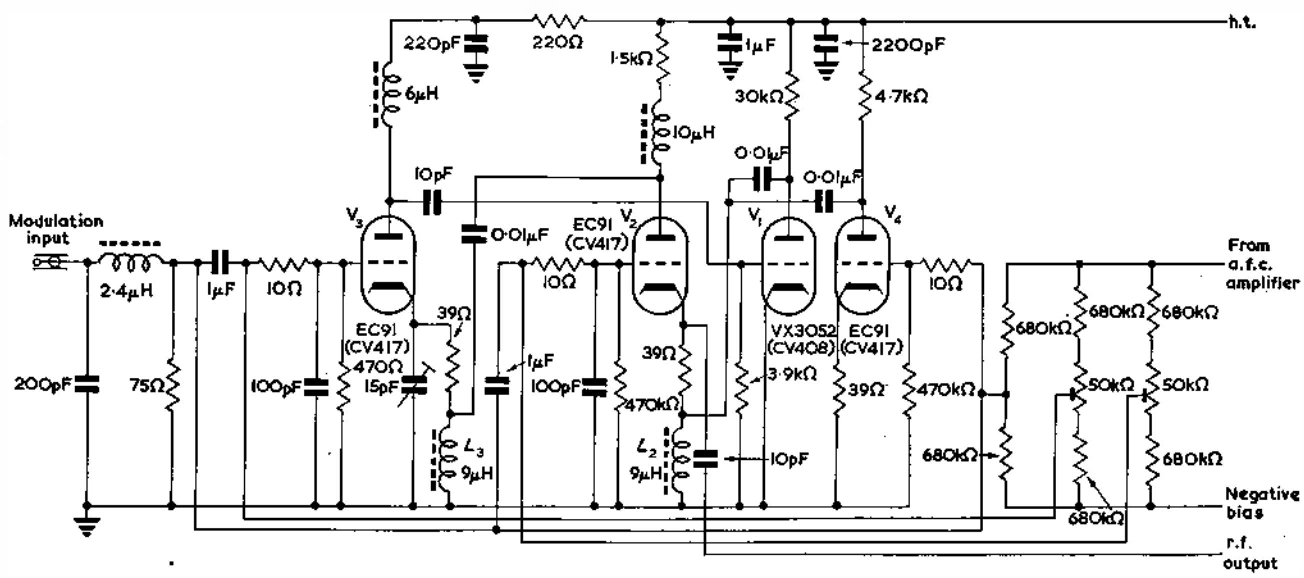


Fig. 6. Typical circuit of frequency modulated oscillator

Waveguide Filters

Filter technique for microwave frequencies has grown in step with the requirements of broad-band systems, and filters can now be designed in waveguide[21] with complete confidence. The design data for post-type waveguide filters has been reduced to nomogram and chart form.

The group delay characteristic in the pass-band is that property of a band-pass filter which is most critical in a broad-band f.m. system, as non-uniformity of this delay leads to intermodulation noise in telephone channels. Much of the development effort in waveguide filters has therefore been concentrated on this aspect of design technique. One of the shortages in test equipment that has not yet been satisfactorily overcome is means for measuring this group delay at 4 000Mc/s to the accuracy of 0·1mμsec required for design work. Accurate group-delay measurements at 4 000Mc/s are among the most difficult of all tests to carry out. Voltage standing-wave ratio measurements, by which waveguide filters are 'aligned', are easy by comparison.

There is still much interesting work for the filter designer in the microwave system field, for example, in connexion with the development of branching filters, used to combine on to one aerial feeder a number of r.f. channels derived from separate transmitters working on different carrier frequencies. As many as six such r.f. channels may eventually have to be combined in the 4 000Mc/s band, but the highest number yet in actual service using branching filters developed in the Laboratories is four, on the television circuit from London to the Isle of Wight.

Carrier Generation

It will have been seen earlier that an automatic frequency control unit forms part of each microwave transmitter and receiver on the Elgin-Wick link[8]. This electromechanical device adjusts the frequency of a klystron oscillator to within ±100kc/s of that of a high stability cavity resonator. The frequencies of the two ends of the link are independent; each has its own reference cavity with an associated a.f.c. unit so that the cumulative error in the receiver i.f. should be less than 200kc/s. Automatic frequency control units are, however, complicated devices and the generation of microwave carriers from highly stable quartz crystal oscillators[22] is now considered better technique.

The enormous increase in frequency from a few megacycles per second, at which quartz techniques are excellent, to 4 000Mc/s cannot be met directly by quartz crystals oscillating in a fundamental or overtone mode. Before microwave carriers could be controlled by quartz crystals it was necessary to await the arrival of valves suitable for frequency multiplication stages working from tens of megacycles to 4 000Mc/s. Only recently have such valves become available in reliable form. As a result, a long-term carrier stability with 1 part in 10^6 can now be achieved at 4 000Mc/s, although this high degree of performance may be unnecessary with present systems using frequency modulation techniques.

The measurement of frequency at microwaves has so far been done by means of precision cavity wavemeters. A cavity with calibrated screw adjustment for frequency determination at 4 000Mc/s has to be a most carefully constructed instrument using the most advanced mechanical techniques; the laboratory's frequency measuring group has recently completed a development project on such a wavemeter[28], suitable for adjusting microwave carrier frequencies to within one part in 10^5.

Microwave Components

The most generally useful size of guide for 4 000Mc/s is the waveguide No. 11 (2·372in × 1·122in inside dimensions) and this has, with few exceptions, been accepted by designers as the standard for the 4 000Mc/s band. Since no components for No. 11 waveguide were available commercially when 4 000Mc/s system development started, the Laboratories were forced to develop most of their own microwave components. For example, development work has been done on the design of bends and twists, on impedance transforming sections and new designs of directional couplers. A good deal of effort has been devoted to the development of frequency-changing devices, both travelling-wave tubes and crystal mixers[23], for use in microwave systems. There is as yet no completely satisfactory engineering solution, but recent development in America on junction diode convertors may lead to the eventual answer.

In the early 1950s there was much rivalry between the development of systems using microwave triodes and those using travelling-wave valves. The Laboratories decided to use the latter[24] for work on 4 000Mc/s broad-band amplifiers, on the grounds that travelling-wave tubes are capable of development to wider bands and higher frequencies, whereas triodes are essentially narrow-band devices that require extremes of manufacturing technique to make them operate in the microwave region. Subsequent events have justified this decision, and on future systems development will be concentrated, as hitherto, around travelling-wave tube amplifiers.

Measuring Equipment and Techniques

In many fields of applied science the rate of progress and the standard of performance which can be achieved are directly dependent on the development and refinement of associated measuring techniques. The development of broad-band radio relay systems is no exception, and over the last ten years almost half of the experimental and development effort available in the Laboratories for work on broad-band radio systems has been devoted to the exploration and improvement of measuring instruments and techniques.

Since system performance tolerances are very stringent, the discrimination required in measurements is of a high order and, in general, is not easy to achieve. Wide frequency bands are normally involved, and the effects, over the whole band, of circuit changes or adjustments in equipment under test must be immediately observable. Scanning methods are always quicker and generally far more satisfactory in this field than point-by-point methods. Measuring instruments using oscilloscopes can be made not only to display very small quantities, but also to permit their measurement just as accurately as laborious point-by-point methods, and scanning methods have been found to be of great value in the improvement of broad-band radio equipment. On the other hand these precision measuring instruments are both complex and expensive to develop and build.

Scanning Oscillators

In any scanning or sweeping oscillator for use in broad-band system measurements, a large and continuously variable width of scan, easy adjustment of centre frequency, an accurate frequency calibration arrangement, low harmonic content and a high degree of uniformity of amplitude over the scanning range may be regarded as essential requirements.

The simplest form of microwave scanning oscillator is probably a reflex klystron with a suitable waveform applied to its reflector electrode; at 4 000Mc/s a scan of 30Mc/s is typical. When larger frequency scans are required (e.g. 50 to 100Mc/s), electromechanical arrangements such as motor-driven rotating paddles[24] or vibrating diaphragms have been used fairly extensively. During the last few years, however, not only carcinotrons but also normal travelling-wave tubes with external feedback have found applications as convenient electronically-controlled scanners for very wide bandwidths.

For i.f. use, many forms of scanner have been tested, and three types have been found particularly suitable for incorporation in broad-band precision test equipment. The first of these i.f. scanners uses a small ferrite core in the inductor of a v.h.f. oscillator and the frequency sweep is obtained by applying an external magnetic field. This arrangement was combined with suitable oscilloscope display facilities in a compact and transportable instrument developed during 1953-4 for accurate measurement of (a) gain/frequency characteristics and (b) reflection coefficients.

The output from this instrument is flat to within ±0·1dB from 45 to 85Mc/s.

A second type of i.f. scanner is based on the phase-shift oscillator principle, using cathode-follower or earthed-grid valves as the variable resistance elements[13]. Variants of this type of scanning oscillator have been embodied in several measuring instruments and they possess the merits not only of high linearity but also of being usable with very high scanning frequencies or complex scanning waveforms. It is of interest to record that the Post Office type of i.f. modulator already discussed started as a by-product of early work on these scanners.

In the third type of i.f. scanner the output signal is the difference frequency of two reflex klystrons, one or both of the klystrons being modulated by electronic, or occasionally mechanical, means. A scanner of this type offers great flexibility of output frequency and has advantages in measuring equipment which may at some time have to operate on a frequency band other than that for which it was originally intended.

Group-Delay Measurement

In measuring group-delay distortion, whether of a simple network or a complete system, it is necessary to determine the departure from uniformity of the slope of its phase/frequency characteristic; i.e. to determine either (a) the variations in the increments of insertion phase shift corresponding to equal intervals of frequency, or (b) the intervals of frequency corresponding to equal increments of phase (e.g. 180° or 360°). The former method has been more widely used in scanning equipment, since the frequency interval can readily be fixed by using a modulated test signal and the problem is then reduced to the measurement of small changes in the phase difference between the original modulation and that extracted from the test signal after it has passed through the network under test; but for the order of delay distortion which is of interest these differential phase changes are very small, and can readily be masked by spurious phase fluctuations. For example, even when the measuring interval is as large as 1Mc/s, a change of 1mμsec in group-delay corresponds to approximately 0·36° change of phase increment.

Early group-delay measurements had to be made by laborious methods, but during the early 1950s a versatile instrument was developed for the display and measurement of the gain/frequency and group-delay/frequency characteristics of networks[25]. This instrument covered a band of approximately 20Mc/s, centred on 60Mc/s, and the discrimination for delay distortion measurements was about 10^{-9}sec, i.e. 1mμsec. A number of instruments of this design, centred on 60Mc/s, were built; but with C.C.I.R. standardization of 70Mc/s as the centre frequency for broad-band i.f. equipment, the original instrument had to be modified for operation over the band 60 to 80Mc/s.

By the addition of a pair of frequency convertors, i.f. group-delay measuring equipment of this type has been extended to be usable for measurements in the 4 000Mc/s band, and useful information has been obtained on the characteristics of microwave filters and amplifiers.

Amplitude modulation methods of group-delay measurement have also been explored and, in work on 70Mc/s a.m. equipment, a discrimination of about 0·1mμsec has been reached. Group-delay measuring equipment using amplitude modulation must, of course, be restricted in application to passive networks or those including only linear amplifiers, and it is not suitable for use with limiting amplifiers. Further work is still in progress on the development of f.m. group-delay measuring sets for both microwave and intermediate frequencies.

For overall testing of systems, the group-delay measuring equipment may have to be split into two parts and the frequency increment is usually reduced to 100 or 200kc/s. Where looping of the system under test is impracticable and no parallel system is available, the reference frequency required at the receiving end has to be provided by such methods as (a) using a separate crystal controlled oscillator, or (b) extracting it from the system under test via an extremely narrow band filter capable of eliminating the sidebands it has acquired.

Measurement of Non-Linear Distortion

A frequency-division multiplex telephony signal having a large enough number of channels can be simulated with reasonable accuracy by using a band of uniform random noise having the same upper and lower frequency limits. The insertion of a narrow band-stop filter can then provide

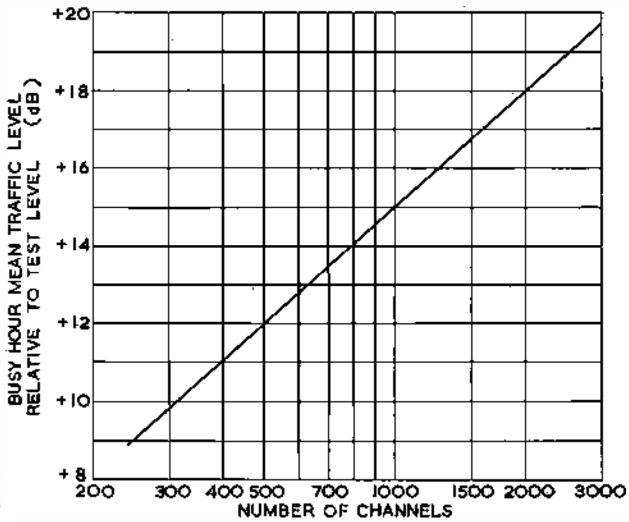


Fig. 7. Noise intermodulation traffic levels and conversion factors. (C.C.I.R. Warsaw 1956)

Conversion factors 4kc/s flat 15·1dB; 3·1kc/s flat 16.2dB; Psophometric 18·7dB

Notes:—Conversion factor is added to the measured noise-power ratio (n.p.r.) to give signal-to-noise ratio. Busy hour mean traffic level = $(-15 + 10 \log n)$dBm, for $N > 240$

the equivalent of an unloaded or quiet channel in an otherwise fully loaded base-band, and at the far end of the system under test a measurement can be made of the total noise and crosstalk appearing in this stop-band[26]. This technique has now been in regular use for several years and has proved to be a most satisfactory and expeditious way of assessing telephony performance of radio or cable systems. The method was accepted in principle by the C.C.I.R. at their 1956 meeting in Warsaw, and for systems of 240 or more channels the level of the test signal to be used was agreed upon. This is shown in Fig. 7.

The generator or sending end of the standard Post Office intermodulation measuring equipment provides bands of noise simulating 120, 240, 600 or 960 channels and incorporates narrow band-stop filters centred on 70, 534, 1 002, 2 438 and 3 886kc/s. The receiver for use at the output of the system under test is crystal controlled and can be switched to any one of these five frequencies: it operates at fixed gain and is preceded by an input attenuator.

The method of making a measurement is simple. The generator is set to the desired number of channels and a flat band of noise is applied at appropriate level to the system under test. The receiver is set to the selected stop-band frequency and its input attenuator adjusted for an arbitrary output level indication. The appropriate band-stop filter is then switched into circuit at the sending end and the receiver input attenuator readjusted for the same indicated level as before. The difference between the two attenuator settings is termed the 'noise-power ratio' (n.p.r.) for that particular loading and that stop-band frequency, and is now widely accepted as an accurate and convenient index of performance. It is important to stress that noise-power ratio should be regarded as a more fundamental and more useful figure than signal-to-noise ratio for assessment or specification of system performance.

Signal-to-noise ratio can be derived from noise-power ratio by adding the appropriate conversion factor under Fig. 7. This conversion factor may be regarded as an allowance for the difference between channel test-tone level and the proportion of the total noise-power loading which is effectively applied to any one channel. For example, to obtain mean busy hour performance data on a 600 channel system, noise loading would be applied at a level of +12·8dB relative to the test-tone level; but when making the upper measurement of a pair for noise-power ratio (band-stop filter out), this noise power is spread evenly from 60kc/s to 2 540kc/s, a total of 2 480kc/s. The noise power in 3·1kc/s is 3·1/2 480 of the total power (i.e. −29dB relative to the total power) and so is +12·8 − 29 = −16·2dB relative to the test-tone level. A conversion factor of 16·2dB must therefore be added to the n.p.r. to get the unweighted signal-to-noise ratio in a 3·1kc/s bandwidth. Psophometric weighting of uniform noise in the standard 300 to 3 400c/s band introduces a further factor of approximately 2·5dB, and the conversion factor which must be added to the n.p.r. to get the weighted signal-to-noise ratio is therefore 16·2 + 2·5 = 18·7dB.

Spectrum Analyser

It is often useful to examine the frequency spectrum of a radio signal and it is desirable that the method adopted should be both accurate and rapid. In 1948 a panoramic receiver was developed which used electro-mechanical scanning to display and to measure signals in the frequency band 55 to 65Mc/s, and which could be extended to other bands with the aid of frequency convertor units.

More recently an improved spectrum analyser has been developed for the display and measurement of signals occupying a larger bandwidth. In this instrument the frequency sweep is produced by varying the field applied to a ferrite core in the inductor of a Colpitt's oscillator. The mean frequency is adjustable in one range from 25 to 140Mc/s and the sweep width is continuously variable from about 1Mc/s up to about 50 per cent of the mean frequency. Alternative 3dB bandwidths of 7kc/s and 35kc/s, a choice of a linear or logarithmic amplitude display and the provision of accurate frequency markers at 10Mc/s or 2Mc/s intervals are other useful features of this compact transportable instrument.

Linearity Test Set for Modulators and Demodulators

Another type of instrument which is essential to progress in the field of broad-band radio-relay systems is one which will check the linearity of a frequency modulator or demodulator. Such instruments are generally arranged to display the derivative of the frequency/voltage or the voltage/frequency characteristic and they are required to be able to measure variations in slope of well under one per cent. Since the dynamic linearity of some frequency modulators and most demodulators varies with base-band frequency, it is important to make this measurement at more than one modulating frequency. Derivative test sets with a fixed test frequency have been in use for many years; these are not really adequate for work on broad-band systems, and more recent Post Office linearity test sets have universally embodied a range of test frequencies. Two-tone testing techniques have also been tried out in instruments which display the change in level of a second or third order intermodulation product as a carrier with two-tone modulation is swept across the frequency band; but it is doubtful whether the additional complications involved are really justified.

Location of Reflections in Waveguide Feeders

The location of mismatches in long waveguide feeders can become a major problem, especially as most radio aerials are erected in exposed sites where high winds and low temperatures make careful mechanical work difficult. An interesting piece of test equipment that will locate mismatches of the order of half per cent in a waveguide feeder has been developed[27]. The principle is akin to that of f.m. radar. A carcinotron oscillator working in the 4 000Mc/s frequency band is regularly swept over a range of up to 300Mc/s to generate a sawtooth of frequency, with a periodicity of 50c/s. The oscillator output is connected to the waveguide through a double-probe detector section, where the beat frequency between the forward wave and any returning reflected wave is detected. The beat frequency is proportional to the distance between the detector and mismatch.

Insertion Loss Measurement

Mention has already been made of the difficulties experienced in measuring the performance of microwave filters with sufficient accuracy. Insertion loss measuring equipment has been developed especially for this purpose. A simple set[21], in which the microwave testing frequency is pulsed at 1kc/s, and using waveguide attentuators, has been in use for many years. A more recently developed set using frequency changers and a piston attenuator at an intermediate frequency is now in use and this is proving a satisfactory precision instrument.

Trends of Development

Performance, reliability and cost per channel are the primary criteria by which a radio relay system must be judged. If the system is to form part of the trunk telephone network, then C.C.I.R. standards of performance are virtually mandatory and reliability and cost per channel remain as the yardsticks of progress. The importance of reliability in large-capacity traffic arteries cannot be stressed too much and its improvement must ever remain a major objective of the designer. Since the cost per channel can be reduced either by lowering the total cost or by increasing the traffic capacity of a system per carrier frequency, two general lines of development become apparent, leading respectively to (a) simpler systems and (b) systems of much higher traffic capacity.

Simplicity is always desirable since it offers not only increased reliability but also lower capital and maintenance costs. If simplicity can be combined with reduced size, and especially with much lower power consumption, building costs and the cost of standby power supplies can be greatly reduced. An h.t. supply of 150 to 180V is adequate for modern i.f. units, and recent developments in power transistors and silicon junction rectifiers offer the possibility of deriving the high voltage required for a permanent-magnet-focused travelling wave valve from this 150 to 180V supply in a simple and efficient manner. The supply voltage and power consumption can be further reduced when transistors suitable for use in 70Mc/s i.f. units are readily available.

As the reliability of the radio equipment is improved, it will be necessary further to improve the reliability of the power supplies, since in many systems the latter is already a limiting factor on the overall reliability.

Travelling-wave valve amplifiers with permanent magnet fields will simplify equipment and save the power required to energize the solenoid. Improved mixers are needed to work with these travelling wave amplifiers. Ferrite isolators will be of great assistance to designers and maintenance engineers in reducing reflection problems and enabling lower degrees of matching to be used in components that have awkward impedance characteristics. The Radio Laboratories are devoting much time to this problem; unfortunately around 4 000Mc/s is one of the more difficult frequency bands for the production of stable ferrite material suitable for this work.

Current techniques of equipment design in the Laboratories provide adequate margins on the C.C.I.R. performance requirements for systems carrying 600 or more telephone channels. With further work on various outstanding problems it may eventually be possible to reach about 1 800 channels (or television plus 600 to 900 channels) within the framework at present agreed of 29Mc/s channel spacing and 70Mc/s i.f.

The combining of several r.f. carriers on to one aerial is already offering a satisfactory way of providing more telephone channels on a route; this technique uses branching filters, by which the various transmitter outputs (each one an r.f. channel) can be combined. Similar branching filters are needed to separate the r.f. channels at the receiving end, or to add or extract channels as required. From this point, which has already been reached, the logical step is to use very wide band aerials capable of dealing with more than one frequency band; for example, one horn aerial system can carry all the r.f. channels from 2 000Mc/s upwards on a major route.

Acknowledgments

An article that surveys the work of a large laboratory organization in even a narrow field necessarily owes a debt to the work of many engineers over a period of many years. It would be invidious to single out any individual by name, but the authors are appreciative of the development work of their colleagues in the broad-band microwave field.

Their thanks are also due to the Engineer-in-Chief of the Post Office for permission to publish this article.

REFERENCES

1. Nancarrow, F. E. Ultra-short Radio Waves and the Cardiff-Weston-super-Mare Radio Link. *P.O. Elect. Engrs' J.* *25*, 303 (1933).
2. Mumford, A. H. The Guernsey-Chaldon Ultra-Short-Wave Radio Telephone Circuit. *P.O. Elect. Engrs' J. 29*, 124 (1936).
3. Gill, A. J. Wireless Section. Chairman's Address. *J. Instn. Elect. Engrs. 84*, 248 (1939).
4. Mumford, A. H. Recent Developments in Communications Engineering. *J. Instn. Elect. Engrs. 93*, Pt. 3, 2 (1947).
5. Sowton, C. W., Hollinghurst, F. Résumé of V.H.F. Point-to-Point Communication. *J. Instn. Elect. Engrs. 94*, Pt. 3A (1947).
6. Merriman, J. H. H., White, R. W. The Application of Frequency Modulation to V.H.F. Multi-Channel Radio Telephony. *J. Instn. Elect. Engrs. 94*, Pt. 2A, 649 (1947).
7. Mumford, A. H., Booth, C. F., White, R. W. The London-Castleton Experimental Radio-Relay System. *P.O. Elect. Engrs' J. 43*, 93 (1950).
8. Corke, R. L., Ephgrave, E. V., Hooper, J., Wray, D. A 4 000Mc/s Radio System for the Transmission of Four Telephony Supergroups (240 circuits). *P.O. Elect. Engrs' J. 50*, Pt. 2, 106 (1957).
9. Froom, R. P., Madder, J. D. C., Hilton, C. G. A 4 000Mc/s Radio System for the Transmission of Four Telephony Supergroups (240 circuits), Part 2. *P.O. Elect. Engrs' J. 50*, Pt. 3, 150 (1957).
10. Bray, W. J. The Travelling-Wave Valve as a Microwave Phase-Modulator and Frequency-Shifter. *J. Instn. Elect. Engrs. 99*, Pt. 3, 15 (1952).
11. Kilvington, T. The London-Isle of Wight Television Link, Stage I. *P.O. Elect. Engrs' J. 48*, Pt. 1, 36 (1955).
12. Hamer, R., Gibbs, C. H. A Broad-Band Intermediate-Frequency Amplifier for use in Frequency-Modulation Microwave Radio Relay Systems. *P.O. Elect. Engrs' J. 50*, 124 (1957).
13. White, R. W., Ravenscroft, I. A. A Frequency Modulator for Broad-Band Radio Relay Systems. *P.O. Elect. Engrs' J. 48*, 108 (1955).
14. Hamer, R., Ravenscroft, I. A. British Patent No. 731592.
15. Ravenscroft, I. A. An Improved Frequency Modulator for Broad-Band Radio Relay Systems. *P.O. Elect. Engrs' J. 50*, 186 (1957).
16. Ravenscroft, I. A. British Patent Application No. 38142/56.
17. Hamer, R., Wilkinson, R. G. A Broad-Band Variable Group-Delay Equalizer. *P.O. Elect. Engrs' J. 50*, 120 (1957).
18. Hamer, R., Wilkinson, R. G. British Patent Application No. 10507/56.
19. Corke, R. L., Hooper, J. Aerials for 4 000Mc/s Radio Links. *P.O. Elect. Engrs' J. 50*, Pt. 3. 178 (1957).
20. Hobbs, J. G. A Pressurised Waveguide System. *P.O. Elect. Engrs' J. 49*, 52 (1956).
21. Floyd, C. F., Rawlinson, W. A. An Introduction to the Principles of Waveguide Transmission, Pt. 2. *P.O. Elect. Engrs' J. 47*, Pt. 3, 157 (1954).
22. Scarbrough, R. J. D., Ferguson, K. H., Searles, A. W. The Generation of Stable Carrier Frequencies in the Range 3 800-4 200 Mc/s. *P.O. Elect. Engrs' J.* To be published.
23. Wray, D. British Patent Application No. 29554/57.
24. Floyd, C. F. An Introduction to Microwave and Waveguide Transmission. *P.O. Elect. Engrs'.* To be published.
25. Whyte, J. S. An instrument for the Measurement and Display of V.H.F. Network Characteristics. *P.O. Elect. Engrs' J. 48*, 81 (1955).
26. White, R. W., Whyte, J. S. Equipment for Measurement of Inter-Channel Crosstalk and Noise on Broad-Band Multi-Channel Telephone Systems. *P.O. Elect. Engrs' J. 48*, 127 (1955).
27. Hooper, J. A Swept-Frequency Method of Locating Faults in Long Waveguide Runs. *P.O. Elect. Engrs' J.* To be published.
28. Tattershall, R. L. O. A High-Accuracy Experimental Cavity Frequency Meter for use at 4 000Mc/s. *P.O. Elect. Engrs' J.* To be published.

G.P.O. Orders First All-Travelling-Wave-Tube Radio Links

The G.P.O. has placed an order with Marconi's Wireless Telegraph Company Ltd for the supply and installation of a single-way u.h.f. radio link between London and a point near Norwich. The link provides two broadband channels primarily intended for the conveyance of television signals, but, if the necessary return channels were provided, they could be used for telephony, each being capable of providing up to 600 high-grade telephone channels simultaneously.

The contract is particularly interesting on three counts: Not only will the equipment handle black-and-white television signals, it has also been designed with a view to transmitting 405- or 625-line colour television signals, should this be required. It is the first order for all-travelling-wave-tube equipment to be placed for use in this country, and will be engineered to C.C.I.R. and C.C.I.T.T. standards for the provision of the 600-telephone-channel facility.

The route will run from the Museum Exchange in London, via repeater stations at Ongar, Sibleys and Ousden, to the terminal station between Norwich and Ipswich. Marconi HM 200 equipment will be used at the terminals, with HM 250 repeaters at the intermediate stations.

It is hoped to bring the link into operation in the spring of next year.

The type HM 200 equipment is designed to meet the requirements of C.C.I.R. and the British Post Master General for long distance international wide-band communications systems. Both television and multi-channel telephony circuits can be carried, each HM 200 radio circuit being capable of carrying either 600 telephone channels or one television channel of 405, 525 or 625 line standards.

A comprehensive supervisory system is available which includes, for the provision of urgent and non-urgent fault alarms, a party line speech channel between all stations, and interrogation of repeaters to identify fault conditions.

Since the HM 200 radio relay equipment is designed to be used in systems employing up to six parallel radio paths providing a total capacity of 6 simultaneous television circuits or 3 600 telephone channels, the supervisory system has been designed as a completely separate entity, which can be used over a separate bearer channel to control a multi-path radio relay system. The use of this facility avoids the necessity for providing independent supervisory systems for each radio circuit. This, in turn, reduces the capital cost of the relay system, and avoids unnecessary duplication of equipment. However, to provide the maximum degree of flexibility, the supervisory circuits can be inserted below the carrier baseband, on a single or twin path multi-channel radio relay system, thus avoiding the need for a separate bearer channel.

A single terminal will occupy three racks of equipment, one for the transmitter, one for the receiver and one for the modulator and demodulator circuits.

SURVEYING FOR MICROWAVE RELAY SYSTEMS

By L. E. Strazza*, B.Sc., Assoc.Brit.I.R.E., and R. C. S. Joyce*

The overall performance of well-planned microwave radio relay links can be predicted with a high degree of reliability; good planning also enables overall system costs to be estimated with considerable accuracy. This article describes the role of surveying in planning telecommunications systems of this kind. Surveys in well-mapped regions may only involve site and path inspections but where the accuracy of path profiles plotted from maps is doubtful path tests are made. The tests may be of short duration if, as is usually the case, they are intended to confirm a calculated mean path loss and hence to confirm tower heights, aerial gains and similar system parameters. Long-term tests are made when the path profile is unfavourable and statistical information on fading is required. Typical methods of testing are described.

THE aim in planning fixed point to point microwave radio links, whether for the transmission of television or multi-circuit telephony signals, is to make an effective compromise between the optimum performance and reliability that can be achieved and economy in capital outlay and maintenance costs.

Since the overall performance of a line of sight microwave system is determined partly by the parameters of the equipment itself and partly by the manner in which the equipment is applied to the system, it is imperative that both the route and installation of the equipment be carefully planned.

It is generally known that microwave radio systems require a line-of-sight path between transmitting and receiving aerials but in practice the received signal is dependent on many other factors.

These factors are as follows:

(a) *There must be adequate clearance between the line of sight and the intervening terrain*

Merely providing a line-of-sight path is not enough since a significant portion of the radiated signal energy does not follow a straight line between transmitting and receiving aerials but occupies a volume embracing several Fresnel zones. A path between two stations must therefore be planned to ensure that there is at least 0·6 × 1st Fresnel zone clearance above all obstacles in the radio path under adverse atmospheric conditions in order to obtain an average path attenuation approximating to the free space loss. A table relating 1st Fresnel zone clearance and path length for propagation in the 2 000Mc/s band is given in the Appendix.

(b) *Multi-path reception*

Certain types of terrain such as water or flat desert areas act as good reflectors of microwaves with the result that the receiving aerial receives the direct signal together with signals that have been reflected. These signals, by virtue of the fact that they have travelled over paths of different lengths, arrive in-phase or out-of-phase to partly re-inforce or cancel the received signal, depending on whether the direct path clearance over the reflection point is equal to an odd or even number Fresnel zone, there being a one half wavelength difference in the reflected signal path between each successive zone. Sites and aerial heights should if possible be chosen so that the reflected signals due to 2nd or higher Fresnel zones are prevented from reaching the receiving aerial by some natural obstruction. It should be borne in mind that under varying atmospheric conditions which cause effective earth bulging or flattening the reflected signal may pass through several Fresnel zones which may cause the received signal level to pass through successive minima if the clearance between the direct path and reflection path is large. Multi-path signal reflections can be obtained from a stratum of air over the radio path in which the dielectric gradient abruptly changes, such as is caused by a temperature inversion. This effect when investigated, sometimes coincides with meteorological con-

* *The General Electric Co. Ltd.*

ditions obtaining at that particular time and the cause of fading can be diagnosed. In a twin path scheme operating with different frequency allocations fading on each path is usually sufficiently uncorrelated to ensure adequate continuity of service although there is a possibility of some echo distortion occurring during a severe fade.

(c) *Fading*

Fading, in addition to being caused by reflections giving rise to multi-path signals, is also due to variations in the earth's atmosphere which cause refraction or scattering of the radio beam. Fading due to these causes is more prevalent in those parts of the world where the air becomes quite still or very humid. Experience has shown that in some territories fading can be estimated with a reasonable degree of accuracy while in other areas prolonged propagation testing is the only way in which accurate data can be obtained.

(d) *Overshoot*

Under certain atmospheric conditions signals may be transmitted over distances considerably in excess of the normal repeater section and care must be taken that significant signals cannot be received at stations other than the adjacent stations.

This 'overshoot' reception as it is called is very probable when the relay stations are sited more or less in a straight line. Therefore stations should be sited wherever possible to form a zig-zag route. If this is not possible then the transmitted frequencies must be selected so that at any station the signal received from the adjacent station and any overshoot signal received are at different frequencies.

Preliminary Stages of Planning—Map Study

If large scale maps of the territory are available then the microwave relay system can be tentatively planned before a visit is made to the sites.

Tentative sites are located on the maps so that the stations on a multi-hop system are of the order of 30 miles apart. In the case of systems with only a small number of hops, stations may be located at slightly wider intervals but in the case of long systems the cumulative effects of fading on several sections at the same time makes shorter sections desirable.

It must be remembered however that in undeveloped country the addition of one or more relay stations constitutes an increased maintenance hazard besides increasing the capital cost of the project.

In addition to ensuring that the stations fulfil the propagation requirements, serious consideration has also to be given to the necessity for siting stations so that access to them is as economic and practicable as possible, both during the initial erection of the station and during its operational life. Therefore a compromise has to be sought.

Once the tentative site locations have been decided profiles of each radio path are prepared on graph paper based on true earth radius in order to ensure that there is at least $0{\cdot}6 \times 1^{st}$ Fresnel zone clearance over any obstacles in the path. Some authorities employ a scale of $4/3 \times$ true earth radius for the profile, but the authors consider it is better to use the more conservative scale of $k = 1$. This will give some extra margin over buildings or trees in the path which cannot always be accounted for by map study. If a scale of $k = 4/3$ is used path clearances should be checked for lower values of k. Information should also be sought at this stage on the heights of any buildings or trees that could obstruct the path. This information may be available to the contractor if he has already had work in the territory or it may be provided by the customer. If it is not known then it must be eventually determined by a visit to the sites.

The preliminary planning enables a rough assessment to be made of system performance and also a budgetary estimate of system costs to be made.

In cases where accurately contoured maps of the territory do not exist or where maps are known to contain inaccuracies an aerial photographic survey is essential and there are contractors specializing in this type of work. This type of survey is however expensive and must be used as sparingly as possible.

Visual Survey of Sites

At this stage the sites should be visited to determine whether visual inspection of the sites will suffice or whether some or all the paths will have to be electrically tested.

The visual survey consists of a visit to each site by a radio engineer and possibly a civil engineer to obtain data about the site itself, how the site can be reached from the nearest road, nature of local obstructions, possibility of reflections from surrounding terrain and all data necessary for the construction of a tower and buildings.

If the radio system is to be located in undeveloped country the task of visiting prospective sites may have to be undertaken partly on foot and this can be both arduous and time consuming. In order to obtain an overall picture of a network as rapidly as possible the use of a helicopter has many advantages. From the air disused and long forgotten tracks, not shown on survey maps, can be quickly identified and this often helps to reduce the costs of building access roads to the site. Also if sites are visited in route sequence the 'plane will fly along the propagation path between two sites and therefore a rough check may be made on obstructions. The photograph on page 262 was taken from a helicopter approaching a proposed site in undeveloped country and it will be noted that topographical features of the site are clearly discernible.

Helicopters, however, cannot be used with safety in some mountainous areas since air turbulence makes them difficult to control.

To summarize, the ground survey should yield answers to the following questions.

(a) Distance from the site selected for the radio station to any existing road and an estimate for building an access road.

(b) Availability of a reliable commercial power supply. If there is a supply near at hand the cost of extending this to the site should be estimated by the power utility. On the other hand if there is no supply or an unreliable one then either dual or standby diesel-electric sets respectively will have to be installed and this will have to be considered when planning the building of access roads. Whereas maintenance personnel can visit the station on a steep graded road by means of a four-wheel drive station wagon the provisioning of fuel oil necessitates an easier gradient and this may lengthen and consequently increase the cost of the road. Sometimes fuel oil is piped from an easily accessible spot to the station itself or, alternatively, the generating plant is housed where it can be reached easily and a high voltage line is run to the repeater station itself; both these schemes simplify the road construction problems but, of course, they have their own drawbacks.

(c) Site conditions. An estimate of the amount of deforestation and levelling that would have to be done before the station and tower could be erected.

(d) Examination of the terrain in the directions of propagation to ensure that the 'lines-of-fire' are free from obstacles which could diffract or scatter the radio signals:

in particular freedom from rising ground formations close to the site or trees which in summer are clothed in dense foliage and which in time could grow high enough to obstruct the path

(e) The type of sub-soil in order to assess the amount of work and cost of preparing tower and building foundations.

(f) Local climatic information. The prevalence of icing and high winds will influence the type and strength of tower which will be needed.

Buildings may require heating or air conditioning depending on the range of ambient air temperatures experienced. Standby diesel-electric generating plant must start up immediately when called upon to do so and the choice must be made between air-cooled and water-cooled engines. Air cooled are preferable at sites with no water supply but in very cold regions the water-cooled engine has the advantage that the water can be warmed by means of a small heater immersed in the jacket to ensure easy starting.

Propagation Testing

Propagation testing is resorted to when confirmation is required that a line of sight path exists or when data is required on fading conditions.

Line of sight confirmation tests need only extend over a period of 24 to 72 hours during which the received signal is measured and the path loss compared with the theoretical free-space loss of the path.

When data is required on fading conditions the propagation tests must be performed over a very much longer period which should include that period of the year when fading can be expected to be most serious. This can be deduced in some cases from local weather records.

Propagation tests, whether short-term or long-term are usually made in the following way. Two stations are established, one at each of two adjacent sites along the route to be surveyed. One station is designated the 'transmit station' and the other is designated the 'receive station'. The sending station is equipped with a transmitter operating on a frequency which is representative of those proposed for the final system. The transmitter is usually a low-power device (that is to say the output power is in the region of 0·5 to 2·0W) and the signal may be c.w., amplitude modulated or frequency modulated—each type of modulation has advantages as will be explained later. Fig. 1 is the block schematic of a typical propagation test transmitter producing about 0·75W at 2 000Mc/s and in which the carrier is 100 per cent amplitude modulated by a 1 : 1 square wave with a recurrence frequency of 5kc/s. It will be noted that it consists of a microwave triode oscillator stage provided with an electro-mechanical a.f.c. system and that the h.t. supply to the unit is switched at 5kc/s by a series modulator. The outgoing signal is monitored by a directional coupler and provision is made for connecting a recording milliammeter so that a permanent record of the output level stability may be produced. The output from the transmitter is connected by means of an appropriate low-loss coaxial feeder to the transmitting aerial, which is usually a paraboloidal reflector driven by a dipole and reflector feed unit or by a broad band launching unit.

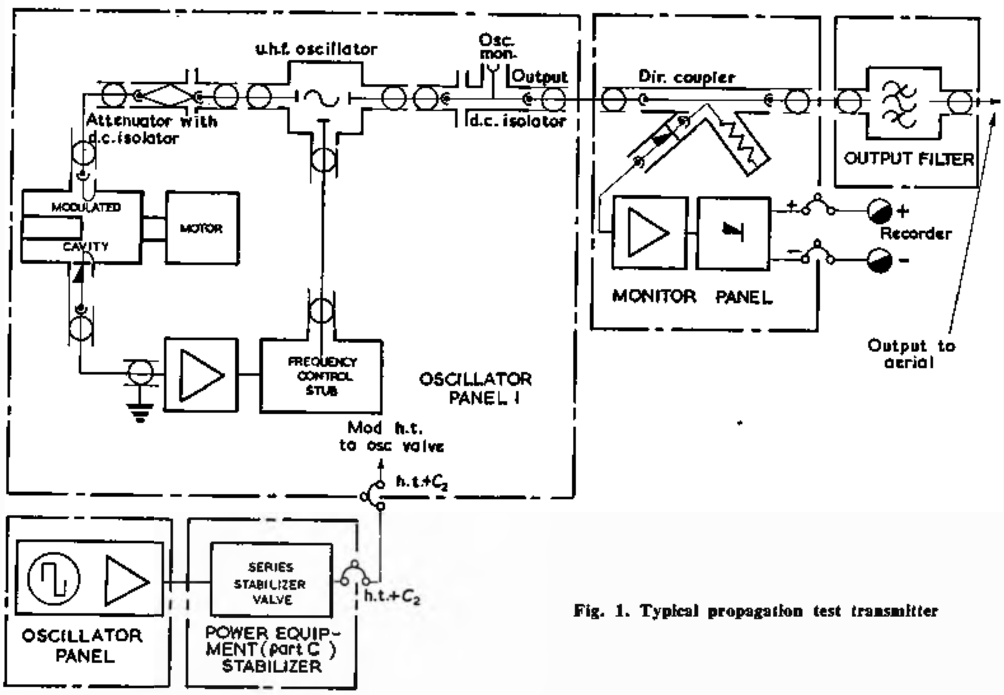


Fig. 1. Typical propagation test transmitter

The receiving station is provided with a receiving system consisting of an aerial (usually identical with the transmitting aerial) a feeder and a receiver. The receiver is generally a fairly conventional superheterodyne receiver employing a low intermediate frequency so that negative feedback can be introduced to stabilize the receiver gain. Fig. 2 is the block schematic of such a receiver in which the i.f. is 2Mc/s and the i.f. amplifier consists of a series of video amplifying 'feedback triples'. Delayed a.g.c. is provided so that a linear relation exists between the input signal level expressed logarithmically and the d.c. output from the receiver which is normally connected to a recording milliammeter. It is usually desirable to record fades as deep as 30 to 40dB and the mean signal level may be abnormally low relative to the level required in the final system, because of the low transmitted power, high feeder losses and low aerial gains, of the survey equipment. On this account the receivers used have a narrow i.f. bandwidth in order to improve the signal-to-noise ratio and hence enable low level signals to be detected.

Other types of receiver are known; for example one authority employs a simple crystal detector followed by a very high gain narrow band audio amplifier. Others, using f.m. transmitters, employ wideband receivers and transmit pulses which, when viewed on a suitable indicating unit can provide information concerning the amplitude and delay characteristics of indirect signals.

Measurement of the Path Loss

Let the level of the signal be P_1 dBw at the transmitter output and let P_2 dBw be the signal level at the input to the receiver. Both these quantities can readily be measured: P_1 by means of a microwave power meter, usually of the bolometer type and P_2 by calibrating the receiver with a microwave signal generator. It is now an easy matter to calculate the path loss L_P (dB) as follows:

The effective power radiated from the transmitting station is P_T where:

$$P_T = P_1 + G_1 - L_1 \,\mathrm{dBw}$$

where:

P_T = Effective power radiated (dBw)
P_1 = Transmitter output level (dBw)
G_1 = Gain of transmitter aerial (dB)
L_1 = Loss in transmitter feeder (dB)

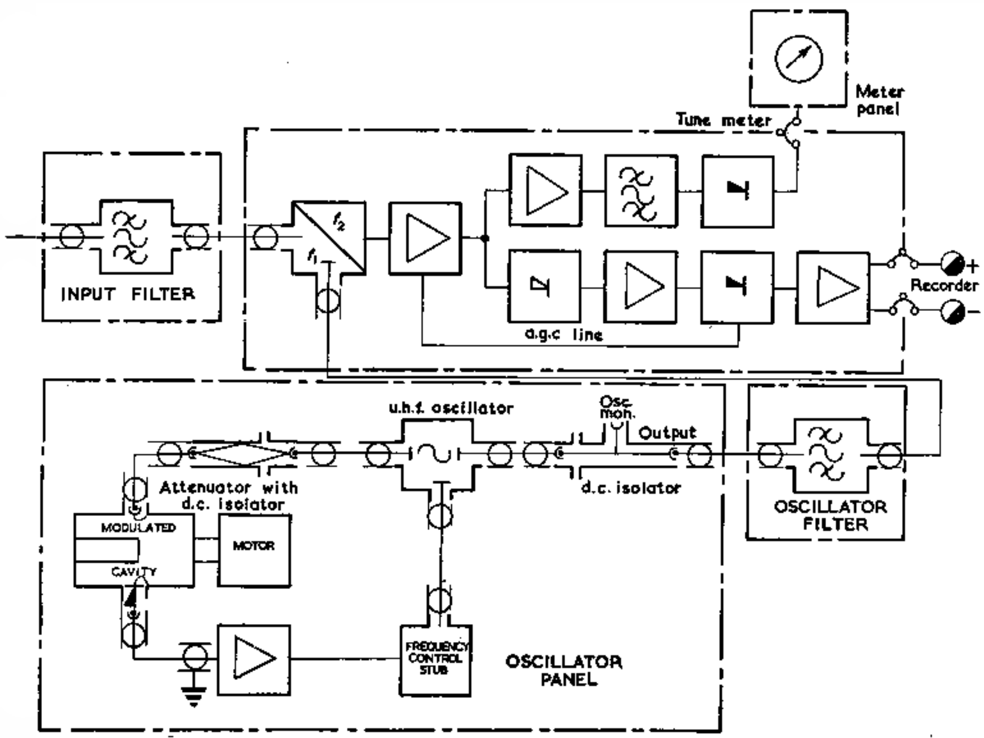


Fig. 2. Typical propagation test receiver

Similarly the effective power received is P_R where:

$$P_R = P_2 - G_2 + L_2 \mathrm{dBw}$$

where:

P_R = Effective power received (dBw)
P_2 = Receive power input level (dBw)
G_2 = Gain of receiver aerial (dB)
L_2 = Loss in receiver feeder (dB)

The path loss L_P is defined as the attenuation between the two stations when isotropic aerials are used, thus:

$$L_P = P_T - P_R \,\mathrm{dB}$$

$$L_P = P_1 - P_2 + (G_1 + G_2) - (L_1 + L_2) \,\mathrm{dB}$$

In the foregoing expression P_1 and P_2 are measured, G_1, G_2, L_1 and L_2 are known constants and hence L_P can be evaluated.

If fading occurs L_P will vary and hence P_2 will vary; the variations of P_2 will be recorded by the recording milliammeter and provided its speed of response and the chart speed are appropriate for the type of fading to be investigated the depth of fading can be determined by calibrating the receiving equipment. Experience shows that if the equipment described in the two drawings is operated from stable a.c. supplies it can run unattended for many weeks without needing re-calibration; this fact is of great practical importance when making long term tests.

Fading analysers are useful adjuncts or even alternatives to chart recorders. These devices accept a signal from the receiver which is linearly proportional to the incoming signal level. By means of a series of say six to eight electronic gates corresponding to six to eight predetermined signal levels (say at 5dB intervals) and of devices indicating elapsed time and frequency of occurrence it is possible to obtain directly an immediate and continuous analysis of the variations of the path loss so that it is easy to produce the familiar fading versus percentage time curves required for predicting system reliability. Facilities are sometimes provided for automatically photographing the time-indicating devices (clocks or counters) so that loss variations over any desired interval of time can be obtained without manning the station continuously. In all this work of course the recorder associated with the transmitter ensures that apparent path loss variations due to variations or interruptions of the transmitter output can be allowed for.

Aerials and Supports for Surveys

One of the greatest problems besetting the survey engineer is the provision of suitable aerials and supporting structures for surveys. On the one hand the system planner may require aerial heights of up to 400ft and will prefer that an aerial of 'system size' should be used for the tests so that the under-path terrain and obstacles are illuminated to the usual extent while on the other hand local conditions may make either or both ideals impossible to achieve on grounds of physical difficulty, cost, time or all three!

It is usually considered important to make so-called height-gain measurements; these are in fact path loss measurements in which the heights of the transmitting and receiving aerials are varied in turn (preferably in a smooth and continuous fashion rather than in discrete steps so that heights corresponding to signal minima (arising from the destructive interference of direct and indirect signals) may be avoided by accurately known amounts. This requirement is a difficult one to fulfil especially if taken in conjunction with the first two (i.e. of great heights and large aerials). The use of balloons to support the aerials is usually impossible owing to the great directivity of large microwave aerials and the impossibility of stabilizing a balloon. Schemes involving servo-stabilized aerials mounted on helicopters have been tried.

At the present time several authorities employ small paraboloids (say 5 to 6ft in diameter) mounted on trolleys which can be run up and down temporary stayed towers or masts by means of tracks and a winch; usually the bulk of the radio equipment is also mounted on the trolley so that only power and monitoring cables are required between the trolley and the ground and the usually cumbersome long low-loss r.f. feeders can be dispensed with.

Survey Equipment

In addition to the basic measuring equipment which is the *raison d'etre* of the survey many other things have to be provided and the survey engineer's art lies in knowing what to provide for each project, taking into account many factors such as climate, topography, transport facilities, availability of labour, the duration of the tests, the degree of development in the territory and so on.

In certain cases it is posible to install all the measuring equipment in a comfortable go-anywhere vehicle in which may also be housed communications equipment (reliable interstation communication is indispensable), workshop facilities, meteorological instruments and crew amenities. Aerials and mast equipment and generating plant and camp equipment may be closely associated with the vehicles and the whole caravan is thus completely mobile and can survey routes at a fast rate.

The capital cost of such a caravan is of course high and cannot often be justified. It is more usual to break the equipment down into small easily-handled units which can be transported by any reasonable local means: it should be remembered that in the undeveloped regions the ultimate means of transporting equipment to a desirable mountain peak may be the human being.

Other Means of Surveying Routes

It is sometimes possible to make visual observations between stations to verify that a line of sight exists and to check that the clearances are adequate. Signalling lamps, theodolites and other devices play their part in activities of this kind.

If an aircraft flies over a path at constant height above sea level (by means of a barometric altimeter) and if the aircraft to ground distance is measured by a radio altimeter or by a radar then an accurate path profile can be produced directly and automatically. This method has been tried in un-mapped regions but, of course, it has its difficulties—such as the need for ground control points, good weather and aircraft servicing facilities.

TABLE 1

DISTANCE d (MILES)	PROPAGATION LOSS (DECIBELS)	DISTANCE d (MILES)	PROPAGATION LOSS (DECIBELS)
6	118·1	41	134·8
7	119·5	42	135·0
8	120·6	43	135·2
9	121·7	44	135·4
10	122·6	45	135·6
11	123·4	46	135·8
12	124·2	47	136·0
13	124·9	48	136·2
14	125·5	49	136·4
15	126·1	50	136·6
16	126·7	51	136·7
17	127·2	52	136·9
18	127·7	53	137·1
19	128·2	54	137·2
20	128·6	55	137·4
21	129·0	56	137·5
22	129·4	57	137·7
23	129·8	58	137·8
24	130·2	59	138·0
25	130·5	60	138·1
26	130·9	61	138·3
27	131.2	62	138·4
28	131·5	63	138·6
29	131·8	64	138·8
30	132·1	65	138·8
31	132·4	66	139·0
32	132·7	67	139·1
33	132·9	68	139·3
34	133·2	69	139·4
35	133·5	70	139·5
36	133·7	71	139·6
37	133·9	72	139·7
38	134·2	73	139·8
39	134·4	74	140·0
40	134·6	75	140·1

Acknowledgments

The authors wish to express their thanks to the manage-

TABLE 2

DISTANCE TX–RX (MILES)	HALF-DISTANCE CLEARANCE (FEET)	QUARTER-DISTANCE CLEARANCE (FEET)	ONE-EIGHTH-DISTANCE CLEARANCE (FEET)
10	80·6	69·8	53·3
11	84·5	73·2	55·9
12	88·3	76·4	58·4
13	91·9	79·6	60·8
14	95·4	82·6	63·1
15	98·7	85·5	65·3
16	101·9	88·3	67·4
17	105·1	91·0	69·5
18	108·1	93·6	71·5
19	111·1	96·2	73·5
20	114·0	98·7	75·4
21	116·8	101·2	77·2
22	119·5	103·5	79·0
23	122·2	105·8	80·8
24	124·9	108·1	82·5
25	127·5	110·4	84·2
26	130·0	112·5	85·9
27	132·4	114·7	87·6
28	134·8	116·8	89·2
29	137·2	118·8	90·8
30	139·6	120·9	92·3
31	141·9	122·9	93·9
32	144·1	124·9	95·4
33	146·4	126·8	96·9
34	148·6	128·7	98·3
35	150·8	130·6	99·7
36	153·0	132·4	101·1
37	155·0	134·2	102·5
38	157·1	136·0	103·9
39	159·1	137·8	105·3
40	161·2	139·6	106·6
41	163·2	141·3	107·9
42	165·2	143·0	109·2
43	167·1	144·7	110·5
44	169·0	146·4	111·8
45	171·0	148·0	113·0
46	172·9	149·7	114·3
47	174·7	151·3	115·6
48	176·6	152·9	116·8
49	178·4	154·5	118·0
50	180·2	156·1	119·2
51	182·0	157·6	120·4
52	183·7	159·1	121·5
53	185·5	160·7	122·7
54	187·3	162·2	123·9
55	189·0	163·6	125·0
56	190·7	165·1	126·1
57	192·4	166·6	127·2
58	194·1	168·1	128·3
59	195·8	169·5	129·4
60	197·4	170·9	130·5

ment of the General Electric Co. Ltd, Telephone Works, Coventry, for permission to publish this article.

APPENDIX 1

PROPAGATION LOSS, AT 2 000Mc/s BETWEEN ISOTROPIC RADIATORS

The free space propagation loss is given by $(4\pi d/\lambda)^2$ where d is the path length between the transmitter and receiver, and λ is the wavelength. d and λ are expressed in the same units of length.

An alternative expression is:

$$\text{Propagation Loss} = (20 \log \frac{\text{Distance in miles}}{\text{Wavelength in cm}} + 126{\cdot}1)\text{dB}$$

For a frequency of 2 000Mc/s this becomes:

$$\text{Propagation Loss} = 102{\cdot}6 + 20 \log (\text{distance in miles}) \text{ dB}$$

The free space propagation loss at 2 000Mc/s for different path lengths is shown in Table 1.

APPENDIX 2

FRESNEL ZONE CLEARANCES AT 2 000Mc/s

The first Fresnel zone is an envelope bounded by points for which the transmission path from transmitter to receiver is greater by one half wavelength than the direct path. Experience in system planning and field tests has shown that approximately free space transmission is obtained with $0{\cdot}6 \times 1^{st}$ Fresnel zone clearance.

The radius of the first Fresnel zone is given by

$$R^2 = \frac{d_1 d_2 \lambda}{d}$$ where all quantities are in the same units.

and d_1 = length of direct path from point above reflection point to transmitter.

d_2 = length of direct path from point above reflection point to receiver.

d = length of direct path between transmitter and receiver.

$$\text{or } H = 2\,280 \sqrt{\left[\frac{A \times B}{(A + B)f}\right]} \text{ feet}$$

where A = length of direct path from point above reflection point to transmitter in miles.

B = length of direct path from point above reflection point to receiver in miles.

f = frequency in Mc/s.

Table 2 gives sufficient points to draw a 1^{st} Fresnel zone envelope at 2 000Mc/s, which is usually sufficiently accurate to cover the band 1 800 to 2 200Mc/s.

Subsequent Fresnel zone clearances are given by multiplying the values by $\sqrt{2}$, $\sqrt{3}$ etc. for the second, third Fresnel zones and so on.

REFERENCES

1. EGLI, J. J. U.H.F. Radio-Relay System Engineering. *Proc. Inst. Radio Engrs.* *41*, 115 (1953).
2. CAMPBELL, R. D. Path Testing for Microwave-Radio Routes. *Commun. and Electronics*, No. 7, 326 (1953).
3. BULLINGTON, K. Radio Propagation at Frequencies Above 30Mc/s. *Proc. nat. Electronics Conference, Chicago.* 2, 130 (1947).
4. VON HAGEN, W., MACDIARMID, A. N., GOLDENBURG, L. V. Path Testing of the Trans-Canada TD-2 Route. I.R.E. Convention, Toronto, Canada. Paper 49 (1957).

Microwave Radio Toll Systems

By E. W. Anderson*

The function of the radio link in telephone networks is outlined and some of the problems requiring solution in the design of a short haul or toll system carrying one supergroup are discussed. In particular a solution to the basic defect of non-linearity, by means of overall feedback, in a frequency modulated system is described. This results in an overall simplification of the repeatered circuit.

A TELEPHONE network may be divided into three categories, the subscriber's line, the toll connexion between local exchanges and the main trunk system. The actual channel capacity and length of these interconnexions depends on the extent and density of the system. The trunk system considered as a standard by C.C.I.T.T. for the purpose of specifying a performance extends for 1 500 miles, made up of nine main sections, capable of carrying six hundred or even eighteen hundred channels. Spur links to this type of trunk carrying toll traffic to outlying towns might be considered to have a capacity of about sixty channels and a range of one hundred to two hundred miles, and this article describes the functions and design problems of equipment for the latter type of system. Spurs of greater importance assume the character of secondary trunk systems for which provision is made in the general frequency allocations for radio networks.

It will be apparent that the integration of line and radio to fulfil these two latter functions immediately places restrictions on the radio equipment. The established multi-channel carrier-telephone practice should be equally applicable to radio circuits, in fact the radio link should be indistinguishable from the equivalent repeatered cable section. This implies comparable bandwidths, stability and freedom from noise. The choice of equipment, whether line or radio, should be made to provide the subscriber with an efficient service at a reasonable cost. This includes not only the cost of the equipment, but also that of installation, buildings, roads, power lines, cable ducts, long term maintenance and possible expansion. The assessment of the efficiency and quality of service or circuit reliability is more difficult. This can only be evaluated by extended trials and finally by operation, though past experience is an invaluable guide. Unreliability has as its immediate result, loss of revenue, prestige and added cost of maintenance.

Clearly the higher performance and reliability required of the trunk system together with the larger channel capacity means that the cost rises steeply. On the other hand, it is equally important to cater for an expected expansion without re-incurring too much of the basic cost. One assumes that this type of equipment will have a use-

* *Marconi's Wireless Telegraph Co. Ltd.*

ful life of twenty years, comparable with that of the associated channelling equipment.

The effort required to design an equipment which is consistent, reliable and easy to maintain precludes the possibility of a wide variety of designs. Standardization increases the scale of production permitting more efficient production methods, lower costs and at the same time, maintenance in the field is simplified. The system to be described, apart from conforming to mechanical standards established for this type of equipment, has a considerable degree of uniformity within itself.

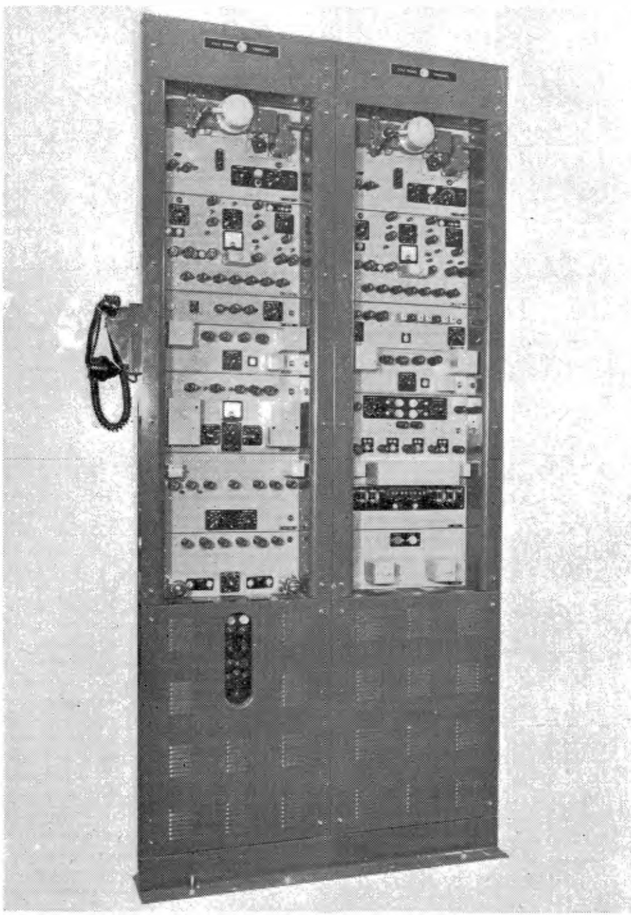


Fig. 1. Typical s.h.f. terminal equipment

Frequency Considerations

For many years the demand for radio links has been fairly well satisfied by a series of v.h.f. equipments, particularly in less developed areas. A situation has now been reached in many territories where difficulty is experienced in allocating frequencies for new systems to cope with traffic expansion and likewise the congestion of the v.h.f. band has precluded the setting up of wide band trunk systems. Also at some densely populated terminals the level of static is too high to allow the desired signal-to-noise ratio to be achieved.

To make use of a wider and less exploited part of the spectrum a system of toll quality in the allocated band of 3 900Mc/s to 4 800Mc/s is appropriate. At these frequencies, the level of interference is very low, but visual paths between stations are required in view of the high attenuation which results from even small obstructions. Fading in this band is generally more pronounced than at v.h.f. particularly over water where ducting and surface reflection can drastically modify the free-space conditions. At the same time, the Fresnel zone clearance is much smaller and the elevation above ground level to give free-space conditions is correspondingly reduced.

A Typical System

A terminal of a typical equipment fulfilling this need is illustrated in Fig. 1. This particular system has a capacity of one supergroup of sixty standard 4kc/s channels with a satisfactory performance over ten hops, totalling about three hundred miles.

Transmitter Design

Owing to the difficulty in providing amplification at these frequencies, a coaxial velocity-modulated valve oscillator, capable of direct frequency-modulation is used, giving an output in excess of half a watt. This provides an adequate system performance when used with 4ft diameter parabolic aerials or their equivalent having a power gain of about a thousand-fold.

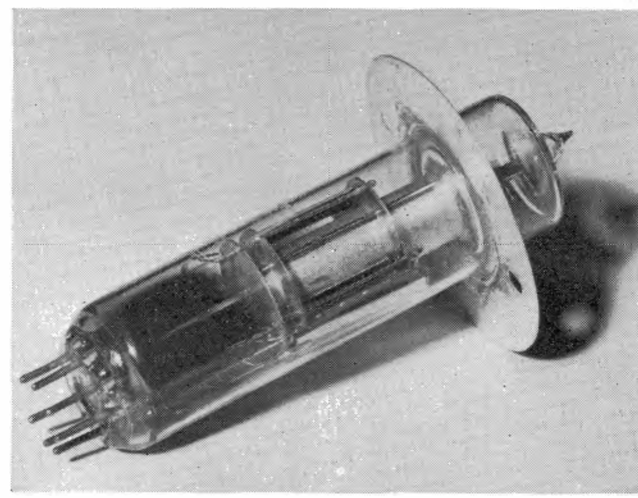


Fig. 2. Heil tube suitable for use at 4 000Mc/s

The acceptance of a group, or super-group, of carrier telephone channels restricts the designer to the use of amplitude or frequency-modulation. The advantages of the latter are well known, and for 60 channels the recommended C.C.I.R. channel deviation of 200kc/s r.m.s. is adequate.

The transmission of multi-channel information requires an extremely linear modulation characteristic to avoid crosstalk between channels. Although the velocity-modulated oscillator is inherently fairly linear, an oscillator when coupled to a long transmission line which is not terminated by a purely resistive load is affected by the phase of the mismatch. Since it is impossible to design a feeder and aerial with such a characteristic over a wide frequency band and maintain it so under varying weather conditions, a directly modulated system is not satisfactory.

A suitable valve for this application is a Heil tube, which consists of a short-circuited coaxial line from which the inner projects to form a radiator (Fig. 2). A slot is cut along the axis. Through both the outer and inner conductors. A cathode and grid assembly is placed at one side of the slot and an anode at the other, so that a flat beam of electrons, focused by a magnet, is passed through the coaxial assembly (see Fig. 3).

The potentials are arranged so that the electrons are accelerated in the space between the cathode and resonator.

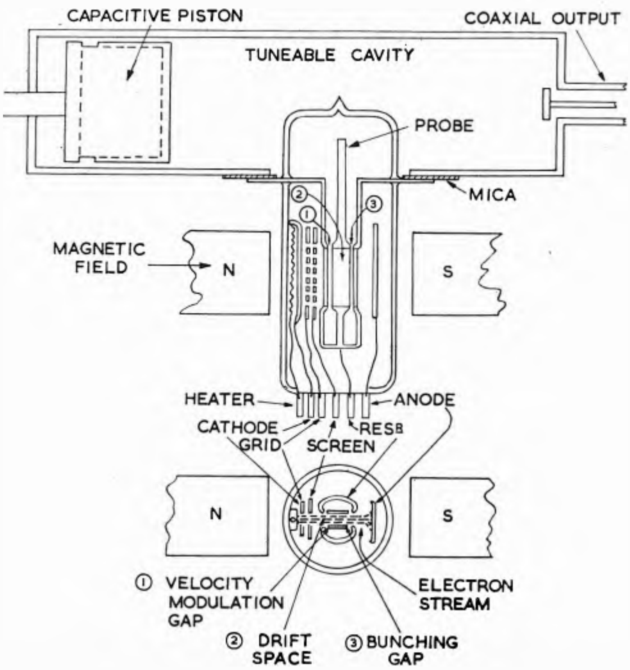


Fig. 3. Diagrammatic representation of velocity-modulated oscillator

The inner line is thick enough to act as a constant potential drift space, equivalent in this case to nine quarter-wavelengths at the operating frequency, allowing electrons with varying velocities in the first gap to bunch before entering the second gap where energy is transferred from the beam to the coaxial line. Some of this energy is expended in modulating the electron velocity in the first gap, maintaining oscillation, while the remainder is radiated from the inner conductor. This is coupled to a tuneable resonator enabling the natural resonance of the system to be controlled and the power coupled into the load to be adjusted. To obtain the maximum output the electron beam voltage must be adjusted so that the bunches arrive in synchronism with the natural resonance of the system. If then the resonator potential is modulated the rate of arrival of the bunches is varied and the frequency of oscillation varied in sympathy.

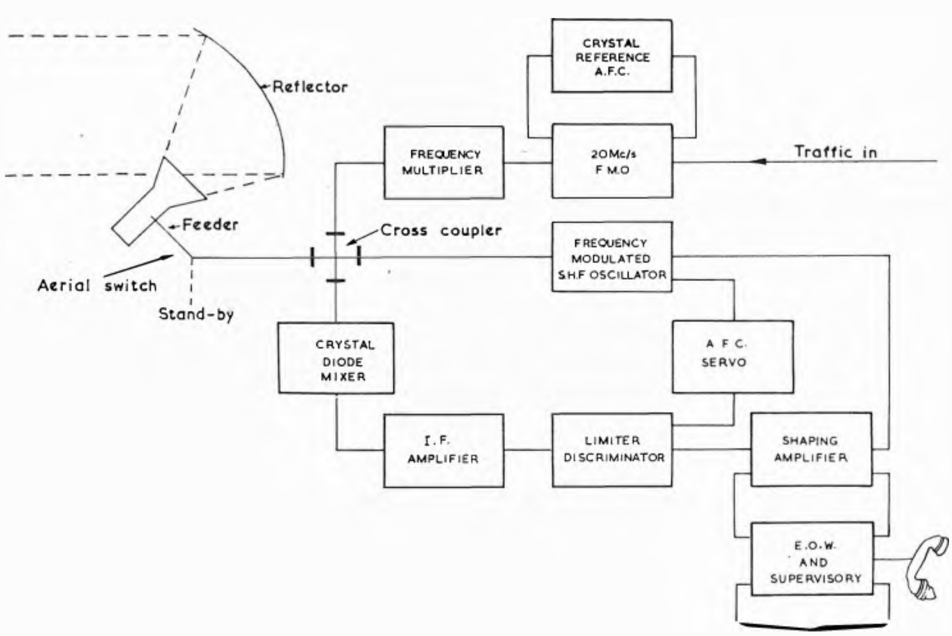
The greater the amount of energy transferred to the load, that is the lower the Q of the tuned circuit, the greater the change of frequency for a given voltage change; excessive load will. however, prevent oscillation. To achieve the desired isolation of the oscillator characteristics, modulation is applied indirectly in a manner shown in Fig. 4. A frequency modulated oscillator operating at approximately 20Mc/s drives a series of frequency multipliers to give a final frequency of approximately 180Mc/s. The output is injected into a silicon diode mounted in a section of waveguide which generates a band of harmonics, some of which propagate along the guide. One of these is heterodyned by the microwave oscillator to produce a 70Mc/s intermediate frequency which is amplified and demodulated, reproducing the original signal used to modulate the 20Mc/s oscillator. This output is applied to the resonator of the microwave oscillator in the correct phase to cause its frequency to deviate in sympathy with that of the harmonic selected. As the final deviation of the transmitted carrier is 200kc/s per channel and the selected harmonic is approximately two hundred times the frequency of the 20Mc/s oscillator the deviation is multiplied proportionately; the basic oscillator therefore is deviated only 1kc/s per channel and is extremely linear. The mid-frequency of the oscillator is stabilized against a quartz crystal so that the harmonic is virtually crystal controlled and has a very linear modulation characteristic.

The microwave oscillator, i.f. amplifier, and discriminator form a closed loop with the characteristics of an amplifier with negative feedback. Analysis will show that depending on the gain of this loop, the deviation of the signal in the i.f. amplifier will be decreased and any non-linearity of the modulation characteristic of the oscillator will also be reduced. In a typical system 20dB of feedback is applied, reducing the deviation in the i.f. amplifier to 20kc/s per channel. The resultant deviation of the oscillator is thus 180kc/s, and is obtained with a 20dB improvement in linearity.

In order to maintain the correct phase within the feedback loop over the wide operating band necessary to accommodate the 60 channel signal (300c/s to 320kc/s) a shaping amplifier is incorporated between the discriminator output and the resonator electrode of the oscillator. If now the transmitter is connected to the aerial, any reactance coupled into the oscillator cavity, which would cause non-linearity, is reduced by the feedback factor of 20dB.

Since the discriminator is tuned to 70Mc/s any departure of the oscillator from the frequency of the harmonic ±70Mc/s will produce a d.c. output proportional to the frequency error. This is applied to a servomechanism which corrects the oscillator frequency, tuning the cavity mechanically and maintaining the resonator voltage at the optimum

Fig. 4. (a) Transmitter



value. In effect the microwave oscillator is crystal controlled and has a highly linear modulation characteristic.

It is still necessary in some cases to use aerials elevated by a considerable amount in order to clear path obstructions. To retain a reasonable length of feeder, these circumstances may be met by mounting the aerial at the base of the tower, directed vertically with a plain reflector at the top at 45°.

The apparent complexity of the transmitter, besides achieving the necessary characteristics, effects a considerable simplification of the system as a whole.

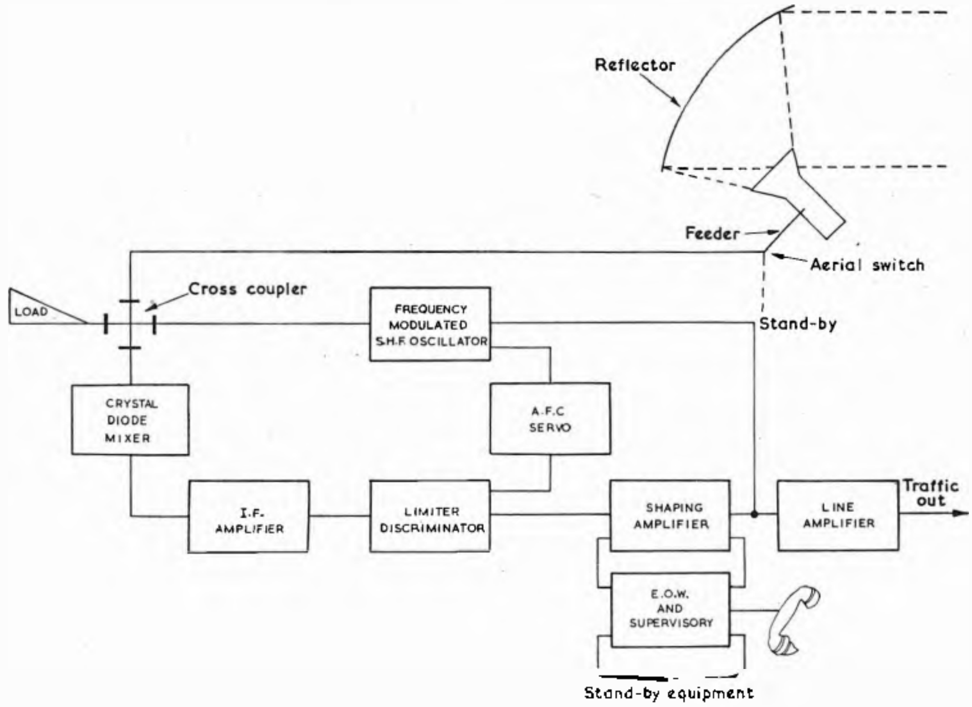


Fig. 4. (b) Receiver

Receiver Design

The receiver is similar to the transmitter except that the signal received from the aerial is fed into the mixer in place of the harmonic of the modulated oscillator, which is absent. The same feedback principle is involved, but the oscillator output is absorbed in a load. The feedback is retained to reduce the distortion in the i.f. amplifier and discriminator, the output from which is fed to line or carrier equipment through an amplifier. The oscillator frequency is locked to that of the incoming signal by a similar servomechanism.

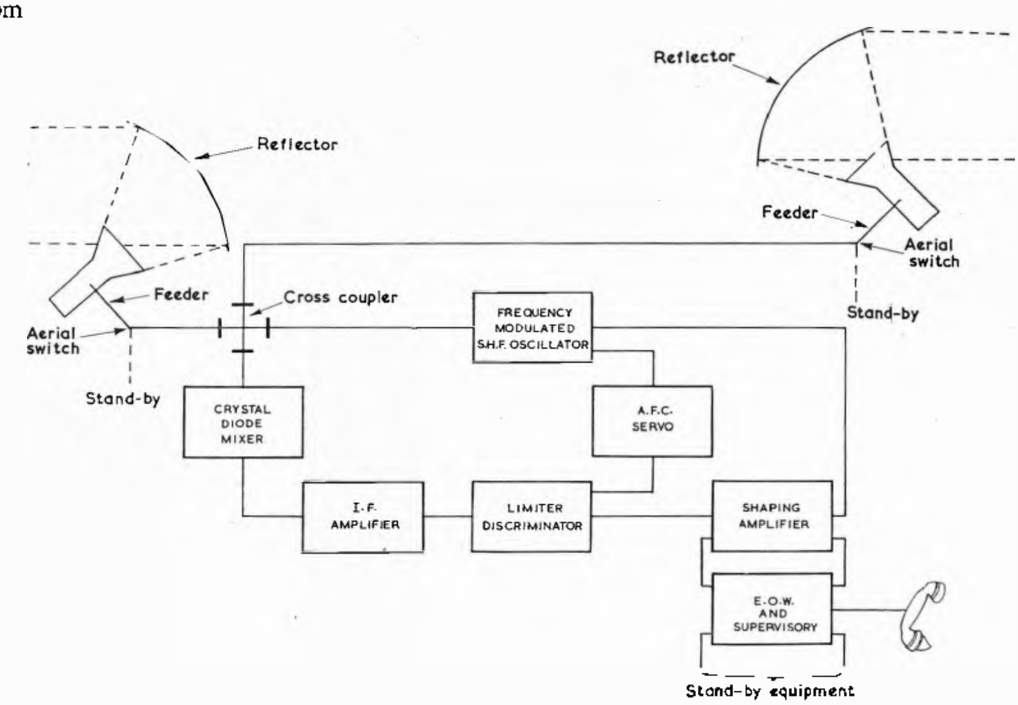


Fig. 4. (c) Repeater

Repeater Design

It will be obvious that if the receiver oscillator is fed into an aerial instead of the load, it will radiate a signal in a similar manner to the transmitter. It thus becomes both receiver and transmitter, serving as a non-demodulating repeater. The use of a simple repeater design imposes a number of system limitations, but these are not important in practice for a toll system of the kind being described. In such a repeater, the transmitter frequency will be removed by 70Mc/s from that of the incoming signal. As with the transmitter 99 per cent of the output from the oscillator is available for transmission.

The deviation is reduced by 10 per cent in the feedback process, representing a loss of 1dB. In consequence, it is normal practice to utilize back-to-back terminals, at about every fifth station, to restore the modulation index. These may be conveniently located at a traffic centre where access to the modulation is anyway necessary. Systems carrying less than the maximum can, of course, be extended further since there is an adequate performance margin, a 24 channel system for instance has at least a 6dB reduction in channel noise over the top channel in a 60 channel system.

This principle of overall feedback not only provides an elegant method of reducing distortion but greatly simplifies the repeater station. It also results in a design which utilizes the same basic units in the terminal transmitter, receiver and repeater, greatly simplifying maintenance (see Fig. 4). The last feature is particularly attractive since relatively few panels are required to provide spares for a complete system.

The terminal transmitter and receiver are each accommodated on a standard 7ft 6in rack. A two-way repeater, consisting essentially of two receiver racks (but without line amplifiers) is also accommodated on two racks.

Reliability

The design of an equipment must not only provide a solution to the technical problems but also ensure that each product provides the same minimum performance for a long period with as little skilled maintenance as possible.

The larger capacity systems are so important that some form of insurance against failure is required. The circuit may fail for a number of reasons, the two basic causes being the loss of signal due to abnormal propagation conditions, and equipment failure. To some extent the system may be designed and planned to allow for loss of signal due to propagation effects so that the interruption of the circuits is reduced to an acceptable percentage of the time An equipment failure, particularly at an unattended station may interrupt vital communications, cause loss of revenue and generally detract from the standard of service required. Trunk systems may warrant the provision of a complete spare active path with standby power supplies providing a 'twin-path' system ensuring continuity save in the event of simultaneous faults on both paths.

An alternative, and generally preferable, arrangement, is the 'quiescent standby' system, which provides alternative equipment at each station which is automatically brought into service in the case of failure. It has the advantage that a fault may develop at each station without permanent interruption of the circuit. The 'twin-path' system obviously requires a duplication of frequency allocations since both are permanently operative. With a 'quiescent standby' system the operational and standby equipments must work one to the other and therefore on a common frequency.

The usual method of operating quiescent standby is to remove the drive from the standby transmitter, and monitor the outputs of both receivers. Failure of the operational receiver switches on the standby transmitter. (An ambiguity exists unless both receivers are monitored since failure of the received signal may be due to a fault in the receiver, a fault in the preceding transmitter, or a deep fade. These difficulties may be overcome by a system of delayed switching to avoid unnecessary changeover, but results in significant delay in restoring the circuit where a number of repeaters are involved).

Since with this equipment a common oscillator is used at repeater stations to serve as local oscillator and transmitter, this arrangement is not suitable. However, a particularly simple solution has been found.

The signal from the receiving aerial is fed directly through a waveguide or coaxial switch to the operational repeater and through a similar switch to the transmitting aerial. A little of the signal is also fed into the standby repeater which is maintained fully active, except that the voltage on the v.m. oscillator is modified to prevent oscillation. As soon as the signal received by the operational repeater falls below a specified level, the standby oscillator is brought into operation. The signal received by the standby repeater is normally lower since it is coupled to the aerial through an attenuator, so that a fade, or failure of the previous transmitter does not cause a changeover. If, however, the standby repeater receives a signal of higher level than that indicated by the operational equipment, the aerial switches are changed over and the faulty equipment switched to the quiescent condition. In the event of power failure the two equipments are switched on when power is restored but the operational repeater takes precedence. This arrangement provides the maximum reliability against failure with a very short changeover time and a low incidence of unnecessary switching.

Provision of a standby equipment is not sufficient in itself. Unless a changeover at an unattended repeater is notified to the control terminal the faulty equipment may be left unrepaired. Consequently, fault indications are transmitted as tones over the engineer's order wire which provides party-line communication between all stations. The order wire occupies a band from 300c/s to 3kc/s and the supervisory tones are located immediately above it.

Two other facilities which have been found desirable as a result of long experience are also available. They are the 'revertive check' and 'turn round' operation. The former switches all stations to 'operational' to check that this path is free from fault or that a changeover has occurred in error. The turn-round facility enables the terminal and each repeater to be checked in turn. An oscillator is used which has a frequency equal to the difference between the transmitted and received signal frequencies, and a little of the transmitted signal is coupled from the aerial feeder into a crystal mixer, together with the output from the turn-round oscillator. The combined frequency is accepted by the receiver, to which it is loosely coupled, thus enabling the terminal to be checked as an entity. The normal circuit tests include signal-to-noise ratio, intermodulation and test-tone levels. At the terminal the turn-round oscillator is manually switched and is normally off. In order to perform the loop test on the first repeater, a characteristic control tone is transmitted which brings the t.r. oscillator into action and likewise each repeater may be turned round from either terminal. Failure of the signal to return identifies the faulty station, while excessive noise or crosstalk may also be located. Rapid identification of a faulty station is very important particularly when both equipments are affected, since the outage time may be governed by the time taken in travelling to the station. Its location enables the maintenance engineer most favourably situated to deal with the trouble. (It is worth noting that a single channel may have a revenue value of several pounds per hour and the loss of sixty channels in a peak period is of some consequence.)

Among other design aspects one must include safety precautions to protect both the equipment and personnel; component identification and layout to facilitate maintenance; quality of components, materials and finishes to prevent deterioration in adverse climatic conditions. The use of s.h.f. techniques introduces a further problem of mechanical accuracy. Many component designs which are otherwise satisfactory require an accuracy of manufacture which is difficult to achieve. The accurate reproduction of complex shapes in high conductivity metals gives scope for the development of new processes. Shell-moulding, precision casting, electro-forming and high-speed precision routing are some of the commonly used methods to avoid expensive and skilled machining operations.

System Performance

The quality of manufacture in many cases can only be confirmed by electrical test. Measurements of a high order of accuracy which initially present a problem in the laboratory must be repeated accurately and quickly in production testing. The design of test equipment and the carrying out of fundamental measurements may involve as much effort as the design of the equipment itself.

Tropospheric Scatter Communication

By G. L. Grisdale*, Ph.D.

By using high-power transmitters, large aerials and sensitive receivers, it is possible to achieve satisfactory multi-channel communication far beyond line-of-sight range. The received signal is characterized by very rapid fading, the effects of which are reduced by using diversity systems. The article reviews the main features of tropospheric scatter communication.

RADIO communication at frequencies above 100Mc/s has usually been confined to transmission paths in which the transmitter and receiver are within sight of each other. Under these conditions quite low powers can provide high-quality transmissions with large modulation-frequency bands for multi-channel telephony, television and radar links. Signal strength fluctuation is not normally a problem if the aerials are carefully sited. The signal in line-of-sight links is normally reliable and steady. When the distance between transmitter and receiver is increased beyond the line-of-sight limit, the signal rapidly becomes weaker, although at the lower frequencies in the v.h.f. and u.h.f. bands the diffraction of waves round the curvature of the earth or over terrain obstructions can provide useful communication with increased transmitter powers. The lower curve of Fig. 1 shows the calculated increase in attenuation with distance for a 3 000Mc/s signal diffracted over a smooth earth.

It has been found that the signals are received at greater strength than would be given by diffraction theory and the points on Fig. 1 show the results of experimental observations made in the United States by Bullington[1]. The measured signal strengths have been plotted in comparison with the signal strength which would be obtained with the same transmitted power and aerial systems in free-space. The signal strength is sufficient for useful communication over paths up to two or three hundred miles, provided that powerful transmitters are used with large aerial systems. It may be necessary to limit the bandwidth transmitted in accordance with the economics of the system.

Propagation Mechanism

The mechanism by which these useful signals are transmitted is called tropospheric scatter. The physical properties of the air do not vary in a smooth and continuous way; there are variations of dielectric constant with height in the form of blobs or layers having values different from the average around that height. When a radio wave, having a sufficiently short wavelength compared with the size of the discontinuity, is sent through the atmosphere, most of the wave energy continues in its original direction, but some is scattered in different directions, in the same way that the blue part of the sun's light is scattered by the atmosphere to make the sky appear blue instead of inky black.

The angle between the direction of the original ray and that of the scattered ray is called the scatter angle; the smaller the scatter angle the more intense is the scattered radiation.

Fig. 2 is an exaggerated picture of a section through the earth, showing a transmitter T and a receiver R, both fitted with high gain aerials pointing at each other, giving narrow beams, the lower edges of which are cut off by high ground. The beams are directed as low as the terrain allows, so that the scatter angle is kept as small as possible, not usually greater than two or three degrees. Where the transmitter and receiver aerial beams intersect, there is a volume of air which acts as the 'scatter volume'; different parts of the scatter volume contribute to the scattered wave, so that the received energy consists of many components which have travelled slightly different distances. This means that

The illustration above shows a 30ft aerial for tropospheric scatter transmission

* *Marconi's Wireless Telegraph Co. Ltd.*

the signal is not of constant strength, as in line-of-sight links, but fades in the same way as long-distance h.f. signals transmitted through different paths in the ionosphere. The fading of tropospheric scatter signals is, however, usually more rapid than that of h.f. signals.

Even with high-power transmitters and large aerials the signal is nearly always weaker than that which is considered necessary for commercial line-of-sight links, and there are two kinds of fading to be taken into account. The 'long-term' fading is the variation of signal strength due to changes in the general condition of the propagation path at different times of the year and day. Provision must be made for the lowest signal strength that is likely to be received, and a usual criterion is that the signal should not be weaker than the lowest acceptable level for more than 1 per cent of the total hours. The long-term fading allowance at 1 000Mc/s for a 200 mile path is about 10dB, compared with the median long-term signal.

In addition to the slow change of signal strength, there is also rapid fading due to the arrival at the receiving aerial of waves from all parts of the scatter volume; these component waves have a range of amplitudes, and their phase relationship is random. They combine to give an output from the receiving aerial which fluctuates; sometimes short, deep fades occur less than a second apart[2]. The characteristic time of this fading may be as short as 0·1sec, and is normally much more rapid than the fading of long-distance h.f. signals. Short-term fading does not necessarily occur simultaneously at aerials spaced a short distance apart, so that spaced-aerial diversity may be used to reduce the time during which the output fades. The spacing of aerials to achieve independent fading is usually taken as one hundred wavelengths, which is not great at frequencies in the u.h.f. band. The distance between aerials and the transmitter and receiver building should be as short as possible to reduce feeder-losses and cost, so that the spacing of aerials for diversity reception should be the minimum that gives independent fading in the two receiving aerials.

Another characteristic of the short-term fading is that if signals are received using the same transmitting and receiving aerials at two different frequencies, the two signals do not always fade simultaneously. This provides the possibility of frequency diversity, if economics and the demands for frequency allocations permit; it also limits the maximum modulation frequency that can be used, for if the whole transmitted frequency band does not fade simultaneously there will be serious distortion of the modulation; this distortion is only too well known with telephony in the h.f. band.

The bandwidth that can be transmitted over a link is determined by the difference in distance between the longest and shortest tracks which transmit energy through the scatter volume; the greater the effective scatter volume, the greater the path differences between different components of the wave and the narrower the band of frequencies which can be transmitted without distortion. If bandwidth is a vital consideration, the effective scatter volume can be reduced by narrowing the aerial beam so that only the most effective part of the scatter volume is illuminated; parabolic dishes of 60ft diameter are already in use, and 120ft dishes have been proposed. The selective fading effect is not at present a serious limitation for multi-channel telephony, but it might cause difficulty if a television relay system were to be considered.

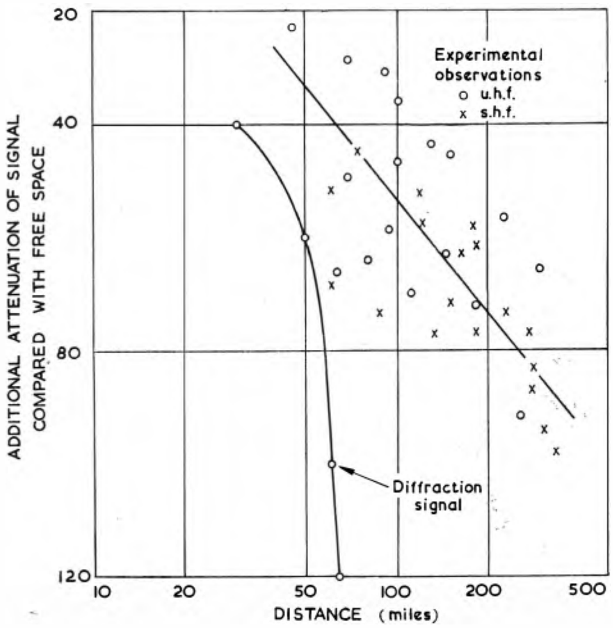


Fig. 1. Calculated and observed curves of attenuation with distance

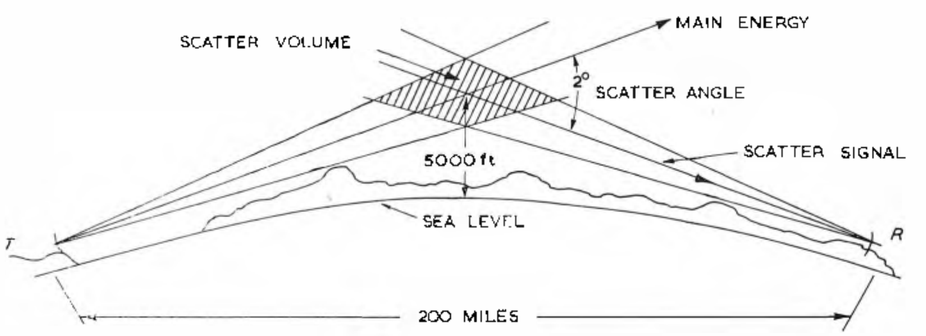


Fig. 2. Geometry of a 200 mile tropospheric scatter circuit

An interesting feature of tropospheric scatter transmission is that the polarization of the signal is preserved, that is, a signal transmitted on a horizontal aerial will not have a vertical component even after passing through the scatter mechanism; this means that polarization diversity reception, which can be used with similar signals in the h.f. band, is of no value with tropospheric scatter communication. The property is used in some diversity systems to attain quadruple diversity with only two aerials at each terminal[3].

The frequency band in which tropospheric scatter is effective extends from 400Mc/s upwards, certainly beyond 5 000Mc/s; at lower frequencies the effect is often masked by other propagation mechanisms. In this range, assuming that the transmitter power and the size of the aerials are kept constant, there is not much variation of performance. The choice of frequency therefore hinges upon the allocation of frequencies by the Administration concerned and the availability of suitable components, notably large transmitting valves. Large multi-cavity klystrons were produced for Band V television, and these have found a welcome outlet for tropospheric scatter transmission. Such a klystron can provide 10kW output at an efficiency of about 40 per cent, and can be adjusted to give a bandwidth wide enough for multi-channel telephony, and even a television link. Fig. 3 shows a 10kW transmitter for tropospheric scatter communication, with the power klystron withdrawn. The fact that the external cavities are designed round the valve leads to a clean assembly. The klystron has high gain, only 10W driving power being required for 10kW output. The input/output characteristic is linear almost to the full output.

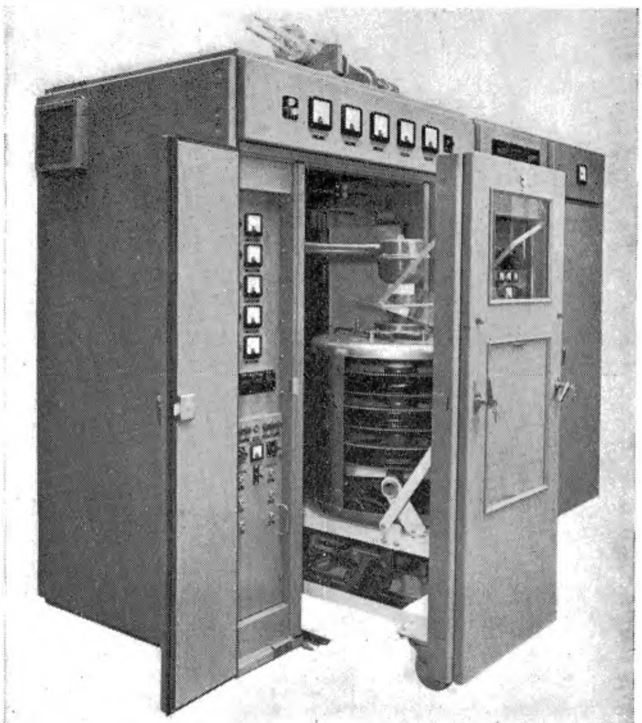


Fig. 3. 10kW tropospheric scatter transmitter

Aerials

The preferred aerials are parabolic reflectors, illuminated by a dipole with a reflector, or a horn radiator. The diameter of dish to give the required gain may be up to 60ft, and the 30ft diameter has become a widely accepted size. Such aerials must be built to stand over an inch of ice coating and gales up to 100 miles/h, and represent a considerable fraction of the total cost of the installation. The surface must have the correct profile to within about 1/16 wavelength, which at 4 000Mc/s gives a tolerance of $\pm\frac{1}{4}$in. A dual diversity system needs two receiving aerials at each end, and these can also serve as transmitting aerials. It is becoming usual to employ two transmitters and four receivers at each terminal to give quadruple diversity with only two aerials; a number of different arrangements is possible, and reliability is improved by using several equipments.

The photograph on page 272 shows such an aerial used for a prototype link; steel has been used throughout the construction, the surface of the dish being sheet metal. The twelve sectors of the dish are assembled on site, the 4 ton dish being then raised on to its gantry. Small angular adjustments are provided in azimuth and elevation. The aerial is fed by a horn with two probes to excite mutually perpendicular polarizations, the probes being adjusted to give the lowest coupling between them, thus protecting the receiver from the high transmitter power. The transmitter feeder for 10kW up to 1 000Mc/s is $3\frac{1}{8}$in coaxial, and the problem of adjusting the feeder over a band of frequencies is formidable. Waveguide could be used if available in the large size needed, but in any case the cost of transmitter feeder is considerable. The receiver feeder in this case has $1\frac{5}{8}$in outside diameter.

Other kinds of aerials are possible at the low-frequency end of the band, but for the most part a high-gain array, such as a group of Yagis, would need so many elements that feeding them would be difficult. Some form of large sheet reflector, such as the parabola, is almost certain to be adopted, but there are possibilities of shaping the side of the building which houses the transmitter to provide the aerial 'dish', or of making the gantry into a building for equipment, which comes to much the same thing.

The effective height of the aerial depends upon the contour of the ground in front of the aerial, and the profile of the terrain along the route must be examined carefully to ascertain that no obstruction causes a serious increase of the scatter angle, and thus adds attenuation to the transmission. The curvature of the earth helps to keep elevated obstructions below the horizon as distance is increased; at 20 miles the curvature drops the effective height by 200ft, which may be significant in fairly flat countries. Raising the aerial to very high ground reduces the scatter angle and gives better transmission, and with a carefully chosen site the aerial can sit with its lower edge on the ground. Many investigations have been made into the calculation of signal strengths over given paths[4] and it is now possible to predict the results over a given terrain with some certainty[5].

Application of Tropospheric Scatter

The most important application of tropospheric scatter transmission is to replace multi-channel v.h.f., u.h.f. and microwave links in cases where intermediate repeater stations cannot be used. The defence network in Northern Canada uses tropospheric scatter communication from very inaccessible sites and where low temperatures exist for a large part of the year. Other possible applications are for links between islands, where it is impossible to erect repeater stations in the sea. The Texas Tower radar installations off the coast of the United States are linked to the mainland by such communication. It has been suggested that Europe and America could be linked by a series of island-to-island hops using tropospheric scatter communication, and that such a link could carry television pictures; calculation shows that such a system would not provide very high quality pictures over this route.

Multi-channel telephony is normally handled in the same way as for low-power microwave links, that is, the speech channels are arranged by frequency changing processes into groups of twelve channels spaced 4kc/s apart, each group occupying a spectrum 48kc/s wide. Thus, if it is desired to transmit 36 telephone channels, they would be converted to single-sideband signals between 12kc/s and 156kc/s, the lower three groups of twelve channels each. This base-band is then regarded as the modulation band for the transmitter, which up to the present has used frequency modulation.

The deviation to be used for the f.m. signal takes account of the fact that the signal fades for instants to a very low level, which does not normally occur for line-of-sight links. During these fades the signal is less than the noise in the receiver, i.e. below the threshold level at which the limiter of the receiver can be effective in reducing noise. Under these conditions short intense bursts of noise are produced in the receiver when the deep fades occur, lasting perhaps 0·1sec. The frequency-modulated tropospheric scatter links have this impulsive type of noise, rather than the smoother noise associated with amplitude-modulated or single-sideband signals.

For an f.m. receiver of given gain the noise power level at the limiter will be proportional to the bandwidth, whereas the average signal level will be independent of bandwidth. For a wideband receiver, to accommodate a large deviation, the signal falls below threshold level more often than in a narrow band receiver and the noise bursts in the output are longer and more powerful. From this point of view the deviation should be as small as possible; however, when the signal is above threshold (most of the time) the signal-to-noise ratio is proportional to deviation. A compromise must be made between the conflicting demands. In general the system parameters are chosen so that the signal does not fall below noise for more than one per cent of the time; this limits the receiver bandwidth and thus the deviation. For a typical 200 mile link, carrying 24 channels, the deviation was 60kc/s for test-tone level in one channel, giving r.m.s. deviation for 24 channels of about 150kc/s; the i.f. bandwidth was 800kc/s, somewhat narrower than it would be for line-of-sight links.

An alternative to frequency-modulation for tropospheric scatter links is single-sideband modulation. The system is essentially the same as that used for h.f. communication, except that there could be many more than the four speech channels which are the maximum for h.f. It has not yet been established whether s.s.b. gives superior results to f.m.; the character of the output noise is different with s.s.b. the noise surges being less spiky, and it is possible that s.s.b. will give better results when signal strength is low. It is likely that many f.m. links will be established before the relative advantages of the systems are resolved.

The standard of signal-to-noise ratio demanded of line-of-sight links is extremely high, and not many tropospheric scatter links over useful distances will achieve this perfection. However, a very satisfactory practical standard is possible and this is achieved by attention to every aspect of the system which can give an improvement to the signal. Receivers in the 600 to 1 000Mc/s band can have 7 to 10dB noise factor; an improvement of 3dB in sensitivity at the receiver is more significant when it is considered as equivalent to doubling the transmitter power output. Again, increasing the transmitting and receiving aerials from 10ft to 30ft diameter is equivalent to increasing the transmitter power by up to 80 times. The improvement by using diversity depends upon the quality of service of the individual channel; the better the performance of the single path, the more the improvement by using two or more paths in diversity. The improvement in signal by changing from double to quadruple diversity is less than that obtained in changing from one path to two paths, but quadruple diversity is likely to become the standard system for high performance links of high reliability.

In tropospheric scatter diversity receivers, the detected outputs of the receivers (at baseband frequency) are applied to a combining unit which adds them together, taking a greater contribution from the receiver which has, for the moment, the greatest signal-to-noise ratio. It has been shown that this form of combining gives a better signal-to-noise ratio in the combined output, than would be obtained by simple switching to the best signal. The improvement results from the fact that in tropospheric scatter links there is no phase difference between the modulation waveforms delivered at the outputs of the diversity receivers, whereas the noise outputs are random and unrelated. With dual diversity the maximum improvement from this form of signal combining is 3dB when the signals in the two channels are equal.

10kW transmitters and 30ft diameter aerials have been mentioned for ranges up to 200 miles. At half this distance, and with the same aerials and receivers, the signal strength would be about 20dB higher, and 100W transmitters could give the same service. Alternatively the size of the dishes could be reduced to about 15ft and the transmitter to 1kW. On the other hand, if ranges much greater than 200 miles are to be covered, the standard of communication would fall off with the powers and aerials now available. Narrow band telegraph circuits have been maintained at distances up to 600 miles.

Conclusion

The first applications for tropospheric scatter communication will certainly be concerned with situations in which the terrain rules out multi-hop microwave links. Tropospheric scatter can give reliable communication, but due to the fluctuating nature of the signal and the limited signal strength over the longer distances the quality of the communication cannot compare with properly engineered line-of-sight links. It is estimated that the cost of a tropospheric scatter link will not be far from the cost of a multi-hop link using v.h.f. or u.h.f., the number of channels being limited. The large cost of high power transmitters, large aerials and diversity receivers is off-set by the number of installations in the multi-hop link. The main interest in tropospheric scatter systems has, up to now, come from the Service organizations, but there can be little doubt that their value in commercial operations will soon be proved.

Acknowledgments

Thanks are due to the Engineer-in-Chief of Marconi's Wireless Telegraph Company for permission to publish this article.

REFERENCES

1. BULLINGTON, K. Radio Transmission Beyond the Horizon in the 40-4000Mc/s Band. *Proc. Inst. Radio Engrs. 41*, 132 (1953).
2. RIDER, G. C. Some Tropospheric Scatter Propagation Measurements and Tests of Aerial Siting Conditions at 858Mc/s. *Proc. Instn. Elect. Engrs.*, Paper 2528R (1958).
3. LONG, W. G., WEEKS, R. R. Quadruple-Diversity Tropospheric Scatter System. *Trans. I.R.E. Communications Systems*, p. 8 (1957).
4. NORTON, K. A., RICE, P. L., VOGLER, L. E. The Use of Angular Distance in Estimating Transmission Loss and Fading Range for Propagation Through a Turbulent Atmosphere over Irregular Terrain. *Proc. Inst. Radio Engrs. 43*, 1488 (1955).

British Television Equipment for Yugoslavia

The Yugoslavian Broadcasting Authorities have ordered from Britain a considerable quantity of equipment for the establishment of a television link between Belgrade and Ljubljana, an overall distance of 569km.

Marconi's Wireless Telegraph Company Ltd are to supply this link, together with an ATE-Marconi engineering order wire system. The approximate value of the contract is £51 000.

The television link has two features of particular technical interest. Working on u.h.f. (in the 1 700 to 2 300Mc/s range) it employs an all-travelling-wave tube technique both in the terminal and repeater stations; the wide band provided by travelling-wave tube amplifiers enables a high-definition television signal to be accommodated, together with its associated sound channel. In this instance the sound channel is conveyed on a pilot carrier inserted above the vision base-band.

The installation, which will include four Marconi Type HM200 terminal equipments and two Type HM250 repeater units, will be routed from Avala (the television transmitting station near Belgrade) via Crveni Cot, Ozren, Psunj, Sljeme (near Zagreb) to Krvavec (Ljubljana). Terminal units are to be provided at two of the intermediate points—Crveni Cot and Sljeme (the latter covering Zagreb)—in order that the vision and sound signals can be extracted at these points if required. Additionally, facilities will exist at Sljeme for inserting programmes originating at Zagreb. These can be transmitted via the link to Belgrade and Ljubljana simultaneously or to either one at will.

S.H.F. Radio Links Using Travelling-Wave Output Amplifiers

By G. Dawson*, B.Sc., M.I.E.E., and T. K. M. Korytko*, B.Sc., A.M.I.E.E.

This article discusses the problems that have arisen in the planning and installation of broad-band s.h.f. radio links in many different countries during the past six years. It describes the techniques which have been evolved in the design of the s.h.f. equipment and the associated supervisory and switching equipment in order to provide transmission circuits to the international standards recommended by the C.C.I.R.

IT is now over six years since the first wideband radio relay system using travelling-wave amplifiers was put into service for television transmission between Manchester in England and Kirk O'Shotts in Scotland, and during this period considerable experience has been gained of the planning, installation and operation of this type of system and further developments have taken place in the design.

It may be of some value to review the experience derived from these systems installed in many different countries, to consider some of the practical problems encountered and to discuss the latest type of equipment and probable future trends in the design.

The travelling-wave tube was invented during the war by Kompfner[1] at Birmingham University and developed at the Clarendon Laboratories, Oxford. It was developed by Rogers[2] into a reliable s.h.f. power amplifier for use in the Manchester-Kirk O'Shotts radio link[3] in 1952. The 4 000Mc/s band was chosen for its initial application from the possible alternatives of 2 000 or 4 000Mc/s allocated for fixed point-to-point wideband radio links, as compact and high-performance waveguide structures and aerials could be used, and the use of this higher frequency band offered the best promise for future expansion of the transmission capacity of radio link systems with the valve techniques then available. This choice has been endorsed by the general trend throughout the world for the large capacity systems to use the 4 000Mc/s band[4,5] (although in the future it is anticipated that the 6 000Mc/s band[6] may be used for even larger capacity systems) and also by the extended use of the travelling-wave tube as the output amplifier in s.h.f. radio relay systems.

The Manchester-Kirk O'Shotts radio link was designed for the two-way transmission of 405 line monochrome television signals only, but the interplay of laboratory developments and field experience has now produced an equipment design which is capable of providing up to six two-way radio channels on a single route in the frequency band 3 800 to 4 200Mc/s, each radio channel being capable of transmitting either 600 telephony channels or a 405, 525 or 625 line monochrome television signal (with the possibility of adaptation for colour) to the international transmission standard of the C.C.I.R. This extension of the technique has only been made possible by intensive laboratory development and close co-operation between the radio link designers and the Administrations who operate the systems.

Frequency modulation is used in all the radio link systems mentioned in this article and the radio repeater design is based on the principle of amplification at intermediate-frequency with reconversion to s.h.f. and the use of a travelling-wave amplifier to deliver the final output power. With the extended development of the travelling-wave valve it is now possible to design a repeater with all the amplification at s.h.f., but there are good reasons, which will be detailed later, for the retention of the present arrangement. It is interesting to note that the recently described Bell TH system intended to carry at least 1 860 channels of telephony employs amplification at 74·1Mc/s followed by a travelling-wave amplifier as a power output stage.

The successful provision and operation of a radio relay system depends upon the transmission performance, the reliability and the cost in both time and money; all of which are to some extent inter-related.

Satisfactory transmission performance, for which the targets are defined by the C.C.I.R, depends not only upon the equipment design but also upon the route planning and the successful choice of suitable sites for the radio stations, and some of the problems which arise are discussed below.

Generally speaking, propagation planning, including good Fresnel zone clearance, a good immediate foreground and a suitable angle between the directions of transmission tends to dominate the selection of sites, but the problems of acquiring the site, accessibility and building construction must also be considered.

In Britain up to fifteen different authorities may have to be consulted before final permission is granted to build a station; even then special action may be necessary as at Blackford Hill on the Manchester-Edinburgh system where the station building was concealed behind the summit of a hill in order to hide it from public view.

On the Madrid-Sevilla 240 channel telephone link similar aesthetic considerations influenced the siting of the aerials on the roof of the Telephone Building at Madrid.

As in the nature of things better clearances and longer hops are obtained by going as high as possible in a hilly terrain, accessibility may become a severe problem. Repeater sites, like Koralpe (2 100 metres) on the Graz-Klagenfurt system in Austria, may become inaccessible in the winter. The position of the sites on the Cabot Strait crossing between Nova Scotia and Newfoundland was influenced by similar considerations.

It is all the more important, therefore, if a compromise has to be accepted, to assess clearly what the choice of an inferior site entails in terms of system performance, and this is no simple matter.

It was noticed on two early systems, Manchester-Kirk O'Shotts and Johannesburg-Pretoria, that excessive s.h.f. coupling could occur between the two directions of transmission, if, as was desirable for reasons of economy of spectrum, the operating frequencies were the same.

The aerials were placed horizontally side-by-side in those days, with one transmit and one receive aerial per direction. It was found that some excessive coupling was

* *Standard Telephones and Cables Ltd.*

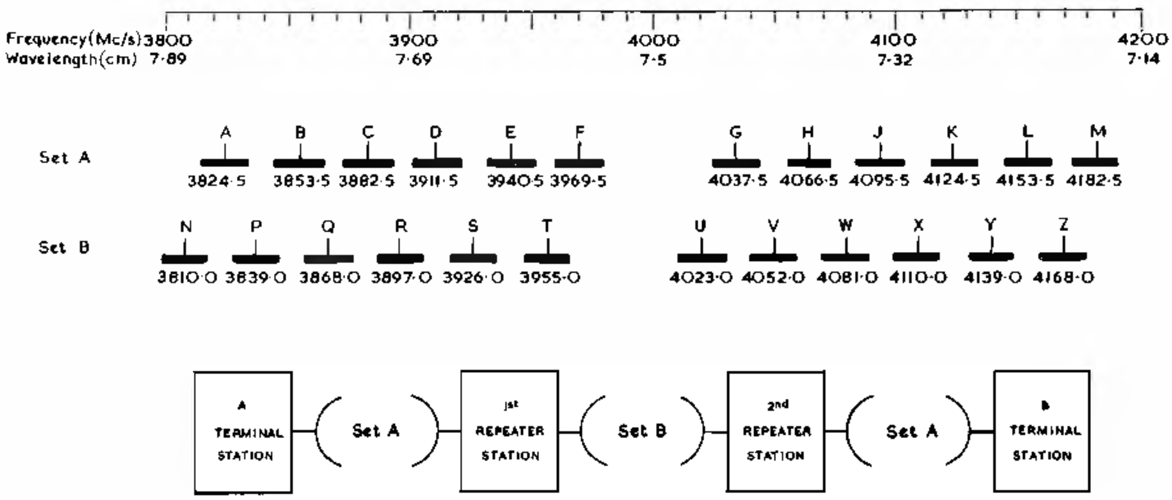


Fig. 1. C.C.I.R. frequency plan and its application

caused by the unwanted signal being reflected from the back of one aerial and entering the adjacent one. This trouble was easily overcome and on later systems the aerials were invariably placed directly above one another, with a horizontal screening plate in between.

Another source of coupling was traced to reflections from the immediate foreground, including the station building itself. Coherent reflections from sharply defined features of the terrain, like cliffs and ridges behaved similarly. A special testing equipment was developed, measuring the magnitude and distance of the reflection, and a method of calculation was finally arrived at, which made it possible to evaluate the effect theoretically once the features of the terrain were known.

It was found, however, that the choice of sites to give adequately low foreground reflections became far more difficult, if the same frequencies were to be used in the two directions of transmission, and a more radical approach was developed.

In the C.C.I.R. frequency plan for wideband systems, there are two sets of frequencies (see Fig. 1) which can be called 'A' and 'B'. The frequency spacing between corresponding radio channels in the two sets is 14·5Mc/s and is too small for both sets to be used on the same hop. It is, however, permissible to use the sets alternately when going along a route (with full conservation of the radio spectrum) so that any foreground or aerial crosstalk effects at a station would then fall outside the signal base-band which considerably eases the site selection problems. This solution has been used in the planning of 600 channel telephony routes in New Zealand (Palmerston North-Hamilton, Fig. 2), Switzerland (Berne-Geneva), and between England and France (Folkestone-Lille). At a branching point the choice is made in such a way as to minimize the total interference: in other words, the two directions with least coupling between them use the same allocation (Fig. 2).

Another kind of unwanted signal can reach a receiving aerial by 'overshoot' propagation. This propagation underlies tropospheric scatter systems; it is known that the field strength beyond the horizon may rise considerably above its mean level at times and that this level is particularly high in tropical countries. Suitable precautions have to be taken in planning the routes to change direction in such a way that there is an adequate angle between sectors *AB* and *CD*, and direction *AD*, where *A*, *B*, *C* and *D* are successive stations, *B* and *D* using the same frequencies. The directional discrimination of both the aerials is thus brought into play, and this can be re-inforced by cross-polarization if necessary. This principle influenced the planning of two routes in Canada (Fig. 3), where sites could have easily been found in a straight line, and this would have simplified access to the stations on these routes.

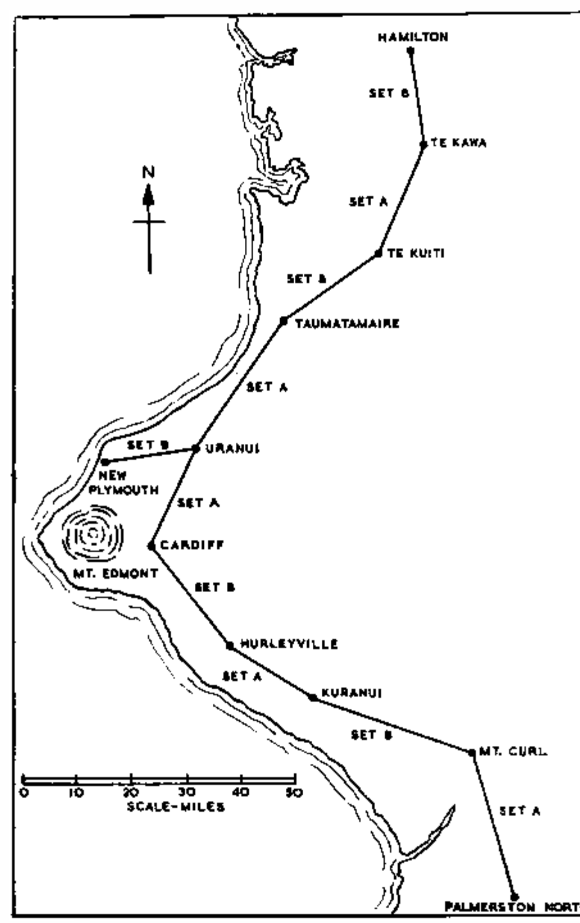


Fig. 2. Route map of a 4 000Mc/s system in New Zealand

The situation becomes more complex if the interfering signals arise outside one's own system or where one has to join two systems using different frequency plans. A difficult case of this kind occurred on the Quebec-Rimouski route in Canada, where a parallel Bell system link existed using the same frequency band (Fig. 4). Detailed agreement was necessary on sites, polarization, aerial directivity and transmitted powers before safe interworking was made possible.

Even after thorough propagation tests unexpected difficulties may arise due to seasonal variations in the terrain. When planning the route between Rio de Janeiro and Sao Paulo in Brazil it was found that the route originally envisaged, which broadly followed a river valley, became completely unsuitable after rains because of flooding and consequent strong specular reflections. Different sites had to be chosen and tested.

Over-water propagation usually leads to specular reflections and severe fading. This was first noticed on the microwave link across the English Channel established in the early 1930's between Lympne and St. Inglevert[3], which used the first microwave equipment in commercial service in the world. It was also found that fading varied with the tide, thus suggesting specular reflections and vertically-spaced diversity as an answer. Extensive tests were carried out after the war across the Firth of Forth and across Torbay and much information was obtained on the behaviour of over-water paths and the use of space-diversity to reduce the effects of fading.

As a result of this, an automatic phasing junction was developed for diversity reception (Fig. 5), which automatically combines the s.h.f. signals from the separate aerials in the correct phase to give the maximum combined signal. It operates continuously and avoids the switching breaks which occur in some systems of diversity combination.

Combination at s.h.f. obviates the necessity for duplication of receivers, so simplifying the equipment. The resulting diversity is of the 'linear addition' type, i.e. if the signals are equal in strength, a 3dB increase in signal-to-noise ratio is obtained over a single aerial; if one of the signals disappears, there is a 3dB degradation in signal-to-noise relative to the case of the other signal alone.

It is essential that a long-distance system carrying many hundreds of telephony channels or several television signals should have an exceptionally high reliability. This is reflected in the need for thorough checking of the propagation conditions along the route, and in the adequate provision of standby power and radio equipment, together with a reliable supervisory and remote control system to ensure its satisfactory operation. As more experience has been gained the methods of standby equipment provision and operation have changed.

On the Manchester-Kirk O'Shotts system each working equipment was provided with a complete standby equipment operating at the same frequency and switching took place at each station separately by means of s.h.f. change-over switches.

On later systems the approach was modified. Provision was made for a separate radio frequency protection channel as a standby for one or more working channels. The route was divided into sections comprising 4 to 6 repeaters with switching carried out at section ends only.

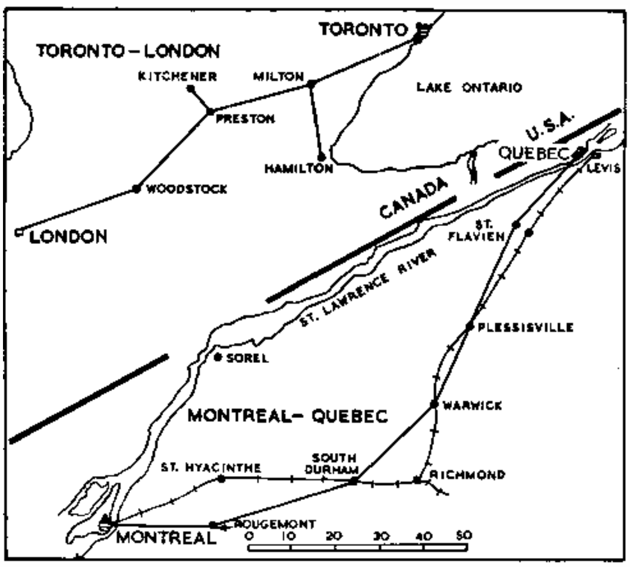


Fig. 3. Route maps of two 4 000Mc/s systems in Canada

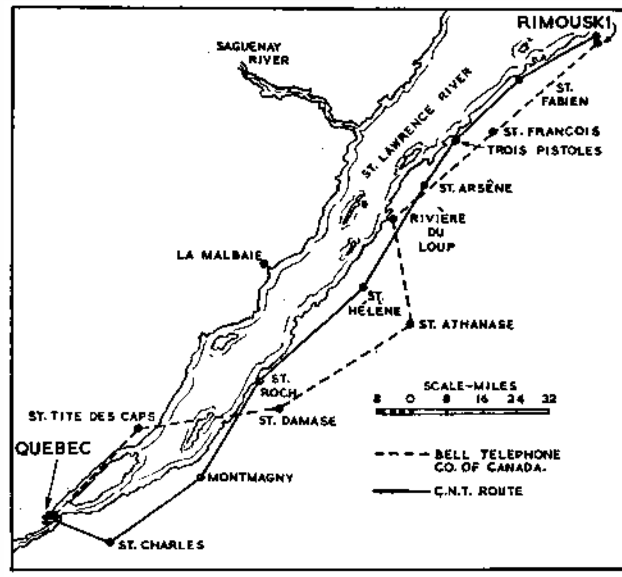
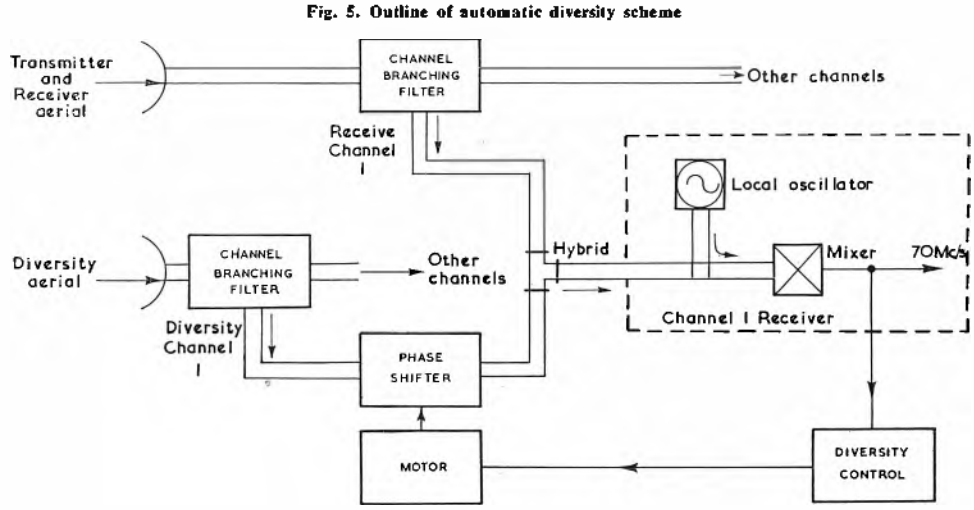


Fig. 4. Route maps of the Quebec–Rimouski 4 000Mc/s systems

Fig. 5. Outline of automatic diversity scheme



Both base-band switching at terminals and intermediate frequency switching at repeaters or terminals have been developed. The initial application of switching at i.f. on a mixed 240 channel telephony and television system with a common standby channel was on the Osaka-Fukuoka system in Japan.

An arrangement frequently used on a system transmitting only one broad-band signal is parallel transmission of the signal on both the 'working' and 'standby' channels, as experience has shown that for effective operation the standby should be switched on and working on full power all the time. Switching is thus confined to the receiving end only and a simple arrangement using base-band pilot-level and noise detectors selects the more desirable of the two signals permanently available.

If more than one broad-band channel is transmitted this arrangement is no longer possible. The spare channel must be normally kept free; switching has to be carried out at both ends, and so the instruction to switch the sending end must be obtained from the receiver. This inevitably means a greater time lag, but even then the complete sequence occupies less than 100-msec and the actual switching is only a few milliseconds. An up-to-date multi-channel switching arrangement is shown in Fig. 6.

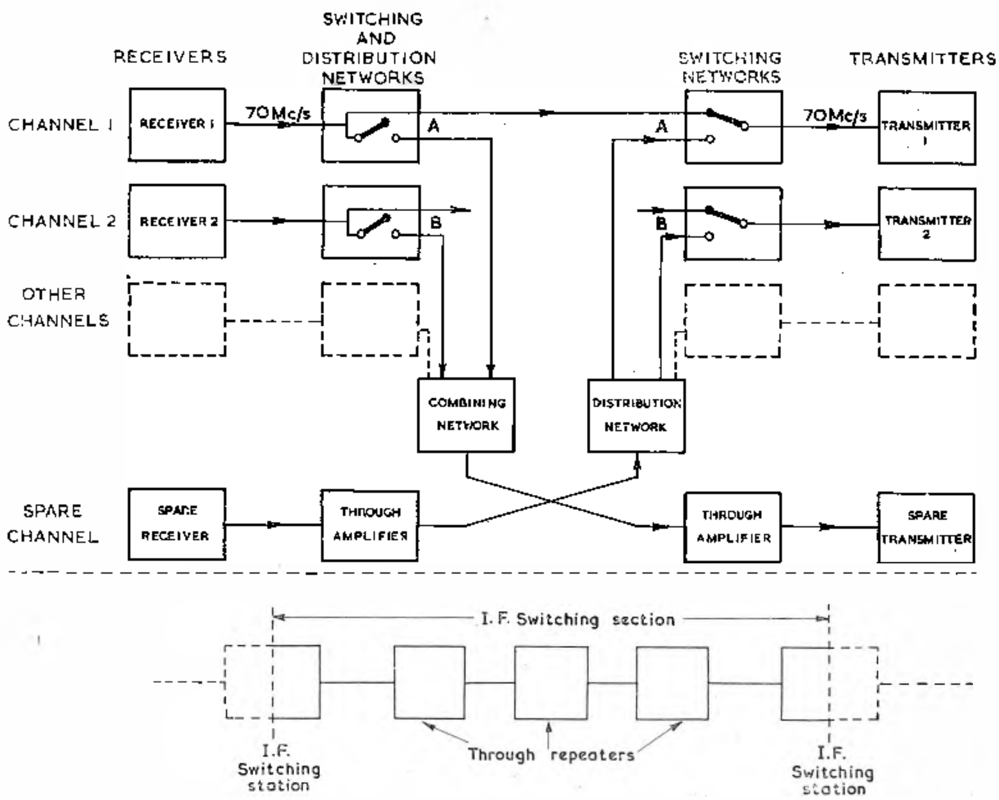


Fig. 6. I.F. switching scheme for several radio channels

While improvements were being devised in methods of standby provision and in switching arrangements, there has been continuous progress in the reduction of the size and complexity of the equipment and in extending the life of microwave valves.

On the Manchester-Kirk O'Shotts link, a uni-directional repeater occupied 3½ cubicles of the same size as the single cubicle housing the present-day repeater and the number of valves has been reduced to about 40 per cent of the original number. It is of interest therefore to examine the transmission reliability figures of the earlier equipments, as they should provide a guide to the expected performance of the later systems.

The 'outage' figures in Table 1 have been recorded within arbitrary periods of approximately one year.

All breaks are taken into account except those caused by major power failures or by engineering working parties.

TABLE 1

METHOD OF STANDBY PROVISION	PERIOD	SYSTEM	OUTAGE TIME	PERCENTAGE OF TOTAL TIME
Separate Equipment ..	Oct. 1956-Dec. 1957	Manchester-Kirk O'Shotts	80min 52sec	0·023
Parallel radio channel ..	Nov. 1956-Dec. 1957	Manchester-Emley Moor	6min 3sec	0·002
Parallel radio channel ..	Jan. 1957-Dec. 1957	Dundee-Aberdeen	20sec	>0·001

The two later systems use the parallel channel standby scheme described earlier. It can be seen that the reliability figures are very good even on Manchester-Kirk O'Shotts with its equipment of early design and in its sixth year of service.

As to travelling-wave amplifiers[10], the life of the CV2188 (W7/2D) one-watt tubes has been progressively improved. Of the eight valves initially equipped on the Thrumster-Kirkwall 240 channel telephone system all but one are still working normally after 10 000 hours. This compares with an average life of 3 000 hours on the early models of the CV2188. The latest equipment uses a five-watt travelling-wave amplifier which incorporates the experience obtained with the CV2188.

The experience on coaxial-line microwave oscillator tubes[11] has been that average working lives of about 6 000 hours were obtained from the early days of link operation onwards.

The lengthening of the travelling-wave amplifier's life substantially reduces the cost of microwave valve replacement, which is one of the major charges on link maintenance.

Great simplification has also been effected in equipment maintenance procedures. Here the major steps of progress have been the elimination of automatic frequency control at repeaters, the introduction of fixed tuned i.f. amplifiers and the advent of the ferrite isolator, which has simplified the impedance matching procedures within the equipment.

The present-day equipment incorporating these features is described below.

General Equipment Description

Aerials, Feeders and Branching Filters

The aerials normally used on 4 000Mc/s main-line routes are paraboloids of 10ft diameter with a smooth metal surface. The aerial gain at that frequency is 39·5dB and so the beam width is only $\pm 1°$ to the half-power points. This makes it necessary to construct a rigid supporting structure, usually taking the shape of a steel lattice mast of rectangular cross-section and a fairly broad base. As the radio stations are frequently situated on high ground and in exposed positions, the structures must be designed to withstand gusts of up to 120mile/h with an aerial beam deflexion well under one degree. Conditions like this are met,

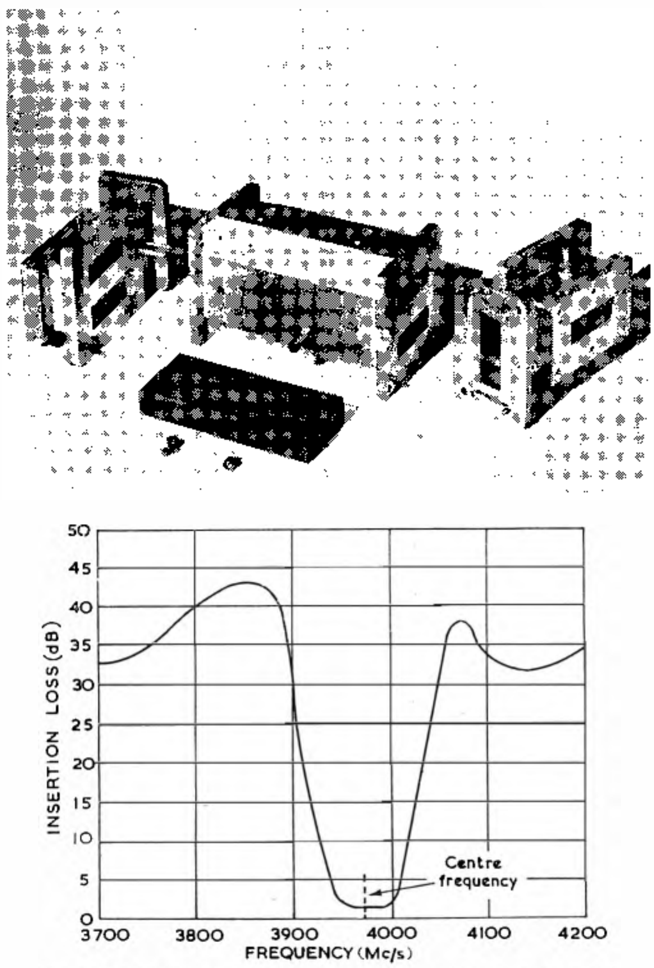


Fig. 7. Exploded view and characteristic of a channel branching filter

for example, on the Thrumster-Kirkwall route and on the Sydney-St. Johns system in Newfoundland.

The aerial reflector is fed from a waveguide horn supported in front of it by a bar parallel to the aerial face, half of which forms the waveguide feed. A rear feed may have certain advantages and its introduction is being considered.

A rectangular waveguide feed of 2in by 2/3in cross-section connects the aerials with the remainder of the equipment situated inside a building at ground level.

The branching filters separating or combining the different radio-frequency channels for common-feeder working are situated inside the building and common transmit-receive working is employed with the three channels of the C.C.I.R. allocation (see Fig. 1) using a common polarization connected to one feeder and aerial.

The other three cross-polarized channels use another aerial with a separate set of branching filters. This arrangement has the advantage of not requiring polarization separators and of easing certain channel crosstalk problems. It also means that in the great majority of cases, where system installation starts with two or three r.f. channels, only one aerial is required per station per direction at first, thus spreading the capital cost of a fully-equipped system. The feeder is simple and no spurious mode problems arise.

The branching filter (Fig. 7) consists of two band reflection filters, spaced a quarter-wavelength apart. Compact waveguide hybrids at each end perform the splitting, dropping and combining functions. The through attenuation of the filter to the dropped channel frequency is 35dB, but at any other channel frequency it is only 0·1dB. The band width of the dropped channel is about 55Mc/s.

The match of the filter to the main run is such that even with several filters in tandem the reflection at the filter end of the aerial feeder does not exceed 4 per cent.

The 'drop' arm of each branching filter is connected to its bay of equipment through a short length of coaxial cord with spot-matched transitions. This imparts mechanical flexibility to the station layout while keeping the cord loss down to about 0·25dB.

5W Repeater Equipment

The arrangement of the repeater is shown in Fig 8. The s.h.f. input signal from the branching filter passes through a four-section input filter and an isolator into the input mixer, where it is converted to the C.C.I.R. recommended i.f. of 70Mc/s. Three i.f. amplifiers follow: the pre-amplifier is designed for the lowest possible noise figure with a conversion gain of some 20dB. With available input converter diodes, the overall noise figure is 11 to 12dB. The main i.f. amplifier, with a maximum gain of some 60dB follows, delivering a signal of 0·5V which is kept constant by a.g.c. action. Finally, the i.f. output amplifier drives the medium level balanced convertor employing two germanium diodes. The medium level convertor is fed with about 0·8W of power from the coaxial-line local s.h.f. oscillator tube via a ferrite isolator, which eliminates the distortion effects of local oscillator pulling. The output from the medium-level convertor, at approximately 20mW, passes through another isolator, then through a four-section waveguide filter selecting the appropriate sideband and into the travelling-wave amplifier. At the output of this tube the signal level is 5 to 7W and, after passing through another isolator, it passes to the repeater output.

The local oscillator tube feeds both the convertors: the transmitting one directly, through the isolator, the receiving one through a shift mixer and two band-pass filters, each tuned to one of the local oscillator frequencies. The shift mixer is driven by a crystal-controlled oscillator-multiplier at the difference frequency of the input and output radio frequencies. This ensures that both local oscillator frequencies are on the same side of their respective signal frequencies, making the repeater frequency shift independent of any drift in the local oscillator. The accuracy of the frequency shift is of the order of 10 to 20kc/s and so the accuracy of the radio frequencies themselves is primarily determined by the terminal transmitter.

At successive repeaters the local oscillators are placed alternately above and below the radio frequencies, thus reversing the sense of the i.f. signal deviation at each repeater. This has the effect of cancelling any systematic component of even order harmonic distortion arising in the i.f. chain.

In earlier systems independent local oscillators were used for the two mixers, each of them controlled by an a.f.c. chain[3]. This could give a cumulative frequency error along a line of repeater stations, apart from added cost and complexity.

The latest local oscillator does not employ an automatic frequency control, but relies on a specially designed coaxial-line tube[11] which has a low pulling figure and low frequency sensitivity on its electrodes. If ageing effects are ignored, the stability of any oscillator depends on temperature, electrode potentials and load impedance. The tube is placed in a simple relay-controlled thermostatic oven kept above the maximum possible ambient, at about 60°C. An air stream is used to couple the cavity to an on-off heater. Temperature control better than 1°C peak-to-peak is thus achieved. The oscillator is fed from a stabilized power supply and feeds into an output isolator.

As a result the frequency stability is at least as good as with the former a.f.c. system: a typical daily frequency excursion is less than 100kc/s peak-to-peak and the drift of the mean frequency is about 100kc/s per month.

The shift mixer is of interest, as it is one of the few elements common to all modern repeaters. As the shift frequency in the C.C.I.R. scheme is relatively high, just over 200Mc/s, a relatively low harmonic (approximately

very few distortion effects that can arise in this tube is a.m.-to-p.m. conversion[13] and the limiting action helps considerably to minimize its incidence. Cumulative limiting at each repeater is also beneficial from the systems point of view, and one of the advantages of conversion-type limiting is that it operates over a wide band and does not introduce the troublesome stray distortion effects associated with some other types of broad-band limiter. The medium-level convertor uses germanium diodes specially designed to give good conversion properties up to 100Mc/s, but to produce negligible harmonics of the 70Mc/s signal which might fall near the output frequency.

The travelling-wave amplifier[10] gives 27 to 29dB of amplification over a bandwidth of some 600Mc/s. Group delay distortion and drift effects are negligible: the band-pass characteristic is entirely determined by the passive filter preceding the tube. There is no tuning as such—merely impedance matching, which in systems employing isolators means adjustment for maximum signal level when setting up.

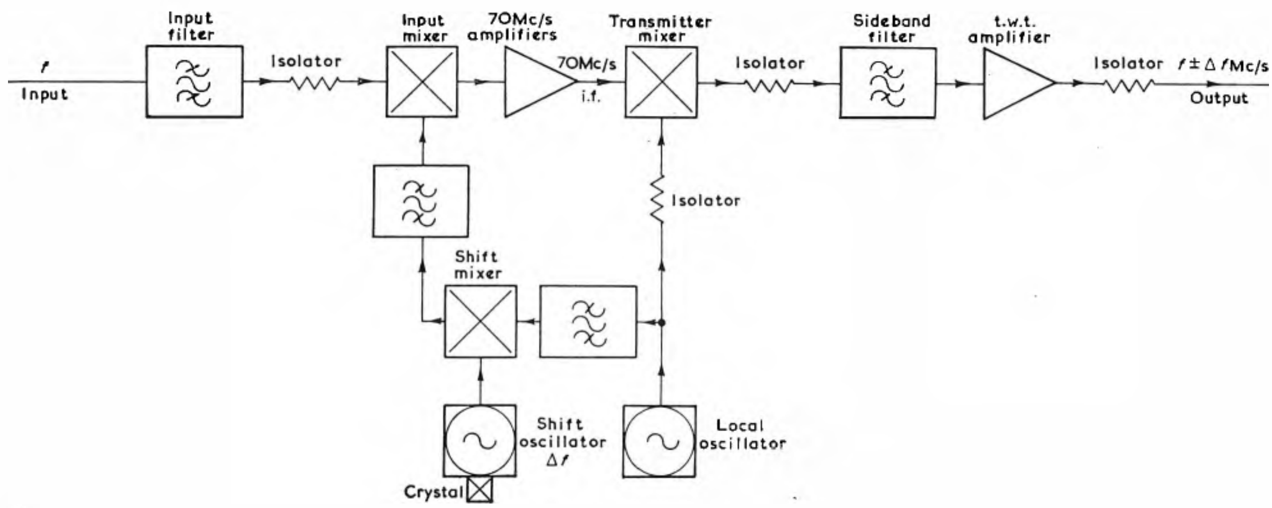


Fig. 8. Block schematic of a 5W s.h.f. repeater

the 20th) might fall close to the input frequency or the receive l.o. frequency. In both cases the harmonic level may be intolerably high in certain r.f. channels if the silicon diode in the shift mixer is driven in the usual way. A circuit has been developed for driving the diode in such a manner that the 4 000Mc/s harmonics are reduced by about 45dB relative to normal drive at a cost of an additional 2dB in s.h.f.-to-s.h.f. conversion loss. This removes all possible restrictions on the choice of shift frequencies and working channels which might be otherwise imposed by interference from harmonics.

All intermediate frequency amplifiers use the same type of inter-stage circuit: a parallel-tuned band-pass transformer with coupling factors between 0·5 and 0·7. Single or double damping is used according to individual circuit requirements. The coils used are true physical transformers and not three-inductor analogues.

The coils are preset and are reproduced against a precision standard initially developed to tune with the mean valve capacitances.

Coils of this type considerably simplify system maintenance, because in the absence of tuning only gain measurements are necessary when changing a tube[6].

The medium level mixer is an important part of the repeater, as it has the effect of limiting the signal just before the output travelling-wave amplifier[3]. One of the

Apart from the above advantages, the travelling-wave amplifier is robust and long-lived. Once adjusted it needs virtually no maintenance and a clear indication of the degree of tube ageing in operation is provided by the electrode voltages, so replacement is a matter of routine.

All waveguide filters in the repeater use quarter-wave spacing between cavities[14] and 'planar' side-by-side grouping of waveguide components into blocks. This is made possible by the development of a planar analogue of a three-dimensional hybrid with considerable band-width. These developments have substantially reduced the size of the repeater.

Operationally, the main advantages of this type of repeater could be stated as follows:

(a) Flexibility of system operation, e.g. if dropping or injection of a television signal is required at any future date, the i.f. point is available.

(b) Compactness, largely due to the small volume occupied by i.f. amplifiers.

(c) Stability of characteristics, caused by absence of active filters with a small fractional bandwidth.

(d) Low tube replacement costs, with only two long-lived microwave tubes in the circuit.

(e) Similarity of the radio equipments at repeaters and terminals (see below).

(f) Cumulative limiting in the medium-level convertors.

(g) Simplicity of maintenance.

5W Terminal Equipment

Transmitter terminals using a klystron modulated oscillator working at the final output frequency, followed by a t.w.t. are being extensively used[3] on systems where the additional terminal flexibility gained by having an i.f. point available is not worth the further complication in the equipment (Fig. 9b). As this type has been fully described elsewhere, the more complicated terminal is described below.

At a terminal with i.f. points available at the transmitter and receiver (Fig. 9a) the equipment can be diveded functionally into two parts: the radio terminal proper and the modulator-demodulator. 70Mc/s points are provided at both the transmitter and receiver—they form the demarcation line between the two parts. Functionally, the radio terminal is little different from the repeater, converting a 70Mc/s signal to 4 000Mc/s at the transmitting end and performing the reverse process at the receiver.

What is essentially peculiar to the terminal is the frequency modulation and demodulation process.

The modulator and demodulator may be divorced from the s.h.f. band in which the equipment is working and their characteristics are dictated solely by requirements on distortion of the information transmitted.

The application of pre-emphasis which reduces the asymmetry and effective deviation of the television picture has made it possible, to produce a terminal equipment common to both television and telephony. The transmitter a.f.c. operates on the mean carrier frequency.

A klystron oscillator is used to produce the modulation and it is mixed with a stable local oscillator to produce the i.f. signal.

The mixer provides an appreciable amount of limiting, as the klystron supplies the high-level signal. A.F.C. is applied mechanically within the klystron cavity.

The base-band amplifier immediately preceding the klystron is flat over the television video bandwidth while at the same time giving negligible non-linear distortion when handling 600 telephone channels.

The i.f. output is 0·5V at 70Mc/s: the frequency is controlled within close limits, as the reference for the a.f.c. is derived from a 70Mc/s crystal-controlled oscillator.

The common demodulator is conventional, consisting of a wideband limiter and discriminator and a wideband output base-band amplifier.

Switches are provided to change the different attenuator, pre-emphasis and output networks associated with either television or telephony.

As the pilot frequency is 3·2Mc/s in the case of telephony and 7Mc/s or 8·5Mc/s in the case of television, automatic network switching can easily be provided on the basis of pilot detector outputs.[7]

Looking ahead, the techniques which have been so well established in the 4 000Mc/s band are already being extended to the 6 000Mc/s band and it is probable that a single wideband s.h.f. channel will be expanded in transmission capacity to about 1 800 telephony channels or a combination of telephony and television signals. The rate of development in this field is noteworthy when it is remembered that it is only fifteen years since the travelling-wave amplifier was invented, but it should be emphasized that the development rests on a solid foundation of field experience in many countries and differing conditions of service.

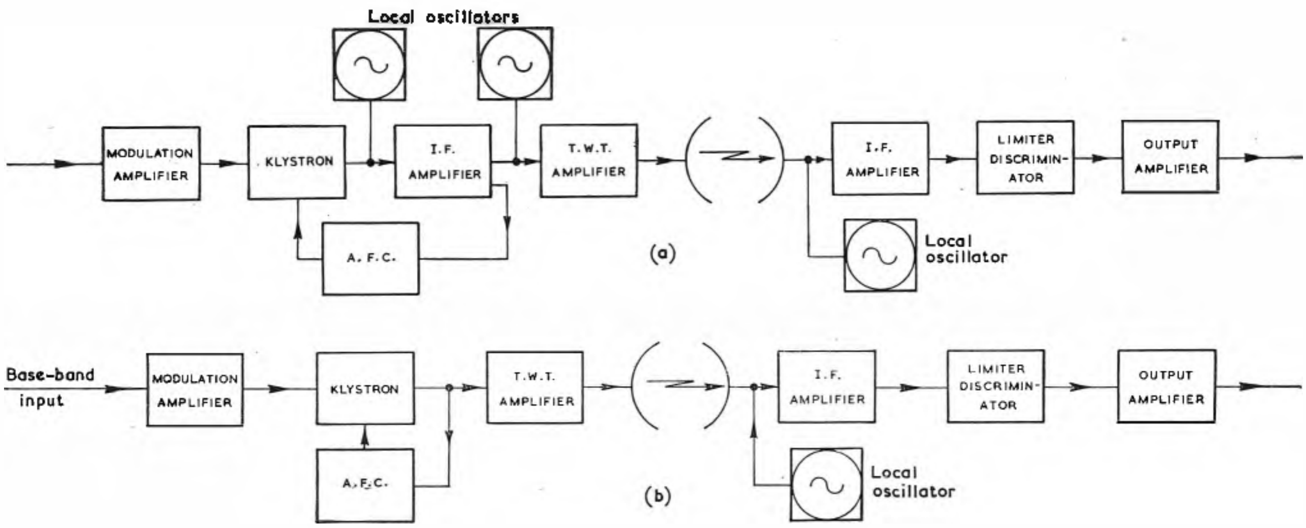


Fig. 9. Block schematics of s.h.f. terminals
(a) I.F. type (b) Direct modulation type

Acknowledgments

The authors wish to thank Standard Telephones and Cables Limited and the Engineer-in-Chief of the General Post Office for permission to make use of the information obtained in the article. Figs. 1, 2, 3 and 4 are reproduced by courtesy of ' Alta Frequenza '.

REFERENCES

1. Kompfner, R. The Travelling-Wave Valve. *Wireless World. 52*, 369 (1946) and *Wireless Engr. 24*, 255 (1947).
2. Rogers, D. C. The Travelling-Wave Tube as Output Amplifier in Microwave Radio Links. *Proc. Instn. Elect. Engrs. Pt. 3, 100*, 151 (1953).
3. Dawson, G., Hall, L. L., Hodgson, K. G., Meers, R. A., Merriman, J. H. H. The Manchester-Kirk-O' Shotts Television Radio-Relay System. *Proc. Instn. Elect. Engrs. 101*, Pt. 1, *43* (1954).
4. Roetken, A. A., Smith, K. D., Friis, R. W. The TD-2 Microwave Radio Relay System. *Bell Syst. Tech. J. 30*, 1041 (1951).
5. Libois, L. J., Thue, M. Les Nouveaux Systemes de Faisceaux Hertziens de l'Administration Francaise des P.T.T. *Onde Electrique.* (Nov. 1957).
6. McDavitt, M. B. 6000Mc/s Radio Relay System for Broad-Band Long-Haul Service in the Bell System. *Alta Frequenza.* (Oct. 1957).
7. C.C.I.R./C.C.I.T.T. Documents.
8. Documents of the 8th Plenary Assembly, C.C.I.R. Warsaw, 1956. Vol. I. (I.T.U. Geneva 1957).
9. Clavier, A. G., Gallant, L. C. The Anglo-French Micro-Ray Link between Lympyne and St. Inglevert. *Elect. Commun. 12*, 222 (1934).
10. Burke, P. F. C. Travelling-Wave Tubes for 4000Mc/s. *Electronic Engng. 30*, 310 (1958).
11. Lambert, D. C. Coaxial-Line V. M. Oscillator Valves. *Electronic Engng. 30* 324 (1958).
12. Lewin, L., Paine, J. Aerial Feeders. *Electronic Engng.* To be published.
13. Laico, J. P., McDowell, H. L., Noster, C. R. A Medium Power Travelling-Wave Tube for 6000Mc/s Radio Relay. *Bell Syst. Tech. J. 35*, 1285 (1956).
14. Craven, G., Lewin, L. Microwave Filters with Quarter-Wave Couplings. *Proc. Instn. Elect. Engrs. 103B*, 173 (1956).

All Travelling-Wave Tube Systems

By S. Fedida*, B.Sc.(Eng.) (Hons.), A.C.G.I., A.M.I.E.E.

The intrinsic features of all travelling wave tube microwave link equipment, of a type similar to the projected London to Norwich television radio relay, and their unique advantages and versatility in applications demanding a high grade of performance, are described in this article. Possible sources of distortion in travelling wave tubes are analysed and methods of reducing their effect to acceptable values, are indicated.

THE development and installation of broad-band microwave line of sight communications systems has been a major feature of the post-war increase in communications facilities, mainly in Europe and North America.

Early designs have tended to concentrate on valves of conventional design, while systems of later vintage have introduced travelling-wave tubes, at least in the output stage. Most of them, on the other hand, have provided the greater part of the gain required in the non-demodulating repeater, by the use of i.f. amplification.

At least one commercial design[1] exists, at the present time, wherein full advantage has been taken of the suitability of the travelling-wave tube for this type of service, to the extent of eliminating i.f. amplifiers altogether in the basic non-demodulating repeater.

The purpose of this article is to highlight some of the more relevant properties of travelling-wave tubes and the advantages of all travelling tube systems as applied to microwave communications links.

The Travelling-Wave Tubes

Although the travelling-wave tube was invented only 14 years ago[2], such is the speed of modern technological development, that it is now taken for granted and regarded as an almost conventional device. While great credit must be given to the pioneers who turned this epoch-making invention into a device available in commercial quantities, it is fair to say that further significant improvements in performance and utilization can be expected, as their use becomes more widespread and their properties better understood.

At the present time, they are available for a variety of duties, and for use at practically all microwave frequencies, from 1 000Mc/s upwards. Power outputs range from 1mW to 20W and gains from 20dB to 50dB, in a single envelope.

The medium power travelling-wave tube (1 to 10W) is generally operated at a helix potential of 2 to 3kV and takes currents of 40 to 60mA, while more recently tubes operating at helix potentials of about 1kV and taking higher currents have been introduced.

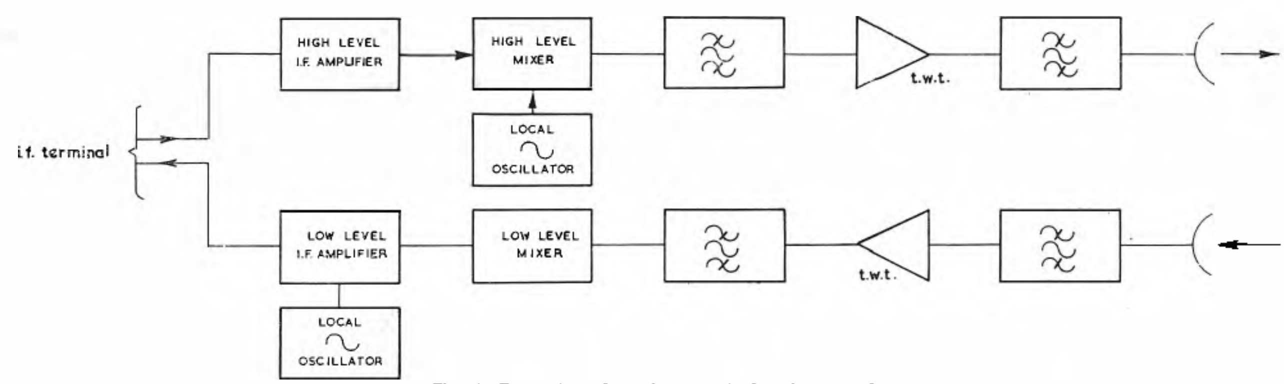


Fig. 1. Transmit and receiver terminals using t.w.t.'s

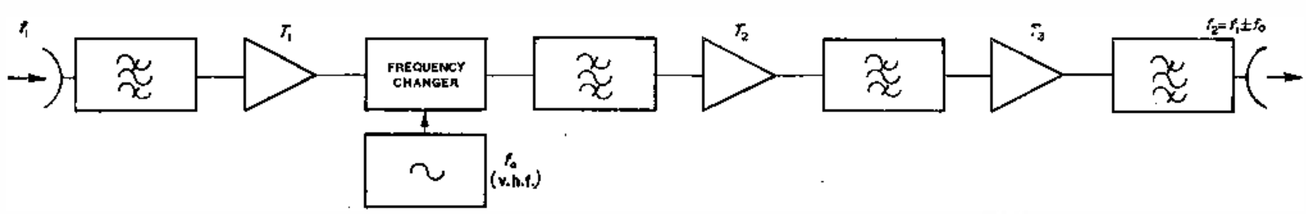


Fig. 2. Basic non-demodulating, all t.w.t. repeater (one way)

The intermediate power and low noise travelling-wave tubes operate at a helix potential of under 1kV, and take currents of 4 to 10mA, and 200 to 300μA respectively. Typical noise figures are 6 to 10dB for receiving tubes and less than 20dB for intermediate power tubes.

Electromagnetic field requirements for travelling-wave tubes vary between 100 and 200W, according to type. Several tube types are also available at present with integral or external permanent magnet focusing.

Structure of Typical Microwave Links

The structure of a typical microwave link featuring travelling-wave tubes is generally quite simple.

Terminals consist essentially of an r.f. amplifier, a frequency changer and associated filters. In transmitting terminals, a high level i.f.-to-microwave convertor is required, while in receiving terminals a low level microwave-to-i.f. convertor is necessary (Fig. 1).

Non-demodulating repeaters, not requiring channel

* *Marconi's Wireless Telegraph Co. Ltd.*

dropping facilities, are, if anything, simpler. They may consist of three or four travelling-wave tubes, a frequency changer and associated filters (Fig. 2).

Channel dropping facilities may be incorporated quite readily, as shown in Fig. 3, at the expense of an additional frequency changer in the through signal chain, but only when channel addition and extraction is to be performed at a standard i.f. frequency.

The structure of the corresponding i.f. repeater is shown in Figs. 4 and 5, where an indication of the increased complexity is evident. This point will be considered again later.

requires circuit realignment, in wideband i.f. amplifiers may, with travelling-wave tubes, be done without difficulty, provided means are adopted to ensure that the return loss at the input and output terminals is maintained at the requisite value.

One way of attaining this objective is a matter for the tube designer. It is a function of the uniformity of the helix, the impedance matching properties of the attenuating elements within the tube, and the design of the helix to transmission line transducers, at the input and output terminals of the tube. The problem is, however, difficult to resolve, particularly in the case of the medium power tube,

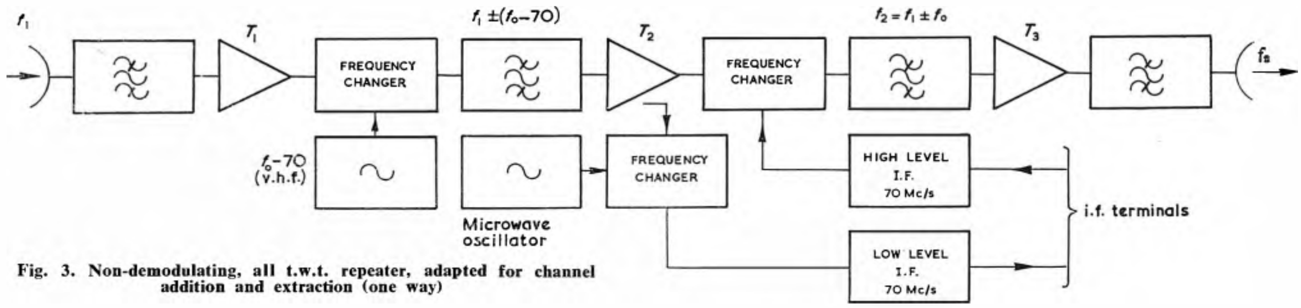


Fig. 3. Non-demodulating, all t.w.t. repeater, adapted for channel addition and extraction (one way)

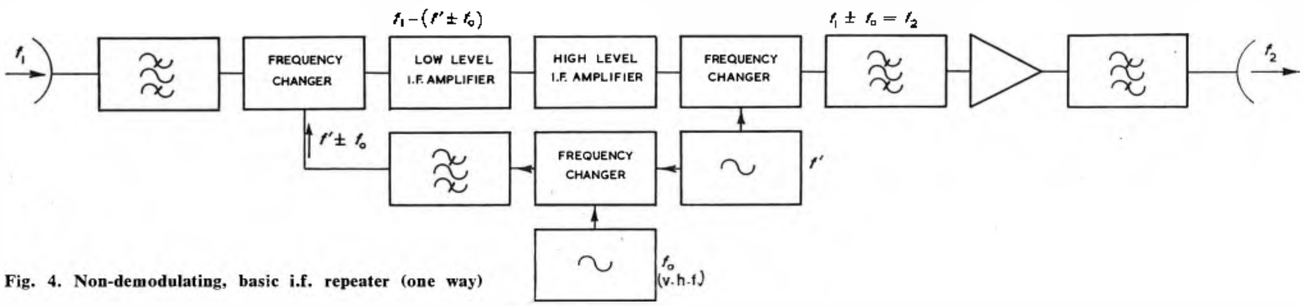


Fig. 4. Non-demodulating, basic i.f. repeater (one way)

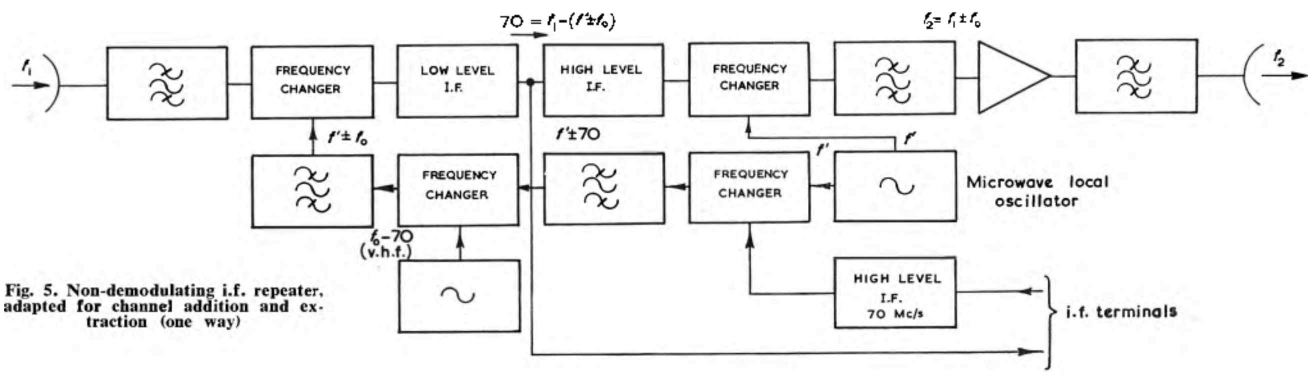


Fig. 5. Non-demodulating i.f. repeater, adapted for channel addition and extraction (one way)

Functional Advantages of Travelling-Wave Tubes

Bandwidth

One of the most well-known, and at the same time the most remarkable property of travelling-wave tubes is their extraordinarily large bandwidth. For example[3], a typical travelling wave tube operating at a centre frequency of 6 000Mc/s has, in the low level signal condition, a bandwidth of well over 2 000Mc/s.

Admittedly, this very large bandwidth is, at present, much greater than may be usefully employed in conventional microwave links, but it has the decided advantage of eliminating the need to use difficult-to-align interstage circuits, which are a source of great difficulties, in wideband i.f. amplifiers.

Another advantage is that tube changing, which often

where the attenuator is situated in the first half of the helix, with the result that reflections at the attenuator are greatly amplified and their rate of change of phase with frequency is very high. In spite of these difficulties, return loss figures of greater than 26dB, over bandwidths of the order of 40Mc/s around any frequency within the operating range of the tube, are currently obtained[4].

A solution which is now gaining considerably more favour is to use ferrite isolators. These have now reached a sufficiently advanced state of development as to be cheap, reliable and readily available.

When isolators are used, tube changing may be accomplished without requiring any readjustments whatsoever to the input and output circuits. Furthermore, they minimize the effect of long line echoes, which are accentuated when travelling-wave tubes are connected directly to long feeders.

Another development of a mechanical nature, intended to facilitate the changing of tubes, is encapsulation. The whole tube, complete with helix to coaxial line transducer, is enclosed in a metal shell fitted with input and output coaxial connectors[5]. Apart from the advantage of rigidity and ruggedness obtained thereby, the changing of tubes may be entrusted quite safely to relatively unskilled personnel, if necessary. There is hardly any need to emphasize the importance of this feature, particularly in installations located in sparsely populated territories.

The penalty for encapsulation is loss of available gain which, in the case of interstage tubes, is quite unimportant. Loss of gain at the input of the low noise tube and at the output of the power tube is seldom acceptable, unless it is taken into account in the early tube and link design stage and other parameters adjusted accordingly to compensate for it.

Encapsulation tends, in certain cases, to increase the capital cost of the equipment. This is due to the fact that encapsulated tubes do not in general incorporate facilities for the adjustment of the return loss at the input and output terminals. In applications where high transmission quality is essential, as in the case of multi-channel telephone systems and television systems conveying a music channel, or colour information, isolators must be used in large numbers and this may be uneconomical.

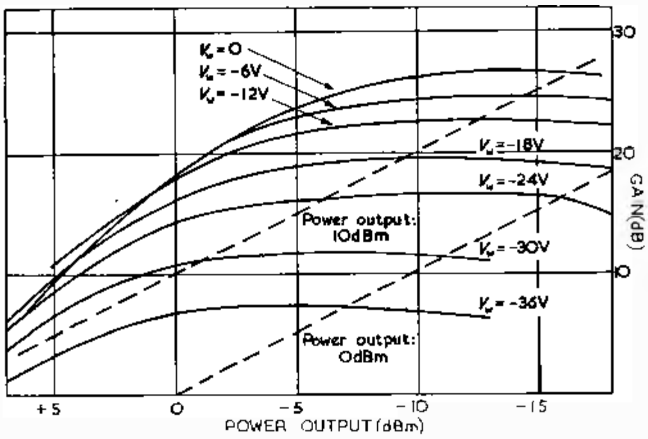


Fig. 6. Gain characteristic of an intermediate travelling-wave tube (At various values of Wehnelt voltage)

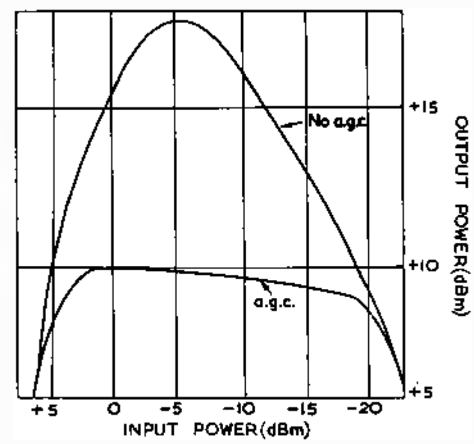


Fig. 7. Gain characteristic, with and without a.g.c. of an intermediate power travelling-wave tube

Automatic Gain Control

The total amount of gain required at microwave link repeaters is usually of the order of 70dB, for favourable propagation conditions. A fading margin of 20 to 30dB must also be allowed for by the provision of additional gain and an automatic gain control arrangement.

The provision of automatic gain control in i.f. amplifiers does not introduce excessive difficulties, provided the control is applied to low impedance stages. Automatic gain control may also be readily applied to travelling-wave tubes, but in a repeater of the type shown in Fig. 2, only one tube (tube T_2) is available for this application. Gain control on the other tubes results in a degradation of noise figure or a reduction in power output, and neither of these results is acceptable.

Typical gain characteristics for a low power travelling wave tube, at various values of Wehnelt voltages, are given in Fig. 6. It may be seen that a gain variation of about 20dB may be obtained for a Wehnelt voltage change of 36V, at an output power level of 1mW. Approximately the same voltage change is also required at an output level of 10mW.

An input/output characteristic for the same travelling-wave tube, with and without a.g.c., is shown in Fig. 7. The output is maintained constant to within $\pm 0{\cdot}5$dB of 10mW, over a power input range of 20dB. A d.c. amplifier having a gain of about 600 was used to amplify the variations in output, which was monitored by a silicon diode rectifier, connected to the output of the travelling-wave tube through a 10dB directional coupler. In a repeater application this monitor should be coupled to the power output tube, provided this tube is not limiting, so as to take advantage of the additional gain provided. Thus the gain of the a.g.c. amplifier need not be more than about 40dB, in order to provide adequate control.

The application of a.g.c. to a repeater has an important bearing on the gain requirements and the power handling capacities of each one of the travelling-wave tube types used. In a typical repeater the input power may be of the order of 1μW, and the output power of the order of 10W. Where a.g.c. is obtained by the application of bias to the Wehnelt of a travelling-wave tube, both gain and power handling capacities are reduced to approximately the same extent. It follows that the ratio of saturation power to operating power in a gain controlled t.w.t. should be at least equal to the a.g.c. range required. It is clearly important, therefore, to reduce the operating output power of this tube to the absolute minimum.

This calls for the removal of the second frequency changer (Fig. 3), from the output to the input of the intermediate tube, and since the output tube is not likely to have a noise figure of much better than about 30dB, it follows that at a drive of 1mW the noise power contribution of this tube is about one-tenth that of the low noise tube, assuming the latter to have a noise figure of 10dB. (It is usually lower). Similarly the drive to the intermediate tube must not fall below about 100μW.

Thus the requirement for a.g.c. in a repeater calls for a power tube gain of 40dB approximately, at an output power of 10W (or 37dB at 5W), for an intermediate tube saturation power of at least 250mW, with a gain of 30dB at 100mW output, and for a low noise tube saturation power of about 10mW, with a gain of 32dB, at an input power of 1μW. This assumes that the loss of each frequency changer is no greater than 6dB.

A block diagram of an all t.w.t. repeater, fitted with a.g.c. and capable of channel insertion and extraction, is given in Fig. 8. Typical power levels along the repeater, under free space conditions, are also given.

The application of a.g.c. to a repeater (Fig. 9) when a total of four tubes is used, is relatively easier and the specifications imposed on the tubes and the frequency changers are easier to meet. These factors have to be weighed up against the cost of the additional tube with its associated mount and power supplies. It is, however, a simple solution which could be justified in the longer run as the cost of manufacturing tubes is progressively reduced. For a desired a.g.c. range of 20 to 25dB, the maximum power handling capacity of the intermediate tube need not exceed 100mW and the frequency changer loss may be up to 10dB

It may also be possible to transmit two broadband channels simultaneously, but the penalties to be incurred in a one-way repeater, in terms of reduced power output, are still to be fully investigated.

Frequency Changing

The use of travelling-wave tubes as frequency changers has already been dealt with in the technical literature[6] and in certain cases it is of advantage to employ them in this duty. There are, however, two major difficulties.

The drive required to operate a general purpose low power tube efficiently in this mode is of the order of 10 to

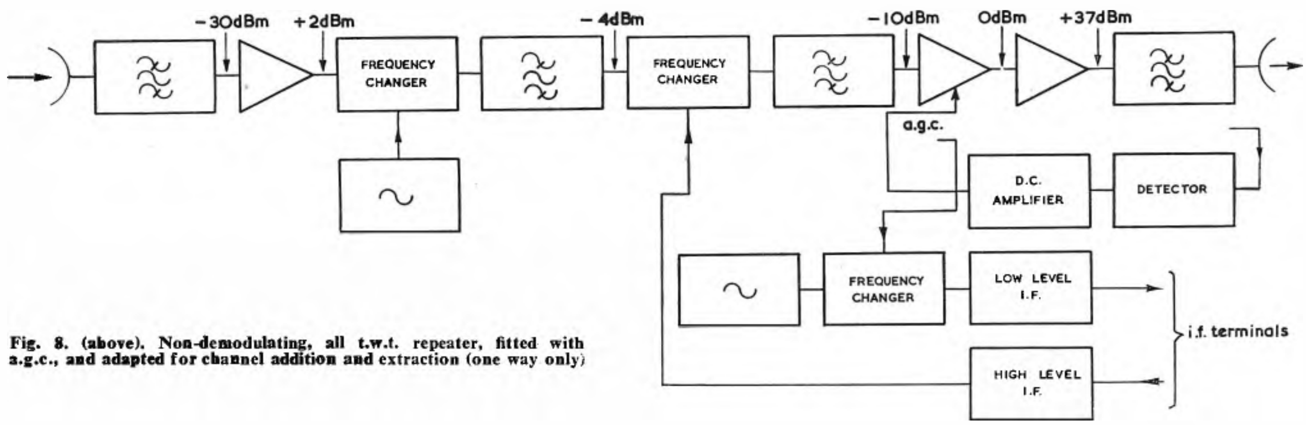


Fig. 8. (above). Non-demodulating, all t.w.t. repeater, fitted with a.g.c., and adapted for channel addition and extraction (one way only)

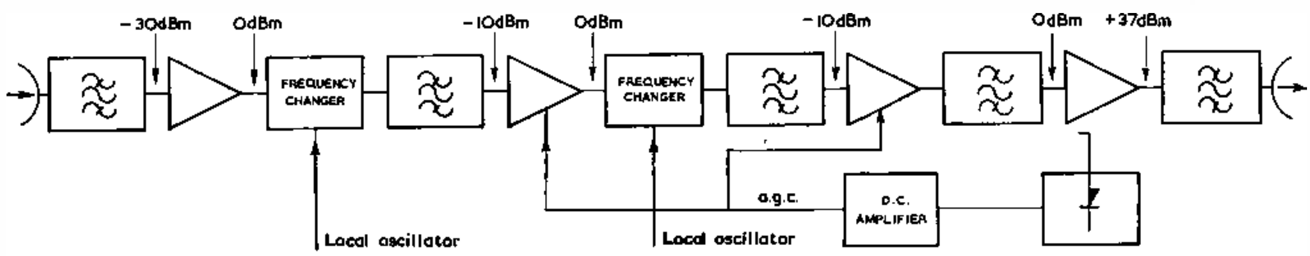


Fig. 9. Application of a.g.c. to four tube repeater

Furthermore the intermediate tubes are operated at approximately 10dB below saturation power.

A 'reflex' mode of operation, which combines the cheapness of the three tube repeater with the high gain of the four tube repeater, is made possible by virtue of the very large bandwidth of travelling-wave tubes. Here (Fig. 10) the signal makes two or more transits through the same tube, and its frequency is changed after each transit except the last one. Intermodulation effects are, of course, liable to occur at certain power levels and the correct choice of tube power-handling capacity must be made, if these effects are to be maintained at an acceptably low value. Recent tests with 600 channel noise loading have indicated that the intermodulation level in a three tube reflex repeater is no higher than in the corresponding four tube repeater.

The above measurements indicate that a one-way t.w.t. repeater may well be used in certain circumstances as a two-way repeater, as indicated in Fig. 11. The potentialities of this arrangement have not been fully investigated yet but it seems that, provided the level of one of the signals is about 20dB lower than the level of the other, the intermodulation noise is not detectable. One possible application of this system is for the transmission of supervisory information and/or a small number of telephone channels along a normally unidirectional link, such as may be used for the relaying of television only.

15V, in a capacitive impedance of about 30 to 40pF. As long as the bandwidth of the drive circuit is not required to be large, i.e. when the local oscillator has a fixed frequency, there are no great difficulties in providing the requisite drive, and the saving of a special microwave mixer is then worthwhile. If, however, the bandwidth must be large, as in the case of the channel insertion mixers, then it is more economical to provide a separate diode mixer, which requires a drive of only about 1V r.m.s.

Another objection to the use of a travelling-wave tube as a mixer results from the reduction in the saturation power by about 6dB.

If a.g.c. is to be applied to a tube operated as a mixer, it follows that the control range is reduced by about 6dB. This would be acceptable in a four tube repeater, but not in a three tube repeater.

Noise Figure and Input Impedance of Repeater

At the present stage of microwave mixer technique the low noise tube has a noise factor advantage of 4 to 6dB over the best silicon wide band mixer available. This is a considerable difference and the resulting advantages for the t.w.t. system are an increased reliability and protection against fading and a reduction in the cost of ancillary equipment such as aerials, towers and feeders, for a given overall performance .A further advantage derives from the fact that the return loss at the input of a low noise travelling-wave tube may be made to exceed 30dB over a bandwidth of 50 to 100Mc/s, given a correct transducer design. While silicon diode mixers have been made with an input return loss exceeding 26dB over bandwidths of the order of 20Mc/s, the performance is subject to appreciable degradation in service and invariably with change of diodes. The readjustment of the mixer, after a change of diode, is not an operation which can be satisfactorily carried out in the field. When isolators are used, the cost of the mixer unit is increased and the noise factor degraded by up to 1dB.

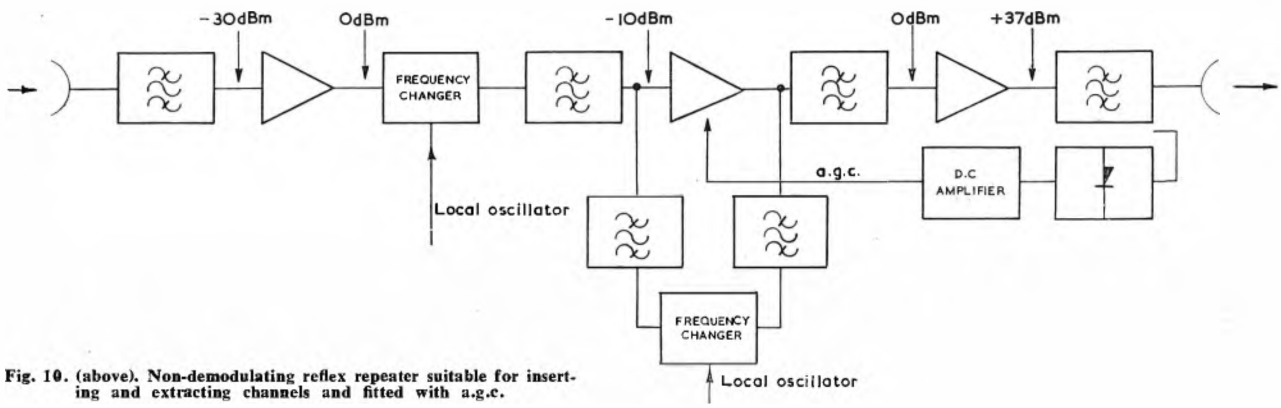


Fig. 10. (above). Non-demodulating reflex repeater suitable for inserting and extracting channels and fitted with a.g.c.

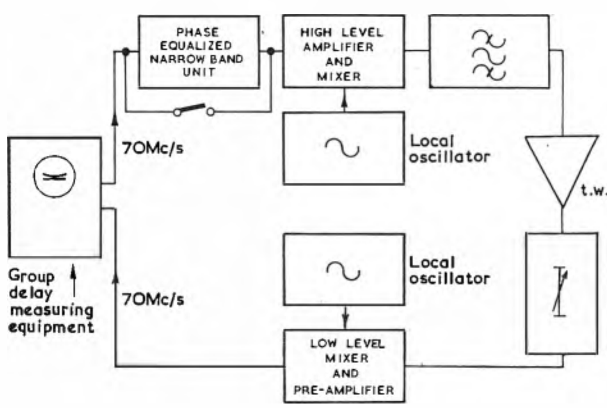


Fig. 11. Measurement of a.m. to p.m. conversion in travelling-wave tubes

Permanent Magnet Focusing

The power consumption of travelling-wave tubes is quite modest. Discounting the output tube, which is normally a necessity in both i.f. and r.f. repeaters, the total power consumption of the additional tubes including cathode heating is less than 30W.

The focusing electromagnets, on the other hand, require a minimum of 100W each and great savings may be made in this direction, by substituting permanent magnets.

Travelling-wave tubes may be focused with uniform fields or periodic fields[7]. The use of the latter has been confined so far to experimental equipment but medium and intermediate power tubes, featuring the former, are available commercially and have been incorporated in commercial equipment.

Low noise tubes[8] may also be focused with permanent magnets, but the design of a suitable periodic structure, capable of giving adequate focusing with no degradation of noise figures, is not easy to achieve. A suitable arrangement using bar magnets and a uniform field presents no difficulties.

A permanent magnet periodic field system has the great advantage over a uniform field system in that the stray field is extremely small and the internal field is consequently very insensitive to the proximity of ferrous materials. On the other hand, and more particularly at the higher microwave frequencies where the diameter of the helix becomes smaller, the tolerances on the uniformity of the individual ring magnets become more difficult to meet.

Experiments performed in this laboratory on the type N1001 tube have indicated that, provided the field at the gun end of a travelling-wave tube is correctly shaped, extremely good focusing performance may be obtained with a periodic structure made up of Magnadur II ferrite magnets and soft iron pole pieces. No difficulties were experienced with tolerances in the magnetization of the individual ring magnets, and a single adjustment was sufficient to obtain good focusing (i.e. better than 99 per cent transmission) when changing tubes.

Recently, experimental tubes featuring internal electrostatic periodic focusing have also been made and one can have no doubt that advances in this and other fields will soon eliminate the need for bulk and expensive electromagnets.

Amplitude to Phase Conversion

An important effect occurring in all electron devices, and particularly in present-day travelling-wave tubes operating near the power saturation limit, consists in a change of carrier phase corresponding to a change of carrier amplitude.

For a typical medium power travelling-wave tube[3], the rate of change of carrier phase with amplitude is of the order of 2·5°/dB at a level of 5W and 6°/dB at a level of 10W, always assuming that the helix potential is set at the value for which the tube gain is at a maximum at the stated output power.

This has two effects. The one is to give an apparent change to the frequency response and the other to the group delay characteristic of the system.

The change in the frequency response ocurs in this way. If it is assumed that the frequency modulated carrier applied to the t.w.t. has a small amount of incidental amplitude modulation, defined by a constant slope of xdB/Mc/s (x may be positive or negative), the t.w.t. will be phase modulated in synchronism with the frequency modulation signals. For a sinusoidal modulation frequency f_m and a peak carrier deviation Δf, the effective peak phase deviation is:

$$\Delta\phi = x \,.\, \Delta\theta \,.\, \Delta f \,.\, (\pi/180) \qquad (1)$$

where $\Delta\theta$ is the rate of a.m. to phase conversion degrees per decibel.

This phase deviation, which occurs at a rate f_m is equivalent to an interfering frequency deviation δf, given by:

$$\delta f = f_m \Delta\phi$$

$$\delta f = (\pi/180) \, . \, x\Delta\theta \, . \, \Delta f \, . \, f_m \quad \ldots\ldots\ldots\ldots \quad (2)$$

The ratio of interference δf to the signal deviation Δf, which is

$$\delta f/\Delta f = (\pi/180) \, . \, x\Delta\theta \, f_m \quad \ldots\ldots\ldots\ldots \quad (3)$$

is clearly only a function of the amplitude slope of the applied carrier, the rate of change of phase with amplitude of the travelling-wave tube and the modulation frequency. The actual frequency deviation Δf is irrelevant.

The resulting output video response is

$$A_t = 1 + (\pi/180) \, x \, . \, \Delta\theta \, . \, f_m \quad \ldots\ldots\ldots\ldots \quad (4)$$

When x is positive the video response rises at high modulation frequencies. It falls at high frequencies when x is negative.

To gain some idea of the order of magnitudes involved, let us assume a typical value of $x = \pm 0{\cdot}1$dB/Mc/s, $\Delta\theta = 2{\cdot}5°$/dB, and $f_m = 4$Mc/s (1 000 channel system). Under these circumstances,

$$A_t = 1 \pm (\pi/180) \times 0{\cdot}1 \times 2{\cdot}5 \times 4$$

$$A_t = 1 \pm 0{\cdot}0175 \, (\pm 0{\cdot}15\text{dB})$$

Thus the video response may have a gain or a loss of up to 0·15dB at 4Mc/s

In practice an amplitude slope x of 0·1dB/Mc/s is considerably worse than may be fairly easily achieved, although one must take into account the effect of ageing and misalignment due to thermal drifts.

It has been assumed, so far, that the frequency response of the circuits preceding the t.w.t.s has a constant slope. This again is not in general likely, and one must think in terms of an amplitude function $A(f)$, with important higher order terms.

$$\text{If } A = A_o + A_1(f - f_o) + A_2(f - f_o)^2 + \ldots \quad \ldots (5)$$

and the phase ϕ of the carrier is

$$\phi = \phi_o + \phi_1 A \quad \ldots\ldots\ldots\ldots \quad (6)$$

the rate of change of phase with amplitude is

$$(d\phi/dA) = \phi_1 \text{ rad/V} \quad \ldots\ldots\ldots\ldots \quad (7)$$

The relationship between ϕ_1 and the $\Delta\theta$ used in equation (1) is

$$\Delta\theta = (180/\pi) \frac{A_o\phi_1}{20 \log e} \quad \ldots\ldots\ldots\ldots \quad (8)$$

since

$$\Delta\theta = (180/\pi) \frac{\delta\phi}{20 \log \dfrac{(A_o + \delta A)}{A_o}} = (180/\pi) \frac{\delta\phi}{20 \log e \, . \, (\delta A/A_o)}$$

$$\Delta\theta = (180/\pi) \, . \, \frac{A_o\phi_1}{20 \log e} \, . \, (\delta\phi/\delta A) \sim (180/\pi) \frac{A_o\phi_1}{20 \log e}$$

The phase response resulting from an amplitude response of the type given above is:

$$\phi = \phi_o + \phi_1[A_o + A_1(f - f_o) + A_2(f - f_o)^2 + \ldots] \quad \ldots (9)$$

If the carrier has a sinusoidal frequency deviation

$$f - f_o = \Delta f \sin pt \quad \ldots\ldots\ldots\ldots \quad (10)$$

the distortion caused by spurious phase modulation is:

$$1/2\pi \, (d\phi/dt) = (1/2\pi)\{\phi_1 A_1 \Delta f \, . \, p \cos pt + 2\phi_1 A_2 \Delta f^2 p \sin pt \cos pt + \ldots\} \quad \ldots (11)$$

and the spurious relative frequency deviations at the fundamental and second harmonic frequencies are consequently:

$$(H_1/\Delta f) = \phi_1 A_1 f_m \quad \ldots\ldots\ldots\ldots \quad (12)$$

$$(H_2/\Delta f) = \phi_1 A_2 \Delta f \, f_m \text{ where } f_m = (p/2\pi) \quad \ldots (13)$$

Again one may here express the coefficients A_1, A_2 etc. in terms of slopes as above.

If x and y are the slopes of the amplitude response, in terms of the first order (linear) and second order (parabolic) variations expressed in dB/Mc/s and dB/Mc/s², then

$$x = \frac{20 \log \dfrac{A_o + \delta A}{A_o}}{\delta f} = \frac{20 \log (1 + (\delta A/A_o))}{\delta f} \quad \frac{20 \log e}{A_o} \, . \, \delta A/\delta f$$

$$y = \frac{20 \log \dfrac{A_o + \delta^2 A}{A_o}}{\delta f^2} = \frac{20 \log(1 + (\delta^2 A/A_o))}{\delta f^2} \sim \frac{20 \log e}{A_o} \, . \, \delta^2 A/\delta f^2$$

and

$$x \sim \frac{20 \log e}{A_o} A_1 \quad \ldots\ldots\ldots\ldots \quad (14)$$

$$y \sim \frac{40 \log e}{A_o} A_2 \quad \ldots\ldots\ldots\ldots \quad (15)$$

therefore

$$H_1/\Delta f = (\pi/180) \frac{20 \log e}{A_o} \Delta\theta \, . \, \frac{A_o}{20 \log e} \, . \, x \, . \, f_m = (\pi/180) \, \Delta\theta \, . \, x \, . \, f_m \quad \ldots\ldots \quad (16)$$

$$H_2/\Delta f = (\pi/180) \frac{20 \log e}{A_o} \Delta\theta \, . \, \frac{A_o}{40 \log e} \, y \, . \, \Delta f \, . \, f_m = (\pi/360) \, \Delta\theta \, . \, y \, . \, \Delta f \, . \, f_m \quad \ldots\ldots \quad (17)$$

Assuming that the response curve is an exact parabola and that it falls by 1dB, below the mid-frequency response, at the edges of a 20Mc/s bandwidth, the value of y is 10^{-2}dB/Mc/s². For the parameters assumed above, and $\Delta f \simeq 1{\cdot}8$Mc/s, one would expect

$$H_2/\Delta f = (\pi/360) \times 2{\cdot}5 \times 10^{-2} \times 1{\cdot}8 \times 4$$

$$H_2/\Delta f = 1{\cdot}57 \, . \, 10^{-3} \, (-56\text{dB})$$

An alternative way of expressing the distortion created by a.m. to p.m. conversion is to regard is as causing a change in group delay. This quantity may be expressed as

$$d\phi/d\omega = (\partial\phi/\partial\omega) + (\partial\phi/\partial A) \, . \, (dA/d\omega) \quad \ldots\ldots \quad (18)$$

It has been seen already that

$$(\partial\phi/\partial A) = \phi_1 (\pi/180) \frac{20 \log e}{A_o} \Delta\theta$$

$$(dA/d\omega) = (1/2\pi) \, (dA/df) = (1/2\pi) \, (A_1 + 2A_2\Delta f + \ldots)$$

$$= 1/2\pi \left[\frac{A_o}{20 \log e} \, x + \frac{A_o}{20 \log e} \, y \, . \, \Delta f + \ldots\right]$$

Therefore

$$(d\phi/d\omega) = (\partial\phi/\partial\omega) + (\Delta\theta/360) \, (x + y\Delta f + \ldots) 10^3 \text{m}\mu\text{sec.} \quad \ldots\ldots \quad (19)$$

The constant slope x in the amplitude response results in a fixed increase of group delay $\Delta\theta x/0{\cdot}360$, while a parabolic shape defined by y results in a skewness in the group delay characteristic of slope $\Delta\theta \, . \, y/0{\cdot}360$.

If $x = 0{\cdot}1$dB/Mc/s, $y = 0{\cdot}01$dB/Mc/s² and $\Delta\theta = 2{\cdot}5°$/dB, then the fixed increase in group delay is

$$d\phi/d\omega = \frac{0{\cdot}1 \times 2{\cdot}5}{0{\cdot}360} = 0{\cdot}25/0{\cdot}36 = 0{\cdot}7\text{m}\mu\text{sec}$$

and the slope of the characteristic

$$d^2\phi/d\omega^2 = \frac{0{\cdot}01 \times 2{\cdot}5}{0{\cdot}360} = 0{\cdot}07\text{m}\mu\text{sec/Mc/s.}$$

The figures for second harmonic distortion and skewness of group delay appear to be consistent with one another and they both indicate that the resulting intermodulation distortion in a 1 000 channel system would be just tolerable.

Tests taken on an all t.w.t. repeater loaded with a 600 channel noise signal have shown that the equivalent noise power ratio is practically independent of the output power of the repeater, over a range of +3dB to −10dB relative to saturation power.

The direct measurement of a.m. to p.m. conversion, in a travelling wave tube, is a difficult operation even if the relatively simple method indicated in Ref. 3 is used. Unless a very high degree of accuracy in the measurement of relative levels can be obtained, this method gives unreliable results.

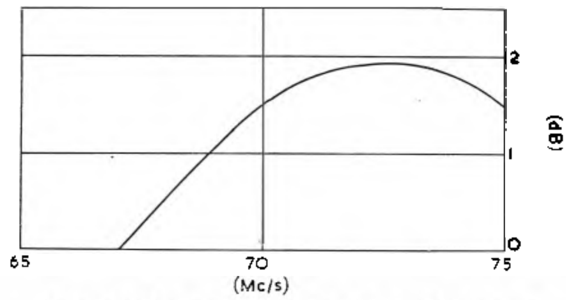


Fig. 12(a). Overall amplitude characteristic at low level

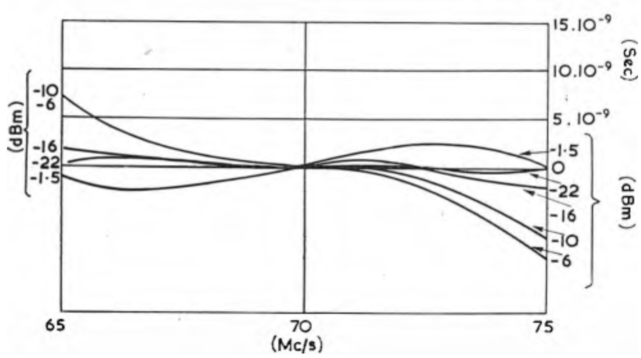


Fig. 12(b). Overall group delay characteristic at various input levels to the travelling wave tube

An alternative method consists in measuring the group delay of a transmission system, including a travelling-wave tube and a phase equalized narrow band circuit, arranged as shown in Fig. 11. The results of some early measurements, which are given in Figs. 12(a) and 12(b), illustrate vividly the change in group delay characteristic, with magnitude of drive to the travelling wave tube.

There are, of course, several ways of minimizing the effects of a.m. to p.m. conversion. The amount and the phase of the a.m. is not, in general, the same from hop to hop, since it is generated within each repeater by the slight misalignments which are bound to occur throughout the working life of the component. It is therefore to be expected that there will tend to be a smoothing out, or cancellation, of the a.m. to p.m. effects from one hop to the next. The major distortion occurs at high frequency and it is possible to insert automatically controlled amplitude and group delay equalizers, at selected points, to give the necessary correction.

However, it is preferable, for the sake of simplicity, to operate all tubes 5 to 10dB below their saturation level, when an order of magnitude reduction in a.m. to p.m. conversion may be gained. This mode of working may be arranged fairly readily with the low noise and intermediate tube, although even with the latter, it may be necessary to strike a compromise with the a.g.c. requirements. It is not in general practicable to operate the power tube as much as 10dB below saturation level and other palliatives have to be resorted to in this case.

One of these consists in amplifying and feeding back to the helix in the appropriate phase, the a.m. ripple existing at the output of the travelling-wave tube. This produces an opposing phase modulation which may be made to cancel out the internally generated one. The gain required in the feedback amplifier is quite low (of the order of 20 to 25dB), as the phase sensitivity of the helix in a medium power tube is of the order of 2°/V. At this value of gain, the feedback amplifier is capable of providing a very wide bandwidth without any difficulty, and a reduction in the level of a.m. to p.m. conversion by an order of magnitude is quite feasible. Since the helix phase modulation sensitivity of the intermediate tube is up to five times higher than that of the output tube, the gain of the feedback amplifier can be reduced appreciably if the control voltage is applied earlier in the chain

It is known that the slope of the phase versus amplitude curve in a travelling-wave tube changes sign beyond saturation. It follows that some degree of cancellation may be obtained by operating the intermediate tube slightly above the saturation level and the power tube slightly below. Also, provided adequate gain has been made available, the power tube could well be operated slightly beyond saturation, where the phase modulation sensitivity falls to zero. The penalty in terms of power output, which is less than 1dB in this mode of operation, should be quite tolerable.

Recently[4] medium power travelling-wave tubes have been built with two helices of different pitches, placed end to end, within the same envelope. The saturation characteristic of a tube of this construction is much more gradual and the difference between low level and saturation gain is only 3dB, as against 7 to 8dB for a conventional tube with uniform helix. The linear gain range is therefore greatly extended and it is anticipated that a lower a.m. to p.m. conversion factor may thereby be achieved at a high power output.

Group Delay Equalization

Apart from the effect of compression within the tubes the other sources of group delay variations, in a non-demodulating repeater, are the microwave filters and the aerial feeders.

The items peculiar to all t.w.t. repeaters are the microwave filters, of which there are between four and six. These are normally adjusted on a swept reflection coefficient instrument, when a v.s.w.r. of under 1dB, over a band of ±10Mc/s, and under 0·25dB over a central bandwidth of ±5Mc/s, is readily obtained.

Because the bandwidth of microwave filters is a very small proportion of the centre frequency, a maximally flat amplitude response allied to a symmetrical group delay response is generally obtainable. The theoretical group delay responses of four and five cavity, maximally flat filters, tuned to 2 000Mc/s, are shown in Fig. 13. Measurements on actual filters, aligned on an impedance basis, show group delay variations approximately equal to the theoretical figures as indicated in Fig. 14.

Group delay curves such as those of Fig. 13 and Fig. 14 are equalized quite readily at i.f. or at microwave frequencies. It is preferable to equalize the delay of repeaters on an individual basis, and in the case of t.w.t. repeaters this should be done at radio frequencies. The advantage of radio frequency equalizers is that their thermal drift may be made to compensate for the thermal drift of the filters and, because their group delay response is very symmetrical, they may be made to match the delay response of the filters quite accurately.

A simple type of r.f. group delay equalizer consists of a 3dB coupler terminated in a pair of cavity resonators at two of the conjugate arms, 2 and 3. It can be shown that the network between the other pair of conjugate arms, 1 and 4, behaves as an all pass quadripole, the phase characteristic of which is that of the terminating resonators.

By adjusting the tuning and the coupling (loaded Q's) of the resonators, a wide range of group delay responses may be obtained. A simple way of adjusting the loaded Q of the resonators, when these are of coaxial construction, is to interpose a variable quarter-wave transformer, made up of a rotatable strip line, between the resonators and the 3dB coupler.

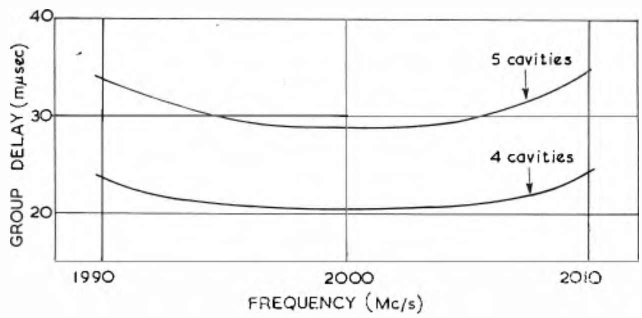


Fig. 13. Group delay of 4 and 5 cavity maximally flat filters adjusted for a return loss of under 26dB over a bandwidth of 20Mc/s

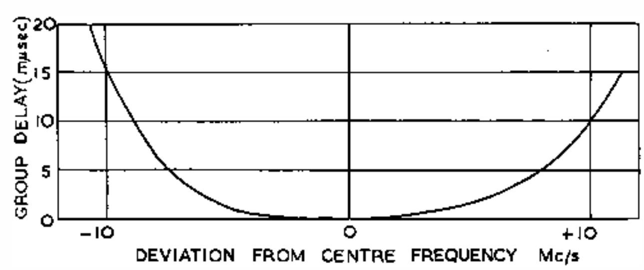


Fig. 14. Measured group delay of 3 travelling wave tubes and four unequalized microwave filters

Discussion

The topics discussed in the preceding paragraphs indicate that all-travelling-wave tube repeaters meet all the specific requirements of microwave link work, including the intermodulation requirements of multi-channel telephone transmission, (at least 600 channels) and 625 line television. Further investigations are necessary in order to extend their transmission capabilities to 2 000 channels and over.

Travelling-wave tubes there must be in any case, whether one is thinking of i.f. or all-t.w.t. repeaters. Admittedly, improved microwave triodes are now available for use at frequencies up to 4 000Mc/s, but they represent, at best, a difficult compromise and the microwave circuits which they require are complicated and expensive to manufacture. A typical example illustrating the use of these triodes is given in Ref. 10, where it is found necessary to use as many as three stages to provide an overall gain of 30dB with an output power of 1·5W. A single travelling-wave tube designed for link work provides 35 to 40dB gain, at an output power of 10W. Compared with i.f. repeaters, all-t.w.t. types score on the ground of simplicity, reliability and cost.

If, now, one examines Figs. 2 and 4, and assuming that the output stage of an i.f. repeater is a travelling-wave tube, the following table can be prepared, in which a comparison is made of the equipment (excluding some common items) involved in t.w.t. and i.f. repeaters:

All T.W.T. Repeater	*I.F. Repeater*
Low noise tube	Low level microwave mixer
Microwave filter	Low level i.f. amplifier
Intermediate t.w.t.	High level i.f. amplifier
High level microwave mixer	High level microwave mixer
Crystal controlled v.h.f. local oscillator	Crystal controlled v.h.f. local oscillator
	Microwave local oscillator
	High level microwave mixer
	Microwave filter

It is clear from this table that, in the case of the basic non-demodulating repeater, the t.w.t. type involves less equipment, is considerably more economical in valves and, provided the t.w.t. focusing field is supplied by permanent magnets, is far more economical in power consumption. All these factors add up to a considerable increase in reliability in favour of the t.w.t. type, particularly when one realizes that the number of working valves in the actual signal path is so much smaller than in the i.f. type, and that the life of well-designed travelling-wave tubes is normally of the order of 10 000 hours or more.

The same conclusions may be derived by comparing Figs. 3 and 5.

In addition to these features, the normal drift in the group delay characteristic of t.w.t. repeaters may be more easily compensated, since the equalization circuits are of the same nature and they operate at the same frequencies as the microwave filters. This is not altogether the case with the i.f. repeaters.

Finally, the very large bandwidth of the t.w.t. allows two-way working in a one-way repeater when one of the signals, because of its narrow bandwidth (supervisory or small number of telephone channels), need not be transmitted at the full carrier level. Further work holds the possibility of achieving two-way working on broad-band signals.

Conclusion

It has been shown that all t.w.t. repeaters are capable of meeting all the known requirements of long haul microwave links and that they are superior to the i.f. type in terms of simplicity and reliability. When travelling-wave tubes are specially designed for this application, this type of repeater should also be more economical in terms of initial outlay and running cost.

Acknowledgments

The author wishes to thank the Engineer-in-Chief, Marconi's Wireless Telegraph Co. Ltd., for permission to publish this article, and his colleagues at the Baddow Research Laboratories and at the Writtle Development Laboratories who have contributed in providing some of the theoretical and experimental results quoted.

REFERENCES

1. Fedida, S. Wideband Microwave Transmission Systems. *Point to Point Telecommun. 1*, No. 3 (June, 1957).
2. Kompfner, R. The Travelling-Wave Tube Centimetre-wave Amplifier. *Wireless Engr. 124*, 255 (1947).
3. Laico, J. P., McDowell, H. L., Moster, C. R. A Medium Power Travelling Tube for 6000mc Radio Relay. *Bell Syst. Tech. J. 35*, 1285 (Nov., 1956).
4. Kleen, W. Les Tubes Hyperfrequence pour un Faisceau Hertzien a Large Bande. *L'Onde Electrique. 37*, 894 (1957).
5. Coulson, R. B. General Review of Travelling Wave Tubes. *Electronic Engng. 30*, 302 (1958).
6. Bray, W. J. The Travelling Wave Tube as a Microwave Phase Modulator and Frequency Shifter. *J. Instn. Elect. Engrs. 99*, Pt. 3. 15 (1952).
7. Mendal, J. T., Quate, C. F., Yocom, W. H. Electron Beam Focusing with Periodic Permanent Magnet Fields. *Proc. Inst. Radio Engrs. 42*, 800 (1954).
8. Chang, Kern K.N. Periodic Magnetic Field Focusing for Low Noise Travelling Wave Tubes. *RCA Rev. 16*, 423 (1955).
9. Beam, W. R., Blattner, D. J. Phase Angle Distortion in Travelling Wave Tubes, *RCA Rev. 17*, 86 (1956).
10. Andrieux, G. Amplifications de puissance a triodes pour 4 000Mc/s. *L'Onde Electrique. 37*, 777 (1957).

Portable U.H.F.-S.H.F. Links

In the BBC Television Service

By T. H. Bridgewater*, M.I.E.E.

The following article discusses the applications of u.h.f. and s.h.f. radio links for television outside broadcast purposes. These links are used in large numbers and to ensure efficiency must be provided with specialized auxiliary and test equipment. There is particular reference to a new BBC-designed 600Mc/s f.m. link, the first of its kind to come into service.

RADIO links are used by the BBC Television Service in connexion with television programmes whenever the equivalent circuit (whether cable or radio) facilities are unavailable from Post Office resources. They may be required for any or all of the following purposes, jointly or separately:

(1) Carrying vision signals over some or all of the distance between the camera site and nearest convenient point of connexion with the national vision network.

(2) Carrying sound signals over some or all of the distance between the microphone at the camera site and the nearest convenient point of connexion with the Post Office or BBC network.

(3) Providing communication for controlling the vision or sound when either or both of these are being conveyed by radio link. Such communication links would usually operate in both directions.

(4) Various short-range applications to replace cables which otherwise would restrict the movement of equipment and personnel either in a studio or, more often, on the site of an outside broadcast. Examples would be:—

(*a*) Radio microphone. This is the case where the microphone is being worn or carried by, say, a commentator who requires freedom of movement without trailing a cable behind him.

(*b*) The equivalent facility applied to a small hand-held television camera.

(*c*) Communications—usually two-way—with each of the above, and also for sundry other local purposes, such as between the Stage Manager in a television studio and his producer in the adjacent control room.

For applications of this kind the range would be measured in hundreds of yards rather than miles.

Types of Link

Many types of radio link are in use for these various purposes with frequencies in the v.h.f., u.h.f. and s.h.f. ranges. Power outputs vary but are usually of the order of a few watts, with a reliance on aerial gain to increase the e.r.p. whenever necessary.

Vision Links

It is simplest to discuss these by frequency ranges.

S.H.F.

The greatest number of links falls into this category. The principal frequencies at present in use are in the range 4 400 to 4 800Mc/s approximately. These allocations are not permanent and later it will be necessary to move to the officially assigned region of 7 050 to 7 300Mc/s. Two main types of equipment are in regular use, viz.: E.M.I., types ML6A, ML6B, ML6C (Fig. 1) and the S.T. & C. type.

These are lightweight portable transmitters and receivers and the aerial system uses paraboloids 4ft in diameter having a beamwidth of $\pm 2\frac{1}{2}°$ to the half-power points. These aerials are fitted on panning heads which, in turn, are carried on substantial metal tripods. The aerials can be traversed through 360° and accurately orientated to within $\frac{1}{2}°$. The gain of each reflector with respect to an isotropic aerial is 33dB. The equipments are divided into separate units for accessibility, ease of handling and convenience of transport. In the case of the E.M.I. ML6A and ML6B equipments, the head units are mounted in front of the paraboloid, but both the new ML6C and existing S.T. & C. equipments have head units mounted on the rear of the paraboloid. The transmitter

* *The British Broadcasting Corporation.*

head units containing the oscillator and modulator are connected to the control and power supply units by means of multi-core cable up to 200ft in length. The receiver head unit, containing the s.h.f. local oscillator, mixer and i.f. pre-amplifier, is similarly arranged and the output from the i.f. pre-amplifier is brought down in a coaxial cable built into the multi-core cable, which connects with the receiver control, discriminator and power supply units. The intermediate frequency used in the ML6A and ML6B equipments is 115Mc/s, in the ML6C is 120Mc/s and in the S.T. & C. equipment it may be 30 or 60Mc/s.

The performance of these equipments complies with the BBC's minimum specification and, in particular, with respect to:—

(a) H.F. response which, with reference to 10kc/s, falls within ±0·2dB up to 1Mc/s and ±0·5dB between 1Mc/s and 3·5Mc/s.

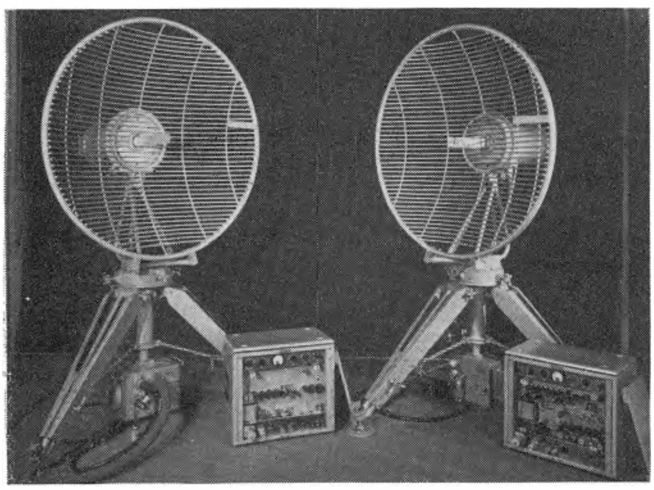


Fig. 1. E.M.I. ML6C equipment

(b) Signal-to-noise ratio which, expressed as

$$20 \log \frac{\text{D.A.P. Signal}}{\text{R.M.S. Noise}}$$

is not less than 58dB over a free-space transmission path of 40 miles.

(c) Group delays and differential phase characteristics which allow the satisfactory transmission of N.T.S.C. colour waveform (as modified to the British standard) even when as many as four links are connected in tandem.

A third type of s.h.f. equipment may be mentioned which, although not used in large quantities, nevertheless performs the important function along with two u.h.f. channels referred to later, of conveying Eurovision programmes in either direction across the English Channel. Eventually this service will be the responsibility of the Post Office (with different equipment and location) but since 1954 the BBC and R.T.F. (Radiodiffusion-Télévision Française) have been jointly responsible for the cross-Channel circuit which operates between Swingate (near Dover) and Cassel (Nord)—a distance of 56 miles.

This equipment, operating on a frequency of 3 750Mc/s, was installed by the French firm of Télécommunications Radioélectriques et Téléphoniques on behalf of Radiodiffusion-Télévision Française. A 3-metre paraboloid and launching unit, mounted at the 50ft level, is connected to the ground equipment via a length of waveguide. The transmitter uses a type CZ klystron with an output power of 10W. The receiver, local oscillator, mixer and i.f. pre-amplifier operate in a similar fashion to that on the ML6C and S.T. & C. television o.b. equipment.

U.H.F.

This is a comparatively new field but one of great value for several purposes. Two main frequency bands are being used for television links, viz.: 522Mc/s to 660 Mc/s (Band V) and 1 700Mc/s to 2 300Mc/s. For the former band, an f.m. link has recently been specially designed by the BBC[1]. This has superseded the use of v.h.f. which is largely having to be surrendered to broadcasting purposes. The advantages of lower frequencies such as these, as compared with s.h.f., are:—

Lighter and simpler aerials, with greater beam-width.

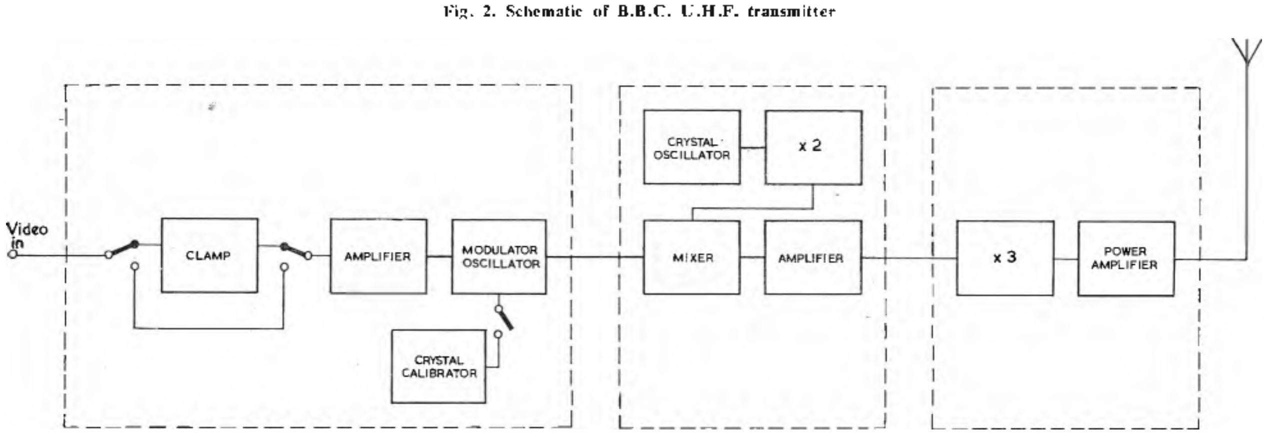
Less serious fading on difficult paths.

This link is used mainly for portable outside broadcast purposes but also for certain fixed links such as Isle of Man—Belfast as well as forming another channel between Swingate and Cassel (Eurovision), in which latter case special aerial systems are employed.

A schematic of the transmitter, which employs frequency modulation, is shown in Fig. 2. This consists of three units: the modulator, frequency changer and u.h.f. stages.

The video input signal of about 1V d.a.p. can be connected either directly to an amplifier which precedes the modulator or through a black level clamp. The output from the amplifier is d.c.-restored in order to restrict the overall peak-to-peak voltage change of the video signal with picture content, and is used to control the reactance value of a frequency modulated phase-shift oscillator. The frequency modulated output signal with a centre frequency of about 60Mc/s has a deviation

Fig. 2. Schematic of B.B.C. U.H.F. transmitter



of 2Mc/s between the bottom of the synchronizing pulses and peak white at a level of 1mW.

The second unit incorporates the drive which is derived from an oscillator employing an overtone crystal operating at about 75Mc/s. After doubling this frequency, the output is fed into a valve mixer stage at a level of about 50mW together with the f.m. signal from the modulator after a single stage of amplification. The upper sideband of the mixer output is selected and the level raised to about 4W in an amplifier having a sensibly flat response over a 6Mc/s band in the range of 200 to 230Mc/s depending upon the channel in use. At this point the deviation is still 2Mc/s.

The third unit comprises two u.h.f. stages, which use planar triodes of the inverted lighthouse type in grounded grid circuits. The valves operate in coaxial cavities which are silver-plated in order to obtain a high Q. They are cooled by an air blast. The first stage multiplies the frequency by a factor of three with a power gain of unity, to drive the final which operates as a class-C amplifier and gives an output of about 17W into the feeder and aerial system. After tripling, the deviation is 6Mc/s in the band 620 to 680Mc/s with an overall 3dB bandwidth of about 12Mc/s.

A monitor, which will operate at either v.h.f. or u.h.f., gives a video signal from the modulator or transmitter output respectively, enabling the operator to check the signal.

A flexible feeder, which has an impedance of 50Ω and a loss of about 3dB/100ft, is used to connect the transmitter output to the aerial. Two types of aerial have been developed, a corner reflector, having a half-power beamwidth of 30° and a six-element Yagi aerial with about the same beamwidth. The gain of both is about 9dB relative to a dipole. Horizontal polarization is normally employed.

For permanent receiving installations a higher gain corner reflector has been developed. This has a measured gain of 12dB.

Fig. 3 shows a schematic of the receiver which also consists of three units: the head, i.f. amplifier and demodulator.

In order to keep the receiving aerial feeder as short as possible to reduce losses the head unit is built into a weatherproof box so that it can be mounted as close to the aerial as practical. This unit contains input band-pass filters, u.h.f. local oscillator, crystal mixer and pre-if. amplifier. Either a 30 or 60Mc/s i.f. can be used depending on local site conditions. The local oscillator uses a planar triode in a quarter-wavelength coaxial cavity and it has been designed to keep frequency drift within the limits needed by the main receiver. The receiver noise figure is 14dB.

A multi-core cable connects the head unit to the main i.f. amplifier. As well as the incoming frequency modulated i.f. signal, this cable carries power, signalling, metering and telephone circuits. The limiter-discriminator following the i.f. amplifier delivers a 1V peak-to-peak video signal into a 75Ω load. These two units are exactly the same as those employed in 4000Mc/s microwave o.b. link equipment currently in service and are commercially available.

The units which comprise the transmitter and receiver are mounted in portable cases. The weight of each individual unit has been kept to a minimum and electronic smoothing is used in the power supplies to reduce their bulk.

2000Mc/s U.H.F. Equipment

This forms another cross-channel link between Swingate and Cassel and uses G.E.C. u.h.f. main line television link equipment with the addition of a travelling-wave amplifier which raises the nominal power output from 2 to 20W. A transmitter and two receivers are installed at each terminal. Three separate paraboloid aerials are used: the first, 8ft in diameter, located at 50ft, the second, 12ft in diameter at approximately 90ft and the third, 12ft in diameter at 200ft.

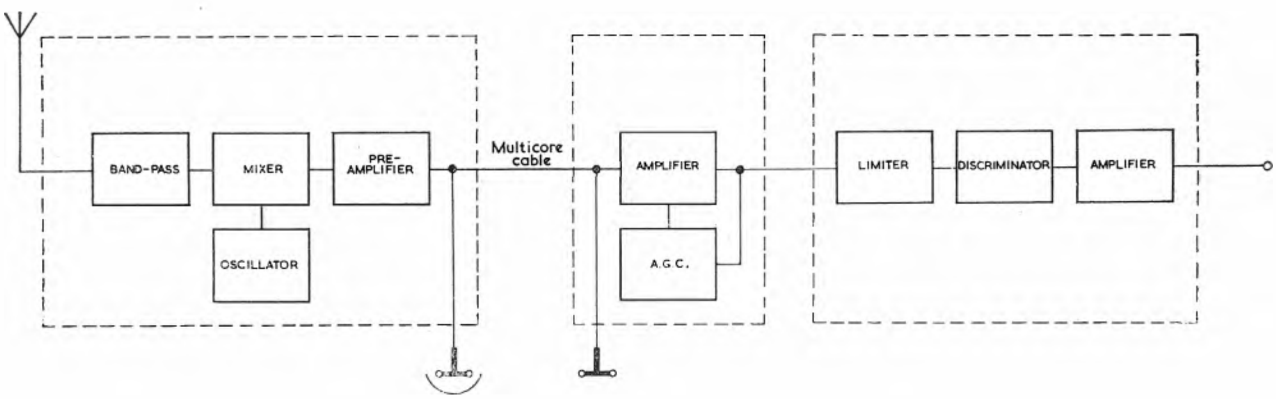


Fig. 3. Schematic of B.B.C. U.H.F. receiver

The aerials are illuminated by a short section of waveguide at the focal point and are connected to the ground equipment by means of helical membrane aluminium sheathed coaxial cables. It is possible to connect either transmitter or receiver to any of these aerials. Normally, when the transmission circuit is incoming from Cassel both receivers are connected to the first and second aerials; the distance between these representing the correct diversity spacing for an improvement in the signal-to-noise during fading conditions. For the outgoing circuit to Cassel, the transmitter is usually connected to the middle aerial. The upper aerial is mainly used in the summer time and during long spells of hot weather, when conditions of super-refraction reduce the usefulness of the lower systems. This link is mainly used to transmit from the British standard 405 lines to the Continental 625 lines, a conversion which is now carried out at Swingate; this 625 line picture is distributed to most Western European television networks except France. A return circuit using similar equipment is provided from Cassel so that simultaneous transmission may take place if required.

Communications

These fall broadly into long-range and short-range.

Long-range

Such links are mainly used for ensuring parallel communication between the two ends of a cross-country vision link. Principal equipments in use are:—Mullard, type GFE 506 and G.E.C. type BRT 143/163.

Both these equipments use f.m. and have an output power of about 12W. Simple unipole aerials are generally used.

The mode of operation is normally two-way simplex. There is a shortage of frequencies for these links, which are distributed over Band I and Band II. Both Bands of course are also used for vision or sound broadcasting so frequencies have to be carefully chosen according to locality.

Because of the congestion in the v.h.f. range, the BBC is experimenting with communication links operating at u.h.f. within the band 660 to 665Mc/s. These will require optical transmission paths and far more care with aerial installations, but in the long run may well prove the best solution to the problem. It is also intended that these equipments should be used for programme links (sound) when required. These links are being built by Mullard to a strict specification which, among other things, calls for the r.m.s. signal/r.m.s. noise ratio to be 50dB when measured over a thirty-mile path with propagation conditions corresponding to 'free space'.

Short-range (vision or sound)

A variety of miniature equipments are used as the control link associated with radio microphones, radio cameras, etc. For the most part, such links are either BBC 'home-made' or are made up from an assortment of commercial units. F.M. is usually employed and, as for the long-range links, the frequencies will fall in either Band I or Band II. Transmitter output power would usually be only a few watts or less.

Sound Programme

The need for links to carry the sound component of a television programme arises much less frequently than in the case of the vision, but transmissions from cars, ships, aeroplanes, etc., call for high-quality radio circuits for which special equipment must be held. The problem is very similar to that of the communication links: the same wavebands are used and largely the same type of equipment, though modified to ensure a higher quality performance.

Special and Auxiliary Equipment

Apart from the essential transmitters and receivers, a good deal of additional equipment is necessary to ensure that portable radio links can be got working in a speedy and efficient manner. Among the more important items are the following.

Mast-head Receivers

It is frequently convenient to exploit the great height of the masts of many of the television broadcast transmitters as reception points for television radio links, particularly at those stations provided with a permanent connexion back into the network. The advantage of, say, a 600ft aerial height for an s.h.f. link requiring an optical path will be easily understood. A probability contour showing the s.h.f. transmitter boundary for a single hop into the Sutton Coldfield s.h.f. receiver is shown in Fig. 4. It would, however, be next to impossible to hoist and install 4ft paraboloids and head units on every separate occasion and so it has been found worthwhile to install permanent s.h.f. mast-top receivers at some dozen of the BBC transmitting stations.

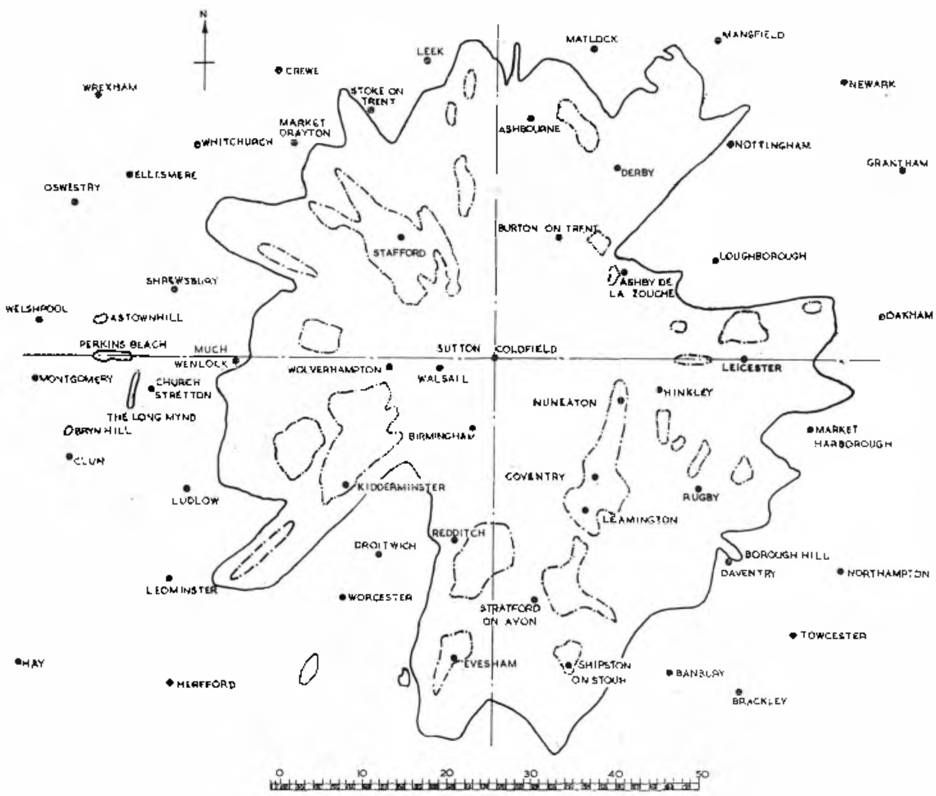


Fig. 4. Range contour. S.H.F. transmitter working into mast-head receiver at Sutton Coldfield

The problems of ensuring the reliability of such equipment all the year round have called for meticulous design and experiment, but the latest version produced to BBC specification is proving eminently satisfactory. It consists of: (1) two 4ft paraboloids remotely controlled (2) a waveguide change-over switch enclosure, and (3) a receiver cubicle, all mounted at high level on the mast.

The paraboloids and slewing units are mounted on a platform which forms part of the mast structure. These platforms are fitted on opposite faces of a rectangular mast and adjacent corners on a triangular mast. Each paraboloid is mounted on a specially designed slewing unit and the whole assembly is produced to stand the hazards of weather and wind loading in exposed places. Radial heater elements are incorporated in each slewing unit assembly to prevent ice building up on the turntable. It is possible to traverse each aerial through 270° by remote control from the ground and accurately orientate this to within $\pm \frac{1}{2}°$ of any desired setting; the actual azimuthal position is registered by a magslip indicator on the remote control panel at ground level.

Each aerial is connected to the waveguide switch and receiver cubicle by a length of waveguide and a short flexible cable. This cable introduces a loss of 0·5dB at about 4 500Mc/s, but enables the paraboloid to be traversed without the complication of waveguide rotating joints.

The waveguide run from each paraboloid terminates on a two-pole waveguide changeover switch which is installed at a convenient place on the mast and the operation of which is remotely controlled. Both outlets from this switch are connected again through a very short length of guide to the s.h.f. receiver cubicle in which two complete local oscillator, mixer and i.f. pre-amplifiers are fitted. The tuning cavities of the local oscillators are motorized and a magslip transmission system is used to indicate the position of the tuning from the control point. The outputs from the pre-amplifiers (which are centred at 30 or 60Mc/s) are connected to the main i.f. amplifiers and discriminators situated at ground level, through a length of coaxial cable. The i.f. pass-band is 20Mc/s and as the equipments are operating in the strong field of the main television transmitter special precautions are necessary and much care has been taken in the design of filters in all the supply leads, screening, etc. The system, as designed, permits two complete and independent receiver chains to be operated simultaneously or either chain may be switched and held as reserve for the other.

A feature of this equipment is that each unit in the receiver cubicle may be removed and replaced by a spare, should a fault develop. Full metering and supervisory circuits are also included and care has been taken to select valves and components which should have an exceptionally long life.

Portable Masts and Towers

Where permanent masts or other convenient high points are not available—and this will be the case more often than not—recourse must be made to suitable means of achieving adequate aerial height at both the transmitting and receiving end of a link. High buildings, water towers, hill tops, etc., are often used for this purpose. Of course, they are not always conveniently to hand and even if they are a great deal of time and effort may be needed in both securing the necesary permissions and then raising and installing the equipment. It has therefore been found indispensable to develop portable telescopic masts and towers which can rapidly elevate an aerial or paraboloid to clear local houses, trees or other obstructions. The BBC uses two main types of such apparatus, namely:—

(*a*) an extensible mast, built like a fire-escape ladder;

(*b*) a telescopic tower.

The former, manufactured by Merryweather Ltd is mounted on a vehicle which also embodies a cabin that houses the radio equipment. The mast can be raised by engine power in a few minutes to a height of over 100ft. Guys are not essential in fair weather and this is an important point since the mast may often be installed on hard ground into which stakes could not be driven. This type of mast is, however, liable to twist and sway very slightly in the wind and this feature renders it unsuitable for s.h.f. paraboloids whose beam width is only a few degrees. It is therefore used for v.h.f. and u.h.f. aerials and for such applications is extremely effective.

Radio Link Vehicle

Since most of the radio links used in the BBC are temporary and being continually moved from place to place, mobility and efficiency are of the greatest importance and the mobile masts referred to above have, of course, been designed to such ends. Another contribution has been the careful development of a special radio link vehicle (see illustration on page 291). This vehicle is in effect a mobile cabin which both transports the equipment and then provides an 'operations centre' on site. Apart from aerials, paraboloids, etc., the link equipment can remain inside the vehicle and so the amount of local installation work is reduced to an absolute minimum. For those occasions where, for one reason or another, it may be impracticable to operate from the vehicle itself—for example, when the equipment is required on board ship—all units can readily be detached, carried from the vehicle and reinstalled elsewhere. All the technical equipment is accommodated in transit cases where possible and located on resiliently mounted shock-proof trays. The ancillary equipment associated with the radio terminal repeater station, such as paraboloids, panning heads, tripods, cables, cable drums, rigging tackle, etc., are carried in compartments or lockers which give access to these items from within or outside the vehicle as circumstances require. The roof platform will support a load consisting of two 4ft paraboloids, tripods and panning heads; special clamps are fitted in the appropriate positions to fasten the tripods to the roof. Stabilizing jacks are provided which allow the chassis of the vehicle to be supported firmly above the springs.

The essential transmitter and receiver equipment, together with test signal generator, picture monitor, distribution amplifier, jackfields, radio/telephone communication equipment and so forth, are arranged so that these are within easy reach and view of the operator.

The same vehicle can be used for a transmitting point, receiving point, or as an intermediate repeater. As the vehicles may have to spend long periods in exposed places under adverse weather conditions, the needs and comforts of the operating staff have been a fundamental consideration in planning the layout and facilities. Adequate heating and ventilation systems are installed and a water supply for domestic purposes is included.

Towing attachments at the rear of the chassis enable a trailer-mounted diesel generator to be coupled up if required.

Mobile Power Units

Radio links are often set up in remote places where mains power is unavailable, and so the BBC has equipped itself with a number of mobile power units. These are diesel-alternators, manufactured by Auto-Diesels Ltd, and mounted on a trailer for towing behind the radio link vehicle.

Transmission Monitor

The need frequently arises, especially under fault conditions, to check that the modulated s.h.f. picture signal is being transmitted satisfactorily. Occasions arise when some doubt exists as to whether the fault lies in the transmitter equipment or the receiver equipment, and therefore a transmission monitor has been specially developed. It receives a small portion of the microwave signal, demodulates it and supplies a video output which can be adjusted to the standard 1V peak-to-peak level. The unit consists of a small tunable cavity which is clamped to the face of the paraboloid: a silicon crystal is coupled to the cavity to demodulate the signal which is passed to the control unit at ground level by means of a length of coaxial cable. The frequency of the cavity is de-tuned to a value which is 0·707 of the transmitted carrier frequency so that it acts as a simple frequency modulated discriminator.

The unit is enclosed in a weatherproofed cast box; for the newer portable link equipment provision is made on the paraboloid mounting to locate this head unit in the optimum position.

S.H.F. Path Simulators

Because of the high aerial gain in an s.h.f. system, the testing of the overall transmission characteristic of the link presents considerable difficulty. If tests between transmitter and receiver were carried out within a building, reflections from adjacent walls and objects would produce most erroneous and peculiar results. In order to overcome these difficulties, a device which is now called a 'path simulator' was developed and this permits a complete link to be tested at base under conditions simulating path lengths of up to forty miles. This consists of waveguide matching stubs to which the head units are attached and correctly terminated, together with waveguide-to-coaxial transformers which enable a finite length of coaxial cable to connect the output from the transmitter head unit through a suitable variable attenuator system to the receiver head unit. The loss introduced into the system by means of the path simulator has been calibrated to provide identical conditions to those when using the link with 4ft paraboloids between fixed points varying from five to sixty miles in length. Full-scale tests can thus be made and overall checks of the equipment carried out without the need to check the equipment over an actual path.

Coverage and Exploitation

Radio links play an important and essential part in the organization for bringing television and sound from points remote from the permanent contribution network. If necessary, by using a sufficient quantity of links in tandem, pictures could be collected from almost any point in the British Isles. The use of two or three links in tandem is quite common practice. Four and five together are by no means unknown and such a number would suffice to reach most parts.

In 1957 nearly 1 000 vision links were established for actual broadcasts, either singly or as part of longer chains, while additional equipment now coming into service will probably increase the number in the present year. This does not take into account the almost daily use of certain point-to-point links being provisionally operated by the BBC pending replacement by a Post Office service; examples being the Anglo-French link, Isle of Man-Belfast, Glasgow-Kirk o'Shotts.

Special Problems

The ordinary run of cross-country radio link presents chiefly the problem of achieving a line-of-sight path; provided this is secured, together with adequate Fresnel zone clearance, satisfactory circuits can usually be established, up to some fifty miles per 'hop'. More difficult conditions do, however, present themselves in cases of transmission across water and transmission from moving sources (e.g., ships, cars, aeroplanes, etc.).

Transmission Across Water

For reasons that are well known, transmission paths over water on u.h.f. and s.h.f. links are liable to fade in certain circumstances. Such paths are often encountered as a result of the many estuaries and islands around the British Isles, and so when links have to be established over more than a mile or two of water it is usually desirable to take special precautions to alleviate fading. The usual remedies, adopted either singly or in combination, are:—

(*a*) To use as low a frequency as possible. For example, a u.h.f. link on 659Mc/s would be safer than an s.h.f. link on 4 500Mc/s.

(*b*) To use space-diversity reception by installing two receivers at different levels.

(*c*) To ensure a signal strength high enough to allow a reasonable fading margin.

(*d*) To use frequency diversity by setting up duplicate links widely spaced in frequency.

(*e*) To take special care in the siting of the transmitting and receiving aerials in relation to local topography, and particularly to avoid reflection from the water by arranging for the reflection point to fall on land. (This technique has been used with success on the cross-Channel Eurovision circuit and between Snaefell (Isle Man) and Divis (Northern Ireland).

Transmission from Moving Sources

The basic problem in the case of moving sources is that of keeping the transmitting and receiving aerials in line with one another. In extreme cases the use of omni-directional aerials can provide the answer, but usually there is insufficient power to permit this, or to permit it at both ends of a link, and so it becomes necessary to arrange that, by one means or another, the two aerials are continuously adjusted to the correct bearing. Even when automatic devices are incorporated such adjustments are seldom precise enough to fall within the few degrees tolerance of an s.h.f. link: one therefore welcomes the wider beamwidth—25°—of the 659Mc/s u.h.f. link for cases of this kind.

Other problems may be associated with moving sources, depending on their nature. Submarines, aeroplanes and helicopters present special difficulties. The 'Roving Eye' (in essence a large car carrying a camera, microphone and transmitters) may want to transmit pictures while driving round built-up areas and so brings one up against non-optical or quasi-optical paths; there is no complete answer to this, but the use of v.h.f. (200Mc/s approximately) rather than u.h.f. does give a noticeable measure of diffraction round minor obstacles that would otherwise form complete obstructions.

Future Developments

The BBC now has a considerable investment in portable radio links and has equipped itself with a wide range of vision and sound apparatus to cater for most of the foreseeable needs of television during, say, the next five years. What the more distant future holds in store is, of course, uncertain, but it is likely that frequency allocations will come under review from time to time in relation to the needs of other users. The general trend is always towards higher frequencies, and this may bring certain new problems, though it is hoped gradually enough to permit the necessary progressive adjustment of technique.

Whether tropospheric-scatter propagation and associated equipment will develop sufficiently to enter the field of portable or temporary television links seems very dubious at the present time. The potentialities, however, are most attractive if thereby one could span, say, 300 miles in one 'hop'. It is possible to map a route between the United Kingdom and North America—via Faroes, Iceland and Greenland—over which none of the sea paths need exceed 300 miles. Thus the range of television coverage might be enormously enlarged and the opposite shores of the Atlantic come within television 'range'.

Acknowledgment

The author thanks the Chief Engineer of the British Broadcasting Corporation for permission to publish this article and also acknowledges considerable help received from colleagues in various departments.

REFERENCE

1. A U.H.F. Television Link for Outside Broadcasts. *B.B.C. Engineering Monograph No. 19* (To be published) (1958).

Broadband Microwave Systems Employing U.H.F. Triodes

By G. W. S. Griffith*, A.M.I.E.E. and B. Wilson*, M.B.E.

The arguments developed in this article indicate that it is more economical and more convenient to arrange for the modulation, demodulation, and as much of the amplification as possible in a microwave system to be carried out at an intermediate frequency.

Further, in the 1 700 to 2 300Mc/s band, microwave triodes are the most satisfactory means of providing amplification, and have been used in conjunction with coaxial transmission lines to produce a very compact and simple equipment using low voltages.

The system described is based on these techniques, and satisfies the performance recommendations of C.C.I.R. for the long haul transmission of 240 telephony channels, or television signals, with an adequate margin for contingencies.

ALTHOUGH investigations into the use of microwave systems for communication services were started in the late 1930's, it was not until after the war that they were extensively used for public services. Many companies on both sides of the Atlantic were developing this type of equipment with the result that different techniques and operating frequencies were employed.

In 1947 an experimental link was set up between New York and Boston operating at 4 000Mc/s. Early American experiments made use of velocity modulated valves, but these were later abandoned in favour of specially designed triodes[1].

The first microwave system in the United Kingdom went into service in December, 1949 to carry 405 line television signals at 900Mc/s between London and Birmingham[2]. This used a range of triodes developed during the war.

A further system for operation at 4 000Mc/s was developed using travelling-wave tubes, the first of these being installed in March, 1952 to carry television signals between Manchester and Edinburgh[3]. Since then many systems have been installed using the various methods of operation, and it is of interest to note that in 1954 when several of the European countries combined in 'Eurovision' for the exchange of television programmes, triode systems on 2 000Mc/s were used in Switzerland, Germany and the South of England, while travelling wave tubes on 4 000Mc/s were used in France and the North of England. Directly modulated klystrons on 4 000Mc/s and 6 000Mc/s were used on some of the interconnecting links.

Types of System

Basically there are two types of broad band microwave system in current use. In the first, all the amplification necessary is carried out in the u.h.f. (or s.h.f.) band. A suitable amplifier for an all radio frequency repeater has been described elsewhere[4]; it employs disk-seal triodes capable of providing a stage gain of 8 to 10dB. It should be mentioned that amplification is also possible by means of travelling-wave tubes, the higher stage gain available enabling a smaller number of valves to provide the necessary overall gain. This is offset by the high voltages required and by having to provide the magnetic field systems necessary for focusing.

However, in spite of the apparent simplicity of this arrangement, most of the microwave equipments so far installed have employed the alternative system in which a large part of the necessary amplification is effected at an intermediate frequency. There are a number of reasons for adopting this procedure:

(1) Gains exceeding 10dB per stage can be readily achieved in amplifiers having bandwidths greater than 30Mc/s to the 3dB points by using modern high figure of merit valves.

(2) The linearity of the phase/frequency characteristic (group delay) can be accurately controlled and can have good symmetry. With the low centre frequency/bandwidth ratio, which is inherent in i.f. systems, temperature, humidity, and ageing effects are minimized, and accurate equalization of i.f. group delay characteristics can be maintained for long periods.

(3) Suitable valves, and their associated circuit elements for use in i.f. amplifiers are comparatively cheap. For example, in an i.f. amplifier providing 80dB gain, the cost of valves required for the necessary six or eight stages would probably be less than that of each of eight u.h.f. triodes, and very much less than each of three travelling-wave tubes that would be necessary to provide the same gain.

(4) Automatic gain control and limiting are comparatively easy to apply to compensate for the large changes in signal level that are involved even with line of sight paths.

In both types of system just described it is necessary to provide a shift of frequency to change the frequency band accepted by the receiver to a different band for re-transmission to prevent direct feedback.

In addition to the amplification question discussed above, it is necessary to consider whether the signal should be modulated and demodulated at the ultra-high or intermediate frequencies. Direct modulation of klystron oscillators has been used, but suffers from the disadvantage that the modulator linearity is dependent on klystron tuning.

Demodulators are usually operated at intermediate frequency because of the difficulty of providing effective limiting of u.h.f. signals over the whole base-band or video range. A further advantage of using an i.f. demodulator is the improvement in the stability of the discriminator characteristic due to the low centre frequency/bandwidth ratio. In a practical system it is advantageous to operate both the modulator and demodulator at the intermediate frequency chosen for the system. As an example, we would mention a recent paper[5] describing a type of modulator (using grounded grid triode valves), which can be frequency modulated over a wide frequency range with good linearity. A number of

* *The General Electric Co. Ltd.*

variants of this general type are possible and have given good results.

System Parameters

Before considering a typical system in detail, it is useful to consider some of the special test methods used to evaluate the performance of a broad-band frequency-modulated radio transmission system. The important parameters are:

(a) A level amplitude/frequency response over the required range with a gradual fall outside this range (in the interests of transient response).

(b) Very good linearity of the input/output signal relationship at all base-band or video frequencies in the required range.

These parameters are influenced by the following factors:—

(1) Modulator linearity—i.e. the maintenance of a linear relationship between the instantaneous amplitude of the incoming signal, and the instantaneous frequency of the modulated output signal.

(2) Amplitude frequency characteristic—This should be level for all parts of the system handling a frequency modulated signal.

(3) Linearity of the phase/frequency characteristic of all parts of the system handling a frequency modulated signal, including the modulator and demodulator.

(4) Echoes—such as may be produced by mismatches in long feeders must be reduced to suitably small proportions.

(5) Reduction of the level of interfering signals to some 65 to 75dB below the wanted signal (these interfering signals may arise from harmonics of shift frequencies in a single channel, or from other channels where more than one r.f. channel is used).

(6) Linearity of the instantaneous frequency/instantaneous amplitude characteristic of the demodulator.

(7) The overall amplitude/frequency response must be level and input amplitude/output amplitude characteristics must be linear in any amplifiers or other devices used in the base-band or video section of the equipment.

(8) The time delay/base-band (or video frequency) characteristic of the equipment must be substantially constant over the frequency band.

Special test equipment has been designed for the development and maintenance of u.h.f. systems. The adoption of an i.f. system simplifies some of this test equipment since all modulators, demodulators, limiters, and i.f. amplifiers operate on the same frequency. Group delay measurements can also be made from an i.f. point of a transmitter to an i.f. point on a following receiver, so that the group delay of the r.f. part of the equipment can be evaluated.

Triode Microwave Amplifiers

So far three methods of providing microwave amplification have been mentioned—klystron, triode, and travelling wave tube. The over-riding advantage of the triode over the klystron is that it is possible to make triode amplifiers that are capable of maintaining the required amplitude and phase characteristic over wide frequency bands. Although the travelling-wave tube is available and capable of dealing with much wider frequency bands, there are disadvantages associated with too great a bandwidth.

The frequency spacing recommended by the C.C.I.R. indicates the desirability of limiting the bandwidth. Adequate

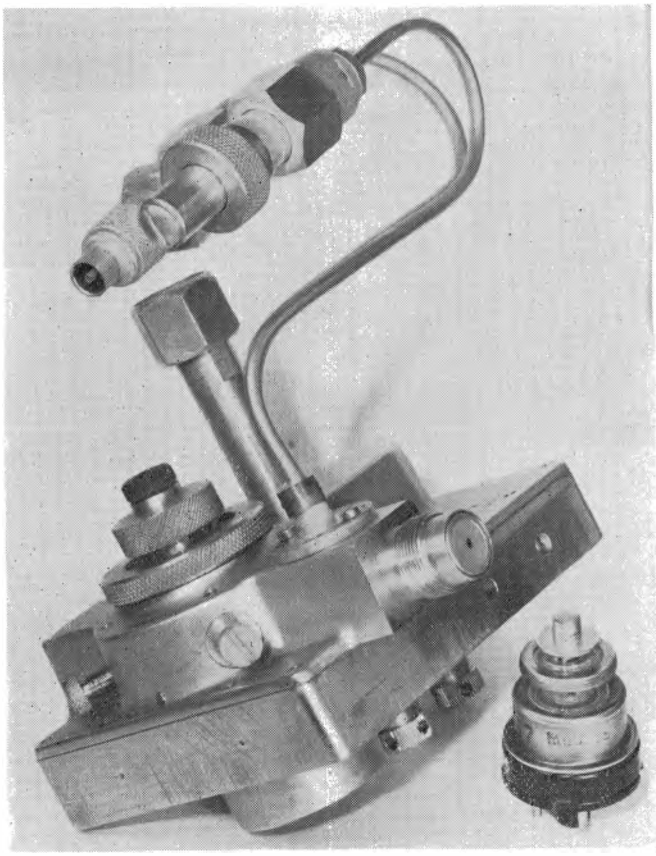
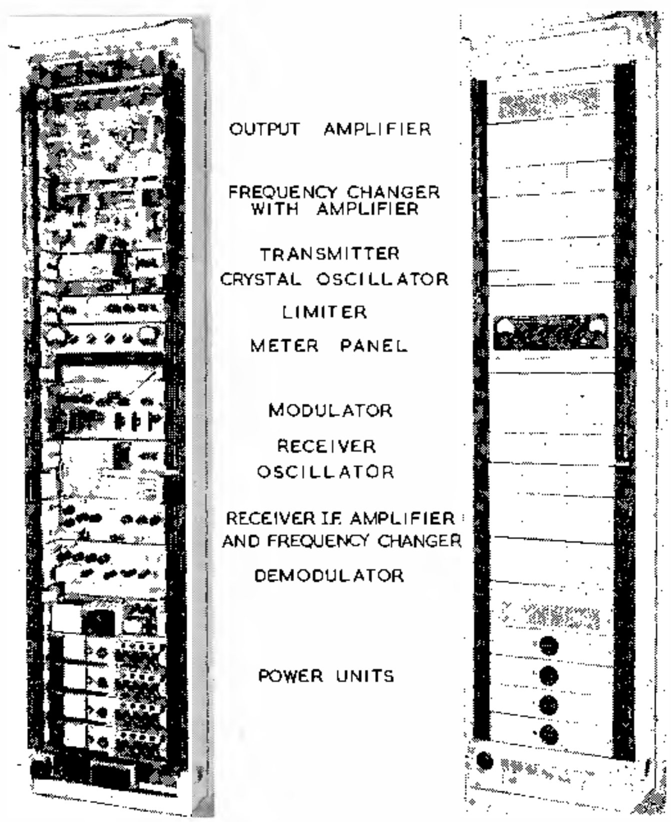


Fig. 1. U.H.F. circuit

Fig. 2. Transmitter-receiver rack without and with front covers



bandwidth restriction is provided by a triode amplifier. Also, in i.f. systems and in shift frequency mixers at repeaters, the r.f. carrier is modulated by a comparatively low frequency (70Mc/s to 250Mc/s), and the appropriate sideband is selected for transmission. The filtering necessary to attenuate the carrier frequency and unwanted modulation products is simpler if triode amplifiers are used because of the inherent selectivity of this amplifier circuit.

A further advantage is that using modern triodes an h.t. supply of 200V with no very stringent requirements on stability or ripple is adequate. This means the power supply units can be cheap and simple without sacrificing reliability. Heater requirements are also easy to meet, but longest life will be obtained if the heaters are operated within 2 per cent of the nominal operating voltage.

The triode has a special advantage in a transmitter mixer stage. This stage provides either the frequency shift at an all r.f. repeater or the translation from i.f. to u.h.f. (or s.h.f.) where i.f. circuits are used. The mixers can use either silicon crystals, triodes, or travelling wave tubes. The silicon crystal with a high level drive suffers from the disadvantage of harmonic generation, and these harmonics of the intermediate frequency can appear at the u.h.f. output terminals at objectionable levels. Triodes and travelling-wave tubes are appreciably better in this respect, the triodes having the advantage that the i.f. power required for efficient operation is considerably less than for travelling-wave tubes.

Typical Triode System

The following microwave system was designed by G.E.C. to transmit at least 240 speech channels, or television signals, in accordance with the recommendations of C.C.I.R. It operates in the frequency band 1 700 to 2 300Mc/s over line-of-site paths.

The linearity of both amplitude and phase characteristics is adequate for good quality transmission of either multi-channel telephony or compatible colour television signals. The bandwidth required for the television signals is of course greater than for the 240 telephone channels.

Because of the advantages enumerated in the first part of this article, most of the amplification is carried out at an intermediate frequency of 70Mc/s. The modulators and demodulators also operate at this

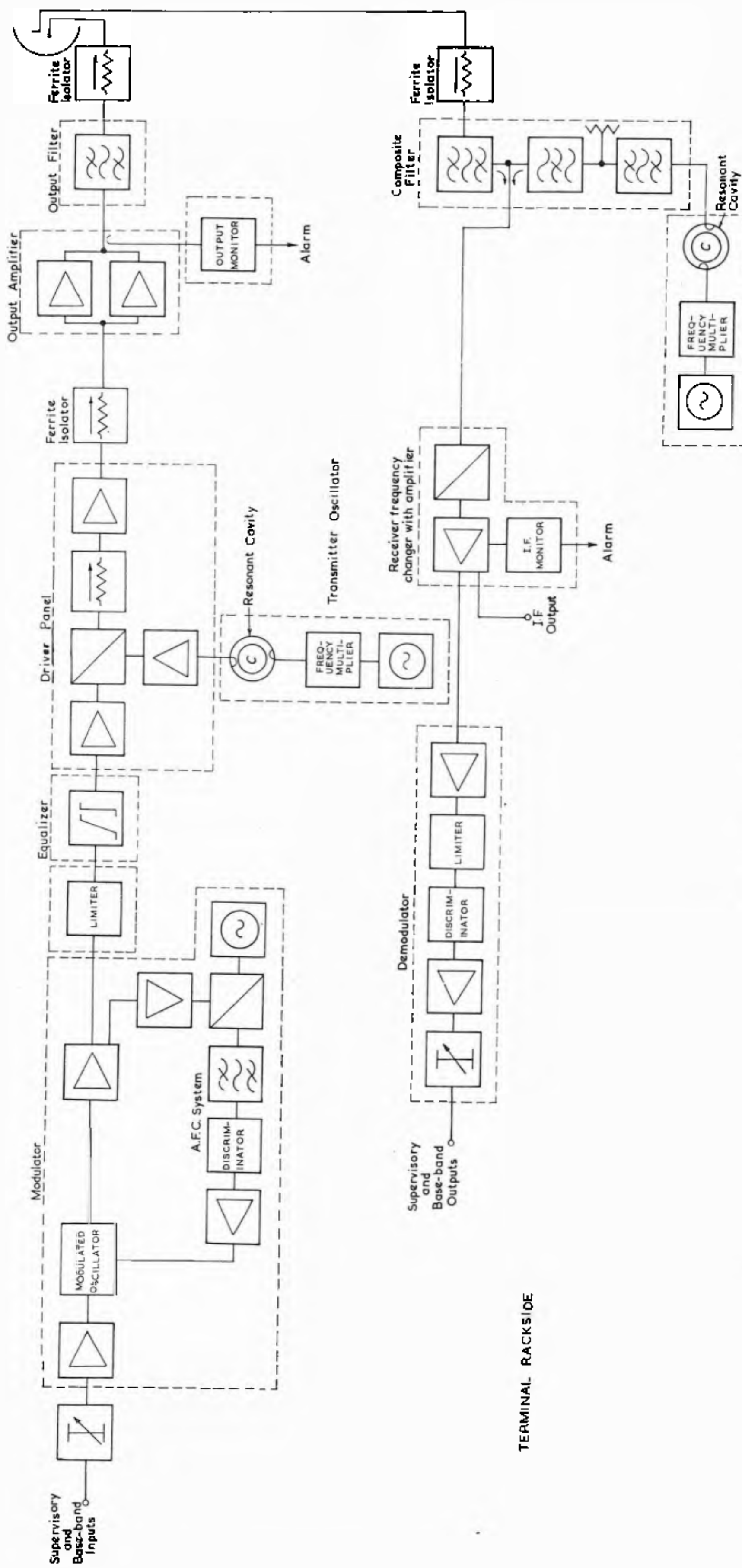


Fig. 3(a). The terminal equipment

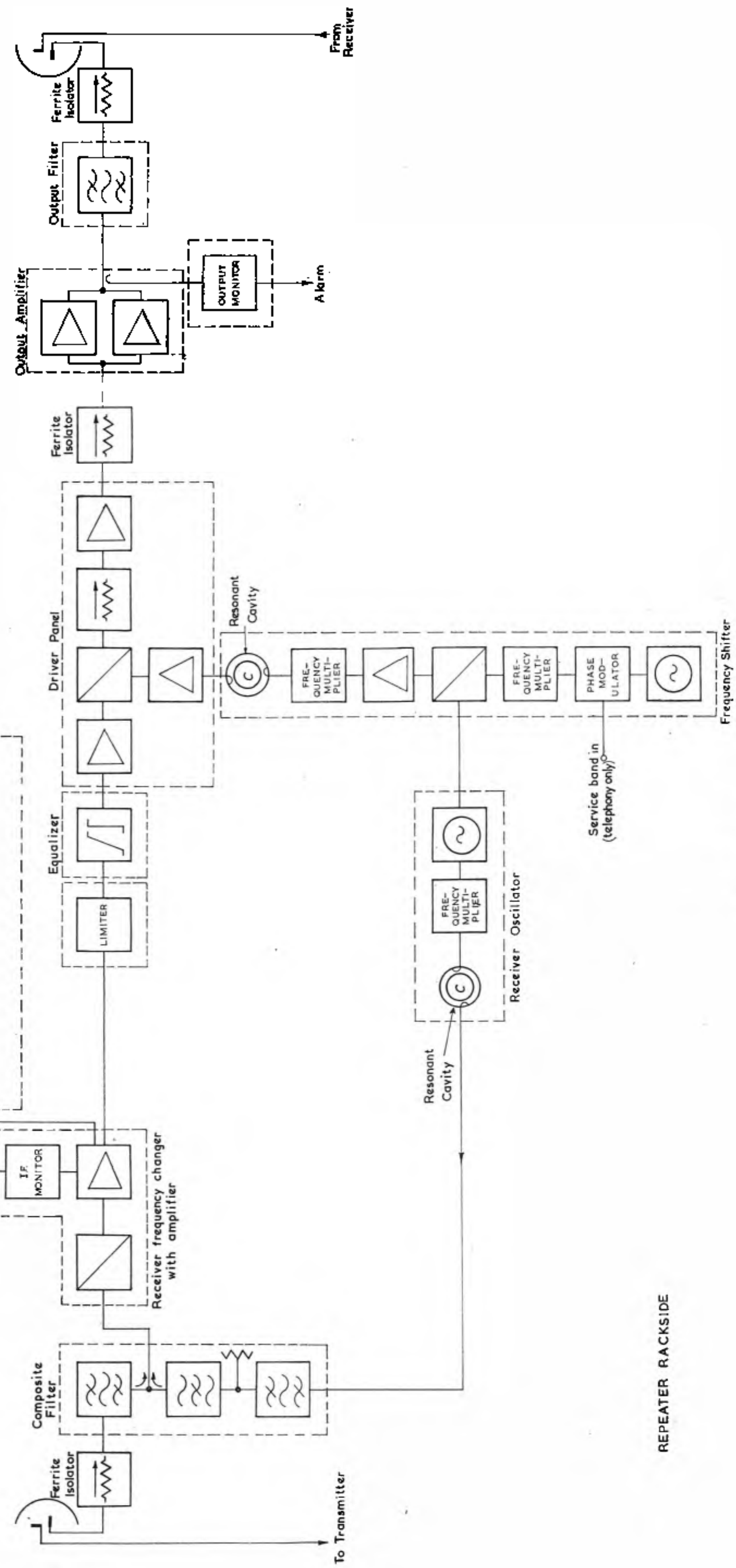


Fig. 3(b). The repeater equipment

frequency. Triodes are used in the transmitter frequency changer and u.h.f. amplifiers. The valves are operated at conservative ratings and have proved to be very reliable and give satisfactory service.

A thin waveguide input circuit is used so that the valve, which is situated rather less than $\lambda/4$ from the short-circuited end of the guide because of the valve input capacitance, effectively terminates the line. The signal is coupled into the waveguide from a coaxial input connexion. Tuning screws are provided to arrange for the waveguide to present a good termination to the coaxial input.

The grid-anode circuit consists of a coaxial cavity, which is capacitively loaded by the grid-anode capacitance of the valve, and which operates in the $\lambda/4$ mode of resonance. The cavity is approximately tuned to frequency by a preselected loading block; final tuning to cover valve replacement variations is performed by a capacitance tuning screw.

The output transmission line is capacitively coupled to the cavity. The anode cavity is designed so that the inner of the circuit makes sufficiently good contact to the valve anode, and is short enough to provide adequate thermal conduction to the outside of the circuit.

The u.h.f. circuits shown in Fig. 1 are manufactured from special aluminium forgings, which provide the internal contours of the cathode circuit as well as the external circuit shape without any further machining. The circuits have been tested with the ambient temperature varied over a range of 58°C. The change in anode circuit resonance over this temperature range was only 0·025 per cent (500kc/s). This variation can be accepted by the equipment without any measurable degradation of the system performance.

Fig. 2 illustrates a rack on which these circuits are mounted. The rack, which is 7ft 6in high, 20in wide and only 8in deep, houses a transmitter and a receiver, their associated power equipment, functional monitoring and built-in metering facilities.

This rack can be used either as a terminal transmitter and receiver providing for simultaneous transmission of signals in both directions or as a repeater. In the latter case two such racks are required to provide a repeater station for simultaneous transmission of signals in both directions.

Figs. 3 (a and b) show in diagrammatic form the arrangement of the terminal and repeater equipment.

At a terminal transmitter the incoming base-band signal is routed to the modulator where it frequency modulates a 70Mc/s sub-carrier. An automatic-frequency control system stabilizes the mean frequency of the modulator. The 70Mc/s signal is then passed through a limiter, which reduces any residual amplitude modulation to negligible proportions, an i.f. phase equalizer and a single stage amplifier, to the transmitter frequency changer. The u.h.f. drive to the frequency changer is produced by a crystal controlled oscillator and multiplying chain employing only three small receiving type valves. This gives an output of a few milliwatts at the required frequency which is selected by a tuneable cavity. As the output from the oscillator panel is too low for high level mixing, a stage of u.h.f. amplification is provided before the frequency changing stage. Local oscillator and i.f. signals are applied simultaneously to the cathode of the mixer which is a triode valve operating on the non-linear part of its characteristic.

The design of the frequency changer u.h.f. circuit is mechanically identical with that of the u.h.f. amplifiers, the anode circuit being tuned to the required sideband. The amplifier following the mixer stage is preceded by a ferrite isolator to reduce interaction between the circuits. A filter network in the anode circuit of this amplifier attenuates the residual carrier and unwanted sideband from the mixer to a low level before being passed via a second ferrite isolator to the output amplifier, which consists of two valves in parallel combined in such a way that they can be tuned independently. Directional couplers are incorporated in the output of each stage so that a simple scanned oscillator and display unit (shown in Fig. 4) can be connected to the transmitter and the response at each u.h.f. stage displayed.

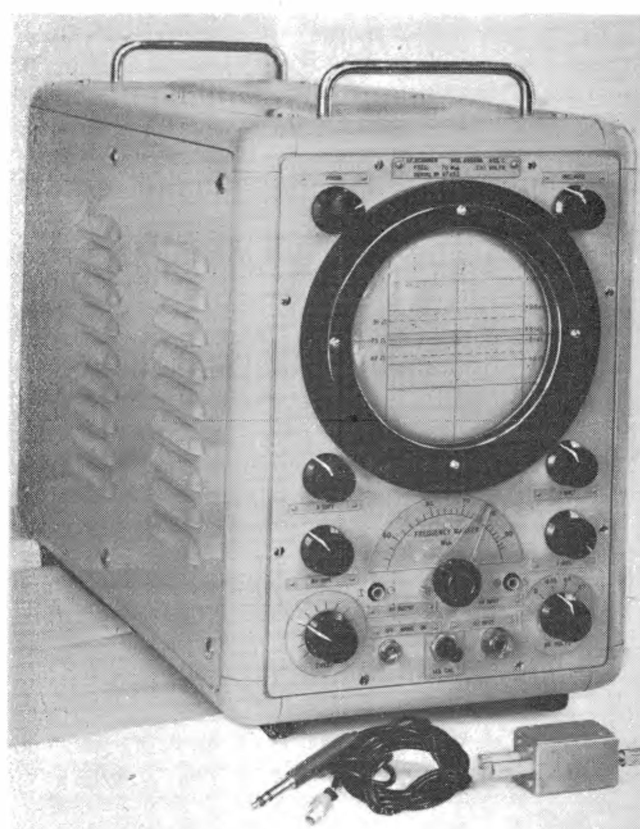
The directional coupler following the output stage is connected to a moving-coil relay which continuously indicates the output power, and provides an alarm signal if the transmitter output power falls below a pre-determined level.

Ferrite isolators are used at the equipment end of the aerial feeders to provide an adequate termination for echoes which may be reflected from the antenna or from feeder irregularities. Semi air-spaced feeders of the helical membrane type are used, since at these frequencies the loss is reasonable, and reflections due to irregularities are at an acceptable level. These feeders have the advantages of lower cost and easier handling than waveguide runs.

The antenna comprises a paraboloidal reflector fed by a waveguide type of launching unit and provides a gain of approximately 34dB at 2 000Mc/s. The launching unit is of the dual polarization type, i.e. both horizontally and vertically polarized waves can be launched or received simultaneously. Multiplexing arrangements can be provided to enable up to three transmitters or receivers to be operated on each polarization of the antenna: i.e. three transmitters and three receivers can be operated on a single antenna.

On the receiver side of the equipment the incoming signal is fed via a branching filter to the receiver mixer. The branching filter is of coaxial construction. The input signal line includes a band-pass filter at signal frequency, while the local oscillator line includes a band-stop filter at signal

Fig. 4. Scanned i.f. oscillator and display unit



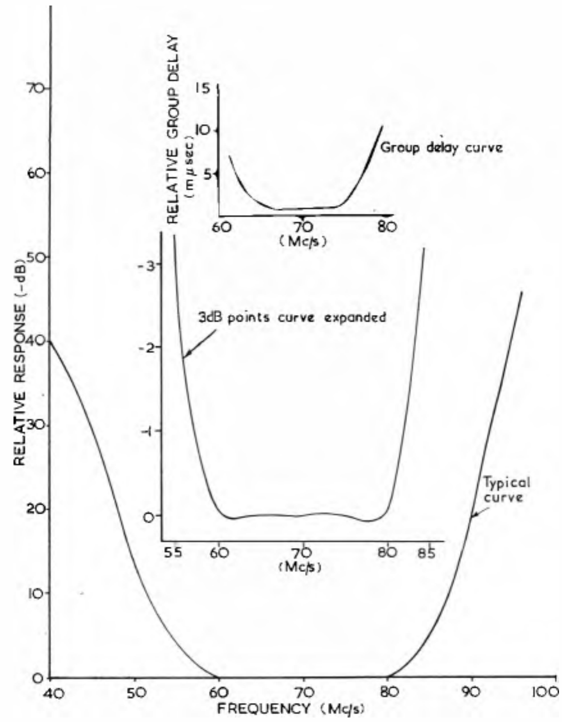


Fig. 5. Amplitude and group delay response of i.f. amplifiers

frequency, a band-pass filter at the local oscillator frequency and a load for the absorption of image power generated in the crystal mixer. The arrangement is shown in Fig. 3. It will be seen that both signal and local oscillator power are fed to the same terminal of the mixer unit, which comprises a coaxial type silicon crystal unit and a matching transformer.

The receiver local oscillator is identical with the transmitter oscillator. The output level of 2 to 3mW is adequate for this purpose, and amplification is not necessary.

The output of the crystal is fed to a low noise i.f. input stage. The overall receiver noise factor is 10dB. Following this low noise input stage are five broad-band amplifying stages. A phase equalizer is fitted in the i.f. chain to compensate for the phase characteristics of the coupling transformers used. The overall characteristics of the i.f. amplifier are shown in Fig. 5.

The output of the i.f. amplifier is fed to the demodulator, while in a repeater station a second i.f. output is fed via a limiter to the transmitter mixer.

The demodulator consists of an amplitude limiter, discriminator, and base-band amplifier. Back-to-back silicon diodes are used in the limiter section, which attenuates any amplitude modulation on the incoming signal by 40 to 50dB before passing it to the discriminator. The discriminator makes use of the amplitude frequency response lying between the series and parallel resonance of a three element network. Values are chosen to provide good linearity and the balanced detector arrangement used has good amplitude rejecting properties at centre frequency. The base-band amplifier is of conventional design.

The reliability of the u.h.f. triodes has proved to be good. Lives varying between 4 000 and 10 000 hours are usual, and in a typical system the fault rate from all causes including normal valve replacement has averaged one per 1 200 hours per transmitter-receiver rack. In only one out of eleven faults was the service interrupted so that even when a system is run without standby the fault liability is low.

The power consumption of one transmitter-receiver rack is less than 0·5kVA.

Acknowledgments

The authors wish to express their thanks to the management of The General Electric Co. Ltd., Telephone Works, Coventry, for permission to publish this article.

REFERENCES

1. ROETKEN, A. A., SMITH, K. D., FRIIS, R. W. The TD2 Microwave Relay System. *Bell Syst. Tech. J. 30*, Pt. 2, 1046 (1951).
2. CLAYTON, R. J., ESPLEY, D. C., GRIFFITH, G. W. S., PINKHAM, J. M. C. The London-Birmingham Radio Relay Link. *Proc. Instn. Elect. Engrs. 98*, Pt. 1, 204 (1951).
3. DAWSON, G., HALL, L. L., HODGSON, K. G., MEERS, R. A., MERRIMAN, J. H. H. The Manchester-Kirk o'Shotts Television Radio-Relay System. *Proc. Instn. Elect Engrs. 101*, Pt. 1. 93 (1954).
4. BOWEN, A. E., MUMFORD, W. W. A New Microwave Triode: Its Performance as a Modulator and as an Amplifier. *Bell Syst. Tech. J. 29*, 531 (1950).
5. RAVENSCROFT, I. A., WHITE, R. W. A Frequency Modulator for Broad Band Radio Relay Systems. *P.O. Elect. Engrs. J. 48*, Pt. 2, 108 (1955).

Travelling-Wave Tubes in Communications

By R. B. Coulson*, B.Sc., B.E.

Microwave repeaters for telephone communication are the first equipments to use travelling-wave tubes in appreciable quantities. The application of t.w.t's for this purpose is discussed.

For commercial equipments, the author expresses the opinion that electromagnets wound with aluminium foil are at present preferable to permanent magnets.

AFTER more than a decade of intensive development, the basic design of helix type travelling-wave tubes is now well established. Tubes now available from many different makers vary only in minor details such as operating voltages and outline. Radar and communication systems incorporating travelling-wave tubes are coming into operation, and equipment engineers can choose, with confidence, a tube to suit their particular requirements from the catalogues.

The tubes most commonly in use may be classified as power, intermediate or low noise tubes, and cover the frequency range 1 000 to 10 000Mc/s with gains up to 40dB.

This article discusses some applications of travelling-wave tubes, their advantages and the problems of installation.

Communication Equipment

The spread of telephonic communication throughout the world has led to the demand for radio links capable of carrying an ever-increasing number of telephone channels. This in turn calls for greater r.f. bandwidths and the extension of carrier frequencies into the microwave region.

The inherent broad bandwidth and high gain travelling-

* *English Electric Valve Co. Ltd.*

wave tubes make them very suitable in link repeaters especially if all the amplification is performed at r.f. by simply cascading three or four tubes. Small capacity systems can utilize ordinary i.f. amplifiers with appropriate frequency changing before and after the i.f. amplification, but usually a t.w.t. is most conveniently used in the output stage giving upwards of 4W. For microwave links with more than six repeater stations carrying 600 or more telephone channels to C.C.I.R. standards, an all travelling-wave tube system is highly competitive[1].

The tube line-up in a typical repeater is a low noise tube, two intermediate tubes and a power output tube. One of the intermediate tubes is used as a frequency changer, say, for example, from 2 000 to 2 100Mc/s, by modulating the helix voltage at the difference frequency, 100Mc/s. The new frequency is selected by means of a bandpass filter, the unwanted frequencies being reflected back into the t.w.t. where they are absorbed. This mode of operation is accompanied by a loss in gain[2] of about 10dB.

There are special problems in the impedance matching of t.w.t.'s for high capacity telephone systems. It has been shown[3] that whether or not t.w.t.'s are used, the v.s.w.r. in some parts of the system, notably the output of the output stage, must be of the order of 1·1 over a bandwidth of 10 or 20Mc/s. Whereas there is little difficulty in adjusting the match of a t.w.t. to this figure when no voltages are applied, the v.s.w.r. often rises when the valve is switched on. This is because a signal reflected back into the output of a t.w.t. travels down the helix and is absorbed except in so far as there are small reflections occurring either at the attenuator or at small irregularities in the helix itself. These reflections, moving now in the same direction as the electrons, are amplified in the usual way by about 20dB. Thus, even though the reflected voltage wave is only 0·2 per cent of the original, a v.s.w.r. of 1·2 appears at the output. In practice the effect is less than this when the t.w.t. is delivering power to the aerial, particularly when the tube is working at saturation, and it is possible to correct for this mismatch over an adequate bandwidth by normal adjustments at the output.

Because the amplification process is involved, the magnitude of the reflected wave will vary as the helix voltage is varied, being a maximum when this is adjusted for small signal operation. This phenomenon can be utilized as a further matching adjustment at the expense of some loss in gain. The helix voltage must therefore be carefully stabilized to reduce variations to less than 1 per cent if small v.s.w.r.'s are required.

This problem is almost completely removed if a ferrite isolator with 20dB or more in the reverse direction is used. The v.s.w.r. at the valve can then rise to 1·5 without affecting system performance. The cost of an isolator is offset by the omission of precise matching controls, and test equipment to measure the v.s.w.r.

Travelling Wave Amplifier Units

Although the t.w.t. has been much publicized for its very wide bandwidth, the applications where full use is made of this property are very few. Fig. 1 shows a packaged t.w.t. complete with power supplies suitable for laboratory or field use where a wide-band instrument is required. The particular unit shown is a low noise amplifier with a noise factor less than 8dB over the band 1 200 to 1 400Mc/s and is suitable for experiments on aerial performance. If the t.w.t. was replaced by an intermediate tube, the same unit could be used to amplify a signal generator for laboratory uses. As the frequency is changed, no adjustments whatever need be made to the t.w.t. The gain may be varied if required by means of the first anode voltage control which in turn varies the beam current.

The box shown in Fig. 1 can be mounted directly in a repeater or radar equipment. This method of installing a tube is to be recommended in situations where only unskilled personnel are available or when specialized test equipment is not kept on the site. Critical adjustments, such as precise impedance matching or setting up for optimum noise factor, can then be made beforehand at

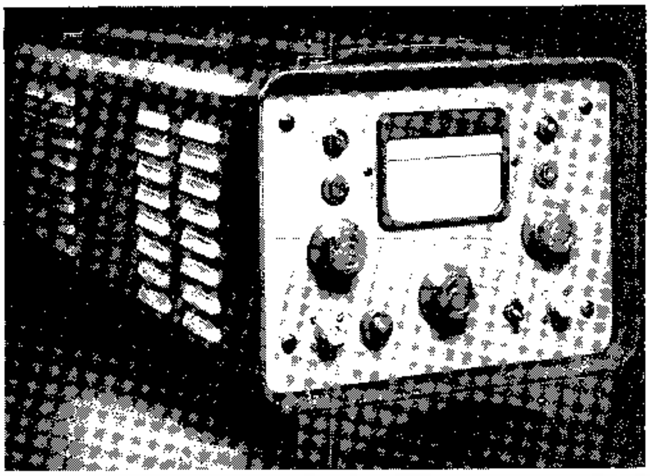


Fig. 1. A travelling-wave tube amplifier complete with power supplies and lightweight electromagnet

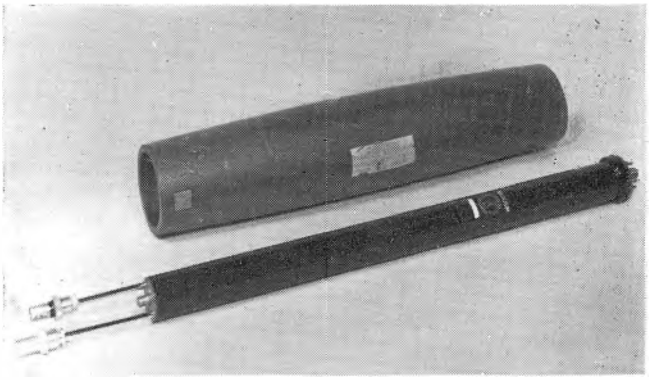


Fig. 2. A uniform permanent magnet shown with the travelling-wave tube for which it is intended

The mechanical arrangements for mounting are not shown

a service depot. In this unit the only hot-cathode tube is the t.w.t. itself. Copper oxide and germanium devices are used as rectifiers in the power supplies for the t.w.t. and solenoid, while the voltage is stabilized by neon stabilizers.

Magnetic Field for T.W.T.'s

As with most devices depending for their operation on a long beam of electrons, the t.w.t. must be placed in a strong magnetic field parallel with its axis. Until fairly recently this has been obtained by an electromagnet wound with insulated copper wire carrying about 2A at 40V. Such electromagnets are very heavy and costly, and it is probable that had this not been so, many more travelling-wave tubes would be in use today than there are. Great efforts have therefore been made to overcome this problem and though modern t.w.t.'s still need focusing fields, the means for obtaining them have been improved.

By far the simplest replacement for the copper electromagnet is a straightforward permanent magnet (Fig. 2). Such magnets overcome the objections of cost, weight and the need for a power supply, and can now be obtained from tube manufacturers for an increasing number of their products. In some applications steps must be taken to screen neighbouring equipment from the stray magnetic field. This tends to make the space taken up by the t.w.t. rather large. The same problem exists with electromagnets to a lesser degree but, in this case, the screens may be brought up close to the electromagnet without seriously weakening the field strength along the axis. This is not possible with permanent magnets.

Another means for focusing the electron beam is the periodic permanent magnet[4] comprising a series of small

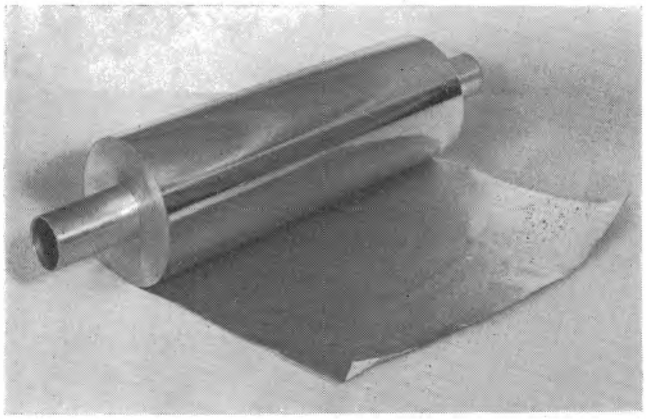


Fig. 3. Aluminium foil for electromagnet

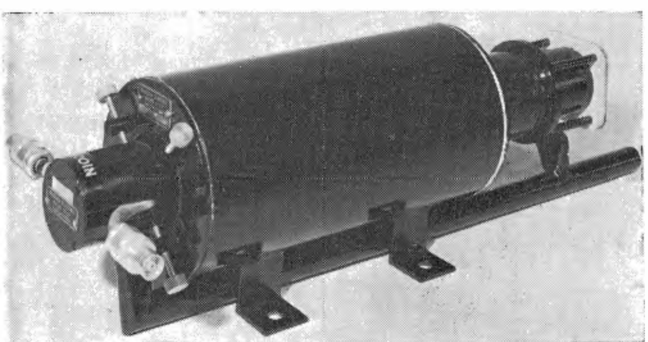


Fig. 4. A completed lightweight electromagnet for a power output t.w.t.
Some of the air needed for cooling the collector is diverted through the centre of the winding allowing a considerable reduction in size and weight. The magnetic field is 375 oersteds

permanent magnets alternating in polarity, so that a plot of the field along the axis is approximately a sine wave. This magnet assembly overcomes all the objections of both copper electromagnets and uniform permanent magnets except that of cost, and this will be almost certainly overcome in time as a result of collaboration between tube designers and the manufacturers of the magnet 'cells'. Until then, there is little likelihood that periodic magnets can be considered seriously for commercial equipments.

Some work is being done on electrostatic focusing[5] for t.w.t.'s but this is still in the development stage, and in any case is not applicable to present-day tubes.

The aluminium foil electromagnet shown in Fig. 3 is a practical solution to the problem of focusing in commercial equipment. The foil is wound tightly on an aluminium tube which forms the connexion to the inner turn of the winding, and is coated on one side only with a resin varnish. The volume is about the same as and the weight less than one-third that of a copper wire coil. The minimum field required to focus any particular t.w.t. is up to 25 per cent less with the foil coil. The heat conductivity is also slightly better, although this is offset to some extent by the higher electrical resistivity of aluminium.

The volume of any electromagnet is determined solely by the safe temperature rise so that, if the coil can be cooled by forced air, the size can be reduced appreciably. Electromagnets obtainable from tube manufacturers are usually designed for running with natural cooling and tend to be conservatively rated. It would be advisable for an equipment designer to consider the magnetic field requirements at an early stage in drawing his layout. Very often an equipment uses forced air for other purposes and, if some can be diverted for cooling the electromagnet, a much smaller design can be made.

Fig. 4 shows an aluminium electromagnet for use with a 4 000Mc/s power t.w.t. It weighs 15½lb and is rated at 14V 8A. In this case, where the collector of the t.w.t. must be cooled by forced air, arrangements have been made for passing a small stream of air down the centre of the coil.

Life Expectancy

The problems of life of oxide cathodes in t.w.t.'s are common to all tubes using long electron beams. One of the main causes of failure is the bombardment and destruction of the cathode surface by positive ions originating in the electron beam. Most of this can be eliminated by rigorous processing and pumping and by the use of convergent guns wherever possible. In power output tubes which have convergent guns, some destruction of the central portion of the cathode does occur, but the area affected is small compared with the total area. Lives well in excess of 10 000 hours are being recorded for such tubes.

Further reduction of ion bombardment can be achieved by means of the 'potential barrier' form of ion trap[6] in intermediate and power tubes.

In tubes so far developed it has not been found possible to combine a low noise factor with either a convergent gun or an ion trap. Nevertheless lives of 5 000 hours are being regularly obtained and it is expected 10 000 hours will eventually become commonplace.

Conclusion

Recently developed lightweight electromagnets and permanent magnets have overcome the obstacles to the adoption of travelling-wave tubes in commercial equipments.

A complete unit has been described in which the t.w.t., solenoid and all power supplies are incorporated into a portable box which can be used as a laboratory test equipment or installed directly into a major equipment such as a radar or radio repeater.

The advantages and some of the problems of the application of t.w.t.'s to radars and radio repeaters have been discussed.

Acknowledgments

The author desires to acknowledge with thanks the permission of the General Manager of the English Electric Valve Company Ltd for the publication of this article.

REFERENCES

1. FEDIDA, S. All-Travelling-Wave Tube System. *Electronic Engng. 30,* 283 (1958).
2. BRAY, W. J. The Travelling-Wave Valve as a Microwave Phase-Modulator and Frequency-Shifter. *Proc. Instn. Elect. Engrs. 99,* Pt. 3, 15 (1952).
3. FEDIDA, S. Some Design Considerations for Links Carrying Multichannel Telephony. *Marconi Rev. 18,* 132 (1955).
4. MENDEL, T. J., QUATE, C. F., YOCOM, W. H. Electron Beam Focusing with Periodic Permanent Magnet Fields. *Proc. Instn. Radio Engrs. 42,* 800 (1954).
5. CHANG, K. K. C. Confined Electron Flow in Periodic Electrostatic Fields of Very Short Periods. *Proc. Instn. Radio Engrs. 45,* 66 (1957).
6. LAICO, J. P., McDOWELL, H. L., MOSTER, C. R. Medium Power Travelling-Wave Tube for 600-Mc Radio Relay. *Bell Syst. Tech. J. 35,* 1285 (1956).

TRAVELLING-WAVE TUBE AMPLIFIERS

By D. H. O. Allen*, B.A.

After a very brief résumé of the fundamental principles of travelling-wave interaction this article discusses the major advances made in the last few years towards improved and more useful travelling-wave tube amplifiers. A better insight into, for example, the noise behaviour of long beam amplifiers and periodic focusing methods has enabled the travelling-wave tube to become an essentially practical device suitable for incorporation in reliable communication equipment. The applications of the t.w.t., especially in microwave radio links, are also discussed together with comments on the distortions that are likely to be introduced under various operating conditions.

A FEW years ago the travelling-wave tube was described as the most publicized and least used of all microwave tubes in existence. Recently, however, circuit and system designers have come to realize that the t.w.t. amplifier can form a useful part of their equipments. One reason for its late entry into the field of practical equipment has, perhaps, been the natural timidity of system engineers to use a new and untried device, but in addition it is clear that until a few years ago valve manufacturers had not brought the t.w.t. to a sufficiently practical form. Why this is so is hard to determine, but one of the major practical disadvantages of the normal t.w.t. is the fact that an external magnetic field is necessary in order to focus the electron beam. Early focusing systems were very bulky and often consumed several hundreds of watts of d.c. power, but the advent of new ideas in the focusing of electron beams has in many cases enabled complete t.w.t. 'packages' to be made small in size and weighing only a few pounds.

Fundamental Aspects of the Travelling-Wave Tube

In principle the t.w.t. is a microwave amplifier depending for its operation on the interaction of an electron beam with an electromagnetic wave (representing the signal) travelling nearly at the same velocity. In practice it is extremely difficult to accelerate an electron beam to near the velocity of light, but the wave can be effectively slowed down by causing it to travel along, for example, a conductor coiled in the form of a helix. An observer on the axis of such a helix would see the electric and magnetic fields associated with the wave travelling past him with an axial velocity of $c \sin(p/2\pi a)$ where c is the velocity of light, and p and a are respectively the pitch and mean radius of the helix.

If the electron beam is projected down the helix at the same velocity as the apparent axial velocity of the wave individual electrons will either be retarded or accelerated according to the direction of the electric fields as they enter. Eventually the electrons travel in 'bunches' but no amplification occurs since an equal number are accelerated and retarded. If, however, the beam enters at a fractionally greater velocity than the wave, more electrons will be retarded and the net effect will be for the beam to give up energy to the wave. In this way the wave (or signal) is amplified.

Basically the t.w.t. amplifier needs merely an electron beam at the correct velocity and a suitable slow wave circuit. In practice, however, the provision of a suitable beam for a low noise or, perhaps, a high power tube may present special problems and these will be discussed below. Different types of slow wave circuits have different properties and some are more suitable for certain applications than others. However, the first t.w.t. made[1] used a helical conductor and the first serious attempts at providing a comprehensive theoretical understanding of the device was carried out by Pierce[2] using a helix type tube as a model. Since those early days several different types of slow wave structure have been considered and these are discussed below.

Slow Wave Structures

The primary function of the slow wave structure in a t.w.t. is to enable the electric fields of an e.m. wave to couple continuously with a slow moving electron beam. The properties of a slow wave structure may be described in terms of the apparent velocity of propagation of the e.m. wave that is injected into it and the manner in which it couples to an electron beam. The broad bandwidth properties of a t.w.t. arise from the fact that the only criterion for useful beam-to-wave interaction is that their relative velocities should be correct. The intrinsic bandwidth is then only limited by the dispersion characteristics of the slow wave structure.

Pierce describes an impedance for a slow wave structure which determines how effectively its fields will interact with an electron stream and this is defined as $K = E_z^2/\beta^2 P$, where E_z is the useful electric field in the beam region, β is the propagation constant and P the power flowing in the structure. The gain of a t.w.t. amplifier is proportional to a parameter C given by $C = (KI_0/4V_0)^{1/3}$ where I_0 and V_0 are the current flowing in the electron beam and the effective voltage of the beam respectively.

Multi-channel communication systems require wide bandwidths and, for minimum distortion, it is clear that over this bandwidth there should be no variation of gain or time delays in any amplifier that may be used. It is therefore of paramount importance that a t.w.t. amplifier used for wide band communications should employ a slow wave structure that is non-dispersive (i.e. has a phase velocity that is independent of frequency), and that also has an effective coupling impedance that is constant over a wide frequency band.

The helical form of slow wave circuit used under the right conditions appears, at the present time, to be the ideal one for communications systems. It is true that several other forms of circuit, such as the disk loaded waveguide and several ladder-like structures have coupling impedances which can be orders greater than that of the helix, but in general these are very dispersive and may have only a few hundred kilocycles of useful bandwidth.

Another form of structure that is worthy of note, is the interdigital (i.d.) line[3]. In its usual form, consisting of a series of intermeshed conducting pins, it is used in backward wave oscillators when its backward fundamental slow wave interacts with the electron beam. However, the fundamental forward space harmonic has relatively little dispersion under its normal operating conditions, but shows only a small coupling impedance to an electron beam. Curves comparing the dispersion characteristics of the helix and i.d. line are shown in Fig. 1. At lower frequencies (for example 1 to 20kMc/s) there is no advantage in using such a structure because of its low coupling impedance, but where very high frequency operation (30 to 100kMc/s) is

* *Mullard Research Laboratories*

considered this is an important advantage in that the i.d. line is easier to make, being somewhat larger than a comparable helical structure at the same frequency.

So far no mention has been made of high power applications. Without special cooling arrangements the helix is capable of operating in t.w.t. amplifiers giving at most a few hundred watts at low frequencies of 500 to 1 000Mc/s. At much higher frequencies where the slow wave structures become small, and resistive loses become greater much lower power levels only can be realized. It would seem, however, that by conventional means a few watts can be obtained at frequencies up to 11·5kMc/s which at present is the highest frequency band envisaged for normal microwave communications systems.

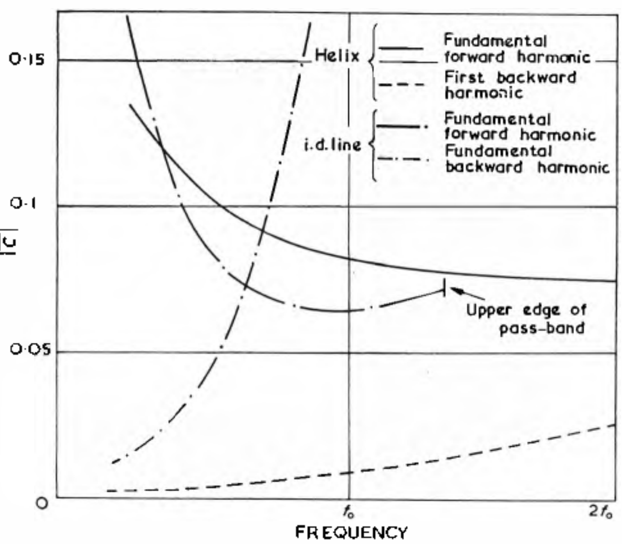


Fig. 1. Phase velocities/frequency calculated for a helix and interdigital line designed for optimum performance at frequency f_0

v/c is the ratio of the phase velocity to that of light

The Noise Problem

The range of a communication system is determined by the maximum output power available at the transmitter, the transmitter and receiver aerial gains, the path loss between transmitter and receiver and the noise factor of the receiver.

In the science and engineering of t.w.t. recent advances have provided a greater understanding of the factors contributing to noise output and noise factor, and the latest theories suggest that the minimum noise factor obtainable may be as low as 3·5dB.

Noise in t.w.t. amplifiers arises from several different causes, some of which are readily eliminated by careful engineering. Thus the familiar partition noise effect found in conventional pentodes and tetrodes is also present in t.w.t.s' if the slow wave circuit is allowed to collect some of the beam current near its input end. Proper alignment of the helix and gun and correct design of the focusing magnetic field can eliminate this without difficulty—at least at the lower microwave frequencies where the structures are reasonably large.

The fundamental source of noise lies in the cathode and is due to the inherent high temperature of the electron-emitting surface and the familiar shot effect which arises from the randomness in time of emission of individual electrons. The elevated temperature of the cathode gives rise to a velocity distribution of electrons which initiates a 'velocity' source of noise propagated on the plasma waves of the electron beam. The shot effect produces a 'current' source of noise also propagated in a similar way.

It would seem that a reasonably complete understanding of the physical mechanism of the methods by which these noise waves may be effectively reduced has been made in recent years by Watkins[4], Bloom and Peter[5], and other workers in the field. As a result a considerable amount of published data on the noise performance of practical tubes is available, at least for amplifiers at frequencies below 7·5kMc/s.

Work by these authors predicts a minimum attainable noise factor of about 6dB, but this figure is largely determined by the assumed 'noisiness' of the electron beam before it enters the noise reduction region of the electron gun. Recent work by several workers has been devoted to acquiring a better understanding of the mechanisms existing in the region between the emitting surface of the cathode and the potential minimum plane and two facts now seem to be clearly established. Firstly, there is no correlation between the velocity and current sources of noise and this has been borne out by experiment. Secondly, it is not a valid approximation in any analysis to consider the electron flow as single-velocity in nature. Theoretical work by Siegman *et al*[6] has confirmed that the multi-velocity nature of the flow in the potential minimum region must be considered. If this is done then noise figures of around 3dB are predicted.

Most of the experimental verification of noise figures of t.w.t. amplifiers has been done at the lower frequency end of the microwave spectrum, generally in the band 2·4 to 4kMc/s. From published results there seems to be no doubt that noise factors as low as 6dB can be obtained consistently in well designed tubes at 3 000Mc/s without very critical adjustments and that reasonably good life may be obtained from such tubes[7].

At higher frequencies there is doubt as to whether the noise theories developed for the cathode-potential minimum regions are still valid, but noise factors of as low as 8·5dB have been reported at X-band. One of the major difficulties at these higher frequencies is that of eliminating partition noise arising from the interception of some of the beam by the very small helix that is employed. Only a tightening of an already very strict mechanical tolerance in the tube engineering can overcome this. A further difficulty at high frequencies is caused by the necessity to go to higher emission densities from the cathode which may well shorten the tube's useful life. One possible solution to this lies in the use of some of the dispenser or impregnated[8] cathodes, but these have the disadvantage of operating at a higher temperature than the conventional oxide-coated type. They do, however, have the advantage of possessing a very smooth uniform emitting surface which may well offer advantage when very small but high density electron beams are required for low noise devices.

Focusing Systems

An electron beam will normally diverge under its own space charge forces, so in order that all the beam may be projected right through a slow wave structure (possibly up to one foot in length) some focusing system must be provided. The most straightforward way of achieving this is to use a magnetic field to confine or focus the beam over the length of the tube. The magnetic field may be provided in two ways, either by means of a solenoid energized from an external d.c. supply or by a permanent magnet system. Whether or not the latter solution is possible is determined largely by the actual field requirements; an electron beam of very high density may require a field that is most impracticable to provide with a permanent magnet system of reasonable size and weight. In this case a solenoid is the more suitable, though this will, of course, require an associated power supply. The size and weight of focusing solenoids for t.w.t.'s have been materially reduced recently by the use of either thin paper or anodizing as the insulation for aluminium foil windings[9].

From the point of view of magnet economy the best method of focusing an electron beam is by the 'Brillouin flow' method[10]. Here, immediately after leaving the electrostatic focusing region of the electron gun the beam is set into rotation by a sudden transverse magnetic field which changes rapidly to a uniform axial field down the length of beam.

It may be shown in the case of an ideal beam that the sum of the centrifugal force, the outward space charge force and the inward magnetic force on an electron in the beam is zero if the axial magnetic field, B, is given by

$$B^2 = 0{\cdot}69 \times 10^{-6} (I / V^{\frac{1}{2}} r_0^2)$$

where I = beam current, V = beam voltage and r_0 = beam radius (all in m.k.s. units).

In practice, due to thermal velocity effects of emission and an inevitable departure from the ideal theoretical input transition from the gun into the axial field, such a condition cannot quite be realized in practice, but satisfactory focusing has been attained with values of about 1·5 times the ideal Brillouin value.

In cases where an electron gun specially designed for optimum focusing cannot be employed, such as in low noise tubes, it is necessary to resort to 'confined flow' or, in popular parlance, 'brute-force focusing'. Here the axial magnetic field is made to be several times the theoretical Brillouin value and extends over the cathode region, but the beam boundary shows undulations which decrease in magnitude as the field is increased. Linear flow occurs only for the limiting case of an infinite magnetic field. Low noise tubes generally employ very low perveance beams of low density and therefore the fields required are realizable ones even if they need to be as high as ten times the theoretical Brillouin value.

Perhaps the most spectacular improvements in the practicability of t.w.t. amplifiers have been with the advent of periodic magnetic focusing systems. The realization that the focusing action of a magnetic field on an electron beam was proportional to the square of the value of the field prompted an investigation into the use of magnetic fields that varied periodically down the length of the tube[11,12]. In fact such a system may be considered as a succession of converging magnetic lenses where, for stable flow, the convergence produced is equal to the divergence caused by the space charge forces. Such a system is somewhat more critical to set up, especially for high perveance beams, but it has several advantages. The fields may be produced by a stack of magnetized disks which are isolated from one another by steel pole pieces and therefore to increase the length of a particular system merely means adding more disks and pole pieces. In a uniform field permanent magnet system the increase in length of the uniform field region would mean an increase to the power of three in both size and weight. In addition, since each successive effective magnetic dipole is reversed in direction the resultant field some distance from the magnet system is very small and most of the field is confined to the axis, where it is required. As a result t.w.t. circuits employing periodic permanent magnet focusing can be as much as ten to fifteen times lighter in weight than the uniform field counterpart. Fig. 2 and Fig. 3 show practical examples of these.

Obviously there are very great advantages in not having a magnet at all and then recourse must be made to an electrostatic periodic system. The problems involved in this have been examined at length, but only recently has come news of a practical tube[13] giving 10dB gain and 100mW of power output at 3 000Mc/s. This particular tube uses concentric bifilar helices to provide the slow wave circuit and the necessary periodic static electric fields to focus the annular electron beam that is made to flow down the gap between the helices. Such a tube is, of course, inherently

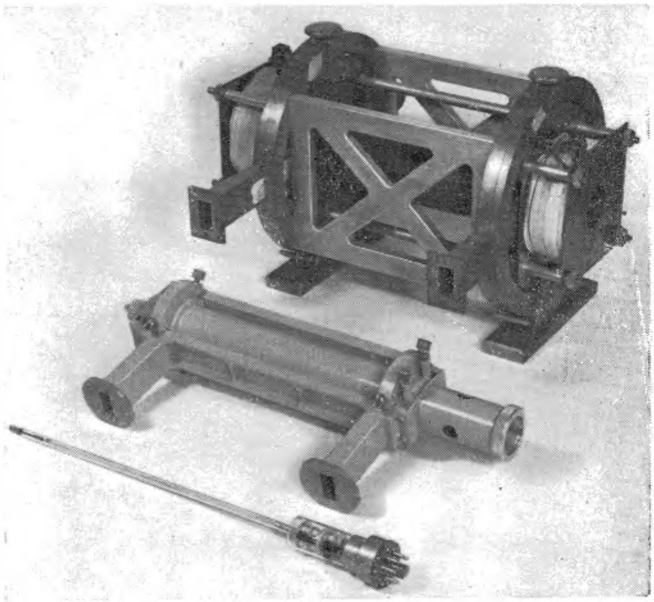


Fig. 2. Solenoid and periodic permanent magnet systems with coupling waveguides for X-band t.w.t. amplifier type VX8506

(gain >20dB from 7 to 11·5kMc/s), the weights are 39 and 4½lb respectively and the coil requires 24V, 4·5A

Fig. 3. Periodic permanent magnet and coupling system for the Mullard LA4-250 intermediate amplifier

This t.w.t. will give over 25dB of gain and 250mW of power in the band 3·6 to 4·2kMc/s

more complex than one focused by conventional means, but the user of such tubes may well be able to offset this against the comparative cheapness of the associated waveguide circuits which will then no longer require magnets or the rather complicated arrangements for moving them for accurate alignment.

Operation at Higher Power Levels

As has been stated earlier the most suitable slow wave structure for a t.w.t. amplifier for communications is the simple monofilar helix and this proves itself entirely adequate for power levels up to a few tens of watts of frequencies in the 4kMc/s band, somewhat less in the 6kMc/s band and a few watts only in the 11kMc/s band. In brief, the limiting factors against high power output are the increased losses and reduced size of the slow wave structures at high frequencies. The first implies the use of higher d.c. powers in the beam and the second the use of electron beams of very high electron density travelling close to the slow wave circuit which means that very large magnetic focusing fields will be required. It must be remembered that heat is dissipated in the slow wave circuit both by the passage of a large r.f. signal along it and by the inevitable collection of some of the d.c. electron beam power.

Several other types of slow wave structures which can be machined from solid copper are undoubtedly capable of higher dissipations but may fail for reasons of bandwidth, group delay etc. However, these may well be suitable for tropospheric scatter systems where the bandwidths employed may not be very large and high powers are a requirement.

At the lower frequencies (below 1 000Mc/s) t.w.t. power amplifiers may well be used as final or penultimate stages in television transmitters and Fig. 4 shows a helix type amplifier which is capable of delivering 2·5kW c.w. at 700Mc/s. In this case size is not the main factor determining the maximum possible power output since the helix is wound from $\frac{1}{8}$in. copper pipe through which cooling water is pumped continuously. The electron beam size is nearly an inch in diameter and even though the current is several amperes the current density is low and thus focusing magnetic fields of only 500 gauss are sufficient.

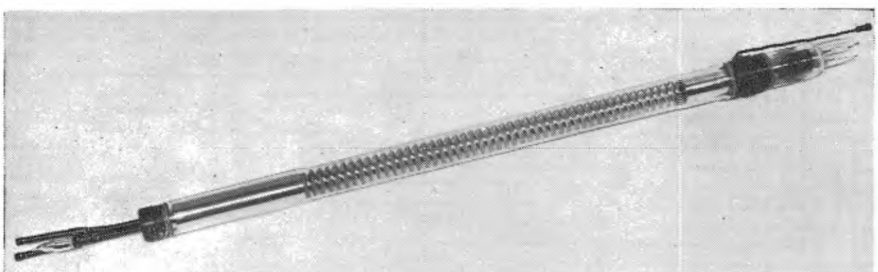


Fig. 4. An example of a higher power c.w., t.w.t. amplifier developed in the laboratories of N.V. Philips, Eindhoven

This tube will deliver 2·5kW of power with 16dB gain at 700Mc/s

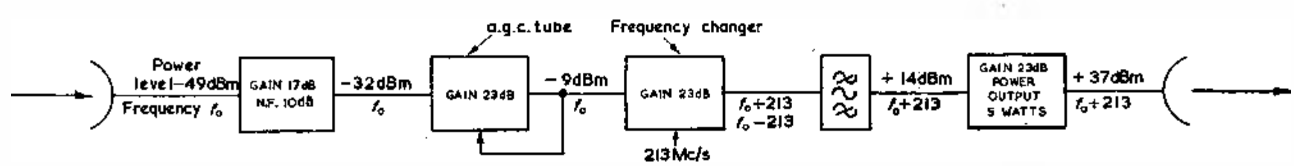


Fig. 5. A possible all t.w.t. microwave repeater containing four t.w.t. amplifiers

Applications

One important requirement of a microwave communication system carrying a large number of telephone and telegraph channels is an amplifier of large bandwidth. Thus the t.w.t. is an obvious choice and has already seen several years of service in the Manchester-Kirk O'Shotts television link[14]. As an example of its use in a multi-channel telephony system there is shown in Fig. 5 a proposed all-microwave repeater at a carrier frequency of, for example, 4kMc/s.

From consideration of path loss between repeater stations, atmospheric fading, thermal noise and noise produced by intermodulation effects it would seem that an ideal repeater, up to C.C.I.R. standards, for a 600 channel link should have a power output of 5W, an input noise factor of 10dB and an overall gain[15] of about 85dB. In addition, a device for side-stepping the frequency by about 200Mc/s in order that there may be no instability arising from feedback between input and output aerials, must be included. Since the electrical length of a t.w.t. may be varied by several radians merely by altering the helix potential, a phase modulation may be imparted to the r.f. signal being amplified and if the necessary filters are included at the tube output to select the wanted sideband the combination may be used as a frequency shifter[16,17]. Two methods of doing this have been discussed using either sine or sawtooth modulation[18] voltages, but at the present time the sine form only is practicable at 200Mc/s and this is normally used. In this case there is a theoretical loss of gain of 6dB, but this has been allowed for in the overall gain calculations for the repeater.

This typical repeater arrangement indicates the versatility of the t.w.t. amplifier as a circuit element—in one particular equipment it is used as a low noise amplifier, as an a.g.c. device to allow for fading, as a frequency shifter and as a power amplifier. In other forms of communication repeater that have been proposed and in some cases are in use, a typical repeater obtains most gain in an intermediate frequency amplifier at less than 100Mc/s using conventional valves and merely uses the t.w.t. as a modulator for applying the information to the microwave carrier, followed by a power-amplifier t.w.t.

A further type of communication system which is rapidly coming into use is that using tropospheric scatter[19]. Frequencies up to 5 000Mc/s have been considered for this technique and although the bandwidth requirements will be relatively small (certainly less than 10Mc/s) there would be an advantage in using a t.w.t. amplifier if it had a noise factor less than that of conventional crystal mixers. From various sources it would seem that the crystal mixer, adjusted for the lowest possible noise factor is not a very reliable device and there would, therefore, be advantages in using a low noise t.w.t. amplifier followed by a medium-noise mixer. On reliability grounds a low noise t.w.t. seems entirely satisfactory; there are published figures[20] of at least 5 000 hours without change of noise factor and, more recently 14 000 hours of life has been recorded without failure.

The power amplifiers considered for tropospheric scatter have been triodes at the lower frequencies and klystrons at the higher frequencies. Although the klystron's bandwidth is just about adequate for this form of communication and its gain is high (perhaps 50dB) the t.w.t. may be more advantageous where price is an important consideration. Generally speaking a glass envelope valve is cheaper to produce than an all-metal one and practical t.w.t.'s of all glass construction have been shown to be capable of providing at least 2·5kW power output at up to 1 000Mc/s. A further application of tubes of this type lies, of course, in the field of television transmitters for the higher frequency bands where gain competition with klystrons must be considered, although where powers of 10kW c.w. are required the all-metal klystron may be the only possibility in view of the cooling requirement.

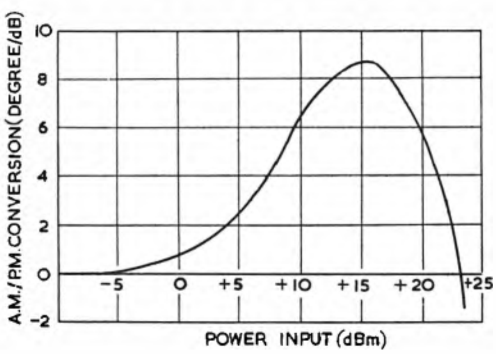


Fig. 6. Typical a.m./p.m. conversion for a t.w.t. amplifier
The phase shift in degrees per decibel change in input level has been plotted against input power

One important feature of any amplifier is the distortion which it may introduce into the signal being amplified. In a multi-channel communication system this will appear as noise which may considerably worsen the signal-to-noise ratio of the overall system. Forms of distortion which may contribute most to unwanted noise in systems are:

(1) Variation of gain across the required bandwidth.

(2) Differential phase distortion across the bandwidth.

(3) An amplitude to phase modulation.

Gain variations across the working bandwidth can be reduced by ensuring that an adequately non-dispersive slow wave structure is used and that the coupling system to the external circuits are adequately matched. Differential phase shifts arise from the dispersive nature of slow wave circuits but with the helix operated in its correct regime this form of distortion is very small and can normally be disregarded.

When a t.w.t. amplifier is being operated near saturation variations in input signal level will alter the velocity of the beam near the output of the tube since more or less energy is being extracted from the beam. As has been discussed earlier, variations of beam velocity cause variations of the phase of the output signal and thus an a.m. component on the input signal can give rise to an a.m. + p.m. component at the output. This will mean that any filters that precede such an amplifier must have an absolutely flat characteristic of transmission versus frequency. If this is not possible then the onus falls upon the tube designer to provide a t.w.t. with a very small a.m./p.m. conversion. Theory suggests that the phase shift at the output per decibel of input amplitude change is given by $9{\cdot}6\ P_{out}/P_{beam}C$ degrees (C in Pierce's gain parameter) as long as saturation is not approached too closely. Nearer saturation the situation worsens and the phase shift is greater than this small signal theory suggests. Published results on one 6kMc/s power amplifier t.w.t. closely follow the theoretical prediction[21]. It is interesting to note that when $P_{out}/P_{beam} = 1/15$ the shift is approximately $2{\cdot}25°$/dB, whereas it has been suggested that a 600 channel system and a colour television link might tolerate only 1°/dB. System requirements on this point appear not yet to have been fully evaluated. It may well mean, however, that all t.w.t. amplifiers in a high quality link may have to run at or below about one-quarter of their normal saturated power output. It will be noted that near saturation the phase shift curve turns over and at saturation the phase shift per decibel becomes zero. In principle this region could be used for low a.m./p.m. conversion, but the mean level in the system would have to be kept very constant against fading. (See Fig. 6).

Conclusions

Without doubt the t.w.t. amplifier is an extremely versatile microwave amplifier with few disadvantages. Its full potentialities are only just being exploited and system designers have only very recently come to regard it as a component engineered to a sufficiently high degree for incorporation into equipments.

The basic principles of the t.w.t. are now well known and understood, although certain work still needs to be done towards a better understanding of its behaviour under large signal conditions. The recent reports of a t.w.t. amplifier using an electrostatic focusing system and having a useful r.f. performance are extremely encouraging and may well mean that certain types of t.w.t. amplifier will become small in size and weight and require no bulky magnet system. The many advantages of the t.w.t. amplifier are now being evaluated by system designers and shortly it will no doubt become an integral part of many communication systems.

Acknowledgments

The author wishes to thank the Directors of Mullard Ltd. and the Director of Mullard Research Laboratories for permission to publish this article.

REFERENCES

1. KOMPFNER, R. The Travelling-Wave Tube Centimetre-Wave Amplifier. *Wireless Engr. 24*, 255 (1947).
2. PIERCE, J. R. Travelling-Wave Tubes. *Bell Syst. Tech. J 29*, 390 (1950), and PIERCE, J. R. Travelling-Wave Tubes (D. van Nostrand Co. 1950).
3. WALLING, J. C. Interdigital and Other Slow-Wave Structures. *J. Electronic Control 3*, 239 (1957).
4. WATKINS, D. A. Travelling-Wave Tube Noise Figure. *Proc. Inst. Radio Engrs. 40*, 65 (1952).
5. BLOOM, S., PETER, R. W. Transmission-line Analog of a Modulated Electron Beam. *RCA Rev. 15*, 95 (1954).
6. SIEGMAN, A. E., WATKINS, D. A., HSUNG-CHENG HSIEH. Density-Function Calculations of Noise Propagation on an Accelerated Multicavity Electron Beam. *J. Appl. Phys. 28*, 1138 (1957).
7. KINAMAN, E. W., MAGID, M. Very Low Noise Travelling-Wave Tube. *Le Vide.* No. 70, 308 (1957).
8. LEVI, R. Improved "Impregnated Cathode". *J. Appl. Phys. 26*, 639 (1955).
9. WORCESTER, W. G., WEITZMANN, A. L., TOWNLEY, R. J. Lightweight Aluminium Foil Solenoids for Travelling-Wave Tubes. *Trans. Inst. Radio Engrs.* ED-3, 70 1956).
10. BRILLOUIN, L. A Theorem of Larmor and Its Importance for Electrons in Magnetic Fields. *Phys. Rev. 67*, 260 (1945).
11. CLOGSON, A. M., HEFFNER, H. Focusing of an Electron Beam by Periodic Fields. *J. Appl. Phys. 25*, 436 (1954).
12. MENDEL, J. T., QUATE, C. F., YOCOM, W. H. Electron Beam Focusing with Periodic Permanent Magnet Fields. *Proc. Inst. Radio Engrs. 42*, 800 (1954).
13. CHANG, K. K. N. Biperiodic Electrostatic Focusing for High Density Electron Beams. *Proc. Inst. Radio Engrs. 45*, 1522 (1957).
14. DAWSON, G., HALL, L. L., HODGSON, K. G., MEERS, R. A., MERRIMAN, J. H. H. The Manchester-Kirk o'Shotts Television Radio-Relay System. *Proc. Instn. Elect. Engrs. 101*, Pt. 1, 93 (1954).
15. FEDIDA, S. Wide Band Microwave Radio Links. *Marconi Rev. 18*, 69 (1955).
16. BRAY, W. J., The Travelling-Wave Valve as a Microwave Phase-Modulator and Frequency-Shifter. *Proc. Instn. Elect. Engrs. 99*, Pt. 3, 15 (1952).
17. STEELE, G. F. The Modulation of Travelling-Wave Tubes. *Electronic Engng.* 29, 429 (1957).
18. CUMMING, R. C. The Serrodyne Frequency Translater. *Proc. Instn. Elect. Engrs. 45*, 175 (1957).
19. BULLINGTON, K. Characteristics of Beyond-the-Horizon Radio Transmission. *Proc. Instn. Radio Engrs. 43*, 1175 (1955).
20. ALLEN, D. H. O., WINWOOD, J. M. A Low Noise Travelling-Wave Tube Amplifier for the 4000 Mc/s Communication Band. *J. Brit. Instn. Radio Engrs. 17*, 75 (1957).
21. LAICO, J. P., MCDOWELL, H. L., MOSTER, C. R. Medium Power Travelling-Wave Tube for 6000-Mc Radio Relay. *Bell Syst. Tech. J. 35*, 1285 (1956).

Travelling-Wave Tubes For 4 000Mc/s

By P. F. C. Burke*, B.A., Grad.I.E.E.

Travelling-wave tubes may be used in microwave radio links as output amplifiers, low-noise receivers or intermediate level amplifiers. The requirements for each of these applications and their influence on the design of the travelling-wave tubes are discussed. The application of these principles is illustrated by description of such tubes developed for 4 000Mc/s microwave links.

THERE has been a great development of wideband microwave radio links since 1950, particularly in the 3 600 to 4 200Mc/s band, and many of these have used travelling-wave tubes. A review of some such tubes developed for this frequency range will give an idea of their present capabilities and also the requirements that must be met for satisfactory operation in communication systems.

While all travelling-wave tubes employ the same basic principle for amplification, and can be used for this function at any stage in a microwave repeater, it will be convenient to consider separately the output amplifier, the low noise receiver and the intermediate level amplifier. Just as different circuit techniques would be used for these stages at lower frequencies, so there are considerable differences in the design of the corresponding travelling-wave tubes. Furthermore it may be desirable to use these tubes in only one or two of these three stages, depending on the requirements and economics of the complete system.

Output Tube

The fundamental requirements for the output tube in a microwave radio link have been extensively discussed before[1], with particular reference to the CV2188. The use of this tube in 1952 in the Manchester-Kirk O'Shotts television link was the first application of travelling-wave tubes in regular commercial service. Considerable experience since has confirmed the requirements foreseen then, with some modifications arising mainly from an increase in the number of telephone channels on each microwave carrier to 600 or more.

The principal factor determining the design of the output tube is the maximum power required. An output of about 5W is generally considered necessary for a 600 channel system, although this will depend on other system requirements such as aerial size, distance between stations and receiver noise factor. It is desirable for the travelling-wave tube to be capable of a rather greater maximum power output than this since the phase delay through the tube increases near saturation owing to the increased slowing-down of the electron beam as it delivers more power. This effect will convert any amplitude modulation on the input into phase modulation. The distortion produced by this effect will vary with the amount of amplitude limiting on the input of the tube. For a 5W operating level a maximum output of 7W to 10W is generally satisfactory and any considerably larger margin would greatly increase the size and expense of the travelling-wave tube and its power supplies.

The focusing of the electron beam is the major problem in obtaining the necessary power output, and has been fully discussed elsewhere[2]. To obtain 5W or 10W output from a travelling-wave tube at 4 000Mc/s the helix diameter will need to be considerably smaller than the cathode diameter if the cathode current density is about 200mA/cm^2, a figure which should give an adequate life from an oxide-coated cathode. This means a convergent electron gun must be used. It first appeared that the best way to obtain this in combination with magnetic focusing of the electron beam was to screen the cathode completely from the magnetic

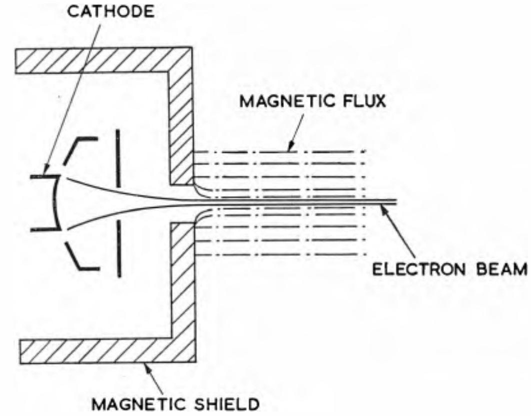


Fig. 1. Electron gun using 'Brillouin focusing'

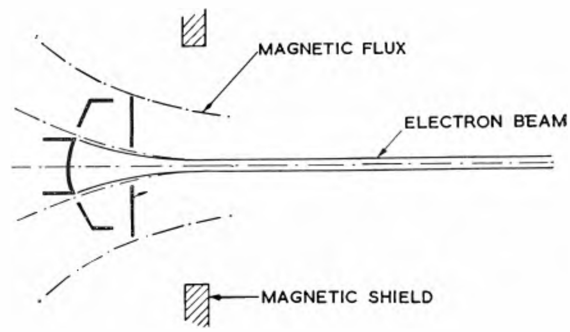


Fig. 2. Electron gun with shaped magnetic field

field, which rose suddenly to the required value near the point where the beam had its minimum diameter. A sketch of this 'Brillouin focusing' is shown in Fig. 1. Theoretically this requires the least field for focusing the beam but, in practice, operation at the theoretical field is rarely obtained and, of considerably more importance, the noise factor may be extremely high. The reasons for this behaviour are not fully understood but it has been found that more satisfactory operation can be obtained if a small magnetic field is present at the cathode. This field may be provided either by such methods as providing additional holes in the magnetic shield,

* *Standard Telephones & Cables Ltd.*

which will permit a small leakage of field into the cathode region, or by designing the magnetic shield to give a gradual increase in field which will not affect the convergence of the electron beam, as in Fig. 2. Both methods give similar focusing but the latter has the advantage that it sometimes makes it possible to bring the magnetic shield outside the tube envelope.

The magnetic field must extend the whole length of the tube in order to keep the beam focused as it travels through the helix. On early tubes this field was provided by solenoids, but their power dissipation and weight make it desirable to use a permanent magnet system if possible. A continuous magnetic field up to 600 gauss can be provided over a length of six or eight inches fairly readily. In order to reduce the bulk, material with a high $(BH)_{max}$ product is used, such as Alcomax III. Since the maximum working coercivity is around 600 gauss, higher fields require a change in shape to something like a horse-shoe magnet, with a corresponding increase in size. Even at this field the stray fields are very large, as the magnetic conductance inside the magnet and surrounding the electron beam is only a small fraction of that outside the circuit. These stray fields may affect other equipment and also mean that any magnetic material in them will affect the focusing field. Consequently, it is necessary to screen the circuit over a considerable volume. Any increase in the length over which the field is required gives a linear increase in all the dimensions and, consequently, an increase as the cube of the tube length in the volume and weight of the magnet.

An elegant solution which greatly reduces the stray magnetic field and changes the total magnet volume from a cubic to a linear dependance on the length of the tube is provided by periodic magnetic focusing. The focusing of the electron beam depends on the square of the magnetic field and hence it should not be affected by sudden reversals in the direction of the field. Perfectly sharp reversals are not possible since they would require pole-pieces of infinite permeability extending to the axis and Fig. 3 shows a typical practical assembly where the field has an approximately sinusoidal shape. If this period is near the wavelength of cyclotron oscillations on the electron beam a resonance effect will occur causing a progressive spreading of the beam with zero transmission. Partial focusing of the beam may be obtained at longer periods than that at which this 'stop-band' occurs, but this will involve electrons crossing the axis and depend critically on the shape of the field. For satisfactory operation it is necessary to use a considerably shorter period than that giving resonance and this is easier at higher voltages. The holes in the pole-pieces are comparable to the magnet length for these short periods and, consequently, the axial field is considerably lower than the coercivity of the magnet material. Even with ferrites a field equivalent to 600 gauss in the uniform case is near the maximum that can be obtained in practice. The reversal of successive magnets gives a very low net magneto-motive force between the ends of the circuit with very small stray fields and a considerable resistance to demagnetization. An increase in length requires only the addition of further periodic elements of the same size and not a corresponding increase in the magnet cross-sectional area.

The principal factors determining the power output and how it may be obtained have now been discussed. The other radio frequency characteristics of importance are the gain and the distortion which may occur, and these will be influenced mainly by the design of the helix and the central attenuating region. The gain necessary will depend on the output of the stage driving the travelling-wave tube, generally a high level mixer, and may vary from 1mW to 25mW depending on the system. There is little difficulty in making the helix long enough to give up to 40dB gain at 5W output but the less gain required the shorter the overall length of the tube can be.

One of the main causes of distortion in frequency-modulated radio links is time delay variation in long lines due to impedance mismatches. One such long line will exist between the mixer stage and the aerial, passing through the travelling-wave tube, and to prevent distortion the attenuation backward through the tube should be of the order of 30dB greater than the forward gain. This attenuation is provided by a central lossy region consisting of a resistive layer on the helix supports. A reverse attenuation of this magnitude is, of course, also entirely adequate to ensure that there is no chance of oscillation or instability caused by feedback.

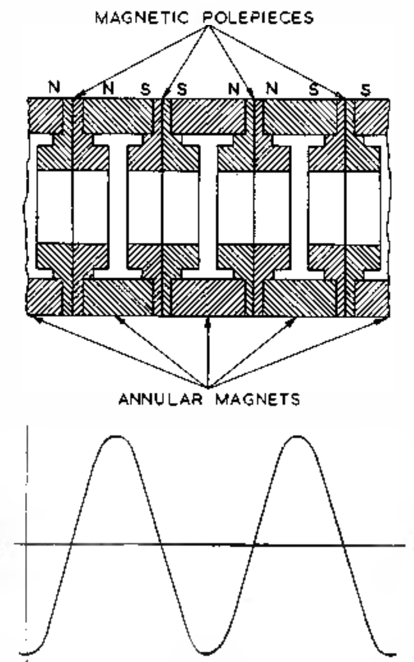


Fig. 3. Element of periodic magnetic circuit

Similar distortion can occur in the waveguide feeders at the input of the tube and between the output and the aerial. This distortion can be minimized by matching the tube very carefully over the required band. A trimming device is generally necessary to obtain a voltage reflection coefficient of less than 2 per cent at the centre frequency and, to keep this less than 5 per cent over a band of about $\pm$10Mc/s or more, the reflections on the helix itself must be eliminated as far as possible. Such reflections are amplified by the usual gain mechanism of the travelling-wave tube and, since they are a number of wavelengths away, cannot be cancelled over a wide band by trimming at the output. This means that the attenuating region described above must be tapered to provide a small reflection coefficient, but not tapered over too great a distance or the gain and efficiency will suffer, and also that any irregularities in helix pitch must be kept very small, since these too can give reflections. The turn-to-turn variation of pitch needs to be about 1 per cent or less. Periodic errors in particular must be avoided since these will give a poor response at those frequencies where the errors are an integral number of half-wavelengths apart.

The matching of both waveguide and coaxial line to the helix in travelling-wave tubes is now well established and although coaxial feeders are preferable for the fullest use of the inherent bandwidth of the travelling-wave tube they offer fewer advantages over the more restricted band, less than a quarter of the operating frequency, generally available

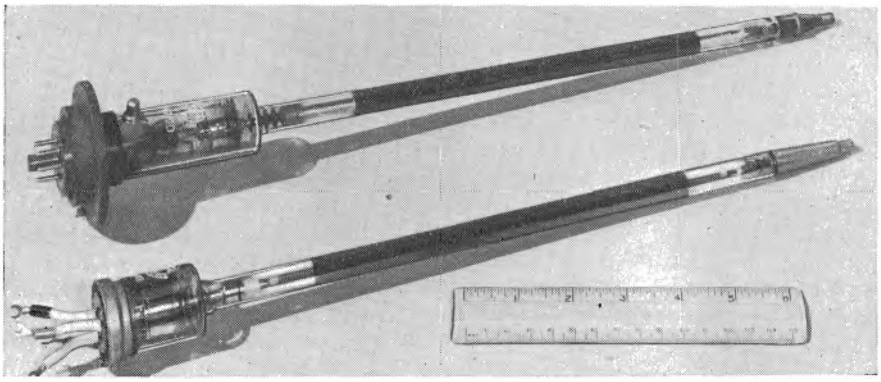


Fig. 4. CV2188 and W7/3G valves

for microwave communications. Since the filters, isolators and aerial feed are almost certain to use waveguide in high quality systems, it seems desirable to use this also for the travelling-wave tube, provided that no great increase in bulk is necessary. This is the case at 4000Mc/s, particularly if a slightly thinner guide than usual, of aspect ratio 3 : 1, can be used.

The user will not be content solely with the basic technical requirements specified above but must also be able to reproduce them economically and with reliability. This means that the tube should have a long life, be easy to set-up and replace, simple to monitor and not too expensive. As regards monitoring, the currents taken by the various electrodes and the voltages necessary are generally a sufficient indication of the satisfactory performance of the tube and also show when it will soon be necessary to replace the tube because of a deterioration in focusing or emission.

The application of these principles is illustrated in the W7/3G, developed from the CV2188. Full descriptions of both tubes have been published elsewhere[1, 3]. Photographs of them in Fig. 4, show the similarities and differences. Table 1 gives the typical average performance figures of both tubes at 4000Mc/s. The cathode area has been increased five times in the W7/3G so that despite the higher beam current the cathode density is considerably less. This has required the use of a convergent electron gun, the field shape in the electron-gun region being controlled by a pole-piece outside the tube. Thus the electron gun, although using ceramics instead of micas for alignment, is basically as simple as in the earlier tube and easily outgassed and processed. Developments in the winding of the helix and the distribution of the central attenuation give a better output match and the maximum gain and efficiency. A modified collector design is used to provide a greater mating area to the forced-air cooler to permit an increased power dissipation.

It is of considerable importance that the location of the valve in its circuit should be positive and central. In both tubes the cooler is held accurately central but allowed to pivot to permit minor alignment errors, one difference being that the cooler for the W7/3G is spring-loaded to the collector so that access to secure it is not normally necessary. At the gun end the CV2188 was held by a plastic flange cemented to the octal base after assembly but the W7/3G uses a flying lead base, to decrease the overall length, surrounded by a "Nilo K" ring. The ring locates the tube in the circuit and can be accurately aligned with the gun and helix during manufacture. Some slight adjustments to the focusing are desirable after the tube has been inserted in the circuit since it is worthwhile to work with the minimum helix current in every case despite variations in the exact alignment of different tubes and circuits. Deflector coils mounted around the electron gun provide a very convenient means of doing this, saving the complication of mechanical adjustments and the need to provide access to the circuit for these. The currents in these coils may be easily controlled from the most convenient point. The use of deflector coils in the W7/3G, despite the convergent electron gun, is possible since the magnetic screening is completely outside the tube and can surround the coils.

TABLE 1.
Typical Performance of CV2188 and W7/3G, (4000 Mc/s)

	CV2188	W7/3G
Helix voltage	3kV	3kV
First anode voltage	1·2kV	2kV
Collector current	14mA	40mA
Helix current	1·5mA	1mA
Maximum output power ..	2·5W	9W
Operating level	1·5W*	5W†
Attenuation	45dB	60dB
Band of output match to 5 per cent reflection	10Mc/s	30Mc/s

* *25mW input.* † *7 mW input*

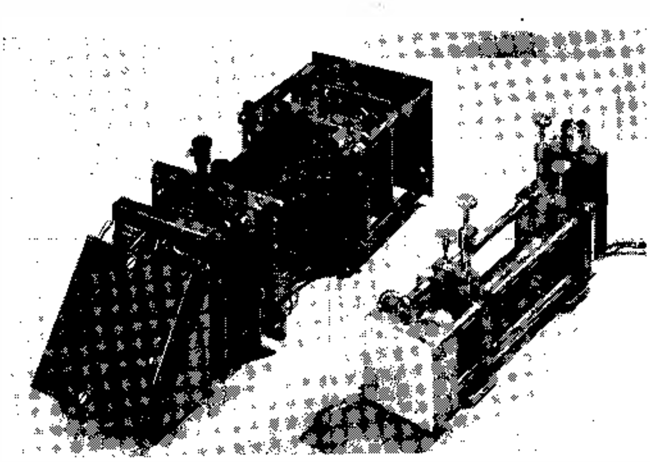


Fig. 5. CV2188 and W7/3G circuits

A separate first anode is used in both tubes to adjust the beam current to the nominal required value and a control circuit keeps this current constant despite any variations in emission, thus ensuring that there is no change in tube performance. The ordinary regulation of the mains supply keeps the other tube voltages sufficiently constant.

A periodic permanent magnet circuit has recently been developed for the W7/3G. Fig. 5 shows this circuit and also the solenoid used for the earlier CV2188. The gain and power output in a periodic circuit is about 1dB worse than that obtained in a corresponding solenoid circuit but this is compensated by the saving of 100W of solenoid power and 60 lb. of total weight. The periodic circuit is not affected by stray fields and it has considerable resistance to demagnetizing forces. Even touching the circuit with iron tools has no permanent effect on it. One substantial difference arises from the fact that focusing is not satisfactory below a certain voltage in a periodic circuit and, consequently, the full voltage must be switched on immediately.

Waveguide input and output is used on both tubes, the internal dimensions being 2 in. × 0·667 in. A flag-tuner in the guide can be adjusted to give a satisfactory match at the operating frequency.

There are other possible approaches to the design of the output travelling-wave tube. One such is the use of a higher perveance beam than that described here leading to a somewhat shorter tube with a continuous field provided by a permanent magnet. Periodic focusing is not usually possible in this case owing to the lack of permanent magnet materials of sufficiently high coercivity. The bulk when a permanent magnet is used is generally greater than that required by periodic focusing, particularly because of the shielding requirements. Moreover the high perveance gun may require screening inside the envelope, which makes the tube more expensive and also some mechanical movement necessary, since deflector coils outside the tube would be ineffective.

Another possibility is the supply of tubes complete with their circuits so that no adjustment of focusing or matching is necessary when changing tubes. In the author's experience, however, operators very quickly become accustomed to the routine for setting-up a travelling-wave tube and the supply of 'packaged tubes', with the added requirement of breaking and making the coaxial or waveguide connexions, would offer few advantages. Against them must be set the added cost of such tubes, particularly in multi-channel systems where separate stocks of spares would have to be kept for each frequency used.

Extensive experience with travelling-wave tubes in systems installed in many parts of the world has confirmed their reliability and ease of operation. Continued development of the W7/2D has suggested minor modifications which have greatly improved the life compared with the first tubes used in the Manchester-Kirk O'Shotts link. For example, results from one recent system indicate an average life of well over 9300 hours.

Low Noise Tubes

The system designer has had no readily available alternative to the travelling-wave tube for the output stage of radio relay links; for the input stage however, the travelling-wave tube must compete with the crystal mixer which has been very extensively developed. It is only in more recent years that better noise factors than those given by crystal mixers have been obtained from travelling-wave tubes and even so the disadvantages of the increased bulk and power supply requirements of the travelling-wave tube may outweigh the advantages of its improved performance. Nevertheless it has been necessary to develop low-noise tubes to decide this question and further improvements, either in noise factor or by reducing the size, may make their use more general.

The fundamental cause of noise in a travelling-wave tube is the current and velocity fluctuations of the electron beam which arise from its emission from a hot cathode. To obtain a low noise factor it is necessary to reduce this noise relative to the input signal and this can be done by appropriate acceleration of the electron beam in the gun region before it enters the helix. A complete theoretical treatment of the problems involved is not yet possible, but by making some plausible assumptions the basis of a design can be arrived at and improved by experiment. The electron gun which results uses a small cathode with a focusing electrode giving a slightly divergent beam, a first anode to control the current and one or more anodes before the final one at helix potential. These adjust the relative magnitude of the fluctuations arising from the velocity and the current variations at the potential minimum in front of the cathode. A drift tube, following the final anode of the gun, preceeds the helix and in this the current fluctuations tend to change into velocity fluctuations and vice versa due to space charge forces. The length of this drift space is chosen to give the optimum phase relations for minimum noise. The anode voltages may be adjusted for minimum noise factor but varying one alone will slightly affect the optimum drift tube length. Hence, if one can vary the potential of several anodes, and the focusing electrode as well, a better noise factor is likely to be achieved, but the setting-up for this, due to the interaction between the electrode potentials, will be considerably more complicated.

It is also essential to have very good focusing since any intercepted current will lead to partition noise. As a general rule, the intercepted current should be less than 0·5 per cent of the total current. This, with the electron gun considerations, forces one to work with small beam currents and, consequently, even at low voltages, the gain per unit length is low.

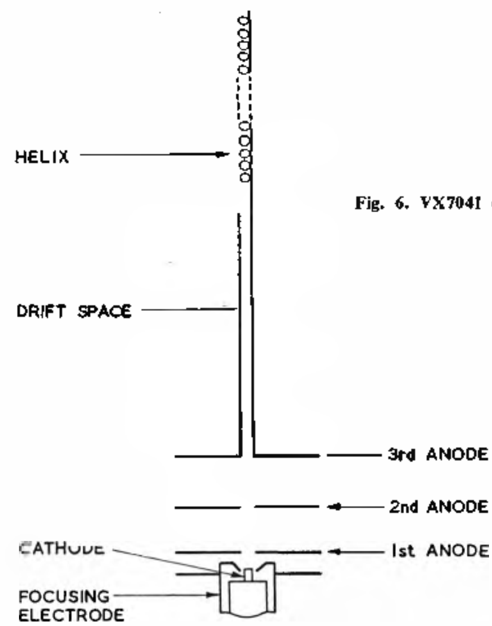


Fig. 6. VX7041 electron gun

A typical low-noise travelling-wave tube, developed with the possibility of use in microwave radio links in mind, is the VX7041*4. Fig. 6 shows the basic elements of the electron gun. While the principles outlined above are employed, the construction has been made as simple as possible to use. The focusing electrode works at cathode potential and in addition to the current controlling anode and that at helix voltage, only one further noise-reducing anode is used. Fig. 7 shows the complete tube and circuit. Apart from the use of a low-noise electron gun, and a smaller helix for operation at 900V, the construction is very similar to that described for the W7/3G. Table 2 gives some typical performance figures. A rather high magnetic field, for this current and voltage, is necessary to give the very good focusing.

In operation the tube fits into the circuit in a similar way to the W7/3G and the focusing is adjusted solely by deflector coils, no mechanical movement being necessary. Adjustments are then made to the first anode voltage to give the required collector current, the helix voltage to give maximum gain and the second anode voltage to give minimum noise output. All these adjustments are substantially independant and are easily and quickly carried out. Experience has shown that the conditions remain stable with life and that any deterioration in noise factor is due to a fall in emission which will also require an increase in the first anode voltage needed to give the usual collector current. Thus, after the first setting-

* *The development of this tube was supported by the Admiralty.*

up, the d.c. currents and potentials are a very reliable indication of a tube's satisfactory operation.

The gain should be sufficient for the noise in succeeding stages to make only a small contribution to the total receiver noise factor. For a travelling-wave tube of 8dB noise factor, 20dB gain is adequate to ensure this. Time delay distortion will arise in the low-noise stage unless precautions similar to those in the case of the output tube are taken. The VX7041 uses identical tuning means to the W7/3G and attention to the total attenuation, the shape of the taper and the regularity of the helix enables a similar specification to be met.

The output level in any travelling-wave tube in a microwave link should never approach the saturation power output of the tube or distortion will arise due to the conversion of amplitude to phase modulation. The maximum output of the VX7041 is about 10mW which easily satisfies this condition.

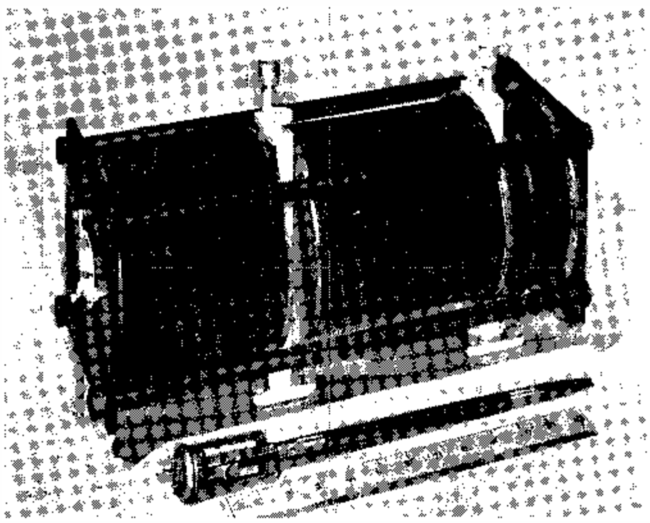


Fig. 7. VX7041 tube and circuit

The solenoid used for the VX7041 is rather bulky and consumes a considerable amount of power. Unfortunately periodic focusing cannot at present be used without a loss in performance. The period at this relatively low voltage must be very short and the ferrite materials at present available do not make it possible to obtain, with this short period, a high enough field to give the very good focusing required. Again the total length of helix and drift tube means that a very bulky magnet would be necessary to give a uniform field. This additional bulk militates against the use of low-noise tubes in microwave radio links under ordinary conditions.

Intermediate Level Tubes

The design of tubes with an intermediate power output, of the order of 100mW is rather simpler than in the case of high power tubes or low-noise tubes since the problems of focusing high perveance beams or of obtaining very low noise factors, the fundamentals of both of which are not fully understood, do not arise so acutely. On the other hand less time has been devoted to the design of such tubes for microwave radio links because of the apparently limited field of application of all travelling-wave tube repeaters, which would be their principal use.

TABLE 2.
Typical Performance of VX7041 at 4 000Mc/s.

Helix voltage	900V
Collector voltage	1kV
First anode voltage	50V
Second anode voltage	200V
Collector current	0·5mA
Helix current	2μA
Gain	20dB
Noise factor	8dB
Attenuation	55dB
Bandwidth of input match to 5 per cent reflection	30Mc/s

The basic features of ease of operation and lack of distortion can be met substantially as described in connexion with other types of tubes. An intermediate level tube can, however, have a much greater gain per unit length than either the low noise tube, necessarily restricted to a low current, or the output tube, which requires a fairly high voltage. The size can then be considerably smaller and permanent magnet focusing presents few difficulties whether periodic, if the voltage is around 1kV, or a continuous field for lower voltages.

An early tube[5] suitable for this application is the CV2358. This has a gain of 25dB, a maximum output about 150mW, and operates at 1·5kV, 5mA. Once again the tube plugs into its circuit and the focusing is adjusted by deflector coils. Using more modern techniques and the construction of the W7/3G, a higher gain and lower voltage operation could readily be obtained.

The CV2358 has been extensively used by the G.P.O. as a frequency changer[6] and this application is important in an all travelling-wave tube repeater. If the helix voltage is modulated and the tube is driven near saturation with the microwave frequency signal, sideband outputs are obtained about 6dB down on the maximum output power. This is a very convenient level for driving an output travelling-wave tube. In practice, the modulation is applied to the cathode which has a lower capacitance to earth than the helix.

Conclusions

Travelling-wave tubes are very well established as output amplifiers in 4 000Mc/s radio links and while economic considerations do not generally justify their use elsewhere in such systems future development may modify these conclusions. For high capacity links at 6 000Mc/s they are again certain to be used and the smaller bulk consequent on the higher frequency will be a considerable attraction. It is likely that even at 2 000Mc/s the ease of setting-up may make them attractive in radio link equipment. At very high frequencies however, above 11 000Mc/s, considerable problems arise in the development of tubes to give appreciably more than 1W output and the use of oscillators directly, where possible, is likely to be more economical.

A microwave radio channel does not generally extend over more than 20Mc/s, which is about 2 per cent or less of the possible bandwidth of travelling wave tubes. It would be particularly pleasing for the valve designer if other systems engineers, encouraged by this practical success in radio links, were to take full advantage of this potential bandwidth.

Acknowledgments

The author is grateful to Standard Telephones and Cables Ltd. for permission to publish this article.

REFERENCES

1. Rogers, D. C. The travelling-wave tube as output amplifier in microwave radio links. *Proc. Instn. Elect. Engrs. 100*, Pt. 3. 151 (1953).
2. Burke, P. F. C., Rogers, D. C. Magnetic focusing of long cylindrical electron beams. *L'Onde Electriqué. 37*, 174 (1957).
3. Burke, P. F. C. A Travelling-wave tube for 4 000Mc/s Radio Links. Paper presented at I.E.E. Convention on Microwave Tubes, May 1958.
4. Burke, P. F. C., Pohl, W. J. A 4 000Mc/s low-noise travelling-wave tube. *Le Vide 11*, 303 (1956).
5. Rogers, D. C. Travelling-wave amplifier for 6 to 8 cms. *Elect. Commun. 26*, 144 (1949).
6. Bray, W. J. The travelling-wave tube as a microwave phase modulator and frequency shifter. *Proc. Instn. Elect. Engrs. 99*, Pt. 3. 15 (1952).

REFLEX KLYSTRONS

By A. H. Atherton*, B.Sc., A.M.I.E.E.

A general description of the construction and operation of reflex klystrons is given, so that the characteristics and performance of this type of valve can be appreciated by the systems engineer. Typical examples of reflex klystrons are described, and brief details are given of practical operation.

AT very high frequencies, and especially in the microwave band above 2 000Mc/s, the use of negative grid valves as oscillators becomes increasingly difficult, mainly due to the necessity for keeping the cathode to grid transit time to below a small fraction of one cycle of the operating frequency. In a normal triode oscillator, amplitude modulation is imposed on the electron beam by means of the negatively biased control grid, and the beam is, therefore, modulated while it has a velocity corresponding to an energy of 1 to 2eV. With such a low electron velocity, the cathode to grid spacing cannot be made small enough to give a low transit time at high frequencies. In the reflex klystron a high speed electron beam, with a velocity corresponding to perhaps 300eV, is velocity modulated by the r.f. field. It is this higher beam velocity, resulting in a lower transit time, which enables the reflex klystron to operate efficiently with reasonable physical spacings.

An upper frequency limit, at present about 35 000Mc/s, beyond which reflex klystrons cannot be efficiently manufactured, is set by the physical dimensions of the resonant cavity, which must be smaller at higher frequencies. The useful range of reflex klystrons may be said then to lie between 2 000 and 35 000Mc/s; within this range, the reflex klystron is unsurpassed as a convenient low-power oscillator which can be both mechanically and electrically tuned.

General Description

In this article, no attempt will be made to give a mathematical theory of reflex klystrons, since this is already available in the literature[1,2]. Instead, a descriptive account of the operation of reflex klystrons will be given, which should enable the user of these valves to understand their operating characteristics and properties.

The essential electrodes of a reflex klystron are shown in Fig. 1, and consist of a resonator, an electron gun, and a reflector. The resonator, which is the r.f. circuit, is a re-entrant cylindrical structure with physical dimensions such that it will resonate at the desired operating frequency; in terms of the normal LC resonant circuit, the inner gap in the resonator may be said to be capacitive, the outer part of the resonator being inductive. The essential features of the resonator are that it is a high-Q, low-loss resonant circuit, which has a high r.f. field at the gap where the electron beam is modulated. Energy from the resonator may be extracted either by an inductive loop coupled to a concentric line, or by an iris coupled directly to a waveguide. The electron gun provides a beam of electrons which is focused through the resonator aperture, passes through the gap and is reflected by the negative field between the reflector and the resonator; the beam shape being such that a high proportion of the original beam is transmitted back through the gap.

In the first passage through the resonator gap, the interaction between the r.f. field and the electrons causes a velocity modulation of the beam, electrons passing through the gap when the field is positive being accelerated, those passing through when the field is negative being decelerated. After leaving the resonator, the different relative velocities of the electrons in the beam result in bunches of electrons being formed so that the velocity modulation is converted into amplitude modulation. If the beam is reflected back through the gap and the phase conditions are such that the bunches are retarded by the r.f. field, energy will be given up to the field and the valve can oscillate.

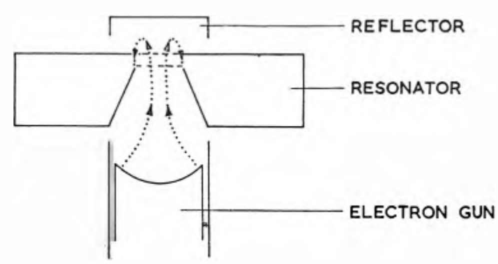


Fig. 1. Essential electrodes of a reflex klystron

Two conditions must be satisfied before the valve will oscillate with maximum efficiency. First, the time taken for the beam to travel from the gap, into the reflector field, and back again (the transit time) must be such that the r.f. field exerts a maximum retardation on the bunches in order to extract as much energy as possible, and it can be shown that the transit time must be $n + \frac{3}{4}$ cycles, where n is 1, 2, 3, 4, . . . The transit time can be varied by adjusting the reflector potential; making the reflector more negative increases the field and reduces the transit time. Since n is multi-valued, the valve will oscillate for several values of reflector potential, the transit time being $1\frac{3}{4}$, $2\frac{3}{4}$, $3\frac{3}{4}$, . . . cycles. A valve oscillating with any particular value of n is said to be operating in the $n\frac{3}{4}$ mode, the valve operating in each mode over a range of reflector potentials which cause the valve to start oscillating, rise to a maximum as the phase condition is satisfied, and then fall off again to zero. As the reflector potential is increased from zero to a high negative value, the power output characteristic is, therefore, as shown in Fig. 2. In setting up a reflex klystron, the reflector potential must be adjusted so that the valve is operated in the chosen mode, and the potential required will depend not only on the mode but on the frequency, since the transit time will vary with frequency.

The second condition that must be satisfied for maximum

* *Research Laboratories of Electric & Musical Industries Ltd.*

power output is that the bunching must be as efficient as possible, i.e. the current density in the bunches must be as high as possible. At any particular time after modulation, the degree of bunching will depend on the relative velocities of the electrons in the beam (the degree of velocity modulation). Since the transit time is determined by the condition that the phase-angle of the returning beam shall be correct, optimum bunching can only be realized by adjusting the degree of velocity modulation to some suitable value. The degree of velocity modulation depends on the r.f. field in the resonator, so the bunching can be adjusted to the optimum value by varying the r.f. field. In practice, this is done by adjusting the coupling to the load until the power output is a maximum; increasing the load decreases the effective resonator impedance and reduces the r.f. field. Since the valve oscillates most efficiently with a high r.f. field at the gap, requiring a low bunching time, the valve is usually more efficient for lower reflector modes, as is seen in Fig. 2, where the peak power in each mode decreases as the mode number increases.

The bunched electron beam corresponds to the amplitude modulated beam of a negative grid oscillator, but it should be noted that in a reflex klystron the amplitude distribution of charge along the beam does not follow a sine or cosine law. A detailed mathematical treatment shows that the modulated beam contains the fundamental frequency and an infinite number of harmonics, the amplitude coefficients being given by a Bessel function of the first kind. Of these, only the fundamental is of importance, since the r.f. losses in the resonator are too great for the harmonics to be sustained.

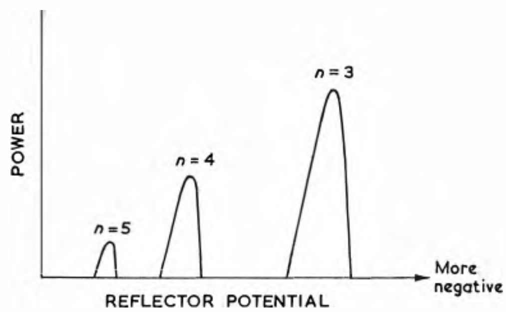


Fig. 2. Power output—reflector potential characteristic

Tuning

Mechanical Tuning

To change the frequency of oscillation, the resonant frequency of the resonator must be changed, and this can be done in two ways. In the first, the capacitance at the resonator gap is varied by physically moving the two ends of the cavity apart. The tuning rate is very high, typically 500Mc/s for 0·001in movement of the gap, so that a very accurate slow motion mechanism is required. The tuning mechanism should not show hysteresis (i.e. after traversing the full tuning range the frequency should return to its original value as the tuner is returned to its original position); it should be rugged, non-microphonic, insensitive to temperature or atmospheric pressure changes, and preferably inexpensive to manufacture. With all these requirements in mind, it is not surprising that the design of this type of tuning mechanism is one of the most difficult parts of the design of oscillators, and has probably not yet been satisfactorily solved. The second method of tuning is to alter the inductance of the resonator, either by changing the effective diameter or by introducing material of a different dielectric constant. Many early klystrons had large diameter screws which entered the resonator wall radially to change the effective diameter, and wide-range tuning devices have been produced in which part of the outer walls of the resonator are formed of movable arms, the lack of circular symmetry apparently having little effect on the electrical performance of the valve. A third method of mechanical tuning is to tune a close-coupled external resonator and so pull the frequency of the klystron cavity. Since the external resonator design is not restricted by considerations of beam coupling, and since it is not in vacuum, the mechanical problems are much less difficult than the gap capacitance tuner to which it is essentially similar. Some loss of valve performance is to be expected, but the improvement in tuner properties, especially in simplicity and ruggedness, is so great that the sacrifice of electrical properties is worth while.

In general, the wider the tuning range demanded from a valve, the more difficult, and expensive, will be the design of the tuner. Except for bench oscillators, where a wide frequency range is obviously necessary, the tuning range demanded from the valve should be no greater than is required for the particular application.

Electronic Tuning

If the transit time in the reflector space is varied slightly from the optimum, by means of a change in reflector potential, the bunches of electrons will no longer return in the correct phase and the beam may be said to have a reactive component. Since the frequency of oscillation is determined by the condition that the total reactance of the beam and cavity must be zero, the frequency of oscillation must change. This frequency change results in some amplitude modulation, because of the increased resonator losses, but tuning ranges of the order of $\frac{1}{2}$ per cent can be achieved between the points where the output falls to 50 per cent of its peak value (the so-called half power points). This electronic tuning of the reflex klystron is one of its most useful characteristics, since the tuning is effected by the high impedance reflector electrode, and low modulating powers are required.

The electronic tuning range varies with the cavity loading and with the reflector mode; in general, the higher the loading and the higher the mode, the greater the electronic tuning. Since the valve efficiency is normally less at the higher reflector modes, some compromise in design is necessary between power and electronic tuning.

Hysteresis

One of the problems encountered in the use of electronic tuning is that of hysteresis, i.e. the multiple dependence of power and frequency on reflector voltage. A typical example is shown in Fig. 3. As the reflector voltage is increased, the power gradually falls off to zero; as the reflector voltage is then reduced, the power curve does not follow the original curve, but suddenly jumps into oscillation as shown in the figure. The frequency may vary continuously or may have discrete jumps, depending on the cause of the hysteresis.

One of the worst causes of hysteresis, multiple electron transits through the resonator gap, is well understood, and is under the control of the valve designer. In a well designed valve, any hysteresis present should not approach the half power points, and hence will not affect the practical performance of the valve. Other causes of hysteresis, however, are under the control of the valve user. If the valve is mismatched into the guide, or if a mismatched load is used, especially if the line is of considerable length, hysteresis or frequency jumping can be expected. A more troublesome cause of hysteresis arises where the resonator can support more than one mode of oscillation (i.e. is resonant at more than one frequency), and the reflector potential for both modes is nearly the same. The transitions from one mode to the other will occur at different points, depending on the direction of change of reflector potential. The effect is more confusing since the receiver may not be sensitive to the frequency of one of the modes, and the observed effect can then easily be confused with multiple transit hysteresis. If the cavity design cannot be modified to eliminate the unwanted mode, some form of mode suppression must be used, which will absorb energy from the unwanted mode, but not from the required mode.

Klystron Design

TYPE OF RESONATOR

The resonator of a reflex klystron may be totally enclosed within the valve envelope (internal resonator), the power being extracted through a vacuum sealed concentric line or through an insulated waveguide output window. Alternatively, the upper and lower walls of the resonator

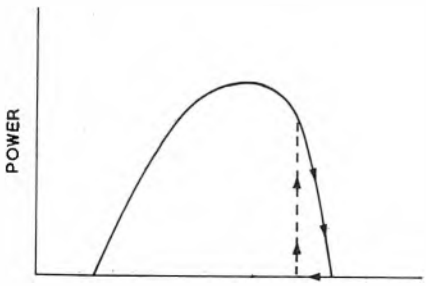


Fig. 3. Typical example of hysteresis

may be sealed through the envelope (disk seal construction), so that part of the resonator is external to the valve and is available for tuning and power extraction.

The advantages of the totally enclosed resonator are that it has less r.f. loss, and the mechanical design of the valve is simplified, particularly where the all-metal valve construction is used; for very high frequency valves where the resonator is very small, this is the only possible construction. The main disadvantage is the difficulty of extracting power from the resonator. The use of a concentric line output is not entirely satisfactory, since it is not possible to control the coupling accurately, with the result that provision must be made for matching the valve into the waveguide to achieve optimum operating conditions. With a waveguide output, the matching from valve to valve can be held quite accurately, and is set by the valve designer so that, when feeding into a correctly terminated waveguide, the valve is working under optimum conditions. This is particularly important in military applications, where adjustments must be reduced to a minimum.

If the disk seal construction is used, the problems of power extraction and tuning are greatly simplified; the main difficulty is of r.f. loss in the part of the glass envelope that lies within the cavity. One method of reducing this is to operate the cavity in a harmonic node, such that the glass is located at a voltage mode. This technique has the additional merit of increasing the diameter of the cavity, so that the physical dimensions of the envelope need not be too small. It may, however, still cause some loss of efficiency, and, since a harmonic cavity can support several modes, it may introduce the possibility of mode interference in the tuning characteristic.

The use of a disk-seal valve makes possible the design of plug-in valves, in which the external part of the cavity is not integral with the valve, but is considered as part of the microwave 'plumbing'. In the valve, the resonator diaphragms are formed of two copper pressings sealed into the glass envelope, and are suitably shaped to enable contact to be made to the external cavity. This technique of making the valve a plug-in replacement, independent of the main cavity, has several advantages:

(a) The resonator and tuning mechanism can be designed for the particular application required, and hence can be designed more efficiently.

(b) Considerable economy results from the fact that valve replacement does not involve replacement of the expensive cavity and tuning mechanism.

(c) One valve will cover a wide variety of applications, leading to more economical manufacture and stores maintenance.

On the other hand, the valve must be tested by the manufacturer to much closer limits, since it may be required to operate in many different types of cavity.

Typical Reflex Klystrons

A list of reflex klystrons of British manufacture has been published[3], and no attempt is, therefore, made to summarize the characteristics of the many types available. Instead, three valves will be described, which will illustrate the electrical characteristics and different methods of physical construction described earlier. The R5146 is an internal cavity valve; the R5222 is a disk-seal plug-in valve, both designed for use in radar receivers and for general bench work. The third valve, the R6010, was designed for use as the transmitting valve in an f.m. centimetric radio link, and is an example of a high-power oscillator in which particular attention has been paid to the linearity of the electronic tuning characteristic.

VALVE TYPE R5146

The R5146 was designed for the 8·5mm microwave band, and is an example of a valve design suitable for the highest frequency for which reflex klystrons are available. The essential electrode structure of the valve is shown in Fig. 4; the complete valve is shown in Fig. 5.

The resonator cavity, which is approximately 3mm in diameter and 1mm high, has a 0·020in aperture through which the electron beam must be focused and returned after reflection. If the aperture is appreciably larger than this, the coupling between the resonator and the beam is reduced, with a corresponding loss of efficiency. The size of the resonator aperture sets a lower limit to the resonator potential required for efficient operation. As the potential is decreased, not only is the coupling between the beam and the r.f. field decreased, but less of the initial beam returns through the gap because of mutual electron repulsion in the beam (space charge spreading). In the R5146, the resonator potential is 2kV, a 10mA beam being used.

Power is extracted through a slot in the outer wall of the resonator, a tapered waveguide providing a smooth transition to standard Q-band guide. The problem of transmitting the power through the metal valve envelope is solved by the use of a glass window designed so that a standing wave ratio of better than 1·2 is achieved over the operating frequency range; ditched chokes are used to prevent leakage of power where the internal and external waveguides face the window. The valve is tuned by changing the gap spacing mechanically. An external screw and lever mechanism is connected, through a flexible diaphragm at the top of the valve envelope, to the reflector support which, in turn, is connected to the flexible upper surface

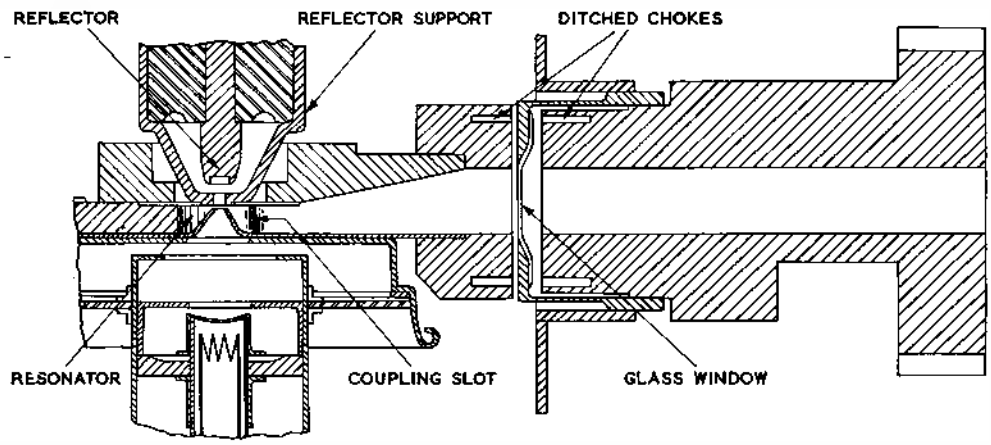
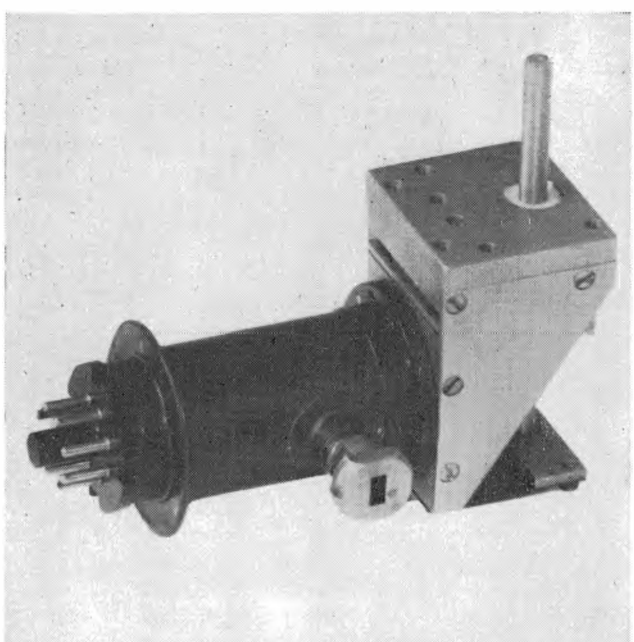


Fig. 4. Essential electrode structure of valve type R5146

Fig. 5. Photograph of valve type VX5023



of the cavity. The coupling between the cavity and the output waveguide is set by the dimensions of the output slot, and is designed so that, when the output power is fed into correctly terminated Q-band waveguide, the valve is adequately matched over the operating frequency range. The only adjustment required in setting up the valve is, therefore, to set the frequency and to adjust the reflector potential for the optimum power. Two reflector modes are normally used (n = 4 and 5). In the $4\frac{3}{4}$ mode the output power is typically 50mW, the electronic tuning range 60Mc/s, and the reflector potential −350V. In the $5\frac{3}{4}$ mode, the power is typically 30mW, the electronic tuning range 90Mc/s, and the reflector potential −225V. The mechanical tuning range is approximately 15 per cent of the nominal frequency. (To cover a wider frequency range, modifications to the cavity design must be made, and a range of valves is available covering 7·5 to 12·5mm.) Over the tuning range, the reflector potential, in the $5\frac{3}{4}$ mode, changes from −190 to −250V in a linear manner, while the electronic tuning rate is of the order of 2Mc/s/V. It should be noted that all voltages are measured with respect to cathode, although in normal operation it is found more convenient to operate the resonator at earth potential.

Valve Type R5222

The R5222 is a 3·2cm (X-band) general purpose reflex klystron with an external resonator cavity. The physical dimensions of of an X-band resonator are sufficiently great to enable a relatively large resonator aperture to be used, the beam coupling being kept high by the use of grids across the two resonator apertures. Because of the large apertures, a low impedance electron beam can be used, 30mA at 300V, so that d.c. supplies for the valve can conveniently be supplied from normal power packs. The power output from the valve, operating in the $4\frac{3}{4}$ mode, is typically 50mW, with an electronic tuning range of 15Mc/s; the tuning sensitivity being 0·45Mc/s/V. By a modification to the design of the cavity, the electronic tuning range may be increased to 30Mc/s, with little or no loss of power, although the mechanical tuning range is restricted. The nominal frequency range of the valve, in suitable cavities, is 8 500 to 10 000Mc/s although, experimentally, the valve has oscillated from 5 000 to 12 000Mc/s.

Special efforts have been made to ensure that the electrical characteristics are reproducible from one valve to another. The limits within which 90 per cent of all valves would be expected to fall are approximately as follows:

Power		±11mW
Frequency	(in standard cavity)	±33Mc/s
Reflector potential		±16V
Electronic tuning rate		±0·08Mc/s

Control of characteristics to such fine limits means that very little adjustment need be made when a valve is changed in an equipment. Ideally, no adjustment should be necessary, and efforts are continually being made, both in valve design and in accuracy of manufacture, to reduce the spread in electrical characteristics.

The R5222 is a plug-in valve, and hence must be inserted into a suitable cavity before it will oscillate. A range of cavities has been designed for the valve, but it is anticipated that the user will, in general, want to design special cavities for specific applications. A fuller account of cavity design for this valve is available[4], and only a brief outline of the possible types of cavity will be given here.

The lower conical resonator electrode (or 'copper') is, in use, clamped between matching surfaces, so that a low-loss r.f. connexion is made, and, at the same time, the valve is located in the external cavity. Contact to the upper copper is made by specially designed spring fingers. The external cavity may take the form of a concentric line or a radial line, the length of the line being $(2k + 1)\,\lambda/4$ where k is 0, 1, 2, 3, . . . The choice of the value of k used will depend on the frequency of operation required, and on the physical limitations of the cavity design. In general, k should be as low as possible, to maintain valve efficiency and reduce the possibility of mode interference. At the centre of the operating frequency range, the valve is designed to operate in a harmonic cavity, with $k = 1$, the voltage node being located near the glass envelope to prevent r.f. losses.

The radial line cavity is less suitable for very wide tuning ranges than a concentric line cavity, but is simpler mechani-

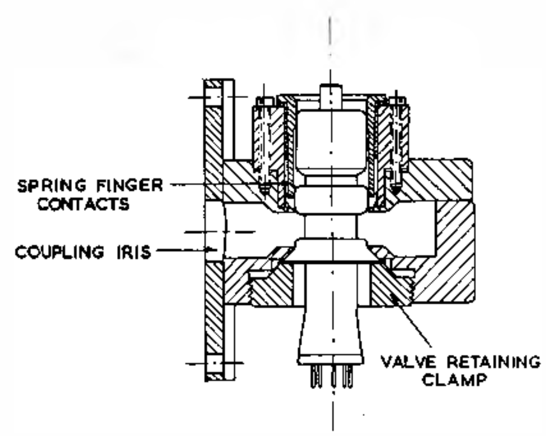


Fig. 6. Fixed frequency cavity for valve type R5222

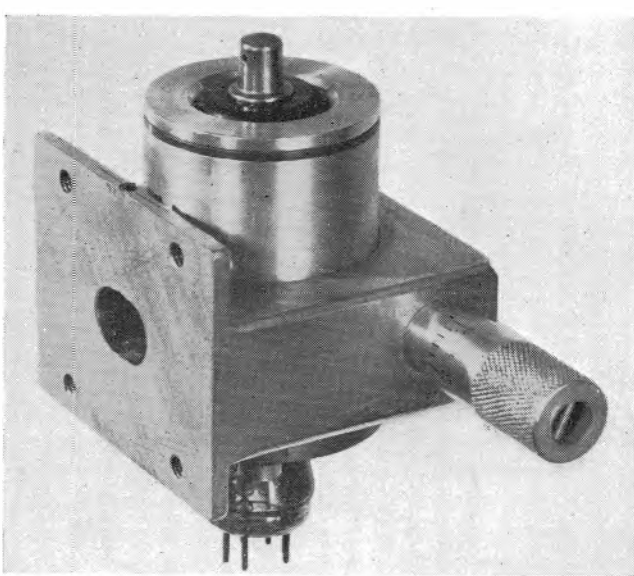


Fig. 7. Tunable cavity for valve type R5222

cally and easier to design. A simple fixed frequency cavity is shown in Fig. 6. The clamping piece for the conical copper and the spring fingers for the cylindrical copper can clearly be seen, together with the coupling iris and waveguide output. A simple tunable cavity is shown in Fig. 7. The tuning is effected by inserting a ceramic rod into the resonator, the higher dielectric constant of the rod causing a reduction in frequency. A micrometer screw thread gives a precise and simple tuning control. A tunable cavity for wider frequency bands is shown in Fig. 8. The two movable side arms are of the non-contacting type, choked to prevent losses. Up to 20 per cent tuning range can be obtained with this type of tuner, which is very suitable for general bench measurements. With the concentric line cavity, the length of the line is varied by means of a non-contacting piston, and the resonant frequency of the cavity can be varied over very wide limits. Where extremely wide tuning ranges are essential, concentric line cavities are to be preferred, but their main disadvantages are the increased difficulty in arranging coupling to a waveguide, the greater possibility of mode interference, and the increased mechanical complexity.

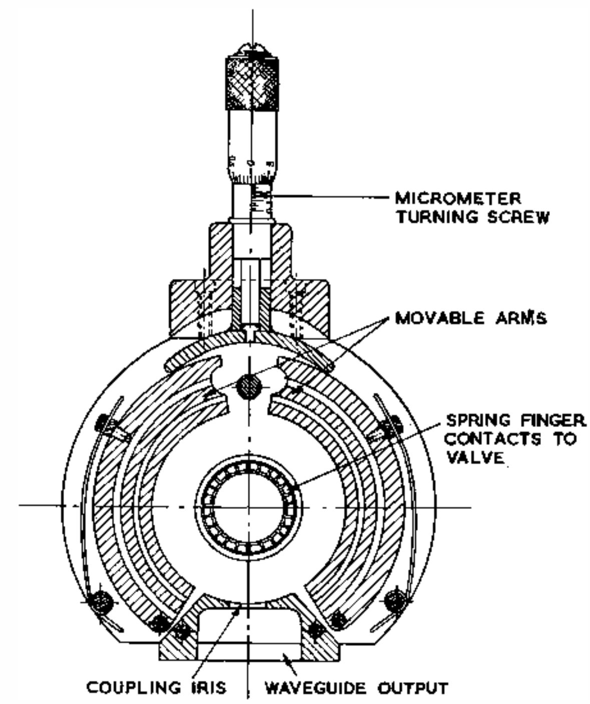


Fig. 8. Wide-band tunable cavity for valve type R5222

In designing a cavity for a plug-in valve such as the R5222, the following rules should be observed:

(a) Choose a symmetrical structure.

(b) Use the simplest approximation to a resonant line, k being kept as low as possible.

(c) Use the simplest possible form of tuner, with as small a tuning range as possible.

(d) Avoid unnecessary r.f. losses, both in the cavity and the contacts.

Valve Type R6010

The R6010 was designed specifically for use as an f.m. transmitting valve, where output powers of the order of 5W are required, and where the reflector voltage-frequency characteristic must be linear in order to prevent distortion and inter-channel modulation. A cross-section of the valve is shown in Fig. 9. The resonator must dissipate 100W of power and is, therefore, made an integral part of the valve envelope to facilitate cooling. The use of a harmonic resonator would impair efficiency and electronic tuning range, so the disadvantages of an internal resonator must be accepted. Power is extracted via a loop and concentric line, although a later version of the valve, for a somewhat higher frequency, has been developed with a waveguide output. The earlier electron optical design of gun and reflector allowed some multiple electron transits, so the hysteresis was present; this has now been corrected by a re-design of the gun. The valve is capacitance tuned, a differential screw mechanism providing the slow motion drive required to change the resonator gap.

Typical electrical characteristics of the valve are as follows:

Mechanical tuning range	4 400 to 4 800Mc/s
Resonator potential	700V
Resonator current	140mA
Reflector potential	–275V
Electronic tuning range	50Mc/s
Tuning rate	0·25Mc/s/V

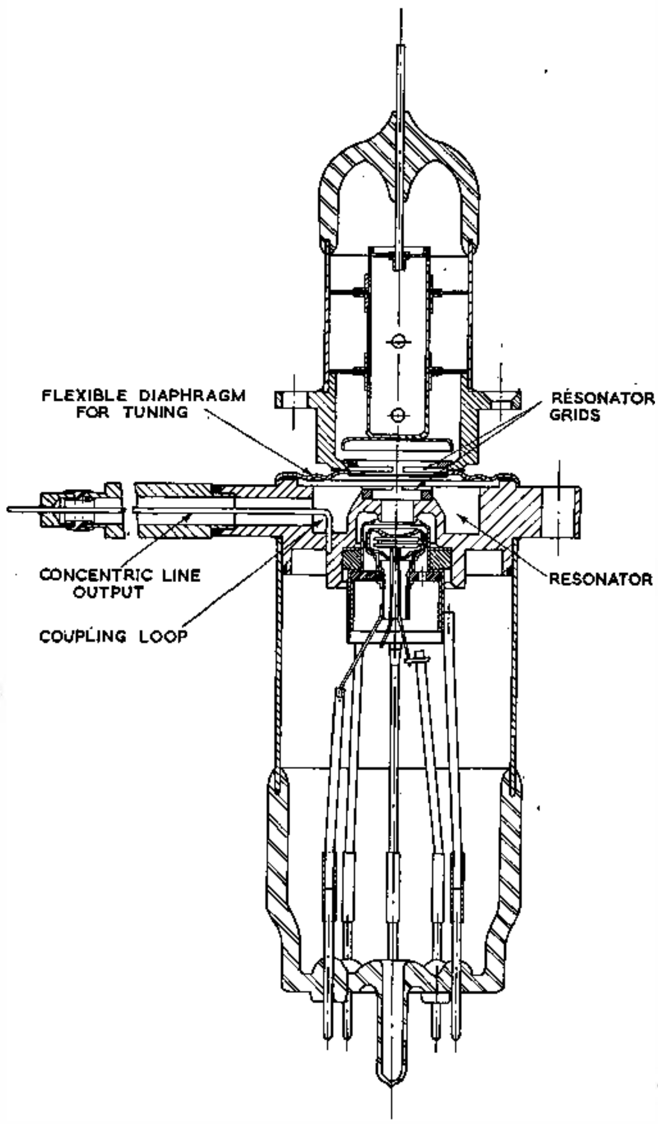


Fig. 9. Cross-section of valve type R6010

The frequency range, over which the tuning rate varies by 1·2/1 is 15Mc/s. The linearity of the tuning characteristic is dependent on the matching into the guide, the standing wave ratio of the load, and on which part of the reflector characteristic is used. Since the optimum linearity is not always achieved under the conditions for maximum power output, some slight sacrifice of power may be necessary if the utmost linearity is required.

Practical Operation

Power Supplies

For general microwave bench testing, it is convenient to have a power pack which is sufficiently flexible to be useful for a wide range of valves. The following supplies are suggested:

Heater.	4 to 8V a.c. continuously variable, rated at 1·5A. Transformer insulated for 500V.
Resonator.	150 to 500V, d.c. rated at 60mA, positive earthed, negative to cathode.
Reflector.	0 to 500V, d.c. rated at 1mA, positive connected to cathode. Transformer insulated for 1kV.
Cathode screen.	As for reflector.

The degree of smoothing required will depend on the type of measurement being made. If the frequency variation with reflector potential is taken as 1Mc/s/V, and the variation of frequency with resonator potential as 0·25Mc/sV, a reasonable estimate of the permissible ripple can be made.

Reflector Scanning

For many measurements, it is useful to have a pulsed output from crystals, wavemeters, or standing wave detectors, since an a.c. amplifier and oscilloscope presentation can then be used. A convenient method of doing this is to modulate the reflector supply with a 500c/s sine wave of amplitude sufficient to scan through one or more complete reflector modes. The oscilloscope picture will then show the reflector modes as in Fig. 2, so that details of mode shape, hysteresis, etc., can be observed. If two inputs are available to the amplifier, the wavemeter output can be superimposed on the power output curve, and frequency or electronic tuning properties can readily be checked.

Operating Precautions

The reflector of a reflex klystron should never be allowed to be positive with respect to the cathode, since this would enable the beam to bombard the reflector, possibly resulting in valve softness. For this reason, the maximum impedance in the reflector supply must be kept below the figure laid down by the manufacturer; in a high-power valve such as the R6010, it is advisable to use a diode clamping circuit to prevent the reflector from going positive.

The lives of reflex klystrons should be comparable with normal valves, and the user should employ the usual precautions to ensure long life. The heater voltage rating should never be exceeded, and may sometimes be slightly reduced to increase life. Maximum ratings should not be exceeded, and cooling the valve by means of an air flow is a definite advantage, even if cooling is not specified by the manufacturer.

Acknowledgments

The author is indebted to the Directors of Electric and Musical Industries Ltd. for permission to publish the article.

REFERENCES

1. Ginzton, E. L., Harrison, A. E. Reflex-Klystron Oscillators. *Proc. Inst. Radio Engrs.* 34, 97 (1946).
2. Barford, N. C., Bowman-Manifold, M. Elementary Theory of Velocity-Modulation Oscillators. *J. Instn. Elect. Engrs.* 94, Pt. 3, 302 (1947).
3. Worsley, P. K. Electron-Tubes for Microwave Applications. *Brit. Commun. Electronics.* 3, 606 (1956).
4. Pearce, A. F., Kreuchen, K. H., Baron, C., Houlding, N., Ratcliffe, S. Plug-in Reflex Klystrons for Microwaves. *J. Electronics Contr.* 3, 535 (1957).

GENERAL REFERENCES

Klystrons and Microwave Triodes. M.I.T. Radiation Laboratory Series. Vol. 7. (1948).

Beck, A. H. W. Velocity-Modulated Thermionic Tubes. (Cambridge University Press, 1948).

MULTI-CAVITY KLYSTRONS

By V. J. Norris*, B.Sc.

A review is given of the factors determining the gain, efficiency and bandwidth of multi-cavity klystrons, together with results which have been achieved in practice. Recent applications and possible improvements are also briefly discussed.

THOUGH the use of reflex klystrons as local oscillators is the best known application of klystrons, it is perhaps in the use of klystrons as amplifiers that the greatest advances have been made in recent years.

Until comparatively recently the magnetron was used exclusively when a high power source was required at microwave frequencies. The use of amplifier klystrons to provide very high levels of power was first demonstrated at Stanford University[1]. Using a three-cavity klystron a peak power of 20MW was achieved at a frequency of 3 000Mc/s with an efficiency of 35 per cent. This was in excess of the peak power which had been achieved previously by any other means.

The ability of the klystron to provide higher peak and mean powers than the magnetron is due to the different geometry. In the klystron the cathode and beam forming region, the structure and the collector are separate and independent regions. The only function of the collector is to dissipate energy and so it can be of any shape and size depending on the requirement. In the magnetron, on the other hand, the three regions are interdependent with the net result that the achievable power is severely limited. The klystron has a number of other advantages over the magnetron; it is used as an amplifier and hence is much more suitable for coherent systems, its mechanism is better understood, and there is no problem comparable with that of moding in the magnetron. Against these advantages must be set the disadvantage of increased size and weight, greater complexity, increased beam impedance leading to higher potentials and the somewhat reduced value of efficiency generally obtained.

The principle of using more than two cavities has been extended still further and valves have been produced using up to six cavities and giving very high values of gain.

In the following sections the factors which determine the gain, efficiency and bandwidth of amplifier klystrons are enumerated. This is followed by a discussion of the applications made possible by recent advances.

Gain in Multi-Cavity Klystrons

In the early stage of a multi-cavity klystron the space charge forces are high in relation to the bunching forces. The modulus of the r.f. current, $|i_{r.f.}|$ due to any gap voltage then varies sinusoidally down the drift tube and may be calculated from the space charge wave theory of Hahn[2] and Ramo[3] as

$$|i_{r.f.}| = \frac{I_0 X \sin hl}{hl} \qquad (1)$$

where h is the modified debunching wave number, X is the bunching parameter, l is the drift length and I_0 is the beam current.

It can then be shown that the gain of an unloaded two-cavity klystron stage is given by

$$G = \frac{2 \cdot 8 \times 10^{-11} \beta^4 R_T^2 K^2 f^2 \sin^2 hl}{h^2} \qquad (2)$$

where R_T is the cavity impedance due to both cavity losses and beam loading, β is the beam coupling factor, K is the beam perveance and f is the frequency.

Equation (2) shows that there is an optimum spacing for maximum gain per stage given by a drift length such that $hl = \pi/2$. The maximum gain per stage is then given by

$$M = \frac{2 \cdot 8 \times 10^{-11} \beta^4 R_T^2 K^2 f^2}{h^2} \qquad (3)$$

From equations (2) and (3) the synchronous, narrow-band gain of a multi-cavity klystron is readily determined in terms of G and the advantage of using extra unloaded cavities to produce high gains becomes apparent. For a two cavity valve the gain is $G/4$. The factor of $1/4$ is due to the necessity to load the output cavity in order to extract power. A two stage valve containing three cavities will have a gain of $1/4\ G^2$. Thus by adding only one extra cavity the gain when measured in decibels is more than doubled.

The VX8507 shown in Fig. 1 provides an example of the improved performance which may be obtained by adding cavities. An experimental two-cavity valve gave a gain of 13dB. The normal five-cavity version[4] has produced a gain in excess of 70dB, at a frequency of 9 375Mc/s.

In general, the maximum useful gain which can be obtained is determined either by the signal-to-noise ratio at the output or by the onset of oscillation which is usually caused by the return of fast secondary electrons from the output region.

Efficiency

In many applications for which the klystron is particularly suitable, such as transmission of radio signals by means of high power tropospheric scatter, the efficiency of the klystron is of great importance. Basically, the bunching mechanism has a high efficiency, though it is not possible to write down an exact expression for the efficiency, owing to the difficulty of predicting the r.f. current in the beam. To a good approximation, however, the r.f. current in the beam is given by[5]

$$|i_{r.f.}| = 2 I_0 J_1 \left\{ \frac{X \sin (hl)}{hl} \right\} \qquad (4)$$

* *Mullard Research Laboratories*

which gives a maximum bunching efficiency of 58 per cent when $X \sin(hl) = 1{\cdot}84$.

It is possible to increase the bunching efficiency still further by bunching with more than one cavity. For example, in a three-cavity klystron, by tuning the centre cavity to a higher frequency than the first cavity, the two bunching voltages may be so adjusted that they are in phase; the bunching efficiency is then increased to 80 per cent[6].

In practice these values of efficiency are not achieved. At low power levels the r.f. current in the beam is insufficient to raise the output gap voltage above the beam voltage. The optimum load impedance is then equal to the cavity impedance, giving a maximum output circuit efficiency of only 50 per cent, half the available power being dissipated in the output cavity. At high power levels the

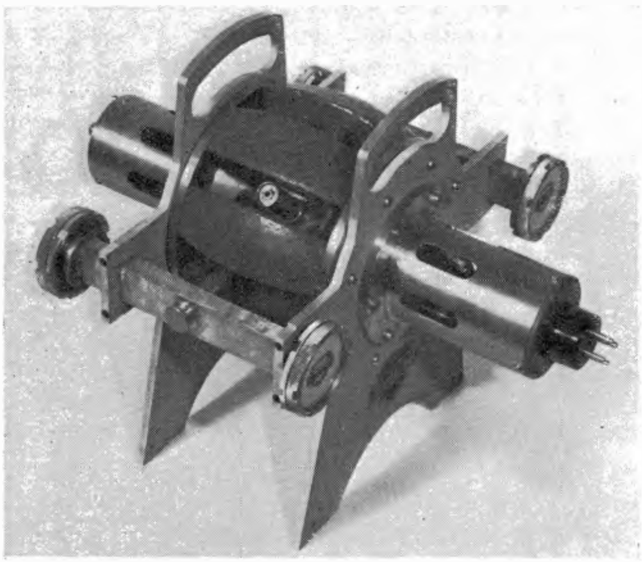


Fig. 1. A high-gain X-band multi-cavity klystron amplifier (Mullard development type VX8507)

circuit efficiency approaches unity as the output load is increased in order maintain the gap voltage at about the beam voltage. However, due to space charge debunching and other effects, these figures are not achieved even at the highest power levels and the klystron, as at present developed, has a somewhat lower efficiency than the magnetron. Values of up to 50 per cent are quoted by manufacturers for some high power tubes used for scatter propagation.

Bandwidth

Although the amplifier klystron is inherently a narrow band device compared with the travelling-wave tube, very useful bandwidths may be obtained by suitably stagger-tuning the cavities in a multi-cavity valve. The bandwidth is adequate for many applications, such as the propagation of television signals or telephony. There are occasions, however, when the maximum attainable bandwidth is required. If all the cavities are assumed to have the same value of unloaded Q, Q_0, and are tuned to the same frequency, f_0, the bandwidth to half power points is given by

$$B = Cf_0/Q_0 \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

C is a constant determined by the number of cavities. (For five cavities $C = 0{\cdot}46$). This synchronous bandwidth will be many times be many times less than that which can be achieved by stagger-tuning the cavities, since in the synchronous case the overall bandwidth is determined mainly by the unloaded cavities. On stagger-tuning, providing that enough cavities are used, the achievable bandwidth is determined by the heavily loaded output cavity. The product of the efficiency, η and the bandwidth B is then given by

$$F = \eta B = k^2\beta^2 (I_0/V_0)(R/Q) \quad \ldots\ldots\ldots \quad (6)$$

where k is the ratio of $|i_{r.f.}|$ to I_0.

In general the load is adjusted to give maximum efficiency and the bandwidth is then determined by the parameter $\beta^2(I_0/V_0)(R/Q)$. Both the value of R/Q, for the optimum shape of output cavity, and the value of β, the coupling factor, increase with decreasing beam and cavity nose

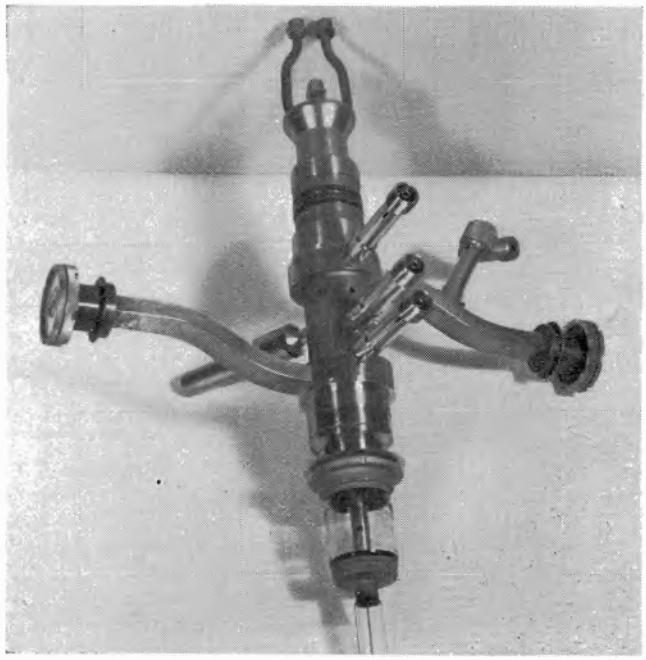


Fig. 2. A high-power X-band multi-cavity klystron amplifier (Mullard development type VX8514)

diameter. Since the beam diameter can be reduced and the beam conductance, I_0/V_0, increased only if the focusing magnetic field is increased, the maximum value of the parameter $\beta^2(I_0/V_0)(R/Q)$ is determined mainly by the magnetic field. The percentage bandwidth available for a given magnetic field increases both with power level and with the wavelength. The VX8514 shown in Fig. 2 is an example of a six-cavity klystron designed to give a peak power output of 0·25MW at X-band. In order to achieve a bandwidth of 1 per cent at this frequency a field of 2 000 gauss is required. The permanent magnet for this valve weighs over 200 lb, but there are promising methods by which the overall weight and size of the permanent magnet in such a valve may be reduced. Among these are the possibilities of using new permanent magnet arrangements or using lightweight aluminium foil coils.

Broadband Tuning

The previous section gave an account of the limitations imposed by the output cavity on the bandwidth of a multi-cavity klystron operated under conditions of high efficiency.

There still remains the problem of determining the resonant frequency distribution of the cavities which is required to produce this bandwidth.

If the frequencies of the cavities are not offset too far from the synchronous condition and if the gain per stage is high enough, then each stage may be considered independent of the other stages and the behaviour is then analogous to that of a conventional amplifier. If the cavity Q's can be adjusted, the tuning and loading of the cavities may be carried out by establishing methods such as Butterworth's and Chebichev's method for i.f. amplifiers. It is then possible to discuss the gain bandwidth product for a given distribution. However, in practice, the loading of the cavities is not adjustable, and the stages are not independent. The major inaccuracy involved in considering the behaviour of a conventional i.f. amplifier analogue is caused by the fact that in a multi-cavity klystron the driving current at any cavity is made up of the vector sum of current produced by the voltages on all previous cavities and not the voltage on the previous cavity alone.

As the amount of broad-banding increases, the importance of the contributions to the total current by the earlier stages increases and the simple approach outlined above must be modified. To a close approximation it can be shown that the overall effect on the frequency response is to add two absorption resonances to the conventional gain-frequency curve[7]. The effect of these unwanted resonances can be minimized by placing the cavities a quarter plasma wavelength apart, the voltage at any gap will then produce no r.f. current two cavities further on. Also, by suitably stagger tuning the cavities it is possible to counteract the resonances to some extent.

The net result of all this is that broad-banding klystrons is a somewhat empirical procedure. The various analyses enable one to predict the bandwidth which can be achieved and how many cavities are required to do this. In practice, providing a suitable swept source for driving the valve is available, the output versus frequency curve may be viewed on an oscilloscope and broad-banding the valve is then neither a lengthy nor a difficult procedure.

Applications

Tropospheric Scatter

In recent years it has been found that the distance over which useful signals can be propagated is not limited by line of sight[8]. Using tropospheric scatter, signals can be transmitted over a wide frequency range up to at least 10 000Mc/s. The attenuation increases rapidly with distance but is relatively independent of frequency. Despite the existence of multi-path transmission leading to fast fading, useful bandwidths of a few megacycles per second may be obtained, provided that narrow beam aerials are used. This fact makes possible a number of applications, particularly in the field of point to point communication over distances of several hundreds of miles. In fact, the first direct commercial telephone link by tropospheric scatter over a distance of 380km has recently been announced[9].

Klystron amplifiers are particularly suitable as transmitters for this application in which as much power as possible is required. Since, by using enough cavities, the gain of the amplifier klystron can be made very large, the required power could be obtained with a reflex klystron providing the driving power. The multi-cavity klystron will also give the required bandwidth for this application. A number of valves have been made, and are available, giving 10 or 20kW of c.w. power with gains up to 50dB, at frequencies from a few hundred megacycles up to several thousand megacycles per second.

Such tubes are also used for broadcasting high power u.h.f. television signals.

Stable Sources at Microwave Frequencies

Another application of multi-cavity klystrons arises when a very stable source at microwave frequencies is required. This requirement arises in many radar and other applications. The best known method of meeting this requirement is that of multiplying the frequency from a crystal-controlled oscillator using conventional valve circuits and following this by a frequency-multiplier klystron. This system is bulky and the stability of the output is reduced due to instabilities introduced in the conventional frequency multiplication stages. By using an amplifier klystron having very high gain a much simpler system is possible. By driving a VX8507 with the 31st harmonic from a silicon crystal driven at 300Mc/s up to 1W of crystal-controlled power may be obtained at 9 300Mc/s.

Improvements

Though the klystron is established as a practical device there are still a number of desirable improvements. Perhaps the biggest disadvantage of the klystron is its high beam impedance, which means that at a given power level the beam voltage is high and that in a pulsed application there is a modulation problem. A great deal of work is going on to develop electron guns giving a higher value of current at a given beam voltage. Where the mean power is high, the modulating power required may be reduced by isolating the anode from the body and collector. The current intercepted by the anode is small and the modulation power is reduced. In some cases where the mean power is low it is possible to use a grid for modulating purposes. The grid is normally run negative and the valve is switched on by applying a positive pulse to the grid.

For some applications the bandwidth of the klystron is too low. It is possible to increase the bandwidth, which is limited by the output circuit, by using coupled cavities as the output circuit and this approach may give an appreciable increase in bandwidth.

Acknowledgments

The author wishes to thank C.V.D., Admiralty, the Directors of Mullard Ltd, and the director of Mullard Research Laboratories for permission to publish this article.

REFERENCES

1. Chodorow, M., *et al.* Design and Performance of a High Power Pulsed Klystron. *Proc. Inst. Radio Engrs. 41*, 1584 (1953).
2. Hahn, W. C. Small Signal Theory of Velocity Modulated Electron Beams. *Gen. Elect. Rev. 42*, 258 (1939).
3. Ramo, S. Space Charge and Field Waves in an Electron Beam. *Phys. Rev. 56*, 276 (1939).
4. Chalk, G. O., Manley, B. W., Norris, V. J. A Five-cavity X-band Klystron Amplifier. *Electronics 2*, 50 (1956).
5. Chodorow, M., Ginzton, E. C., Nalos, E. J. Debunching of Electron Beams Constrained by Strong Magnetic Fields. *Proc. Inst. Radio Engrs. 41*, 999 (1953).
6. Warnecke, R., Guenard, P. Tubes a Modulation de Vitesse. p. 120 (Paris: Gauthier Villars, 1951).
7. Curnow, H. J. A Study of the Frequency Response of a Six-cavity Klystron. *S.E.R.L. Tech. J. 46*, 14 (1956).
8. *Proc. Inst. Radio Engrs.* Scatter Propagation Issue *13*, 1175 (1955).
9. *L'Onde Electrique*, No. 367, 907 (1947).

Coaxial-line Velocity Modulated Oscillator Valves

By D. E. Lambert*

The principle of operation of the coaxial-line velocity-modulated oscillator valve is described together with a discussion of the advantages and limitations when compared with other v.m. oscillators. A description is given of the use of this type of valve in various equipments operating in the 800 to 5 000Mc/s frequency range. A list of valves which are in current manufacture is included giving operating details and performance.

THE type of velocity-modulation oscillator to be described uses a coaxial-line resonator as part of the velocity modulation system to generate microwave power. Work on this type of valve was started in the laboratories of Standard Telephones and Cables Limited in 1938 and active development has continued ever since. In 1939 sealed-off non-tunable valves achieved an output of a hundred milliwatts or so at frequencies in the region of 2 500Mc/s. During the war years this type of valve was used in signal generators, wide-band search receivers, some experimental radar equipment and in a radio-link for military use.

Since 1946, it has been actively developed to fulfil a number of roles, mainly in the Company's microwave radio link systems in the 3 600 to 4 200Mc/s and 4 400 to 5 000Mc/s frequency bands.

Principles of Operation

The essential requirements for a velocity modulated oscillator are one or more gaps in a resonator system for interaction between electron beam and r.f. fields and adequate coupling between these gaps, if more than one is used, to sustain oscillations. In the coaxial line resonator used for this type of valve, two gaps are used one on either side of the centre conductor, between inner and outer conductors, see Fig. 1, and the electron beam is projected across a diameter perpendicular to the resonator axis. The electric fields in the two gaps are equal in magnitude but opposite in phase, and as they form part of the same resonant circuit, the coupling between them is unity. The valve operates in the single-transit mode, the electron beam traversing the resonator once. Velocity-modulation occurs in the first gap, electron bunches forming in the drift-tube separating the gaps, i.e., in the inner conductor, and energy is abstracted from the bunched beam in the second gap. The beam is collected on an anode external to the resonator. The valve resonator is coupled to another circuit, external to the valve envelope, which serves as a means of tuning the frequency of the oscillator and from which the generated r.f. power can be coupled.

There are two methods of changing the oscillatory frequency of a v.m. oscillator, mechanically and electrically. For mechanical tuning, the physical constants of the circuit coupled to the valve resonator are changed so as to affect the resonant frequency, and the coupling between valve resonator and tuning circuit is dependent on the amount of tuning required. The mechanical-tuning range of this type of valve can be made as much as 2·5 : 1. For electronic-tuning, the susceptance of the bunched electron beam is changed by variation of the beam velocity, and this varies the resonant frequency of the valve resonator. The electronic-tuning range is limited and is usually around $\pm\frac{1}{4}$ per cent.

* *Standard Telephones & Cables Ltd.*

Method of Construction

The method of construction employed is similar on the valve types to be described and they only differ in detail. The three types of resonator assembly employed are illustrated in Fig. 2. In these drawings, the electron gun and anode assemblies have been omitted for clarity.

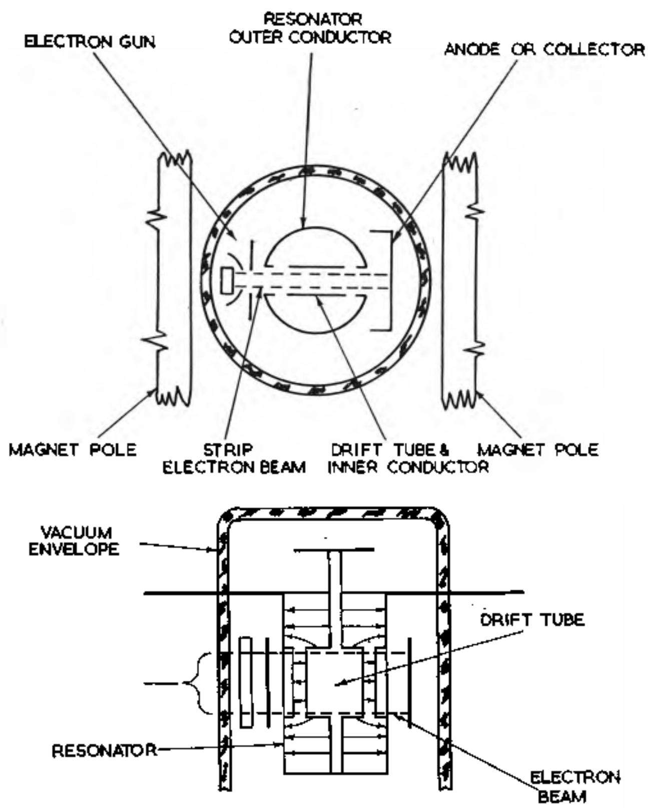


Fig. 1. Cross-section of coaxial-line v.m. oscillator

In Figs. 2(a) and 2(b) the centre-conductor supporting the drift tube is mounted on a short-circuit so positioned that there is a maximum of electric field in the interaction gap region. The two types illustrated in Fig. 2 differ in the manner in which the valve resonator is coupled to the external tuning-circuit. For wide mechanical-tuning range, or where electronic tuning is of secondary importance, the centre-conductor is extended beyond the outer conductor which is connected to a disk-seal and forms a means of coupling to the external circuit, see Fig. 2(b). The form of tuning circuit employed depends on the particular application, but the most common are the coaxial-line, shown in Fig. 3, and waveguide, shown in Fig. 4.

When as much electronic-tuning range as possible is required it is essential that the valve resonator operates on the fundamental mode, in this case $\lambda/4$, and the output circuit is capacitively coupled to the drift-tube, see Fig. 2(a). The coupling is adjusted so as to obtain adequate mechanical-tuning range, and is usually much less than that obtained with the construction shown in Fig. 2(b).

For valves to operate at frequencies below about 1 500Mc/s, it is possible to support the drift-tube on an insulator, as shown in Fig. 2(c). The drift-tube is then at a voltage

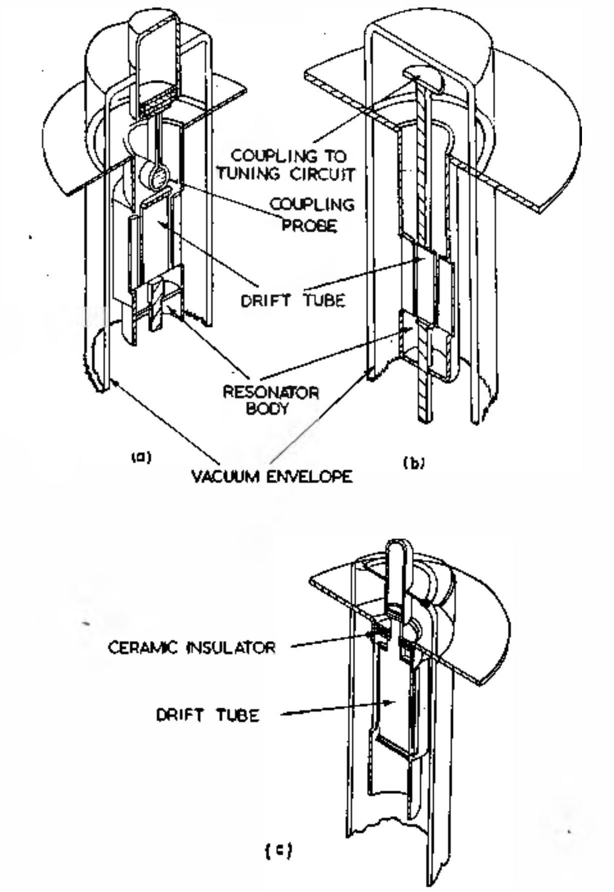


Fig. 2. Constructional details of coaxial-line v.m. oscillator

maximum at all frequencies of operation, the coaxial-tuning circuit being connected between resonator disk and top cap, and the valve operating in the $\lambda/4$ or $3\lambda/4$ circuit mode.

To obtain good frequency stability the outer conductor of the resonator is usually made of copper which gives a good thermal path between drift-tube and resonator disk which is connected to the external circuit to provide a good heat sink.

To take full advantage of the r.f. field in the coaxial line a strip electron beam is used, and to maintain the beam shape during passage through the resonator a uniform magnetic field is used provided by a small permanent magnet. Adjustment is not made to the position of this magnet when replacing valves in equipment, as once having been set correctly when the equipment is made, the three holes that can be seen punched in the resonator disk locate on pins on the magnet plate.

Typical Performance Characteristics

The coaxial-line v.m. oscillator is characterized by low operating voltage, comparatively high efficiency and high stability. The low voltage and high efficiency are a direct consequence of the use of the single-transit mode of operation and a magnetically focused strip electron beam. To illustrate the kind of performance characteristics obtained, Figs. 5 and 6 show power output and operating voltage versus frequency for a number of valve types designed to operate in the 700 to 5 000Mc/s frequency range.

Fig. 5(a) refers to a group of valves designed to cover the 4 400 to 5 000Mc/s band between them with around $\pm$10Mc/s electronic tuning range. The operating cathode current in each case is 50mA, and the valves operate in a tuning circuit as shown in Fig. 4.

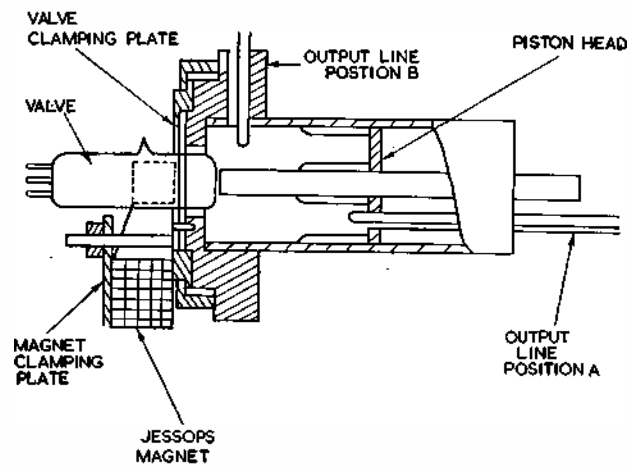


Fig. 3. Coaxial-line tuning circuit

Fig. 6(a) shows the performance of a valve designed to have a wide mechanical tuning range, and operates in a coaxial line tuning circuit as shown in Fig. 3. The performance shown is for a d.c. input of 15W, except at the lower frequencies where the input is limited by the maximum permissible cathode current of 65mA.

Fig. 6(b) shows the performance of a valve operating in the 700 to 1100Mc/s frequency band. The operating cathode current was 80mA, and it is to be noted that the overall efficiency is in excess of 10 per cent over most of the band. The valve operates with a coaxial-line tuning circuit and the electronic-tuning range is about $\pm 2\frac{1}{2}$Mc/s between frequencies of 800 and 1 000Mc/s.

Fig. 5(b) shows the performance of the latest valve to be designed for use in the 3 600 to 4 200Mc/s communications band, and it can be seen that this valve delivers a power output in excess of 1W. The tuning circuit is of the waveguide variety with coaxial-line output, and is illustrated in

Fig. 4. Waveguide tuning circuit

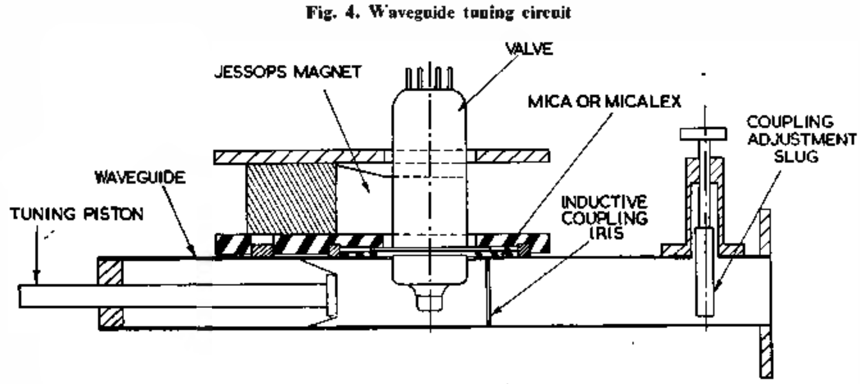


Fig. 8. All of these valves are conduction and convection cooled and do not require any air-blast cooling.

The characteristics shown are for typical operating conditions but the cathode current, and hence the performance, can be varied by means of the voltage applied to the screen electrode of the electron gun. This current control has the added advantage that the r.f. performance of the valve can be maintained almost constant during the life of the valve by adjusting the screen voltage so as to keep the cathode current constant.

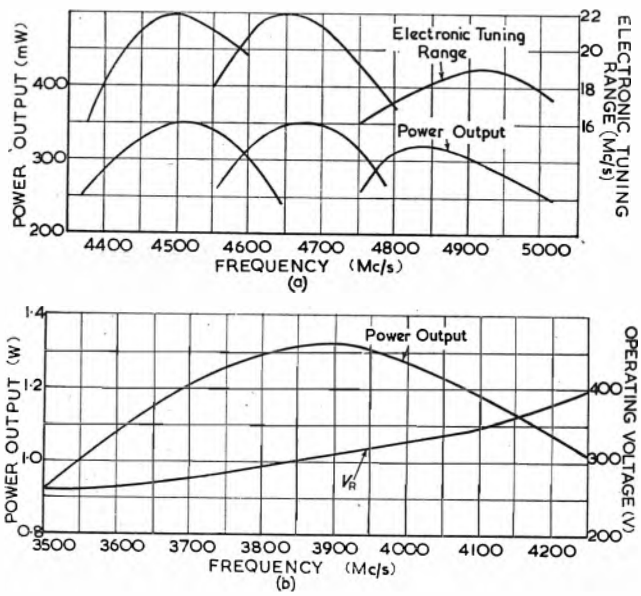


Fig. 5. (a) Power output and electronic tuning range for valves type V245C/1K and V249C/1K

(b) Power output and operating voltage versus frequency for V238A/1K valve in waveguide tuning circuit

A list of valves of this type in current production is given in Table 1.

Comparison with other Types of V.M. Oscillator

It is of interest to compare the characteristics and performance of the coaxial-line oscillator with other v.m. oscillators. In doing this, it is not intended to show that this type of valve is an answer to all the problems of the microwave system designer, but to show that it is complementary to other types and to show the particular fields of application to which it is best suited.

The other type of valve commonly used in microwave systems is the reflex klystron. This valve uses an electrostatically focused electron beam which leads to limitations in power input, especially at low voltages. In valves operating at comparable voltages to the coaxial line types described, the interaction gaps are usually gridded to keep the coupling

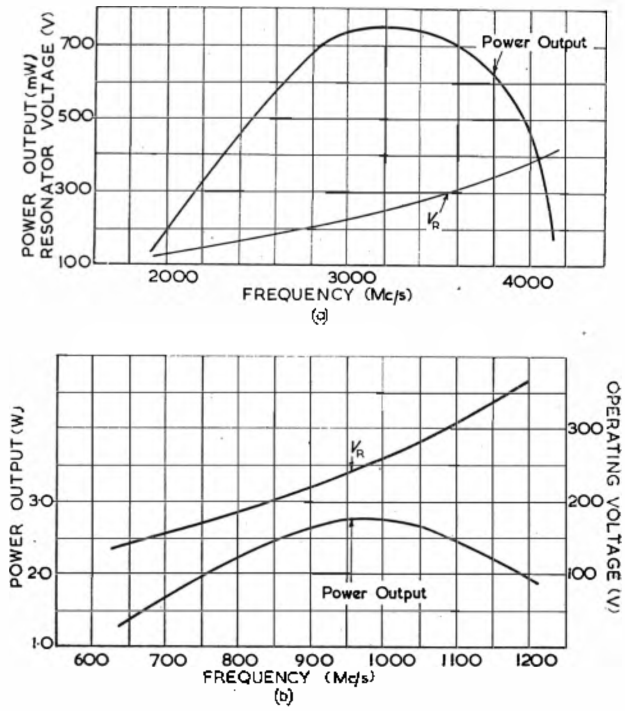


Fig. 6. (a) Power output and operating voltage versus frequency for V233A/1K (CV2190) valve in coaxial-line tuning circuit

(b) Power output and operating voltage versus frequency for V190C/1M valve in coaxial-line tuning circuit

between electron beam and r.f. field sufficiently high. This, combined with the electron-optical problems of the reflector in returning the beam through the gap results in a lower fraction of the cathode current being used to generate power. The use of the magnetically focused strip electron beam in the coaxial-line tube permits higher current densities to be used and this combined with the single-transit mode of operation leads to higher efficiency. The absence of grids in the coaxial-line structure makes it less sensitive to dimensional changes and leads to a higher frequency stability.

The mechanical-tuning ranges of the two types of valve are

TABLE 1

Velocity Modulated Oscillators

COMMERCIAL CODE	CV NUMBER	BASE	HEATER VOLTAGE (V)	FREQUENCY RANGE (Mc/s)	MINIMUM POWER OUTPUT (W)	RESONATOR VOLTAGE RANGE (V)	CATHODE CURRENT (mA)	TYPICAL ELECTRONIC TUNING RANGE (Mc/s)
V190C/1M	—	B8G	6·3	800-1000	2.0	180-270	80	±2·0
V230A/1K	234	B7G	6·3	1700-3700	0·3	100-300	50-65	±1·0
V233A/1K	2190	B7G	6·3	2700-4200	0·3	190-380	50-65	±1·0
V235A/1K	2221	B7G	6·3	2700-4000	0·5	190-350	50-65	±1·0
V237C/1K	—	B7G	6·3	3560-3820	0·35	225-285	45	±4·0
V238A/1K	—	B7G	6·3	3550-4250	0·9	250-400	50	—
V239C/1K	—	B7G	6·3	3780-4040	0·35	225-285	45	±7·0
V240C/1K	—	B7G	6·3	4000-4100	0·35	240-280	45	±9·0
V240C/2K	2189	B7G	6·3	3950-4050	0.35	240-280	45	±9·0
V241C/1K	—	B7G	6·3	4000-4240	0.35	225-285	45	±7·0
V245C/1K	—	B7G	6·3	4400-4630	0·2	230-265	50	±10·0
V246A/1K	228	B7G	6·3	4400-4850	0·45	200-250	65	±6·0
V246A/1K	485	B7G	6·3	4560-4860	0·2	200-230	40	±4·0
V247C/1K	—	B7G	6·3	4570-4750	0·2	230-265	50	±10·0
V249C/1K	—	B7G	6·3	4760-5000	0·2	240-290	50	±9·0

similar. Although most reflex klystrons have built-in tuning mechanisms for convenience, external cavity valves are made with tuning ranges in excess of 2 : 1. In frequency ranges where the two types of valve can be compared from this point of view, the coaxial line valve usually operates with a shorter transit-time in the bunching region, which reduces the possibility of mode interference.

There is a lot to be said for the use of external tuning circuits as they enable the circuit to be tailored to a specific application and reduce the cost of the valve.

The method of cooling the valve is of importance in some applications and if powers in excess of $\frac{1}{4}$W to $\frac{1}{2}$W are required, most reflex-klystrons will require some air blast cooling. All the coaxial line valves described in this article need no air-blast cooling for operations up to their maximum ratings.

The major limitations to the coaxial line type of valve are the limited electronic-tuning range possible and the frequency range in which valves can be made to operate with an attractive performance.

The frequency limitation arises because of the increasing losses in the coaxial line resonator as the frequency is raised which, coupled with the reduced physical length results in the performance falling rapidly above about 6 000Mc/s. The low frequency limit is set by physical size but the V190C/1M has been operated successfully at 600Mc/s. The reflex klystron has a similar low frequency limit but the high frequency limit is nearer 60 000Mc/s.

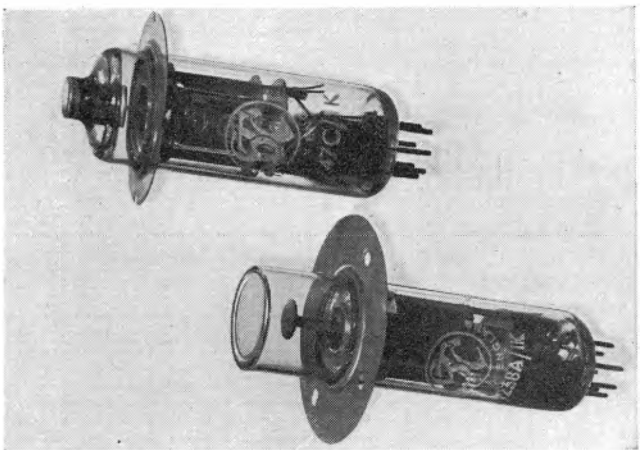


Fig. 7. V247C/1K and V238A/1K coaxial-line oscillator valves

The electronic tuning range of a v.m. valve is a function of both bunched electron beam and valve resonator characteristics. In the case of the reflex klystron, the amount of bunching that takes place is a function of both transit-time and reflecting field shape, enabling the bunching characteristic to be so adjusted that wide electronic tuning-ranges are obtained. The valve resonator should be designed so that the interaction gap capacitance is a minimum and the coaxial line resonator is also inferior to the type of resonator commonly used for the reflex-klystron, in this respect. Although the higher current density usable at a low beam voltage for the coaxial line valve offsets these effects to some extent, the net result is that much wider electronic-tuning ranges, ± 1 per cent, can be obtained with the reflex klystron.

To summarize, for applications in the frequency range 600 to 6 000Mc/s where low voltage operation, power output in excess of $\frac{1}{4}$W, moderate electronic tuning range and good frequency stability are of importance the coaxial line valve will have some advantage.

To illustrate some of these points, a number of successful applications of the coaxial line type of valve, mainly in the radio-link field during the past ten years, will be described.

Mobile F.M. Radio Links

This equipment[1] is intended for the transmission of television or multi-channel telephony signals for distances up to 30 or 40 miles and has been in use by the BBC since 1948. The equipment operates in the frequency range 4 400 to 5 000Mc/s and three valves[2] are used in the same waveguide tuning circuit to cover this range. One of these, the V247C/1K, is shown in Fig. 7, and a view of the valve in the equipment in Fig. 8.

The valve is used both as the transmitter and local oscillator in the receiver. The modulation linearity is adequate for television or up to 60 telephone channels.

This particular application is of interest as it illustrates the good frequency stability of this type of valve. No automatic frequency control is used on either transmitter or receiver, and after the initial warming-up period will operate for long periods without adjustment.

The average life of the valves used is in the 3 000 to 4 000 hour region.

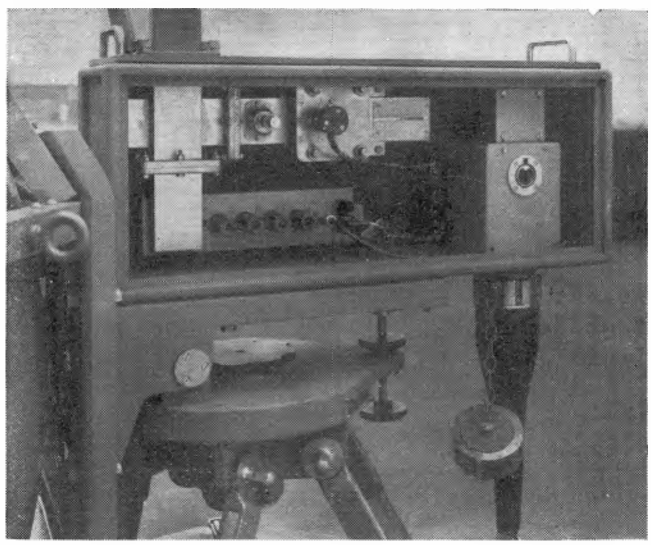


Fig. 8. View of V247C/1K valve in receiver of portable radio-link

Manchester—Kirk-O'-Shotts Television Link

This was the first of a number of similar equipments[3, 4] for permanent relay of television or telephony signals and has been in use since 1952. Other equipments of a similar nature are in operation in the United Kingdom, Japan, Canada, Brazil, Austria and Spain.

The coaxial line valve is used as the local oscillator in all the receivers at both terminal and repeater stations. In the type of repeater used in these systems, in which most of the amplification takes place at frequencies in the region of 70Mc/s, oscillators are required to drive the mixers which convert from the incoming microwave frequency to i.f. and back again after amplification, to the outgoing microwave frequency. A crystal mixer is used for the i.f. to microwave frequency conversion and has to produce enough power to drive the travelling-wave output amplifier. As this type of mixer has appreciable conversion loss, 12 to 15dB, about $\frac{1}{2}$W of microwave power is required from the local oscillator. This is supplied by a coaxial line oscillator V240C/2K, and in later equipment V237C/1K, V239C/1K and V241C/1K forming a series to cover the complete 3 600 to 4 200Mc/s communications band. These all operate at 300V with a cathode current of 45mA and the average life in service is around 6 000 hours.

New Wide-Band Radio Link

Field experience with the equipment described above showed that it would be advantageous to eliminate complex a.f.c. circuit arrangements in future equipments, thus simplifying the maintenance and reducing the size of the equipment.

In the latest system[4], a V238A/1K valve (see Fig. 5) is used to drive both incoming and outgoing mixers and

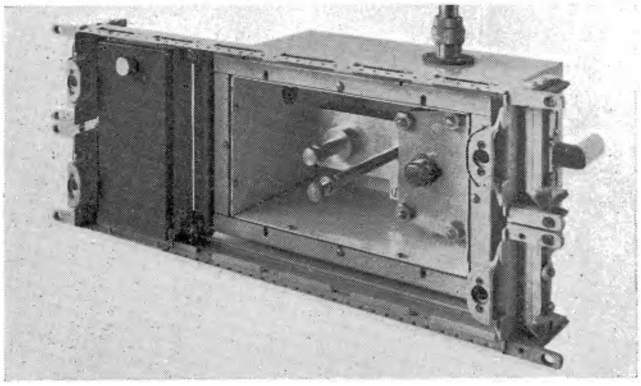


Fig. 9. View of V238A/1K in waveguide tuning circuit with lid of oven removed

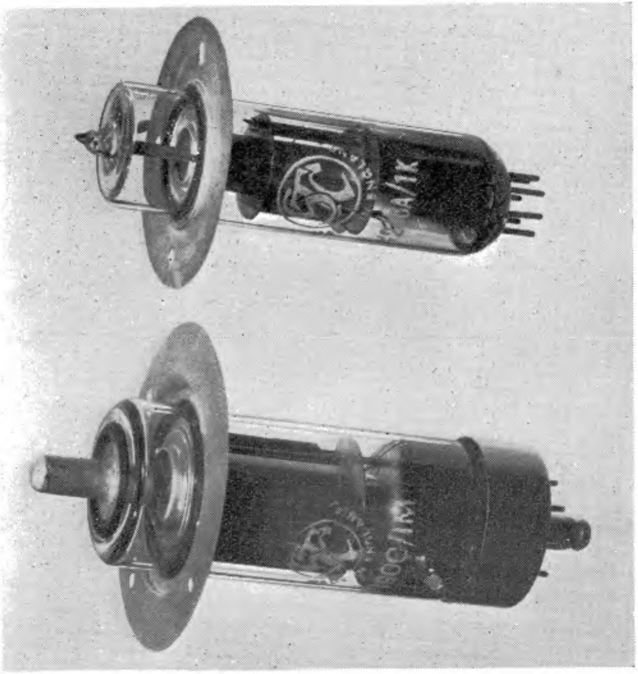


Fig. 10. V190C/1M and CV485 coaxial-line oscillator valves

operates without automatic frequency control. To achieve the high frequency stability required for these systems, the valve and tuning circuit is operated in a temperature controlled oven and fed by carefully regulated power supplies. By this means, a long-term frequency stability of better than ±100kc/s in the frequency range 3 550 to 4 250Mc/s can be obtained.

To avoid distortion due to frequency pulling of the valve by the high-level mixer some buffering must be introduced between them, the amount being dependent on the pulling-figure* of the valves, which should be a minimum. The V238A/1K has a pulling figure of about 1¼Mc/s but the use of ferrite isolators has now reduced the importance of this characteristic. The average life is similar to the V240C/2K valve used previously.

A view of the valve in the equipment with the lid of the oven removed is shown in Fig. 9. Equipments of this type are about to be installed in the United Kingdom, Switzerland, New Zealand, Newfoundland, Australia, South Africa and Malaya.

900Mc/s F.M. Radio-Links

The V190C/1M valve, see Figs. 6 and 10, has been used in a multi-channel f.m. link designed by S.T. & C. Pty, Australia. The valve is used both as a local-oscillator and transmitter valve to drive the 2C39A output amplifier.

This valve has also been used in experimental f.m. equipment at 600Mc/s for the transmission of television where the low voltage, 130V, and 1W output were of importance. The tuning circuit is reasonably small at this frequency, 2in long, as the valve plus circuit operate in the $\lambda/4$ mode.

Lightweight Communication Set

The CV485 valve, which is a later version of the V246A/1K valve, see Fig. 10, has been used in the British Army portable multi-channel f.m. telephony link manufactured by Marconi Wireless Telegraph Company. The valve is used as both transmitter and local oscillator. The operating voltage is 230V and with a cathode current of 40mA has an output power of about 200mW.

Signal Generators

It has been mentioned that wide mechanical tuning ranges are possible with this type of valve and this characteristic has found application in the design of signal generators. A typical example is the CV2221 which is used in the type MW40/5 S-band oscillator manufactured by Decca Radar Ltd. The particular point here is that the high power output enables the generator output to be extended to about 10mW while still maintaining linear attenuator calibration.

The circuit used is of the coaxial-line variety and the valve performance is similar to that shown for the CV2190 in Fig. 6.

Conclusions

The construction, design and performance of a range of v.m. oscillators utilizing a coaxial line resonator has been described. It has been shown that for some applications, this type of valve will have some advantage over the reflex-klystron and examples of its use in microwave systems has been given.

As the trend in new microwave radio link systems is to higher frequencies, around 7 000 and 11 000Mc/s, it is unlikely that much further development of this type of valve will take place.

Acknowledgments

The author wishes to thank Standard Telephones and Cables Limited for permission to publish this article.

REFERENCES

1. Dawson, G. Portable equipment for a microwave television link. *Proc. Instn. Elect. Engrs.* *99*, Pt. 3A, 231 (1952).
2. Lambert, D. E. A coaxial-line velocity-modulated oscillator for use in f.m. radio-links. *Proc. Instn. Elect. Engrs.* *99*, Pt. 3A, 421 (1952).
3. Dawson, G., Hall, L. L., Hodgson, K. G., Meers, R. A., Merriman, J. H. H. The Manchester-Kirk o'Shotts television radio-relay system. *Proc. Instn. Elect. Engrs.* *101*, Pt. 1, 93 (1954).
4. Dawson, G. S.H.F. Radio Links using Travelling-wave Output Amplifiers. *Electronic Engng.* *30*, 276 (1958).

* *The pulling figure is defined as the maximum change in oscillation frequency that occurs when the phase of a v.s.w.r. of 1·5, in the output line, is changed through 180°.*

Backward Wave Oscillators

By A. G. Stainsby*, B.Sc., A.Inst.P.

The backward wave oscillator has developed rapidly since its invention six years ago. It combines rapid electronic tuning with a frequency range sometimes in excess of an octave. There are two distinct types of oscillator, the O type which is mainly suited to low power operation as a local oscillator or test source and the M type which is capable of delivering high powers with good efficiency and is thus suited to use as a transmitter. The characteristics and mechanism of operation of these two types are explained and typical examples are described.

THE feature which particularly distinguishes a microwave valve from those used at lower frequencies is the inclusion of the circuit or at least a major part of it within the evacuated envelope. Thus if the valve is to be tuned, this must be done by modifying what remains of the circuit external to the valve or by modifying the circuit inside the valve by a mechanism operating through the wall. For example a magnetron may be tuned by moving metal pins in the resonant cavities using a mechanism operating through metal bellows or by coupling a reactance to the output waveguide external to the valve. Reflex klystrons are tuned by plungers in the part of the resonant cavity external to the valve or by means of a piece of ferrite material inside, or they can be tuned over a few per cent electronically, i.e. by altering the reactance of the electron beam. All these methods are, however, limited in range or the rate at which tuning can be achieved. It was the invention of the backward wave oscillator which at once offered rapid tuning over a wide frequency range.

The backward wave oscillator principle was discovered independently by B. Epsztein of the Compagnie Générale de T.S.F. in Paris and R. Kompfner of Bell Telephone Laboratories. Epsztein showed that a condition for oscillation should occur under certain conditions of a beam interacting with a circuit, and later found such oscillations, whose frequency varied with beam velocity, in a travelling-wave amplifier[1,2] of the M type. To the resulting oscillator was given the name 'Carcinotron'. Kompfner, on the other hand, had recognized the possibility of such oscillations during early work on travelling-wave tubes and found oscillations which tuned between 6 and 7·5mm wavelengths in a travelling-wave tube designed by Millman[3,4].

The backward wave oscillator immediately offered a microwave device capable of being tuned rapidly over a frequency range of first 50 per cent and soon over more than two octaves[5] and thus opened up new possibilities for systems employing frequency modulation with an exceptionally wide deviation and of measuring equipment for covering bandwidths of an octave in a single display. Transmitting types of oscillator have been developed[6,7], giving powers in the ten centimetre region of 500W. Receiving types are now available to give continuous coverage between 1 000Mc/s to 15 000Mc/s, each tube having an electronic tuning range of 50 per cent to 100 per cent[8,9] and coherent oscillations with an experimental valve[10] have been obtained up to a frequency of 200Mc/s.

Description

What, then, is the characteristic of the oscillator which has made such remarkable progress in so short a time? In the first place it contains no resonant circuits, the circuit consisting of an artificial line which ideally is matched at each end. The basic features of the device, as shown in Fig. 1, are similar to those of a travelling-wave tube. An electron stream passes along the tube parallel to a periodic structure constituting a delay line which is matched at each end either by an external load or an internal termination. The output connexion is at the gun end of the delay line and in this way the tube differs from the travelling-wave tube. Now the electron stream can interact with any wave whose phase velocity is approximately the same as its own and it can be shown that a wave travelling along a periodic structure may be analysed into components having phase velocities in the same and the opposite direction to that in which the energy is propagating.

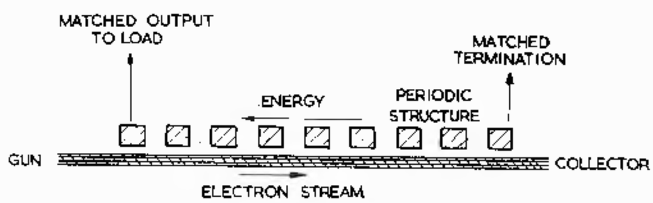


Fig. 1. The basic structure of the backward wave oscillator

A picture of the interaction may be obtained by considering a wave passing along a periodic structure. Imagine that the gaps between successive bars are small. Then, if the wave propagates forward at the same speed as the electrons, an electron will experience a force of the same magnitude and direction at all gaps, e.g. if it is retarded at the first gap it will be retarded at subsequent gaps so long as synchronism is maintained. In this way there is continuous interaction along a tube and this represents the mode of operation of an amplifier. If, however, an electron travels in the opposite direction to the wave, but this time slower, there is again a possibility of the electron experiencing the same retarding field at consecutive gaps provided that the velocity is correct. For example, in the case illustrated in Fig. 2 if the phase difference per segment is $\pi/2$, i.e. there are 4 segments per wavelength, then for an electron to synchronize with the wave at the gaps it must travel back one segment while the wave travels forward 3 segments. In general the relation is that if the electron velocity v is required to synchronize with a wave whose phase change per segment is ϕ in the forward direction, then, to synchronize in the opposite direction, i.e. with a reverse component of the wave the electron velocity must be reduced to $(v\phi)/(2\pi - \phi)$. It is now seen to be possible for the electron stream to couple to a component of a wave whose energy is travelling in the opposite direction, for

* *Research Laboratories of The General Electric Co. Ltd, Wembley*

example, in Fig. 1 while the electrons are travelling from left to right and causing a build up of energy towards the entry end of the tube, there is thus a mechanism to feed back energy into the beginning of the line which modulates the beam giving rise to oscillations.

There are two conditions for oscillation. Firstly that the beam shall be substantially synchronous with the wave on the circuits and secondly that the total phase shift around the tube shall be such as to reinforce oscillations. Since the tube is many wavelengths long, this second condition need not concern us here since, for a very small change of wavelength, the total phase shift can be changed by a substantial amount. The condition of most interest is thus the

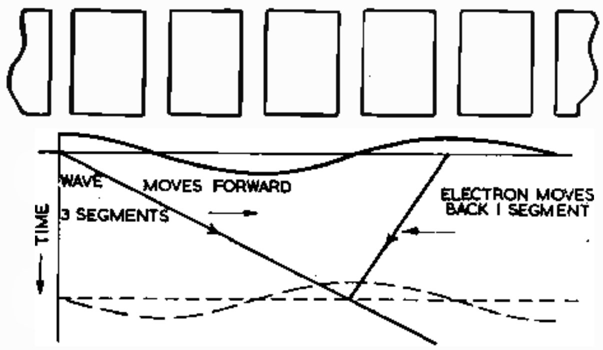


Fig. 2. The wave on the periodic structure analysed into a forward and backward component

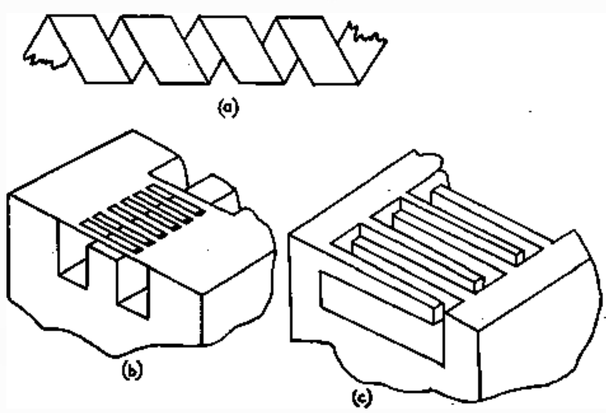


Fig. 3. Slow wave structures
(a) Tape helix
(b) Loaded ladder
(c) Interdigital

synchronizing condition. Since the velocity of the beam may be varied at will by varying the voltage, it will be necessary for the phase velocity of the wave component to alter in order to provide the condition of synchronism and so of oscillation and, as there are no other variables which are free to change, this can only occur if the frequency changes. How, then, does the phase velocity of a wave component depend on frequency? To answer this one must consider the type of slow wave structure which is to be used.

Any periodic structure, for example in Fig. 3 the helix (a) ladder line (b) interdigital line (c) can propagate waves over various bands of frequencies. They can be regarded as filters whose distributed capacitances and inductances combine to provide pass and stop bands as in lower frequency theory. The filters can be described in terms of the phase and frequency diagram in which the frequency (multiplied by $2\pi\ \times$ distance between segments) is plotted as a function of the phase change per segment. Examples of the two general types are shown in Fig. 4, they are in fact typical of band-pass filters. Fig. 4(a) shows a high frequency cut-off when $\beta\rho = 0$, or 2π and a low frequency cut-off where $\beta\rho = \pi$ with propagation for intermediate frequencies. Where β is the phase constant representing the change of phase per unit length.

Now for any structure the phase velocity (the velocity to which electrons couple) is given by ω/β and for a frequency corresponding to the point K in the diagram it is given by the slope of the line OA. The group velocity, or the velocity with which energy is propagated along the structure is given by $d\omega/d\beta$ and is shown by the slope of the tangent AT at the point A. From the curve it is seen that these two slopes are opposite and consequently the group velocity is in the opposite direction to the phase velocity. Furthermore it is seen that as the phase velocity

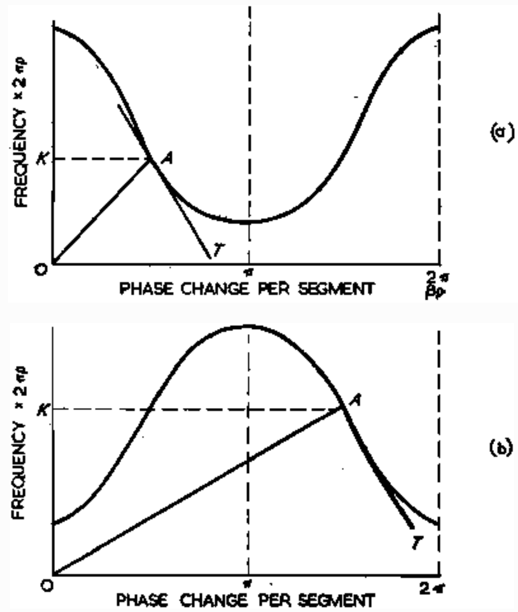


Fig. 4. Dispersion curves
(a) Backward fundamental wave
(b) Forward fundamental wave

is altered (a rotation of the line OA) corresponding to the electron velocity altering, the frequency must alter to maintain synchronism. The interdigital line Fig. 3(c) will give a dispersion curve of this type and since the fundamental component has a phase velocity in the opposite direction to the group velocity this is called a backward wave structure.

If the structure is of the opposite type, as illustrated in Fig. 4(b), i.e. having a low frequency cut-off at 0 and 2π and a high frequency cut-off at π, the phase velocity of the fundamental component has the same direction as the group velocity, since the slope is positive, and this is called a forward wave structure. It may, however, still be used in a backward wave oscillator by utilizing the component wave or space harmonic which gives between π and 2π phase change per segment. Thus one might operate a backward wave oscillator at the point A. This has the advantage that a larger structure may be used for a given wavelength, but the coupling impedance of the wave to the beam is reduced.

If, then, the velocity of an electron beam passing a slow wave structure is known, one can immediately predict to a first order the frequency of oscillation assuming the dispersion curve of the structure is known. The exact frequency depends to some extend on space charge and also depends on the type of backward wave oscillator which is being considered.

Types of Backward Wave Oscillator

In the discussion so far it has been assumed that a beam was being injected parallel to the periodic structure, but no consideration has been given as to how the electrons were travelling or how the beam was produced. There are, however, two distinct types of beam which are usually described as the O and M types.

The O type (shown in Fig. 5(a)) employs an electron gun in which electrons are accelerated to a given velocity on passing through one or more anodes. The electrons then continue to travel parallel to the periodic structure in a region where no d.c. fields are present. A focusing magnet may be employed, but it is not fundamentally essential to the operation of the tube. In producing oscillations the beam gives up a proportion of its kinetic energy to the r.f. field on the line and power may be extracted from the end nearest to the beam entry. At the opposite end of the line there is a matched resistive load which absorbs any power reflected from the output.

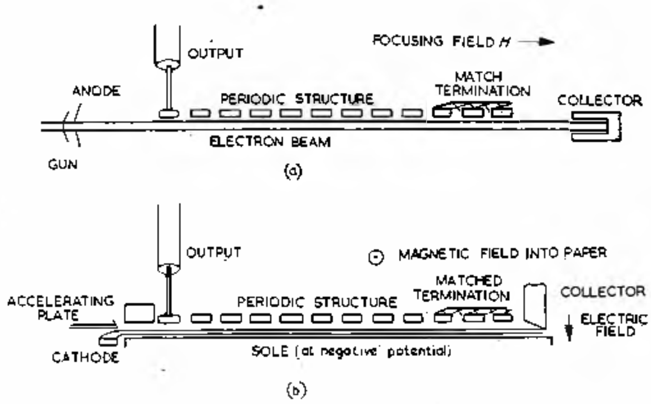


Fig. 5. Backward wave oscillators
(a) O type
(b) M type

In the M type, Fig. 5(b), the magnetic field, as in a magnetron, is at right-angles to the direction of propagation and to an electric field which exists between the line and a negative electrode usually called the sole. The cathode lies parallel to the sole above which is the accelerating plate controlling the beam. The beam is accelerated by the voltage on the plate and is deflected by the magnetic field so that it enters the interaction space between sole and line at a velocity given by the ratio of electric field to magnetic induction E/B. The electron beam now travels parallel to the line and in doing so interacts with the fields of any components of the r.f. wave which are travelling with the same velocity. In this case, however, the beam does not give up its kinetic but its potential energy and gradually approaches the delay line where it is absorbed. These two devices, while operating on the same general principles, are sufficiently different for each to be treated separately.

O Type Oscillator

As stated above an O type tube employs a directly accelerated beam which gives up some of its kinetic energy to the r.f. wave as it passes down the circuit.

The energy is given by the equation $eV = \frac{1}{2}mv^2$. Where v is the velocity of an electron of mass m and charge e, and V is the accelerating voltage. Thus the velocity is proportional to the square root of the voltage. If the phase velocity on the circuit is a linear function of the frequency of the wave, the variation of frequency with voltage will be rapid at low voltages, but changing more slowly at high voltages. Further, unless the current is varied across the frequency band by the use of a grid, the power input will vary considerably across the frequency band. The O type oscillator is thus primarily suited to use as a low power oscillator, for example as a local oscillator in a receiver or low power modulator. A typical example of such a tube is the Mullard VX8509 which covers the frequency range 7·0 to 11·5kMc/s. This is shown in Fig. 6. It consists of an inner tube which slides into the permanent magnet required for focusing. The valve employs an interdigital line which is terminated internally by a resistive attenuator while the other end passes through a fin-line to the waveguide output. The gun produces two flat beams which pass down either side of the interdigital line. The characteristics of frequency and power as a function of line voltage are shown in Fig. 7. It is seen that the frequency increases steadily with increase of the voltage, but the rate of change of frequency per volt falls off at the high voltage end. The power output changes from 40mW at low frequencies to nearly 200mW at the high frequency end of the band. This power output can, however, be controlled by varying the collector current with the aid of the grid. Since the waveguide output is isolated from the line the gun may be operated at earth potential which makes modulation of the control grid much easier than if the line were earthed. By this means the power may be held down to a constant level to within about 1dB, depending on the rate of fluctuation of power of the individual sample and the rate of sweeping.

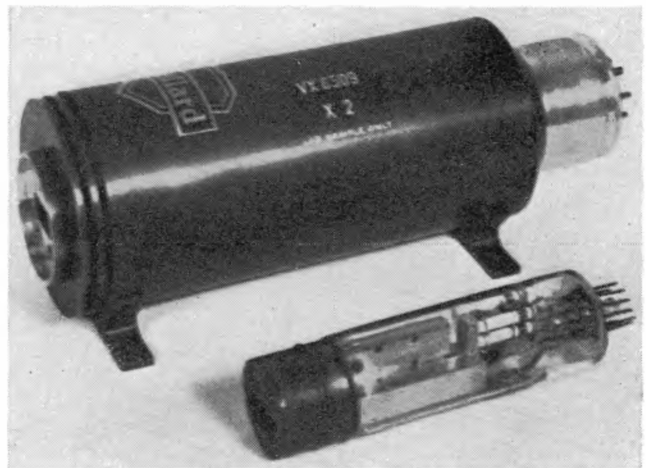


Fig. 6. O type backward wave oscillator VX8509
(Photograph by courtesy of Mullard Research Laboratories)

The tape helix Fig. 3(a) has also proved a useful structure for use in O type oscillators. A helix is essentially a forward wave structure but by use of a broad tape the spatial harmonics are accentuated. The amplitude of the backward wave component increases from zero at the centre of the tube to a maximum close to the helix. Thus in the design of such tubes the beam is injected close to the helix and frequently a hollow beam is used. The connexion for the output may be taken to a waveguide or coaxial coupling using techniques similar to those used for travelling wave amplifiers.

Because of their wide tuning range O type backward wave oscillators have been designed to cover all frequencies between 1 000 and 15 000Mc/s using interdigital or tape helix lines and for still higher frequencies the loaded ladder line has been used. This has the attraction of being easy to construct with great precision, and A Karp[10] has used this type of line to obtain oscillations up to 200kMc/s which represents the highest frequency so far obtained. In his construction he overwound a molybdenum former with gold plated molybdenum tape to produce a line of the type shown in Fig. 3(b). By furnacing, the whole could be brazed together to yield an accurate and robust structure. This gave a forward structure and oscillations were obtained using a reverse spatial harmonic.

The facility for rapid electronic tuning carries with it the disadvantage that it is difficult to use the tube as a stable oscillator. The variation of frequency in the 10 000Mc/s region may be of the order of 4Mc/s/V. Thus if a voltage stable to one part (r.m.s.) in 10^6 is available, the residual frequency modulation due to this power supply is still 1kc/s. Because of this feature, there does seem to be a need for an oscillator with rather smaller frequency coverage at X-band where it will become increasingly necessary for ships' radars to change their frequency to avoid interference. Here a compact tube using some circuit such as was used by Karp may offer a local oscillator covering a bandwidth of, say, 10 per cent with consequently less voltage sensitivity than that of the wide small range tubes,

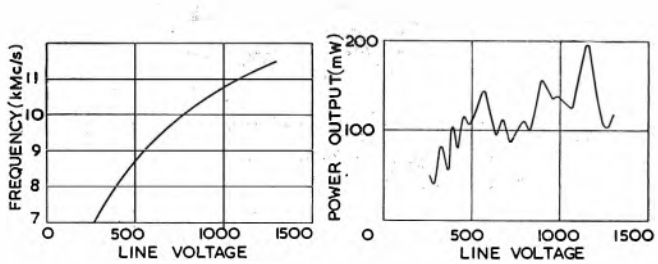


Fig. 7. Characteristics of frequency and power as a function of line voltage for typical VX8509

the voltage sensitivity being the main objection to the use of backward wave oscillators as local oscillators in receivers.

The tuning curve of frequency/voltage will be smooth if the line is perfectly made and matched at either end. If however there is a mismatch at both ends or along the line length the curve will no longer be perfectly smooth. This is because the two mismatches make the line to some extent resonant, with a consequent change of power output and slope of the tuning curve when the resonant frequency is attained. If the product of the reflection coefficients is too great there can even be a discontinuity. This condition is, however, unlikely to occur with practical tubes since the internal attenuator can be very well matched to the line. In the ideal case the pulling figures (as defined by the frequency excursion when a given mismatch usually 0·66:1 is moved through all phases along the line) is zero, but in practice this may amount to a few Mc/s at 10 000Mc/s. Very large mismatches can, in fact, be tolerated with little frequency deviation, thus when the valve is used as a local oscillator there is no need for much padding or an isolator.

M Type Oscillator

A typical example of the M type oscillator is the VX3510 made by the M.-O. Valve Co. Ltd which is shown in Fig. 8, together with the magnet required for its operation. The internal construction is similar to that shown in Fig. 5(b) with the difference that the whole system is curved into the arc of a circle. For the curvature used, however, this does not affect the operation of the tube. The gun containing an indirectly heated cathode injects a stream of electrons into the interaction space between the copper sole and the interdigital line which is brazed to the water-cooled outer block. The output is taken from the first finger of the line and matches into a standard ⅜in coaxial line of 50Ω impedance. The internal transformer is designed to give a match of better than 0·5:1 across the tuning range. At the other end of the line the fingers are sprayed with a resistive coating to provide the internal termination.

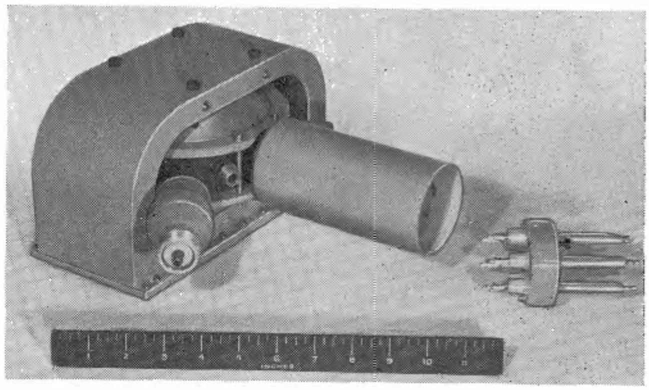


Fig. 8. M type backward wave oscillator type VX3510

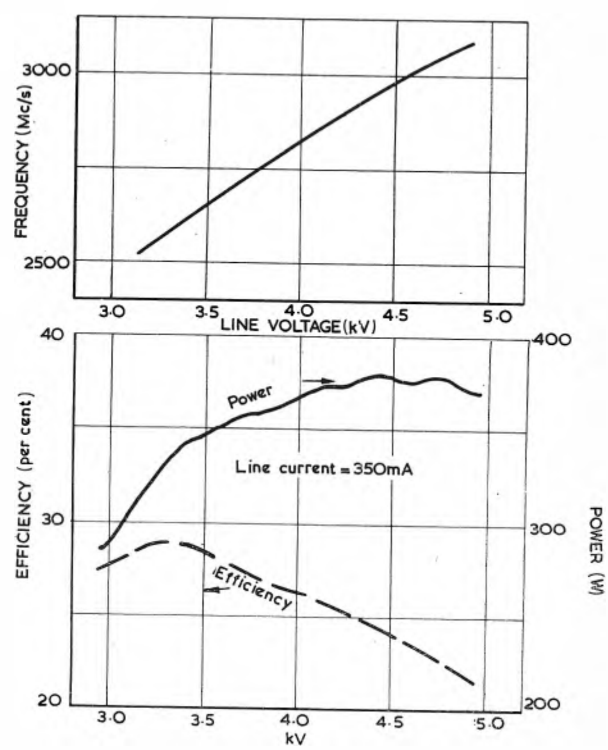


Fig. 9. Characteristics of frequency, power and efficiency as a function of line voltage for typical VX3510

Fig. 9 shows a typical operating characteristic at a line current of 350mA. Here the power, efficiency and frequency are plotted as functions of line voltage for constant current from the cathode. It is seen that unlike the O type, the M type has an almost linear frequency characteristic. This is because the electron velocity is given by E/B and is thus directly proportional to the line voltage. Assuming then that the dispersion curve for the line is linear, the tuning curve is also linear. To approach a linear tuning curve it is necessary to put the upper and lower cut-off frequencies for the line far apart. In the case of the VX3510 these are at appproximately 8 000 and 1 500Mc/s respectively.

It is seen that the efficiency falls from near 30 per cent at the low frequency end of the band to about 20 per cent at the high frequency end. This trend is in line with theory although the absolute value of efficiency is well below the theoretical maximum. To explain how this maximum arises the mode of operation of the valve will be considered.

In the absence of any r.f. field the electrons will travel parallel to the sole along an equipotential line. If, however, an r.f. field is excited on the line the tangential component of the r.f. field will force the electrons closer to or further from the line. In the former case the electrons give energy to the wave and in the latter they extract it. As the electrons which have given up energy approach the line they will experience stronger fields and give up energy faster; the unfavourable electrons which extract energy will move further away and their interaction will become less. It should be emphasized that in the first order theory no density modulation occurs but the beam is drawn up into waves, the space charge density and the velocity remaining the same. This is in contrast to the O type in which density modulation is essential. Fig. 10 illustrates the form taken by the beam after appreciable interaction has taken place.

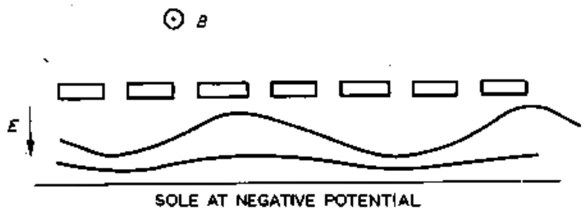


Fig. 10. Electron beam configuration in an M type oscillator

It is seen that when an electron arrives at the line it has given up the whole of its potential energy in passing from the original equipotential given by $V_1 - 1/2\ m/e\ (E/B)^2$ to the potential of the line, or putting it in another way, since the electrons are trapped by the wave they cannot arrive at the line with less than the energy corresponding to their precessional velocity E/B. The maximum possible efficiency of the device is thus given by:—

$$1 - (KE/PE) = 1 - \frac{\frac{1}{2}m(E/B)^2}{eV_0} = 1 - (m/2e)\frac{V_0}{d^2B^2}$$

where V_0 is the voltage between cathode and line and d is the separation of the zero equipotential from the line. This will be the sole-line spacing if the sole is operated at cathode potential, but this is not the usual case. It is seen from the expression that the efficiency is increased by increasing B the magnetic induction for a given voltage of operation or alternatively reducing voltage for a given magnetic induction.

Examination of the efficiency curve in Fig. 9 shows that it slopes in the direction suggested by this theory. However, putting figures for the VX3510 into the equation the maximum possible efficiency at the high frequency end of the band should be of the order of 80 per cent, whereas 20 to 25 per cent is, in fact, realized. The extreme efficiency will, however, be realized only with linear orbits. If the orbits were pure cycloids the expression for efficiency becomes:

$$1 - 2(m/e)\frac{V}{d^2B^2}$$

since the velocity of the electron at the top of a cycloid is $2(E/B)$ which indicates a maximum efficiency of only 20 per cent. Clearly something in between these two extremes is realized.

There is a further effect, that of variation of line current. If the characteristics of Fig. 9 had been taken at a higher current, the efficiencies would also have been higher.

However, as compared with any other oscillators, particularly any with an appreciable tuning range, the efficiency is high and this feature makes the M type valve very suitable as a power oscillator.

If the gun is operated at constant bias the current will be constant and the power input will be a rising function of line voltage. Because the efficiency falls with increasing voltage, as is shown by the formula above, the change of output power over the frequency band is not so marked, amounting only to about 10 per cent for a 20 per cent frequency change. Being a transmitting valve the extremes of the tuning range are not exploited in the M type valve, e.g. although a 2:1 frequency range can be easily obtained, the power output at the low frequency end of the band becomes too small to be useful.

The M type b.w.o. is thus an ideal power oscillator for use where rapid tuning or frequency modulation is required.

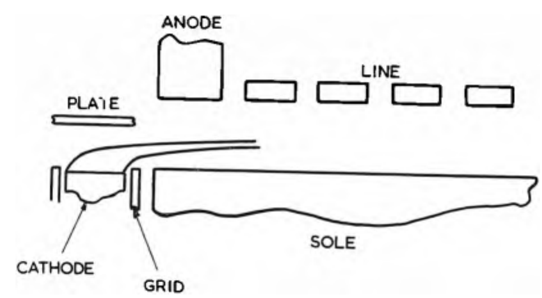


Fig. 11. Gun system for M type oscillator

By doubling the line voltage the frequency may be altered by 25 to 30 per cent if an interdigital line is employed, the variation of frequency being almost linear. Furthermore, frequency modulation with high modulation frequencies may be employed. For example in the 3 000Mc/s region deviations of 150Mc/s at a modulation frequency of 5Mc/s have been measured with little distortion.

As with the O type device there is very little change of frequency for variation of the output load impedance. For example in the 3 000Mc/s region a mismatch of 0·33:1 may only pull the frequency by 2 or 3Mc/s when moved through all phases. The figure may be considerably less than this and depends on the exact matching of the attenuator of the particular sample tested. The valve continues to oscillate if the output is completely short-circuited, but of course this condition cannot be tolerated at full power, as in this case all the power would be absorbed either in the attenuator or the output window. Tests at low power have shown that the pulling with a complete short-circuit is still only in the region of 5 to 10Mc/s for a 3 000Mc/s oscillator.

The gun system used in M type devices is completely different from the more usual O type guns. This is because of the effect of the magnetic field. The aim is to inject the electrons into the interaction space with the natural velocity of motion for that region, i.e. E/B. Electrons are accelerated from a cathode (Fig. 11) by applying a voltage to the plate immediately above it. The electrons start off normal to the cathode, but are immediately bent round in cycloidal orbits by the magnetic field. When they reach the top of their orbit they pass into the region between line and sole, where the electric field is higher, and travel approximately along an equipotential with a velocity in that direction of E/B. It should be noted that even if the velocity is not correct they will still have the same average velocity, the difference being that instead of following a linear path they will perform oscillations about that line. It is this fact which allows of a very simple gun being used, for in practice no compensation is made for variation of line voltage. The remaining electrode, the grid, consists of a lip in front of the cathode which can be biased to alter the initial direction of travel of the electrons. It will also modulate the current, but in general the plate gives better control.

The first bar of the line is called the anode and is solidly connected to the block of the valve. It serves as an earth shield to the first finger of the line and also intercepts a large number of stray electrons which might otherwise melt the first finger.

As the plate voltage is varied the line current increases if the line voltage is kept constant (Fig. 12 curve (a)). As the current increases no oscillation occurs until a critical value

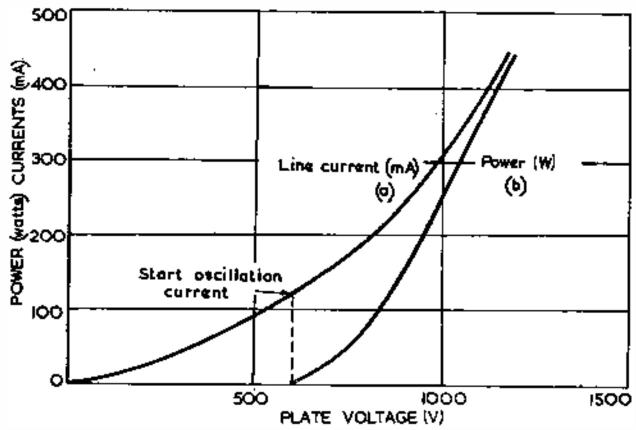


Fig. 12. Line current and power as a function of plate voltage for M type oscillator

is reached called the start-oscillation current. This is the condition in which the gain of the tube as an amplifier is exactly equal to the circuit losses. Above this current the power rises steeply (curve b). Thus it is seen that by variation of the plate voltage amplitude modulation may be obtained. As the current is increased the efficiency also increases, and in general maximum efficiency is obtained by working as far from the start oscillation current as line dissipation or cathode emission will allow.

Variation of line current is also accompanied by a small variation in frequency, the frequency increasing as the line is increased. This 'frequency pushing' as it is called is due at any rate in part to a variation of space charge density altering the electric field strength in the interaction region.

A further feature which, although common to both types of oscillator, is more frequently seen on the M type, is the occurrence of sidebands. These are not seen in the usual working range of the device, but may occur at lower voltages or higher line currents. The signal is seen to be accompanied by two sidebands displaced by a few tens of megacycles, from the carrier in the 3 000Mc/s region. This displacement remains constant when the valve is tuned. These sidebands are due to a second mode of oscillation which has a higher start-oscillation current than the main oscillation. If, however, the valve is operated at high currents this oscillation may be started. It will also beat with the main oscillation to produce the opposite sideband.

Conclusions

It is seen that the two types of backward wave oscillator, the O and M types form an unusual pair of valves having properties which are largely complementary. Each is capable of being tuned rapidly over a wide frequency range; but while the former is mainly suited to low powers because of its modest efficiency and wide variation of power over the frequency band, the latter is essentially a transmitting valve capable of delivering high mean powers with good efficiency and maintaining its power output reasonably constant over the tuning range.

These two valves are relatively new and their capabilities have not so far been fully exploited. They have however already established themselves in certain applications. Whenever a panoramic display system, such as a reflectometer or spectrum analyser, is required there is no doubt that the O type valve is supreme. It enables broad frequency bands to be covered at high sweeping speeds. Furthermore, the power output can be sampled and a signal fed back to the grid of the valve to control the beam current and hence the power output. Here the rate of variation of power with voltage sets a limit to the sweep speed and the feedback which can be applied. As oscillators with smoother characteristics become available it will be possible to operate to higher sweep frequencies.

The b.w.o. also appears attractive as a local oscillator due to the ease with which it may be tuned. So far, however, two difficulties have appeared; first there is a considerable difficulty in producing a voltage supply which is sufficiently stable to maintain the frequency constant, and it may well prove necessary to introduce a mechanically tuned cavity to which the oscillator is locked if very high stability is required. In the second place there has been difficulty with noise close to the carrier on some valves. It is well known[11] that spurious modulation of electron beams can occur at frequencies between a few kc/s and a few Mc/s due to residual gas or secondary emission phenomena. This may give rise to noise modulation which will debase the performance of a receiver unless a high intermediate frequency is employed.

The M type valve has not as yet been exploited to any extent; but it has obvious possibilities as a transmitter. For example in a simple f.m. radio link where the possibility of wide frequency modulation should allow the accommodation of many channels. Again for use in f.m. radar the wide and almost linear tuning range offers possibilities which have not been available previously. Once again the noise problem becomes important, but measurements of a.m. noise have shown that it is not excessive[7], the noise per cycle ranging from -135dB at 10kc/s to -160dB at 100Mc/s away from the carrier. Once again the challenge is to the systems designer to make use of the great potentialities of these devices.

Acknowledgments

The author is indebted to the Admiralty for permission to publish this article. Fig. 6 is reproduced by courtesy of the Mullard Research Laboratory.

REFERENCES

1. Epsztein, B., British Patent No. 699893.
2. Guenard, P., Doehler, O., Epsztein, B., Warnecke, R. Nouveaux tubes oscillateurs a large band d'accord électronique pour hyperfréquences. *Comptes Rendus Acad Sc. 235*, 236 (1952).
3. Kompfner, R., Williams, N. T. Backward-wave Tubes. *Proc. Inst. Radio Engrs. 41*, 1602 (1953).
4. Millman, S. A Spatial Harmonic Travelling-Wave Amplifier for Six Millimeters Wavelength. *Proc. Inst. Radio Engrs. 39*, 1035 (1951).
5. Sullivan, J. W. A Wide-Band Voltage Tunable Oscillator. *Proc. Inst. Radio Engrs. 42*, 1658 (1954).
6. Warnecke, R., Guenard, P. Some Recent Work in France on New Types of Valves for the Highest Radio Frequencies. *Proc. Instn. Elect. Engrs., 100*, Pt. 3, 351 (1953).
7. Warnecke, R. R., Guenard, P., Doehler, O., Epsztein, B. The 'M'-Type Carcinotron Tube. *Proc. Inst. Radio Engrs., 43*, 413 (1955).
8. Palluel, P. Recent Developpements dans le Domaine des Tubes 'Carcinotron'. *L'Onde Electrique, 36*, 318 (1956).
9. Palluel, P., Goldberger, A. K. The O-Type Carcinotron Tube. *Proc. Instn. Radio Engrs. 44*, 333 (1956).
10. Karp, A. Backward-wave Oscillator Experiments at 100 to 200 Kilomegacycles. *Proc. Instn. Radio Engrs. 45*, 496 (1957).
11. Cutler, C. C. Spurious Modulation of Electron Beams. *Proc. Instn. Radio Engrs. 44*, 61 (1956).

Triodes and Tetrodes For U.H.F.—S.H.F. Operation

By C. A. Tremlett,* B.Sc., Grad.I.E.E., and A. D. Williams,* B.Sc.(Eng.)

This review deals briefly with the advances made over the last ten years in the development of space charge control valves for operation above 300Mc/s. The improvements in technology are described with particular reference to some valve types. Two tables are included, the first giving data on valve geometry, the second summarizing general information (static characteristics, r.f. performance, etc.) regarding a wide selection of valves.

IN the last decade or so, significant advances have been made in all branches of electronic engineering, not least in the development of triodes (and some tetrodes) for use at very high frequencies. For example, the upper frequency limit for disk-seal triodes has been advanced from just over 4 000Mc/s to almost 10 000Mc/s, glass-based valves are now available which will operate efficiently at over 1 000Mc/s, and very impressive increases in power output have been reported, especially in the band 500 to 1 000Mc/s. These improvements have been obtained entirely from advances in the technology of valve making, for the valves are invariably used in conventional circuits.

From theoretical considerations it is well known[1,2] that the lead inductances and resistances, especially those to the grid, must be as low as possible to ensure electrical stability; that the inter-electrode spacings must be as small as possible in order to minimize noise and improve the high frequency performance, by reducing the time of flight of the electrons; and that the cathode must be operated at the highest current density, consistent with life, to ensure optimum efficiency. It is in the means of achieving these, and other, desirable features that these major advances have been made.

The increasing complexity of electronic equipment has called for a considerable increase in the standard of realiability demanded for all the components, and much effort has been expended on this aspect of valve design. It is not always appreciated how difficult it is even to specify requirements for life, especially for triodes, which tend to be general-purpose valves and the same type may be used in a wide variety of equipments. For example, the life and reliability requirements for valves employed in the repeater stations of a submarine cable, and in guided missiles, are quite different.

This article will deal first with some of the more important advances in valve technology, and will then review the performance of some of the resulting valves.

Developments in Technology

The Envelope

The maximum temperature to which a valve can be baked during processing is about 400°C if it employs a 'soft' glass, or about 550°C if it uses a 'hard' glass, these values being determined by the viscosity of the glass.

Considerable advances have been made in the precision with which the glass sealing operations are carried out, particularly in the case of the disk-seal valve envelopes. For example, jigs are in use which result in the planes of the electrodes being in a position at right-angles to the axis of the system to an accuracy of ±·003in per inch of radius. This may be improved by a factor of about five by subsequent machining.

If the glass can be replaced by a ceramic, the maximum baking temperature may be increased to about 750°C, and is determined by the vapour pressure of the metal used to braze the ceramic to the metal components. This higher temperature enables a much better vacuum to be obtained in the finished valve, with corresponding improvements in its life and performance. However, the greater temperature range coupled with the different non-linearity of metal and ceramic expansion coefficients make it much more difficult to match the metal and ceramic components up to the brazing temperature[3] of about 900°C.

Fig. 1. Ceramic triode valve U.L.11. (Photograph reproduced by permission of Ferranti Ltd.)

Although the pioneer work on ceramic/metal envelopes was done in Germany over twenty years ago[4], it is only quite recently that reliable solutions to some of the technical problems have been found, and several types of ceramic-envelope disk-seal valves are now becoming available, for example, the U.L.11 (Fig. 1) in this country and the 6181, 2C39B and 6BY4 in the U.S.A.

As ceramic components are conveniently ground to close tolerances, and as no distortion occurs during brazing to the metal parts, the mechanical tolerances on the valve envelope are small. This may sometimes be achieved without using jigs. In addition, the loss factor is less than for glass, an important point in the design of high power valves, and the higher outgassing temperature which may be employed with ceramic valves enables higher operating temperatures to be used[5].

As ceramic materials generally have low coefficients of expansion, and a considerably higher thermal conductivity than glass, their ability to withstand thermal shock is superior. They are also better able to resist mechanical shock and their use has made possible the construction of more robust envelopes. A typical valve using such an envelope is the U.L.11, which will operate with the anode temperature at 350°C.

New Brazing Alloys

In recent years new brazing alloys have been developed as a result of pressure from valve engineers[3]. The demand for them arises from several reasons, probably the most important

* *Research Laboratories of The General Electric Co. Ltd., Wembley.*

of which occurs when a number of successive brazes are made during the assembly of a valve. This process, known as step brazing, was, until recently, quite difficult to accomplish in view of the large difference between liquidus and solidus temperatures of some long-established alloys. In addition, these standard alloys have been found unsuitable in valves of ceramic construction as the vapour pressure of one or more constituent metals is too high. Improvements have also been made in wetting properties and alternatives have had to be found for the gold alloys, which when used to braze copper suffer the handicap of too high a rate of diffusion.

As examples of these new materials, the range of several silver-copper-palladium alloys may be quoted as having very good wetting properties. The use of indium to make ternary alloys with silver and copper, has extended the lower end of the temperature range, while similar alloys of indium, copper and gold have lower vapour pressures and similar melting points to the standard silver-copper alloys. The importance of making all vacuum-tube brazing alloys to better standards of purity has only been appreciated relatively recently. For example, traces of carbon, present in hitherto unobjectionable quantities, have been shown to cause regions of poor wetting, and more dangerously, voids in the brazing material, with the attendant danger of leaks.

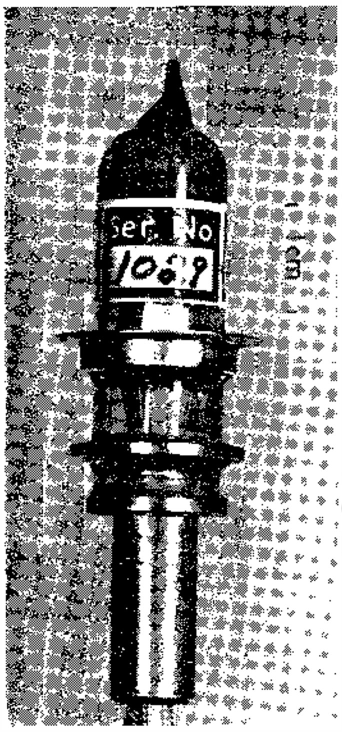


Fig. 2. Triode valve DET29

Cathodes

Of the total time an electron takes to travel from cathode to anode, the major portion is spent in moving from the cathode to the plane of the grid. This cathode grid transit time (T *sec*) is approximately given by the relationship $T = 6{\cdot}7 \times 10^{-10} (d/j)^{1/3}$ where d is the cathode grid clearance in cm and j the current density in A/cm². This clearly calls for high current density and small grid cathode clearance.

For almost twenty years prior to 1950, the conventional cathode consisted primarily of a mixture of barium and strontium oxides on a nickel base. Over this period research and development effort succeeded in explaining more or less satisfactorily the mechanism of operation of these cathodes, and as a result they are consistently produced under mass-production conditions with a high degree of uniformity of emission, following a very short period of activation. From an oxide-coated cathode a mean current density of 300 to 400mA/cm² may be taken in c.w. class-C operation, and a life of several thousand hours be obtained (e.g., DET29, Fig. 2). During the past eight years, however, several new types of cathode have been evolved and although none of these is likely to supersede the oxide cathode for general use, they are proving very useful in improving the performance of high frequency valves.

In the L cathode[6] the barium and strontium atoms reach the cathode surface by diffusing through a porous tungsten plug. This design not only removes the lossy reservoir of oxides from the r.f. field, but permits the current density from the cathode to be increased from about 350mA/cm² to about 750mA/cm². Development work is still proceeding on variations of this basic design, and at least one triode currently in production, the EC57 (Fig. 3) uses this type of cathode.

It is under pulse conditions that the significance of high current density is most apparent, an increase of 5 to 20:1 over c.w. operation being possible[7]. This may be illustrated by comparing the performance of the ACT27 (Fig. 4) with that of the geometrically identical ACT28 (CV2163).

	$V_{a\,(max)}$ (kV)	$I_{a\,(max)}$ (A)	Frequency at which efficiency has fallen to 25% (Mc/s)
ACT27			
c.w. operation	1·5	1·5	400
ACT28			
pulse operation up to 5μsec	11·0	50·0	900 (estimated)

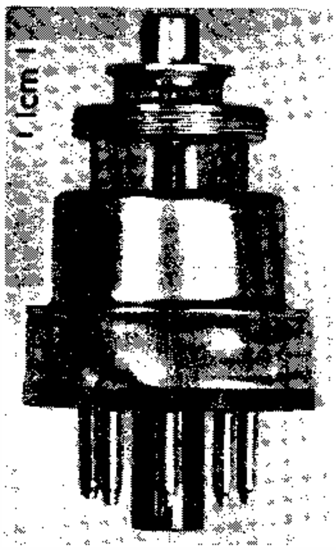


Fig. 3. Triode valve EC57 (Photograph reproduced by permission of Mullard Ltd.)

An additional advantage of high current density is that a smaller cathode area can be used to obtain the same mutual conductance, thus giving some compensation for the increased capacitance resulting from closer clearance.

While demands for increased output power have led to attempts to employ higher anode potentials, experience has shown that, for oxide coated valves, it is dangerous to exceed 1·5kV in c.w. operation. This is because the contamination of other electrodes by barium during life leads to electrical breakdown which disrupts the cathode coating. L cathodes are much more able to withstand the effects of such discharges, though the incidence of these is not materially altered by the use of such cathodes.

Advances in metallurgy have resulted in the development of sheet thoriated tungsten, and this has enabled unipotential cathodes to be made from this material for power valves, as, for example, in the Eimac 4W 20 000A. In addition, this sheet material has made the manufacture of strip filaments possible, thus leading to more efficient electron gun designs in beam power tubes. Thoriated tungsten cathodes also permit the application of high voltages on c.w. and this has been a significant factor in the development of the R.C.A. valve type[8] 6448 (Fig. 5) and its later version the 6806. Their development types [13] A2346 and A2335, which also employ such a cathode, are also able to operate at high voltages. Such valves as these owe a considerable amount to the work done on the Resnatron[9] which is a high power tetrode with an r.f. circuit constructed within the envelope. An interesting lower power transmitting valve[10] with a thoriated cathode is the TBL 2/300 which is capable of operating with V_a = 2·5kV below 200Mc/s. This voltage must be reduced by stages up to the maximum frequency of 900Mc/s due to increasing losses in the glass. The use of a high anode voltage permits designs with larger grid anode clearances, smaller output capacitances and thus larger bandwidths.

Grid Materials

The majority of the valves for use above 300Mc/s either use plane grids, in which the wires are brazed to a rigid frame, or squirrel-cage grids. In either case it is usual to use molybdenum or tungsten wire, owing to their desirable properties of good thermal and electrical conductivity, low coefficients of expansion and high hot strength. Tungsten can be drawn down to 10 micron diameter and has been electro-etched down to 3 micron diameter, while all sizes can be gold-plated before use in grid manufacture. The purpose of gold plating is to provide a surface which, at the elevated operating temperature of the wires, will absorb barium as it evaporates from the cathode and so prevent the formation of barium oxide, which may cause grid emission. The maximum temperature at which such gold-plated grids can operate is in the region of 600°C, values higher than this causing gold to evaporate with subsequent rapid loss of

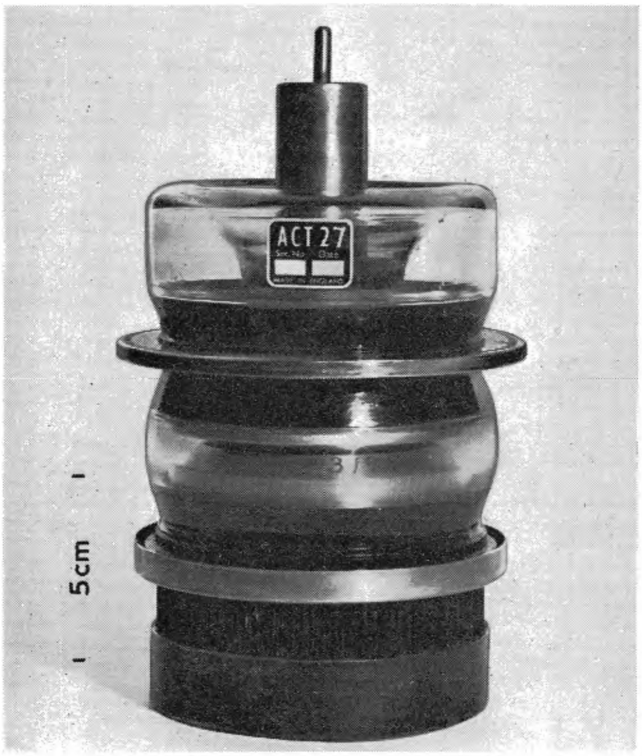


Fig. 4. Triode valve ACT27

emission. For medium power valves employing a squirrel-cage type of grid in which the length/diameter ratio of the wires is small, gold-plated copper wires have been successfully used. The dissipation may, in this way, be well over twice that which can be tolerated in a more conventional molybdenum grid for the same level of primary emission; grids such as this are employed in the ACT27 and the 6181. The use of either molybdenum or tungsten, clad with a layer of platinum of a thickness of about 10 per cent of the radius of the wire, has become almost universal for lower frequency high power valves during the past ten years, and is now being used for valves within the u.h.f. range which employ thoriated tungsten cathodes. The platinum absorbs the evaporated thorium in the same way as gold absorbs evaporated barium. The maximum temperature of operation is in the region of 1 200°C and the maximum dissipation about 10W/cm². More recently methods have been developed to increase the emissivity of the surface and have raised this figure to the region[11] of 15W/cm².

New Grid-making Techniques

The two most important mechanical features in any microwave triode are the accuracy to which the grid is made, and the accuracy to which the grid cathode clearance is maintained.

The majority of close-clearance valves use planar frame grids, and the wires are permanently stretched by inelastic deformation of the frame. In this way a very rigid structure can be made, with wire as fine as can be handled. It is not only in obtaining a reasonable accuracy of pitch at around 500 turn/in, but also in the handling of the very fine wire, of between 5 and 10 microns (0·0002 in to 0·0004 in) diameter, that difficulties arise, and an interesting new method of doing this has recently been developed[12]. At the present time the best that can be achieved in production values is a grid-cathode clearance of 15 micron using a grid wire of about 5 micron diameter.

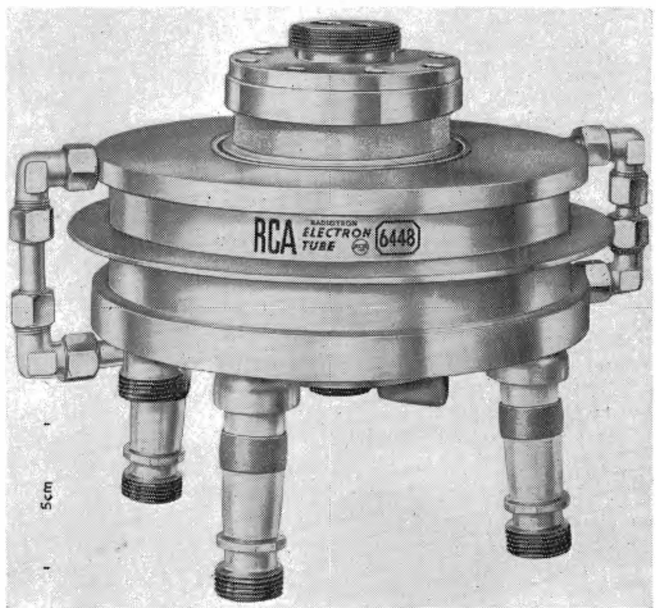


Fig. 5. Tetrode valve 6448 (Photograph reproduced by permission of Radio Corporation of America)

Recent work with cylindrical grids using a mesh formation has shown that with such a construction greater rigidity may be obtained than with the squirrel cage. It should therefore be possible to use finer wires and to make medium power u.h.f. valves with this form of grid construction.

New methods of fabricating copper and copper-alloy grids for medium power valves have been developed. For example, beryllium copper has been used by R.C.A. to form squirrel-cage grids by their 'Uniscan' process, which consists essentially of simultaneously rolling and spinning the material into grooves cut in the surface of a cylindrical mandrel parallel to the axis. This material formed into the grooves constitutes the grid wires after the web joining them has been carefully etched away in acid. The mandrel is, of course, so shaped as to leave a supporting band for the wires at each end.

The present state of constructional techniques is illustrated by Table 1, which gives the important dimensions of some valves which have appeared in recent years.

Cooling Methods

Medium power valves in the u.h.f. band rely mainly upon forced air to dissipate heat at the anode. The performance of the type of cooler fitted to valves such as the 6181 and ACT27 represents the probable limit which can be achieved with this kind of design, being in the region 1W/cm² of fin surface. The influence of such considerations as the maximum seal temperature and the intricacy of the conduction path, make this figure only a guide. In a number of the smaller u.h.f. triode coolers the fins consist of thin circular washers arranged perpendicular to the axis of the anode, this general design being considerably less efficient.

The highest power valves[13] such as the R.C.A. development types A2346 and A2335 may be taken as examples of the best loadings which have been or are likely to be achieved by water cooling of relatively large areas. The dissipation is probably in the range of 6-8kW/cm², and improvements on this are almost certainly precluded by the thermo-mechanical stresses which would be set up in the structure. With small dissipation areas of up to perhaps 5cm², as much as 10kW/cm² can be reached[14].

Life Considerations

Satisfactory emission may be obtained from cathodes with a barium surface for many thousands of hours provided reasonable care is exercised during both valve manufacture and valve operation. However, barium is deposited on the grid wires by evaporation from the cathode, increasing the wire diameter and thus changing the valve characteristics. While this effect is negligible in most valves, it becomes an increasingly serious problem as the diameter of the grid wire is reduced, and leads to failure considerably before the end of emission life. The difficulty can be reduced to some extent by careful choice of operating conditions[15], but it is probable that this grid wire growth, rather than mechanical difficulties, will determine the ultimate limit for grid wire diameter. For example, the use of 10 micron grid wire forces the valve manufacturer to give the problem close attention, and it is unlikely that grid wire finer than 5 micron can be used in a sound design.

As far as manufacture of microwave triodes is concerned there is no doubt that the establishment of dust-free conditions is a *sine qua non* of success. In parallel with the measures taken to achieve these conditions[16] the use of ultrasonic cleaning is now steadily growing and is likely to have the widest application.

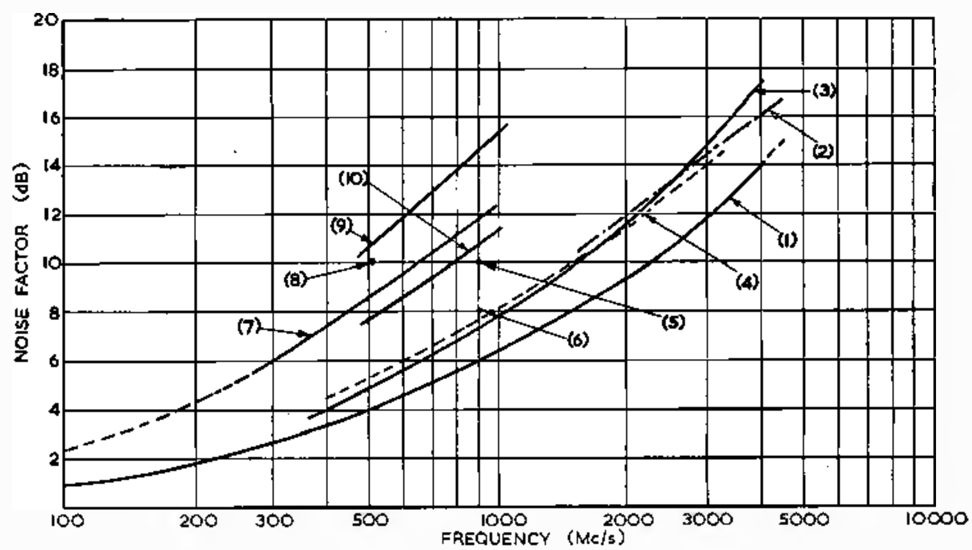


Fig. 6. Variation of noise factor with frequency

(1) 416A and 416B disk-seal triodes
(2) DET29 disk-seal triode
(3) EC57 disk-seal triode
(4) GL-6299 ceramic disk-seal triode
(5) EC56 disk-seal triode
(6) 6BY4 ceramic disk-seal triode
(7) A2521 glass-based valve
(8) CV354 disk-seal triode
(9) 6AJ4/6AM4 glass-based valves
(10) 6BA4 disk-seal triode

In addition, new assembly techniques are being developed. For example, the use of argon-arc welding and high speed r.f. brazing considerably reduces the dangers of contamination which are inherent in the older methods of torch brazing and glass sealing.

TABLE 1
Some Electrode Dimensions of Selected Microwave Valves

VALVE NO.	GRID/CATHODE CLEARANCE (MICRON)	GRID WIRE DIAM. (MICRON)	GRID (TURNS/IN)	CATHODE CURRENT DENSITY (A/Cm²)	
A2521	40	10	506	0·15	Glass Base Valves
6AJ4 / 6AM4	51	20	250	0·2	
416B	15	7·5	1000	—	Low Power Disk-Seal Valves
6BY4	15	7·5	1000	0·20	
DET29	35	10	500	0·32	
6BA4	37	20	360	—	
EC57	40	7·5	530	0·72	
UL11	125	50	Mesh	0·15	
CV354	200	30	240	0·20	
2C39B	80	50	170	0·2	Medium and High Power Disk-Seal Valves
4X150G / 4CX150G *	250	120	39	0·1	
ACT27	600	300	24	0·075	
6448*	500	90	36	0·4	

* *Tetrode.*

Application Considerations

Triodes for u.h.f.-s.h.f. service are usually considered to be general-purpose devices and the present tendency is to publish information covering as many applications as possible extending over the widest frequency range. The need to do this arises from the fact that in most cases it is difficult to calculate the r.f. performance at all accurately, though a fair estimate can be obtained[17,18].

Noise Factor

For a number of well-known reasons[19,20] the significance of noise factor has assumed more and more importance, and the above remarks concerning the difficulty of accurate calculation, particularly apply. The valve design factors which tend to improve noise factor are similar to those which lead to good r.f. performance in general. The measurement of the noise factor of triodes in the u.h.f. region is usually carried out using noise diodes, and significant discrepancies occur between the published results of different workers. A great deal of effort has been made to reduce these discrepancies[20, 21]. Noise generators have been extensively

TABLE 2
Performance Data on Selected Space Charge Control Valves for Operation in the Microwave Frequency Range

VALVE TYPE	TYPE OF CATHODE	PULSE VOLTS	MAXIMUM ANODE RATINGS			MUTUAL CONDUCTANCE (mA/V at Ia mA)		μ	CAPACITANCES (pF)			FREQUENCY (Mc/s)	OSCILLATOR OUTPUT WATTS	SMALL SIGNAL AMPLIFIER		LARGE SIGNAL AMPLIFIER		OUTPUT WATTS
			D.C. VOLTS	D.C. mA	WATTS				Cgk	Cga	Cak			BANDWIDTH (Mc/s)	GAIN (dB)	BANDWIDTH (Mc/s)	GAIN (dB)	
6AJ4 Glass-based low noise triode c/c	Oxide	—	150	20	2·0	10·0	16	42	4·4	2.8	0·16	500 900	—	10 10	8·3 7·2	—	—	—
6AM4 Glass-based low noise triode c/c	Oxide	—	200	20	2·0	9·8	10	85	4·6	2·8	0·16	500 900	— —	10 10	7·2 7·0	—	—	—
A2521 Glass-based low noise triode c/c	Oxide	—	250	20·0	2·5	15·0	15	66	3·5	1·6	0·06	500 600 900 1250	0·7 — — 0·15	— 10 10	— 18·0 14·0	—	—	—
6AF4 Glass-based oscillator triode c/c	Oxide	—	150	28	2·25	6·6	16	15	2·2	1·9	0·45	500 900	0·4 0·3	—	—	—	—	—
EC81 Glass-based oscillator triode c/c	Oxide	—	170	30	5·0	5·5	30	16	1·7	1·5	0·5	500 1000	3·0 1·25	—	—	—	—	—
6BY4 Ceramic disk-seal low noise triode c/c	Oxide	—	300	5·5	1·1	6·5	5·0	100	2·0	0·7	0·007	900	—	10	15	—	—	—
GL6299 Ceramic disk-seal low noise triode c/c	Oxide	—	200	11·0	2·0	12·0	10	115	3·5	1·7	0·01	1200 3000	— —	10 10	16·0 11·0	— —	— —	— —
416B Disk-seal triode c/c	Oxide	—	270	30	7·5	50·0	30	200	11·1	1.45	0·019	4000 4000	— —	— —	— —	100 70	10·0 5·5	0·065 0·57
EC57 Disk-seal triode c/c	Lemmens	—	300	70	10·0	20·0	60	43	3·3	1·6	0·04	4000	3·0	—	—	100	4·0	1·5
DET29 Disk-seal triode c/c	Oxide	—	400	40	10·0	16·0	10	65	3·4	1·2	0·03	2300 4000	3·2 2·1	50 50	14·0 12·5	50 50	7·5 4·5	1·75 1·0
UL11 Ceramic disk-seal triode c/c*	Oxide	5000	500	200	30	14·5	50	19	6·5	5·5	—	400 1000	5·5 3·5	— —	— —	— —	— —	— —
2C39 Disk-seal triode forced-air-cooled	Oxide	—	1000	125	100	22·0	125	100	6·5	1·95	0·035	500 2500	— 15·0	— —	— —	— —	6·5 —	27 —
TBL2/300 Disk-seal triode forced-air-cooled	Thoriated Tungsten	—	1750 at 470 Mc/s 1300 at 900 Mc/s	400 — —	260 at 470 Mc/s 300 at 900 Mc/s	18·0	—	32	9·0	4·0	0·12	470 470 900	405 — 155	—	—	20 20	12·0† 5·0‡	—
LD9 Ceramic disk-seal triode forced-air-cooled	Oxide	—	1500	175	300	23	100	—	9·0	3·0	0·025	1700 3000	40 15	—	—	—	—	—
ACT27 Disk-seal triode forced-air-cooled	Oxide	—	1500	1500	1500	40	500	50	30	28·0	0·5	400	900	—	—	—	—	—
ACT28 Pulse version of ACT27 forced-air-cooled	Oxide	11000	—	50000 (Pulse)	1500	—	—	—	—	—	—	600	250000 (Pulse)	—	—	—	—	—
									C_{IN}	C_{OUT}	C_{a/g_1}							
4CX150G Ceramic and 4X150G glass tetrode disk-seal forced-air-cooled	Oxide	— 7000	1250	250 6000 (Pulse)	150	12·0	200	—	16·0	4·5	0·03	750 1200	100 20000 (Pulse)	—	—	—	—	—
4CX300A General-purpose ceramic tetrode forced-air-cooled	Oxide	—	2000	250	300	12·0	250	—	29·5	4·6	0·035	500	225	—	—	—	21·0	410
6181 Disk-seal tetrode forced-air-cooled	Nickel Mush	—	2000	1750	2000	—	—	—	44	22	0·1	900	—	—	—	—	8·0	1200
6448 Water-cooled power beam valve	Thoriated Tungsten	—	7000	7000	26000	—	—	—	335	30	0·1	400 900	— —	— —	— —	— —	15·5 11·0	14000 11000
A2346 Development triode water-cooled	Thoriated Tungsten	—	—	—	$0·9 \times 10^6$	—	—	—	—	—	—	600 400	600000 25×10^6 (Pulse)	—	—	—	—	—
A2335 Development triode water-cooled	Thoriated Tungsten	—	—	—	120000	—	—	—	—	—	—	900 1000	80000 5×10^6 (Pulse)	—	—	—	—	—

* Increased performance can be obtained with forced air cooling. c/c=Convection/Conduction Cooled.
† Grounded Cathode. ‡ Grounded Grid.

investigated as they are the reference standards by which the noise level in valves is determined. Of the two types of generator used, the discharge tube has been employed successfully in the microwave region[22,23], while coaxial noise diodes such as the CV2341 are used up to 1 000Mc/s. The published variation of noise factor with frequency on a number of available receiving valves is shown in Fig. 6. As far as can be ascertained the noise factors at frequencies below 1 000Mc/s have been measured using diode noise generators and those at higher frequencies using discharge tube generators. It is assumed that the valves were measured under optimum running conditions and that the figures quoted apply to the first stage alone.

CIRCUITS

As the frequency rises it becomes impossible to consider valve and circuit separately, and the larger the valve, the lower the frequency at which this holds. The great majority of circuits used in the u.h.f.-s.h.f. band are of the common grid type. For small glass-based valves 500Mc/s represents the approximate limit at which lumped capacitances and inductances may be used with reasonable efficiency. In recent years good performance has been achieved with valves such as the A2521 (Fig. 7) up to 1 000Mc/s when operated in trough-line circuits.

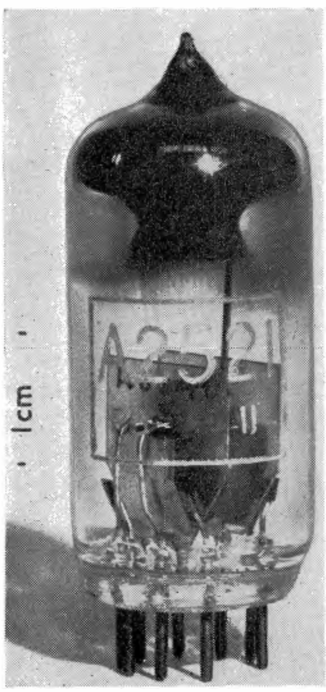


Fig. 7. Triode valve A2521

The design of resonators and coaxial lines used with disk-seal valves depends upon the particular application[24]. Quarter-wave mode operation is always desirable, but its mechanical realization is not always possible at the higher frequencies. Higher modes of operation have the disadvantage of reduced bandwidth and tend to increase circuit losses.

A number of valve properties make it difficult to calculate circuit dimensions accurately. For example, the metal components of the envelope and the internal electrode support structures become part of the circuit, thus increasing its length. The capacitance loading is then considerably less than the static capacitance and approaches that of the working area alone. This point is illustrated by data[25] on the EC57.

	Static capacitances (pF)	*H.F. capacitances (pF)*
C_{gc}	3·3	1·8
C_{ag}	1·6	0·6

Also the configuration of the envelope introduces electrical discontinuities, by virtue of both geometry alone and the presence of glass or ceramic insulators. Typical examples of modern u.h.f.-s.h.f. circuits have been described for use with the 416A[26], the 6448[8] and the TBL 2/300[10]. The first of these illustrates how a number of such circuits can be used in cascade to give high gain, while the second employs an interesting method of overcoming the problem of mismatch between valve and circuit at high power level. It has been shown possible to increase the power which may be obtained from a particular valve type by employing a number of them in parallel in an annular circuit[28,29].

Future Developments

At the end of the article on valves in the 1950 v.h.f. issue of this journal[29], it was suggested that triodes would probably be made eventually to operate down to a wavelength of 3cm. This has now been achieved by the most advanced of the small disk-seal valves described here, although the limit of their useful power output is probably at about 4·5cm.

The lower limit of grid wire diameter which can be used appears to be in the region of 3 microns, and if grids with from 1 000 to 2 000 turns/in can be successfully made with this diameter of wire, then a further step forward in s.h.f. performance will be made.

The use of multiple beam systems has, in this period resulted in about a thirtyfold increase in power level in the region of 1 000Mc/s, while the ratio at somewhat lower frequencies has been even more impressive. It is likely therefore that the same design principles, when applied to valves for operation at higher frequencies in the u.h.f. band will lead to further increases in r.f. power output.

Acknowledgment

The authors wish to thank the M.O. Valve Company Limited for permission to publish this review.

REFERENCES

1. BECK, A. H. W. Thermionic Valves. (Cambridge University Press, London, 1953.)
2. BENHAM, W. E., HARRIS, I. A. The U.H.F. Performance of Receiving Valves. (MacDonald, London, 1957.)
3. KOHL, W. H. Materials Technology for Electron Tubes. Ch. 16 (Rheinhold, New York, 1951.)
4. ESPE, W., KNOLL, M. Werkstoffkunde und Hochvakuumtechnik, Ch. 25. (J. Springer, Berlin, 1936.)
5. BEGGS, J. E., The Use of Titanium Metal in Vacuum Devices. *I.R.E. Trans. Electron Devices, ED-3*, No. 2, 93 (1956).
6. LEMMENS, H. J., JANSEN, M. J., LOOSJES, R. A New Thermionic Cathode for Heavy Loads. *Philips Tech. Rev. 11*, 341 (1950).
7. WRIGHT, D. A. A Survey of Present Knowledge of Thermionic Emitters. *Proc. Instn. Elect. Engrs. 100*, Pt. 3, 125 (1953).
8. BENNETT, W. P. New Beam Power Tubes for U.H.F. Service. *I.R.E. Trans. Electron Devices ED-3*, 57 (1956).
9. MCCREARY, R. L., ARMSTRONG, W. J., MCNEES, S. G. An Axial-Flow Resnatron for U.H.F. *Proc. Inst. Radio Engrs. 41*, 42 (1953).
10. PAPENHUIZEN, P. J. A Transmitting Triode for Frequencies up to 900Mc/s. *Philips Tech. Rev. 19*, 118 (1957-58).
11. ESPERSON, G. A., ROGERS, R. W. Studies on Grid Emission. *I.R.E. Trans. Electron Devices ED-3*, 100 (1956).
12. WIDDOWSON, A. E., GORE, D. C., BUTCHER, C. H. A New Method of Making Accurate Fine-Wire Grids for Use in Radio Valves. (*To be published.*)
13. KLEEN, W. On the Present Limitations of Electron Tubes with Respect to Power Frequency and Noise. *Elektrotech. Z. 77*, 769 (1957).
14. HICKEY, J. S. Heat Transfer at High Power Densities. *J. Appl. Phys. 24*, 1312 (1953).
15. METSON, G. H. A Study of the Long-Term Emission Behaviour of an Oxide Cathode Valve. *Proc. Inst. Elect. Engrs. 102B*, 657 (1955).
16. JENKINS, D. E. P. Clean Rooms for Special Valve Assembly. *Instrum. Pract. 8*, 1085 (1954).
17. POHL, W. J., ROGERS, D. C. U.H.F. Triodes. *Wireless Engr, 32*, 47 (1955).
18. GIACOLETTO, L. J., JOHNSON, H. U.H.F. Triode Design in Terms of Operating Parameters and Electrode Spacings. *Proc. Inst. Radio Engrs. 41*, 51 (1953).
19. THOMSON, J. Radio Communication at Ultra-High-Frequencies. (Methuen, London, 1950).
20. VAN DER ZIEL, A. Noise. (Chapman and Hall, London, 1955.)
21. HOULDING, M. Valve and Receiver Noise Measurement at V.H.F. *Wireless Engr. 31*, 15 (1954).
22. HUGHES, V. A. Absolute Calibration of a Standard Temperature Noise Source for Use with S-Band Radiometers. *Proc. Instn. Elect. Engrs. 103B*, 669 (1956).
23. SUTCLIFFE, H. Noise Measurements in the 3cm Waveband Using a Hot Source. *Proc. Instn. Elect. Engrs. 103B*, 673 (1956).
24. GRIFONE, L. Dimensionsanento della Cavita Coassiali per i Tubi ad Elettrodi Piani. *Alta Frequenza 23*, 357 (1954).
25. DIEMER, G., RODENHUIS, K., VAN WIJNGAARDEN, J. G. The EC57, A Disk-Seal Microwave Triode with L-Cathode. *Philips Tech. Rev. 18*, 317 (1956-57).
26. BOWEN, A. E., MUMFORD, W. W. A New Microwave Triode, Its Performance as a Modulator and as an Amplifier. *Bell Syst. Tech. J. 29*, 531 (1950).
27. PRIEST, D. H. Annular Circuits for High Power Multiple Tube R.F. Generators at V.H.F. and U.H.F. *Proc. Inst. Radio Engrs. 38*, 515 (1950).
28. SAYER, W. H., MEHRBACH, E. U.H.F. Transmitter Uses Beer-Barrel Cavity. *Electronics 24*, 125 (Dec. 1951).
29. A Survey of V.H.F. Valve Developments. *Electronic Engng. 22*, 310 (1950).

Ferrite Components in Microwave Systems

By B. L. Humphreys*, B.Sc.

This article considers some of the many ways in which ferrite components are used in microwave communication systems, indicating present possibilities and limitations.

THE development of microwave ferrite components has given the systems engineer a range of passive circuit elements having properties which were previously unobtainable except by using active elements and mechanical components, or which could not be obtained at all. These circuit elements lead to techniques of great value both in experimental work and in engineered equipment.

The outstanding feature of microwave ferrite components is that, although they are passive elements they can have non-reciprocal characteristics. It must be appreciated, however, that not all microwave ferrite components have non-reciprocal properties and in fact a reciprocal characteristic may be desirable in some cases.

The most commonly used microwave ferrite elements such as the isolator or the circulator have a constant applied magnetic field, but by using a variable magnetic field elements for switching, modulating, a.g.c., a.f.c. and other purposes can be designed[1].

Although much of the experimental work on microwave ferrite components has been done in the 10 000Mc/s region, the results are applicable in principle at all frequencies. Components have been made at 50 000Mc/s and at 200Mc/s, and the frequency limits within which components can usefully be made are probably much wider than this.

Ferrite components have not only been made in hollow waveguide systems but isolators based on strip-lines and coaxial lines have also been constructed.

Even though the full scope of passive ferrite devices has not yet been explored, active elements such as amplifiers and mixers are also being developed. The reported results indicate that, in some respects, these new devices are superior to existing valves, through their lower noise figure, or their higher power handling capabilities. This development is, however, at a too early stage for a full assessment to be made and a better appreciation of their possibilities will only be obtained after more widespread use of these elements.

Components With Constant Applied Field

The Gyrator

The first non-reciprocal passive element was termed the gyrator and was introduced in a theoretical paper by Telegen[2] as a new circuit element. It has the property of transmitting a signal unchanged in one direction of propagation, but reversing its phase for the other direction (Fig. 1).

In practice this property can be achieved at microwave frequencies only in association with a reciprocal phase shift. This constant term, however, does not affect the essential performance of the component and it is on this element that the circulator is based. It may be said, in fact, that the design of all non-reciprocal passive elements may be based on the gyrator.

* *Mullard Research Laboratories*

The Isolator

This circuit element is probably the most commonly used of all microwave ferrite components. It is a two port, four terminal network, Fig. 2, with non-reciprocal transfer impedances. Its basic property is that the attenuation between ports is dependent upon the direction of propagation. Ideally the isolator has zero attenuation for one direction of propagation and infinite for the other, while maintaining the correct input and output impedances. In practice these ideals are by no means obtained, and the choice of isolator must be determined by the application requirements.

The isolator may be of two types, those which absorb power within the ferrite itself, such as the resonance absorption type[3] and those which divert power into an additional load, such as the field displacement isolator[4] or those based on the circulator[5].

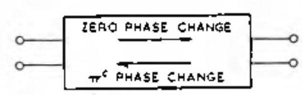


Fig. 1. The gyrator

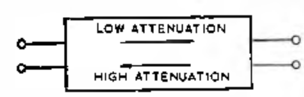


Fig. 2. The isolator

Their principal use derives from their property to absorb the reflected power without attenuating the incident power. Thus a power source working into an isolator will have a substantially constant load independent of changes in the system. Isolators have been designed to operate at all frequencies between 200Mc/s and 70 000Mc/s and a typical isolator at X-band will have an insertion loss of less than 1dB and an isolation of greater than 20dB, for a 5 to 10 per cent bandwidth, although very much better performances have been quoted[6]. Such an isolator reduces the incident power by only 20 per cent and yet changes a reflection coefficient of unity to 0·1, i.e. a zero v.s.w.r. becomes 0·82, assuming the isolator itself does not introduce reflection.

In the laboratory this non-reciprocal property is of great value in the protection of stable power sources, as the frequency and amplitude of the signal from most microwave sources is dependent to more or less extent on the load into which they are operating. The same degree of stability can of course be achieved by using a reciprocal attenuator of one half the isolation value of the isolator, but the available output power is then reduced by this amount and with this, the sensitivity of the system. The use of an isolator to increase the power available from a signal source may result in a great saving of size, weight and cost. When the power requirements are such that attenuator pads can be used, the degree of isolation can be increased until the insertion loss of the isolators is equal to that of the pad. If for example a 6dB pad is used, so that the total isolation is 12dB, isolators of the performance quoted above could be used to give an isolation of 120dB without reducing the forward going power level.

The reduction of the reflected power will reduce the frequency pulling effect of aerial systems on magnetrons or klystron oscillators, especially in the case of scanned aerials, where changes in the reflection coefficient will result in a frequency modulation of the transmitter. Frequency control either by electrode potential or by cavity tuning is simplified by isolators, as the oscillator then operates into a substantially constant and nearly resistive load and unwanted frequency pulling effects of the system are practically eliminated.

In a communications system where microwave amplifiers may be cascaded, each stage may be fed from and into substantially matched terminations by the use of isolators between stages. This is of particular application to microwave triode amplifiers where there is considerable coupling between output and input circuits.

The isolator also enables two power sources to be coupled together in a system, without the one affecting the performance of the other. This has been used to great effect in a swept frequency signal generator, where the output signal was obtained by mixing two microwave signals.

Isolators are also valuable in systems when it is necessary to keep reflections to a minimum, such as in communication systems, where time delays in long waveguide runs may lead to channel intermodulation. In this application the isolator itself must be well matched as it is clearly of no value to absorb one reflection, only to replace it by another and possibly larger one.

The considerations so far have paid no attention to power levels but apart from electrical breakdown within the guide, changes in the performance of the isolator can occur with temperature. Isolators for high power use are usually different in design from those for low powers, the difference arising from the need to keep the temperature of the ferrite as constant as possible. In this context low powers refer to levels of up to a few watts. The problem of heating is reduced either by obtaining the best possible contact between the ferrite and the heat sink, as in the high power resonance isolator, or by air cooling in the Faraday rotation circulator where the ferrite is supported centrally along the axis of the waveguide.

At X-band isolators have been made which will handle peak powers of the order of 250kW and absorb mean powers of the order of 100W.

The degree of isolation required clearly depends on the application, and one figure of merit of an isolator is the ratio of the isolation to the forward insertion loss both expressed in decibels. While 20 to 30 are typical values for this ratio, isolators in which the ratio is over 200 have been constructed. This increase in the ratio is often accompanied by a decrease in the bandwidth and an increase in the sensitivity to external influences. The bandwidth of an isolator has been defined as the frequency range over which the isolator has greater than one half the peak attenuation, but once again it is the application that determines its usefulness, the specification usually being that a certain minimum performance over the entire frequency range is required.

The Circulator

The circulator is not analogous to any simple electronic system and is a multi-port component in which power may be transmitted unidirectionally between consecutive ports, while maintaining a high degree of isolation between all other pairs of ports. The three and four port circulators Figs. 3 and 4, are the simplest forms, from which the multi-port circulator can be derived by cascading. It may be noted that each additional four port circulator adds two more ports per element whereas the three port circulator only adds one, and that a three port circulator can be derived from the four port element by short-circuiting the unwanted port.

The most commonly used circulator is the four port component based either on Faraday rotation or differential phase shift. The former have been constructed with low insertion loss, less than $\frac{1}{4}$dB, and with high isolation through unwanted paths, greater than 40dB for bandwidths of about 1 per cent at 10 000Mc/s. The phase shift circulator, appears to be inherently of a broader band nature,

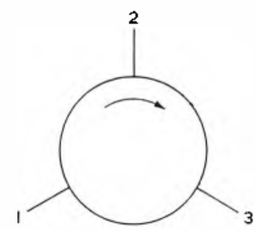


Fig. 3. Three port circulator

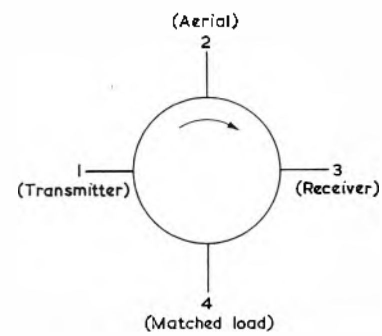


Fig. 4. Four port circulator showing its use for duplexing

5 per cent at X-band but quoted figures suggest the isolation is reduced to between 20 and 30dB. A broad-band version of the Faraday rotation circulator[7] has also been made with a bandwidth of about 10 per cent at X-band for a 30dB isolation.

The principal use of the circulator is in systems where it is required to receive power into a two terminal network from one part of a system and to transmit power to another part of the system, from the same two terminal network, without coupling the two parts of the system. A particular instance of this is the Maser amplifier when using the same termination for both input and output[8].

Similarly the circulator is used for duplexing[9] in either c.w. or pulsed systems. As is apparent from Fig. 4 it provides paths of low attenuation between the transmitter and the aerial, and between the aerial and the receiver, yet giving a high degree of isolation between the transmitter and the receiver. While the isolation due to the circulator may be between 30 and 40dB, the isolation is limited in practice by the aerial mismatch which may easily result in reflected powers greater than −20dB relative to the transmitter power. Further discussion on duplexing is included in the section on switching.

As indicated previously a circulator may be used as an isolator by terminating the two additional ports, 3 and 4 of Fig. 4, with matched loads. The power handling capacity of the circulator may then be limited more by the loads than by the circulator itself. Once again operation at high power raises the problem of change of performance, but a four port circulator of the Faraday rotation type has been made with a through-put of 100W mean at X-band. The phase shift type of circulator should be capable of handling higher powers as the problem of cooling the ferrite is simpler.

The present lower frequency limit on the circulator is in the region of 3 000Mc/s. This limitation arises not only from the difficulty of obtaining suitable ferrites but also from the problem of physical size.

Components With Varying Magnetic Fields

An extension of the application of constant magnetic field ferrite components lies in the use of varying applied magnetic fields, and the possibility of components for switching and for modulation is at once apparent. By varying the magnetic field of a resonance absorption isolator a variable attenuator results. By switching the direction of the applied field in a circulator a microwave switch is obtained. Many such components can be devised and have been described in the literature.

Components with varying magnetic fields frequently operate through low and zero applied fields and it may be important that magnetic microwave losses in the ferrite material shall still be small. It becomes necessary in these circumstances to use materials which do not exhibit low field loss. This is a function of frequency and for a given material disappears at a sufficiently high frequency. Suitable materials do exist for frequencies as low as 3 000Mc/s and therefore components may be designed for frequencies at least as low as this.

Switches

Switches may be of several types, those which either transmit power or absorb it, those which transmit power or reflect it and those which change the path of transmission.

A switched resonance isolator would be of the first type and a switched circulator would be of the third type. The design of the second type can be based on Faraday rotation, but instead of the 45° angle of rotation used in the circulator 90° is used instead[10,11].

The applications of microwave switches are many and the use of switched magnetic fields enables the switching time to be many orders of magnitude less than is obtainable mechanically. The order of magnetic field required in an X-band Faraday rotation circulator is tens of oersteds, that required for resonance isolators is thousands of oersteds. It is clearly more practical to make use of the Faraday rotation effect for switching than the resonance absorption effect.

Such a switch has been made[15] for X band with a rejection of about 35dB and an insertion loss of less than 1dB. The bandwidth of the switch was only a few tens of megacycles per second but had switching times of less than 0·1μsec and could be switched at repetition rates of over 1Mc/s.

A switched circulator can clearly be used to transmit signals through alternative paths such as to one or other of a pair of aerials, or to obtain a pulse modulated signal.

The Phase Shifter

The original non-reciprocal passive element, the gyrator, is a particular case of a non-reciprocal phase shifter giving a differential phase shift of π^c. Ferrite phase shifters can be reciprocal or non-reciprocal and for a given magnetic field and size of material the differential phase change and the total phase change are functions of position. As with other ferrite components, the design is dependent upon the application and the optimum configuration for differential phase shift is not necessarily the optimum for maximum unidirectional or total phase shift.

At X-band variation in phase shift with magnetic field of the order of $2\pi^c$ per inch with insertion loss of less than 0·5dB, and differential phase changes of $\pi/2^c$ per inch with less than 0·2dB insertion loss, are obtainable.

The $\pi/2^c$ differential phase shift has been used in the design of circulators, while the larger variable phase change has been used in the tuning of cavities and for modulation.

Modulation

Phase Modulation

The simplest form of modulation to achieve with ferrite devices is phase modulation, the phase shift through ferrite loaded guide being a function of the magnetic field. The most attractive phenomenon on which to base such a modulator is Faraday rotation as the magnetic fields required are small. One form of a phase modulator consists of a pair of identical Faraday rotation elements in cascade, but

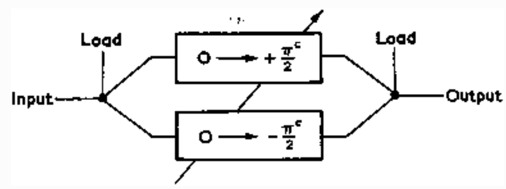


Fig. 5. Amplitude modulator using coupled variable phase changers

having contrary directions of rotation[16]. In this way the resultant rotation is zero but the phase change is directly related to rotation recurring in each part. This technique results in phase modulation with substantially no amplitude modulation and which is sensibly linear with the applied magnetic field. The maximum modulation angle will be determined by the saturation rotation of the ferrite and the degree of non-linearity permitted. The saturation rotation is a function of the size of ferrite and may change rapidly with diameter.

Amplitude Modulation

The problem of amplitude modulation is by no means so easily solved as a variation of magnetic field results in a change of phase shift as well as other wanted effects. A resonance isolator with a varying magnetic field will give phase shift. The output from one arm of a circulator is proportional to the sine or cosine of the angle of rotation but as previously stated this is accompanied by a phase change. This unwanted phase change can be removed by a coupled phase changer, but it clearly depends on the application as to whether this added complexity can be tolerated.

An alternative arrangement is to use a pair of phase shift sections coupled by a hybrid-T, Fig. 5. Here the output is proportional to the sine or cosine of the change in phase and is not phase modulated. In both systems the modulation is non-linear with phase angle and with the applied magnetic field, so except for small modulation amplitudes correcting networks will be required.

Single Sideband Modulation

A component has been described by Cacheris[12] which is used for displacing the input frequency. This displacement is achieved by a rotating magnetic field and the displacement frequency is twice the rotation frequency. As described, the component is narrow band, a few per cent at X-band, and gives a carrier and undesired sideband suppression of greater than 30dB. Frequency displacements of over 20kc/s have been reported.

Valve Tuning

The variation of phase change of a ferrite loaded section with applied magnetic field leads to the possibility of cavity tuning and thence to the tuning of microwave valves[9].

In principle the variable phase shifter takes the place of a mechanical tuner and at X-band a klystron with an 800Mc/s mechanical tuning range to half power points has been tuned over 600Mc/s with a ferrite phase shifter in similar waveguide structures. The output power in the latter condition was within 1dB of that obtained with the mechanical tuner.

The immediate advantage of ferrite tuning is the increased sweep rate over mechanical methods and the increased frequency sweep over reflector tuning. The system lends itself to the design of a rugged tunable signal source as it involves no moving parts. Care must, however, be taken to ensure that the magnetic field control system does not introduce unwanted variation.

Magnetic field control does not lend itself to modulation at high frequencies and unless the required frequency sweep is larger than is obtainable with the reflector potential, ferrite tuning does not offer any real advantage. However, for frequency modulation involving wide frequency sweep, ferrite tuning offers a useful possibility.

Automatic Gain Control

The possibility of a magnetically controlled microwave attenuator is at once of importance in microwave a.g.c. systems. This component could make use of low field loss which decreases with increasing applied magnetic field. A variation of up to 30dB without prohibitive residual attenuation has been reported. The use of a Faraday rotation or phase change system for controlling amplitude would give up to 30dB of attenuation with a residual loss of less than 0·5dB.

Aerial System

The use of ferrite rods for microwave aerial system has been described in a paper by Reggia, Spencer, Hatcher and Tompkins[18], which shows the advantage to be gained in size, over the conventional dish reflector system. They show that narrow beam-widths with low level side lobes are obtainable. The feeder system is more complex than for the conventional array but does lend itself to scanning by electrical means instead of mechanical, with the consequent possibility of very fast scanning rates.

Filters

The use of ferrites for phase changers and cavity tuning results in its use for tunable microwave filters. The filters may be non-reciprocal and contain resonant circuits which have resonance frequencies dependent upon the direction of propogation. In his paper on the subject N. C. E. Nelson[17] describes a passive duplexer, which not only has the behaviour of a circulator, but also protects the receiver from signals outside a certain frequency range. The order of magnitude of the change in the resonant frequency of an X-band cavity which he reports is a few hundred megacycles per second.

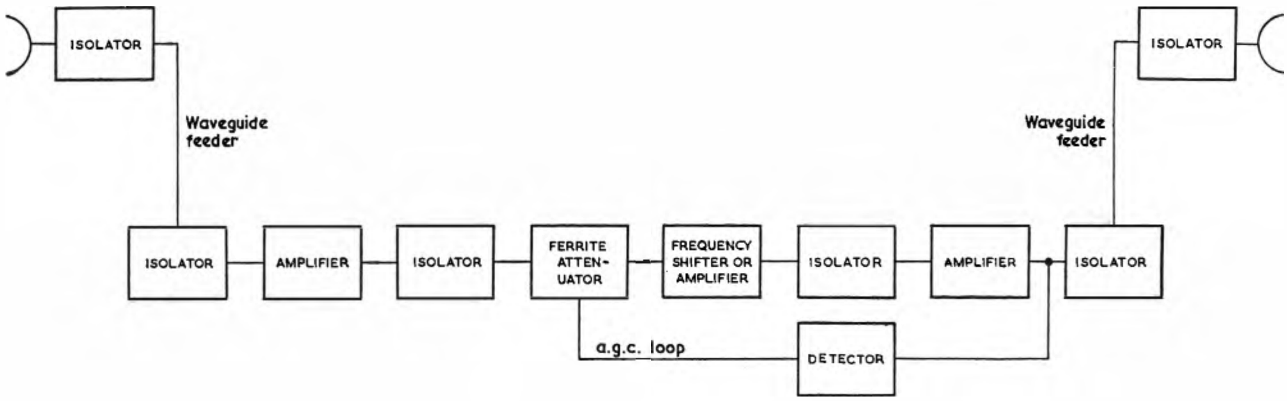


Fig. 6. Schematic of communications relay

Noise Properties of Ferrite Components

The non-reciprocal nature of some ferrite components has given rise to suggestions that the noise associated with them may be more or less than that associated with the equivalent attenuator. It may be argued on thermodynamical grounds, however, that as the component is a passive element the noise properties must be those of resistive attenuators. The non-reciprocal nature of some ferrite components does however, lead to the result that noise radiation due to the ferrite itself may be different in the two directions of propagation. This is, however, compensated by similar differences in the noise absorption.

Active Ferrite Components

The more recent developments in the field of ferrite components have resulted in the use of ferrites for amplication[8], mixing and detection[19] of microwave signals. Of these the amplifier has received the greatest attention and noise figures of 3dB at room temperature have been quoted for an X-band amplifier. This noise figure drops to 0·06dB at 4°K. The operating temperature of the amplifier extends from the lowest temperatures obtainable to several hundred degrees centigrade. The noise improvement over existing valves is clearly of importance in all fields of communication where a high sensitivity is required.

The sensitivity of the mixer and the detector is less well established, but present results are promising and indicate that further investigation should be profitable.

Conclusions

The use of ferrite components has resulted in the realization of microwave systems otherwise impossible to design, and can improve the performance of others. Ferrite components such as isolators and circulators are now well established in their use in equipment as well as in the laboratory but the capabilities of other components have still to be fully ascertained. The limitations of these components are not well established and a full appreciation of their uses can only come through their widespread employment in microwave systems.

In principle the techniques which result from their use are not frequency dependent, but the range of available ferrites imposes limits, particularly where the material is used in an unsaturated condition.

The presently available ferrite components, isolators, circulators etc. are all passive elements but active ferrite components are being developed which give advantages over existing valves.

To illustrate the possible uses of ferrite components a microwave communication link is shown in Fig. 6.

The link includes a.g.c. using a ferrite attenuator and isolators in the aerial feeders to prevent intermodulation.

Acknowledgements

The author wishes to thank the Directors of Mullard Limited and Mr. P. E. Trier, Director of the Research Laboratories, for permission to publish this article, and Messrs. P. E. V. Allin, R. M. Godfrey, F. W. Smith and other colleagues for the benefit of numerous discussions and suggestions.

REFERENCES

1. Fox, A. G., Miller, S. E., Weiss, M. T. The Behaviour and Applications of Ferrites in the Microwave Region. *Bell Syst. Tech. J. 34*, 1 (1955).
2. Tellegen, B. D. H. The Gyrator, A New Electric Network Element. *Philips Res. Rep. 3*, 81 (1948).
3. Beljers, H. G. Non-Reciprocal Waveguide Elements using Ferrites. Nat. Lab. Report No. 3026 (1953).
4. Weisbaum, S., Seidel, H. The Field Displacement Isolator. *Bell Syst. Tech. J. 35*, 877 (1956).
5. Hogan, C. L. The Microwave Gyrator. *Bell Syst. Tech. J. 31*, 1 (1952).
6. Vartanian, P. H., Melchor, J. L., Ayres, W. P. Broadband Ferrite Microwave Isolator. *I.R.E. Convention Record 4*. Pt. 5. 79 (1956).
7. Ohm, E. A Broadband Microwave Circulator. *Trans. Inst. Radio Engrs.* MTT-4, 210 (1956).
8. Damon, R. W. Maser Shows Promise, Some Drawbacks. *Aviation Week.* P. 76 (Aug. 1957).
9. Godfrey, R. M., Humphreys, B. L., Allin, P. E. V., Mott, G. Applications of Ferrites at 3cm. Wavelength. *Proc. Instn. Elect. Engrs. 104B.* Supplement No. 6 (1957).
10. Sullivan, R. E., LeCraw, R. C. A New Type of Ferrite Microwave Switch. *J. Appl. Phys. 26*, 1282 (1955).
11. Barry, J., Clarke, W. Microwave Modulator uses Ferrite Gyrator. *Electronics 28*, 139 (May 1955).
12. Cacheris, J. Microwave Single Sideband Modulator using Ferrites. *Proc. Inst. Radio Engrs. 42*, 1242 (1954).
13. Chait, H. N. Non-reciprocal Microwave Components. *I.R.E. Convention Record*, Pt. 8. 82 (1954).
14. Jones, G. R., Cacheris, J. C., Morrison, C. A. Magnetic Tuning of Resonant Cavities and Wideband Frequency Modulation of Klystrons. *Proc. Inst. Radio Engrs. 44*, 1431 (1956).
15. Godfrey, R. M., Humphreys, B. L., Lusted, A. R. A Ferrite Duplexer for 3cm. Wavelength. Mullard Research Lab. Report No. 256.
16. Scharfman, H. Three New Ferrite Phase Shifters. *Proc. Inst. Radio Engrs. 44*, 1456 (1956).
17. Nelson, C. E. Ferrite Tunable Microwave Cavities. *Proc. Inst. Radio Engrs. 44*, 1449 (1956).
18. Reggia, F., Spencer, E. G., Hatcher, R. D., Tomkins, J. E. Ferrod Radiator System. *I.R.E. Convention Record. 4*, Pt. 1. 213 (1956).
19. Jaffe, D., Cacheris, J. C., Karayianis, N. Ferrite Microwave Detector. *Proc. Inst. Radio Engrs. 46* 594 (1958).

U.H.F. Power Meter for Operation in the 2000Mc/s Communication Band

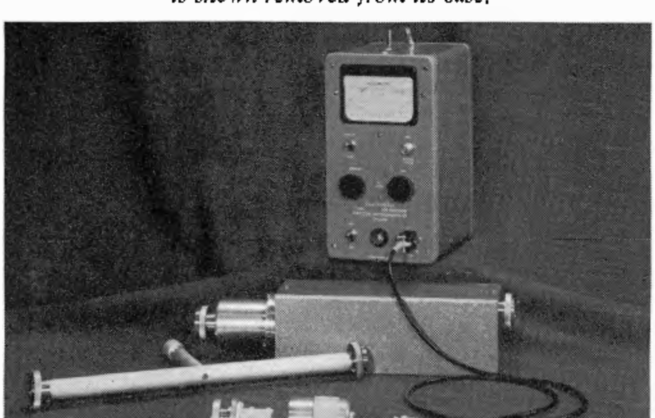
By J. K. Murray*

In order to assess the performance of circuits operating at frequencies above, say 1 000Mc/s it is often more convenient to measure r.f. power than to attempt to measure voltage or current in a direct manner. Such circuits, of necessity, involve the use of mechanical assemblies within which the voltage and current distribution cannot readily be ascertained. This article describes some of the problems encountered in the design of a direct-reading r.f. power meter for use in the frequency range 1 700 to 2 300Mc/s.

THE design project called for a direct-reading r.f. power meter having three sensitivity ranges, 0 to 1mW, 0 to 1W, and 0 to 20W, for use in the frequency band 1 700Mc/s to 2 300Mc/s. After due consideration it was decided to employ a semi-absolute method of power measurement in which the r.f. power would be dissipated directly in a temperature-sensitive element forming one arm of a bridge. This method, utilizing the bolometer principle, offered the choice of three possible alternatives—a lamp, a barretter, or a thermistor.

The use of a lamp did not receive more than fleeting consideration, because such a device only operates satisfactorily from about 1W upwards. This led to more serious consideration of barretters and thermistors.

The Marconi u.h.f. power meter type TF 1202 together with the 1W and 20W range extension units. The thermistor head is shown removed from its case.



Barretters have a positive temperature coefficient of resistance and, in the interests of quick response and sensitivity, employ very fine wire in their construction. These two factors tend to render the barretter very susceptible to overload burn out. In addition the barretters under consideration had to be biased to offer a resistance of not less than 200Ω; this presented a problem of impedance matching in order to terminate a 50Ω line correctly. Four such barretters could have been used for this purpose, but this introduced further problems in the form of greatly increased shunt

* *Marconi Instruments Ltd.*

capacitance and the possible necessity for tuning controls.

Attention was eventually turned to the possibility of using thermistors. From the outset the prospects looked favourable. The likelihood of overload burn out is greatly reduced due to such devices possessing a negative temperature coefficient of resistance. Under overload conditions the resistance of the thermistor drops to such a low value that, provided the indicating meter is suitably protected,

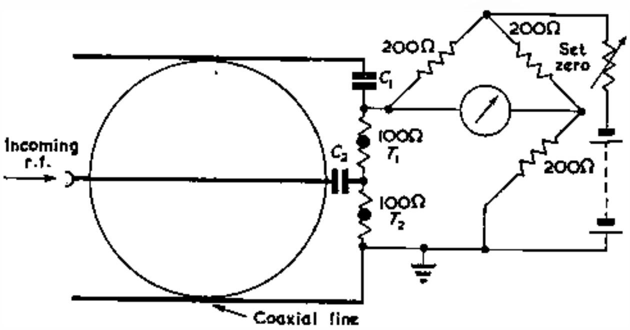


Fig. 1. The basic thermistor head, in which two thermistors are connected in series so far as the bridge is concerned, but effectively in parallel to r.f.

the resulting mis-match tends to restrict the thermistor overload to a relatively safe value. In the interests of quick response it was essential to choose a type of thermistor having very small mass. The type chosen had a 'bead' diameter of approximately 0·01in. Further, these thermistors could be biased to offer a resistance of 100Ω, so that two of them in parallel would form a satisfactory impedance match to a 50Ω line.

Fig. 1 shows the basic idea of a 50Ω coaxial thermistor head in which two thermistors are connected in series, so far as the d.c. bridge is concerned, but effectively in parallel to r.f. When using this system, d.c. bias is applied to the series-connected thermistors and adjusted, by means of a 'set-zero' control, until the bridge is balanced. When this state is achieved the two thermistors have a total series resistance of 200Ω and, from an r.f. point of view, form a terminating load of 50Ω.

Ideally, a balanced bridge would enable the thermistors to offer a constant 50Ω terminating load regardless of the r.f. power being measured. A 'slide-back' system using a calibrated meter-zero control would form a satisfactory solution. Use of this system was precluded, however, by the requirements for a direct-reading power meter.

Experiments with unbalanced bridges indicated that a resistance change from 100Ω to about 88Ω could be expected with the average thermistor for a power increase of 1mW. In the envisaged circuit, this would entail the 50Ω line being correctly terminated, when no r.f. power was being measured, and mis-matched, due to the load resistance falling to 44Ω, when maximum r.f. power was applied. This degree of mismatch was regarded as tolerable but led to further thoughts of a tapered coaxial section forming a correct match to a 47Ω load; in this case a near-perfect termination would be achieved when an r.f. power of 0·5mW was applied, and the mismatch at an r.f. power level of 1mW would be halved.

Practical considerations of mechanical layout led to the conclusion that, although the thermistors were satisfactorily small, there was going to be considerable difficulty in connecting the hair-like lead-out wires without introducing a large amount of unwanted inductance. This, together with a similar problem concerning the inductance associated with capacitor C_1, was overcome by dividing the coaxial outer longitudinally and, from a d.c. point of view, isolating the two halves from each other. These two halves, together with the coaxial inner conductor, taper to very small physical dimensions while still retaining a constant characteristic impedance—this taper may also be modified to offer a characteristic impedance of 47Ω at the load-end of the taper, as already discussed.

Thermistor Head

Fig. 2 shows the theoretical circuit of the final thermistor head, while Fig. 3 shows a cutaway view of its practical construction.

As illustrated in Fig. 3, the tapered outer part of the coaxial line is attached to the outer part of the input socket by means of screws passing through an aluminium washer. This washer is anodized on one surface only. The anodized surface faces the tapered section so that it forms an annular capacitor between the tapered section and the input socket

Fig. 2. Theoretical circuit of the final thermistor head

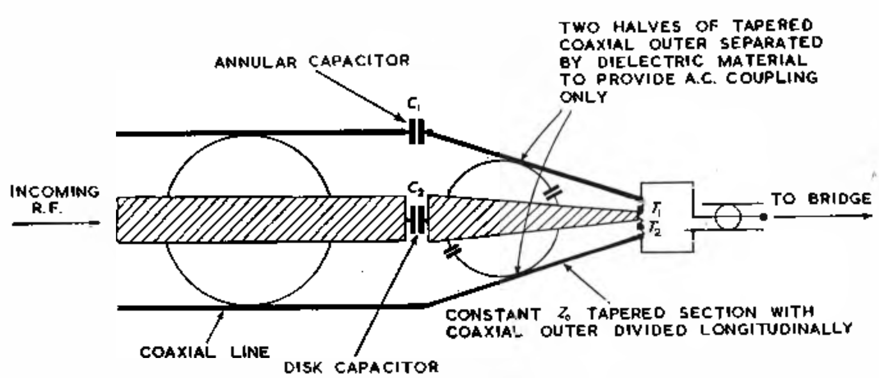
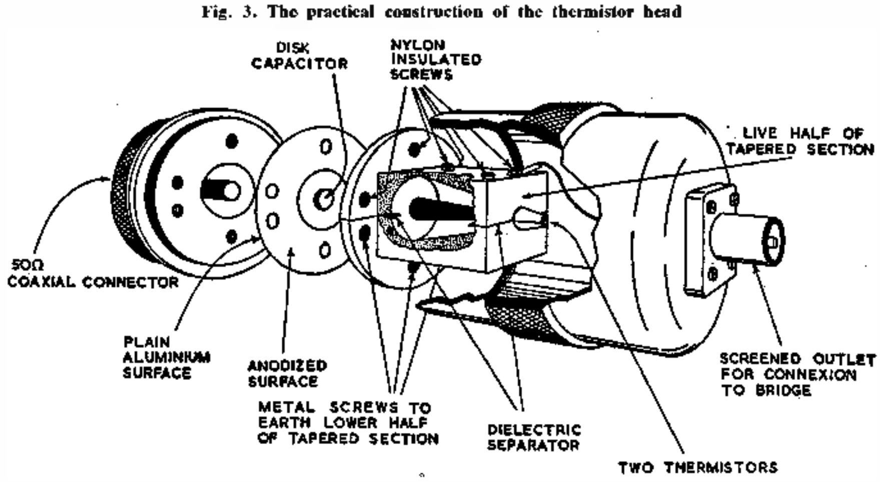


Fig. 3. The practical construction of the thermistor head



—the aluminium oxide forming the dielectric. Nylon screws are used to secure the upper half of the tapered section to the input socket while the lower half is attached with ordinary metallic screws. In this way, a capacitor, having very low residual inductance, is created between the upper half of the tapered section and the input socket, while the lower half is directly connected and therefore at earth potential.

The space between the inner and outer sections of the coaxial assembly is occupied by a solid dielectric of p.t.f.e. which, quite apart from its electrical properties, ensures that concentricity is maintained during mechanical assembly. The thermistor mount is contained within a machined metal cover so that the whole head assembly is effectively screened both electrically, and against sudden changes in ambient temperature.

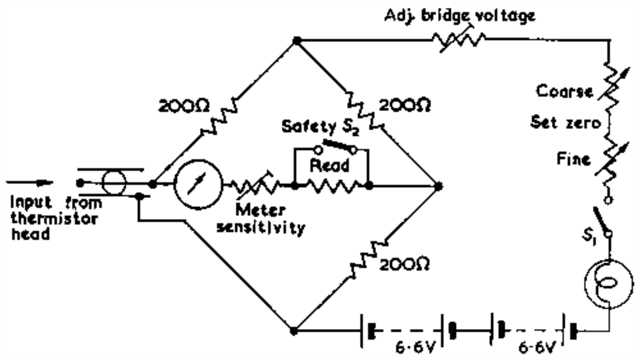


Fig. 4. Theoretical circuit of the bridge unit

Connexion from the bridge to the thermistor head is made via a conventional screened socket, the live part of this socket being internally connected to the upper part of the coaxial tapered section.

Bridge Unit

Fig. 4 shows the theoretical circuit of the complete bridge unit. This part of the instrument follows conventional lines and needs very little comment.

The two dry batteries supplying the bridge are housed within the meter case. In the interests of reliability these batteries are of the mercury variety, each delivering approximately 6·6V and possessing a total capacity of 3·4Ah.

A low-voltage bulb, in series with the battery supply, serves the dual purpose of acting as a pilot light and of providing a measure of current stabilization.

'Coarse' and 'fine' set-zero controls are provided for balancing the bridge while an additional preset control, 'adj. bridge volts', enables the instrument to be adjusted so that balance is achieved with the 'coarse' control at approximately mid-position throughout the life of the batteries.

During initial calibration the sensitivity of the meter is adjusted, by means of an internally-situated preset control, to give full-scale deflexion for an r.f. input of 1mW. When setting up this control, and for all subsequent r.f. measurements, switch S_2 is closed to the 'read' position. This switch is a biased toggle switch, and is normally in the open position so that a series resistor lowers the meter sensitivity and provides protection against accidental overload; this feature is particularly useful when the bridge is first being brought to a state of balance.

1-Watt Range Extension Unit

An add-on coaxial attenuator had to be designed for the measurement of r.f. power in the range 1mW to 1W. This presented certain problems before a really satisfactory unit for insertion between the source of r.f. power and the thermistor head was produced (Fig. 5).

The unit consists of a rigid coaxial line, some 15in in length, at the mid-point of which a coaxial stub-line is situated. The free end of the stub-line is short-circuited by means of a piston; the position of this piston is adjustable by means of a micrometer screw, thus providing a means of altering the effective length of the stub-line. The centre conductor of the main coaxial section is loaded with dust-iron material to introduce the required attenuation.

To provide reflection-free absorption of the r.f. energy, special precautions are taken over the construction of the centre conductor. The carefully-graded dust-iron material is arranged to offer the maximum absorption in the proximity of the stub-line, while the outer ends are gradually tapered to provide a smooth transition between the loaded and unloaded parts of the coaxial assembly.

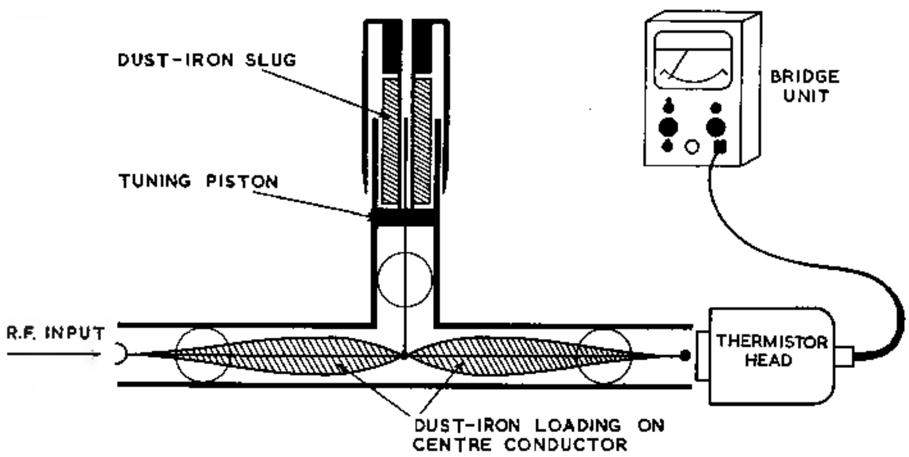


Fig. 5. Theoretical representation of the 1W range extension unit

The attenuation resulting from a loaded centre conductor of this type, varies with frequency. This is overcome by varying the position of the piston, and thus the admittance at the centre of the attenuating section. In this manner exact adjustment of the attenuation is achieved at any frequency within the band 1 700 to 2 300Mc/s. With each instrument the micrometer screw, controlling the position of the piston, is individually calibrated in frequency. The possibility of calibration inaccuracy, due to unintentional r.f. leakage past the piston, is minimized by a large dust-iron slug situated between the piston and the calibrated adjusting spindle.

20-Watt Range Extension Unit

The principle utilized in the design of the 1W range extension unit could not be satisfactorily applied in the case of the 20W unit—mainly due to considerations of physical size. The problem was therefore tackled in a different manner.

A unit was designed within which a high-dissipation resistor provides a 50Ω coaxial termination. A minute fraction (1/20 000) of the total power dissipated in this load is used to energize the thermistor head.

When designing the terminating load, use was made of the fact that, in a coaxial transmission line having a linearly resistive centre conductor with a total resistance equal to Z_0, a near-perfect termination may be obtained by tapering

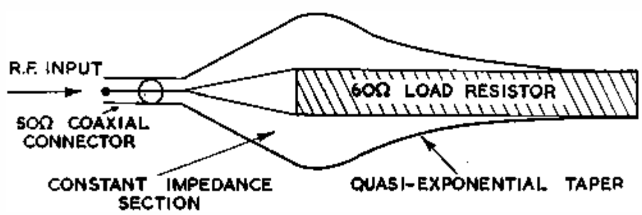


Fig. 6. Illustrating the principle of a 50Ω coaxial termination

the outer conductor in quasi-exponential fashion so as to join the inner conductor at the end remote from the source. When using this system, however, it is important that the dimensions of the coaxial line at the input end of the load resistor should be equivalent to Z_0. This requirement tended to conflict with the idea of using a large diameter resistor in order to dissipate the required 20W. The solution to this problem takes the form of a constant-impedance tapered section to increase the physical dimensions of the coaxial line to correspond with the dictates of resistor diameter.

Fig. 6 shows, in much simplified form, the construction of a terminating load using this principle. The heavy-duty, high stability, resistor employed has a tubular ceramic former with a conducting outer coating of cracked carbon. Provided the outer conductor is tapered in the correct manner, such a device presents a remarkably pure input impedance over a very wide frequency range –the necessity for tuning controls being completely eliminated.

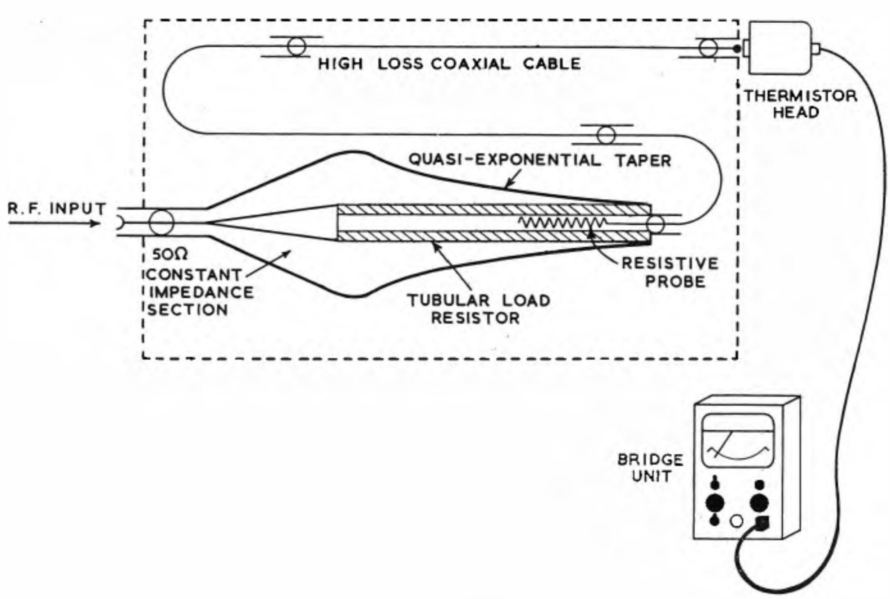
The basic construction of a terminating load having been settled, the next problem involved methods of 'tapping off' a small proportion of the energy in order to operate the power meter. It was also necessary, at this stage, to consider ways and means of adjusting the output level, so that an exact 1mW would be available for the thermistor head when 20W was being dissipated in the main load.

A 'tapping clip' on the load resistor, quite apart from its inaccessibility for adjustment purposes, was out of the question on the grounds of having to maintain a good v.s.w.r. The method finally adopted was simple in the extreme; it consisted of a probe inserted, to a greater or lesser degree, up the hollow centre of the 20W load.

Although this method of coupling seemed highly desirable, it failed to meet one requirement—that of matching the 50Ω thermistor head. In the final design, therefore, the probe itself is made of resistive material, and a three-foot length of lossy coaxial cable is interposed between the probe and thermistor head. This, of course, necessitates slightly increased coupling between the probe and the r.f. load resistor, in order to offset the attenuation due to the cable. However, the vast improvement in v.s.w.r., attendant upon using the lossy cable, completely outweighs the very slight disadvantages of using increased coupling.

As a matter of experience, rather than design, the lossy cable employed in the instrument has a characteristic impedance of 75Ω. To match this cable to the 50Ω thermistor head, a quarter-wave, 75/50Ω, coaxial transformer forms part of the outlet socket. The non-availability of suitable 50Ω lossy cable was entirely responsible for this arrangement and, because of this, these particulars are not included in the final diagram (Fig. 7).

Fig. 7. Theoretical representation of the 20-Watt range extension unit



Performance

The complete equipment meets the original specification requirements for an instrument to indicate mean r.f. power in the frequency band 1 700 to 2 300Mc/s with an accuracy of ±5 per cent of full scale deflexion on a direct-reading meter. The instrument is completely independent of external power supplies—apart from the power being measured—and, using the accessories described, the basic range of 0 to 1mW can be extended to measure powers of 0 to 1W and 0 to 20W. Lastly, under all conditions of measurement, the input impedance is 500Ω with a v.s.w.r. of better than 1·2 : 1.

Acknowledgment

The author wishes to record his thanks to the Management of Marconi Instruments Ltd. for permission to publish the details of the instrument described.

The White-Noise Method of Measuring Crosstalk and Noise Interference in Multi-Channel Telephone Link Systems

By J. F. Golding *

This article deals with a method, originally developed by the Post Office[1], of obtaining a realistic measure of noise and intermodulation products in multi-channel telephone-link systems. The use of a noise signal permits the simulation of full-traffic load conditions for the test. The Marconi white noise test set, an example of a commercially available equipment for this type of measurement, is also described.

Most of the noise and interference produced in multi-channel systems using frequency-division multiplexing is caused in two ways. The first and obvious cause is the noise energy generated in components of the link system itself. The second is intermodulation between channels due to various types of distortion in the system.

A wideband telephone system normally carries between 60 and 600 channels, each channel being formed by a single sideband of a carrier spaced 4kc/s from that of the adjacent channel. The sideband occupies a bandwidth of 3kc/s with 1kc/s separation from the next channel. The channel is, however, normally regarded as occupying the full 4kc/s. Originally, this type of transmission was passed through land lines; and non-linear distortion was the predominant cause of intermodulation. The crosstalk so produced normally becomes audible to the subscriber as an interference resembling random noise.

Providing the intermodulation is produced almost entirely by non-linear distortion, a fairly satisfactory measure of the relative noise and intermodulation products can be obtained by:—

(1) Applying a single-frequency tone and measuring both the harmonic distortion and the signal-to-noise ratio in a 4kc/s bandwidth.

(2) Applying two tones simultaneously and measuring the cross modulation between them.

From the practice of using these two methods, a more or less standard way of evaluating this aspect of the performance of these transmission systems has arisen; viz., the ratio between unwanted noise and test-tone level in a 4kc/s channel.

These methods do not, however, provide a realistic simulation of conditions during full-traffic periods of the systems. They are, indeed, particularly unsatisfactory when applied to radio links, which employ frequency or phase modulation almost exclusively. With these types of transmission low-level crosstalk is produced, not only by non-linear distortion, but also by what is known as delay distortion. This is selective phase distortion, whereby some of the frequency components of the transmitted signal are delayed relative to others. Thus the multiple sidebands of the frequency- or phase-modulated signal reach the receiver in the wrong relative phase. Delay distortion can be caused by reflection in transmission lines and waveguides. The intermodulation so produced adds to that caused by harmonic distortion and to the noise signals generated in the system and thus raises the interference level in the final output signal.

The white noise method of measuring the interference suppression automatically takes all these factors into account and gives, as a measured result, a noise-power ratio figure which can either be used directly as a measure of the relative freedom from interference or converted to the more familiar ratio already mentioned by the application of a specified factor.

To carry out the test a band of random noise, derived from a special generator, is injected at the input of the transmission system in place of the multi-channel traffic signal. The bandwidth of this noise signal is equal to that of the base-band frequency range of the traffic signal—i.e., the total frequency range of the number of channels carried; and the mean level of noise energy is adjusted to be approximately equal to the signal level exceeded during one per cent of the busy hour.

The noise energy is passed through a number of band-stop filters which produce 'holes' or quiet channels in the output frequency characteristic of the noise generator.

The output from the receiving end of the link system is fed to a noise receiver, having a narrow acceptance band whose centre frequency can be adjusted to coincide with the centre frequency of any one of the band-stop filters in the noise generator. With the receiver set to the centre frequency of a particular filter, the output from the link system is measured with the filter in and out of circuit, and the ratio between the two output levels measured is expressed in decibels.

The figure obtained as a result of this test is known as the noise-power ratio of the system. To convert it to the ratio of test-tone level to unwanted noise in a 4kc/s channel, the factor obtained from the following expression should be applied:—

$$R = \left(P - 10 \log \frac{A}{4} \right) \text{dB}$$

Where R is the conversion factor

P is the multi-channel signal power exceeded for 1 per cent of the busiest hour, measured at a zero-level point in the system expressed in dBm

A is the bandwidth of the multi-channel signal in kc/s.

This factor is expressed in decibels and should, therefore, be added to the noise-power ratio obtained by the measurement.

The Test Set

The following description of the arrangement of the Marconi white noise test set type OA1249 will serve to illustrate in more detail a practical application of the method of test. The basic arrangement of the equipment is shown in Fig. 1 as a block schematic diagram. From this diagram it can be seen that the test set comprises two main units—the noise generator and the noise receiver.

The Noise Generator

The noise signal originates from a noise diode. This type of noise source produces a band of random noise considerably wider than the base-band frequency range of the types of link equipment for which the test set is intended. It operates in a conventional circuit, in which the anode is held at a positive

* *Marconi Instruments Ltd.*

potential with respect to the cathode in order to ensure that all the electrons leaving the cathode reach the anode. The level of noise voltage developed across the diode load is directly proportion al to the d.c. diode current, which is, in turn, proportional to cathode temperature. This is controlled by a variable resistor in series with the d.c. filament supply.

The noise signal from the diode is applied to the input of a wideband amplifier. This is the first of a pair of amplifiers and may be called the low-level amplifier for the sake of convenience. It comprises four *RC* coupled stages in cascade feeding a cathode-follower output valve. Correction of the response to the higher frequencies within the band is effected by the use of an *LC* circuit in series with the anode loads of the amplifier valves. The cathode-follower output valve delivers the noise signal from a low source-impedance to match the filter circuits that follow the amplifier.

The noise signal from the cathode-follower is fed, via a capacitive coupling, to a high-pass filter having a cut-off frequency of 60kc/s. It is then passed through a low-pass filter having a cut-off frequency of 2·54Mc/s. The combined effect of these two filters is the production of a noise bandwidth extending from 60kc/s to 2·54Mc/s.

By the use of plug and socket connexions, the output signal from the second of these filters can either be fed directly to the input of the second wideband amplifier or one of three additional low-pass filters can be introduced. The frequencies of these additional filters are (1) 1·052Mc/s giving a noise bandwidth corresponding to a 240 channel base-band, (2) 522kc/s, corresponding to a 120 channel base-band, and (3) 300kc/s, corresponding to a 60 channel base-band.

The second—high-level—amplifier comprises three *RC* coupled voltage amplifying stages feeding a transformer-coupled output stage. A single loop of negative feedback from the output stage to the second amplifying stage carries an *RC* network which provides part of the necessary frequency-response correction. There is also a preset variable inductor, in series with the anode load of the output valve, and this provides the additional correction required.

The output from the high-level amplifier is passed through a network of switches and band-stop filters, so arranged that the noise signal can be passed through one or more filter or fed directly to an output attenuator. There is a separate switch for each filter.

The centre frequencies of the stop-bands of these filters are 70kc/s, 290kc/s, 534kc/s, 1 002kc/s, and 2 438kc/s respectively. Each stop-band is 4kc/s wide and introduces a loss of 80dB over this range. The switches are of rather a specialized nature, with earthed contacts acting as screens. This refinement is necessary in order to avoid capacitive coupling across the switch to a degree which may seriously reduce the 80dB insertion loss introduced by the filter.

The noise signal from the filter network is fed to the coaxial outlet via a 75Ω attenuator whose insertion loss is variable in 0·5dB steps over the range 0 to 61·5dB. The input to this attenuator is monitored by a diode voltmeter—comprising a moving-coil meter and two germanium diodes—which is calibrated in arbitrary noise units.

When the test set is in use, the noise output is adjusted by means of the diode current control, to deflect the meter to a predetermined reference indication. With zero attenuation, the mean noise output is then 15dBm for the 600 channel bandwidth, 12dBm for the 240 channel bandwidth and 10·5dBm for the 120 channel bandwidth. The output attenuator may be adjusted to produce a final mean noise level which is equivalent to the input level at full-traffic load on the link system.

The general form of the noise output characteristic is shown in Fig. 2. With all band-stop filters switched out of circuit, the characteristics is as shown in Fig. 2(a). The introduction of the filters modifies the characteristic to that shown in Fig. 2(b), for one filter in circuit, or Fig. 2(c) for all the filters in circuit.

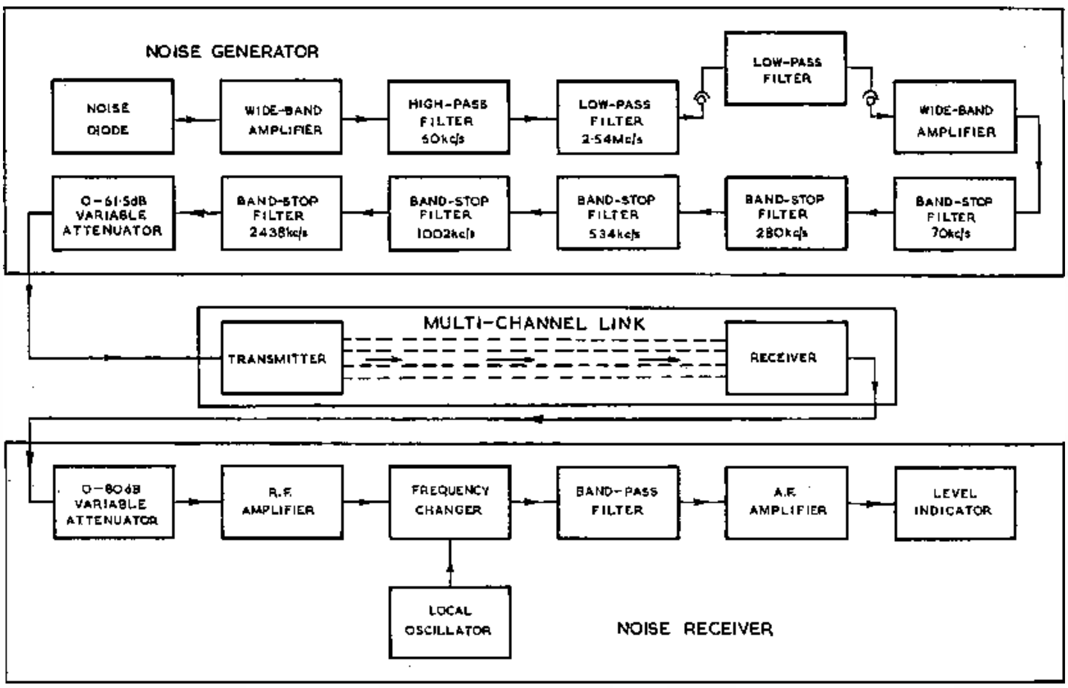


Fig. 1. Block schematic diagram of Marconi white-noise test set

The Noise Receiver

The narrow bandwidth and low internal noise, which are mandatory to the successful operation of the receiver, have necessitated the use of a somewhat unorthodox design.

The signal from the equipment under test enters the receiver via a 75Ω attenuator which is variable in steps of 0·5dB over a range 0 to 80dB. This attenuator plays an important part in the slideback method of measuring noise-power ratio.

From the attenuator, the noise signal goes to an over-coupled band-pass filter having a very steep-sided response with a certain amount of double-humping. This filter forms the grid circuit of the pentode input stage of the receiver. The tuned anode load of this valve has a very narrow bandwidth and, since it resonates at the centre frequency of the filter, the combined effect is one of producing an extremely well defined input-stage pass-band with an attenuation of off-tune signals exceeding 80dB. There are five input-filter/anode-load pairs selected by means of a frequency-selector switch, their centre frequencies corresponding to those of the five band-stop filters in the noise generator.

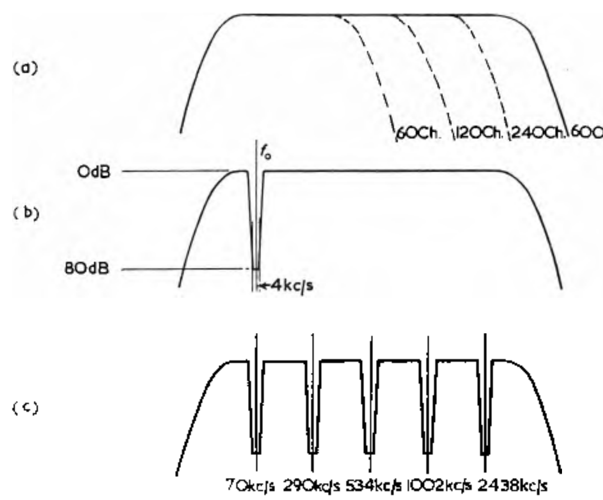


Fig. 2. The form of the noise-output characteristic of the noise generator
(a) With no bandstop filters in circuit
(b) With one bandstop filter in circuit
(c) With all the bandstop filters in circuit

The input stage is followed by a frequency changer comprising a hexode mixer and a separate oscillator. This oscillator's tuning inductor is selected by means of a wafer on the frequency-selected switch. The frequency of the oscillator coincides with the centre frequency of the input-stage pass-band; and the output of the mixer is therefore composed of both the upper and lower sidebands which add together over a frequency band covering from zero to about 1·2kc/s. However, the noise power from the mixer is equivalent to that contained in the total bandwidth of the input stage, which is, of course, twice that of the mixer output.

This output then passes through a low-frequency band-pass filter having a pass-band of 200 to 800c/s. It is this filter that finally determines the acceptance band of the receiver; and, because the signal passing through it is the sum of the upper and lower sidebands of the local-oscillator 'carrier', its 800c/s upper cut-off frequency produces an overall receiver bandwidth of 1·6kc/s. This is well within the '80dB' bandwidth of the bandstop filters in the noise generator, so that, for any given setting of the frequency-selector switch, the receiver's response can be regarded as 'fitting into' the stop-band of the appropriate filter in the noise generator. The lower cut-off frequency (200c/s) of the receiver's low-frequency filter produces a 400c/s gap in the middle of the overall response characteristic, so that the amount of noise energy actually reaching the l.f. stages which follow is equivalent to that contained in a noise bandwidth of 1·2kc/s.

This noise signal is amplified by means of a high-gain l.f. amplifier and applied to a diode voltmeter. The l.f. amplifier provides almost the whole of the receiver gain—some 80dB. The gain of the input stage is deliberately kept low because a high level of input to the mixer may produce spurious noise and cause errors in the final measurement.

Measurement of Noise-power Ratio

The method of using the white noise test set is simplified by the fixed alignment of the receiver frequency with the stop-band of each quiet-channel filter in the noise generator.

The output of the noise generator is applied to the input of the system under test, the noise generator's output band and level having been previously set up as described earlier in this article, and the noise output of the system is fed to the receiver.

The quiet-channel frequency at which the test is to be made is then decided, and the receiver's frequency-selector switch is set accordingly. The appropriate band-stop filter is switched into circuit at the noise generator and, with the receiver's input attenuator set to zero insertion loss, the receiver gain is adjusted for a convenient reading on the panel meter. The bandstop filter is then switched out of circuit, and the receiver's input attenuator is adjusted to return the meter pointer to the same reading.

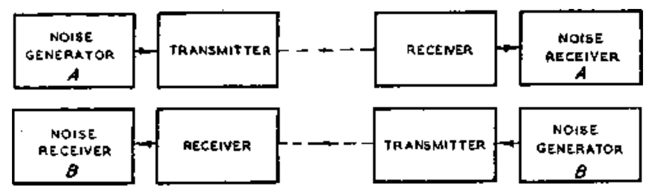


Fig. 3. Two white-noise test sets in use with two-way link system. Noise generator A is associated with noise receiver A, and noise generator B with noise receiver B

The noise-power ratio is then indicated directly in decibels by the setting of the input attenuator.

The Practical Installation

When this type of test set is used in connexion with a transmission system which is in operational use, the noise generator and its associated receiver are inevitably separated by some considerable distance. The test procedure, in these circumstances, must necessarily be carried out by two operators having some means of communication between them.

Normally, a telephone-link system is two way, and the arrangement shown in Fig. 3 may be used. This diagram is self explanatory regarding electrical arrangement. Noise generator *A* is associated with receiver *A*, and noise generator *B* with receiver *B*. Where the test equipment forms part of the permanent installation, generator *A* and receiver *B* would probably be mounted in a rack together at one terminal station, with generator *B* and receiver *A* assembled similarly at the other.

The test equipment is, however, also used on radio link equipment at factories and stores, with the receiving terminal separated from the transmitter by only a short physical distance. A complete white noise test set comprising the noise generator and its associated noise receiver may then form a single rack-mounted assembly.

Acknowledgment

The author wishes to thank the Management of Marconi Instruments Ltd for permission to publish details of Marconi equipment and for the use of photographic facilities in producing this article.

REFERENCE

1. Equipment for measurement of non-linear distortion and noise on 120, 240, or 600 channel systems. P.O. Engineering Department Radio Report No. 2339.

BOOK REVIEWS

Transistor Electronics

By David Dewitt and Arthur L. Rossoff. 381 pp. 80 figs. Demy 8vo. McGraw-Hill Publishing Co. Ltd. 1957. Price 60s.

TO cover the ground between the structure of the atom and the design of transistor amplifier and switching circuits in less than 400 clearly-printed pages requires a certain amount of condensation. It is no reflection on the authors that this is achieved by a lack of rigour and occasional superficiality. This book is intended as an introduction to transistors for the electronic engineer; for those who require a deeper treatment of particular points there is an adequate bibliography.

The main topics are: the Bohr Atom; quantum mechanics and the semiconductor; the pn junction; the transistor; the small-signal equivalent circuit; d.c. conditions and stabilization; the audio power amplifier; general amplifier circuits, feedback (but with little discussion of feedback problems peculiar to transistor amplifiers), and the use of matrices; high frequency amplification; large-signal transient response and switching; mixers and demodulators; radio receivers; other transistor types, including an analysis of the drift transistor; and transistor noise. A feature of the book is that the mathematics, of which there is necessarily a large amount, is from first to last illuminated by constant reference to practice and to fundamental physical processes. As a further aid to comprehension some problems on the content of each chapter are given at the end of the book. The appendices give the physical properties of germanium and silicon, the values of some physical constants, and matrix and determinant interrelations. A list of symbols would have been useful, and might have helped to avoid some minor lapses.

The diagrams are sometimes rather small but always clear, and are placed appropriately in the text. Printing errors are few. The book is American, but peculiarly so only in the circuit symbols used and occasionally the grammar and terminology (e.g. one-shot multivibrator); there is no loss of clarity.

The number of books on transistors is growing rapidly; it seems unfortunate, but is perhaps symptomatic, that the authors chose a title which has been used before. However, this book is undoubtedly among the best of its kind; by its nature it will not date rapidly; and both the student and the practicing engineer who wishes to refresh his memory or seeks an introduction to some unfamiliar aspect of transistors will find it of great value.

F. HIBBERD.

L'Automatique des Informations

By F. H. Raymond. 188 pp. 51 figs. Demy 8vo. Masson & Cie, Paris. 1957. Price Fr.1 600

"L'Automatique des Informations" deals chiefly with the digital computer. It is, however, not with electrical and mechanical details, nor with mathematics, that the author concerns himself. His emphasis is on fundamental notions—the representation of numerals in various coded forms each having their different merits, logical combination theory, the basic capabilities that a computer must have, and the manner in which it can be programmed.

These topics are minutely explored, classified, illuminated. At times the approach strikes one as tedious, but in general one is grateful to the author for having called a halt, as it were, in a rapidly-growing field and devoted so much patience to a loving exposition of its foundations. An introductory tribute speaks of the pages as "dénudant le corps et l'âme d'une calculatrice." It is an apt comment and, for those who are employed in programming or in computer design, this sifting of the essentials may well throw a refreshing light on concepts grown over-familiar.

The book includes a chapter on analogue techniques. It also gives a thoughtful survey of the field of automata in general and of the history of their development - governors, servo mechanisms, and data-handling and computing devices.

T. L. CRAVEN.

Technology of Instrumentation

By E. B. Pearson. 202 pp. 95 figs. Demy 8vo. English Universities Press Ltd. 1957. Price 25s.

SUCH a wide subject as the title imputes cannot be covered in two hundred pages. In the General Editor's Foreword it is stated that the book sets out to answer the problem—how to design an instrument which will measure . . . ., but the author refers to automatic control, computing and dynamic measurements of time-variant quantities . . . ., a much more limited scope. The book really deals primarily with the theory of servo-mechanisms for position control, which it does very thoroughly: it may prove to be a reference work as well as a text-book. Theory is predominant, with little reference to instruments in general use; for example some of the principles of selsyns and magslips are given, but neither name is quoted, Readers looking for design information on electrical measuring instruments, for example, will find nothing on the electrical side, but an excellent theoretical treatment of torque, damping, natural frequency, etc. Feedback, both mechanical and electrical is very fully treated mathematically and there is an introduuction to ways by which some simple mathematical operations can be performed by inanimate devices, such as differentiation and integration, and including an outline of analogue computers. The terminology is mainly correct, but the author's physics training is obvious. In some instances it is difficult to decide if 'capacity' is meant for physical capacity or is a mistake for capacitance. Within its actual scope the book is very good.

E. H. W. BANNER

Mechanical Resolution of Linguistic Problems

By A. D. Booth, L. Brandwood and J. P. Cleave. 306 pp. 19 figs. Demy 8vo. Academic Press, Inc., New York. Butterworths Scientific Publications, London. 1958. Price 50s.

IT is only twelve years since the use of a machine to translate one language to another was first suggested. Today one has the impression that ambiguity of word-meaning is the only problem that still presents serious difficulties. It seems fairly certain that all problems of syntax can be made amenable to handling by computer and that, at least if the subject matter be restricted to a given field, the finished translation can be of such a standard as to require little or no editing after it emerges from the machine. The syntactical rules devised for the machine are often artificial—utility, we are told, is their only criterion. It must be remembered, though, that even in the spontaneous development of a language it is utility that is the unconscious guide.

Mechanical Resolution of Linguistic Problems is an account of some of the results which have been obtained at Birkbeck College Computational Laboratory on the application of digital computers to translation and some other linguistic problems.

The early pages give a historical review, and then proceed to outline the general principles of data-processing machinery, including the APEXC computer used at the college. One feels that the operation of this computer should have been explained more clearly in view of the extent to which an ability to understand its programmes is necessary in reading some later parts of the book. In particular, the table of computer orders seems as if taken in its entirety from some other source, and could with advantage have been made more explicit. The reader must deduce for himself that the third column, mysteriously headed "C_s", refers (or so your reviewer believes) to the setting of the counter. For that matter, it is only the first and the last of the five columns that appear to be necessary to the reader.

Two interesting chapters deal with the use of computers in literary research. To take an example, it appears that an author's works, of dates unknown, can sometimes be tentatively arranged in chronological order by noting a changing trend in the use of certain words or certain sentence structures. The discovery of such trends demands an enormous amount of counting and sorting, and for this a computer is invaluable.

There follows a clear exposition of the approach to the basic problems of machine translation—the arrangement of the dictionary and the dictionary-search procedure, recognition of stems and endings and the correct interpretation of the latter, word-order re-arrangement, idioms, and the resolution of ambiguities. Then come some accounts of investigations into machine translations of specific languages.

These accounts, which form the greater part of the book, exhibit an unfortunate variation of approach and size of the chapters. The different approaches are understandable—one reason is that the scope of the investigations has varied—but, as to size, the reader must be warned that no less than 162 of the 306 pages in this book are given to one sole chapter which presents some of the main problems to be faced in the machine translation of German. There is, in consequence, a lack of balance. One feels that one has been reading the laboratory reports available to date. And yet that, perhaps, is a very good thing—there has been so little published on this subject, and things are moving fast.

T. L. CRAVEN.

Elements of Tape Recorder Circuits

By H. Burstein and H. C. Pollak. 223 pp. 60 figs. Demy 8vo. Gernsback Library Inc. 1957. Price $2·90 paper cover, $5 hard cover.

THIS book is written for both the audiophile and technician. Basic design principles, with a minimum of arithmetic, are included so that the reader may modify his own equipment to suit his particular needs.

The various chapters are concerned with the electronic functions of the tape recorder, why these functions are called for and the means of performing them. The fundamental problems of frequency response, distortion and noise are dealt with in terms of design and of extracting optimum performance from a given recorder. Requirements for high quality performance are given and basic circuitry is described and evaluated.

Electronic Engineer's Reference Book

Edited by L. E. C. Hughes. 1 303 pp. 800 figs. Demy 8vo. Heywood & Co. Ltd. 1958. Price 84s.

AFTER the preliminary pages which include two articles on the history of electronics, the body of the text has been divided into nine main parts: Fundamentals, Radiations, Electrics, Valves, Materials, Vibrations, Computers, Automatics, and Miscellaneous.

This reference book endeavours to put before engineers in industry and in development laboratories some of the latest knowledge and techniques.

Nuclear Energy in Industry

By J. G. Crowther. 168 pp. 98 figs. Demy 8vo. George Newnes Ltd. 1957. Price 17s. 6d.

THIS book is the outcome of the author's attendance at the International Conference on the Peaceful Uses of Atomic Energy held in Geneva in 1956. It tells the story of development culminating in the building of nuclear power stations in this country and overseas, and describes the role of the radioisotope in industry and its effect on techniques and processes. The book makes interesting reading for the layman and is recommended for those interested in the future of industry.

Control Engineering Manual

Edited by B. K. Ledgerwood. 188 pp. 80 figs. Royal 4to. McGraw Hill Publishing Co. Ltd., New York and London. 1957. Price 56s. 6d.

THIS manual consists of articles that have appeared in Control Engineering. It deals with specification of control systems, synthesis of control systems, evaluating control systems, generating control functions, design techniques, analogue computers, non-linearity in control systems, etc.

Electronic Designer's Handbook

By R. W. Landee, D. C. Davis and A. P. Albrecht. 800 pp. 80 figs. Royal 8vo. McGraw-Hill Publishing Co. Ltd, New York and London. 1957. Price £6 8s.

THIS handbook gives fundamentals and data to aid in the design of various types of electronic equipment. A large number of the circuits used by engineers is covered, with theoretical and technical discussions and explanations, design examples to show application of theory, and graphical and tabular data needed in day-to-day design work.

There are twenty-three sections ranging from vacuum tube fundamentals and voltage and power amplifiers to such topics as computer and servomechanism techniques and waveform and network analysis.

Installing Electronic Data Processing Systems

By R. G. Canning. 190 pp. 40 figs. John Wiley & Sons, Inc., New York. Chapman & Hall Ltd, London. 1957. Price 48s.

THIS book deals with the high-cost aspects of installing electronic data processing systems, training of personnel, programming, physical installation of the system, and the early phases of operation. It is written in non-technical language and assumes that the reader is relatively unacquainted with electronic computers.

Frequency-Modulated Radio

By K. R. Sturley. 120 pp. 40 figs. Demy 8vo. 2nd edition. George Newnes Ltd. 1958. Price 15s.

THE purpose of this book is to explain the general principles, theory, design, construction and servicing of v.h.f./f.m. receivers. Each stage of a v.h.f./f.m. receiver is analysed in detail, and a chapter is also devoted to the special features of combined a.m./f.m. receivers.

In this new edition, information has been included on the counter-type detector and other revisions have been made.

ELECTRONIC EQUIPMENT

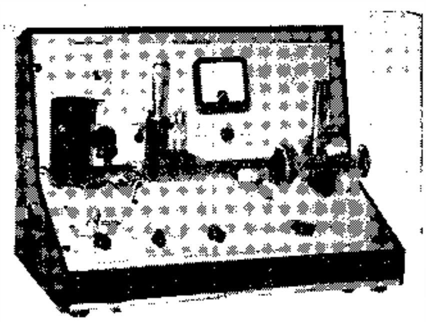
A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

MICROWAVE TEST BENCH

(Illustrated below)

Microwave Instruments Ltd, West Chirton Industrial Estate, North Shields, Northumberland

THE Micro bench type 3930 is essentially a complete X band test bench for the teaching and demonstration of microwave technique, with complementary uses in industry, for microwave measurement and research. Basically it comprises a high quality stabilized power unit (type 3950), on to whose sloping front panel is mounted the waveguide assembly (type 3900). All those instruments necessary for microwave measurements are incorporated into the short No. 16 waveguide run and these are:—klystron mount, setting attenuator, frequency meter, directional coupler, and sampling crystal, measurement attenuator and standing-wave meter, design frequency being 9 375Mc/s (3·2cm) and the range of the 723 A/B klystron 8 500 to 9 660Mc/s.



The standing wave meter which is the only detachable item on the waveguide is capable of measuring v.s.w.r. in the order of 0·99 (1·01) and is controlled by a knob on the front of the instrument. Variation of probe penetration to a maximum of 3mm is provided and the tuned line can be adjusted to 3·2cm at the centre of its travel. Each instrument is capable of accurate measurement and the supply power unit is specifically designed to suit the requirements of the klystron used.

The resonator and reflector voltages are shown on a 3½in meter which also serves to indicate the cathode current of the klystron and as a monitor for the u.h.f. generated in the waveguide. A separate meter is needed to measure the output from the standing wave indicator and for this purpose a matching unit is available, the transistor amplifier T.A.2.B. having a sensitivity of 2μV r.m.s. for full scale deflexion.

500W R.F. POWER METER

(Illustrated in next column)

Marconi Instruments Ltd, St. Albans, Hertfordshire

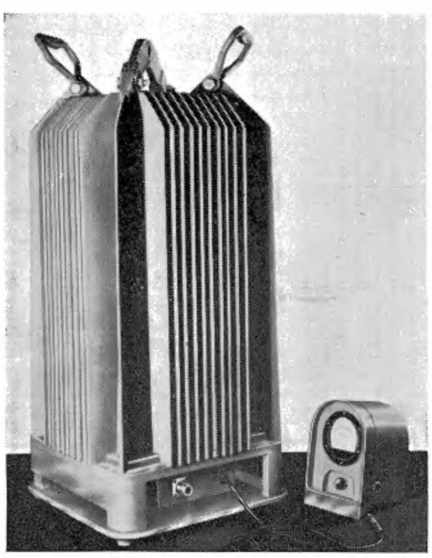
WITH this instrument, Type TF 1205, direct measurement of powers up to 500W may be made at any frequency from d.c. to 500Mc/s. The equipment has commercial as well as military applications, satisfying the requirements of the ever increasing number of transmitter operators who are working on higher powers and frequencies in the v.h.f. and u.h.f. bands.

An absorption-type instrument, the TF 1205 is ruggedized, completely self-contained and fully transportable. The dissipative element is oil immersed and cooling of the finned outer casing is by free air convection. The design obviates the need for water or forced-air cooling and yet retains the advantages of compactness and a sealed construction.

The directly calibrated meter has a substantially linear scale and indicates true mean power, irrespective of waveform. Intended for direct connexion to a coaxial cable, the power meter has an input impedance of 50Ω.

The dissipative element in the TF 1205 consists of a heavy-duty high-stability resistor of tubular form, so mounted that it forms the central conductor of a slab or parallel-plate line. The input power is fed to the 'live' end of the resistor by an outward-tapering section which preserves a constant impedance between the connector and the relatively large-dimensioned resistor. From 'live' to 'earthy' end of the resistor, the broad metal plates which form the outer conductor have an inward taper so that continuous matching is maintained along the whole length of the resistor. Efficient cooling is obtained by immersing the complete slab-line assembly in a tank of transformer oil. Cooling fins provide a total surface area of over twenty square feet. By this means the overall temperature rise of the oil is restricted to approximately 30°C.

The measuring circuit comprises a vacuum thermocouple, fed from a tap on the load resistor, and a moving-coil meter mounted in a separate indicator unit. The latter can be attached to the top of the main assembly or used remote from it up to the full six feet allowed by the connecting cable.

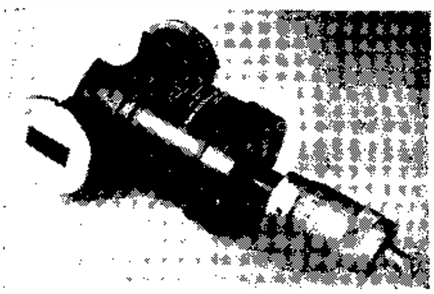


WAVEMETER

(Illustrated below)

Flann Microwave Instruments Ltd, 133 Munster Road, Teddington, Middlesex

THIS instrument (type 16/7) consists of a length of waveguide that supports, on one narrow wall, a high Q resonant cavity operating in the TE_{01} mode. Coupling between the waveguide and the cavity is through two symmetrical holes at the end of the cavity, in the common waveguide and cavity wall. Spurious mode resonances are inhibited by interruption of current paths in the appropriate points.



Tuning of the cavity is effected by means of a micrometer head that is coupled to a spherical plunger. A disk of lossy material behind the plunger is fitted to damp out resonances that may otherwise occur. The cavity is hermetically sealed to guard against changes of humidity.

Although the wavemeter can only be used from 8·4kMc/s to 10kMc/s without ambiguous readings, it is possible to use the instrument up to 12·4kMc/s if the frequency is roughly known within 500Mc/s.

The instrument is finished in silver and rhodium plate with external surfaces in dark green.

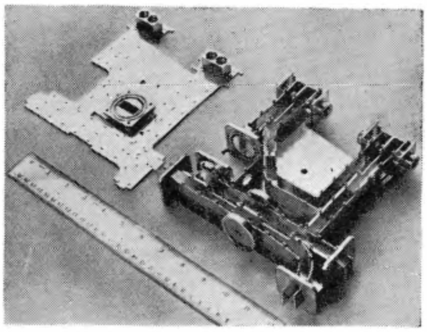
The waveguide is 0·9in × 0·4in, the unloaded Q 20 000 and the loaded Q 10 000. The calibration accuracy 1Mc/s and the descrimination 200kc/s.

DIP-BRAZED ALUMINIUM WAVEGUIDE

(Illustrated above right)

General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

A TECHNIQUE for the ready manufacture of almost any shape of waveguide components, providing much lighter and more flexible designs, has been developed by the G.E.C. The process involves the use of specially coated aluminium sheet which is punched and assembled by an interlocking tag and



slot construction, thus dispensing with the necessity of assembly and brazing jigs. This feature enables the whole unit to be brought up to temperature quickly and uniformly by means of salt-bath brazing, eliminating the distortion experienced with torch brazing methods. In addition, the salt dip-braze produces cleaner and more uniform joints, and on X-band waveguide, tolerances can be maintained to within ·002in.

Using this technique, it has been possible to replace a complex system of aluminium die-cast waveguide components for an airborne radar transmitter-receiver. The net reduction in weight was from 8lb to less than 1lb.

A.M.-F.M. MODULATION METER

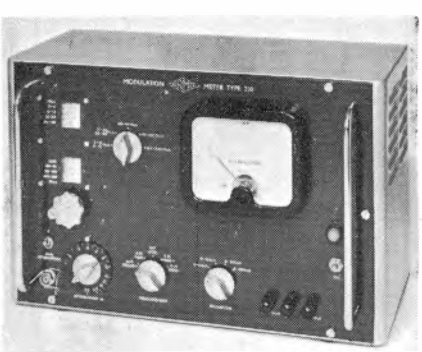
(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire

THE modulation meter type 210 may be used to measure the percentage modulation of amplitude modulated signals and the peak deviation of frequency modulated signals in the carrier frequency range 2·25Mc/s to 300Mc/s. Operation with reduced sensitivity is possible up to a frequency of 600Mc/s. Modulation depths up to 100 per cent and frequency deviations up to ±100kc/s. may be measured in the modulation frequency range 30c/s to 15kc/s.

Outputs at the intermediate frequency of 750kc/s and at low frequency are available from terminals on the front panel. These outputs enable the modulated envelope of the input signal and the demodulated signals to be observed on an oscilloscope.

The limiting action of the instrument is so effective that it can be used to measure spurious frequency modulation occurring on amplitude modulated signals. Furthermore, changes of mean carrier level when amplitude modulation is applied can be measured to an accuracy of better than ±1 per cent.



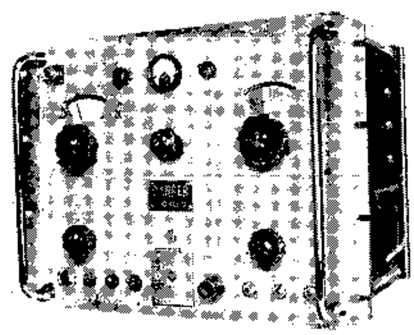
One of the most attractive features of the instrument is its simplicity of operation. The tuning control is adjusted until a meter reading is obtained, and the input attenuator adjusted for full scale deflexion. It is then only necessary to switch to the a.m. or one of the f.m. positions to obtain a direct reading of modulation.

MICROWAVE SIGNAL GENERATORS

(Illustrated below)

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Hertfordshire

A RANGE of signal generators are available to provide oscillation over the band 2kMc/s to 12kMc/s. A typical one (illustrated) is the XT.314 used to measure the performance of receivers and for general laboratory measurements in the frequency range 7 to 12kMc/s. A piston attenuator provides a calibrated output adjustable between +10dBm and −100dBm at a 50Ω type N socket on the front panel. An internal power set control



is provided to establish a reference power level of 1mW. Facilities are provided for internal square and pulse modulation at a frequency of 3kc/s together with a synchronizing signal output. There is also provision for external pulse modulation from a source delivering a 5V positive pulse with a duration greater than 0·1µsec, and for frequency modulation.

NEW KLYSTRONS

English Electric Valve Co. Ltd, Chelmsford, Essex

THE K345 is a newly-developed E.E.V. medium power reflex klystron oscillator giving an output of 1·2W and is an equivalent of the American type VA220. The K345 has been designed particularly for microwave links. Good linearity of the frequency versus reflector voltage characteristic ensures low f.m. distortion. Frequency stability is improved by the addition of an external tuning cavity.

There are seven frequency variants which together cover the band from 5 925 to 8 025Mc/s.

The high power K339 for c.w. operation in the u.h.f. range 1 400 to 1 600Mc/s gives an output of 10kW at 30 per cent efficiency with a bandwidth of 5Mc/s. This valve is a four cavity design with water cooled collector and drift sections. Up to 55dB gain is obtainable.

The K350 is a two resonator klystron oscillator which has been specifically designed to have low noise modulation and good frequency stability making it especially suitable for the latest airborne f.m. radar applications. Output power is approximately 1W over the frequency band 8 500 to 10 000Mc/s. This valve operates at a beam voltage varying from 650 to 950V over the frequency band and has an electronic tuning range of about 12Mc/s between half power points.

The K343 is a low voltage reflex klystron giving 20mW minimum output for an input of 350V, 20mA. Mechanical tuning covers 12 000 to 14 500Mc/s. The output connexion is American type UG419/U into No. 18 waveguide. The moulded base and flying leads permit high altitude operation without pressurization. The K346 is similar except that its tuning range is 14 500 to 17 000Mc/s.

R.F. REFLECTOMETER

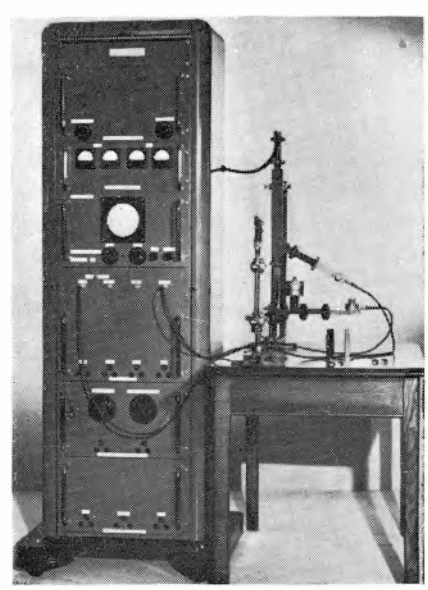
(Illustrated below)

General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

THE use of this instrument substantially reduces the time needed to test wide-band waveguide and coaxial line components. Hitherto a laborious process taking a matter of days, work of this nature can now be accomplished in a few hours. The reflectometer displays the magnitude of the reflection coefficient of a microwave device on a cathode-ray oscilloscope, and can also be used to show transmission coefficient.

The three essential components of the reflectometer are a microwave generator with power supplies, electronic frequency scanning supply and a regulating feed-back circuit; an r.f. circuit comprising level-setting attenuator, wavemeter, and two directional couplers each with a crystal detector and matched load on the coupled arm, and an oscilloscope, with power supplies, amplifiers and display circuits.

The complete equipment is at present available in waveguide 16 for X-band (7 500 to 11 000Mc/s), and in coaxial line for S-band (2 000 to 3 200, or 2 400 to 4 000Mc/s).



Short News Items

H.R.H. The Duke of Edinburgh has consented to be patron of the Electronic Computer Exhibition and Business Symposium to be held at Olympia, London, from 28 November to 4 December 1958.

'Hazel', a new zero-energy reactor, is now operating at the Atomic Energy Research Establishment at Harwell. 'Hazel' (homogeneous assembly-zero energy) uses enriched uranium fuel in the form of a salt of uranium curanyl fluoride which is dissolved in the heavy water used as a moderator. The core of the reactor is a stainless steel cylinder (7ft high by 2ft diameter) surrounded by a graphite reflector. The fuel solution is pumped into the cylinder from two nearby storage vessels. The system is controlled by adjusting the level of the fuel solution in the reactor vessel and by moving a vertical neutron-absorbing cadmium plate into the gap between the steel cylinder and the graphite reflector. Two cadmium plates similar to the control device are used as 'shut-off' rods: two additional 'shut-off' rods can be dropped vertically into the fuel solution in the reactor vessel. The reactor will be operated at a power of less than a watt and cooling is not necessary. 'Hazel' will be used to obtain basic nuclear information. An earlier system at Harwell, ZETR (zero energy thermal reactor) was used to study solutions of nuclear fuel in ordinary water.

Standard Telephones and Cables Ltd have received an order from the National Physical Research Laboratory of the Council for Scientific and Industrial Research of South Africa for an electronic digital computer. The machine is called 'Stantec Zebra' and will be used to carry out calculations for pure and applied research problems by the N.P.R.L.'s Applied Mathematics Section. Work in a variety of fields will be undertaken including nuclear physics, crystallography dynamics, telecommunication and statistical problems. It will also be made available to industrial organizations in the Union of South Africa. Similar machines have been sold to universities and research centres in England, Holland and Switzerland.

The BBC has announced that with the approval of the Postmaster General, low power television stations are to be installed at Swingate, near Dover, and in Folkestone. The difficulties that had arisen in finding wavelength channels on which these stations can operate without causing interference elsewhere have now been overcome, and both stations will be put into service as quickly as possible.

Mr. Graham Phillips has been seconded to the Kenya Government Broadcasting Service as Chief Broadcasting Engineer.

A radio telecommunications system, the largest of its type in the world, was inaugurated recently in Nigeria by the Nigerian Minister of Communications and Aviation. Marconi's Wireless Telegraph Co Ltd have been responsible for the construction of the whole system, which is capable of providing an ultimate capacity of over 50 000 channel miles. Marconi v.h.f. multi-channel equipment has been used at all the 14 terminal and 25 repeater stations. With the completion of this scheme internal communications in Nigeria have made a significant advance. It is now possible to telephone from Lagos to many provincial towns whose local services have not previously been connected into a national network. The important centres now linked by first class trunk services include Lagos, Ibadan, Benin, Onitsha, Enugu and Kaduna.

The Mechanical Engineering Research Laboratory, East Kilbride, Glasgow, is to hold Open Days on Wednesday, 4 June, and Thursday, 5 June. Representatives of any organization with engineering interests will be welcome. Applications for invitations, stating which day is preferred, should be sent to the Director. The whole of M.E.R.L. will be open for inspection, including two new laboratories for research on mechanisms and engineering metrology, and for research on heat transfer.

ARTICLES HELD-OVER

Due to the limitation of space, the following articles submitted for this issue have had to be omitted.

"Aerial Feeders" by L. Lewin and J. Paine (Standard Telephones and Cables, Ltd)

"New Equipment for Testing the Linearity of Modulators and Demodulators in Multi-Channel F.M. Transmitters and Receivers" by G. C. Davey (Marconi Instruments, Ltd)

"Two Automatic Impedance Plotters" by R. S. Cole and W. N. Honeyman (E.M.I. Electronics, Ltd)

"Use of Scale Model Techniques in the Design of V.H.F. and U.H.F. Aerials" by F. J. H. Charman, J. Thraves and E. F. Walker (E.M.I. Electronics, Ltd)

They will, however, be published in the near future.

"Letters to the Editor" have also had to be held-over until next month.

A two/three day Introductory Course on Radioisotopes is to be held at the North Herts Technical College from 14-16 May. The fee will be £3 3s., inclusive of luncheons and light refreshments. Further details may be obtained from the Principal of the College, Broadway, Letchworth.

A Summer School in Programme Design for Automatic Digital Computing Machines will be held in the University Mathematical Laboratory at Cambridge during the period 15-26 September. It will be along similar lines to those held in previous years, but the material has been thoroughly revised and students will have access to EDSAC 2, a new and fast machine now installed in the Laboratory. A detailed syllabus and form of application for admission may be obtained from Mr. G. F. Hickson, Secretary of the Board of Extra-Mural Studies, Stuart House, Cambridge, to whom the completed application form should be returned not later than 16 June.

The Royal Institution have chosen four films made jointly by Mullard Ltd and the Educational Foundation for Visual Aids for showing in the International Science Hall at the Brussels Exhibition. These are The Principles of the Transistor, The Linear Accelerator, Discharge Through Gases, and Mirror in the Sky. The latter film won a premier award at the recent International Industrial Film Festival at Harrogate.

The Electronic Machine Co Ltd have announced the merger with the Ellis Optical Co. The merger means that in addition to their having now entered the optical and scientific instruments field, the Electronic Machine Co Ltd will be able to increase the scope of their present business. Expansion has meant the creation of two new executive posts. Mr. W. A. Burnside has joined the company as General Sales Manager and Mr. J. Wardle has been appointed Contract Manager.

Elliott Brothers (London) Ltd have announced the conclusion of an agreement with the Holtzer-Cabot Motor Division of the National Pneumatic Co Inc of Boston, Massachusetts, U.S.A., for the manufacture of many of the Holtzer-Cabot instrument motors in the United Kingdom. The types of motors to be manufactured, to meet the special requirements of the European and British markets, include precision electric servomotors for strip recording instruments, timing devices and electronic control systems of all kinds, and also general-purpose ranges of industrial induction and synchronous fractional-horsepower motors for industrial purposes.