WS No. 19 Mark III

CARRIER TELEPHONY

Errata

Note: This page 0, Issue 1, must be filed immediately in front of Page 1, Issue 2, dated 24 Sep 68.

1. The following amendments must be made to the regulation:

2. Page 5, para 8

    Delete: 16Hz Insert: 16kHz
    Delete: 48Hz Insert: 48kHz

3. Page 15, para 33, line 5

    Delete: 180kHz Insert: 108kHz

4. Page 20, para 54, line 4

    Delete: nearer Insert: hearer

5. Page 20, para 55, line 5

    Delete: 30dB Insert: 80dB

6. Page 22, para 67

    Delete: Fig 3 Insert: Fig 13

7. Page 24, para 75.a., line 2

    Amend $\frac{L}{C}$ to read $\sqrt{\frac{L}{C}}$

8. Page 24, para 75.b., line 1

    Amend $\frac{L}{WC}$ to read $\frac{1}{WC}$

9. Page 25, para 76

    Amend $\frac{L}{C}$ to read $\sqrt{\frac{L}{C}}$

EME8/3047/Tels

END

CARRIER TELEPHONY

Note: This Issue 2, Pages 1-28, supersedes Issue 1, Pages 1-8 dated 18 Jan 43. It has been completely rewritten.

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INTRODUCTION

1. This EMER is divided into three sections: The first two, General Principles and Circuit Techniques are concerned exclusively with carrier telephony. The last Telephone Transmission Principles, deals with principles which, whilst not exclusively applicable to carrier telephony, are employed extensively in carrier systems.

GENERAL PRINCIPLES

FREQUENCY DIVISION MULTIPLEX

2. In order to permit the simultaneous transmission of several discrete telephone conversations over one transmission path a system of multiplexing is required so that the conversations may be separately channelled at the receiving end. This may operate either on a time-sharing basis (Time Division Multiplex) or else by translating the voice frequency channels to individual higher frequency bands so that they occupy different positions in the frequency spectrum and thus may be separated at the receiving end by selective circuits. This is Frequency Division Multiplex.

CARRIER TELEPHONY

3. The technique is very similar to that used for the transmission of radio telephony where voice frequencies modulate a high frequency carrier and for this reason it is usually called 'carrier telephony'. However, whereas in radio telephony the process is employed to translate audio to frequencies sufficiently high to be efficiently radiated from an antenna, in carrier telephony the carrier frequencies are generally kept as low as practicable in order to reduce losses in the transmission path. Each voice channel modulates a different carrier frequency, thus producing sidebands separated by the difference between the carrier frequencies.

TRANSMISSION PATH

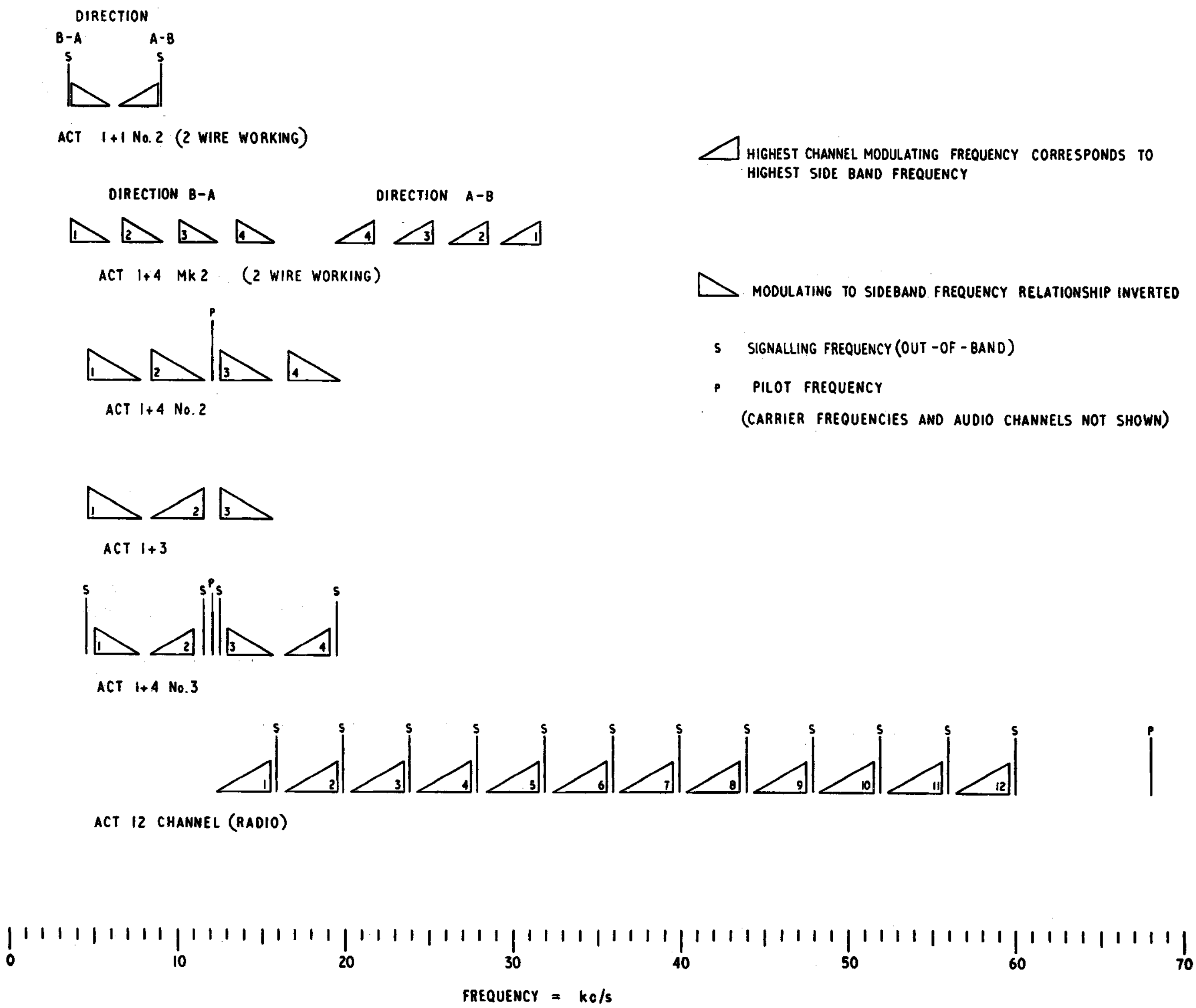
4. The transmission path may be open wires or cables, in which case the carrier sideband frequencies are transmitted over the path. It may also be a radio path, in which case the carrier frequencies are made to modulate the radio frequency carrier. In general, the radio frequencies will be v.h.f. or higher in order to provide a sufficiently wide band to accommodate all the carrier channels. Either a single path (2-wire; most commonly used with open wire construction) or separate paths for each direction (4-wire; quad cable or radio) may be provided.

MODULATION SYSTEM

5. In carrier equipments, the single-sideband, suppressed-carrier system is almost invariably used; all the information is contained in one sideband only and bandwidth and power handling capacity are wasted if anything else is transmitted. This means however that the carrier frequency of each channel must also be generated at the receiving end to demodulate the sidebands. When automatic gain control is required it is also necessary to transmit a single pilot frequency at a fixed level; this may also be used to operate an alarm when received levels vary outside acceptable limits.

CHANNEL SPACING

6. For modernhigh clarity speech channels an audio bandwidth of 300 to 3400Hz is considered necessary; with a single-sideband and allowing for practical filter



3047/8

Fig 1 - Frequency spectra of Army carrier equipments limitations it is feasible to space voice channels evenly 4kHz apart. Either upper or lower sidebands may be selected; indeed, the carrier set, telephone 1+4 No 3 selects upper and lower for alternate channels, thus requiring only two carrier frequencies at 8 and 16kHz.

7. It is now usual to provide all carrier frequencies from a single crystal-controlled master oscillator by frequency division and mixing (frequency synthesis) but in older equipments (eg ACT 1+4 Mk 2) individual self-excited carrier oscillators are used for each channel, the channels are unevenly spaced and restricted to 300 to 2600Hz audio bandwidth.

NUMBER OF CHANNELS AND BASEBAND

8. If a separate transmission path is available in each direction (4-wire or radio path) four channels will require approximately 16Hz of bandwidth and twelve channels approximately 48Hz. The total bandwidth occupied by all channels in one direction of transmission is termed the 'baseband'. If one path only is available (2-wire working) the bandwidth requirements will be doubled since the 'go' and 'return' transmissions of each channel must also be separated in the frequency spectrum. Fig 1 shows the frequency spectra for some Army carrier equipments. All are 4-wire unless otherwise stated.

Channel nomenclature

9. Carrier systems with one to four channels were usually required to provide additional channels on an existing telephone pair. This gave rise to the nomenclature '1+1' or '1+4' indicating one (or four) carrier channels added to the audio channel. Where twelve or more channels are provided this nomenclature is rarely used even when an audio channel is included because the transmission path is usually tailored to the requirements of the carrier equipment or the latter may be designed to go 'over the top' of an existing three or four channel equipment on the same path. The audio channel is commonly used as an 'Engineering order wire' (EOW) to facilitate technical cooperation between terminals.

BASIC CARRIER SYSTEM

10. Fig 2 shows an outline block diagram of one terminal of a 4-wire or radio system; the other terminal will be identical. For 2-wire working all carrier frequencies in one direction of transmission are made higher than those in the opposite direction; either different carrier supplies are provided for the modulators and demodulators or a second stage of modulation (group modulation, described in para 31) is applied in one direction. The send and receive lines are then connected to the 2-wire line by a directional filter consisting of a high-pass and a low-pass filter.

Two-wire to four-wire (voice) termination

11. Referring to Fig 2, one channel only is shown but the others are, except for the carrier frequencies used, identical. The 2-wire exchange line is connected to a hybrid transformer with its balancing network thus converting the 2-wire line to a 4-wire system.

Send direction

12. The speech frequencies modulate the carrier frequency and the desired sideband is selected by the bandpass filter. All sidebands produced by the four speech channels are then amplified in the transmit amplifier to a suitable level for feeding

to the send line. A high-pass and low-pass filter combination permit connection of the audio channel to the line without diverting the carrier frequencies.

Receive direction

13. The received carrier frequencies are separated from the audio circuit as before and directed by the parallel-connected channel bandpass filters to the appropriate channels. After demodulation it is usual to provide a stage of amplification for individual channels to allow their levels to be adjusted to local requirements.

Signalling

14. Since the speech channels are designed for 300 to 3400Hz, low frequency (17Hz or 50Hz) ringing current from the exchange is converted to 500Hz so that it may be transmitted over the carrier circuits. This is performed in the ringer panel R. Other signalling systems may be employed, eg out-of-band signalling, in which the exchange signals cause carrier frequency oscillators to be connected to the system whose frequencies lie between individual carrier channels, thus allowing supervisory signals to be sent during conversations and avoiding the risk of false operation due to speech or noise voltages.

Auxiliary equipment

15. Apart from the obvious omission of power supplies, Fig 2 omits several units necessary for effective working but not essential to illustrate the principle of carrier telephony. The more important of these are:-

a. Attenuators, variable and fixed (the latter usually called 'pads') to adjust levels and to minimise the effects of impedance variations.

b. Transformers for impedance matching.

c. Equalisers (networks with a 'shaped' frequency response) to correct for uneven responses in the transmission path or to improve filter operation.

d. Compensating networks at A and B Fig 2 to correct for the parallel connection of several bandpass filters.

e. Transmission measuring set (t.m.s.) for checking levels at various points in the system. This is often a separate item of equipment.

CIRCUIT TECHNIQUES

DOUBLE-BALANCED BRIDGE-RING MODULATOR

16. The modulators used in carrier telephony are required to handle low power only and rectifier type modulators, designed to give suppressed carrier modulation, are almost invariably used. A circuit, which suppresses both the carrier and the modulating signal, is shown in Fig 3.

17. The operation of the modulator is as follows: the amplitude of the carrier is large compared with that of the modulating signal; consequently, on the half cycle which makes point A positive, MR1 and MR2 are biassed 'on' with MR3 and MR4 biassed 'off'. On the next half cycle, MR3 and

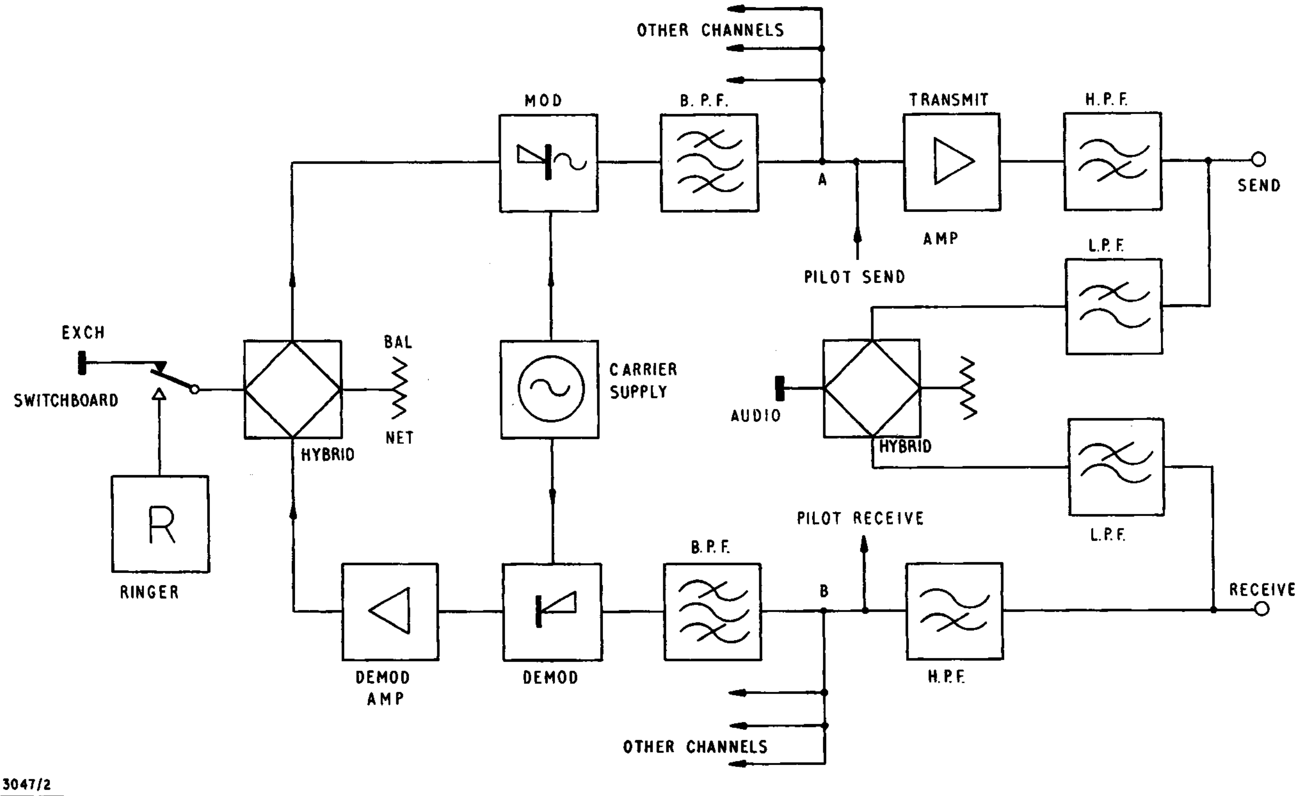


Fig 2 - Basic carrier terminal

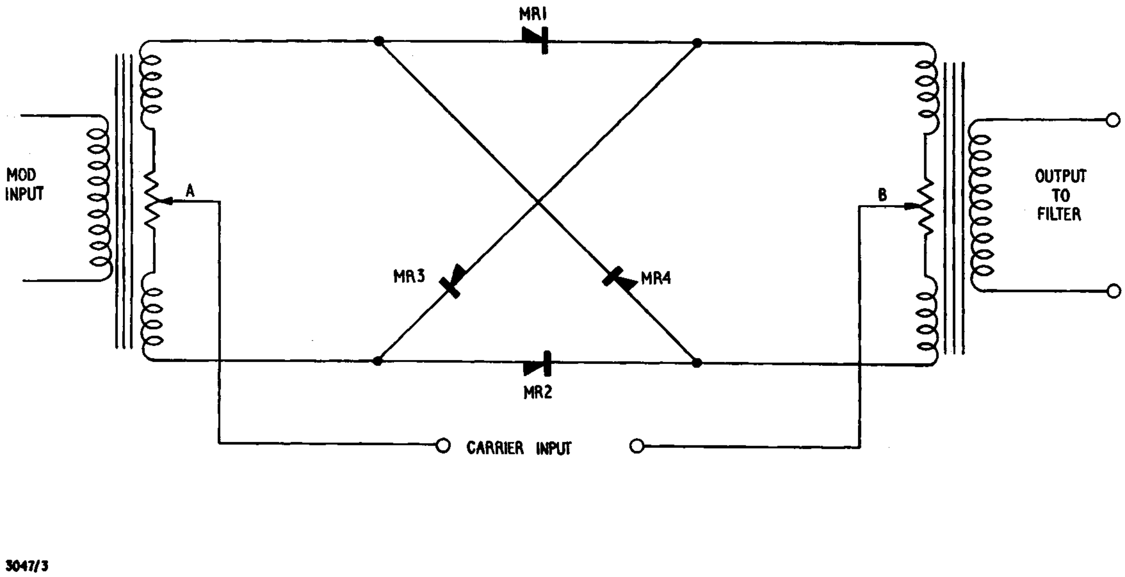


Fig 3 - Double-balanced, bridge-ring modulator

MR4 are conducting whilst MR1 and MR2 are biassed 'off'. The modulating signal at the output of the circuit is thus being switched in polarity every half cycle of the carrier frequency. No carrier frequency appears in the output since the carrier current divides equally between the phase-opposed halves of the transformer windings producing zero resultant flux. Fig 4(a) shows the result when a carrier is modulated by a frequency of one tenth the carrier frequency. From Fig 4(b) it can be seen that this very closely approximates to a square-wave modulation by a frequency of one tenth the square-wave frequency. In general terms, if we allocate unity amplitude to the square-wave it may be expressed by the Fourier Series:

$$\frac{4}{\pi}(\sin \omega t + \frac{1}{3}\sin 3\omega t + \frac{1}{5}\sin 5\omega t + \ldots\ldots\ldots\ldots)$$

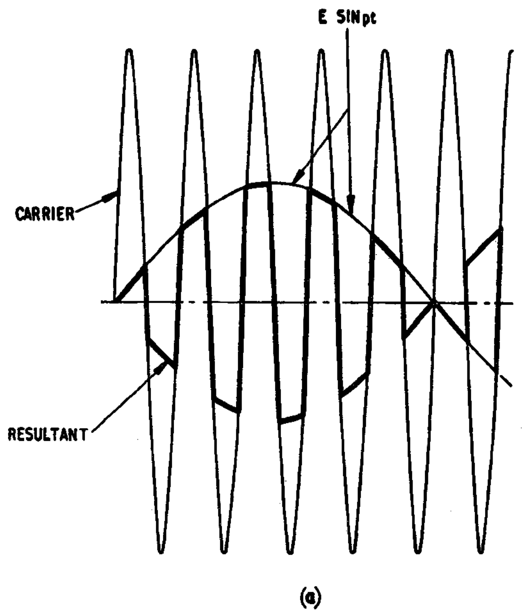
Multiplying this by the modulating frequency E sin pt we get:

$$\frac{4E}{\pi}(\sin pt \cdot \sin \omega t + \frac{1}{3}\sin pt \sin 3\omega t + \frac{1}{5}\sin pt \sin 5\omega t + \ldots\ldots\ldots\ldots\ldots)$$

$$= \frac{2E}{\pi}(\cos(\omega - p)t - \cos(\omega + p)t) + \frac{2E}{3\pi}(\cos(3\omega - p)t - \cos(3\omega + p)t)$$

$$+ \frac{2E}{5\pi}(\cos(5\omega - p)t - \cos(5\omega + p)t) + \ldots\ldots\ldots\ldots\ldots$$

18. This gives a difference frequency term and a sum frequency term representing respectively lower and upper sidebands together with sum and difference terms involving odd harmonics of the carrier and the modulating frequency, but no carrier or modulating frequency terms. All save the selected sideband are removed in the band-pass filter at the output of the modulator.



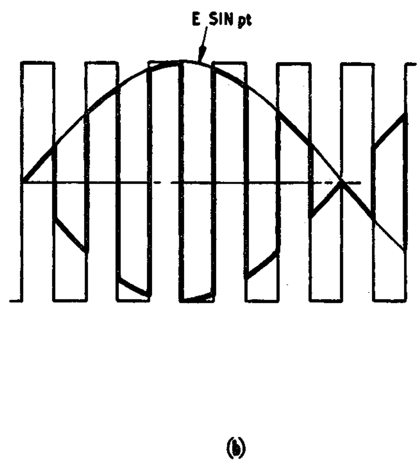


Fig 4 - Modulation waveforms

Demodulation

19. The same circuit may be used at the receiving terminal for demodulation and a circuit such as this is sometimes called a 'modem'. In this case the received sideband (eg the upper sideband, $E \sin (\omega + p)t$) takes the place of the modulating frequency and a locally generated carrier of the same frequency as the carrier at the transmitting terminal is applied to the centre taps as before. The output is a square-wave of the injected carrier frequency multiplied by the upper sideband frequency producing:

$$\frac{4E}{\pi}(\sin(\omega + p)t \sin \omega t + \frac{1}{3} \sin(\omega + p)t \sin 3 \omega t + \ldots\ldots\ldots\ldots)$$

$$= \frac{2E}{\pi}(\cos pt - \cos (2\omega + p)t) + \ldots\ldots\ldots\ldots$$

Thus the modulating frequency $(f = \frac{p}{2\pi})$ is recovered provided that the carrier frequency injected is identical with that at the sending end; if not the frequency at the demodulator output will differ by the same number of cycles, producing no serious distortion of speech provided the difference is small and still within the filter passband. A slight drift from exact synchronism of the carrier (say, 20Hz) is not therefore significant. The phase of the injected carrier is not important provided there is only one sideband involved.

Carrier leak

20. To assist in suppressing the carrier current in the output of the modulator a potentiometer is included at the centre of the balanced output and/or input transformer as shown in Fig 3. These are adjusted for the minimum carrier 'leak' at the output. They are not normally fitted to demodulator circuits since the output of these is usually at a frequency far removed from the carrier.

CARRIER FREQUENCY GENERATION

21. As previously stated, the generation of carrier frequencies by means of individual oscillators has now been superseded by a system using one high-stability crystal-controlled oscillator as the primary source for all carrier frequencies for both channel and group modulation and demodulation. Thus only one high-grade oscillator is required and only one frequency check need normally be made however many channels are included in the system. Standard channel spacing is 4kHz and hence the oscillator operates at this or at a multiple of this frequency.

Frequency synthesis

22. There are basically two methods of obtaining the required frequencies; frequency division using a relatively high frequency crystal, and frequency multiplication using a low frequency crystal, ie 4kHz. The latter method has only become practical in comparatively recent times, as previously no high-stability crystals were available at frequencies as low as 4kHz. However, it is clear that frequencies at 4kHz intervals cannot be obtained either by successive frequency multiplication or division and to obtain any desired multiple it is necessary to combine multiplication and/or division with mixing. With a 4kHz crystal amplified harmonics may be strong enough up to say the fifth (20kHz) but with both high and low frequency crystals, the sum or the difference frequencies obtained by mixing harmonics will give any desired multiple or dividend of the crystal frequency.

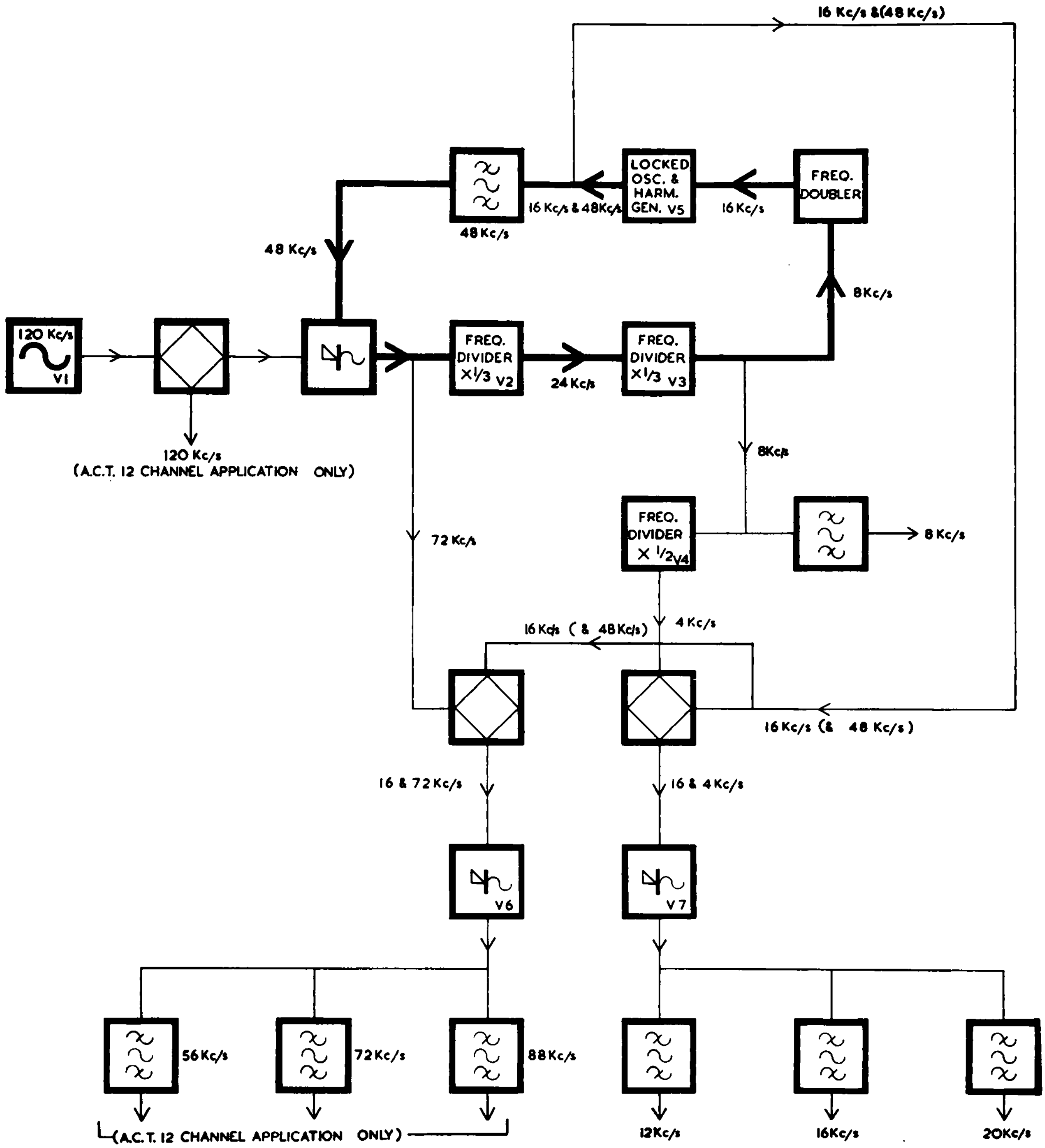


Fig 5 - Generation of carrier frequencies (120kHz crystal)

23. Examples illustrating both methods are shown in block diagrams Fig 5 and Fig 6; both produce the same range of frequencies from, in one case 120kHz, and in the other 4kHz crystals.

24. Fig 5 shows the technique used in the ACT 1+4 No 2 (cable system): see EMER Tels R 162. In this equipment a 120kHz crystal is used as the primary source. This is mixed with 48kHz from the third harmonic of the locked oscillator V5 and the 72kHz difference is extracted. The locked oscillator V5 also serves as a generator to give 16kHz, and although self-excited, is firmly locked by the 16kHz dividend. The frequency dividers are in this case all locked oscillators though multivibrator and Eccles-Jordan circuits can also be used for this purpose. Note the use of hybrid transformers to isolate separate sources feeding a common input.

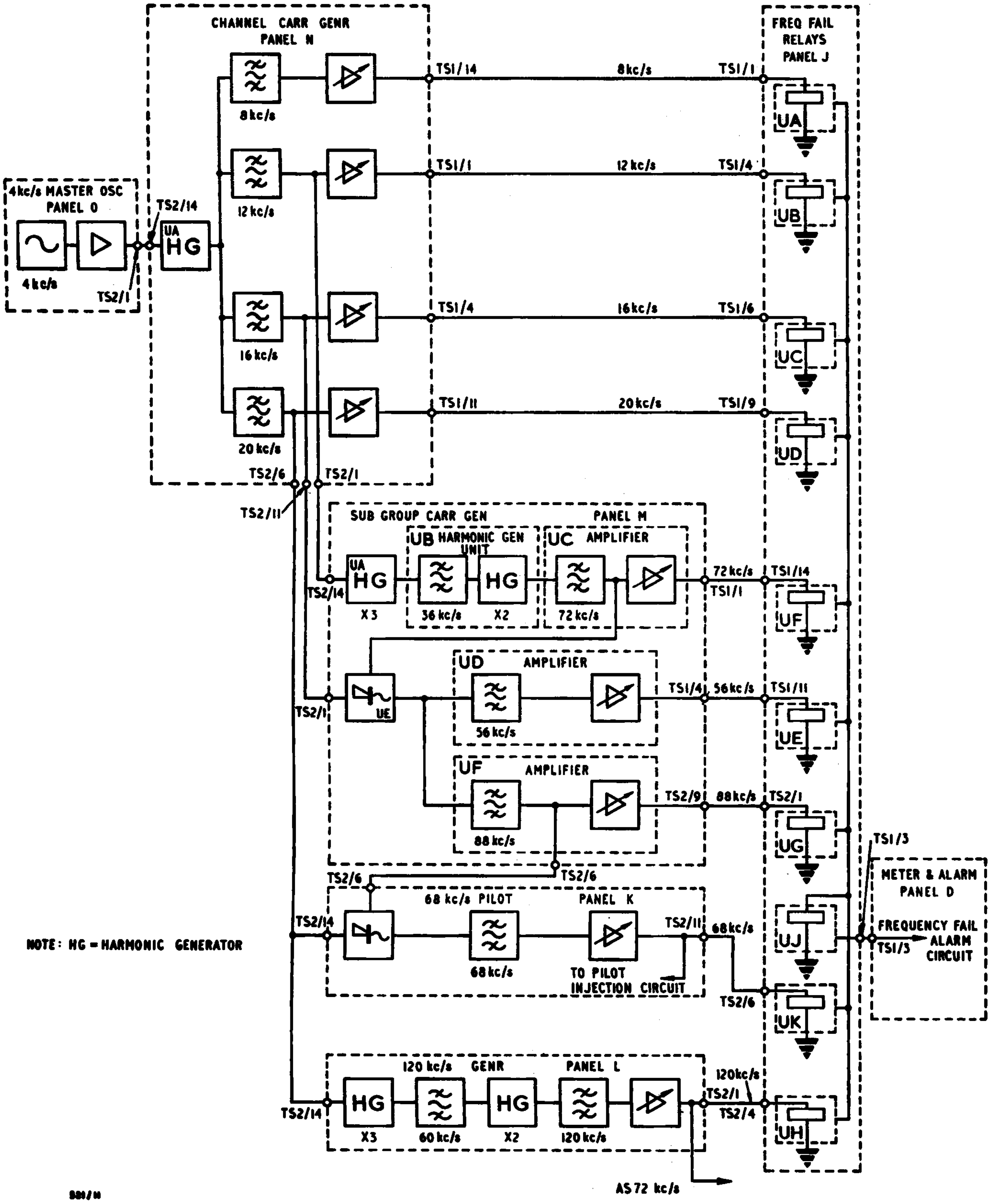
25. Fig 6 is taken from the ACT 12 channel (radio) equipment EMER in which a 4kHz crystal oscillator is used; the buffered output of this drives a harmonic generator which in this case is an over-driven amplifier. In the output circuit four bandpass filters are connected thus selecting the 2nd, 3rd, 4th and 5th harmonics for amplification by separate amplifiers. Further stages of frequency multiplication and mixing supply the Sub-group, Pilot and Group carrier frequencies (EMER Tels R 172 refers).

AMPLIFIERS

26. The frequency response of amplifiers is deliberately limited to essential bandwidth to reduce noise and stability problems. Amplifiers used after the final stage of demodulation (Demod amplifier) are simply audio amplifiers carrying one speech channel. It is usual to use a moderate amount of negative feedback in the interests of stability and freedom from noise, and gain adjustment is usually made easily accessible so that individual channel levels may be set to local requirements. A maximum gain of 20-40dB is usually adequate.

Carrier amplifiers

27. The requirements for these are more stringent. A transmit amplifier which amplifies all carrier frequencies sent to line, is almost invariably provided; a receive amplifier, performing a like function for all received carrier frequencies is less common. The reason for this may be deduced from para 60 on repeaters. Where however group modulation is used, it is usual to provide an amplifier after group and sub-group demodulation and the design of these is similar to the transmit amplifier. The most important requirement for a carrier amplifier is freedom from intermodulation distortion; this will infer also freedom from harmonic distortion. The reason is that, since a considerable number of discrete channel sideband frequencies are being passed through the same amplifier, any inter-modulation or harmonic distortion would cause serious crosstalk (para 53) between channels. This means that the amplitude distortion of the amplifier must be negligible, ie there must be constant gain for all levels of input over the working range. In addition, gain must be level over the full working range of frequencies, though this will not generally involve any considerable number of octaves as explained in para 33. If an audio channel is provided, this will by-pass the carrier amplifiers. Maximum gain is in the order of 40-65dB. With modern components there is no difficulty in providing ample gain with very low levels of distortion of all kinds by using ample negative feedback in a two or three stage amplifier. Combinations of voltage and current feedback are often used to give input and output impedances of the required value. The negative feedback confers additional benefits such as inherent noise reduction and gain stability.



NOTE: HG = HARMONIC GENERATOR

Fig 6 - Generation of carrier frequencies (4kHz crystal)

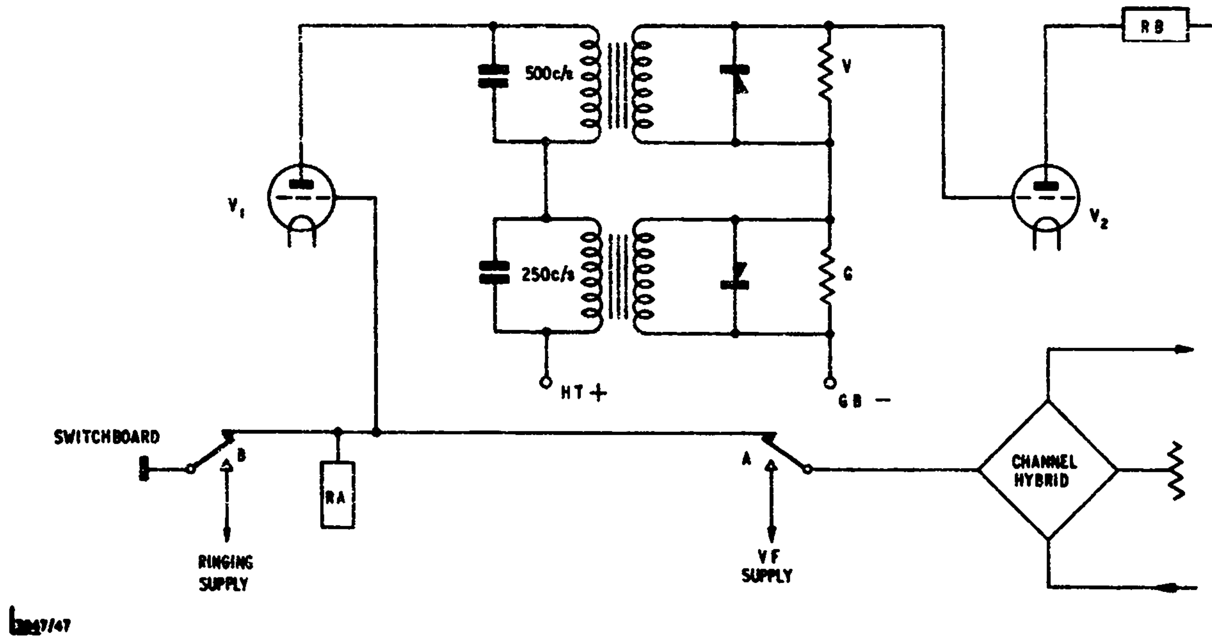


Fig 7 - Voice frequency signalling on a channel

SIGNALLING

28. In addition to the transmission of speech it is obviously essential to pass signals over the system to call the distant exchange, indicate when answered and show when a call has finished. Since it is not possible to pass either low frequency ringing current or d.c. (the usual form in which signals are given from a switchboard) directly over the carrier system, these signals must be converted either to voice frequency and thence transmitted over the appropriate channel in the same way as the speech, or else direct to a carrier frequency outside the channel sideband frequencies; the second method is called out-of-band signalling.

Voice frequency signalling

29. The essentials of this system are shown in Fig 7. On receipt of ringing current from the switchboard, relay RA operates to change over the channel hybrid to the voice frequency (500Hz) which is transmitted in lieu of speech. On receive, the VF tone (from the distant terminal) passes through the channel hybrid and is amplified in V producing a positive bias across the resistor V which backs off the negative bias on V2; anode current then operates RB and changes over the switchboard line to ringing supply. 'G' is a guard-circuit which prevents false operation on speech and is tuned to a frequency such as 250Hz or 660Hz which is a major frequency component of most human voices. Thus speech produces a voltage across resistor G also which opposes that of V. This type of signalling is generally satisfactory but, despite the guard

circuit, false operation is possible and results in interruptions to speech. It is also not possible for exchanges to signal during conversations so that a positive supervisory condition cannot be provided.

Out-of-band signalling

30. In this case the signalling frequency is just outside the transmit band of the associated channel; for example, channel sideband frequencies 16.6-19.7kHz, signalling frequency 16.175kHz. On receiving 17Hz ringing current from the switchboard, a relay, operating in a similar manner to relay A (Fig 7) connects the out-of-band signalling supply through an isolating hybrid to the output of the appropriate carrier channel transmit. The received out-of-band signal is similarly diverted through a suitable filter and operates a relay whose contacts perform a similar function to relay B. An example of an equipment using out-of-band signalling will be found in EMER Tels R 172. A somewhat different technique is described in EMER Tels R 192; a 3.5kHz signal is modulated and demodulated in the speech channel Modem, separation by filtering taking place after demodulation. This signal is sent continuously during idle periods, modulated with 17Hz on 'call' and cut off on 'answer'.

GROUP MODULATION

31. This is a system which arranges identical sets of channels (eg four channels at 4kHz spacing) one above the other by further stages of modulation applied successively, thus forming, for example a twelve-channel system. Super groups may be formed to extend the system further. An example of group modulation as applied to the ACT 12 channel (radio) (EMER Tels R 172) is given in Table 1.

Table 1 - Group modulation, 12-channel equipment

Mod. stage	Channel No.				Channel No.				Channel No.			
	(1)	(2)	(3)	(4)	(1)	(2)	(3)	(4)	(1)	(2)	(3)	(4)
	Channel carrier kHz				Channel carrier kHz.				Channel carrier kHz.			
1st	20	16	12	8	20	16	12	8	20	16	12	8
	1st Sub-group carrier kHz.				2nd Sub-group carrier kHz.				3rd Sub-group carrier kHz.			
2nd	88				72				56			
	Sub-group channel kHz.				Sub-group channel kHz.				Sub-group channel kHz.			
	108	104	100	96	92	88	84	80	76	72	68	64
3rd	Group carrier frequency = 120kHz											
	Final group channel frequencies (kHz)											
	12	16	20	24	28	32	36	40	44	48	52	56

32. The frequencies shown in Table 1 are carrier and virtual carrier frequencies only, the final group channels being the upper sidebands of the frequencies shown. The first stage of modulation is that of the 4-channel equipment of which there are three identical sets. Each channel is then translated to a different frequency at 4kHz intervals by sub-group modulation with three different sub-group carriers. These are then all combined in a final stage of group modulation.

33. The advantages are those of manufacture and maintenance which accrue from building up a system largely from standard 'bricks' with a reduction in the number of separate carrier frequencies which must be generated and the number of octaves

which filters, amplifiers, etc, must cover. Since no very high frequencies are involved this is (except for a very large number of channels) more important than limiting the top frequency. Thus 12 channels at 4kHz spacing require 48kHz bandwidth. From 0.3 to 48.3kHz this would cover more than seven octaves, but from 12 to 60kHz it covers only about two and a half octaves and from 60-180kHz less than one octave. The system also permits carrier systems to be 'stacked' above each other on the same path; a special case of this is the use of group modulation to translate all carrier frequencies in one direction to a higher band on a 2-wire system. With a very large number of channels, however, the top frequency limit must be restricted; eg a 600 channel system sends 60-2540kHz to line and requires a coaxial cable.

INERT TERMINAL

34. It is possible to operate a single-carrier channel without power supplies at one end of a 2-wire line, and this may be an advantage in remote locations. The system uses the so-called 'inert terminal' and it is necessary that the normal 'active' carrier terminal transmit the carrier as well as one sideband. The inert terminal may also be bridged across a line between two active terminals, giving party line facilities.

35. As shown in Fig 8, the received carrier channel is separated from the audio channel by filters HP and LP1; HP passes both sidebands and the carrier. The carrier channel is demodulated in modem M and the derived audio channel passes through LP2 to the switchboard line No 1.

36. Speech from switchboard line No 1 passes through LP2 and modulates the carrier received from the distant terminal; both sidebands are transmitted through filter HP to line so that both A and B active terminals can receive the transmission.

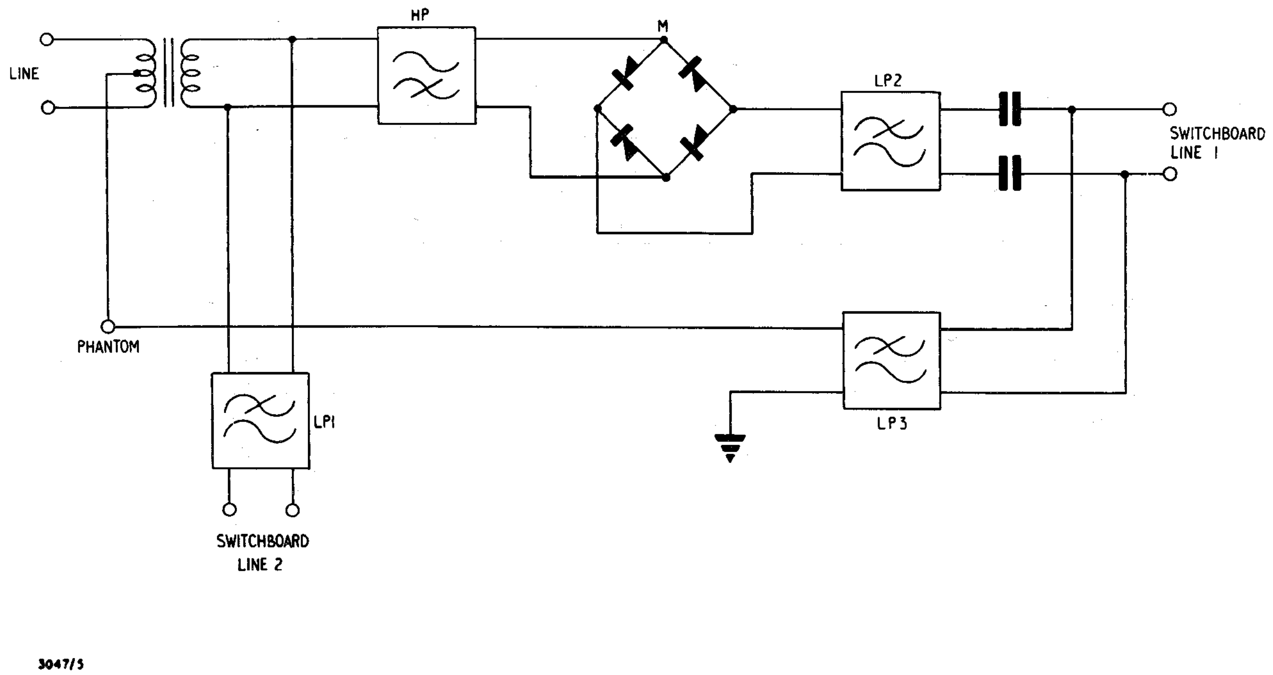


Fig 8 - Inert carrier terminal

37. The normal audio channel passes through LP1 to switchboard line No 2.

38. Ringing is by 17Hz current through LP1 for switchboard line No 2 and through LP3, phantom and earth for switchboard line No 1.

39. Examples of inert terminal equipments will be found in EMERs Tels R 122 and R 132.

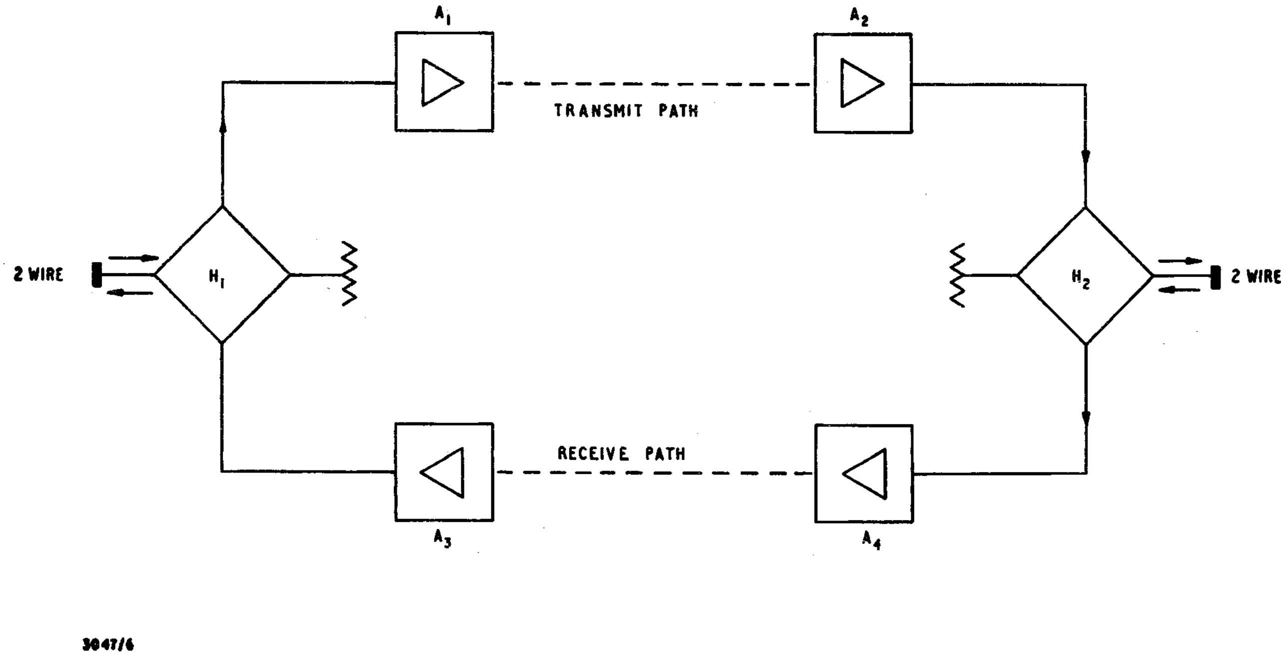


Fig 9 - Two to four-wire terminations

TELEPHONE TRANSMISSION PRINCIPLES

INTRODUCTION

40. Carrier circuitry and techniques involve the use of many principles common to the transmission of speech over long distances whether this be done by carrier telephony or otherwise. For example there is the problem of path attenuation which often varies considerably with frequency. Attenuation can be countered by amplification and equalizers can be used to compensate for the uneven frequency response of lines whose characteristics can be modified, within limits by loading coils.

41. Amplification increases the problems of crosstalk between circuits. Moreover, an amplifier is inherently a unidirectional device whereas telephone exchange lines are almost invariably bi-directional.

TWO TO FOUR-WIRE TERMINATION

42. This problem can be overcome by a hybrid transformer which transforms a 2-wire (bi-directional) transmission path into a 4-wire (one transmit and one receive) path or vice-versa; the amplifiers are inserted in one or both of the unidirectional paths.

43. Consider the general case shown in Fig 9; if at any frequency the total gain of the amplifiers A_1 to A_4 equals or exceeds the total attenuation of transmit and receive paths plus the attenuation of the hybrids H_1 and H_2 between these paths, oscillation will build up and the circuit will 'sing round' at the frequency of zero overall phase shift. The introduction of frequency-changing, as in carrier telephony, or of radio 4-wire paths into the circuit does not affect the fundamental loop gain condition of stability. The hybrids must therefore introduce maximum attenuation between the transmit and receive paths but minimum attenuation between these and the 2-wire lines.

Hybrid transformer

44. The hybrid is essentially a three-winding transformer and Fig 10(a) shows a perfect (loss-free) transformer with three identical windings. The impedances S, T, B and R are also identical. If a current is generated in series with R, then, since impedance S equals impedance B points X and Y are at equal potential and no voltage is developed across T. The power in the upper network will divide equally between S and B.

45. When the generator is in series with S as in Fig 10(b) the cyclic currents shown are generated. Since this is a perfect transformer the flux due to I_1 and I_2 will be equal and opposite to that due to I_3 and since all windings have the same number of turns:

$$I_3 = I_1 + I_2$$

$$I_3 = {}^e/R \text{ thus, } e = R\,(I_1 + I_2) \quad \ldots\ldots\ldots (1)$$

From Kirchoff's second law the algebraic sum of the p.d.'s and e.m.f.'s in the r.h. network is zero.

$$I_2B + e + I_2T - I_1T = 0$$

$$e = (I_1 - I_2)\,T - I_2B \quad \ldots\ldots\ldots (2)$$

Equating (1) and (2):

$$R\,(I_1 + I_2) = (I_1 - I_2)\,T - I_2B$$

$$I_2\,(R + T + B) = I_1\,(T - R)$$

Since T = R the r.h. side is zero and so the l.h. side must also be zero. But (R + T + B) cannot be zero; therefore I_2 must be zero. Thus no current flows through B and the output from the generator at S divides equally between T and R.

46. The practical hybrid transformer is an elaboration of this, the windings being divided to give balanced inputs as shown at Fig 11(a) which should be compared with Fig 10. Half the power (3dB) is lost from switchboard to transmit side and also from receive side to switchboard, the other half in the latter case being lost in balance network. An alternative version using two transformers with one reversed winding (sometimes called a two transformer terminal unit) is shown at Fig 11(b).

47. From para 44 and a comparison of Fig 10(a) and Fig 11(a) it is clear that the effective attenuation between receive (R) and transmit (T) depends on the degree of equality between the impedance of the balance network (b) and that of any line connected at the switchboard (S). In practice the characteristics of these lines will differ so the balance impedance can only be a compromise. Referring to para 43, the hybrid transmit-to-receive path attenuation will therefore vary as different lines are connected at the switchboard. This limits usable amplification to that which gives less than unity loop gain when the match between balance (B) and switchboard line (S) is worst. In practice, an overall loss (2 wire-to-2 wire, ie including both hybrid 2/4-wire losses) of 3 to 4dB is usually accepted.

48. Hybrids may be used in any application where it is required to couple two circuits to a common circuit without cross coupling between the two circuits. When unequal windings and impedances are used the device is sometimes called a 'skew hybrid'.

TRANSMISSION LEVELS

49. Since the minimum satisfactory level of received signal will depend on the ratio of this to total noise level (including path noise and noise generated in the receiving terminal) it would seem desirable to send at a high level. 'Noise' in this context means any electrical currents in the output not forming part of the wanted information. Such currents are caused, inter alia, by random electron motion in wires and components and in radio paths by electrical storms and man-made static.

50. Where radio is used as the transmission path the sending level is effectively the radiated power of the radio transmitter and the usual conditions applicable to radio transmissions will apply. Assuming the carrier and radio equipment are sited fairly close together the output level of the carrier station itself may then be low.

51. Where, however, the carrier signal is to be sent over a long open-wire line, or a cable, a higher level will clearly be desirable. There are, however, limitations on the sending level which can be used (it rarely exceeds 50 milliwatts in practice) and this limits the distance between carrier terminals or alternatively necessitates the introduction of repeater stations (which receive, amplify and re-transmit the signals) at intervals along the transmission path. These limitations should not be confused with the overall loop gain discussed in para 43; overall loop gain could be kept down by introducing attenuation at the receiving terminal but this would not affect the signal-to-noise ratio.

52. The main practical limitation to improving the signal-to-noise ratio by raising the sending level is near end crosstalk.

Crosstalk

53. The unwanted transfer of signal energy between channels on the same route is called crosstalk. This is generally caused by stray inductive or capacitive coupling or inefficient filtering which results in overhearing of conversations (intelligible crosstalk); there is also a type of crosstalk caused by intermodulation in non-linear circuits common to several channels (eg amplifiers) which is often unintelligible but nevertheless objectionable.

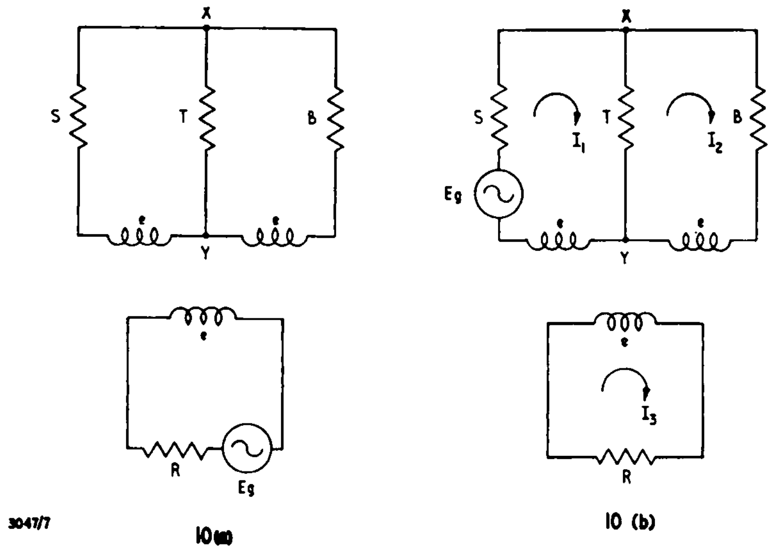


Fig 10 - Derivation of hybrid transformer

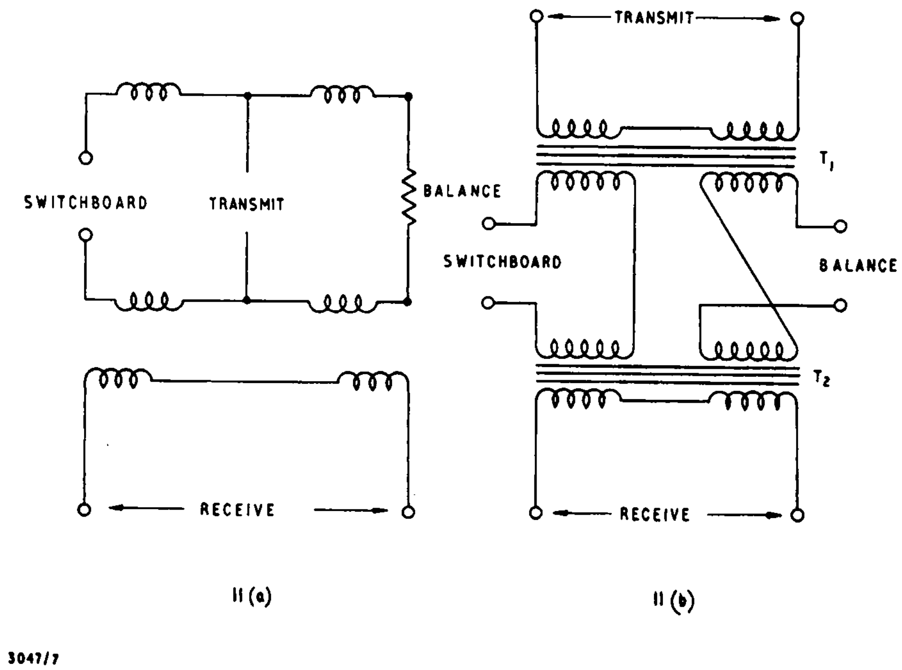


Fig 11 - Hybrid transformers

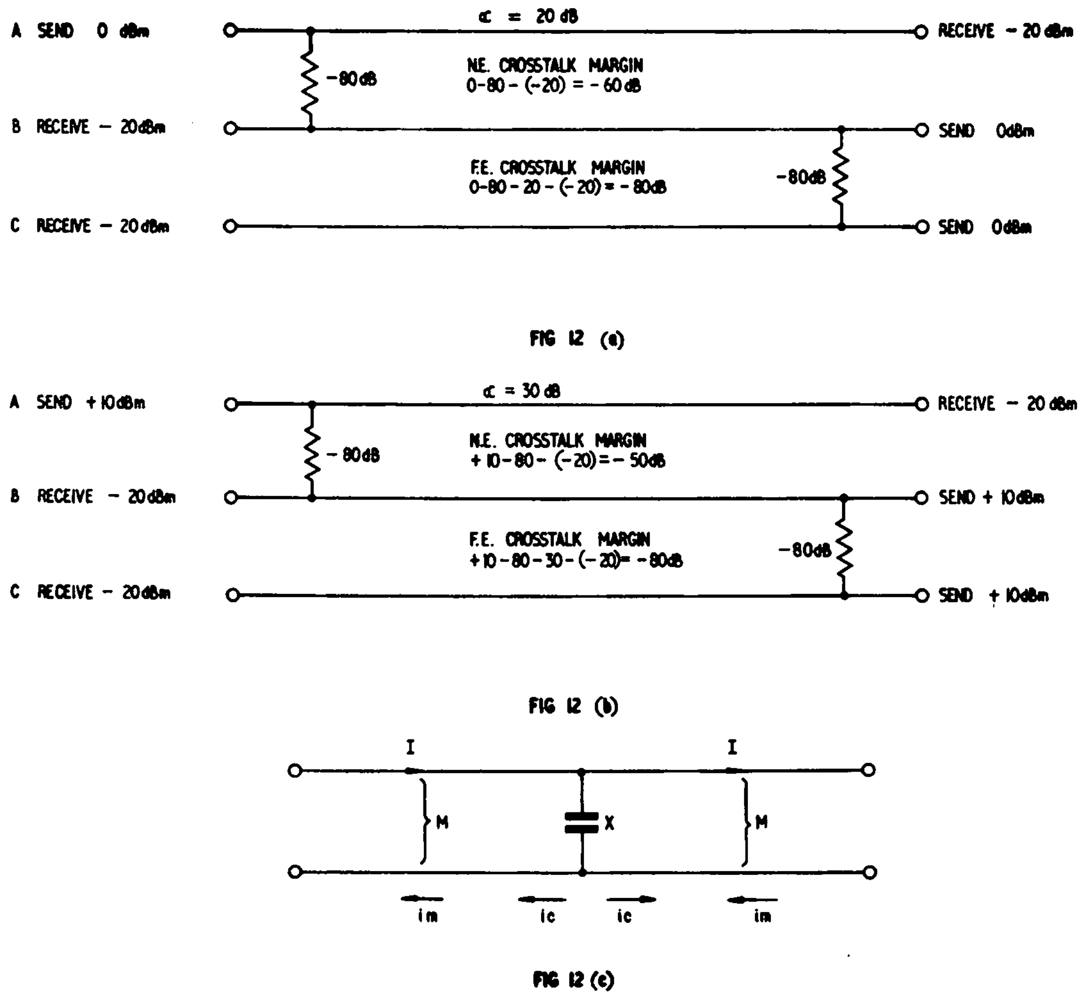


Fig 12 - Near end and far end crosstalk

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Near end and far end crosstalk

54. Near end crosstalk is crosstalk in which the disturbing channel is transmitting in the opposite direction to the disturbed channel. Far end crosstalk occurs when both channels are transmitting in the same direction. The terms 'near' and 'far' indicate the position of the crosstalk source with reference to the nearer.

55. Fig 12(a) shows three adjacent signal paths A, B, C. Consider the two paths A, B. The crosstalk in B due to A is the near end crosstalk. Assuming a crosstalk path of 80dB attenuation, this crosstalk is 60dB below the signal level. Consider now paths B, C. The crosstalk in B due to C is the far end crosstalk and is 30dB below the signal level. Near end crosstalk is the worse by an amount equal to the attenuation of the signal path.

56. Fig 12(b) shows the ineffectual result of attempting to make up for an additional 10dB path loss by increasing the sending level by 10dB. The far end crosstalk margin is unaffected but the near end crosstalk is worse by the increase in path attenuation.

57. Crosstalk occurs on any type of path including radio paths; in this latter case, increased radio frequency separation between 'go' and 'return' transmitters, improved circuit selectivity, linearity in common amplifiers, suppression of spurious emissions, and the avoidance of image frequencies all improve crosstalk. Near end crosstalk can also be reduced by using separate receive and transmit antennae, each having high directivity and maximum physical separation.

58. Where overhead lines or cable pairs are used an additional cause of increased near end crosstalk is evident from Fig 12(c). The capacitive currents flowing through the out-of-balance capacitance, X, divide in either direction; the current due to inductive coupling between the lines is induced in a direction opposite to that of the inducing current. Hence near end crosstalk is $i_m + i_c$ and far end crosstalk is $i_c - i_m$. Almost complete isolation of the go and return paths can be achieved by using separate screened cables for each, in which case only the far end crosstalk need be considered.

9. Consider again Fig 12(a) Paths B and C: the far end crosstalk in the terminal load (l.h. end) of the disturbed channel B has travelled the same distance as the speech in the terminal load of the disturbing channel C and is therefore in phase with it. By injecting into channel B load in reversed phase a small portion of the speech in the terminal load of channel C it is possible largely to cancel out the far end crosstalk in channel B.

Repeaters

60. The longer the path, whether radio or line, the greater will be the attenuation of the signal and the greater the amount of noise introduced. The ratio of signal-to-noise will thus progressively deteriorate and eventually reach unacceptable proportions. Once this has happened, amplification is of little use since this will raise the levels of both noise and signal and their ratio will be unchanged. Amplification must therefore be introduced at intermediate points along the transmission path where the signal-to-noise ratio is still satisfactory.

61. If the path is two-wire, hybrids are required at each repeater resulting in the losses mentioned in para 46, but in the four-wire (and radio) path these are required ly at the terminals. Radio repeaters usually translate the radio signal to intermediate frequency before amplification in either direction. If fading is deep it may be impracticable to work near zero loss as the automatic gain control range is limited and loop gain may be excessive at signal peaks.

PASSIVE NETWORKS

62. Passive networks are widely used in telephone transmission equipment to control signal levels or the direction and selection of signals. Levels may be reduced equally at all frequencies by means of attenuators or differentially with reference to frequency by means of equalizers. It is usual for amplifier to be given a flat characteristic (this simplifies negative feedback problems) and to use external passive networks to vary frequency response. Where different frequency bands are concerned selection and direction of signals is achieved with filters.

Attenuators

63. These are essentially 'lossing' devices for reducing equally at all frequencies the power transferred between circuits. The simplest form is the well known potentiometer or potential divider but this presents a different impedance in either

direction; see Fig 13(a). As shown in Fig 13(b) and (c) the impedance can be made the same or any desired value in either direction by the introduction of a third resistor thus producing a T or π attenuator from the potentiometer, which is, in fact, an L attenuator. Alternatively, one can consider the T and π as two L sections back-to-back.

64. In all three cases the attenuation will change if the load impedance is changed but not if the source impedance is changed; however, the input to the attenuator (and therefore the output) will change in the latter case. The connection of the load impedance will also affect the input impedance; it will have least affect on this where the attenuation is high because then the series arm(s) will be high and the shunt arm(s) low impedances. Thus an attenuator can be used to mask load impedance variations.

65. The attenuation of T or π attenuators is usually controlled by the tandem connection of more (or fewer) fixed attenuators (known as 'pads') or sometimes by the bridged T device of Fig 13(d).

66. Attenuators may also be used to 'match' impedances, ie to ensure that the circuit 'sees' the desired impedance looking either way. This involves a loss, of course, and the loss is always greater than the simple mismatch loss; one would not therefore use this as a device for transferring the maximum amount of power but it could be used, for example, to secure correct filter or equalizer load conditions and hence correct operation. The L attenuator gives minimum loss when used for matching.

67. An example is given in Fig 3(e): the L matching pad gives a matching loss of approximately 11.1/2dB when matching 600Ω to 150Ω. The simple mismatch loss would be 4dB.

Equalizers

68. By using reactances or a combination of reactance and resistance for the attenuator elements, attenuation can be made to vary with frequency, and thus, by suitable design, to correct for the unequal response of transmission circuits, particularly open lines and cables. The loss/frequency characteristic should clearly be the inverse of the line characteristic. Such a network is termed an equalizer.

69. In amplified transmission circuits a uniform response is particularly important, as will be seen from para 43. Even small peaks may result in the loop gain exceeding unity at that frequency and hence it will then be necessary to tolerate increased overall loss to achieve freedom from 'singing round'.

Loading coils

70. The losses in transmission cables normally increase with frequency but this may be modified by increasing their inductance. With telephone cables this is achieved by inserting inductance coils at regular intervals. This 'lumped' inductance reduces losses at the middle and lower frequencies but increases them rapidly above these because it forms with the line capacity a low-pass filter (Fig 14). Hence the technique is of use in carrier circuits only where relatively few channels are required. The cut-off frequency can be raised either by reducing spacing between coils (which is expensive, increases total resistance and may thus increase overall loss) or by reducing the inductance of the coils (which reduces the improvement below cut-off frequency) or preferably by a combination of both. Typical loading for carrier quad cable is 4.6mH every 440 yards as compared with 88mH every 2000 yd for audio cable.

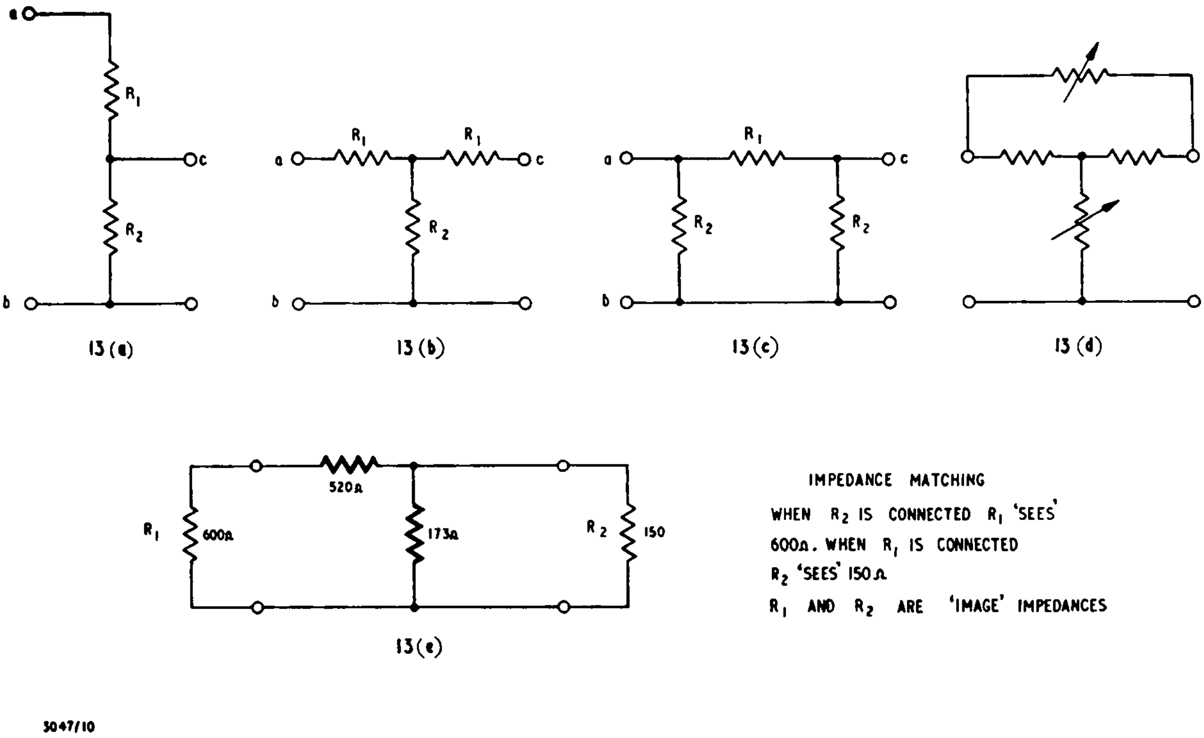

Fig 13 - Attenuators

Filters

71. Like the equalizer, the attenuation of a filter varies with frequency, but whereas the former's frequency attenuation curve is a more or less continuous slope, the ideal filter passes all wanted frequencies without attenuation and, at some predetermined frequency (the 'cut-off' frequency) changes abruptly from zero to infinite attenuation; its curve consists, in fact, of horizontal and vertical lines only. No practical filter ever operates like that and the problems of filter design are largely concerned with getting as close as possible to that ideal within the bounds of practical economy, bulk and weight. Some of the practical problems and their solution are briefly discussed in para 72 to 80; although a low-pass filter is considered, the general conclusions apply to all types.

Types of filter

72. The four main types of filter, low-pass, high-pass, band-pass and band-stop are easily recognized and their names explain their functions. Four examples, all T configuration, single-section, are illustrated in Fig 14. To obtain high attenuation it is usual to connect several sections in cascade.

Characteristic impedance

73. Since it is a reactive attenuator, a filter's attenuation at any frequency depends on the value of the load impedance (para 64). Further it will generally be necessary that its insertion in a circuit shall not alter the impedance at that point. Thus it will be designed so that its load impedance is that of the circuit and also its own characteristic impedance eg 600Ω. The characteristic impedance is the interative

impedance of a symmetrical network, ie that value of impedance which when connected across one end of the network causes the same value of impedance to be presented at the other end, thus fulfilling the insertion requirement above.

Simple filter section

74. Consider the T section filter in Fig 15(a). If Z_0 is to cause the impedance looking in at the other end to be Z_0 also, its value can easily be calculated since Z_{IN} is equal to the series/parallel network. Setting this equal to Z_0 we can evaluate Z_0 in terms of Z_1 and Z_2:

$$Z_o = \sqrt{Z_1 Z_2 + \frac{Z_1^2}{4}}$$

Applying this to the low pass-filter of Fig 15(a) where $Z_1 = j\omega L$ and $Z_2 = \frac{-j}{\omega C}$

$$Z_o = \sqrt{\frac{L}{C} - \frac{L^2\omega^2}{4}}$$

75. Consider how Z_o will vary with frequency in this filter:

a. When $\omega = o(f = o = d.c.)$ then $Z_o = \frac{L}{C}$ This is the characteristic impedance of a 'distortionless' line and is a pure resistance.

b. When $\frac{L^2\omega^2}{4} = \frac{L}{C}$ Z_o is zero and $\frac{L\omega}{4} = \frac{L}{\omega C}$. The equation of inductive and capacitive reactance denotes the resonance of the filter; $\frac{L}{4}$ represents the two $\frac{L}{2}$ coils effectively in parallel since input and output impedances are zero. This value of ω gives f_c, the cut-off frequency.

c. As f increases above this value, and as $\frac{L}{C}$ is constant, Z_o becomes a continuously increasing inductive reactance.

Z_o is therefore required to vary from a pure resistance to zero and then to an ever-increasing inductive reactance; clearly, a correct termination cannot be found for all frequencies and in practice it will usually be a resistance numerically equal to

$$\sqrt{\frac{L}{C}} = \sqrt{Z_1 Z_2}$$

The expression $Z_1 Z_2$ is constant (denoted by K) in filters such as these; hence they are called Constant K filters.

Attenuation

76. As a result of terminating the filter in $R = \frac{L}{C}$ and the fact that the inductances have resistance, the calculated attenuation curve for the practical filter will be of reduced slope as C in Fig 15(b) and there will also be some attenuation in the pass band. In carrier circuits, where channels are spaced as closely as possible we shall often need to attenuate heavily frequencies close to f_c. More attenuation at higher frequencies, such as would be given by additional constant K sections would be neither an economic nor an effective substitute and would also increase attenuation in the pass band.

M-derived section

77. To overcome this problem we can introduce either a series tuned (acceptor) circuit in place of C or parallel tuned (rejector) circuits in the series arms as Fig 16. The values of L and C required for this will be m times those for the constant K section and, as shown, additional reactances, also a function of m, will be required to resonate the acceptor or the rejectors. Since resonance is required at a frequency above f_c, m always has a value less than one and on this value depends the frequency of maximum rejection, f_r; see curve d in Fig 15(b). By putting m-derived half sections at either end of the constant K section(s) and choosing a suitable value of m, we can also present a more constant input and output impedance quite close to $\sqrt{\frac{L}{C}}$, within the pass band.

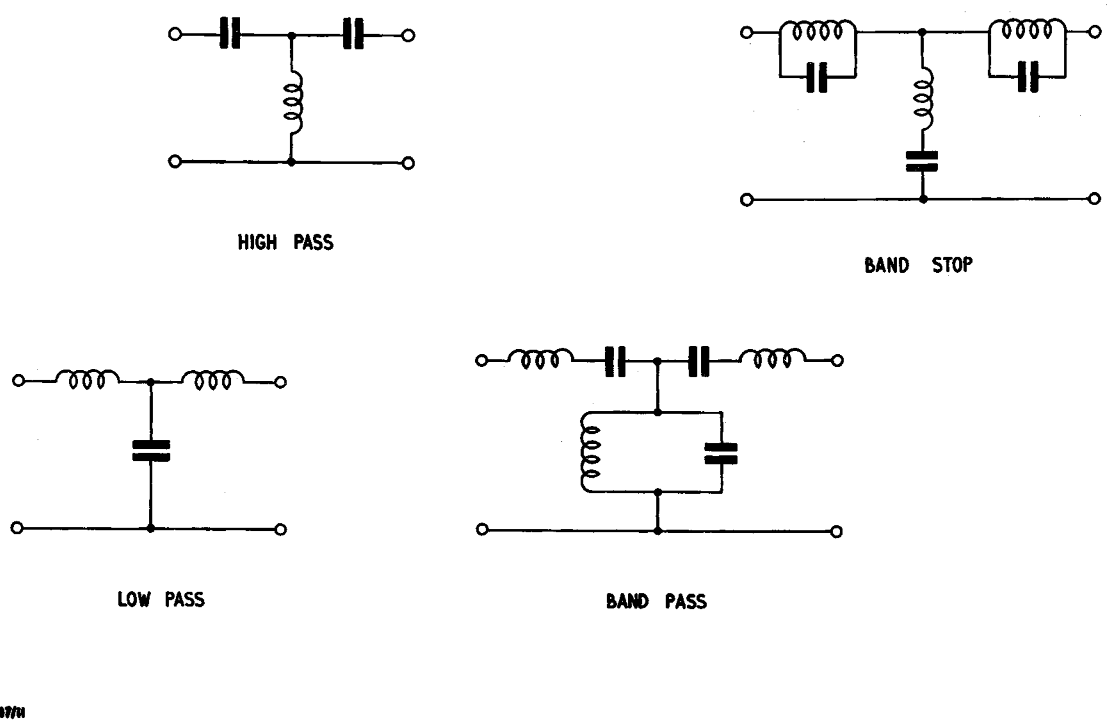

Fig 14 - Types of filters

Parallel connection of filters

78. In carrier circuits as we have seen a number of filters are often connected in parallel and the reactances of their end sections must obviously mutually affect each. Where a high pass and a low pass filter are paralleled it is generally sufficient to modify the value of the final inductor/capacitor in each filter so that it forms, with the parallel reactance of the other filter, an m-derived half section. A similar technique can be used for a group of channel band pass filters in respect of each filter for which there is at least one other filter above and below in cut-off frequency. For the end filters, ie a. lowest, and b. highest frequencies, the necessary compensating network must be provided in the form of a series-resonant circuit tuned for a. just below, and for b. just above the respective cut off frequencies of the two filters.

79. One result of this is that filters which have been designed to work in parallel will not give their specified performance if used separately.

80. In modern practice increasing use is being made of RC filters with integrated amplifiers. These have near-linear phase characteristics and fit in better with modern electronic circuitry. Future carrier equipments may include such devices.

POWER MEASUREMENT

81. In telephone engineering practice we are usually concerned to measure relative rather than absolute levels of power. This is because the absolute levels are generally continually varying with the speech input level. We also require to be able to assess the net effect of a number of amplifiers, attenuators etc in cascade and this task is made easier if we can use addition and subtraction instead of multiplication and division. Also the response of the human ear to sound intensity is approximately logarithmic; that is, it judges increasing levels of sound intensity on a scale in which each equal interval indicates x <u>times</u> the last interval, rather than x <u>added</u> to the last interval. It is clear therefore that a logarithmic scale would best suit power measurement in audio practice.

Decibels

82. The decibel, as its name suggests, is one tenth of a bel. A bel is the logarithm (to the base 10) of the ratio of two quantities (generally two power levels) so that a ratio of 100:1 (10^2) is 2 bels and thus 20 decibels; similarly 10:1 is 10 decibels. Since division is indicated by a negative logarithm, one tenth is -10 decibels (dB). The expression '10dB' therefore conveys nothing beyond 'ten times' and naturally raises the query 'ten times what?' In fact, it is the milliwatt which is the usual reference level, and to specify a power level quantitatively requires an expression such as '6dBm', (6dB reference one milliwatt) which (since $10^{\cdot 6} = 4$ approx) is four times one milliwatt, ie 4mW.

83. Where, however, it is required to take account only of gains or losses in an equipment, <u>relative</u> levels at different points may be shown as 'dBr' (decibels relative) or dBO, decibels relative to a point at zero dB; this point, in the case of carrier equipment, is usually the 2-wire audio input, ie the speech input; for test purposes a level of zero dBm at this point will make all other indicated levels dBm also.

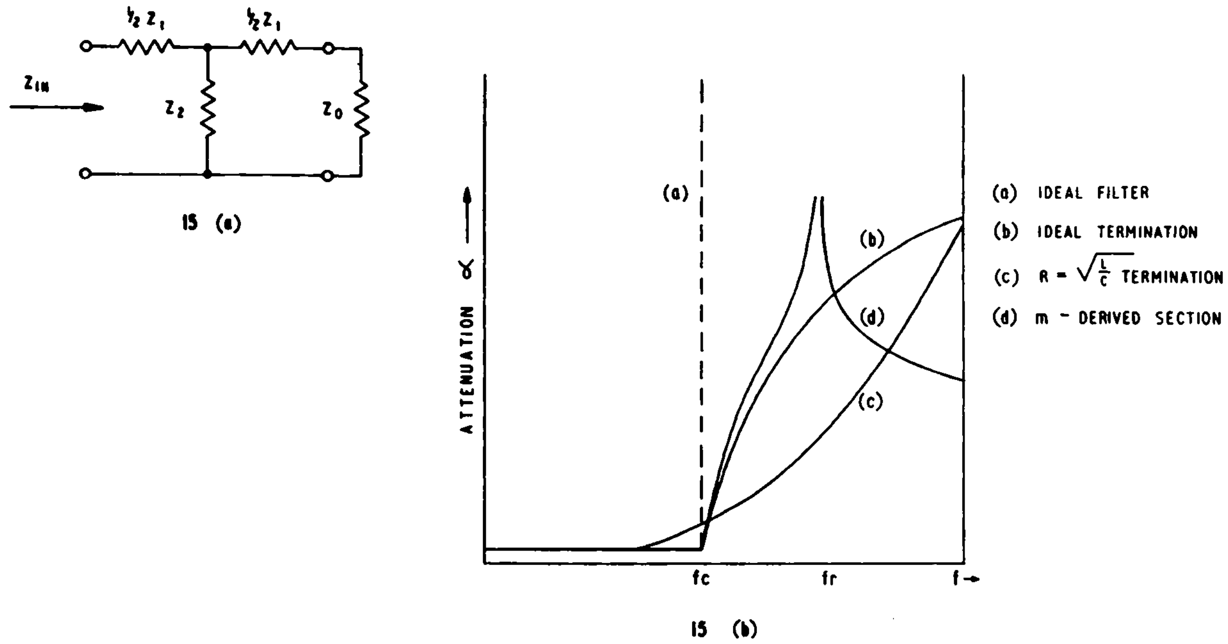

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Fig 15 - Characteristics of low pass filter

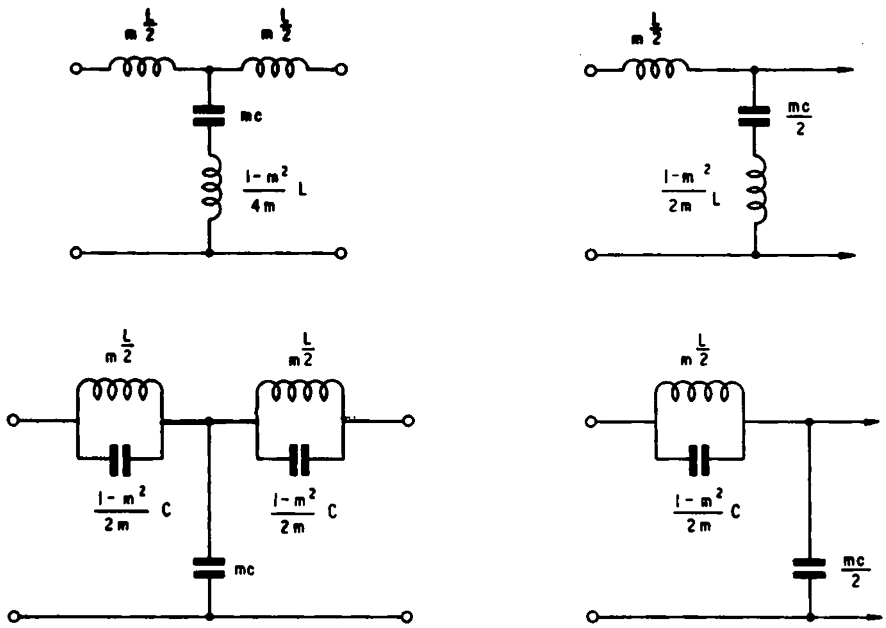

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Fig 16 - M-derived sections and half sections

84. As the response of the ear to different levels of sound tends to be logarithmic, and at normal levels, 1dB is about the smallest change it is possible to detect by immediate comparision; about 3dB is the smallest where memory is involved. This means that a loss of half the power would produce an only just perceptible drop in signal level, and in view also of the high gains achieved by modern amplifiers, explains why major power losses in hybrids, matching pads etc are readily accepted in telephone engineering to secure some other advantage such as circuit stability.

Nepers

85. This is also a logarithmic ratio but normally a current (or voltage) ratio and uses the natural or Naperian logarithm which is to the base e = 2.71828. Under conditions which make power proportional to I^2 or E^2, 1 Neper = 8.7dB, approximately. Like the bel, it is rather too large a unit for general use.

DB meter and TMS

86. The usual decibel meter is simply an a.c. voltmeter whose scale has been calibrated so that there is (for example) a 10dB interval between points $\sqrt{10}$ volts apart; other points follow a similar logarithmic relationship.. When used across the design impedance, the readings will represent an absolute level of power. Thus across 600 ohms 0.775 volts gives 1 milliwatt (OdBm) and -20dBm at 0.0775 volts. If used across any other impedance each reading will indicate correct levels relative to each other reading, but not relative to one milliwatt.

87. A transmission measuring set in a carrier equipment, usually consists of a dB meter capable of reading correctly at audio and the carrier frequencies involved, pads, terminating resistor(s) and a stable audio oscillator to supply a signal to the 2-wire audio channel inputs.

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