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THE MARCONI REVIEW

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WIDE BAND MICROWAVE RADIO LINKS

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

A broad survey of the techniques used in the construction of wideband microwave links is given, with particular emphasis on the applications of travelling wave tubes, in these links.

Some of the design requirements are examined in the light of the conclusions of the C.C.I.R. Study Group IX, meeting at Geneva, at the end of 1954.

Historical

EXPERIMENTS made before the war[1] have shown that the use of radio links at VHF and UHF was quite practicable for the carrying of telephone and telegraph traffic. The development of these links, after the war, was, in the main, concentrated in the VHF bands, and the quality and reliability of these links was found to be comparable with that of coaxial cables, while the initial cost was considerably lower. Furthermore they provided an almost ideal medium for the conveyance of traffic, in relatively undeveloped countries, where the establishment of wire circuits is both an expensive and hazardous operation.

However, with the very considerable expansion of telephone traffic and the added requirement for the transmission of television signals, it was quite clear that only the UHF and SHF ranges could provide sufficient spectrum space to accommodate the very wide bandwidths that were becoming necessary. The VHF range was becoming overcrowded to such an extent that the design of equipment capable of carrying sufficient traffic of adequate quality was becoming extremely difficult and onerous, mainly because of bandwidth limitations.

In the case of radio links carrying 600 telephone channels, the signal bandwidth is approximately 2·5 Mc/s, while for a 625 line monochrome television signal, the highest modulation frequency is 5 Mc/s. The need for a high order of linearity in multichannel telephony systems, at any rate, requires the use of frequency modulation, and, on the assumption of a modulation index of approximately unity, the necessary R.F. bandwidth is of the order of 15-20 Mc/s. Hence the shift to the microwave frequencies.

The great development in the technique of microwave circuitry, during the war years, was such as to make a preliminary system design quite feasible. The position as regards suitable valves was, however, by no means a favourable one. Few power klystrons were in existence and the life of the microwave triodes which were then available was not such as to allow the design of links having the necessary degree of reliability for continuous, trouble-free operation.

The Bell Telephone Laboratories developed a new range of triodes(2) for use at SHF, by pushing conventional techniques to very high orders of perfection in terms of mechanical precision and electrical performance. At the same time, the invention of a new broad band amplifier, the travelling wave tube, suitable for use at microwave frequencies, ushered in a new era in the development of reliable links capable of providing, economically, the very large bandwidths that are necessary. In fact, the bandwidth provided by these tubes is so large as to prove somewhat of an embarrassment in certain cases. The travelling wave tube also provided a means of eliminating, at least partially, the complicated IF circuits where simplicity has to be thrown overboard in order to attain the necessary degree of phase linearity and consistency of gain, over the required bandwidths which are essential in the case of systems of large capacity.

Choice of Frequencies

The frequency bands allocated for point to point links, by the International Convention at Atlantic City(3), are as follows:—

352-420 Mc/s.
1700-2300 Mc/s.
3600-4200 Mc/s.
5850-8500 Mc/s.

It is possible, by working at relatively high frequencies, to take advantage of the increased directivity and gain of aerials of a given size. In the last three ranges also, waveguide components become smaller and easier to handle as the frequency increases. On the other hand mechanical tolerances in almost every microwave component become more difficult to achieve and fading becomes more severe.

On the whole the choice between the last three ranges is one of convenience and it is, in the last resort, dictated by the availability of suitable valves. As the frequency increases, the power output of travelling wave tubes, or of other valves, tends to decrease and it is no more difficult, for example, to provide 20 watts at 2000 Mc/s than it is to provide 5 watts at 4000 Mc/s, so that the increased aerial gain at higher frequencies is roughly offset by the reduced power output of the transmitting valves.

In the same way the need for better mechanical tolerances at the higher frequencies is partly compensated by the reduced relative bandwidth over which certain microwave components have to operate. For example, at 4000 Mc/s a relative aerial bandwidth of about 15% is required to cover the whole of the available band, while double this figure is necessary at 2000 Mc/s.

The use of microwaves limits the transmission of information to line of sight paths. Adequate terrain clearance has to be provided in order to avoid excessive attenuation due to ground reflections and due to shadow effects.

Even with adequate terrain clearance however, reflection from the ground may be quite troublesome. In serious cases bad echoes and excessive crosstalk may be produced. Atmospheric absorption is reasonably low in the communication bands, but fading due to multipath propagation is likely to be more severe as the frequency increases[4].

The Microwave System

Of the known methods of modulation, i.e. amplitude, frequency, phase or pulse, frequency modulation is at present generally accepted, for reasons discussed in a subsequent paragraph as the most practicable system for use in conjuction with high capacity radio links intended for the conveyance of telephone traffic or television signals.

Because of the very high frequencies involved, wideband radio links have to be operated under line of sight conditions. In the case of a long link, therefore, it is necessary to use one or more repeaters between terminals, when the latter are not within optical distance of each other. Given normal terrain configurations the distance between any two repeaters is generally between 20 and 40 miles. A typical link arrangement is shown in Fig. 1.

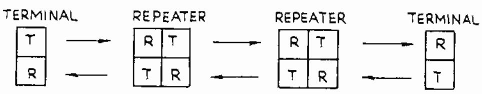

FIG. 1

Typical 3-hop radio link. R = *Receiver.* T = *Transmitter.*

The signal, which it is intended to convey, is applied at the transmitting terminals to a frequency modulator, operating at some frequency f_m, generally different

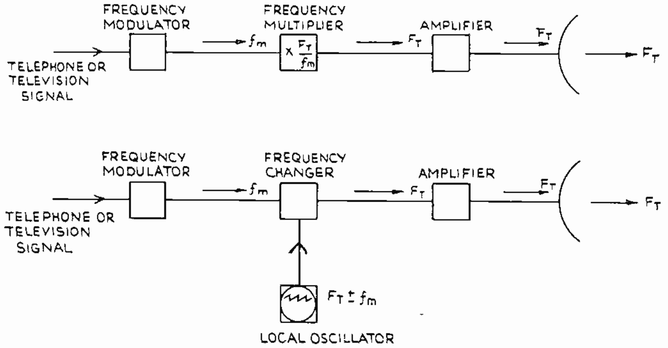

FIG. 2

Block diagram of typical terminal transmitters.

from the microwave carrier frequency F_T. The latter is obtained from f_m by frequency multiplication, frequency changing or a combination of both. Block diagrams of typical terminals are shown in Fig. 2.

The receiving terminal, Fig. 3, consists of a microwave amplifier and a frequency changer, to shift the carrier frequency f_R to some selected IF frequency f_i, followed by a frequency demodulator, which recovers the wanted signal from the modulated IF carrier. Demodulation is, at present, generally carried out at IF, as no microwave demodulator of adequate quality is as yet available.

Intermediate repeaters are usually of the non-demodulating type wherein the received carrier is amplified, frequency changed, amplified again and retransmitted as shown in Fig. 4. A demodulating type of repeater consists of a terminal receiver and a terminal transmitter connected back to back as shown in Fig. 5. This type of repeater is normally avoided because of the excessive distortion produced by the

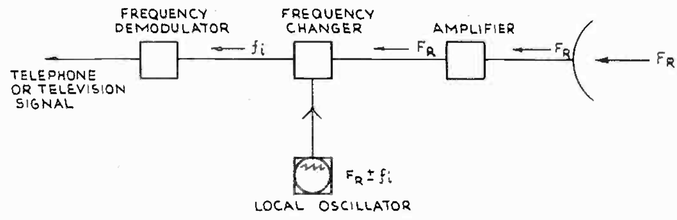

FIG. 3

Block diagram of typical terminal receiver.

repeated modulation and demodulation processes. It is only used where full terminal facilities, i.e. access to the transmitted signal and reinsertion of a fresh one, are required. In the case of multichannel links, this type of repeater provides for the dropping of some, or all of the channels and the reinsertion of fresh ones.

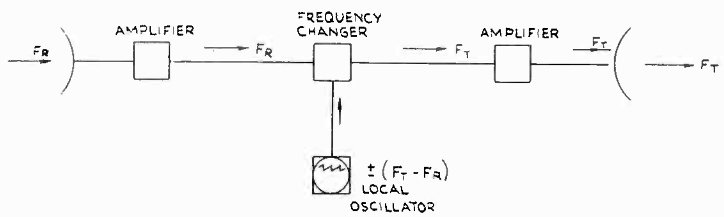

FIG. 4

Block diagram of non-demodulating repeater.

The facility for dropping and inserting channels at repeaters is sometimes required and this may be done, without necessitating the use of back to back terminals, provided the band occupied by the channels it is desired to insert is not already

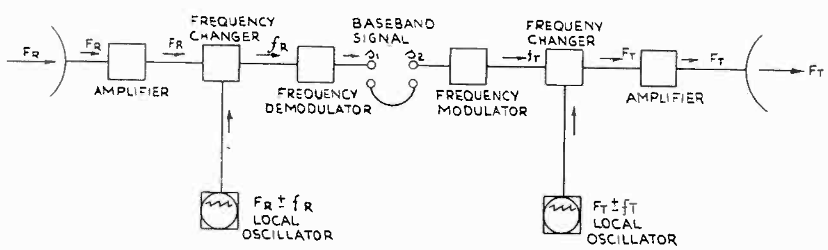

FIG. 5

Block diagram of demodulating type of repeater.

occupied. For example, in a system carrying ten supergroups between A and C (Fig. 6) it is possible to insert nine supergroups at A and the tenth at B. This arrangement provides then a traffic capacity of one supergroup between B and C without the necessity of modulating and demodulating the through traffic at B.

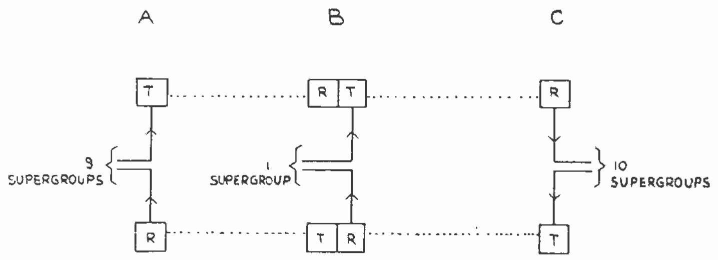

FIG. 6

Link providing for the extraction and insertion of channels at an intermediate repeater.

A block diagram of a typical repeater, used for inserting or dropping channels, is shown in Fig. 7.

In the case of a link carrying a television signal there is no question of dropping or inserting channels. It may, however, be required to make the television signal available at B for rebroadcasting, for example, in which case the lower part of Fig. 7 would apply when a television signal is being transmitted from B to A. Further

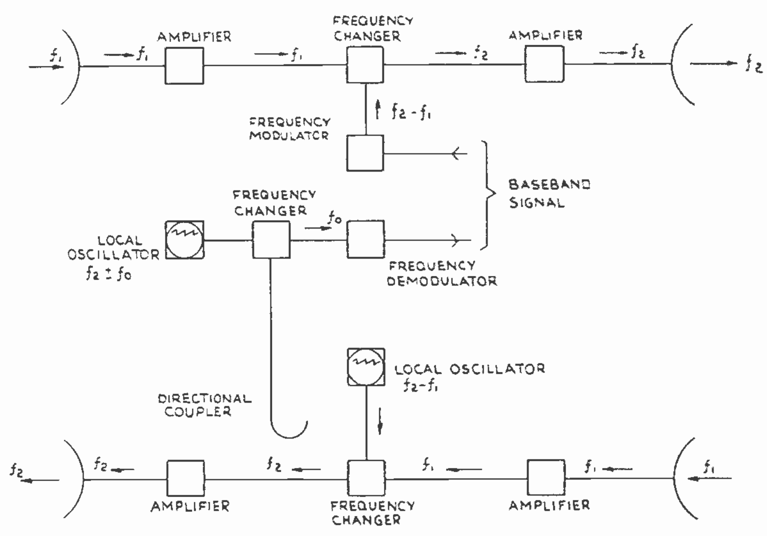

FIG. 7

Through repeater providing the facility for inserting and dropping channels.

facilities for inserting a television signal at B, an outside broadcast, for example, may be provided as shown in the top part of Fig. 7, provided no signal is being applied at A.

The Microwave Terminals

Mention has been made above of the various ways in which the modulation may be applied to the carrier.

A method used in systems of fairly low capacity consists of applying the intelligence signal to a frequency modulator operating at some relatively low frequency, such as 20 or 30 Mc/s. The modulated carrier is then applied to a frequency

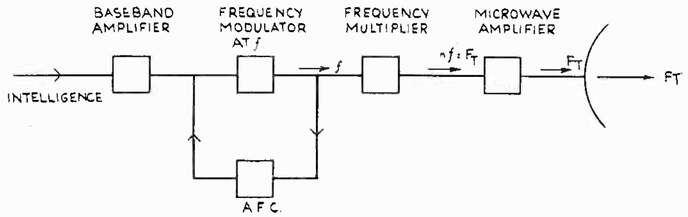

FIG. 8

Block diagram of terminal using low frequency modulator and frequency multiplier.

multiplier, the output frequency of which is the wanted carrier frequency. A block diagram of such a system is shown in Fig. 8. This type of system has the advantage

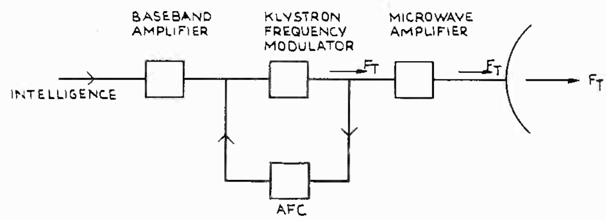

FIG. 9

Block diagram of terminal using klystron modulator operating at the required output microwave frequency.

of requiring the modulator to operate at 1/nth only of the frequency deviation of the microwave signal, where n is the frequency multiplication. For example if a multiplication of 100 times is used in the multiplier and the required rms deviation of the output carrier is 1 Mc, then the rms deviation at the modulator is only 10 kc/s. The design of a suitable modulator is thus considerably eased.

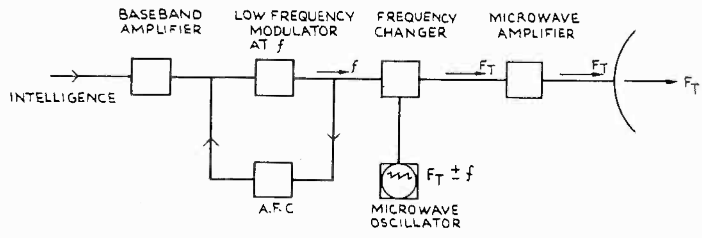

FIG. 10

Block diagram of terminal using low frequency modulator followed by frequency changer.

The high value of frequency multiplication required in this system is only satisfactorily obtained in low capacity systems. It also has the disadvantage of requiring a modulator amplifier chain that is capable of being tuned over wide bands in order to generate the required output frequencies.

An alternative method of obtaining the necessary carrier modulation at a terminal is by direct frequency modulation of a klystron generator[4]. Klystrons may be designed and operated in such a way as to be quite linear. The block diagram of such a system is shown in Fig. 9.

A third method consists in using a low frequency modulator as in the first method operating at some frequency f, and in frequency changing, subsequently, to any desired microwave frequency F. A block diagram of such a system is given in Fig. 10. The design of a low frequency modulator capable of providing the large deviations demanded in television and the reduced deviations, with the required degree of linearity, demanded in multi-channel systems of high traffic capacity is not easy. The problem is sometimes solved[5], by obtaining the required deviation, with a klystron, at some fixed microwave frequency and beating down to the wanted low frequency f as shown in Fig. 11. This latter method is rather cumbersome but it does provide a fully modulated signal, at some selected IF frequency, when this cannot be obtained by other means with the required degree of linearity. The rest of the microwave terminal may then be as shown in Fig. 10, with the advantage of providing any desired output frequency by the suitable selection of a microwave driving source. A microwave terminal may then be provided with input terminals at a fixed IF (say 70 Mc/s) independently of the exact value of the output frequency. Similarly a microwave receiving terminal is best designed with output terminals at a standard IF which could then be demodulated in a separate part of the equipment. A block diagram of a microwave terminal, built up along these lines, is given in Fig. 12. It provides transmit and receive terminals at what may be agreed upon as standard frequencies, impedances, input and output levels of 70 Mc/s, 75Ω, 0·35 v. rms and 0·70 v. rms respectively.

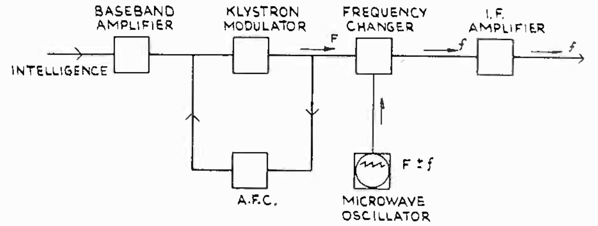

FIG. 11

Block diagram of frequency modulated oscillator supplying full deviation at low frequency f.

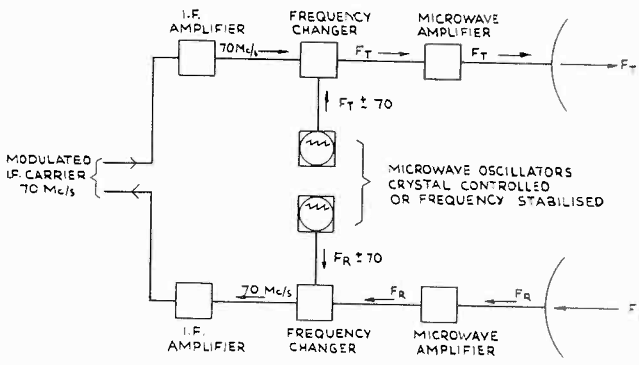

FIG. 12

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The frequency changer shown in the top of Fig. 12 may be a high level balanced crystal mixer[4, 5], or a travelling wave tube amplifier, operated as a phase modulator[6]. The operation of these two types of frequency changers is adequately explained in the above references. It may, however, be useful to recall that, in the case of the travelling wave tube, the modulated IF carrier is effectively applied between the helix and the rest of the gun-end electrodes. Sometimes the helix is

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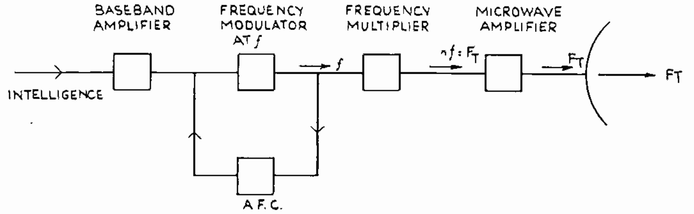

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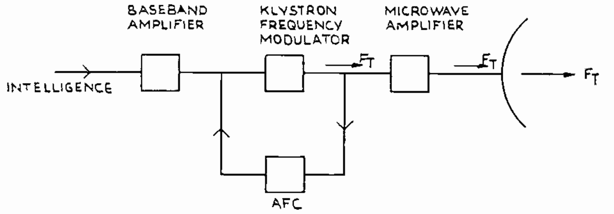

FIG. 9

Block diagram of terminal using klystron modulator operating at the required output microwave frequency.

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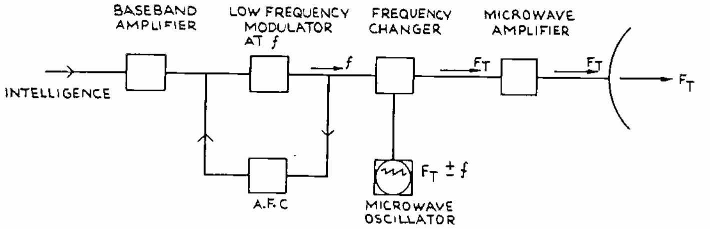

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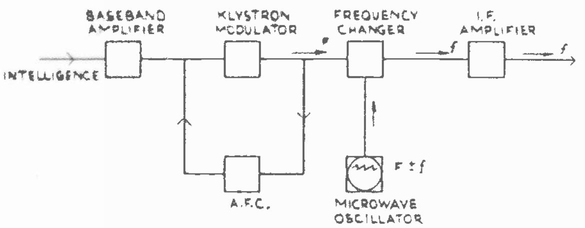
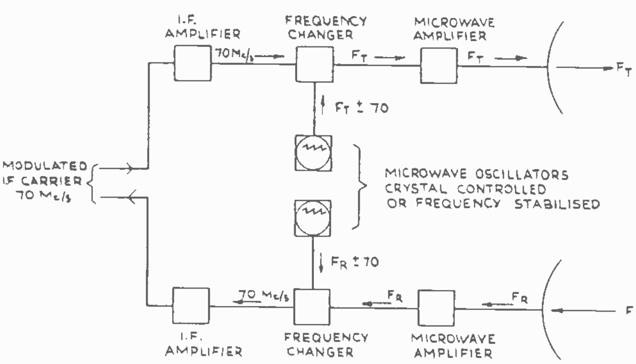


FIG. 11

Block diagram of frequency modulated oscillator supplying full deviation at low frequency f.



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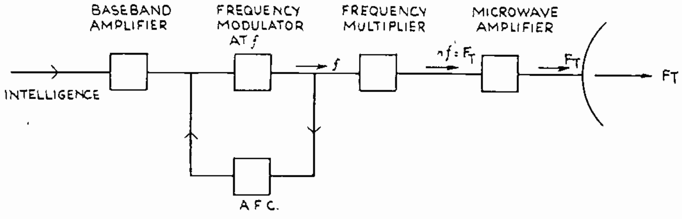

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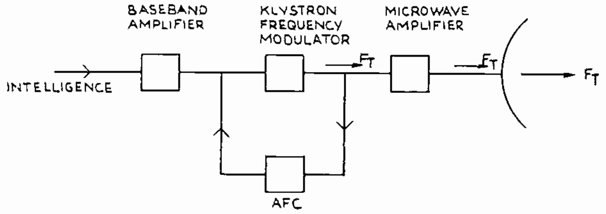

FIG. 9

Block diagram of terminal using klystron modulator operating at the required output microwave frequency.

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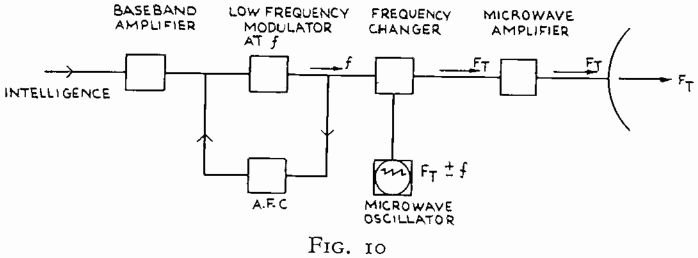

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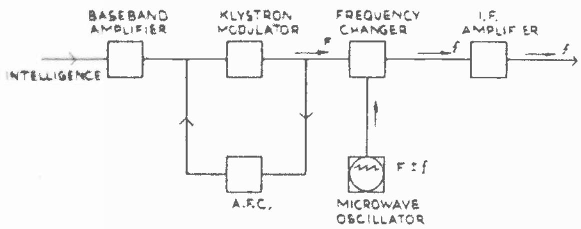

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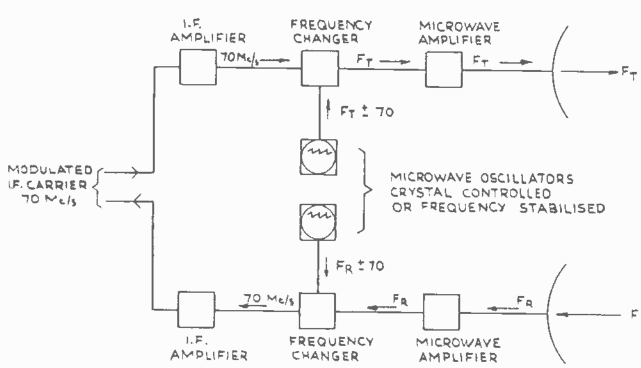

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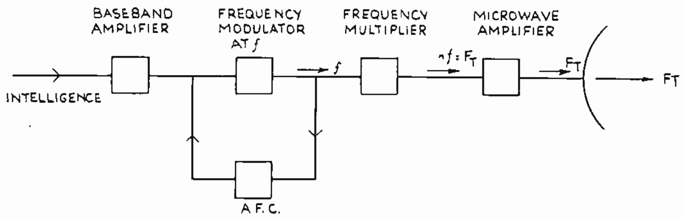

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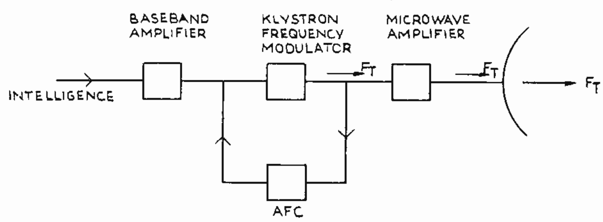

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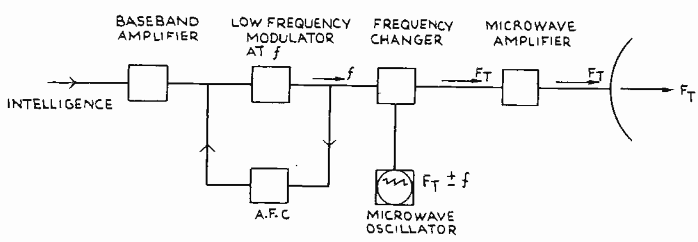

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The high value of frequency multiplication required in this system is only satisfactorily obtained in low capacity systems. It also has the disadvantage of requiring a modulator amplifier chain that is capable of being tuned over wide bands in order to generate the required output frequencies.

An alternative method of obtaining the necessary carrier modulation at a terminal is by direct frequency modulation of a klystron generator(4). Klystrons may be designed and operated in such a way as to be quite linear. The block diagram of such a system is shown in Fig. 9.

A third method consists in using a low frequency modulator as in the first method operating at some frequency f, and in frequency changing, subsequently, to any desired microwave frequency F. A block diagram of such a system is given in Fig. 10. The design of a low frequency modulator capable of providing the large deviations demanded in television and the reduced deviations, with the required degree of linearity, demanded in multi-channel systems of high traffic capacity is not easy. The problem is sometimes solved(5), by obtaining the required deviation, with a klystron, at some fixed microwave frequency and beating down to the wanted low frequency f as shown in Fig. 11. This latter method is rather cumbersome but it does provide a fully modulated signal, at some selected IF frequency, when this cannot be obtained by other means with the required degree of linearity. The rest of the microwave terminal may then be as shown in Fig. 10, with the advantage of providing any desired output frequency by the suitable selection of a microwave driving source. A microwave terminal may then be provided with input terminals at a fixed IF (say 70 Mc/s) independently of the exact value of the output frequency. Similarly a microwave receiving terminal is best designed with output terminals at a standard IF which could then be demodulated in a separate part of the equipment. A block diagram of a microwave terminal, built up along these lines, is given in Fig. 12. It provides transmit and receive terminals at what may be agreed upon as standard frequencies, impedances, input and output levels of 70 Mc/s, 75Ω, 0·35 v. rms and 0·70 v. rms respectively.

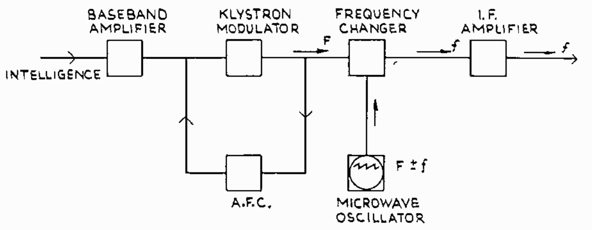

FIG. 11

Block diagram of frequency modulated oscillator supplying full deviation at low frequency f.

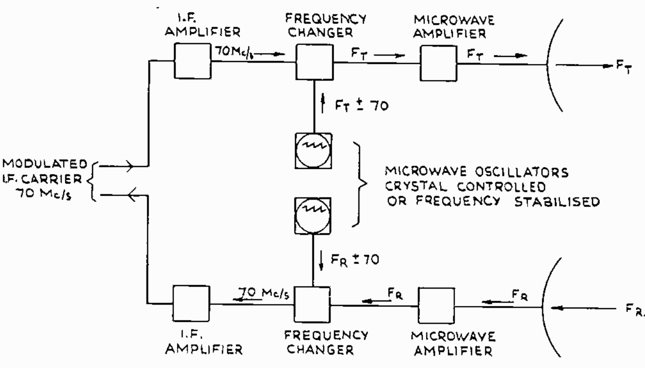

FIG. 12

Block diagram of microwave terminal with receive and transmit terminals at IF.

The frequency changer shown in the top of Fig. 12 may be a high level balanced crystal mixer(4, 5), or a travelling wave tube amplifier, operated as a phase modulator(6). The operation of these two types of frequency changers is adequately explained in the above references. It may, however, be useful to recall that, in the case of the travelling wave tube, the modulated IF carrier is effectively applied between the helix and the rest of the gun-end electrodes. Sometimes the helix is

A method used in systems of fairly low capacity consists of applying the intelligence signal to a frequency modulator operating at some relatively low frequency, such as 20 or 30 Mc/s. The modulated carrier is then applied to a frequency

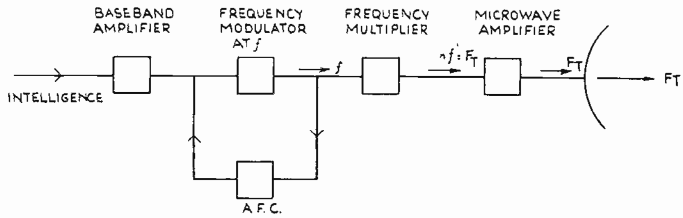

FIG. 8

Block diagram of terminal using low frequency modulator and frequency multiplier.

multiplier, the output frequency of which is the wanted carrier frequency. A block diagram of such a system is shown in Fig. 8. This type of system has the advantage

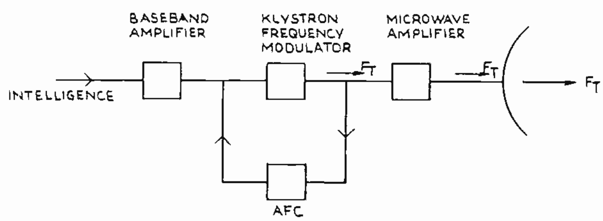

FIG. 9

Block diagram of terminal using klystron modulator operating at the required output microwave frequency.

of requiring the modulator to operate at 1/nth only of the frequency deviation of the microwave signal, where n is the frequency multiplication. For example if a multiplication of 100 times is used in the multiplier and the required rms deviation of the output carrier is 1 Mc, then the rms deviation at the modulator is only 10 kc/s. The design of a suitable modulator is thus considerably eased.

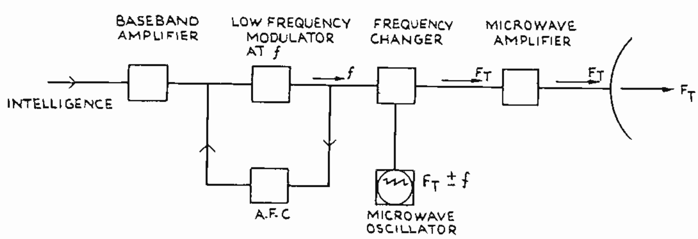

FIG. 10

Block diagram of terminal using low frequency modulator followed by frequency changer.

The high value of frequency multiplication required in this system is only satisfactorily obtained in low capacity systems. It also has the disadvantage of requiring a modulator amplifier chain that is capable of being tuned over wide bands in order to generate the required output frequencies.

An alternative method of obtaining the necessary carrier modulation at a terminal is by direct frequency modulation of a klystron generator[4]. Klystrons may be designed and operated in such a way as to be quite linear. The block diagram of such a system is shown in Fig. 9.

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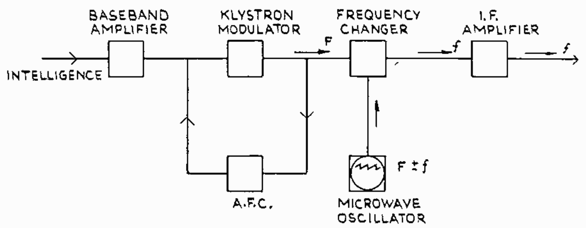

FIG. 11

Block diagram of frequency modulated oscillator supplying full deviation at low frequency f.

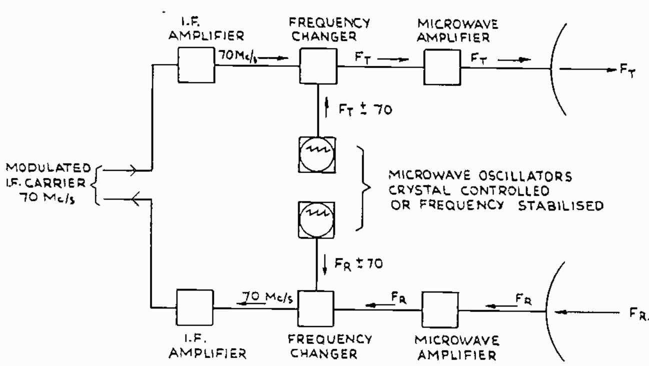

FIG. 12

Block diagram of microwave terminal with receive and transmit terminals at IF.

The frequency changer shown in the top of Fig. 12 may be a high level balanced crystal mixer[4, 5], or a travelling wave tube amplifier, operated as a phase modulator[6]. The operation of these two types of frequency changers is adequately explained in the above references. It may, however, be useful to recall that, in the case of the travelling wave tube, the modulated IF carrier is effectively applied between the helix and the rest of the gun-end electrodes. Sometimes the helix is

connected, within the tube, to a structure of high capacity to earth and it is desirable, in such cases, to earth the helix and to apply the modulating IF voltage to the cathode, heater and anode, strapped together, at IF.

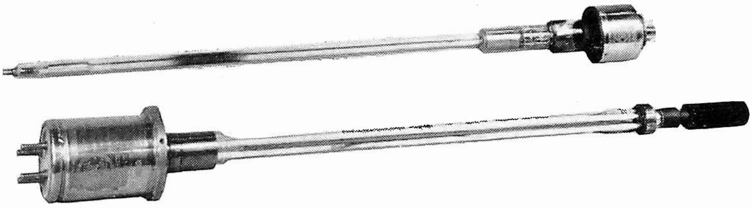

FIG. 13
Travelling wave tubes operating in 2000 Mc/s band.

If we first assume that the applied IF carrier is not modulated, the sinusoidal variations in the helix potential, cause a periodic variation in the velocity of the electron beam which results in a periodic phase change of a microwave carrier of frequency F, which is simultaneously applied to the tube. Thus the microwave carrier becomes phase modulated, at a rate equal to the IF frequency f, and energy is generated at sideband frequencies $F \pm nf$, where n is the order of the sideband. The required frequency $F+f$, say, may then be selected at the output of the travelling wave tube with the aid of a filter. The conversion loss of a travelling wave tube frequency changer is of the order of 5 db when the first sideband is selected and it compares favourably with the conversion loss of crystal mixers. When the IF carrier f is frequency modulated, the selected sideband $F \pm f$ is likewise frequency modulated.

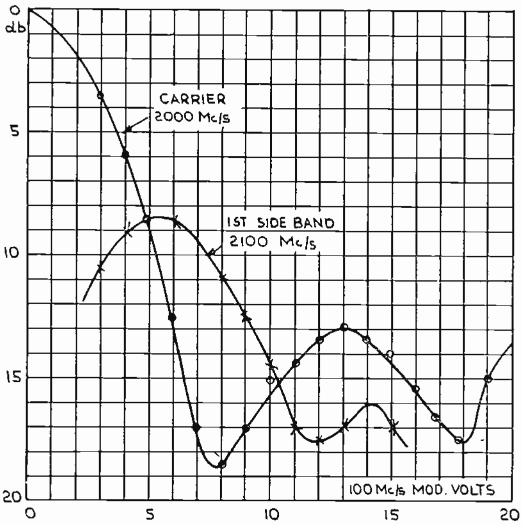

FIG. 14
Relation between IF drive and 1st sideband amplitude for a typical frequency changing travelling wave tube.

The amount of phase modulation produced in a travelling wave tube depends on the amplitude of the IF carrier. The latter is determined by the beam velocity, which is a function of the accelerating voltage, and by a number of other parameters depending on the construction of the tube. However, the lower the beam voltage, the lower is the required IF drive at the helix for a given modulation index. The frequency of the IF drive is also of importance as it determines the ratio of the IF voltage applied to the external terminals of the tube to the IF voltage developed between the actual helix and cathode of the tube. This ratio depends on the internal construction of the tube and it has to be determined empirically. Fig. 14 gives the relationship between applied IF drive and 1st sideband amplitude obtained in one type of travelling wave tube. It is obviously desirable to keep this drive to the absolute minimum, in order to ease the design of the driving amplifier. The phase change $\delta\phi$ of a microwave carrier, applied to the input of travelling wave tube, consequent upon a small change δV_0 of the helix to cathode potential is

$$\delta\phi = -A\delta V_0 \text{ where } A = \frac{\pi lc}{\lambda v V_0}$$

l=effective axial length of helix, c=velocity of light, v=beam velocity, λ=wavelength and v_0=helix to cathode potential.
If n is the effective length of the helix wire in terms of the wavelength λ, we have

$$\frac{l}{v} = \frac{n\lambda}{c}$$

and therefore

$$A = \frac{\pi n}{V_0}$$

For a typical power type travelling wave tube, operated in the small signal condition, n=30, V_0=2400V, and therefore $A \doteqdot 0{\cdot}04$ radians per volt. In the case of a typical low power, general purpose tube, n=40 and V_0=650, so that $A \doteqdot 0{\cdot}2$ radians per volt.

The minimum conversion loss (4·7 db) is obtained for a modulation index of 2·86 and the selection of the first order sideband. Thus the rf drives required for the above tubes are approximately 35 volts rms and 6·5 volts rms respectively.

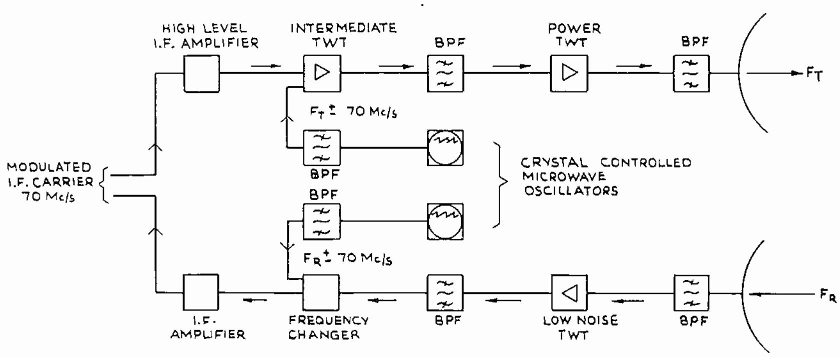

FIG. 15
Block diagram of typical microwave terminal using travelling wave tubes.

The block diagram of a typical microwave terminal, using travelling wave tubes is shown in Fig. 15.

The receiving microwave terminal comprises a low noise travelling wave tube amplifier, followed by a low level silicon crystal mixer. There are great advantages in using a microwave amplifier of the travelling wave tube type, ahead of the crystal mixer. The matching of a travelling wave tube to the input aerial feeder is considerably easier and gives a broader band than that of a crystal mixer and the noise figure is, in general, better than that of a crystal mixer followed by a broad band IF amplifier.

The Microwave Repeater

Microwave repeaters have so far been of the IF type, wherein the input signal is frequency changed down to some IF frequency, amplified in conventional IF circuits and frequency changed back to another microwave frequency. A block diagram of a non-demodulating IF type of repeater is given in Fig. 16. This consists, essentially, of a microwave receiving terminal and a microwave transmitting terminal placed back to back. This type of repeater lends itself very easily to the extraction and reinsertion of a signal at IF (for example at points AA), so that it is possible, at any repeater station, to have access at IF to the traffic signal which is being conveyed and to insert a fresh one, for example a TV outside broadcast, or an additional one, if required.

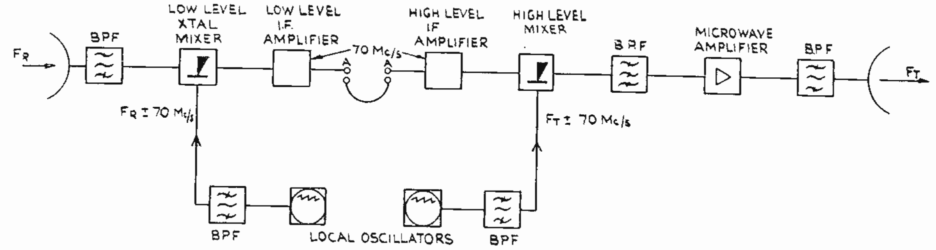

FIG. 16

One-way, non-demodulating, IF type, microwave repeater.

A simpler type of non-demodulating repeater, not using conversion to, and amplification at, IF, uses travelling wave tubes as amplifiers and frequency changers. A block diagram of this type of repeater, using four tubes, is shown in Fig. 17. Here three types of tubes are used; a low noise tube T_1, two intermediate tubes T_2 and T_3 and a power output tube T_4.

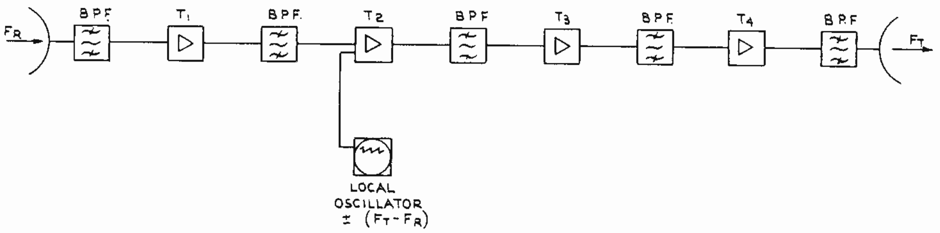

FIG. 17

One-way, non-demodulating, all TWT repeater.

An illustration of an experimental all-travelling wave tube, non-demodulating repeater, operating in the 2000 Mc/s band is shown in Fig. 18. This type of repeater does not lend itself readily to the extraction or the insertion of additional or alternative traffic, without the provision of ancillary equipment. It is possible, of course, to monitor the traffic signal by extracting a small fraction of the microwave signal at some high level point at the output of T_3 or T_4. This signal may then be demodulated down to IF or to video, if required, by the addition of a microwave mixer, local oscillator and IF amplifier. The traffic signal may be recovered by the use of a limiter-discriminator and a baseband amplifier.

Injection of an additional signal may be achieved by modulating the frequency of the heterodyne oscillator. Since it is desirable to insert traffic at a fixed IF (70 Mc/s say), a double heterodyne is necessary to ensure that the desired frequency separation $F_T - F_R$ is obtained.

It is not normally necessary to provide full modulation and demodulation facilities at all repeaters of a microwave link, so that the arrangement of Fig. 19 is

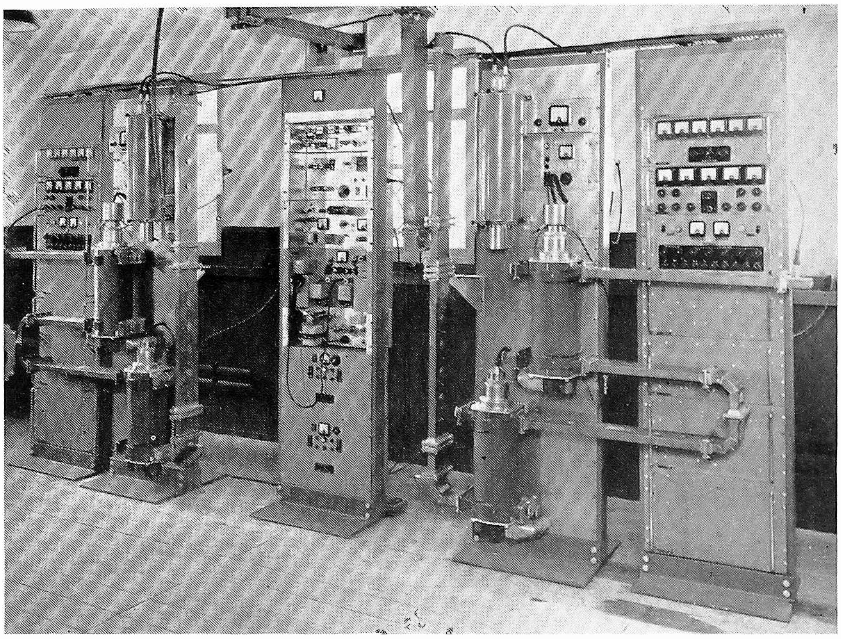

FIG. 18
Experimental all TWT non-demodulating repeater, operating in 2000 Mc/s band.

only likely to be used on rare occasions. At important traffic centres it is more likely that a full terminal would be used.

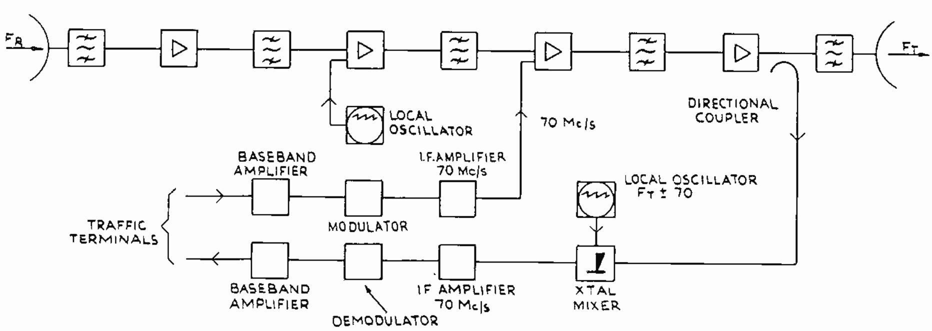

FIG. 19
Non-demodulating repeater providing for the extraction and insertion of traffic, additional to through traffic.

Type of Travelling Wave Tubes

A range of travelling wave tubes, suitable for use in wideband links, is now available for the 2000 Mc/s band (Fig. 13). The power tube is fully described in a paper by Coulson and Robinson(7). A short summary of the important characteristics of the tubes in this range is given here.

The low noise tube N1002 has a gain of about 23 db and a noise figure of the order of 10 db, under operational conditions. The limiting power of an N1002 is of the order of 1 mW.

The intermediate type tube N1013 has a gain of about 38 db, when operated under low level conditions, i.e. when delivering not more than 50 mW of power and a gain of 27 db when delivering over 200 mW. The noise figure of this tube is of the order of 20 db.

The power tube N1001 has a gain of 27 db, when delivering a maximum power output of 20 watts. A four tube repeater has, therefore, an average gain of about 110 db, which is considerably more than required under normal operating conditions. Since one of the intermediate tubes is to be used as a frequency changer, a loss of about 10 db should be allowed, so that the overall gain of a repeater is 100 db, when maximum power is being delivered. It is in general desirable to frequency change at the second tube and not at the third, since in addition to the loss in gain obtained by frequency changing there is also a loss in maximum output power. Tube type N1013 is, however, designed fully to saturate a type N1001 tube, even when used as a frequency changer.

The noise figure of a travelling wave tube frequency changer is 4 to 7 db higher than its noise figure when used as a straight amplifier, according to whether the lower or upper sideband is to be selected. In either case the output noise is made up, in the main, of five noise contributions. They are the equivalent input valve noise powers at the input, output, first order image and second order image frequencies; the weighting of these contributions being as $0{\cdot}58^2$ (input frequency), $0{\cdot}40^2$ (output frequency), $0{\cdot}43^2$ (1st order image frequency), $0{\cdot}29^2$ and $0{\cdot}28^2$ (2nd order image frequencies). These weighting figures are obtained from Fig. 5 of Ref. 6. The selected output signal has a weighting of $0{\cdot}58^2$ or $0{\cdot}43^2$, according to whether the lower or upper sideband is desired.

When mixing takes place in the second tube of a repeater, the difference between the noise figure of the complete repeater and that of the low noise tube alone, is between 0·5 and 1·0 db for the average performance figures quoted above.

Power, Signal/Noise Ratios and Gains of Repeaters

In an FM system the output rms signal to rms noise ratio W_r'/W_n', after demodulation, referred to band of width b, is related to the input rms signal to rms noise ratio W_r/W_n, referred to the same bandwidth by the formulae:

$$\frac{W_r'}{W_n'} = \frac{3}{2}\frac{W_r}{W_n}\left(\frac{\triangle F}{f_b}\right)^2 \qquad (1)$$

$$\frac{W_r'}{W_n'} = \frac{1}{2}\frac{W_r}{W_n}\left(\frac{\triangle F}{f_b}\right)^2 \qquad (2)$$

Equation (1) applies when the signal to noise ratio is referred to the whole of the baseband from zero frequency to the top modulation frequency f_b, as is normally the case for television systems. $\triangle F$ is then the peak deviation caused by the television signal. For multi-channel systems the signal to noise ratio is normally referred to a channel bandwidth b of 4 kc/s. In this case $\triangle F$ is the channel test tone peak deviation and f_b is also the top modulation frequency, since the top channel is usually the noisiest for systems without pre-emphasis.

The microwave system may therefore be designed on the basis of a carrier signal to noise ratio before demodulation, referred to a channel bandwidth of 4 kc/s for telephone systems, or a video bandwidth of 5 Mc/s for a monochrome 625 line, television system.

The calculation of the carrier signal to noise ratio will be made for a microwave system having the following characteristics.

Frequency ...	2000 Mc/s.
Transmitter Power	20 watts.
Path length	35 miles (56 km).
Aerial diameter	10 ft.
Filter and feeder loss	4 db (both ends).
Receiver noise figure	12 db.
Channel bandwidth	4 kc/s (telephony).
	5 Mc/s (television).

The path attenuation L_p between half-wave dipoles is

$$L_p = 45\,\mathrm{D.f.}$$

where D=distance in miles and f=frequency in Mc/s. For D=35 miles and f= 2000 Mc/s.

$$20 \log_{10} L_p = 130 \text{ db.}$$

The power gain of a parabolic dish aerial, over a dipole, is

$$G = \frac{K}{1{\cdot}65} \frac{4\pi A}{\lambda^2}$$

where K is the efficiency of illumination, usually taken as 50%, A is the area of the aperture and λ the wavelength. For A=10 ft. and λ=15 cm.

$$10 \log_{10} G \doteqdot 30{\cdot}5 \text{ db.}$$

From this it follows that the effective attenuation between the transmitter output terminals at one repeater and the receiver input terminals at the next repeater is 69 db for normal propagation conditions and no losses in feeders and filters and 73 db when the assumed filter and feeder losses are taken into account. The latter figure specifies the normal repeater gain. The normal receiver input power, for a transmitted output power of 20 watts is 1 μW (−60 dbW), and the input valve noise power, at average room temperature, referred to a bandwidth of 4 kc/s, is, for a noise figure of 12 db

$$10 \log_{10} W_n = -156 \text{ dbW.}$$

It follows that the carrier signal to noise ratio, in a single hop is

$$10 \log_{10} \frac{W_r}{W_n} = 96 \text{ db for a 4 kc/s bandwidth}$$

and

$$10 \log_{10} \frac{W_r}{W_n} = 65 \text{ db for a 5 Mc/s bandwidth.}$$

As an illustration of the significance of these figures we shall assume a rms channel test tone deviation of 200 kc/s, for a 120 channel telephone system and a 6 Mc/s, d.a.p.* deviation (including sync.) for a 625-lines monochrome television system.

* d.a.p Double Amplitude Peak.

The signal to noise ratios after demodulation, are:

$$10 \log_{10} \frac{W_r'}{W_n'} = 10 \log_{10} \frac{W_r}{W_n} - 3 - 6$$

$$= 87 \text{ db for the noisiest telephone channel}$$

and

$$10 \log_{10} \frac{W_r'}{W_n'} = 10 \log_{10} \frac{W_r}{W_n} + 1{\cdot}75 - 7{\cdot}5$$

$$= 59 \text{ db for the average over a 5 Mc/s video band of a television system.}$$

The latter figure corresponds to a ratio of d.a.p. signal to d.a.p. noise of 50 db, for one hop only. For 600 channel operation and the same channel test tone deviation, the signal to noise ratio in the noisiest channel reduces to

$$87 - 20 \log_{10} \frac{2{\cdot}5}{{\cdot}5} = 73 \text{ db}$$

It will be shown, in a companion paper to be published in the next issue (No. 119) that a channel signal to noise ratio of 69 db, should be obtained in a one-hop link, when only the noise contribution of the microwave receiver is taken into account. It follows that for 120 channel operation an 18 db fading margin, per hop, may be allowed without degrading the system performance, as regards valve noise. In the case of a 600 channel system, the fading margin, per hop, is reduced to 4 db.

The fading margins that may actually be tolerated on systems of this type, are, in fact, considerably greater than the figures mentioned above, because fading at UHF[10], is usually characterised by very fast variations in signal strength, deep fades being usually of very short duration.

The figure of 69 db, mentioned above, is obtained on the basis of a 45 hop link, 1600 miles in length and corresponding to the CCIF " circuit fictif de reference " for coaxial cable circuits.

If N is the valve noise contribution of each hop and W is the signal power at the receiving end of each hop, then assuming that W is kept rigidly constant, on non-fading hops, by the automatic gain control action of the repeaters, the overall signal to valve noise ratio is $W/45N$, for normal conditions of propagation. If we have an 18 db fade in every hop, then the overall valve noise power may rise to $45 \times 63N = 2830N$, without any degradation in the link performance. It follows that a fade on a single hop at a time, may be permitted to reach a value of $2830 - 44 = 2786$ times or 34·5 db (18·5 db for 600 channels), provided it does not last for more than 0·02% of the time, without degrading the link performance. It is assumed of course that all repeaters have a gain margin of at least 35 db, in order to be capable of generating the rated maximum output of 20 watts, when receiving a low signal. This automatic gain control action need not be perfect, since a drop of a few db in the output power of the repeater, receiving the low signal, cannot add appreciably to the overall noise of the link.

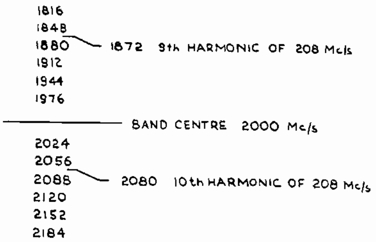

FIG. 20

Frequency allocation according to CCIR.

Frequency Allocations and Filter Selectivities

A preferred frequency allocation plan for frequency division telephony and television links has been agreed by Study Group IX of the C.C.I.R. meeting in

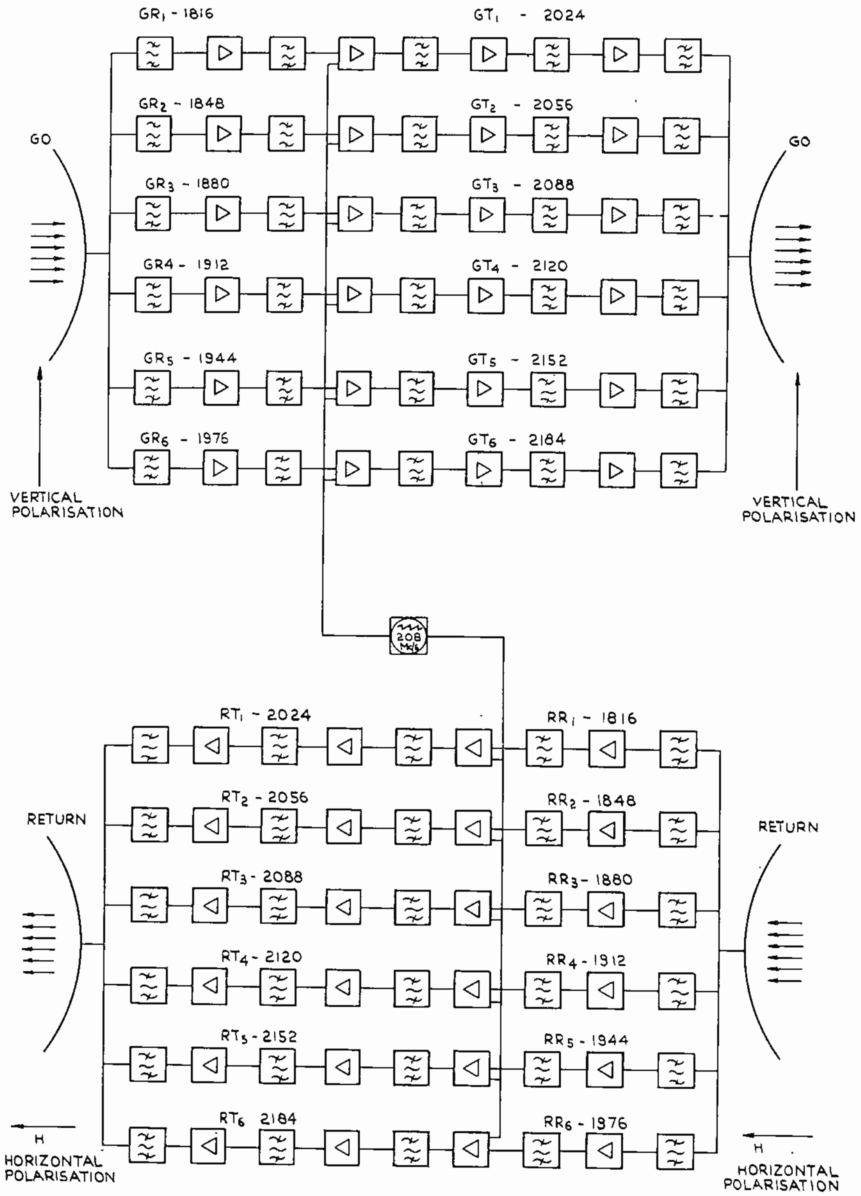

FIG. 21

6-channel repeater using four aerials.

Geneva in 1954. This plan provides for six two-way, high-capacity systems, within the last three communication bands listed on page 70.

Taking the case of a particular repeater station, there are six receiving frequencies and six transmitting frequencies. All the receiving frequencies are grouped together with a separation between consecutive frequencies of 32 Mc/s. Similarly all the transmitting frequencies are grouped together with a separation of 32 Mc/s. The separation between the nearest receive and transmit frequencies is 48 Mc/s. Thus a six-channel system occupies a bandwidth of 400 Mc/s, the centre of which is the centre of the applicable band and the difference between incoming and outgoing carriers at a repeater station is 208 Mc/s. A frequency allocation plan for the 2000 Mc/s band is shown in Fig. 20.

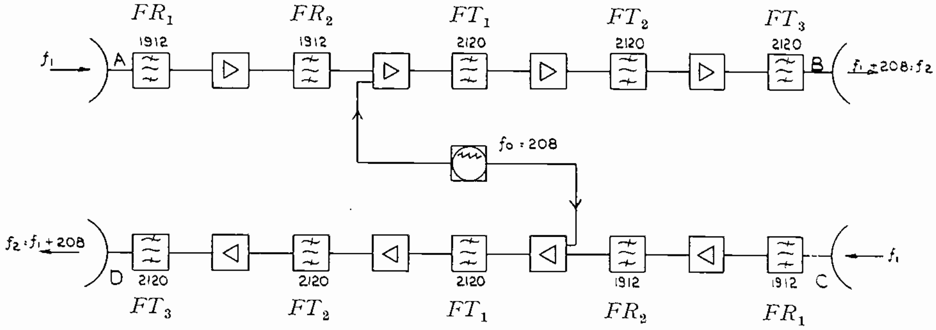

FIG. 22

Single channel, 2-frequencies repeater, using four aerials (one of 6 repeaters).

Provision is also made to receive from the two directions at the same frequencies but at rectangular polarisations in order to increase the back to front aerial protection. Transmission in the two directions is also at the same frequencies and at rectangular polarisations. A block diagram of a six-channel all travelling wave tube repeater station, using four aerials, is given in Fig. 21.

The necessary filter selectivities at repeater stations may be derived by applying a number of criteria as given below. It will be assumed that an individual repeater has the configuration of Fig. 22. Actual frequencies have been selected from the frequency plan of Fig. 20 by way of illustration. For the purpose of brevity the attenuation of a filter, at a frequency f away from its resonant frequency will be denoted by $A_{(f)}$, followed by the symbol denoting the particular filter. For example the attenuation of filter FR_1 at a frequency 208 Mc/s away from its centre frequency, is denoted by $A_{(208)} FR_1$. The side by side aerial attenuation is denoted by A_s, the back to back attenuation by A_b and the back to front ratio by A_{fb}.

Prevention of Oscillations

The loop gain of the travelling wave tubes, used in a repeater, is, under non-limiting conditions, approximately 200 db. Therefore the attenuation of the filters, aided by the protection of the aerials, should be greater than 200 db, at any frequency. The transmitted frequency f_2 suffers the attenuation of two sets of receive filters, added to the protection of the two sets of aerials, before arriving at the starting point. Thus we should have:—

$$2A_{(fo)} FR_1 + 2A_{(fo)} FR_2 + 2A_s \nless 200 \text{ db.}$$

If we assume $A_s = 50$ db, then

$$A_{(fo)} FR_1 + A_{(fo)} FR_2 \nless 50 \text{ db.}$$

The transmitted frequency F_2, at B, may, after breaking through the two receive filters of branch CD, be frequency changed to f_1. Thus we should have also:—

$$2A_s + A_{(fo)} FR_1 + A_{(fo)} FR_2 + A_{(fo)} FT_1 + A_{(fo)} FT_2 + A_{(fo)} FT_3 \nless 100 \text{ db}$$

Obviously this condition is nearly met by the aerial attenuation alone.

The frequency f_2 at B, may arrive at A through the back to back coupling of the aerials. This condition gives

$$A_b + A_{(fo)} FR_1 + A_{(fo)} FR_2 \nless 100 \text{ db}$$

Again if we assume $A_b = 50$ db, then we obtain as before

$$A_{(fo)} FR_1 + A_{(fo)} FR_2 \nless 100 \text{ db}$$

Prevention of Cross Modulation

The power of the transmitted signal breaking through to the input of the receiver, on the same side, must be less by about 15 db, than the power required to cause the input tube to limit. This power is about -55dbW and, since the maximum power of the transmitter is about $+13$ dbW, the first receive filter should have an attenuation of about 33 db, at the transmit frequency. The most difficult case is that of GT_1 and RR_6, and RT_1 and GT_6 Fig. 21, where the separation between transmit and receive frequencies is 48 Mc/s only. An allowance of 50 db has already been made for the side by side aerial protection. The most stringent case is therefore satisfied if

$$A_{(48)} FR_1 \nless 35 \text{ db.}$$

Another mechanism by which cross modulation may occur is as follows. The transmitted signal at B may, by passing through the branch CD, appear at A, at the same frequency as the wanted signal. This breakthrough signal is of the nature of an echo, with considerable phase and amplitude distortion. Provided, however, its level does not exceed -60 db reference the level of the wanted signal, it need not cause any distortion. It follows that the attenuation of branch CD should exceed 230 db at a frequency 208 Mc/s away from centre frequency. This figure should be reduced by twice the aerial protection A_s to give:

$$A_{(fo)} FR_1 + A_{(fo)} FR_2 + A_{(fo)} FT_1 + A_{(fo)} FT_2 + A_{(fo)} FT_3 \nless 130 \text{ db.}$$

Another form of cross modulation or excessive distortion is caused by the passage of a given carrier frequency through the wrong repeater. Since incoming frequencies are spaced 32 Mc/s apart, it follows that the attenuation of a path such as AB, excluding the gains of the travelling wave tubes, should be about 100 db. Thus

$$A_{(32)} FR_1 + A_{(32)} FR_2 + A_{(32)} FT_1 + A_{(32)} FT_2 + A_{(32)} FT_3 \nless 100 \text{ db.}$$

Reduction of Spurious Signals and Noise

The breakthrough of the transmitted frequency f_2, into branch CD, appears after the frequency changer at approximately the same frequency as the wanted signal. The beat frequency appearing at the terminal receiver must be reduced to the noise level. Thus an attenuation of about 80 db below the level of the wanted signal is required; and we should have

$$A_s + A_{(fo)} FR_1 + A_{(fo)} FR_2 \nless 80 + 90 = 170 \text{ db.}$$

on the assumption of a difference in level, between incoming and outgoing carriers, of 90 db max. The above condition may be rewritten

$$A_{(fo)} FR_1 + A_{(fo)} FR_2 \nless 120 \text{ db.}$$

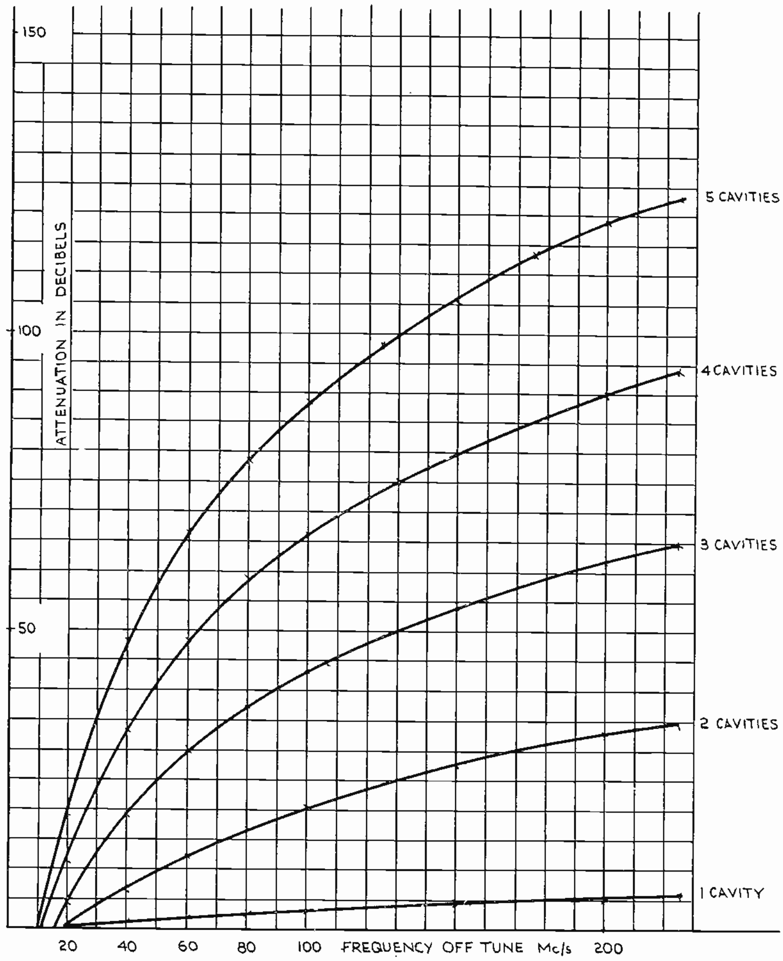

FIG. 23

Selectivity curves for maximally flat filters, tuned to 2000 Mc/s and having a bandwidth of 15 Mc/s to V.S.W.R. of 1 db.

Similarly the breakthrough of the carrier f_1 at A through AB to C, should be about 80 db below the level of the wanted carrier at C. Thus we should have

$$A_s + A_{(fo)} FT_1 + A_{(fo)} FT_2 + A_{(fo)} FT_3 \nless 80 + 100$$

and

$$A_{(fo)} FT_1 + A_{(fo)} FT_2 + A_{(fo)} FT_3 \nless 130 \text{ db}$$

assuming that the full gain of the branch AB is 100 db.

Again the received carrier at A may get to C through the back radiation of the C aerial. This therefore calls for an aerial back to front ratio of about 80 db, measured on site, to take into account unwanted reflections from existing obstacles.

The overall requirements on the filter selectivities are summarised below:—

$$A_{(48)}\, FR_1 \nless 35\,\text{db}$$

$$A_{(208)}\, FR_1 + A_{(208)}\, FR_2 \nless 120\,\text{db}$$

$$A_{(208)}\, FT_1 + A_{(208)}\, FT_2 + A_{(208)}\, FT_3 \nless 130\,\text{db}$$

$$A_{(32)}\, FR_1 + A_{(32)}\, FR_2 + A_{(32)}\, FT_1 + A_{(32)}\, FT_2 + A_{(32)}\, FT_3 \nless 100\,\text{db}$$

we can satisfy all these conditions by making

$$A_{(48)}\, FR_1 = 40\,\text{db}$$

$$A_{(32)}\, F \quad = 32\,\text{db}$$

assuming that all filters are identical, maximally flat, with a bandwidth of 15 Mc/s to a VSWR of 0·9. It will be seen from Fig. 23 that four cavity filters are adequate.

The additional difficulty arising at terminal stations is due to the reception into any one receiver of unwanted carriers spaced 32 Mc/s away from the wanted frequency. If we assume that there are two four cavity filters preceding the crystal mixer, the nearest unwanted carrier will be attenuated by about 60 db, so that the spurious frequencies, 38 Mc/s and 102 Mc/s, generated at the output of the mixer, are also 60 db below the wanted IF of 70 Mc/s. Harmonic and beat frequencies of 38 Mc/s and 102 Mc/s will be at least 100 db below the wanted carrier and at this level there should be little danger of causing cross modulation or spurious responses.

Filter Design

Filters used in microwave links operating in the upper UHF band and in the SHF bands, are usually of the waveguide type[10]. Three-quarter wave coupled, maximally flat, waveguide filters, using inductive rods or irises are normally employed. They suffer however from a number of defects. For one thing they are generally designed for, and used at, fixed frequencies. The tuning of these filters over wide frequency bands is not possible. Again, although the band pass response closely approximates to the theoretical at frequencies close to the design frequencies, they exhibit transparencies at frequencies not very far removed from the design pass band. These transparencies may cause serious difficulties in repeaters using travelling wave tube amplifiers exclusively, in the signal path, because of the very large band-widths of these tubes. One way of filling up these filter transparencies consists in inserting a number of very wide band resonant irises in series with these filters. Fig. 24

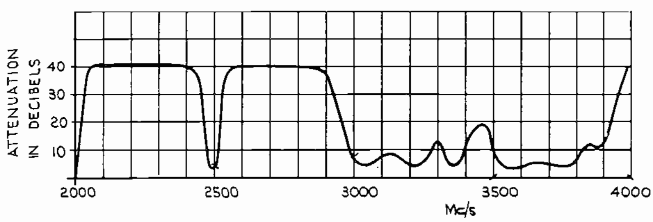

FIG. 24

Transmission in the stop band of 3λ/4 coupled waveguide band pass filters.

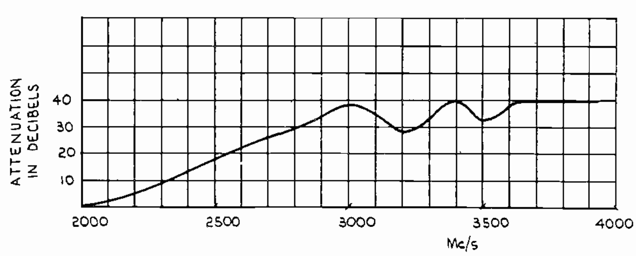

FIG. 25

Response of a pair of resonant irises.

and 25 show the response of some typical, $3\lambda/4$ coupled, waveguide filters, in the band stop region and the response of a pair of tuned irises, which must be placed in series with the waveguide filters in order to fill up the transparencies. It is possible, by using pairs of resonant irises separated by $\lambda/8$, to obtain a reduction in the losses which are normally experienced with these irises.

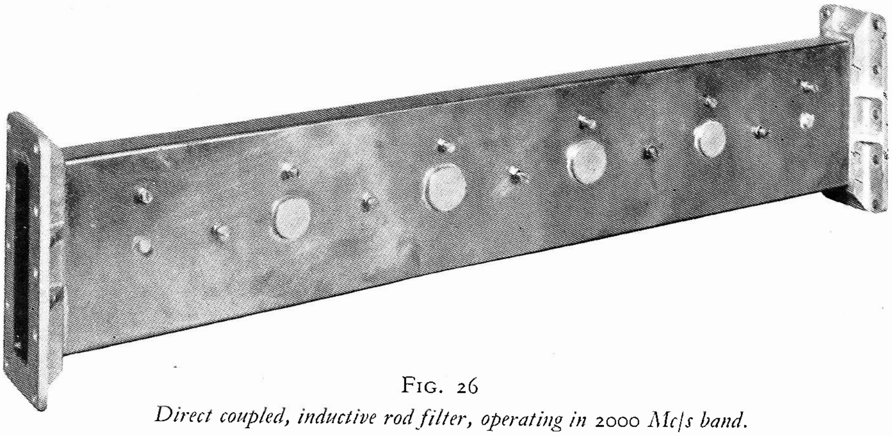

FIG. 26
Direct coupled, inductive rod filter, operating in 2000 Mc/s band.

Three-quarter wave coupled filters in waveguide operating in the 2000 Mc/s band are excessively bulky and it is preferable to use direct coupled waveguide filters. The saving in size is almost two to one. Direct coupled waveguide filters are, however, difficult to design and tune. They also suffer from unwanted transparencies in the stop band but not quite to the same extent as three-quarter wave coupled filters. A picture of a direct coupled waveguide filter for use in the 2000 Mc/s band is given in Fig. 26.

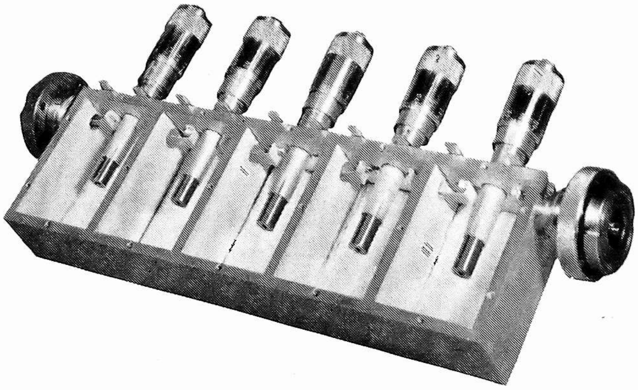

FIG. 27
Direct coupled, coaxial type filter.

Even direct coupled waveguide filters prove to be excessively bulky in the 2000 Mc/s band and their tuning to frequencies other than the design frequencies is even more difficult than the $3\lambda/4$ coupled filters.

Coaxial type filters, as used in the VHF bands, may be used successfully in the 2000 Mc/s band where the tolerances and the losses do not appear to be prohibitive. Apart from the advantages of greatly reduced size and ease of manufacture, they are tuneable over wide frequency bands, they do not suffer from unwanted transparencies closer than the third harmonic of the design frequencies and they lend themselves very easily to interconnection between travelling wave tubes. A coaxial type filter using five coupled cavities is shown in Fig. 27. The ease with which they lend themselves to interconnection between travelling wave tubes is rather important, as matching considerations may require that the position of the waveguide input and output terminals of travelling wave tubes be adjustable with respect to the tube mounts. This adjustment would of course create great difficulties if waveguide filters are used.

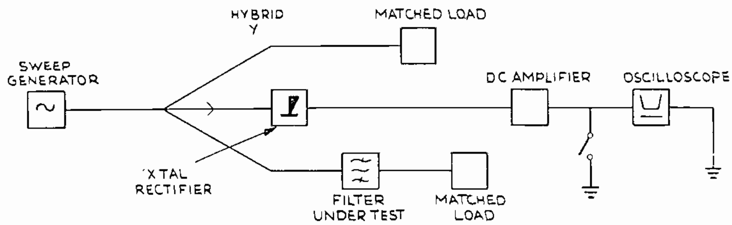

FIG. 28
Sweep equipment used for the tuning of waveguide filters.

It is possible to tune correctly designed direct coupled filters by tuning adjustments made at the design centre frequency, using Dishal's method[11]. The standing wave ratio thus obtained has a maximum value of about 0·9 over the design band.

If a better impedance match is required, it is necessary to resort to sweep methods. A successful arrangement used for setting up filters, with the help of a sweep generator, is shown in Fig. 28. A response curve obtained by the use of this tuning procedure is shown in Fig. 29.

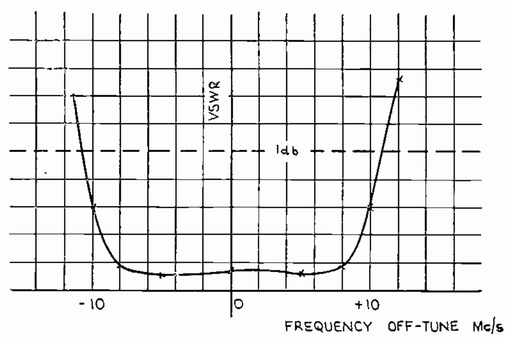

FIG. 29
Input impedance of typical coupled cavity filter, tuned by the sweep method.

Branching Filters

Branching systems for combining more than two radio channels on to a single aerial feeder are fairly complicated. A successful type evolved by the Bell Telephone Laboratories[12] makes use of sets of hybrid Y's separated by band-stop filters.

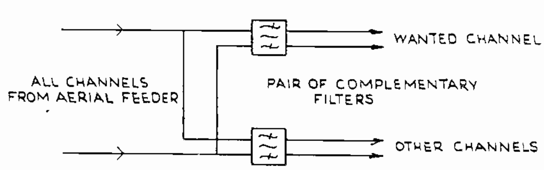

FIG. 30
Branching system using complementary filters.

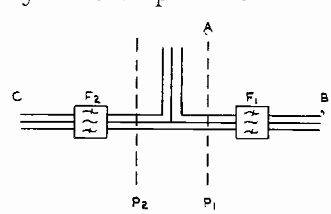

FIG. 31
Simple type of branching filter for use with two RF paths.

An important property of successful combining systems is that the combining device, used for one particular channel, should not affect in any way the devices used for combining the other channels. The Bell scheme possesses this property.

Another possible system is based on constant resistance filter pairs[13]. One such filter pair would be made up of a band stop filter and a band pass filter, in series or in parallel, according to the diagram of Fig. 30. This branching system is also applicable to any number of channels and it possesses the property mentioned above.

The difficulty met with this type of combining filter is the realization of a suitable structure in waveguide or coaxial lines.

In a case where only two radio channels are to be combined in a single aerial feeder, simpler branching networks are desirable. One such network, for use with

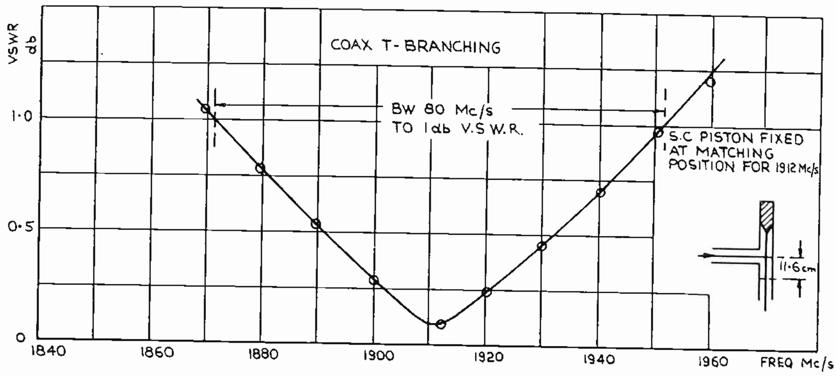

FIG. 32(*a*)

Response of coaxial T junction.

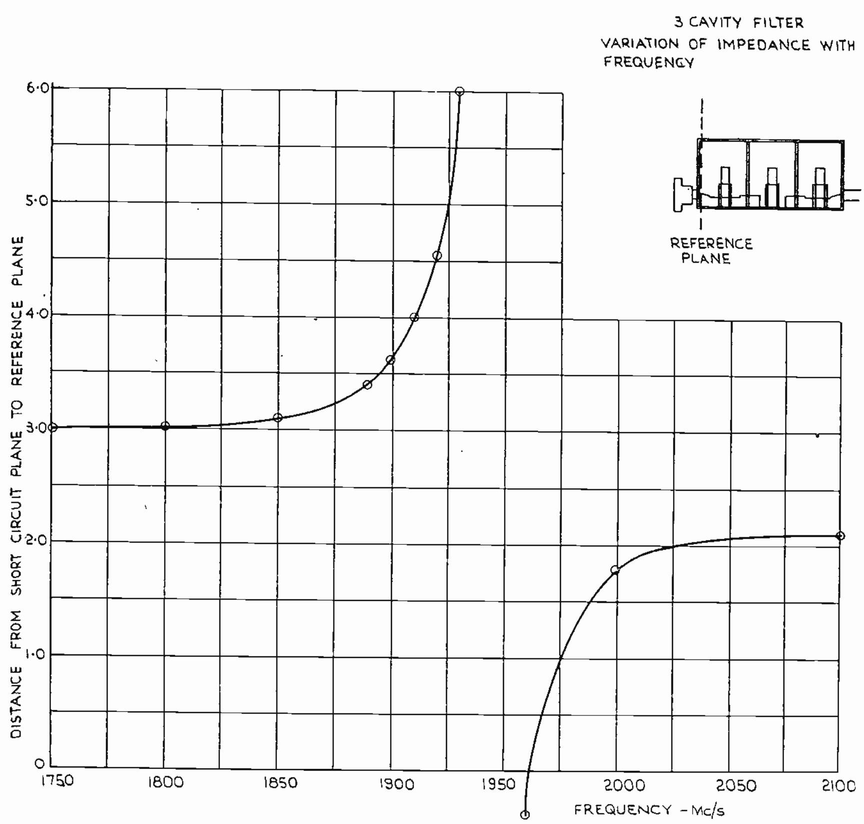

FIG. 32(*b*)

Input impedance of coaxial filter in stop bands in terms of position of virtual short circuit (resonant frequency 1944 *Mc/s).*

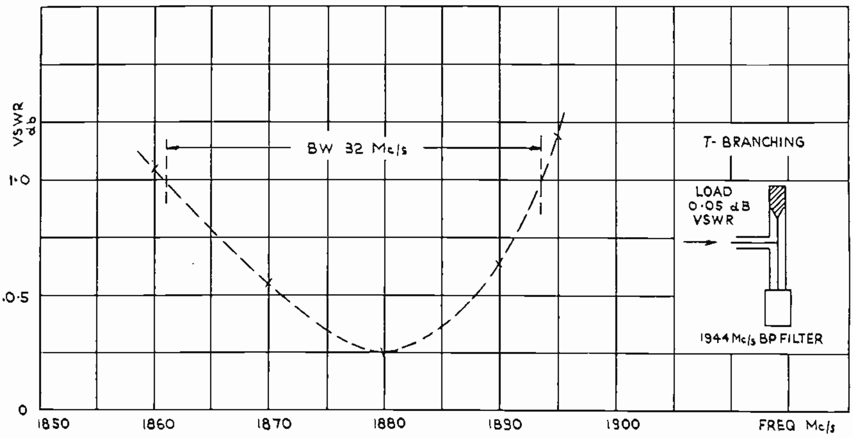

FIG. 32(*c*)

Input impedance of coaxial-T branch with filter at one end. 64 Mc/s away from resonant frequency of filter.

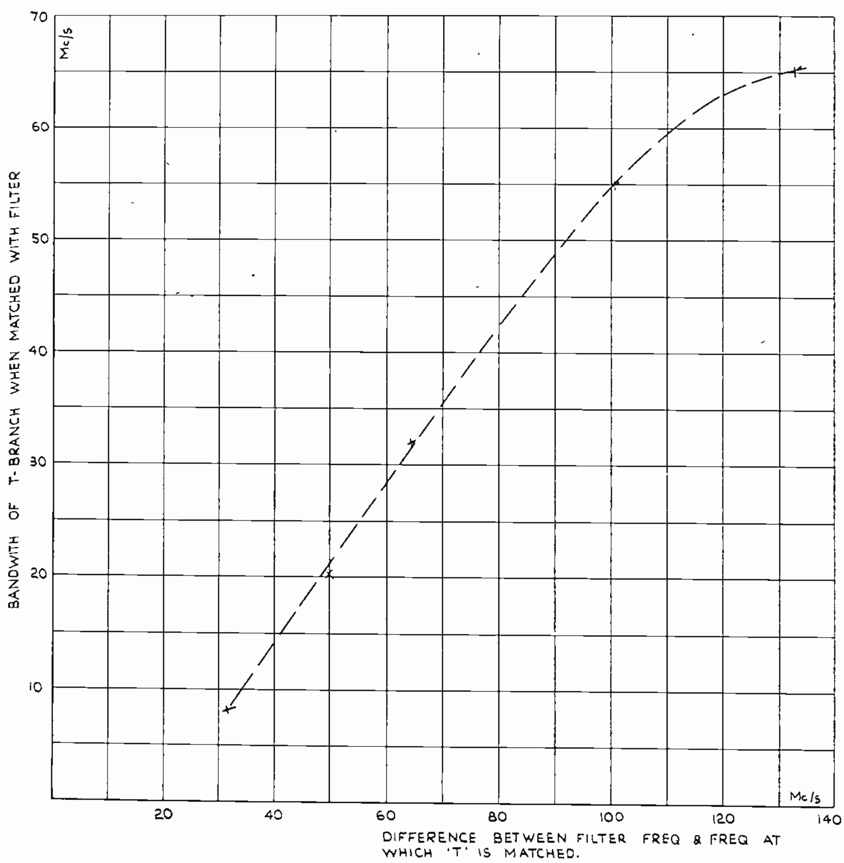

FIG. 32(*d*)

Bandwidth (to VSWR of 1 db) of coaxial-T branch as a function of separation from filter resonant frequency.

The difficulty met with this type of combining filter is the realization of a suitable structure in waveguide or coaxial lines.

In a case where only two radio channels are to be combined in a single aerial feeder, simpler branching networks are desirable. One such network, for use with

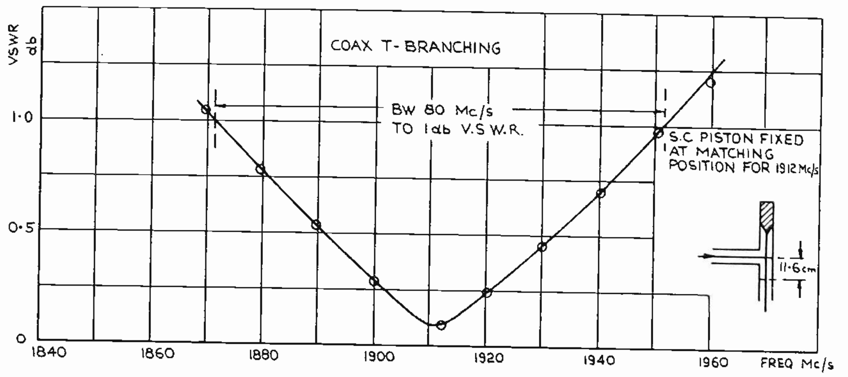

FIG. 32(*a*)
Response of coaxial T junction.

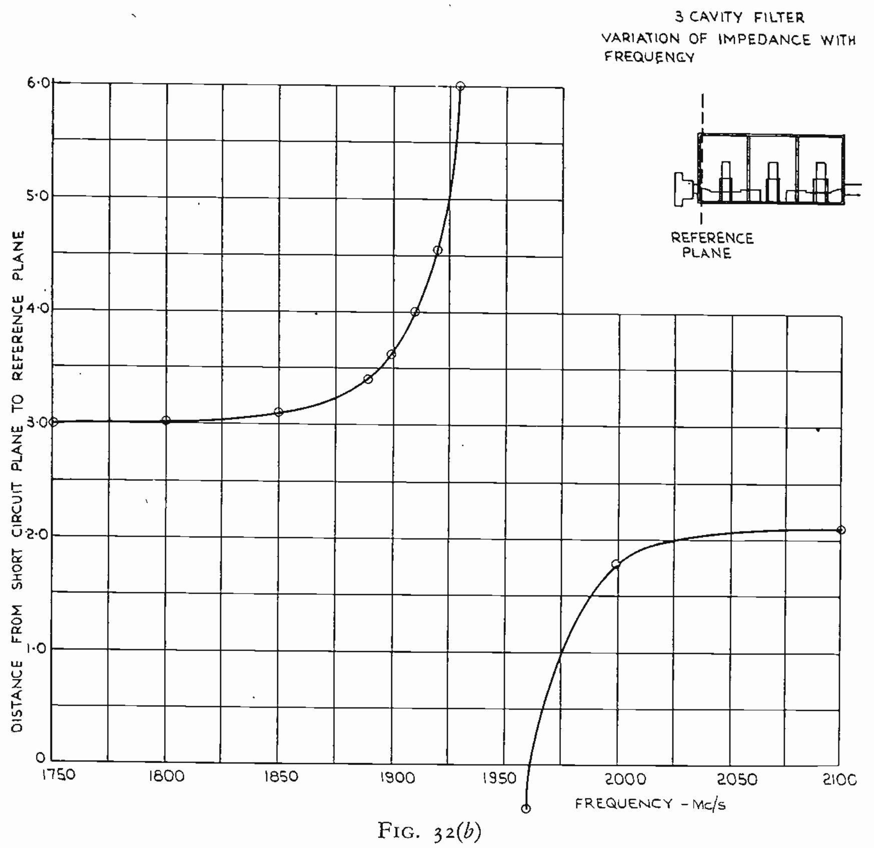

FIG. 32(*b*)
Input impedance of coaxial filter in stop bands in terms of position of virtual short circuit (resonant frequency 1944 *Mc/s).*

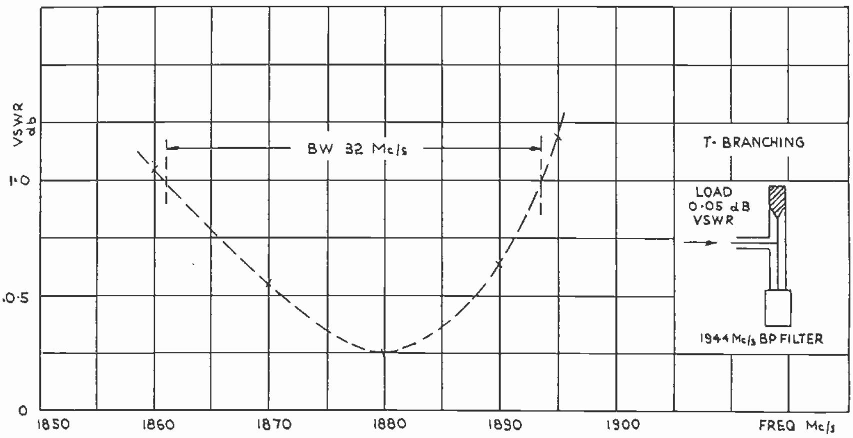

FIG. 32(*c*)

Input impedance of coaxial-T branch with filter at one end. 64 Mc/s away from resonant frequency of filter.

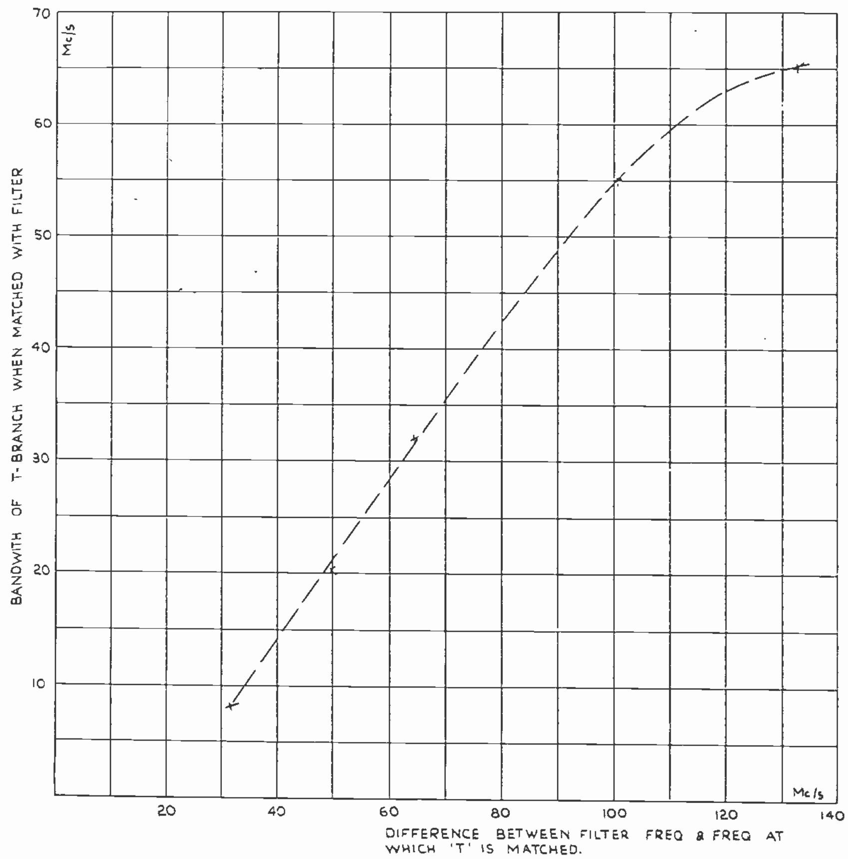

FIG. 32(*d*)

Bandwidth (to VSWR of 1 db) of coaxial-T branch as a function of separation from filter resonant frequency.

coaxial filters, consists of a coaxial T, connected to two filters as shown in Fig. 31. The operation of this type of combining system depends on the simulation with one filter, F_1 for example, of a short circuit at some plane P_1, such that energy applied at A_1 over the pass band of filter F_2 is directed into branch C. Similarly, filter F_2 simulates a short circuit at some plane P_2, so that energy applied at A_1 over the pass band of filter F_1 is directed into branch B.

This simple type of branching filter may be easily designed and constructed. Because, however, of the sensitivity of the coaxial T, regarding the position of the virtual short circuit, it cannot be applied to separate signals having too close a frequency separation, in terms of the required filter bandwidths. Bandwidth curves of the coaxial T under various conditions of use are given in Fig. 32.

Aerial Systems

The frequency plan, discussed on page 70, requires the use of very wide band aerials, capable of operating over bands of at least 400 Mc/s, with a standing wave ratio of not less than about 0·95. If only two aerials are used at each repeater station, then each aerial must be able to handle two rectangular polarizations simultaneously and the combining system at the equipment end of the aerial feeder must be designed to handle up to twelve different frequencies. It is doubtful whether such a combining system can be designed and adjusted to give a sufficiently small loss and standing wave ratio over the required band and it appears more economical to design aerials on the basis of four units per repeater, as indicated in Fig. 21.

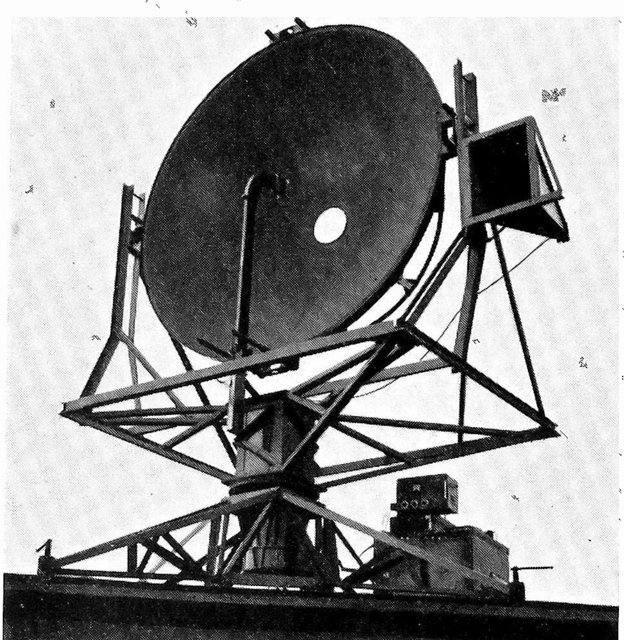

Fig. 33
10-*foot Aperture Parabolic Aerial.*

This eases considerably the design of the combining system, which has to handle up to six channels only and of the aerials, which only need to handle a single polarization and to be well matched over bandwidths of 200 Mc/s (10% at 2000 Mc/s).

A type of aerial which can, with suitable design, give such a performance is the parabolic dish, with a centrally placed horn feed, and vertex matching plate (Fig. 33). The vertex plate matching technique[14, 15, 16] appears capable of reducing reflections into the feed horn, due to the aerial, to an equivalent VSWR of 0·99 over wide bands. The bends and flanges making up the feed waveguide may be made to cause a VSWR of not less than 0·985 over a wide band. This leaves an allowance of 0·975 for the VSWR of the feed horn proper.

Experimental horn designs have shown that such an impedance match is possible to obtain over wide bands. A Smith Chart plot of one such design is shown in Fig. 34.

The design of the feed horn must be a compromise between the impedance matching requirement and that of the desired primary feed pattern. The use of an angle of illumination of 180°, which is desirable in order to minimize the coupling between aerials placed side by side, renders the design of a suitable horn appreciably more difficult. Smaller angles of illumination would simplify the design of the horn at the expense of aerial cross-talk.

The use of an offset feed simplifies both the design of the horn and that of the aerial matching device, again at the expense of crosstalk and reduced back to front ratio.

The material used for the making of aerial dishes is generally aluminium but experimental dishes of glass fibre-resin laminates have also been made. These dishes appear to have good weathering properties, but complete tests in tropical and arctic climates would have to be made before this point can be definitely ascertained. They do, however, lend themselves very well to experimental work, because of the relative ease with which it is possible to produce the required profiles.

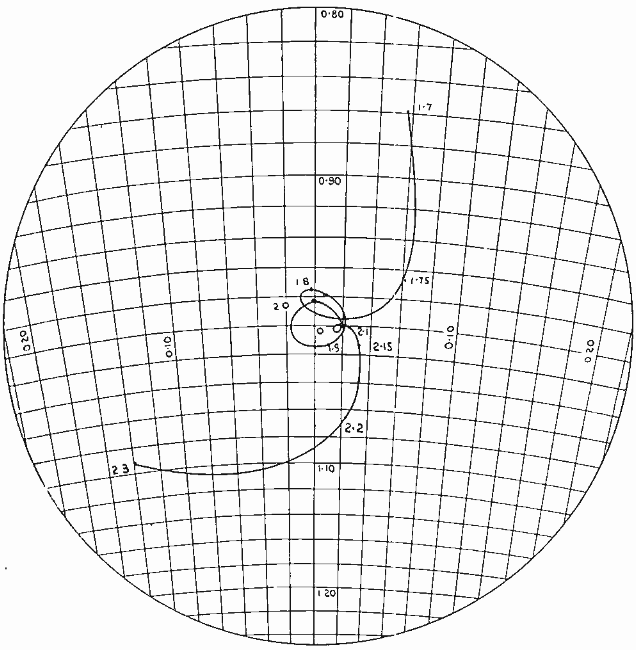

FIG. 34

Typical impedance plot of experimental horn.
(All figures are in 10^9 *cycles.)*

Aerial Feeders

The complete feeder system, from aerial terminals at the top of the mast to the combining filter network, must be as reflectionless as possible, if undesirable cross-talk is not to occur. A limit of 0·95 for the standing wave ratio of the whole of the feeder system is desirable for high capacity systems.

Low loss coaxial feeder such as Telcon HM4AL and ¾ inch F and G Styroflex are quite practicable in the 2000 Mc/s band from the loss point of view. It is doubful, however, whether they are suitable for high capacity telephony systems, because of unavoidable cable irregularities. Furthermore they require the use of very wide-band waveguide to coaxial line transformers at the aerial terminals and at the terminals of the combining system. The latter may not be necessary if an all-coaxial combining system for multiple radio channels can be devised.

All-waveguide feeders appear to present the better alternative. When, however, this is examined in detail it is apparent that a considerable amount of high precision work is required before an all-waveguide feeder can be made to the required degree of perfection. In general it is not possible, in a practical installation, to arrange the connection between the aerial and the equipment to be in a single straight line. Bends, both E and H, are required, and these should be extremely well matched over the whole of the feeder bandwidth. Since waveguide feeder is generally not supplied in lengths longer than about 14 feet it is necessary to use a number of connecting flanges, which themselves introduce appreciable mismatches.

A 150-foot long feeder is likely to contain as many as twenty pairs of flanges and perhaps three or four bends of various types. If the allowable reflection is divided equally between bends and flanges, then the allowable VSWR of each bend should be better than 0·995 over the required band, and that of a pair of flanges better than 0·999 over the same band.

The best type of flange appears to be the simple butt flange, with a thin copper shim, between the two surfaces. Typical standing wave ratio obtained with this type are of the order of 0·997 over the whole of the 2000 Mc/s band.

Conclusion

It is clear, from the above survey, that the design of broad band systems presents a number of difficult problems, particularly with regard to the impedance matching requirements of aerials, feeders, branching systems and some of the interconnecting microwave components. The requirements as regards the modulation and demodulation equipment seem to be equally, if not more, difficult to meet.

Whilst it is not wise to be dogmatic in such matters, it appears that these requirements can be met by further refinements in known techniques, although new advances in the microwave field may well make these refinements pretty well unnecessary.

References

(1) A. G. Clavier and A. C. Gallant. "The Anglo-French microray link, between Lympne and St. Triglevert. *Electrical Communications* 12, 222, 1934.

(2) T. A. Morton and R. M. Ryder. "Design factors of the Bell Telephone Laboratories, 1553 triode." *B.S.T.J.* 29, 496-530, October, 1950.

(3) International Radio Regulations. Chap. 3, Article 5, pp. 54-58.

(4) A. T. Starr and T. H. Walker. "Microwave radio links." *J.I.E.E.*, April, 1952

(5) H. T. Fries. "Microwave Repeater Research." *B.S.T.J.* 27, 183-246, April, 1948.

(6) W. T. Bray. "The travelling-wave valve as a microwave phase modulator and frequency shifter." *J.I.E.E.*, Part II. 99, 57, pp. 15-20, January, 1952.

(7) L. B. Coulson and F. W. H. Robinson. "A Medium Power Travelling Wave Tube for 2000 Mc/s." *Marconi Review* 117, p. 48, 2nd Quarter, 1955.

(8) I. I. Egli. "Radio Relay System Engineering." *Proc. I.R.E.*, January, 1953, pp. 115-124.

(9) A. B. Crawford and W. C. Yakes. "Selective fading of microwaves." *B.S.T.J.* 31, 68-90, January, 1952.

(10) W. W. Mumford. "Maximally flat filters in waveguide." *B.S.T.J.* 27, 684-713, October, 1948.

(11) Milton Dishal. "Alignment and adjustment of synchronously tuned, multiple resonant circuit filters." *P.Proc. I.R.E.* 39, p. 1448, November, 1951.

(12) W. D. Lewis and L. C. Tillotson. "A non-reflecting branching filter for microwaves." *B.S.T.J.* 27, 83-95, January, 1948.

(13) E. L. Norton. "Constant resistance networks with application to filter groups." *B.S.T.J.* 16, 178-193, April, 1937.

(14) Silver. "Microwave Antennas."

(15) British Patent 581457.

(16) British Patent 675245.

MICROWAVE TELEVISION TRANSMISSION SYSTEMS

By W. L. Wright, B.A., A.M.I.E.E.

The performance requirements of Television Radio-relay Links are discussed The dependence of overall video signal-to-noise upon transmitted power, aerial gain, path length, receiver noise factor, frequency deviation and video bandwidth is examined and the conclusion reached that typical requirements may be met using microwave equipment employing travelling-wave amplifiers.

Modulation and demodulation equipment used in broadband television links is reviewed. Amongst the units described, the design of frequency-modulated oscillators with their associated automatic frequency control systems receives special attention.

Introduction

THE need for relaying television signals over fixed links was realized at the time when the first public television service was opened in 1936. Within a year of this date a short coaxial cable television link was in operation connecting the West End of London to Broadcasting House and the Alexandra Palace transmitter.

As an alternative to cable the television radio-relay link was early in the field. Pre-war prototypes were in operation in the London area operating on frequencies of about 60 Mc/s and were used for outside broadcasts.

The phenomenal advances made in microwave radio technique during the war have resulted in the radio link becoming an even more serious rival to the coaxial cable than it had been in previous years. Post-war experimental television link systems were set up by engineers of the British Post Office between London and Castleton, near Cardiff. Subsequently, a television radio-relay link was installed between London and Birmingham,[1] and four years later, in March, 1952, the 4,000 Mc/s microwave link between Manchester and Kirk O'Shotts[2] was put into public service.

The development and installation of long-distance microwave television and multichannel links in America and also on the Continent of Europe took place over the same period.

The present trend in this Country is to develop and extend broadband link facilities by the use of both coaxial cable and radio relay systems.

The Marconi Company has been active in the television and multichannel radio relay field throughout the post-war years and, to-day, their V.H.F. multichannel telephone radio relay systems are in operation in many different countries.

Radio link activities in the Company during the past few years have been directed towards the use of the 2000 and 4000 Mc/s communication bands. Technical problems, associated with the amplification of microwave signals of less than one microwatt up to the level of several watts, over an accurately matched, wide, radio frequency band, have been considerably simplified by the use of travelling wave valves, developed by the English Electric Valve Company to meet the stringent linearity requirements of radio relay links.

These link systems have been designed to carry either television or multichannel telephony over the same microwave equipment. The modulator and demodulator

(" Modem ") equipment is different for the two uses, so that in practice, either the television modem apparatus or the multichannel modem apparatus is connected with the microwave equipment forming one broadband channel.

The radio frequency portions of broadband microwave links have been described in a companion article[3] and will only be further discussed here in relation to the particular requirements associated with the transmission of the spectrum produced when the system is frequency modulated with a television signal.

Requirements of a Television Relay System

A fixed television relay link is normally required to transmit one television signal, of video bandwidth up to 5 Mc/s, in each broadband microwave channel. Along important routes a number of these microwave channels may be required to operate side by side, housed in common buildings, and sharing common aerials. Simultaneous two-way facilities are often required necessitating the duplication of all the equipment in the reverse direction. Proposed adjacent channel carrier spacing in the 2000, 4000 and 6000 Mc/s bands is 32 Mc/s. In a frequency modulated television link the instantaneous frequency of the microwave signal during the transmission of the bottom edge of the synchronizing pulses is required to be held constant, within about ± 200 kc/s of a specified radio frequency. This usually requires the provision of automatic frequency control at the terminal transmitter operated from a measurement of frequency during the line synchronizing interval.

Stand-by channels are required to provide against the failure of a working channel. It is often desirable to provide points at certain repeater stations along a route at which video signals may either be inserted into a broad band channel or else tapped off.

Added to this there are the requirements of interconnection between channels, remote switching, supervisory control and monitoring with automatic switch-over to a stand-by channel when a particular channel fails or becomes too noisy. Reliability requirements are such that, with unattended operation of repeater stations, time lost due to the failure should be a minimum and comparable with that expected of cable relay systems.

This impressive list of requirements relates, of course, to a highly developed and comprehensive system, but it is desirable, when planning even a one-microwave channel radio relay link, to consider possible future extensions to such a channel, and to allow for this in the initial design.

Transmission Performance

When specifying the transmission performance of a link it is understood that the figures refer to a definite path length of p miles, divided into m hops, each of average length p/m miles. Performance figures may conveniently be given assuming free space propagation conditions, the figure for signal-to-noise ratio being sufficiently high so that, after degradation by an estimated fading allowance of β db, the resulting performance will still be acceptable.

In a F.M. system, the repeater gain should not be below $(\alpha+\beta')$ db, where α represents, for a particular hop, the free space path loss plus aerial gains and feeder losses, and β' is a prescribed maximum fading margin for the hop. This will ensure that the repeater output power will remain unchanged during a fade, up to a maximum of β' db. At normal carrier level, during the absence of fading, there will be β' db of compression in the repeater equipment, either in the form of limiting or due to A.G.C. action.

In a television link the following transmission characteristics, measured from a video input point to a video output point, are of importance:—

(i) The attenuation/frequency characteristic.

(ii) The phase/frequency characteristic.

(iii) The amplitude linearity and overload level.

(iv) The signal-to-noise ratio in the video band.

(v) The gain stability of the video signal.

(vi) The waveform response.

In an amplitude linear system (i) and (ii) taken together determine the transient response to a given test pulse, provided that the measurements are made over a frequency range covering the complete spectrum of the pulse. Nevertheless a separate wave-form response specification is normally given, since the test pulses are designed to simulate the actual television signal, and so to form a good practical system test. The choice of a practical form for a test pulse is, however, often a compromise, for, if it possesses a sensibly flat spectrum at the high end of the pass band, and if it is reasonably free from overshoot, it will also contain frequency components above the pass band. In this case those portions of the gain and phase characteristics of the system which lie outside the nominal video band will also influence the shape of the observed output waveform. This may not be a desired feature of the test.

The faithful reproduction of low frequency waveforms (e.g., of a square waveform of 40 milliseconds duration, with line synchronizing signals added) is also dependent upon (i) and (ii) so long as no form of level clamping is introduced at the end of the line period. In many cases, however, gated clamps are employed at transmit and receive ends of the system, and in these systems the low frequency waveform response is largely a measure of the efficiency of the clamps.

Acceptable figures for the transmission characteristics enumerated above are:—

(i) Attenuation/frequency characteristic; within ± 1 db between 20 c/s and 5 Mc/s. The top frequency limit of 5 Mc/s allows pictures of up to 625 lines to be transmitted.

(ii) The linearity of the phase/frequency characteristic between 200 kc/s and 5 Mc/s should be such that the variation of group delay (i.e., the change in the value of the derivative of this curve) does not exceed ± 50 millimicroseconds. Between 20 c/s and 200 kc/s the phase characteristic should be linear enough for proper transmission of the low frequency waveforms (see paragraph (vi).

(iii) Linearity of input/output characteristic. About $\pm$ 5% change in slope is allowable over the range of levels occupied by the complete signal in systems which carry one television signal only per broadband microwave channel. This applies to a signal adjusted for normal deviation level. If for some reason an increase in drive level of up to 3 db should occur, no marked degradation of linearity or other performance factor should occur.

(iv) An acceptable ratio of D.A.P (double amplitude peak) signal amplitude to peak-to-peak noise amplitude is about 50 db, before allowance has been made for fading, over an approximate 180 mile path. The noise is measured in a 0 to 5 Mc/s bandwidth and the D.A.P. signal amplitude is measured between black and white levels.

(v) Gain stability over long periods of the order of $\pm 0{\cdot}25$ db is desirable for the video signal. "Aircraft flutter" should be well below this figure.

(vi) The required transient response for the proper transmission of a 0–5 Mc/s video signal is:—

Build-up time $= 0{\cdot}1\mu s$.

Overshoot on a step waveform of build-up time $= 0{\cdot}12\mu s$ should not exceed about $\pm 4\%$ of the undistorted step amplitude.

The low frequency response will be adequate if the slope of the top and bottom of the received 25 c/s square waveform does not exceed about $\pm 3\%$.

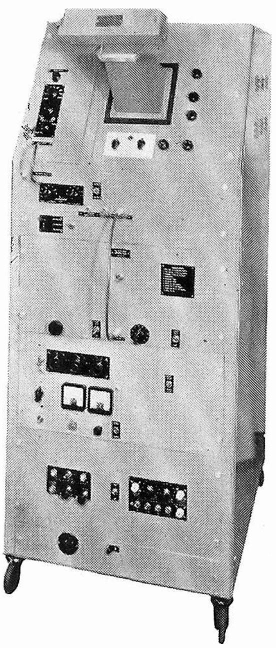

FIG. 1

Typical apparatus for the generation and display of the various test waveforms required for measuring the video transmission characteristics of a link system is illustrated in Fig. 1.

Design Considerations for a Television Radio Relay Link

The type of link described below attempts in its design to meet the requirements detailed in the preceding section. No description is here given of the considerable amount of developmental engineering work necessary to convert an experimental prototype link, which satisfies the initial transmission tests, into a fully engineered multi-broadband-channel system, complete with supervisory control, automatic changeover to standby during failure, remote switching and monitoring, and fitted with the accurately matched filter branching networks necessary for common aerial working. Only points of special interest in the electrical design of a television link, with special reference to the modulation and demodulation equipment will be given.

The initial consideration involves the calculation of transmitter power and aerial gain necessary in order to obtain the required signal-to-noise ratio over a television system, p miles long, between demodulation points.

It is of interest to consider the case where the link length is equal to that between demodulation points on the 2500 km C.C.I.F. coaxial cable reference circuit[4]. This distance is 173 miles approximately. For the present calculation it will be assumed that the average length per hop is 25 miles making the number of hops equal to 7.

For a good quality television circuit the required S/N ratio is about 50 db for the seven hops, before making allowance for fading.

Let this equal 20 log N where

$$N = \frac{\text{D.A.P. video signal amplitude}}{\text{D.A.P. noise amplitude in the video band.}}$$

If 20 log N_1 is the S/N ratio for one hop $\left(\frac{N_1}{N}\right)^2 = 7.$

Therefore
$$\begin{aligned} 20 \log N_1 &= 20 \log N + 8{\cdot}45 \\ &= 50 + 8{\cdot}45 \\ &= 58{\cdot}45 \text{ db} \end{aligned}$$

or $N_1 = 837.$

It is required to find the power output and aerial gain which would give this S/N ratio under free space propagation conditions.

(a) Path attenuation

Let the radio frequency, $f = 2000$ Mc/s, and let feeder losses, $\gamma = 3{\cdot}4$ db total, transmit and receive. Hop distance, $d = 25$ miles. Let α db = the free space path attenuation between isotropic aerials. Then fundamentally

$$\alpha = 10 \log \left(\frac{4\pi d}{\lambda}\right)^2 \text{ db.}$$

where λ is the wavelength.

If d is expressed in miles and f is the frequency in Mc/s, this expression reduces to

$$\alpha = 20 \log (67{\cdot}4\, df)$$

$$\text{or} \quad \alpha = 36{\cdot}6 + 20 \log (df) \text{ db.}$$

If we add to this the feeder losses, γ, then path and feeder attenuation

$$(\alpha + \gamma) = 40 + 20 \log (df) \text{ db.}$$

In this case $(\alpha + \gamma) = 40 + 20 \log (25 \times 2000)$ db.

$$= 134 \text{ db.}$$

(b) Relation between video S/N and Signal Power

In a F.M. system the S/N power may be expressed by the following equation[6]:—

$$\frac{\text{R.M.S. Signal Power}}{\text{R.M.S. Noise Power}} = \frac{3}{2} \cdot \frac{W_s}{FKTB} \left(\frac{\delta f}{B}\right)^2 \qquad (1)$$

where

W_s = received signal power, Watts.

F = noise figure of receiver.

$KT = 4{\cdot}2 \times 10^{-15}$ Watts per Mc/s bandwidth.

B = width of baseband from zero to top modulation frequency.

$\pm\delta f$ = maximum frequency deviation in Mc/s.

From (1) it may be shown that

$$N_1 \equiv \frac{\text{D.A.P. signal amplitude}}{\text{D.A.P. noise amplitude}} = \frac{\sqrt{3}}{8} \frac{\triangle f}{B} \left(\frac{W_s}{FKTB}\right)^{\frac{1}{2}} \qquad (2)$$

where $\triangle f$ = Double Amplitude Peak (Peak to Peak) signal deviation = $2\delta f$. The crest factor for noise has been taken to equal 4, the components being of random relative phase and following a normal-error law distribution of amplitude as a function of time [5, 6].

If the total deviation of television picture plus synchronizing pulses = 6 Mc/s, the picture information will have a deviation = $\cdot 7 \times 6 = 4{\cdot}2$ Mc/s from black to white levels.

Let $F = 10$, $\quad B = 5$ Mc/s.

Substituting these values into (2)

$$N_1 = \frac{\sqrt{3}}{8} \times \frac{4{\cdot}2}{5} \left(\frac{W_s}{10 \times 4{\cdot}2 \times 10^{-15} \times 5}\right)^{\frac{1}{2}}$$

Therefore

$$N_1^2 = (1{\cdot}575 \times 10^{11})\ W_s$$

or

$$10 \log W_s = (20 \log N_1 - 112) \text{ db W} \tag{3}$$

Equation (3) relates Watts into the receiver with S/N for this system.

The required value for $N_1 = 58{\cdot}45$ db, given above, therefore

$$10 \log W_s = -53{\cdot}55 \text{ db } W, \quad \text{or} \quad W_s = 5\mu \text{ W approx.}$$

Since the path attenuation + feeder loss = 134 db, transmitter power plus aerial gains should together equal (134—53·55 db W) to give a received signal of 5μ W i.e. required aerial gains and transmitter power

$$= 80{\cdot}5 \text{ db W}$$

for a S/N of 50 db over 7 hops each 25 miles long.

This figure could be obtained by using paraboloidal aerials 10 ft. in diameter with area efficiency = 0·6, and employing an output power of 20 watts. These aerials would each have a gain = 33·8 db at 2000 Mc/s, since

Gain, $G = \dfrac{4\pi A}{\lambda^2}$ where A is the effective area, or

$$10 \log G = 13{\cdot}8 + 20 \log (rF), \text{ for } 0{\cdot}6 \text{ efficiency} \tag{4}$$

where r = radius of the parabolic mirror in feet, and F = frequency in Kilo Mc/s. When $r = 5$ and $F = 2$,

$$10 \log G = 13{\cdot}8 + 20 \log (5 \times 2)$$
$$= 33{\cdot}8 \text{ db.}$$

Thus aerial gains (10 ft.)+transmitter power (20 Watts)

$$= (33{\cdot}8 \times 2) + 13 \text{ db W}$$
$$= 80{\cdot}6 \text{ db W as required.}$$

(*c*) *Number of Hops*

It is of interest to examine the effect of a change in the number of hops on signal-to-noise ratio when the total distance from end to end of the circuit is kept constant. If the S/N ratio $= x_1$ db when p miles are traversed in one hop, then the new S/N ratio, x_m db which obtains when p miles are covered in m hops, each p/m miles long, is given by

$$x_m = x_1 + 10 \log m. \tag{5}$$

In the above example $x_7 = 50$ db, therefore

$$x_1 = 50 - 10 \log 7 = 41{\cdot}55 \text{ db}$$

thus for a 5 hop system each hop of which = 35 miles:

$$x_5 = 41{\cdot}55 + 10 \log 5 = 48{\cdot}54 \text{ db.}$$

This represents a loss of only 1·46 db in S/N ratio with reference to that for the 7 hop system; the change would therefore prove economical, provided that 35 mile optical paths were possible, without the need for excessively high aerial towers.

(*d*) *Relation between Radio Frequency and S/N.*

Let the attenuation between transmitter output and receiver input be λ db.

Summing the expressions derived above for path attenuation and aerial gain we obtain:

$$\lambda \equiv \alpha - 10 \log (G_T \,.\, G_R) + \gamma$$
$$= (69 + 20 \log d + \gamma) - 20 \log (r_T \,.\, r_R \,.\, F) \qquad (7)$$

where F is in Kilo Mc/s, d is in miles, r_T and r_R are respectively the transmit and receive mirror radii in ft., and γ db = feeder loss. An aerial efficiency = 0·6 has been included in this expression. Other things being equal, we see that

$$\lambda_2 - \lambda_4 = 20 \log \frac{4}{2} = 6 \text{ db}$$

where λ_2 and λ_4 represent the values of λ at 2000 Mc/s and 4000 Mc/s respectively.

Thus for the same aerial diameters, path distance and feeder losses, equivalent S/N performances would be expected when working with 20 watts transmitter power at 2000 Mc/s or with 5 watts at 4000 Mc/s.

MODULATION AND DEMODULATION

(a) Direct Modulation at Microwaves

The simplest and most direct method of producing a frequency modulated microwave signal for television purposes having the required deviation of about ±3 Mc/s, is to employ a velocity modulated self-oscillator tube, with the video modulating signal superimposed upon the direct potential applied to its frequency-control electrode. The reflex klystron offers a distinct advantage over some other types of velocity modulated tube in that its oscillation frequency may readily be controlled by means of the potential applied to the reflector electrode, which itself draws no current and presents only a small capacitative load. A klystron valve type CV 2161 has been successfully used as a frequency modulated oscillator for T.V. purposes. The modulation sensitivity is such as to give a frequency change of between 0·5 and 1·5 Mc/s for one volt change in reflector voltage. It is well known that the slope and linearity of the reflector volts/frequency characteristic is a function of various parameters within the control of the user. These are briefly noted below:—

(i) The nature of the complex impedance presented to the valve by the output coupling and load. This helps to determine the working Q of the cavity, together with its phase/frequency and amplitude/frequency characteristics.

(ii) The position of working along the particular reflector-volts/cavity-tuning curve for the valve, drawn for a constant desired centre frequency of operation. The choice of operating point on this curve also affects the power output and the level of spurious amplitude modulation.

(iii) The klystron operating mode.

(iv) The cathode-to-cavity volts affect the modulation characteristic but to a lesser degree.

By means of a careful adjustment of cavity tuning, reflector volts, output load and cavity volts it is possible to obtain a modulation characteristic which is linear to within ±1% over a frequency sweep of about 8 Mc/s. Greater linearity than this is possible over smaller deviations (e.g., a 70 db harmonic margin has been obtained at ±0·5 Mc/s deviation) but any drift in tuning or in operating voltages will usually result in a marked falling off from the initial linearity of the characteristic.

The valve type CV 2161 will deliver about 30 mW at 2000 Mc/s into a 50 ohm resistive output load, when the coupling takes the form of a simple loop of correct size, using no tuning stubs or other form of reactive adjustment. A flat cylindrical type of tuned cavity may be used with the valve, tuned over the 1750 to 2300 Mc/s band by means of plungers screwed into the circular wall.

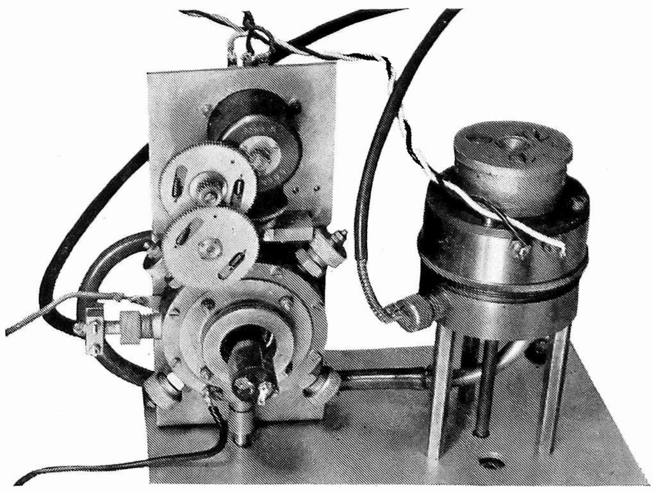

FIG. 1(*a*)

Fig. 1*a* shows a photograph of an experimental arrangement of the CV 2161 klystron modulator unit consisting of the valve and tuned cavity, fitted with a motor-driven vane for automatic frequency control and an invar cavity which serves as a frequency reference.

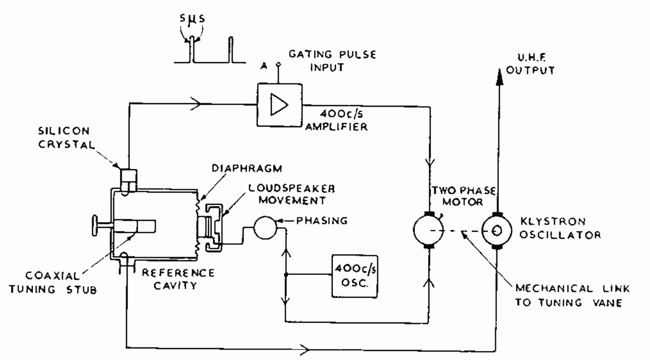

FIG. 2

Schematic Diagram of Microwave A.F.C. Unit. (Jenks System.)

An important part of any frequency modulated oscillator drive used in a television relay system is the automatic frequency control unit. The principle of the A.F.C. system shown in Fig. 1*a* has been described by F. A. Jenks[7] and was successfully used by engineers of the Post Office Radio Branch in the London-Wenvoe experimental television link. The unit illustrated in Fig. 1*a* employs motor-driven tuning in conjunction with a single-cavity discriminator. A schematic diagram is shown in Fig. 2. From this it will be seen that the reference cavity is supplied with microwave energy fed from the klystron oscillator. A silicon crystal rectifier, loosely coupled to the cavity, produces a direct current, the amplitude of which is greatest when the klystron frequency and the resonant frequency of the cavity coincide.

The cavity resonant frequency is swept over about ± 200 kc/s by means of a moving-coil loud-speaker unit which vibrates a diaphragm, fitted at one end of the cavity. The moving coil unit is driven from a 400 c/s oscillator which also supplies one winding of a two-phase motor. The other winding of this motor is energized through a narrow-band 400 c/s amplifier from the signal derived from the crystal diode associated with the reference cavity. The 400 c/s component of the crystal diode output is zero when the klystron frequency coincides with the mean frequency of the reference cavity. When the klystron frequency drifts from this mean frequency, a 400 c/s error signal is produced, the phase of which differs by 180° according to whether the klystron frequency is higher or lower than the mean reference frequency. This 180° difference of phase of the 400 c/s error signal determines the direction of rotation of the motor-driven tuning vane, within the klystron cavity, and so produces a properly sensed tuning correction.

The Jenks system possesses the following advantages over many other types of microwave A.F.C. system:—

(1) The single cavity arrangement, using only one crystal diode, avoids errors associated with two crystal systems in which the discriminator centre-frequency is determined by balancing the output of one crystal against another. In such systems a centre frequency shift occurs when, for any reason, the relative sensitivity of the crystals changes, either by ageing or through a change in R.F. level, since the crystal characteristics cannot be identical in practice.

(2) In the single-crystal system the gain and phase stability of the amplifier circuit is not very important.

(3) A motor drive system effects a correction which remains effective after removal of the error signal.

(4) The absence of a frequency correcting voltage on the modulator frequency-determining electrode reduces the possibility of change of working point on the modulator characteristic. (This is, of course, only true if, as is usual, the chief cause of the original frequency error is drift of the resonant frequency of the klystron cavity.)

The chief disadvantages associated with the Jenks system are:—

(1) Change in the dynamic centre position of the diaphragm may take place, either over a period of time as mechanical stresses settle down, or as a result of a change in amplitude of the 400 c/s e.m.f. applied to loud-speaker movement.

(2) Very loose coupling is required between klystron and reference cavities in order to avoid undue pulling between the two. This necessitates the use of a high gain amplifier between crystal and motor. The amplifier has to have a narrow bandwidth to minimize thermal noise, hence a good stability figure for the 400 c/s oscillator is required.

A.F.C. Gating and Line Clamping

In a television radio relay system it is normally required to stabilize the frequency of the microwave signal within about ± 200 kc/s at the instant of transmission of the tip of the line synchronizing pulses. One way of accomplishing this is to gate the A.F.C. system so that its action is cut off at all times except during this period. In Fig. 2 this operation is seen to be performed by means of a gating pulse applied to the 400 c/s amplifier at point A. This pulse reaches an amplitude of $+25$ volts about 1 μs after the initiation of the line synchronizing pulse. Its duration is 5 μs.

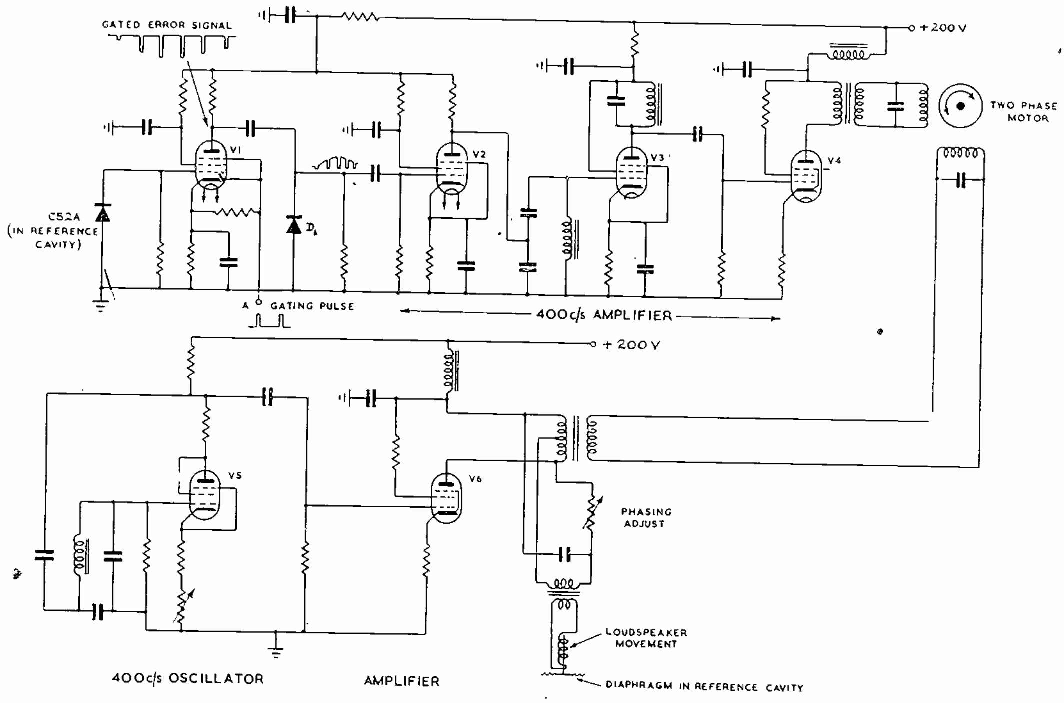

FIG. 3

A.F.C. Circuits. 400 *c/s Oscillator and Amplifier.*

The circuit of the 400 c/s A.F.C. oscillator and amplifier is shown in Fig. 3 in which the gating pulse is seen to be applied to the suppressor grid of the first amplifier valve V_1 via an R-C network. Due to rectifying action at this point the suppressor grid develops a negative potential which keeps the valve cut off at all times except during the 5 μs when the gate pulse swings positively. In the anode circuit of V_1 negative pulses appear of line repetition frequency, whose amplitude is modulated by any 400 c/s error signal present. Picture components appearing at the grid of V_1 during video modulation of the klystron are not present at V_1 anode, due to the gating action. The negative pulses at the anode of V_1 are d.c. clamped at their negative tips by the crystal diode D_1. Any 400 c/s component present is thereby transposed to the top edge of the pulses; action equivalent to that of pulse lengthening is thus obtained. (See waveforms shown at the grid of V_2, Fig. 3.) The three narrow-band 400 c/s tuned-amplifier stages which follow, filter off all components except the wanted 400 c/s error signal.

In addition to keying the error signal amplifier it is necessary to clamp the klystron reflector potential during the period of the tip of the line synchronizing pulses. Unless this is done, variations in picture content of the video signal will cause changes in the instantaneous values of the voltage applied to the klystron reflector during successive tip-of-synchronizing-pulse periods. The A.F.C. system would try to correct this error, but if its time constant were longer than that of those networks

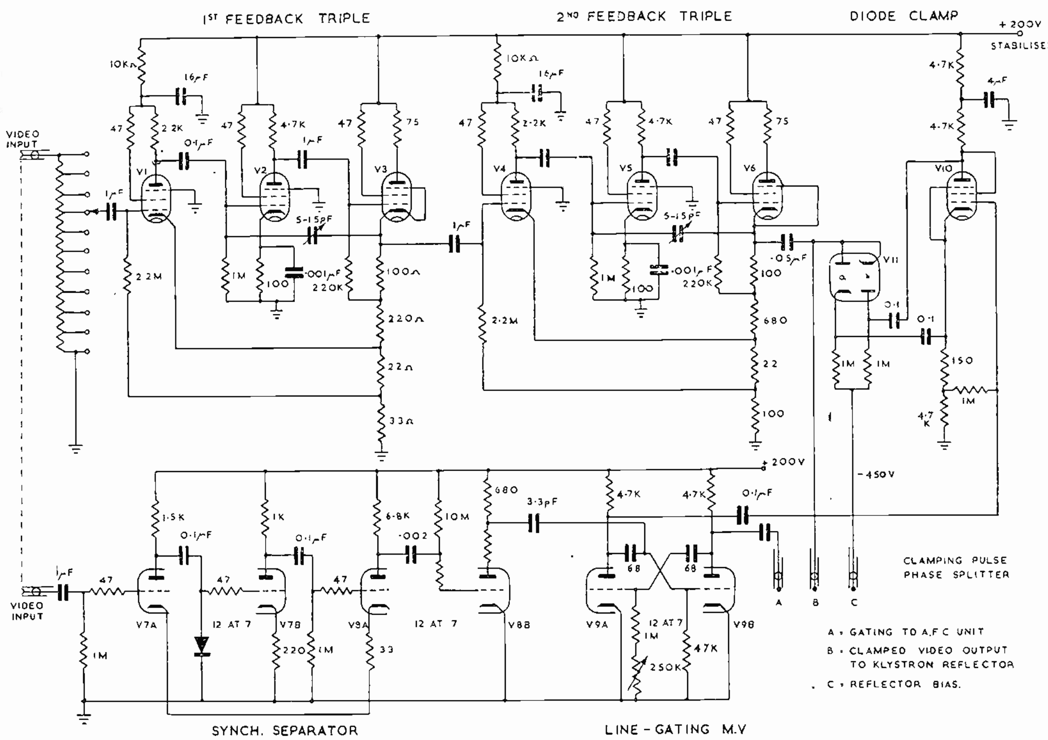

FIG. 4

Transmit Terminal: Video Amplifier, Synch. Separator, Gating-Pulse Generator and Line Clamp.

preceding the klystron (up to the last point at which the signal was clamped), the correction would only be partially effective. In any case, during a moving scene, the motor would be continuously correcting for changes in picture content, which is undesirable.

The clamping of the klystron reflector is performed in the circuits associated with the video amplifier. These are described below. The degree of frequency stabilization obtained using the experimental equipment shown in Figs. 1*a*, 2 and 3 was:—

(*a*) Long-term, over many months: ± 170 kc/s.

(*b*) Short-term, 1 day: ± 50 kc/s.

Transmit Video Amplifier

The unit remaining to be described in the transmitting terminal is the video amplifier. A theoretical circuit of this is shown in Fig. 4 together with the synch-separator and gating-pulse generator stages required for clamping the klystron reflector and keying the A.F.C. unit.

The input level is nominally 1 volt D.A.P. for a full white signal; the required output is of the order of 12 volts D.A.P. working into the capacitative load of the klystron of about 4 pF.

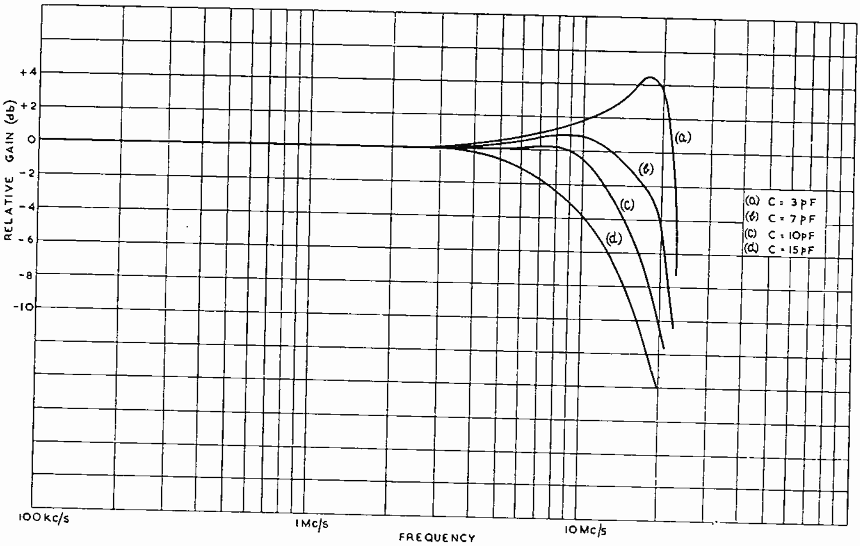

FIG. 5

Response measured over first feedback-triple of Fig. 4. *Curves* (*a*) *to* (*d*) *show effect on frequency response of increasing the value of the feedback capacitor connected between the cathode of* V_3 *and the grid of* V_2 *in the circuit of* Fig. 4.

The video amplifier has a frequency response which is flat within $\pm 0{\cdot}5$ db from 1 c/s to 10 Mc/s when measured at an output signal level $= -10$ db, reference the normal working level of 12 volts D.A.P. In Fig. 4 two three-stage loop-feedback amplifiers are shown coupled in tandem. Double loop-feedback is employed in each feedback triple. The return path of each main loop is through a potentiometer network coupling the cathodes of the first and last stages (i.e. V_1 to V_3 cathode and V_4 to V_6 cathode). The auxiliary loop in each triple takes the form of a small coupling capacitance of value between 5 and 15 pF connected between the cathode of the last feedback stage and the grid of the preceding valve. This small capacitative feedback serves to reduce the phase-lag at high frequencies over the second and third stages. The effect of this second loop is to provide greater overall stability [8, 9], enabling 30 db of stable feedback to be obtained at low frequencies over the main loop. (The fall-off in forward gain at high frequencies reduces the loop-feedback to about 20 db at 5 Mc/s.)

The influence of the auxiliary feedback capacitance upon frequency response is illustrated in the curves of Fig. 5. With zero capacitance, oscillation would occur at about 18 Mc/s, since in this frequency region the forward path phase-lag is 180°. As the value of the auxiliary feedback capacitance is increased, oscillation ceases, but there is at first a rise in response at this frequency. Further increase in capacitance causes the rise and the characteristic to flatten out and finally to become a depression as shown in Fig. 5. Thus the shape of the response characteristic can be controlled at the high frequency end by varying the value of the feedback capacitor.

The amplifier gain/frequency characteristic falls off gradually above 10 Mc/s so that the phase-frequency characteristic is very linear over the pass-band between 50 c/s and 5 Mc/s. As noted above, the amplifier is required to step up from 1 volt D.A.P. to 12 volts D.A.P., i.e., a gain of 22 db is required, neglecting impedance differences between input and output. The amplifier gain is 34 db, so that at normal video signal level, the input attenuator is set at 12 db.

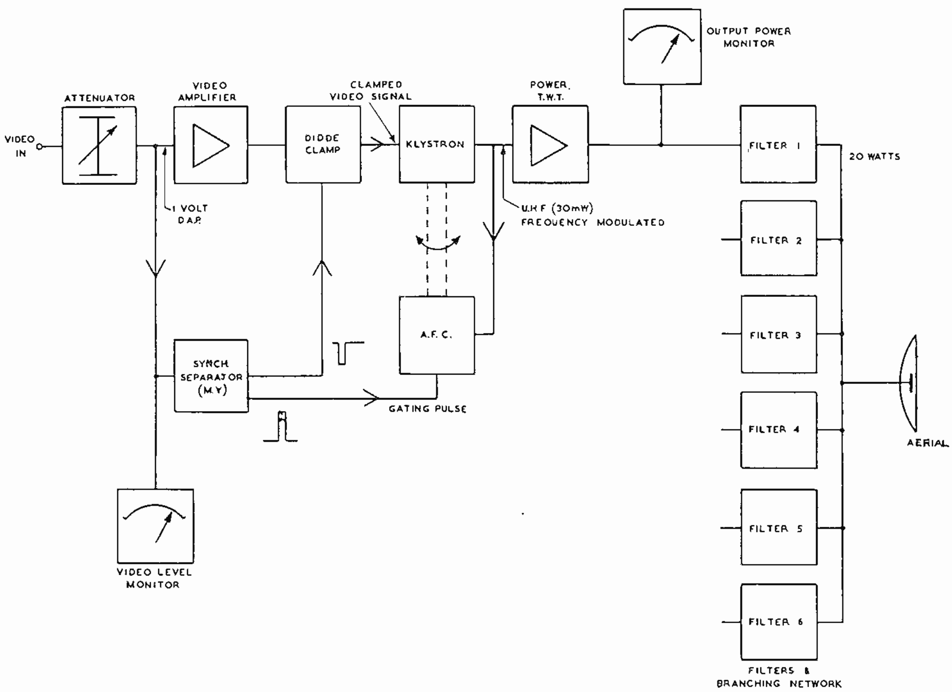

FIG. 6

Arrangement of Units: Klystron T.V. Transmit Terminal.

The low-frequency response of the amplifier is maintained flat down to 1 c/s in order to ease the work required of the gated clamp. In Fig. 4 valve stages V_{7a}, V_{7b}, V_{8a} and V_{8b} perform the function of synch-separation. V_9 is a self-running multivibrator generating pulses 5 μs wide at line repetition frequency. It is locked from the video signal line-synch pulses by connecting the grid of V_{9a} and the anode of V_{8b} by means of a differentiating R-C network. V_{10} is a phase-splitting stage and supplies positive and negative pulses of about 30 volts peak to key the line clamp diodes V_{11a} and V_{11b}.

The reflector electrode of the transmitter klystron is connected directly to the diode clamp; the negative potential applied to the latter is adjusted to equal the particular value desired to be present during the tip of the synchronizing pulses.

Signal components are impressed on the klystron reflector through a 0·05 μF capacitor which connects to the output of the video amplifier. Design factors for keyed clamping circuits have been given in recent literature[10, 11].

Television Relay Transmitting Terminal

The inter-connection between the klystron modulated-oscillator and the remainder of the transmitting terminal equipment is shown schematically in Fig. 6. The klystron is coupled directly into the matched 50 ohm input to the power travelling wave tube. This tube amplifies the signal from 30 mW to 20 watts. At this high level of drive the tube (E.E.V. Co. type N.1001) is working under

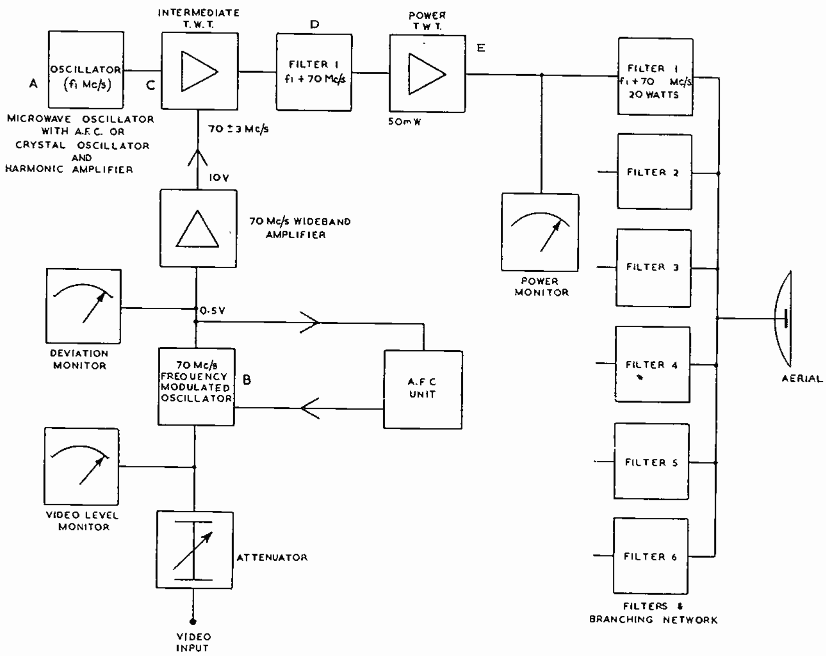

FIG. 7

T.V. Link Transmit Terminal with F.M.O. at Intermediate Frequency.

compression (i.e., the working point is above the bend on the input/output characteristic) so that amplitude limiting of the output signal takes place.

(b) Modulation at Intermediate Frequency

An alternative method of generating a frequency-modulated microwave signal is to perform the frequency modulation at an intermediate frequency (e.g., at 70 Mc/s) and then to frequency-change this signal to the required microwave band. The frequency changing operation may conveniently be performed as indicated in Fig. 7. A stable unmodulated microwave source (A), of frequency f_1 is connected to the input of a medium power T.W.T. (C). The electron beam passing through this tube is at the same time velocity modulated by the 70 Mc/s signal originating in the frequency modulated oscillator unit (B), this signal being impressed between helix and the gun electrodes of the T.W.T. (C). The resulting phase modulation of the microwave carrier f_1 produces sidebands ($f_1 \pm 70$ Mc/s), ($f_1 \pm 140$ Mc/s), --- ($f_1 \pm n \times 70$ Mc/s), where n is a whole number. Each of these sidebands will follow the instantaneous frequency of the 70 Mc/s signal which caused them. A microwave filter (D) connected at the output of tube (C) is arranged to have its passband centered on any desired sideband ($f_1 + n_1 \times 70$ Mc/s) where n_1 is positive or negative. Usually n_1 is either $+1$ or -1 when the selected centre frequency passed by the filter is either $f_1 + 70$ Mc/s or $f_1 - 70$ Mc/s. By choosing the appropriate amplitude of the 70 Mc/s signal applied to the helix of the T.W.T., an optimum level of the required sideband may be realized, this being a function of the operative modulation index according to well known F.M. formulæ[12]. In Fig. 7 the output signal from the filter (D) is shown feeding into a power-amplifier T.W.T. (E). This supplies a signal of frequency (f_1+70 Mc/s $\pm \triangle f$) via a branching network to the aerial feeder, where $\pm \triangle f$ represents the frequency swing produced by the video signal in the modulated oscillator (B).

Advantages

Advantages of the I.F. Modulation System are:—

(i) Increased system flexibility. A 70 Mc/s F.M.O. unit may be connected into a broadband link system at any intermediate frequency input point. This facilitates the injection of video programme material at any repeater station at which such a point is available. The I.F. modulation system also enables any 70 Mc/s Frequency Modulated Oscillator unit to be connected into any microwave transmitting terminal rack situated at a given station, irrespective of the output radio frequency of the equipment.

(ii) Modulation at I.F. avoids the use of microwave modulated-oscillator tubes with their associated tuned cavities and stable high voltage supplies. The receiving-type valves used in the I.F. modulator, have a long life and are cheap to replace.

(iii) Automatic frequency control requirements are less stringent for the modulated-oscillator working at I.F. A stability of ± 200 kc/s at 2000 Mc/s, required of a microwave Frequency Modulated Oscillator necessitates an A.F.C. accuracy within ± 1 point in 10^4, whereas in the 70 Mc/s I.F. system the required A.F.C. accuracy for a similar microwave stability is only ± 1 part in 350.

Disadvantages

Disadvantages of the I.F. Modulation System are:—

(i) The I.F. modulated-oscillator operates at a greater fractional deviation than that of the microwave oscillator (e.g., $\triangle f/f_0 = \pm 3/70 = \pm 4{\cdot}3\%$ in the case of the I.F. oscillator, whereas $\triangle f/f_0 = \pm 3/2{,}000 = \pm{\cdot}15\%$ for a 2000 Mc/s Frequency Modulated Oscillator (the deviation in each case being ± 3 Mc/s). The I.F. oscillator thus has to work with the same linearity as the microwave oscillator but over a considerably greater ratio of deviation to centre frequency.

(ii) The ratio of top modulation frequency (5 or 6 Mc/s) to modulator centre frequency (70 Mc/s) is much smaller in the case of the I.F. oscillator, rendering the unit more liable to interaction between I.F. and video frequencies. A difficulty is encountered in maintaining adequate by-passing of the 70 Mc/s signal at the video input circuit of the modulator. If filtering is inadequate instability of centre-frequency results.

(iii) More complex circuitry is involved in the I.F. system. A comparison between Fig. 6 and Fig. 7 shows that the I.F. system uses additional equipment comprising a T.W.T. stage, a wideband 70 Mc/s I.F. amplifier unit and a frequency stabilized microwave source.

The final choice between the two systems depends largely upon the relative importance with which the user views the conflicting factors of cost and flexibility of use. Technically, the success of the I.F. system is largely dependent upon the development of a 70 Mc/s Frequency Modulated Oscillator possessing the required large fractional deviation with sufficient linearity and flatness of frequency response over the video band.

Modulators which have been found to meet these requirements are discussed in the following section.

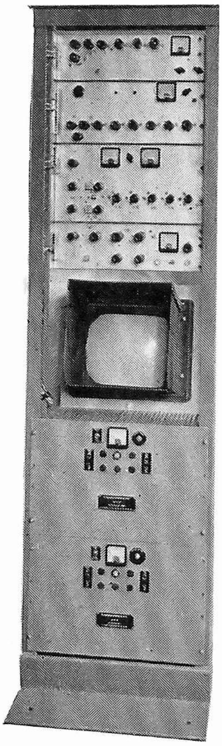

FIG. 7(*a*)

An illustration of a television link transmit and receive modem employing the I.F. system of modulation is given in Fig. 7a.

70 *Mc/s Frequency Modulated Oscillators*

Conventional reactance modulators produce insufficient deviation for T.V. purposes when used in conjunction with oscillators centred on 70 Mc/s. Incidental stray capacitanes, including those within the valve, have such low reactance values at this frequency that it is necessary to compensate by introducing one or more inductances into the circuit. This results in circuits in which the oscillatory loop has a phase/frequency characteristic with an excessively high slope at 70 Mc/s. In other words, the group delay around the positive feedback loop at 70 Mc/s is too great to permit adequate electronic deviation.

Frequency-Modulated Oscillators, employing positive loop feedback over several stages, in which the phase of the feedback e.m.f. is electronically shifted by an amount proportional to the instantaneous applied modulating voltage, produce much larger frequency shifts, since it is possible to design such circuits to have a comparatively short group delay over the I.F. band. The governing factors in the operation of variable phase-shift Frequency-Modulated Oscillators have been given in a paper by O. E. De Lange[13] whilst practical circuits employing cathode follower phase-shifting stages have been described by M. Ames[14] and A. Cormack[15].

Fig. 8 illustrates the principle of a variable phase-shift Frequency-Modulated Oscillator. M is a phase modulator, A is an amplifier, L is a limiter. Let ω be the instantaneous angular frequency of self-oscillation.

v be the instantaneous applied modulating voltage.

ϕ_m be the instantaneous phase measured across the phase-modulator, M, Fig. 8.

ϕ_a be the total instantaneous phase measured across the amplifier and limiter circuits at angular frequency ω.

Amplifier A is of sufficient gain to produce unity gain around the loop, taking into account the action of the limiter L.

The phase of the oscillation must equal zero around the loop so that $\phi_m = -\phi_a$ at all frequencies.

It is required to evaluate $d\omega/dv$.

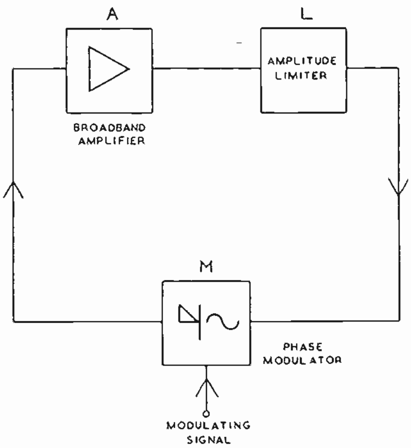

FIG. 8

Principle of the Variable Phase-Shift Frequency-Modulated Oscillator.

We have

$$\frac{d\omega}{dv} = \frac{d\omega}{d\phi} \cdot \frac{d\phi}{dv}$$

where $\phi \equiv \phi_m = -\phi_a$

or

$$\frac{d\omega}{dv} = \frac{d\phi_m}{dv} \div \frac{-d\phi_a}{d\omega}$$

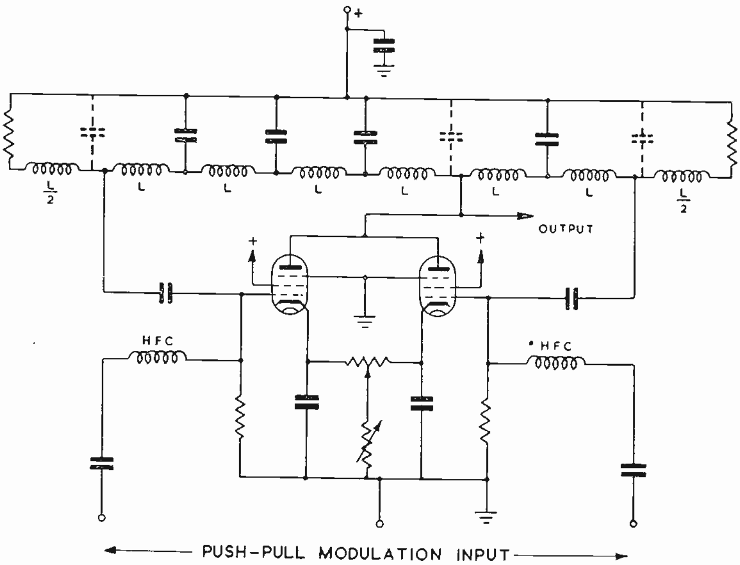

FIG. 9

Delay-line Frequency-Modulated Oscillator.

i.e., frequency deviation per unit modulating volts = phase deviation per unit volts applied to the phase modulator $\div$ the group delay around the feedback loop.

It is thus important in designing a wide-range frequency modulated oscillator

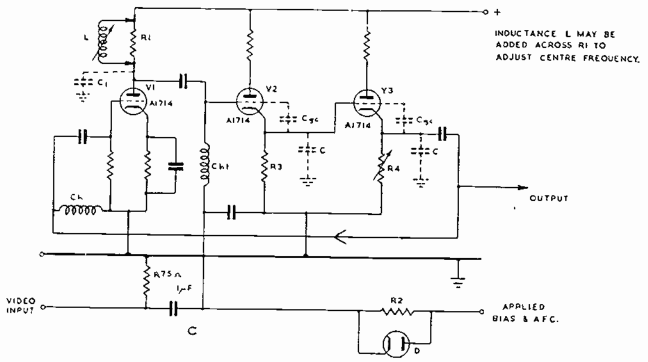

FIG. 10

70 Mc/s Frequency-Modulated Oscillator employing cathode-follower phase-shifting stages.

to try to obtain a large loop phase-shift with the applied modulating voltage and to keep the group delay around the loop as short as possible.

These factors have been taken into account in the Frequency-Modulated Oscillator shown in outline in Fig. 9.

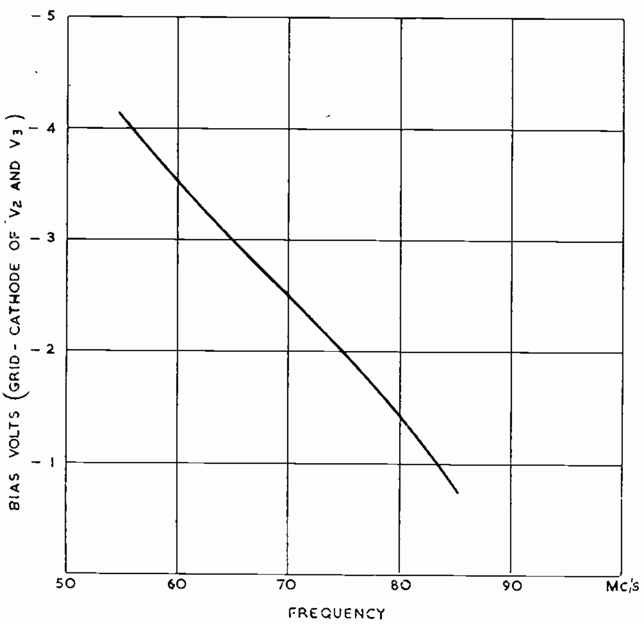

FIG. 11

Typical Frequency-Modulation Characteristic of circuit of Fig. 10.

A delay line is used in conjunction with a pair of valves connected as a push-pull modulator. The anodes are commoned and the grids are phased one at $-120°$ and the other at $-240°$ with reference to the e.m.f. at the anode at centre frequency, by including line lengths = $\frac{1}{3}\lambda$ and $\frac{2}{3}\lambda$ respectively between anode and grids. The anode and grid stray capacitances to earth are taken into account by including them in the delay line and making allowances in choosing the values of the shunt capacitors at the points along the line at which these electrodes are connected. The impedance of the line is chosen to be as low as possible, whilst still maintaining adequate oscillator loop gain. This enables the line to be designed with a sufficiently large number of sections to keep the cut-off frequency well above the centre frequency of the oscillation.

The video modulating signal is applied in push-pull to the modulator grids so that the mutual conductance of each valve is changed by an amount $\pm \triangle g$ dependent upon the amplitude of the video signal.

If ω_0 represents the centre frequency when each valve has a mutual conductance g_0 and ω is the instantaneous frequency when the mutual conductances have

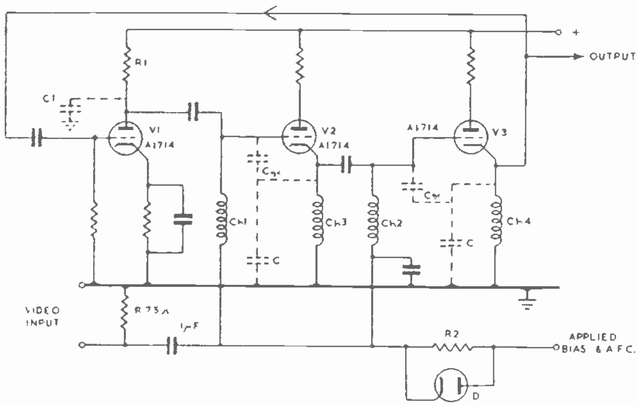

FIG. 12

F.M.O. circuit of Fig. 10 with alternative method of applying modulation voltages to cathode-follower stages.

changed to $(g_0 + \triangle_g)$ and $(g_0 - \triangle_g)$ respectively, it may be shown that

$$\cos\left(\frac{2\pi}{3} \cdot \frac{\omega}{\omega_0}\right) = -\frac{1}{2} \cdot \frac{g_0 + \triangle_g}{g_0 - \triangle_g}$$

Larger deviations (but with reduced centre-frequency stability) are obtained when using circuits in which the phase shift is produced over one or more cathode-followers included in the loop-feedback.

Fig. 10 shows a typical circuit using two cathode-follower electronic phase shift stages[11] and an amplifier stage connected to form an oscillation loop. About 60° phase-shift is allowed in each stage.

When using a CV 138 amplifier valve and two CV417 valves as cathode-followers, the oscillation frequency of this circuit tends to centre at about 45 Mc/s. Using valves type A1714 in all stages the circuit may be operated at a centre frequency of 70 Mc/s, when working the cathode-follower valves at a suitable grid bias potential. Fig. 11 shows a typical modulation characteristic for this Frequency-Modulated Oscillator, from which it will be seen that sufficient linear deviation is obtained for application as a T.V. frequency-modulated drive unit.

The circuits given by Ames and Cormack require modification in order to permit modulation at frequencies up to 5 Mc/s. The modulation input circuit shown in Fig. 10 enables video signals to be impressed upon the grid of the first cathode-follower without interference with the path of the 70 Mc/s oscillation at this point. The second cathode follower also receives video modulation on its grid, since it is connected to the cathode of the previous stage. Correct termination of the video input cable is achieved by means of the resistance R, Fig. 10, C forms a blocking condenser for the application of bias and A.F.C. potential via R_2 and diode D to the modulator valves. Ch1 is a high-frequency choke, self-resonant at 70 Mc/s.

An alternative modulation-input circuit is shown in Fig. 12. By means of additional high-frequency chokes Ch2 to Ch4 the full amplitude of the input video signal is made to appear across the grid and cathode of each of the two phase-shifting valves. Also by selecting the inductance values of Ch2 and Ch4 so that part of the cathode-to-earth capacitative reactance is balance out at centre frequency, it is possible to adjust the phase-lags at these points independently of the applied grid bias. This expedient enables the circuit to be worked at a higher centre frequency, but since the group delay is increased by the addition of parallel inductance, the permissible deviation is correspondingly reduced. An inductance L, may be placed across anode resistance R_1 to serve a similar purpose.

In order to make an assessment of the influence of the various elements in the circuit of Fig. 10, we may examine the equations given by Cormack for phase angle and gain of a cathode follower stage, when $R \geqslant 1/\omega C$.

$$\text{Phase angle, } \theta \doteqdot \tan^{-1} \frac{-\omega C g_m}{g_m^2 + \omega^2 C_{gc} (C + C_{gc})} \tag{i}$$

$$\text{and Gain, } \frac{e_0}{e_1} \doteqdot \frac{g_m + j\omega C_{gc}}{g_m + j\omega (C + C_{gc})} \tag{ii}$$

Where C = cathode-to-earth capacity, and C_{gc} = grid-cathode capacity.

Let us first determine the requirements for linearity of the frequency-modulation characteristic. From equation (i) we may write

$$\omega = \frac{g_m}{C} \cdot \frac{-1 \pm \sqrt{1 - 4K^2 C_r}}{2KC_r} \tag{iii}$$

where

$$C_r = \left(\frac{C_{gc}}{C} + \frac{C_{gc}}{C^2} \right)$$

and $\quad K = \tan\theta.$

Equation (iii) shows that $\omega \propto g_m$ provided that θ, the phase angle for each cathode-follower stage is constant at all oscillation frequencies. This is true if the other phase lags in the circuit also remain constant, since

$$2\theta + \theta_1 = -180^\circ \text{ during oscillation}$$

where θ_1 the phase angle of the amplifier stage

$$= \tan^{-1} - \omega C_1 R_1.$$

For θ_1 to remain constant, either $\omega C_1 R_1$ must be so great that $\theta_1 \doteqdot -90^\circ$ for all practical values of ω, or alternatively ω should be inversely proportional to C_1R_1. In either case we obtain the desired result, $\omega \propto g_m$. It follows that, by making $R_1 \propto \frac{1}{g_m}$, we should also be satisfying the requirement, $\omega \propto \frac{1}{C_1 R_1}$, to give constant phase angle θ_1. This effect was to a certain extent achieved by Cormack by the use of an additional cathode follower valve placed across R and modulated on its control grid. The present analysis shows that this device also results in improved linearity of the ω/g_m characteristic of the oscillator, especially if the g_m of the valve which modulates R_1 is adjusted to match that of the cathode follower stages.

Dealing now with the question of the generation of spurious amplitude modulation, from equation (ii) it would appear that the gain over a cathode follower stage remains constant provided that $\omega \propto g_m$. Thus if $g_m/\omega = K$ then

$$\frac{e_0}{e_1} = \frac{K + j\,C_{gc}}{K + j\,(C_{gc} + C)}$$

which is constant with frequency. If the oscillation loop comprises two cathode followers and an amplifier stage, the loop-gain will equal

$$\frac{R_1\,g_{m1}}{[1 + (\omega\,C_1R_1)^2]^{\frac{1}{2}}} \cdot \left[\frac{K + j\,C_{gc}}{K + j\,(C_{gc} + C)}\right]^2$$

where g_{m1} is the mutual conductance of the amplifier stage.

For this to be constant, again $\omega\,C_1R_1 \gg 1$ and $g_{m1} \propto \omega$, or alternatively if R has been modulated as described above so that $R_1 \propto \frac{1}{g_m} \propto \frac{1}{\omega}$ then constant gain will result when $g_{m1} \propto g_m$.

This result would indicate that it is desirable to apply modulation voltages to the amplifier stage also, in order to make $g_{m1}, \propto g_m$, thus equalizing the loop-gain at all frequencies and avoiding spurious amplitude modulation. This is not a valid conclusion, however, at high frequencies of operation, since the effect of a negative-resistance component in the cathode-follower input impedance has been neglected in the simplified gain equation (ii). The negative-resistance component results in a rising gain characteristic at high frequencies, over a cathode-follower stage in which the g_m is increased in proportion to ω. This compensates at certain frequencies for the falling gain/frequency characteristic of the amplifier stage and so results in low spurious A.M. generation even without the application of modulation potentials to the grid of V_1. This latter expedient however has been employed in certain 70 Mc/s, frequency-modulated oscillators used in television modems in order to minimize the generation of spurious amplitude modulation.

(c) Demodulation Circuits in the Television Link Modem

The spectrum arising from a carrier frequency-modulated with a 625 line television picture, having a peak-to-peak deviation of 6 Mc/s, requires a flat bandwidth of about 16 Mc/s in any network through which it passes, in order to transmit the signal without undue cutting of significant sidebands. A choice between the many methods of achieving this at intermediate frequency at once confronts the designer.

Very good results may be achieved with transformer-coupled stages but difficulty is experienced in the production and initial setting up of such amplifiers if primary and secondary are wound on the same former. For ease of tuning some form of bottom-coupling is to be preferred, which enables each coil to be wound on a separate former, tuned by its own slug. Capacities have to be kept to a minimum and valves of high figure-of-merit must be used. As alternatives to the transformer-coupled amplifier either the staggered-tuned amplifier or the feedback-pair amplifier are useful circuits.

In a typical T.V. link modem the I.F. amplifier is required to accept a signal of about 0·1 μW in 75 ohms from the output of the mixer and pre-amplifier stages. (This signal may be about 20 db greater if the mixer is preceded by a low-noise T.W.T. amplifier.) The output of the I.F. amplifier as it feeds into the limiter stages is of the order of 0·5 volts into 75 ohms. Thus an amplification of about 46 db is required. It is usual to allow for about 20 db fade at the amplifier input. To this end A.G.C. is fitted and the initial amplifier gain is increased by this amount. Care must input cable is achieved by means of the resistance R, Fig. 10, C forms a blocking condenser for the application of bias and A.F.C. potential via R_2 and diode D to the modulator valves. Ch1 is a high-frequency choke, self-resonant at 70 Mc/s.

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Because of the steep slope of the sides of the I.F. amplifier characteristic it is desirable to provide phase-equalization within the amplifier. This becomes a

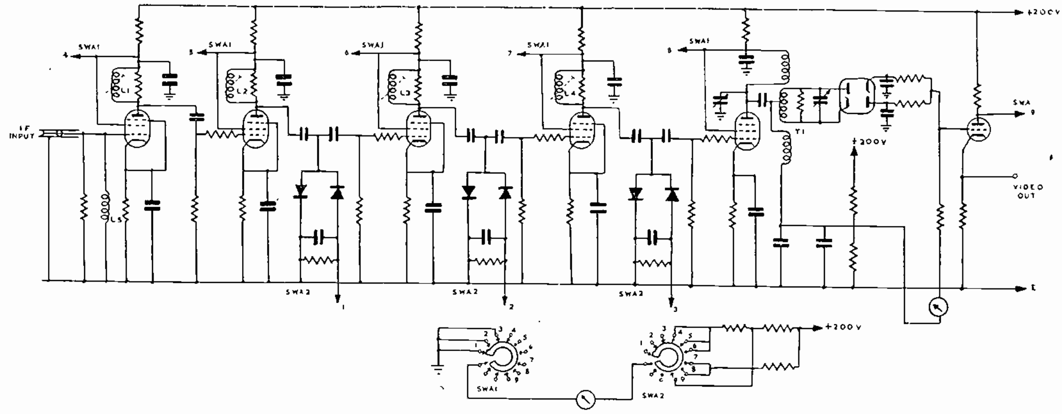

FIG. 13
Receive Terminal: 70 Mc/s Limiter Circuits and Discriminator.

necessary requirement in the case of a circuit which is carrying a colour sub-carrier, as in the N.T.S.C. system. The group delay/I.F. frequency characteristic of individual components of a long link must be extremely flat in order to achieve the overall constancy of about ± 4 mμ seconds required for colour sub-carrier transmission.

Design problems associated with the limiter stages of a television modem may be summarized as follows:—

(1) Wide bandwidth requires low Q circuits which provide but little gain.
(2) The final limiter must deliver a signal of about 10 volts rms to the discriminator. This signal should be of level constant within $\pm 0 \cdot 1$ db over a frequency range of about ± 8 Mc/s.
(3) The limiting action should be fast enough to smooth out amplitude variations which may have components of frequencies up to about 10 Mc/s.
(4) Spurious phase modulation should be minimized.

These requirements are substantially met in limiters of the type shown in the circuit of Fig. 13. Valves of high figure-of-merit (g_m/C) are again used; in addition, the final stages should be capable of producing sufficient anode current swing to provide the necessary drive to the discriminator stage.

Staggered tuning is used to obtain the required bandwidth. Limiting is achieved by the use of crossed-crystal diodes, fitted with auto-bias circuits to provide amplitude gating.

These crystal circuits are fast acting and tests have shown that they produce less spurious phase-modulation and less phase distortion than many other types of limiters.

Discriminator circuits for T.V. modems tend to be either of the well-known Foster-Seeley type or of the staggered tuned-circuit class.

The latter are easier to make and are probably simpler to set up in production, but the Foster-Seeley discriminator will usually deliver a larger output signal for a given deviation, and with care may be made to have a very linear characteristic over a deviation range up to about three-quarters of the frequency separation between peaks. In the T.V. case this peak separation is usually about 20 Mc/s.

Special methods of winding the discriminator transformer have to be adopted when working at frequencies of the order of 70 Mc/s to ensure maximum symmetry regarding the disposition of the windings and a minimum of out-of-balance capacitance between windings, or to earth.

A split-primary is used, one coil being wound on either side of the centre-tapped secondary, the two primary coils being then connected in parallel as shown in Fig. 13. Careful adjustments of coupling and resistive damping have to be made to realize optimum linearity.

A television-link discriminator must deal faithfully with a wide band of video components extending up to about 5 Mc/s. The design of the detector circuit is therefore most important; the combining of the video signals derived from each rectifier has to be performed with proper regard to phase at all video frequencies. A minimum of capacitance has to be used for I.F. by-pass purposes, otherwise the high-frequency response will not be maintained within the required limits.

In the circuit of Fig. 13 the combination of the two detector outputs is very simply achieved by direct connection through 1KΩ resistors, no valves being used for this purpose. A cathode-follower stage directly connected to the junction of the detectors ensures a minimum of loading at this point.

The output signal from the T.V. modem is supplied through a video amplifier and line-clamp very similar to that described above (Fig. 4) in connection with the transmitter equipment. Two 75 ohm output points are provided each capable of supplying a video signal of level = 1 volt D.A.P. to line.

Conclusion

The chief electrical performance requirements of present-day fixed television links have been given. It has been shown that signal-to-noise level requirements may be met over a long link using travelling wave tubes, at present available, in each broadband repeater, and at the terminal stations.

The circuitry used in the modems has been fairly well established for some years and has proved capable of meeting the detailed video performance specifications given above.

The newest unit in the television modem is probably the 70 Mc/s Frequency-Modulated Oscillator. For this reason more space has been devoted to this unit than to the other modem components. Work is still in progress on improved types of modulated-oscillator circuit in which grounded-grid circuits are used in place of cathode-followers. This idea was first suggested to the author by engineers of the British Post Office to whom acknowledgement is made.

Acknowledgements

The author wishes to acknowledge the work of his colleagues in the Marconi Company's Microwave Link and Television Laboratories who have worked together to produce link equipment from which much of the practical information contained in this article has been obtained.

References

[1] R. J. Clayton, D. C. Espley, G. W. S. Griffith, J. M. C. Pinkham. "The London–Birmingham Television Radio-Relay Link." Proceedings I.E.E., 1951, 98, Pt. 1, p. 204.

[2] G. Dawson, L. L. Hall, K. G. Hodgson, R. A. Meers, J. H. H. Merriman. "The Manchester-Kirk O'Shotts Television Radio-Relay System." Proceedings I.E.E., Vol. 101, Pt. I, No. 129. May, 1954.

[3] S. Fedida. "Wide Band Microwave Radio Links." Marconi Review No. 118, p. 96.

[4] C.C.I.F. XVth Plenary Assembly, Paris, July, 1949, Tome III.

[5] V. D. Landon. "The Distribution of Amplitude with Time in Fluctuation Noise." Proc. I.R.E., February, 1941, p. 50-55.

[6] M. G. Crosby. "Frequency-modulation Noise Characteristics." Proc. I.R.E., April, 1937, p. 472-514.

[7] F. A. Jenks. "Simplified Microwave A.F.C." Electronics, November, 1947, p. 120 and December, 1947, p. 132.

[8] W. L. Wright, British Patent Specification 707,029. "Wideband Thermionic Valve Amplifiers."

[9] H. W. Bode. "Network Analysis and Feedback Amplifier Design." D. Van Nostrand Co. Inc.

[10] K. R. Werdt. "Television D.C. Component." R.C.A. Review, Vol. IX, p. 85, March, 1948.

[11] R. N. Rhodes. "Factors in the Design of Keyed Clamping Circuits." R. C. A. Review, Vol. XV, No. 3, September, 1954.

[12] W. J. Bray. "The Travelling-wave Valve as a Microwave Phase-Modulator and Frequency-shifter." Proc. I.E.E., Vol. 99, Pt. III, No. 57, January, 1952.

[13] O. E. De Lange. "A Variable Phase-shift Frequency-modulated Oscillator." Proc. I.R.E., November, 1949, p. 1328-1330.

[14] A. Ames. "Wide-range Deviable Oscillator." Electronics, May, 1949.

[15] A. Cormack. "Wide-range Variable-frequency Oscillator." Wireless Engineer, Vol. 28, September, 1951, p. 266.

BOOK REVIEWS

Introduction to Solid State Physics, by C. Kittel. (Chapman and Hall. 56/-.)

In the preface the author says that this book is intended for senior and junior graduate students in physics, chemistry and engineering, and whilst a general familiarity with modern atomic physics at the undergraduate level is assumed, a course in quantum mechanics is not requisite to the understanding of the text. The book starts on familiar ground with a classification of solid and crystal structures. The physical, thermal electrical and magnetic properties of solids are then discussed. The last part of the book is concerned with metals, semi-conductors and imperfections in solids. A series of appendices deal with more advanced topics which require a formal background of quantum mechanics for their understanding. It is proper to ask therefore: how well has this book fulfilled the author's intention?

The method of presentation of these subjects is clear and concise, and the mathematics used present no great difficulty. One of the best features of the book, however, is the large number of questions set at the end of each chapter. These questions have been nicely chosen, not only to make sure that a thorough understanding of the text is necessary for their solution but that, in most cases, the results obtained from the problems amplify the text.

Although it is intended as an introduction to the subject and is not supposed to be a reference book the inclusion of detailed bibliographies and references at the end of each chapter enable a reader interested in pursuing any one topic in greater detail to do so.

The author may be criticised on the score that in some cases the treatment is so concise that understanding is difficult. As the text attempts to cover so large a ground in such a small space this defect is almost inevitable.

Generally, however, the book accomplishes effectively what the author set out to do, and it will be welcomed by those new graduates in the electronics industry who, because of modern research in transistors, phosphors and ferrites, require a good grounding in solid state physics, a subject which, in the main, is not given its full significance in the Universities of this country.

Active Networks, by V. C. Rideout. (Constable and Co., London. Price 42/-.)

This book, dedicated to the many scientists and engineers upon whose work it is based, is the fourth of the Prentice-Hall Electrical Engineering Series under the editorship of W. L. Everitt. Its aim is to enable those already having some acquaintance with the physics of vacuum tubes and passive linear networks to proceed to the more difficult study of the combination of these elements into active networks. This aim is admirably fulfilled and the book should take its place among those and how few they are—that can be confidently recommended to the average graduate at the start of a career in communication engineering to assist him to bridge the gap between his academic studies and the practical day to day work of an engineering development laboratory: in addition, it will be found a convenient reference book for the more experienced.

The author has successfully welded together material from many sources into a very readable form in which the treatment is both concise and in general adequate to a sound understanding of the operation and design of the circuits described. A keen sense of the practical is strong throughout. Whilst recognizing the difficulties associated with the choice of material in such a book it is to be hoped that in another edition more space will be found for the discussion on the various methods of modulation and for additional sections dealing with harmonic amplifiers, used so frequently in U.H.F. communication, and feedback pairs for wide band amplifiers. The gas tube noise reference source deserves more than scant mention.

Although steady state operation is the main concern, a short chapter is devoted to transient operation. The final chapter on Noise and Information Theory succeeds in presenting with clarity the essentials of both subjects and leaves the reader in no doubt as to the fundamental task with which he is faced in the sphere of communication. Each chapter throughout the book is followed by numerous references (among which British publications are adequately represented) and a collection of problems.

Among the weaknesses, the use of cycles for cycles-per-second is deplored. One meets it first in the Definitions on p. 10 and not until p. 29 is a footnote inserted justifying the use on grounds of common practice. In speech it may tend to be so, but in print it seems inexcusable. Perhaps the author is not really convinced for on p. 340 we read " peak deviation $= 162 \times 25$ cps $4{\cdot}05$ Kc."

Some confusion is bound to be caused, in the discussion on triodes on p. 41 et seq., by the definition of μ as the ratio of grid-cathode capacity to plate-cathode capacity. It is believed that this can be traced back through the literature to van der Bijl's paper of about 1930 in which an earthed grid connection is implicit. For the normal earthed cathode connection μ may be more accurately defined as the ratio of grid-cathode capacity to grid-plate capacity, i.e. the ratio of input capacity to that between input and output terminals, a definition which is equally effective for tetrodes and pentodes (and is in fact used on p. 52).

The use of the symbol K for voltage gain of an amplifier (G is used for power gain) is not a happy choice. Within a few pages K is used, in addition, to represent a cathode and as the constant in Child's equation. On p. 443 it appears several times though obviously as a misprint for the Boltzmann constant k. Similarly B represents gain-bandwidth product in sect. 3.20, but bandwidth only when discussing noise.

The parameter Q is defined in sect. 5.2, but is first used without explanation in sect. 3.12 (the shunt compensated interstage) where it could without apparent disadvantage be replaced by the single variable L/CR^2 thus avoiding the use of the two variables Q_0 and ω_0. In discussing Cowan and ring modulators it is stated incorrectly that side-frequencies of harmonics of the carrier are present in the output; side-frequencies of even harmonics are in fact suppressed.

It is most unfortunate that the treatment of delay, a most welcome feature and one that is rarely mentioned in texts, is seriously in error, for delay is here defined as θ/ω rather than $d\theta/d\omega$; as a result equations 3—24 and 3—55 are incorrect, as is Fig. 3—15 which has been derived from them. By accident delay is correctly given by equation 3—56 though a more compact form is obtained with ω_2 as parameter instead of ω_0. It is to be hoped that when the text is revised this treatment, which has been restricted to the low pass amplifiers, will be extended to other circuits.

The book appears to have been reproduced by a photolithographic process. In places the print appears grey rather than black whilst in others excess of ink makes some subscripts nearly unreadable. Fig. 5—14 is inverted and the rulings on three of the sets of valve characteristics in the Appendix are sketchy. Apart from these few blemishes format and typography are excellent, diagrams adequate and clear, and by present day standards the price is reasonable.

The Amplification and Distribution of Sound, by A. E. Greenlees (Third revised edition 1954, Chapman and Hall, 35s.)

This book is not, as might perhaps be inferred from the title, a textbook for the high fidelity enthusiast, but rather a handbook for those concerned with the planning, installation and operation of sound reproducing equipment; as such it can be recommended. The author has carried out the revision and enlargement with care and discretion. The text, re-arranged and in part re-written, is, as in the first edition, mainly descriptive with occasional simple calculations; more than thirty new figures have been added.

As in the earlier editions a general survey of the principles of operation of the component parts of basic equipment is supplemented by a wealth of detail on the practical aspects of distribution both for indoor and outdoor installations; rather more space could perhaps have been devoted to the special problems associated with equipments relying of necessity on their own power supplies as well as to a more critical discussion of the requirements for satisfactory reproduction from records running at the lower speeds.

The book has been re-set in new type with advantage to the reader.

Radar and Electronic Navigation, by G. J. Sonnenburg (Second Edition, pp. 282 + 4 Charts; London: George Newnes Ltd., 31/6 net.)

This book is concerned throughout with navigation at sea, not in the air, and is intended particularly for the use of Masters and Navigating Officers. Emphasis is on the operational rôle of the systems described and description of actual circuits is avoided.

After the General Introduction the chapters cover Echo Sounders, Direction Finders, Loran, Consol, Decca and Radar. The author considers that " Radar . . . is of course the most important electronic aid to navigation," and allocates more than one-third of the book to the chapter dealing with radar. Although many would disagree with his opinion, this space is needed in order to deal adequately with the subject, particularly in the interpretation of radar displays with due warnings of possible false echoes. Many excellent diagrams and photographs are included, and the association of a portion of a chart, two air photographs and a photograph of the p.p.i. of the Marconi Radiolocator IV, showing the mouth of the New Waterway at the Hook of Holland, is particularly pleasing.

Unfortunately accuracy sometimes suffers for the sake of simplicity, and the following remark, though correct, is dangerously misleading: " . . . even if the bearing . . . is constant at, for instance, 3°, a collision may nevertheless take place, because the ship can be driven out of her course by the current." There is an implication that the ship would be quite safe if the bearing were constant at some larger angle.

On the other hand one appreciates many homely and useful touches which reveal the author's familiarity with the equipments he is describing, and the book is confidently recommended as an introduction to the use of these electronic aids to navigation.

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