COURSE OF TECHNICAL INSTRUCTION BOOK

LONG LINE EQUIPMENT 1

PUBLISHED BY
THE POSTMASTER-GENERAL'S DEPARTMENT

THE AUSTRALIAN POST OFFICE

COURSE OF TECHNICAL INSTRUCTION

Engineering Training Section, Headquarters, Postmaster-General's Department, Melbourne C.2.

LONG LINE EQUIPMENT 1

Issue 2, 1951.

The subject of Long Line Equipment is presented in three books -

LONG LINE EQUIPMENT I includes the elementary theory of transmission, principles of carrier telephony and telegraphy, details of the apparatus used and information about crosstalk and power plant.

LONG LINE EQUIPMENT II includes voice frequency repeaters, signalling on trunk circuits, description of carrier telephone and telegraph systems and radio programme transmission over trunk lines.

LONG LINE EQUIPMENT III includes long line installation, maintenance and testing notes, line considerations, and transmission measurements.

LONG LINE EQUIPMENT I.

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COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
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COURSE OF TECHNICAL INSTRUCTION.

LONG LINE EQUIPMENT I.

ELEMENTARY THEORY OF TRANSMISSION.

CONTENTS:

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1. INTRODUCTION.

1.1 This subject of Long Line Communication deals with the theory and practice of providing Telephone and Telegraph Transmission and Signalling over lines (generally trunk lines) of considerable length. Broadly speaking, Long Line Communication is concerned with two problems -

(1) Improving the performance of line plant so that it transmits a wider range of frequencies over distances longer than those normally encountered in the local connections studied up to the present and in such a manner that all frequencies in the range, or predetermined portions of it, are affected equally by the line characteristics.

(2) Increasing the efficiency of line plant so that a number of communication channels can be obtained from each wire pair, or physical circuit as each wire pair is termed.

1.2 It is desirable that the student should have some idea of what is involved in each problem before proceeding with a detailed study, otherwise the different sections dealt with may not appear relevant. For this reason, the two problems outlined above are expanded, using general terms only, as follows -

Problem (1).

Earlier studies have shown that intelligible speech involves the transmission of a band of frequencies extending from about 200 c/s up to about 3,000 c/s. This transmission of a band of frequencies, rather than a single frequency, is because speech sounds are extremely complex and contain many frequencies present simultaneously, the band mentioned above containing the most important. The lower frequencies determine the pitch of the speech whilst the higher frequencies, that is, the harmonics, govern the intelligibility. By varying the number and the amplitude and phase relations of the harmonics, different consonant and vowel sounds are produced on the same fundamental. To preserve intelligibility, the amplitude and phase relations present in the original speech must be maintained during transmission. All telephone lines contain series inductance and resistance and shunt capacitance and leakance and, with this in mind, four effects are produced on a band of speech frequencies transmitted over a line.

(i) The series inductance of a line "irons out" the higher frequencies to a greater extent than the lower frequencies, whilst the shunt capacitance "drains" a greater proportion of the higher frequencies away from the receiving equipment, so producing incorrect amplitude relations at the equipment and impairing intelligibility.

(ii) The reactive elements mentioned above produce phase angles between current and voltage which vary with frequency, so further impairing intelligibility.

(iii) An equal proportion of the applied voltage is "dropped" across the series resistance and inductance of each section (say, mile) of line, and an equal proportion of the current sent into the line is drained away by the shunt resistance and capacitance of each similar section. This occurs at all frequencies, and means that the length of any type of line over which satisfactory transmission is possible is limited, the limit being reached when the received current and voltage become too small to be useful.

(iv) The preceding effect increases as the frequency rises, because inductive reactances increase whilst capacitive reactances decrease as the frequency rises.

It is apparent that there is an upper limit of frequency above which transmission over a line is impossible without the aid of special equipment. The first problem, therefore, consists largely of examining how the behaviour of a line varies with frequency and evolving methods of correcting that varying behaviour.

Problem (2).

The efficiency of line plant is raised for economic reasons. Long trunk lines are extremely costly, and it is desirable, for economic reasons, to obtain as many communication channels as possible from each physical circuit. As discussed more fully later, this is done by translating the voice frequency bands from a number of telephones up into different bands of higher frequencies. These bands are simultaneously transmitted over the same physical circuit, suitable selective circuits directing each band of frequencies to its appropriate equipment for retranslation prior to applying the resultant voice frequencies to the appropriate telephone. This means that lines over which this equipment operates must be capable of transmitting a much wider band of frequencies than a line over which only the voice frequency band is transmitted. The two problems outlined above are not separate, therefore, from a lines point of view,

/as

as the line problems remain much the same no matter what frequency range is involved. The equipment, however, is a special study, as the arrangements for frequency translation and retranslation are not necessary in voice frequency circuits.

1.3 For convenience, the subject has been divided into three parts, each part being published separately as Long Line Equipment I, II and III. Long Line Equipment I deals with fundamentals, a thorough knowledge of which is necessary before studying Long Line Equipment II, which deals with Carrier Systems, etc. Long Line Equipment III deals with Transmission Tests, Maintenance Tests, and the like.

1.4 To appreciate fully some of the subject matter dealt with in these books, a knowledge of the Alternating Current Theory dealt with in Applied Electricity III is necessary, also a knowledge of the Trigonometry dealt with in Practical Mathematics II. The Alternating Current Theory and the Trigonometry dealt with in these books is simple, and only enough to enable students to gain a good knowledge of how equipment functions. For an Engineering treatment of the subject, students are advised to consult such works as Communication Engineering by Everitt, Outline Notes on Telephone Transmission Theory by Palmer, etc.

2. PRIMARY CONSTANTS OF A TELEPHONE LINE.

2.1 Communication signals are alternating in nature, so that the impedance of a line is of some importance in considering the effect of the line on the signals sent over it. A telephone line has four components which determine the magnitude and nature of its impedance. These four components are termed the Primary Constants of the line when measured in terms of unit line length. They are -

(i) Series Resistance, R ohms per unit length. This resistance is merely the conductor resistance per loop mile, this being the unit of length used.

(ii) Series Self-Inductance, L henrys per loop mile. If the position of the two sides of a line coincide, then the resultant magnetic field produced by the line is zero because the flux produced by one side of the line is equal and opposite to that produced by the other side and, as the two lots of flux are produced in exactly the same space, they neutralise exactly. When the two wires do not occupy the same position, as is the case in practice, it is not possible for the lines of force produced by each side to occupy exactly the same positions, so that what is left over from the neutralising process appears as a species of leakage flux to induce voltages across the side of the line opposite to that which produced it. As with series or conductor resistance, the unit length is the loop mile.

(iii) Shunt Leakance, G mhos per mile. It is not possible to obtain an insulating medium with an infinite resistance to separate the two sides of a line. The reciprocal of this insulation resistance per mile of line is termed the Shunt Leakance or Shunt Conductance.

(iv) Shunt (or Mutual) Capacity, C farads per loop mile. The two sides of a telephone line form the plates of a condenser, with the insulating medium separating them acting as the dielectric. The capacity of the condenser so formed by one mile of line is termed the Shunt, or Mutual, Capacity.

2.2 Typical examples of the four primary constants for some different types of line are given in Table 1. The constants given in the tables used in this book are taken from sources which are acknowledged in the tables - they are not to be taken as the Australian standards, which are being prepared at the time of writing.

/Table 1.

Type of Line	R, Ohms	L, Henry	G, µMho	C, µF
100 lb. Copper Aerial	17.6	0.0039	1	0.0081
200 lb. Copper Aerial	8.8	0.00366	1	0.0086
300 lb. Copper Aerial	5.9	0.00355	1	0.0089
600 lb. Copper Aerial	2.9	0.00331	1	0.0096
10 lb. Cable	176	0.001	1	0.065
20 lb. Cable	88	0.001	1	0.065
40 lb. Cable	44	0.001	1	0.065

TABLE 1.

(From Telephony, Vol. I. Herbert & Proctor.)

2.3 From the above four constants a network could be made up to simulate a short length of line. Fig. 1 shows the network which simulates one mile of 200 lb. copper line, the values being those given for this class of line in Table 1.

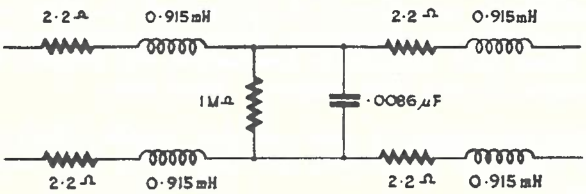

FIG. 1. NETWORK TO SIMULATE ONE MILE OF 200 LB. H.D.C. LINE.

This network is accurate enough for examination purposes, but does not accurately take the place of one mile of line as the constants are "lumped" together and not evenly distributed over a length of one mile.

3. ELECTROMAGNETIC WAVES.

3.1 In wire transmission, the electrical energy generated by the transmitting equipment is guided to the receiving station by the line conductors. It is well known that the passage of current through a conductor produces a magnetic field about the conductor and that the difference in potential between two conductors produces an electrostatic or electric field between them. The lines of force of these two fields are at right angles to each other and also at right angles to the conductors, and therefore to the direction in which energy is being transmitted. This is shown in Fig. 2, which also shows the electric field produced by the difference in potential between two line conductors and the magnetic field produced by the current flowing through the conductors.

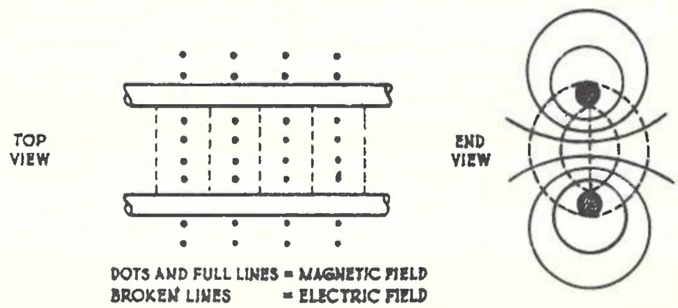

FIG. 2. FIELDS PRODUCED ABOUT LINE CONDUCTORS.

3.2 Thus, the energy transmitted is not confined to the wires themselves but is stored in the electric and magnetic fields surrounding them, the wires merely serving to guide the fields. The fields travel along the line in the form of waves, the combination of the two fields being termed an electromagnetic wave.

3.3 It is important to have some idea of the production and behaviour of these waves, as the behaviour of circuits to the transmission of electromagnetic waves is exactly the same as that of any other medium conducting energy in the form of waves.

4. PHYSICAL CONCEPTION OF WAVE PROPAGATION ALONG LINES.

4.1 Before proceeding further with this Paper, a physical picture of the production of electromagnetic waves is given, thus enabling the student to appreciate more readily subsequent sections.

4.2 An electromotive force can be regarded as the force necessary to move an electron from one atom to another in much the same way as a mechanical force is the force necessary to move an object from one place to another. This movement of electrons, of course, constitutes an electric current. It seems impossible to imagine either force being transmitted at an infinite velocity.

4.3 As a familiar example, an engine applies the mechanical force necessary to move a long line of railway trucks. When the force is applied the engine starts to move, the truck coupled to the engine starts to move a little later followed by the next truck, and so on down the line. The truck at the rear moves last. This is responsible for the series of "clatters" which are transmitted down a train when it starts to move. In other words, the engine transmits a mechanical force down the line of trucks to move them and, as the last truck moves some time after the engine, the force is transmitted at a finite velocity, this velocity being the length of the line of trucks divided by the time interval between the movement of the engine and the movement of the last truck.

4.4 To illustrate the electrical case, a direct current example is used first.

Fig. 3 represents an extremely long line having a length of, say, one million miles.

FIG. 3. PROPAGATION OF ELECTRICAL ENERGY - DIRECT CURRENT CASE.

At the precise instant of closing the switch, the electrons of only those atoms near the switch move. It takes some time for the electrons of atoms remote from the switch to move because the electromotive force is transmitted down the line at a finite velocity, just as the mechanical force is transmitted down the line of trucks. This means that ammeter 1 reads as soon as the switch is closed, ammeter 2 some time later, and ammeter 3 some time later still.

4.5 When alternating voltages are transmitted a similar behaviour is experienced.

/Fig.

Fig. 4 shows what happens in the alternating current case. In Fig. 4a, assume that the switch has been closed just at the instant the generator develops its maximum positive voltage. Under the influence of this voltage, a number of electrons starts to move in the direction indicated only at the end of the line near the switch. The number of electrons moving depends on the maximum voltage developed by the generator and the impedance which the line offers to that voltage.

4.6 One quarter of a cycle in time later, the generator voltage falls to zero (Fig. 4b), so that no electrons are moving along the line near the switch. During this time, however, the original positive maximum has had time to travel some distance along the line, for example, nine miles, so that the maximum number of electrons and, therefore, maximum current, flows at that point.

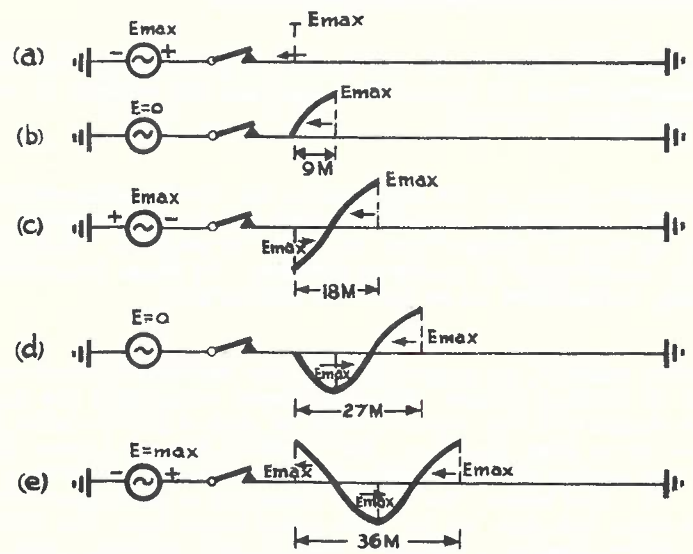

FIG. 4. PROPAGATION OF ELECTRICAL ENERGY - ALTERNATING CURRENT CASE.

4.7 In Fig. 4c, 4d and 4e, the position for successive quarter cycle intervals of time is indicated. It is noted from Fig. 4e that the current is flowing in opposite directions in different sections of the line at the same instant. Thus, in the first nine mile section, the direction (electronically) is from right to left, in the next 18 mile section the direction is from left to right, and in the next nine mile section the direction is from right to left again. The position may be more easily seen by considering the train analogy. The engine starts to move the trucks backwards but, before the energy impulse has travelled right down the line, the engine has reversed its direction so that the trucks remote from the engine are moving backwards whilst those near the engine are moving forwards. If the engine maintains this backward and forward movement, the trucks along certain sections of the train move backwards and, at the same time, the trucks along the remaining sections move forwards.

4.8 A similar position exists on a long transmission line. The finite velocity of propagation means that the generator voltage reverses in direction before a generated positive half cycle has reached the remote end, so that electrons along some sections of the line are moving in one direction whilst electrons in the remaining sections are moving in the opposite direction. This produces the current flowing in different directions along different sections of the line simultaneously. As a consequence, in many alternating current circuits the electrons constituting the current flow do not make a complete journey around the circuit as in the steady direct current case, but merely oscillate backwards and forwards about their normal position.

4.9 Thus, as the generator develops its successive cycles of alternating voltage, these cycles are applied to the line and travel along it in exactly the same way as successive

/cycles

cycles of sound travel through a conducting medium. The fact that a voltage and its effect, which is a current flow, travel along the line at a finite velocity means that, in alternating current transmission over long circuits, the voltage and current travel along the line in the form of waves. Likewise, the energy stored in the electric and magnetic fields produced by the voltage and current travels in the form of waves, the combination of the two fields being termed an electromagnetic wave.

4.10 The length of line occupied by one complete cycle of alternating current or voltage at the frequency applied is termed the "Wavelength" at that frequency, and the length of any line can be expressed in wavelengths at that frequency. A little consideration shows that, assuming the line of Fig. 4 is purely resistive, doubling the applied frequency halves the wavelength.

5. REFLECTION.

5.1 All energy transmitted in the form of waves undergoes reflection if the conducting path is not infinitely long and uniform. A familiar example is the reflection of sound waves producing an echo. Here, a source of sound transmits acoustical energy in the form of waves, and these waves, in travelling out from the source, encounter an irregularity in the conducting medium. For example, these waves are transmitted through the air for some distance and then encounter a brick wall. Some of the energy is transmitted onwards through the wall, the remainder, being reflected, travels back to the sound source again. Thus, at the sound source, an echo is heard. The production of this echo means that not only must the original sound wave be present there but also that reflected. The resultant sound heard at the source at any instant, therefore, is that due to the vector sum of the transmitted and reflected sound waves.

5.2 The electromagnetic waves transmitted over a telephone line likewise undergo reflection if the conducting path for these waves is not infinitely long and uniform. If reflection takes place due to an impedance irregularity, the current and voltage waves are reflected from the irregularity back to the sending end. The resultant voltage and current at the sending end, therefore, is the vector sum of the transmitted and reflected waves at that end. As the length of the line in wavelengths to the irregularity causing reflection is different for different frequencies, the phasing of the reflected current and voltage waves relative to those applied at the sending end by the source of supply varies with frequency. Thus, at some frequencies, the two voltages are in phase and the two currents largely out of phase, and vice versa. The former effect of increased voltage and decreased current is equivalent to an increase in line impedance. This means that an impedance irregularity causes the impedance of the line to vary with frequency in the manner shown in Fig. 5.

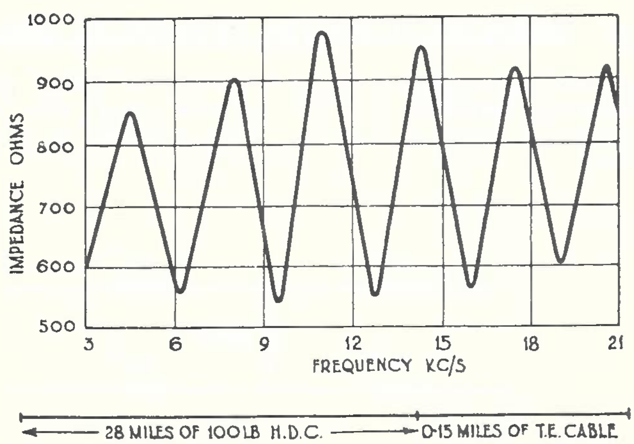

FIG. 5. ACTUAL IMPEDANCE VERSUS FREQUENCY CURVE.

6. CHARACTERISTIC IMPEDANCE.

6.1 From what has been dealt with above, it is found that one undesirable effect of reflection is to cause the impedance of a line to vary widely with frequency. In Fig. 5, the line impedance varies between about 600 ohms and about 900 ohms at 1.5 kc/s intervals. It is desirable, therefore, to prevent reflection, and this can be achieved by making the line infinitely long and uniform.

6.2 Uniformity is secured by employing the same gauge of wire, uniform wire spacings, and so on, throughout the entire length of a line. Where this cannot be done, as in Fig. 5, other arrangements to be discussed later are employed. It is not possible, of course, to make any line infinitely long. However, from a study of the behaviour of an infinitely long line, it is possible to arrive at a method by which any finite length of line can be made to behave electrically as though it is infinitely long, even though that finite length may be only a few yards. To arrive at this method, it is necessary to know the impedance of an infinite length of any type of line to to dealt with, this impedance being termed the "Characteristic Impedance" of the line and designated Z_0.

(Note: The reason given above is only one reason for endeavouring to make lines behave as though infinitely long. Another reason is that, if a line is terminated in its equivalent characteristic impedance, the impedance of the line, viewed from all points along the line and in either direction from those points, is always the characteristic impedance. This makes for simplicity of measurements, as the ratio of measured voltage and current at any point is always the same.)

6.3 The fact that an infinitely long line has a finite impedance is perhaps most easily demonstrated by considering a short length of line, for example, one mile, and then gradually lengthening that line. This is done in the following treatment.

6.4 A short length of telephone line, say one mile, can be represented at some frequency, say 1,000 c/s, by the network shown in Fig. 6.

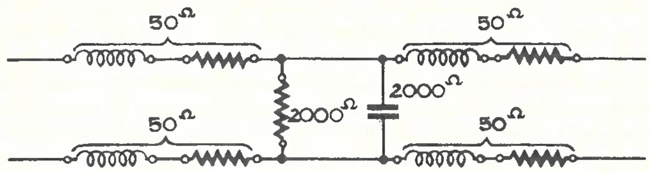

FIG. 6. NETWORK FOR ILLUSTRATING CHARACTERISTIC IMPEDANCE OF LINES.

For simplicity, the values chosen do not represent an actual case. Neglecting phase and considering magnitude only for the time being, Fig. 6 is replaced by the network of resistances shown in Fig. 7. This is called an H network because of its configuration.

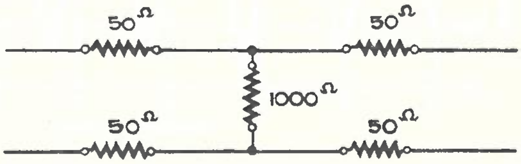

FIG. 7. H SECTION DERIVED FROM FIG. 6.

Fig. 8 is obtained by combining the two series elements on each side of the shunt element. This is called a T network. In regard to magnitude, this T network behaves exactly as the one mile of line did at 1,000 c/s and, therefore, is an equivalent network or circuit at that frequency.

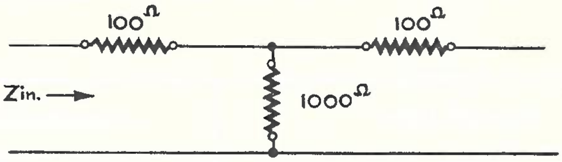

FIG. 8. T SECTION DERIVED FROM FIG. 7.

6.5 If the impedance of the line is measured from one end with the distant end open, the impedance is 1,100 ohms. This impedance is, in future, referred to as the input impedance and designated Z_{in}.

Under similar conditions, a line 2 miles long, shown in Fig. 9, has an input impedance of 100 ohms + (1,000 ohms in parallel with 1,200 ohms), that is, 100 ohms + 545 ohms = 645 ohms.

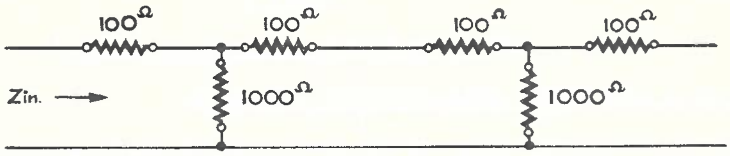

FIG. 9. TWO MILES OF LINE.

By similar reasoning -

A line of 3 miles has an input impedance of 527 ohms.
A line of 4 miles has an input impedance of 485 ohms.
A line of 5 miles has an input impedance of 468 ohms.
A line of 6 miles has an input impedance of 462 ohms.
A line of 7 miles has an input impedance of 459 ohms.
A line of 8 miles or longer has an input impedance of 458 ohms.

An examination of these figures shows that, as the length of a line increases, its impedance decreases rapidly, at first, but later with diminishing rapidity until a point is reached beyond which the impedance of the line does not change as its length is increased. This is shown in Fig. 10, which is a graph of line impedance versus line length using the figures quoted above. In the case chosen, the impedance did not change after 8 miles, so that, if the line is made infinitely long, its impedance is that of 8 miles, that is, 458 ohms.

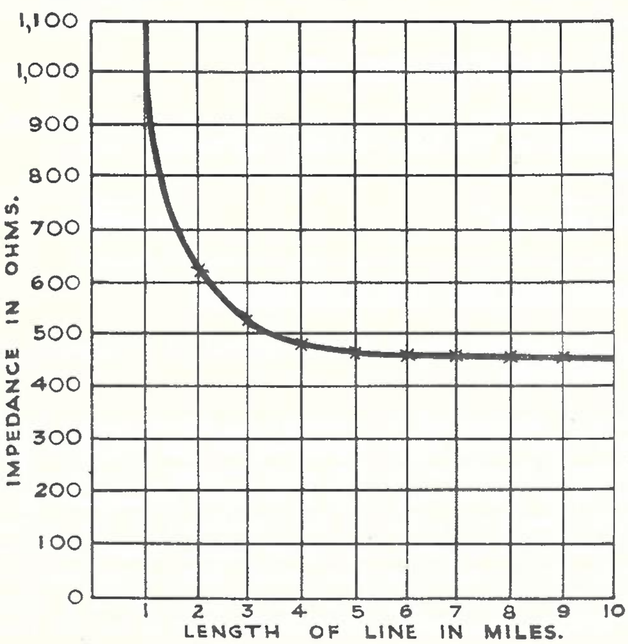

FIG. 10. VARIATION OF IMPEDANCE WITH LENGTH.

6.6 From this it is seen that an infinitely long line has a finite impedance, this impedance being termed the Characteristic Impedance of the line.

6.7 As the shunt and series elements of the initial T section considered contain reactive components, the characteristic impedance of a line varies with frequency. Further, as these shunt and series elements

/depend

depend on the spacing between the conductors, the dielectric constant of the insulating medium and the gauge of wire used for the line conductors, the characteristic impedances of different types of lines have different values. Table 2 gives a number of characteristic impedances for different types of lines calculated at 800 cycles.

Type of Line	Characteristic Impedance
100 lb. Copper Aerial	804 $\backslash\underline{20.3^{\circ}}$ ohms.
200 lb. Copper Aerial	687 $\backslash\underline{12.1^{\circ}}$ ohms.
300 lb. Copper Aerial	647 $\backslash\underline{8.4^{\circ}}$ ohms.
10 lb. S.Q. Cable	699 $\backslash\underline{44.2^{\circ}}$ ohms.
20 lb. S.Q. Cable	516 $\backslash\underline{43.3^{\circ}}$ ohms.
40 lb. S.Q. Cable	366 $\backslash\underline{41.7^{\circ}}$ ohms.

TABLE 2.

(From The Telephone Handbook, Poole.)

7. TERMINATION OF A LINE IN ITS CHARACTERISTIC IMPEDANCE.

7.1 If a line is infinitely long and any finite length (such as one mile) is cut off, the line is still infinitely long, as any finite value subtracted from infinity leaves the infinite line unchanged in length. As an example, one mile of line cut from a line two miles long reduces the length of the line by 50%, or halves the length of the line, but one mile cut from a line one hundred million miles long makes no practical difference to the length of the line.

Fig. 11 shows an infinitely long line. The input impedance of this line is its characteristic impedance (458 ohms, using the type of line worked on previously).

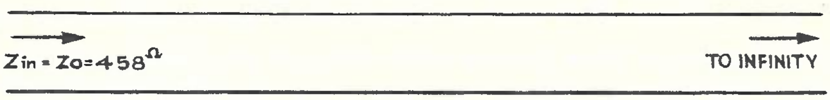

FIG. 11. INFINITELY LONG LINE.

Fig. 12 shows this infinite line with one mile cut off. The input impedance is still its characteristic impedance, as the line is still infinitely long.

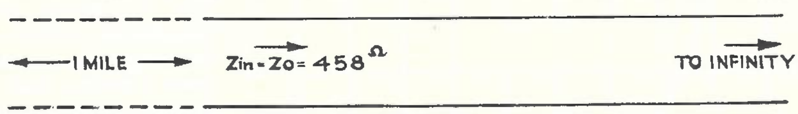

FIG. 12. INFINITELY LONG LINE MINUS ONE MILE.

Fig. 13 shows the one-mile length cut-off drawn as its equivalent T network, the input impedance of which is 1,100 ohms.

Fig. 14 shows this one-mile length terminated in a single impedance equal to the characteristic impedance of the line. The input impedance of the one-mile length is now 458 ohms, the characteristic impedance of the line because, in Fig. 11, the one-mile length was terminated in an infinitely long line whose impedance was 458 ohms, whilst in Fig. 14 a single impedance equal to the impedance of the infinitely long line replaces the infinitely long line. In other words, there is no way to determine by measurements within the one-mile section under discussion whether the line was infinitely long beyond the one-mile point or was terminated in a single impedance equal to its characteristic impedance. The truth of this can be proved

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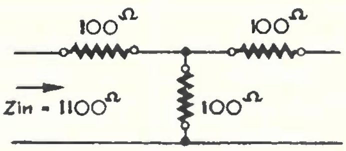
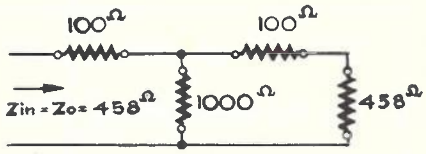
by calculating the input impedance of Fig. 14, which is 100 ohms + (1,000 ohms in parallel with 558 ohms), that is, 100 ohms + 358 ohms = 458 ohms.



ONE MILE OF LINE.

FIG. 13.



ONE MILE TERMINATED IN Z_0.

FIG. 14.

7.2 Since the input impedance of a line terminated in its characteristic impedance is equal to that of an infinite line, terminating any length in its characteristic impedance produces an input impedance equal to the characteristic impedance at the other end. In other words, the input impedance of a line terminated in its characteristic impedance is the characteristic impedance of the line, and the line, although physically short, behaves electrically as though infinitely long.

7.3 The characteristic impedance of a line is one of its Secondary Constants, others of which are dealt with later.

8. WAVELENGTH AND WAVELENGTH CONSTANT.

8.1 The length of line occupied by one complete cycle of current or voltage, or the length of line through which the current or voltage is shifted in phase through an angle of 360°, is termed the "wavelength" and is signified by the Greek letter (lambda).

8.2 This distance is shown in Fig. 4e by the distance travelled by the original positive maximum in the time taken for the generator to develop one cycle, that is, a distance of 36 miles. Under these conditions, the phase shift will be $\frac{360}{36} = 10^\circ$ per mile. This value of wavelength will apply at a particular frequency only, for example, 1,000 c/s. Increasing the frequency to 2,000 c/s will cause an increase in phase shift to, say, 15° per mile, with a corresponding wavelength of $\frac{360^\circ}{15^\circ} = 24$ miles. Thus, as the frequency rises, the phase shift increases, resulting in a decrease in the wavelength.

8.3 As mentioned above, another way of regarding the wavelength is to think of it as the length of line through which the voltage or current shifts in phase an angle of 360°. In Fig. 4, the current shifts through 360° in 36 miles, or 10° in one mile. The angle through which the voltage or current shifts per unit length, generally one mile, is called the Wavelength Constant or Phase Constant. This constant may be expressed either in degrees or radians, for example, $10^\circ = 0.1745$ radian.

8.4 Table 3 shows the Wavelength Constants, Wavelengths and Velocities of Propagation for different types of lines calculated at 800 c/s.

/Table 3.

Type of Line	Wavelength	Wavelength Constant	Velocity of Propagation
100 lb. Copper Aerial	208 Miles	$1^{\circ}44'$ or 0.030 Radian	165,000 miles per second
200 lb. Copper Aerial	220 Miles	$1^{\circ}39'$ or 0.0288 Radian	173,700 miles per second
10 lb. S.Q. Cable	35 Miles	$10^{\circ}21'$ or 0.1807 Radian	27,700 miles per second
20 lb. S.Q. Cable	50.5 Miles	$7^{\circ}8'$ or 0.1245 Radian	40,000 miles per second
40 lb. S.Q. Cable	69.5 Miles	$5^{\circ}11'$ or 0.0903 Radian	55,300 miles per second

TABLE 3.

(From The Telephone Handbook, Poole.)

8.5 As shown earlier, increasing the frequency of the voltage applied to any line increases the phase shift (and, therefore, the phase constant), decreases the wavelength and increases the velocity of propagation. Table 3, therefore, applied at only one frequency - 800 c/s.

9. VELOCITY OF PROPAGATION.

9.1 By purely mathematical reasoning, Maxwell, in 1869, proved that electromagnetic waves are identical with light waves and are, in fact, light waves of a frequency too low to affect the eye. Electromagnetic waves, therefore, travel at the same maximum velocity as light and, as with light, this velocity is different in different mediums.

9.2 The velocity at which these electromagnetic waves travel along a transmission line is called the "velocity of propagation."

9.3 The velocity of light is 186,000 miles per second in a vacuum, slightly lower in air and much lower still through water. The maximum velocity of propagation of electromagnetic waves along a transmission line is 186,000 miles per second over a purely resistive circuit, that is, one with no capacitance or inductance in its make-up. Over actual circuits, however, the velocity is lower than 186,000 miles per second, because such circuits inevitably contain series inductance and shunt capacitance, the velocity decreasing as these increase.

9.4 This may be seen by considering the effect of shunt capacitance on a transmission line to which a direct current voltage is applied, as shown in Fig. 15.

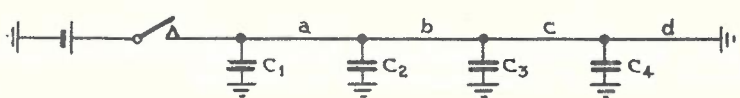


FIG. 15. EFFECT OF LINE CAPACITY FOR DIRECT CURRENT CASE.

Capacitances C_1, C_2, C_3 and C_4 represent the wire to earth capacity distributed along the line. Before any current can flow in section "a" of the line, capacity C_1 must become charged and, as this takes time, the current flow into section "a" is slightly delayed. Similarly, no current flows in section "b" until capacity C_2 has been charged, and so on. Thus, on the switch being closed, and with shunt capacity present, the electrons in those parts of the line remote from the battery move much later than they normally would if no shunt capacity were present. In other words, the effect of shunt capacity is to slow down the velocity of propagation. The effect of series inductance is to slow down further the velocity of propagation by limiting the charging rate into the shunt capacities below that normally due to the line resistance alone. A similar effect on the velocity of propagation is observed, of course, if the direct current supply were replaced by an alternating current supply.

9.5 As shown earlier, the wave in passing along the line progresses at a velocity of one wavelength in the time taken for the generator voltage to pass through one complete cycle, that is -

$$\text{Velocity of Propagation} = \frac{\text{Wavelength}}{\frac{1}{\text{Frequency}}} = \lambda f$$

$$\therefore \text{Wavelength} = \frac{\text{Velocity}}{\text{Frequency}}$$

Assuming that at 1,000 c/s the wavelength is 36 miles, the time taken to generate one cycle is 1/1,000 second. The velocity of propagation at 1,000 c/s, therefore, equals 36,000 miles per second. Reverting to the previous example, the wavelength at 2,000 c/s was found to be 24 miles. The velocity of propagation at 2,000 c/s, therefore, equals $\frac{24}{1/2,000}$ = 48,000 miles per second.

From this, it can be seen that different frequencies travel with different velocities of propagation.

10. TRANSMISSION MEASURING UNITS - THE DECIBEL AND NEPER.

10.1 Any method used for calculating the attenuation or gain produced by the large number of items of equipment through which communication signals have to pass should not be too involved. For example, when two subscribers connected to exchanges in different capital cities are conversing via a carrier channel, the signals pass through many items of equipment, each of which affects the amplitude of the signal and causes it to either increase or decrease.

10.2 The simplicity of the method used in practice is shown by the following case -

Fig. 16 shows three items of equipment connected in cascade.

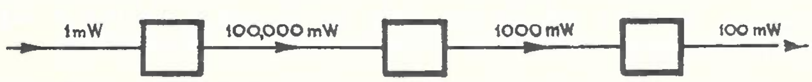


FIG. 16. CIRCUIT ELEMENTS ALTERING POWER LEVEL.

One milliwatt of power is applied to the input of the first item, an amplifier, the output power from which is 100,000 milliwatts. This 100,000 milliwatts is then applied to the second item which reduces the 100,000 milliwatts to 1,000 milliwatts, and this 1,000 milliwatts is then applied to the third item which reduces the 1,000 milliwatts to 100 milliwatts. The first item causes signals to be increased in power 100,000 fold or 10^5, the second item causes signals to

/be

be reduced in power to 1/100th or 10^{-2} of their input value, and the third item causes signals to reduce the power output to 1/10th or 10^{-1} of the input power. Thus, for any value of input power, the output power from the whole system is calculated by multiplying the input power to the first item by 100,000 dividing the result by 100 and then dividing this result by 10. For example, if the input power is 75 milliwatts, the output power will be -

$$75 \text{ mW} \times 100{,}000 \times \frac{1}{100} \times \frac{1}{10} = 7{,}500 \text{ mW}.$$

This calculation is simple because of the values chosen. The values encountered in actual practice do not always produce such a simple calculation - the calculations are usually a fairly involved multiplication. Multiplication is simplified by using logarithms, which involves only the addition of the logarithms and the extraction of the antilog of the result. Thus, in the case taken above, the output power with 75 mW input power is -

$$\begin{aligned} & 75 \text{ mW} \times \text{antilog } (\log_{10} 100{,}000 + \log_{10} \tfrac{1}{100} + \log_{10} \tfrac{1}{10}) \\ &= 75 \text{ mW} \times \text{antilog } (5 + -2 + -1) \\ &= 75 \text{ mW} \times \text{antilog } 2 \\ &= 75 \text{ mW} \times 100 \\ &= 7{,}500 \text{ mW}. \end{aligned}$$

10.3 In practice, the factor by which the power is altered by an item of equipment is expressed as the logarithm of that factor. For example, instead of regarding the first item, the amplifier, as having an amplification of 100,000, it is regarded as having an amplification of 5, which is the common log (that is, the log with 10 as a base) of 100,000. Similarly, the second item produces a reduction of 2, which is the common log of 100, and the third item produces a reduction of 1, which is the common log of 10. The name given to the common log of the factor by which the power is altered by an item of equipment is the Bel, so that the amplifier has an amplification or gain of 5 bels, the second item a loss of 2 bels and the third item a loss of 1 bel.

10.4 From this, it may be seen that, to express the gain or loss in bels produced by an item of equipment, the following formula is applied -

$$\text{bels} = \log_{10} \frac{P_1}{P_2}$$

where P_1 and P_2 are the powers involved. Generally, the larger power appears in the numerator.

10.5 In practice, the bel is too large for most purposes, and the decibel (abbreviated "db" and equal to 1/10th of a bel)is used. Thus, as there are 10 db in 1 bel -

$$\text{db} = 10 \log_{10} \frac{P_1}{P_2}$$

10.6 It should be understood that the bel and decibel have no physical significance as have the ampere, the volt and the ohm. The bel is merely the log of the ratio of two powers, this being extracted in order to simplify calculations.

10.7 Whilst the decibel is fundamentally a unit of power ratio, this ratio can be expressed in simplified form from current and voltage ratios. The power input to the first item of equipment is -

$$I_{in}^2 Z_{in} \text{ or } \frac{E_{in}^2}{Z_{in}}$$

and the power output -

$$I_{out}^2 Z_{out} \text{ or } \frac{E_{out}^2}{Z_{out}}$$

/Here

Here E, I and Z are the input and output voltages, currents and impedances, respectively, as designated by the subscripts. In general, communication circuits are matched, that is, $Z_{in} = Z_{out}$, so that the ratios of the two powers reduce to the ratio of the squares of the two currents or the two voltages as follows -

$$P_1 = I_{in}^2 Z_{in}$$

$$P_2 = I_{out}^2 Z_{out}$$

$$\therefore \text{db} = 10 \log_{10} \frac{I_{in}^2 Z_{in}}{I_{out}^2 Z_{out}}$$

As $Z_{in} = Z_{out}$

$$\text{then db} = 10 \log_{10} \frac{I_{in}^2}{I_{out}^2}$$

$$= 10 \log_{10} \left(\frac{I_{in}}{I_{out}}\right)^2$$

$$= 20 \log_{10} \frac{I_{in}}{I_{out}}$$

$$\text{Similarly db} = 20 \log_{10} \frac{E_{in}}{E_{out}}$$

$$\text{thus db} = 10 \log_{10} \frac{P_1}{P_2} = 20 \log_{10} \frac{I_1}{I_2} = 20 \log_{10} \frac{E_1}{E_2}$$

It is important to notice that the current or voltage ratios are used only for matched impedances. When this does not obtain, the power ratio must be used, that is -

$$\text{db} = 10 \log_{10} \frac{I_1^2 Z_1}{I_2^2 Z_2} = 10 \log_{10} \frac{\frac{E_1^2}{Z_1}}{\frac{E_2^2}{Z_2}} = 10 \log_{10} \frac{E_1^2 Z_2}{E_2^2 Z_1}$$

10.8 When referred to some arbitrary amount of power, the decibel is used as a unit of absolute value of power. For example, in telephone testing, the arbitrary amount of power (termed a zero or reference level) is fixed at 1 milliwatt, and other amounts of power are referred to this level. Thus, a power level of +30 db referred to 1 milliwatt would be a power 1,000 times the standard or 1 watt. Similarly, -30 db referred to 1 milliwatt would be a power 1/1,000 times the zero level or 1 microwatt. For testing in radio systems, including telephone lines feeding them, a zero level of 6 milliwatts has been adopted as standard.

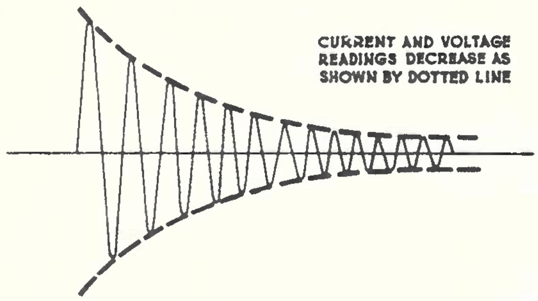
10.9 Another transmission unit, also logarithmic in its basis, is used on the Continent. This unit is the neper and, instead of being a power ratio, it is a current ratio. Also, instead of using the common logarithm of the current ratio, logarithms having a base e are used. e = 2.71828, and is highly important in mathematics. This subject is not concerned with the derivation of e or the reason for its use as a base in the neper. It is sufficient merely to know that such a unit exists, as the decibel is used exclusively in Australia. Thus -

$$\text{nepers} = \log_e \frac{I_1}{I_2}$$

/11.

11. ATTENUATION AND ATTENUATION CONSTANT.

11.1 As the voltage and current waves progress along the line, their amplitudes are reduced. Fig. 17 shows how the voltage and current readings taken along a line at a particular instant decrease in amplitude due to the series and shunt losses, respectively.



ATTENUATION AND PHASE SHIFT ALONG A LINE.

FIG. 17.

11.2 An equal proportion (not an equal amount) of the voltage applied to the line is dropped across the series impedance of each mile of line. Similarly, an equal proportion of the input current is drained away by the shunt impedance of each mile of line. For example, if the input current to a line is 10 mA and the leakance present in each mile of line reduces the current by half, then the current at the end of the first mile will be 5 mA, that at the end of the second mile 2.5 mA, that at the end of the third mile 1.25 mA, and so on.

11.3 The gradual reduction in the amplitude of the voltage and current, and therefore of the power, as the wave progresses along the line is called attenuation. The attenuation per unit length of a line is expressed in either decibels or nepers, and is called the Attenuation Constant.

11.4 Typical values of attenuation for different types of lines, calculated at 800 c/s, are shown in Table 4.

Type of Line	Attenuation in Decibel per Mile
200 lb. Copper	0.061
100 lb. Copper	0.106
40 lb. S.Q. Cable	0.760
20 lb. S.Q. Cable	1.01
10 lb. S.Q. Cable	1.56

TABLE 4.

(From Telephony, Vol. I, Herbert & Proctor)

11.5 The attenuation and attenuation constant values increase as the frequency rises, because the series inductive reactance increases and the shunt capacitive reactance decreases as the frequency rises. This is indicated in the curves of Fig. 18.

/Fig. 18.

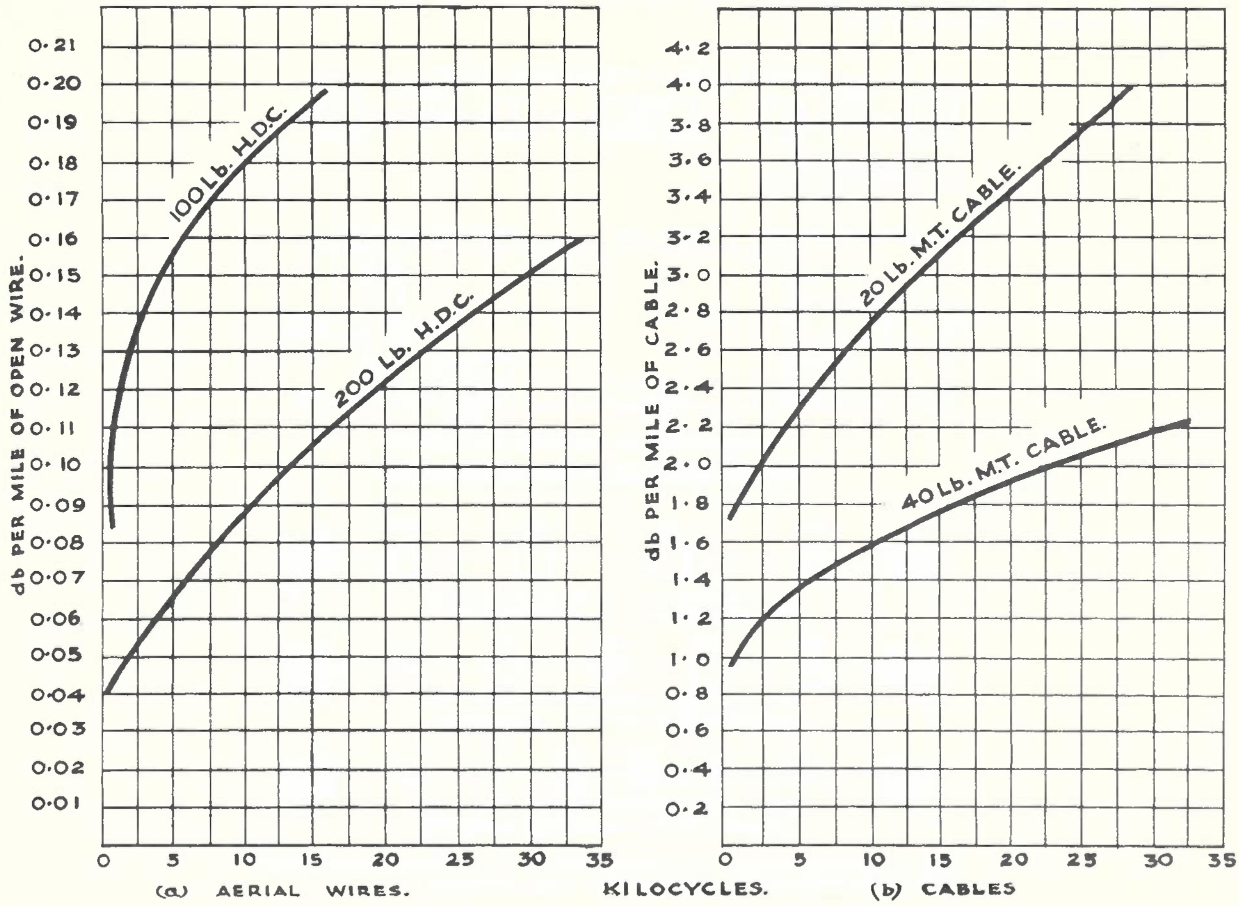


FIG. 18. RELATIONSHIP BETWEEN FREQUENCY AND ATTENUATION.

12. STANDARD GRADE OF OVERALL TRANSMISSION.

12.1 In order to keep transmission losses within reasonable limits and to ensure the same standard of transmission on all classes of calls, whether they are local, intrastate, interstate, and so on, it is necessary to have some standard circuit producing definite losses on which to base the design of all classes of circuits, that is, local exchange networks, junction networks and trunk line networks. This circuit is called a Standard Reference Circuit, and the grade of transmission which it supplies is called the Standard Grade of Overall Transmission.

12.2 Fig. 19 shows, in skeleton form, the Department's "Standard Reference Circuit." This diagram shows that each connection is considered as made up of three units, two of them being the separate local lines and instruments at the ends of the connection. The third unit is the transmission line joining the local line units and consisting of junction circuits, switching exchanges and trunk line transmission aids, for example, repeaters, as required. It is noted that this transmission line is to be equivalent to a 15 db attenuator.

/12.3

12.3 Fig. 19 also gives three instances of typical connections conforming to this standard of overall transmission.

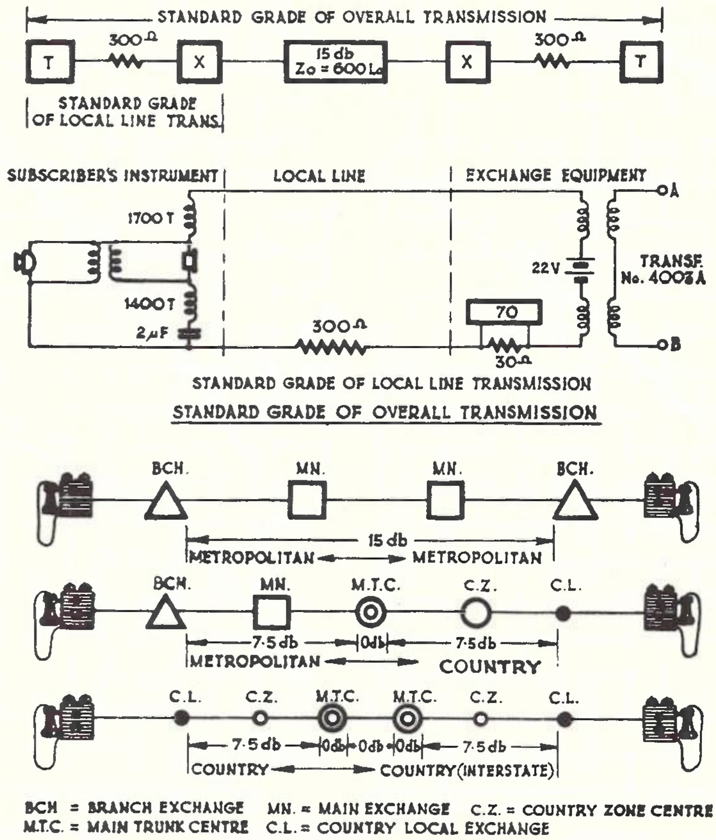


FIG. 19. TYPICAL CONNECTIONS OF STANDARD OVERALL GRADE.

12.4 The telephone system is designed so that any connection between two subscribers provides transmission at least equal to that of this reference circuit. In this circuit, the loss introduced by the apparatus and lines connecting the two telephones is approximately 29 db. The telephones of the reference circuit are of the handset type (Telephone 162). When less efficient telephones are used, the allowable loss in the local lines must be reduced to provide the same grade of transmission. Similarly, more efficient exchange transmission bridges, etc., allow an increase in the loss in the lines.

12.5 The allowable limit of resistance of the subscriber's line with 250/250 ohm battery feed relays and transmitter C.B. No. 1 is 275 ohms whilst, with 50/50 ohm nickel sleeve relays and resistor, barretter or ballast lamp at the exchange with transmitter inset No. 10 in the telephone, the line resistance may be 540 ohms.

12.6 The loss between the telephones is divided up into two portions for the purpose of arriving at the maximum losses to be permitted over each portion -

(a) Subscriber's circuit to exchange, and
(b) Junctions and trunks between any two exchanges.

The loss in case (b) must not exceed 15 db, leaving, in the case of the reference circuit, about 14 db for the sum of the losses from telephone to exchange (sending loss) and from exchange to telephone (receiving loss). The sending loss is approximately 10 db and the receiving loss is approximately 4 db.

/The

The reasons for the difference in the figures quoted for sending and receiving losses are that the receiving loss consists of attenuation only, whereas the sending loss is determined by the attenuation and the direct current in the transmitter.

12.7 The reason for choosing a 29 db overall loss from telephone to telephone in Fig. 19 is that the average electrical power output from the transmitter of a Telephone 162, when actuated by a normal voice, is approximately 200 microwatts. The power necessary at a telephone receiver to produce a good signal is approximately 0.2 microwatt. The ratio of transmitted to received power for good signals from the normal voice, therefore, is about 200 to 0.2 or 1,000 to 1, meaning a permissible loss of about 30 db.

13. PROPAGATION CONSTANT.

13.1 From what has been dealt with, it is apparent that two undesirable effects of a transmission line are, first, to reduce the amplitude of the voltage and current waves as they progress along the line and, secondly, to produce a phase shift, these varying with frequency. An expression containing the attenuation constant and phase constant of a line enables the behaviour of the line to be fully determined, as from these can be calculated the wavelength, the velocity of propagation and the current and voltage at any point along the line. Such an expression is called the "propagation constant," and can be likened to the full specification for an impedance, that is, a value in ohms together with an angle, except that the propagation constant contains the attenuation constant in nepers and the phase constant in radians.

14. USE OF TRANSFORMERS IN SECURING UNIFORMITY.

14.1 It is not always possible to have a line with exactly uniform characteristics right throughout its length. For example, a 200 lb. copper aerial line is used as a trunk line between two trunk centres. On the outskirts of the trunk centres, however, the aerial line usually terminates, the line being carried into the offices concerned in what is called a "trunk entrance cable." As shown earlier in this Paper, the characteristics of cable differ markedly from that of aerial wire.

To secure a uniform line throughout, transformers are connected at the junction of the cable and the aerial wires, the transformer employed being such that, viewed from the cable, the characteristic impedance of the aerial line approximately equals that of the cable whilst, viewed from the aerial line, the characteristic impedance of the cable approximately equals that of the aerial line. In other words, the transformer is used as an impedance transforming device as in Fig. 20.

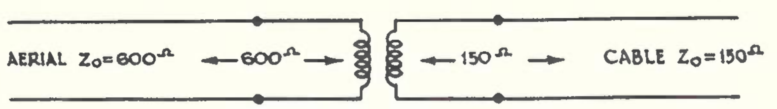


FIG. 20. TRANSFORMER CONNECTED BETWEEN AERIAL LINE AND CABLE.

In Fig. 20, the characteristic impedance of the aerial line has been assumed to be 600 ohms and that of the cable 150 ohms over the range of frequencies to be transmitted over the cable.

/15.

15. TEST QUESTIONS.

1. List the Primary Constants of a transmission line and state, in general terms only, the effects these have on the transmission of a band of frequencies over the line.

2. What is meant by the Characteristic Impedance of a line? What is the effect of terminating a line in its characteristic impedance? Give reasons for so doing.

3. Define -

(a) Attenuation Constant
(b) Wavelength
(c) Wavelength Constant, and
(d) Velocity of Propagation.

4. Five milliwatts of power are sent into a line 8 miles long, and 2.5 milliwatts are received at the distant end. Assuming equal input and output impedances of 600 ohms, calculate -

(a) The attenuation of the line in db, and
(b) The attenuation constant in db.

(Answer : (a) 3 db, (b) 0.375 db.)

5. A transmission line produces a loss of 7 db. If 1 milliwatt is applied to the input of the line, what will be the output power?

(Answer : 0.2 milliwatt.)

6. An amplifier has a gain of 56 db. The input impedance is 600 ohms and the load in the output is 10 ohms. What will be the current in the load when an alternating voltage of 1 volt is applied to the input?

(Answer : 8.15 amperes.)

7. The input impedance of a piece of equipment is 600 ohms and the output impedance is 500 ohms. It is found that when an alternating voltage of 10 volts is applied across the input, a current of 10 milliamperes flows in the output. Calculate the power loss in the equipment in db.

(Answer : 5.2 db.)

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

PRINCIPLES OF CARRIER TELEPHONY AND TELEGRAPHY.

CONTENTS:

1. INTRODUCTION.

1.1 As mentioned previously, one aspect of this subject of Long Line Equipment is concerned with increasing the efficiency of line plant so that a number of messages, either telephonic or telegraphic, can be simultaneously transmitted over any circuit. This increased efficiency is obtained by means of Carrier Telephone and Telegraph Systems, and this Paper gives a broad outline of Carrier Telephony and Telegraphy. The aim is to give here, in general terms only, the functions of the items of equipment required, together with the build-up of a typical system, so that the student can appreciate more readily the matter dealt with in the different Papers of this book.

2. MODULATION AND DEMODULATION.

2.1 In Telephony, intelligible speech involves the transmission from speaker to listener of a frequency band extending from about 200 c/s to about 3,000 c/s. Whilst any speaking voice produces frequencies above 3,000 c/s and below 200 c/s, this band, if "picked off" from the whole range produced, will produce good, intelligible results. Any telephone communication channel, therefore, will have to be capable of passing a band of frequencies about 2,800 c/s wide.

2.2 Fig. 1 shows two telephones, A and B, connected by a line. During speech, the transmitters at A and B may be considered as generating alternating currents which are transmitted over the line. From what has been discussed above, the line need not transmit any frequencies higher than 3,000 c/s insofar as communication between A and B is concerned.

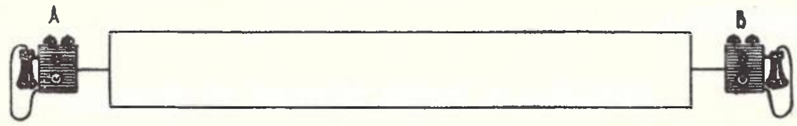


FIG. 1. VOICE FREQUENCY CHANNEL.

2.3 To obtain an additional telephone channel from the existing line between A and B by means of Carrier Telephony, frequency translation is resorted to. Fig. 2 shows the principle. Communication between A and B is carried on in the normal manner, using the voice frequency band extending up to 3,000 c/s. In future drawings, single line circuits will be used for simplicity. It should be remembered that each single line is a two-wire circuit.

FIG. 2. ONE-WAY CARRIER CHANNEL.

2.4 For communication between telephones C and D in the direction C to D, a frequency translating device is used at C which will translate the voice frequency band 200 c/s to 3,000 c/s output from the transmitter at C into a higher frequency band, for example, from 10,200 c/s to 13,000 c/s. This frequency band is passed over the line and, before entering the telephone at D, it is applied to a frequency retranslating device which will bring the frequency band 10,200 c/s to 13,000 c/s down to the original 200 c/s to 3,000 c/s band output from the transmitter at C. This original band is then passed to the telephone at D to actuate the receiver there.

2.5 One process of translating the voice frequency band up into a higher frequency band is called Amplitude Modulation because, to do this, the voice frequency currents or voltages vary or modulate the amplitude of a high frequency current or voltage in much the same way as the speech input to a telephone transmitter varies or modulates the amplitude of the direct current passing through it. There are other methods by which modulation may be achieved, but the method always used in Carrier Telephony is Amplitude Modulation.

2.6 The reverse process of retranslating the high frequency band back to the voice frequency band is termed Demodulation, the translation device being known as the Modulator and the retranslation device as the Demodulator. The high frequency current or voltage whose amplitude is varied is called the Carrier Frequency Current or Voltage (usually abbreviated Carrier) as, when modulated, it appears to "carry" the signal in much the same way as the direct current to a telephone transmitter does when a signal is impressed on it by the transmitter.

2.7 Any telephone channel must be two-way in operation, so that a modulator is necessary at D and a demodulator at C for communication in the direction D to C. Fig. 3 shows the arrangement, the arrows indicating the direction of operation of the modulators and demodulators.

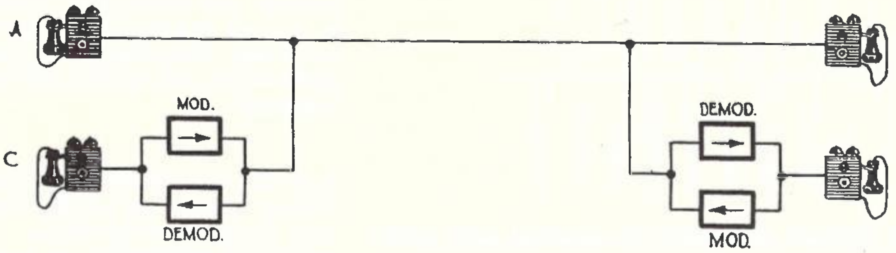
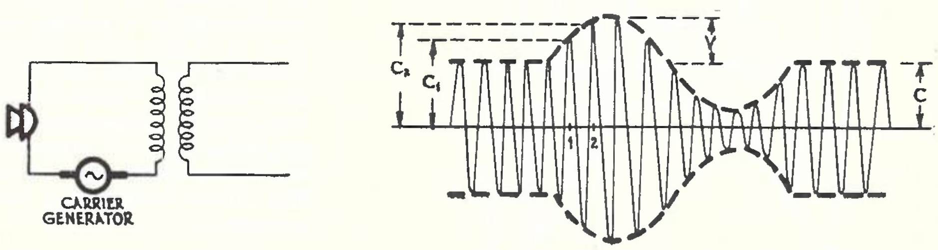


FIG. 3. TWO-WAY CARRIER CHANNEL.

2.8 A general idea of the way the modulator acts as a frequency translation device is desirable before proceeding further. Fig. 4 is a comparison of the operation of two telephone transmitters, one of which modulates the amplitude of a direct current and the other the amplitude of a carrier current, the input sound in each case being a single frequency.

(a) Transmitter Modulating a Direct Current.



(b) Transmitter Modulating a Carrier Current.

FIG. 4. EXAMPLES OF AMPLITUDE MODULATION.

2.9 A mathematical analysis shows that an amplitude modulated carrier of the type shown in Fig. 4b, that is, a carrier frequency current or voltage modulated by a single voice frequency current or voltage, can be regarded as containing three frequencies, namely -

fc, the original carrier frequency,

(fc + fv), a frequency equal to the sum of the carrier and modulating voice frequencies, and

(fc - fv), a frequency equal to the difference between the carrier and modulating voice frequencies.

2.10 As an example, assume that the carrier frequency is 10,000 c/s (10 kc/s) and that the modulating voice frequency is 1,000 c/s (1 kc/s). The resultant modulated carrier contains frequencies of 10 kc/s and 9 kc/s. In other words, the 1 kc/s modulating /frequency

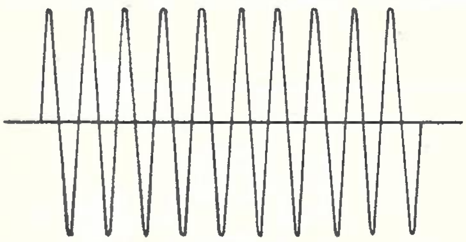
frequency has been translated upwards into two higher frequencies, one lying above the carrier frequency and the other below it. The higher frequency to which the modulating voice frequency has been translated, that is, the sum of the carrier and modulating frequencies, is called the Upper Sideband, whilst the lower frequency, that is, the difference, is called the Lower Sideband.

2.11 The amplitude of the carrier during modulation is the same as that prior to modulation. Thus, amplitude modulation is more conveniently thought of as frequency translation, as the action of varying the amplitude of a carrier apparently does not cause its amplitude to vary but generates new frequencies (the sidebands) which, when added to the constant amplitude carrier, give the appearance of an amplitude modulated carrier. This is shown in Fig. 5 where the three frequencies dealt with above, that is, a 10 kc/s carrier with a constant amplitude together with a frequency of 11 kc/s and one of 9 kc/s, are added vectorially for all instants over one-thousandth of a second. The resultant produced by adding the three components varies in amplitude at 1 cycle during one-thousandth of a second and, therefore at 1,000 c/s. This addition could be carried out on Fig. 5 by the student. There is justification, both mathematically and graphically, for thinking of the amplitude modulated carrier of Fig. 4b as containing three frequencies -

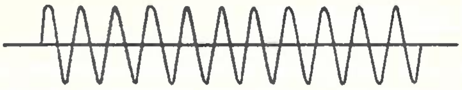
The carrier frequency (10 kc/s),

The upper sideband frequency (11 kc/s), and

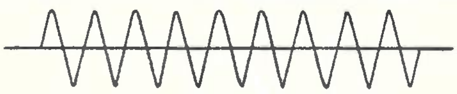
The lower sideband frequency (9 kc/s).



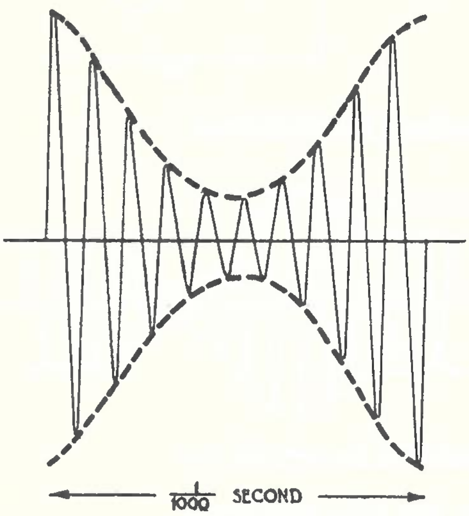
10 kc/s Carrier Frequency of Constant Amplitude.



Upper Sideband 11 kc/s.



Lower Sideband 9 kc/s.



Amplitude Modulated Carrier.

FIG. 5. ADDITION OF CARRIER AND SIDEBANDS.

2.12 In telephony, carriers are modulated by a band of frequencies extending from about 200 c/s to about 3,000 c/s, which would produce the following frequencies in a modulated carrier, assuming that the carrier frequency is again 10 kc/s -

(i) The 10 kc/s carrier at a constant amplitude.

(ii) A frequency range equal to 10 kc/s + (200 to 3,000 c/s) = 10,200 to 13,000 c/s, that is, the upper sideband.

(iii) A frequency range equal to 10 kc/s - (200 to 3,000 c/s) = 7,000 to 9,800 c/s, that is, the lower sideband.

/Modulation

Modulation, therefore, translates the voice frequency band to two higher frequency bands, one immediately above the carrier frequency and the other immediately below it. Either of these sidebands contains a band of frequencies equal in width to the modulating voice frequency band, that is, the voice frequency and the upper and lower sidebands are all 2,800 c/s wide. Further, as the amplitude of any frequency component in the modulating voice frequency band varies, so does the amplitude of the corresponding frequency in the upper and lower sidebands. This means that as the amplitude of the 200 c/s component of the modulating range varies, so does that of the 9,800 c/s and 10,200 c/s components in the sidebands, these components being produced by the 200 c/s component of the modulating range.

2.13 Demodulation is a process exactly similar to modulation in that, if the sideband used for transmission is caused to amplitude modulate a carrier of the same frequency as that applied to the modulator, the process will produce sidebands, one of which will be the original voice frequency band. As an example, using the lower sideband of an amplitude modulated 10 kc/s carrier, that is, 7,000 c/s to 9,800 c/s, the sidebands produced by this frequency range modulating a 10 kc/s carrier will be -

(i) The upper sideband, 10 kc/s + (7,000 c/s to 9,800 c/s) = 17,000 c/s to 19,800 c/s.

(ii) The lower sideband, 10 kc/s - (7,000 c/s to 9,800 c/s) = 200 c/s to 3,000 c/s, that is, the original voice frequency band.

3. FILTERS.

3.1 Whilst the ability to translate the voice frequency range upwards to any desired frequency range is the basis of carrier systems, these systems would be unworkable if it were not possible to separate the different frequency bands at the terminals and direct them to their appropriate circuits. For example, in Fig. 3, the system is successful only if, at the terminals, the voice frequency range can be prevented from gaining access to the carrier system and directed to the telephones at A and B. Similarly, the band of frequencies to which the voice frequencies from C and D have been translated must be directed to the carrier system and denied access to the telephones at A and B. As either sideband contains all of the intelligence to be transmitted, only one sideband needs to be transmitted so that some means of selecting the required sideband and suppressing all other products of modulation is required. Thus, circuits are necessary which will discriminate strongly in favour of a particular band of frequencies and against all others. Such circuits can be designed, and are called Filters.

3.2 The arrangement shown in Fig. 3, therefore, requires filters with the following characteristics for the following purposes -

(i) A filter which will pass with little or no attenuation all frequencies up to 3,000 c/s and reject frequencies above this figure to be connected in each voice frequency terminal, as in Fig. 6. Such a filter is called a Low-Pass (L.P.) Filter, and will allow only voice frequencies to enter the telephones at A and B. A low-pass filter is also necessary in the output of the demodulator to pass the voice frequency range, whilst rejecting all other products of demodulation.

(ii) A filter which will pass with little or no attenuation all frequencies above 3,000 c/s and reject frequencies below this figure to be connected in the carrier circuit at each terminal, as in Fig. 6. Such a filter is called a High-Pass (H.P.) Filter, and will allow only carrier frequencies to enter the carrier terminals. The combined effect of the low-pass and high-pass filters, described above, will be to separate the voice frequencies from the carrier frequencies and direct each range to the appropriate equipment.

/(iii)

(iii) A filter which will pass with little or no attenuation the range of frequencies included in the sideband selected for transmission but will reject all other products of modulation. For example, if the carrier frequency supplied to the modulators is 10 kc/s and the lower sideband is selected for transmission, a filter which will pass a band of frequencies extending from 7,000 to 9,800 c/s is required to immediately follow the modulators, as in Fig. 6. Such a filter is called a Band-Pass (B.P.) Filter.

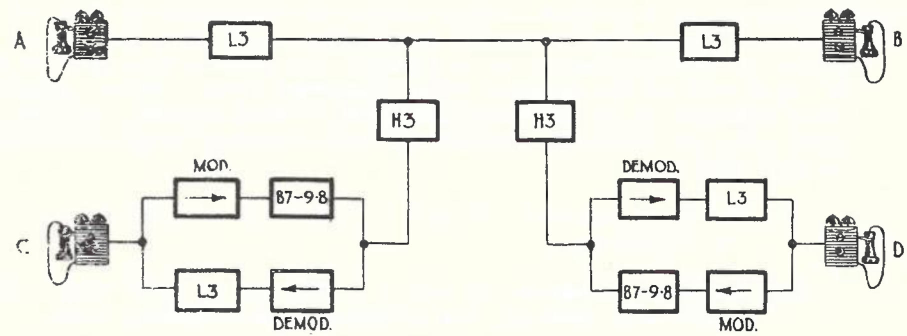


FIG. 6. FILTERS REQUIRED FOR ELEMENTARY CARRIER SYSTEM.

4. OSCILLATORS.

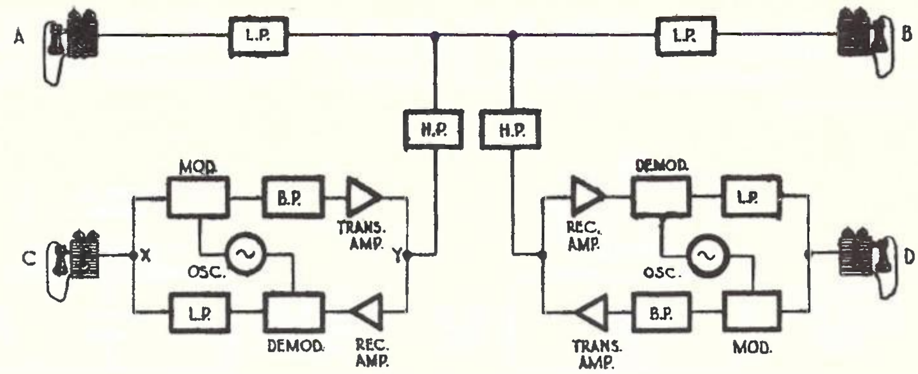
4.1 As mentioned previously, frequency translation is effected by the voice frequencies modulating the amplitude of a carrier frequency. The item of equipment which generates the carrier frequency is called an Oscillator, that is, it produces oscillations of the required carrier frequency. The main requirement of the oscillator is a high degree of frequency stability, that is, the design must be such that there is very little deviation from the carrier frequency it is required to develop.

5. AMPLIFIERS.

5.1 The amount of power in the sidebands to which the voice frequencies are translated by modulation is very small - so small that communication over a long circuit on their power is not satisfactory. It is necessary, therefore, to raise the level of the power in the sideband selected for transmission before applying it to a line, this being done by means of a Transmitting Amplifier. The name "amplifier" suggests its function, that is, to raise, enlarge or amplify. Similarly, at the receiving terminal a receiving amplifier is necessary as, in general, the amplitude of the sideband received from the line is too low to produce an appreciable voice frequency output from the demodulator. The receiving amplifier raises the level of the received sideband before it is applied to the demodulator, so that the voice frequency output from the demodulator is high enough to operate satisfactorily a telephone receiver.

5.2 The carrier system, complete with all of the items discussed up to the present, is shown in Fig. 7.

/Fig. 7.



Mod. = Modulator.
Osc. = Oscillator.
Demod. = Demodulator.
Rec. Amp. = Receiving Amplifier.
L.P. = Low-Pass Filter.
H.P. = High-Pass Filter.
B.P. = Band-Pass Filter.
Trans. Amp. = Transmitting Amplifier.

FIG. 7. MAIN ELEMENTS OF A CARRIER CHANNEL.

6. SEPARATING OPPOSITE DIRECTIONS OF TRANSMISSION.

6.1 An examination of the carrier terminals of Fig. 7 will bring out a source of instability. Take, for example, the C terminal. Voice frequencies from the transmitter at C are applied to the modulator input via junction X. The amplified sideband output from the modulator is applied to junction Y where it divides, some passing to the distant terminal via the H.P. filter and line, and the remainder being applied to the input of the demodulator via the receiving amplifier. Assuming that the carrier frequencies applied to the modulator and demodulator are the same, the demodulator will retranslate that portion of the transmitted sideband applied to it back to the original voice frequency range and apply this range to the junction X. At some voice frequency the phase shift from X through the modulator, modulator band-pass filter, transmitting amplifier, receiving amplifier, demodulator and demodulator low-pass filter at C may be 360° or some integral multiple thereof. At this frequency, the demodulator output will appear at X in phase with the original signal from C, so reinforcing it. This means that the signal from C could be withdrawn and there would still be some input to the modulator. This action would continue indefinitely, producing a sustained "singing" tone in the telephones at C and D, because portion of the demodulator output at the C terminal reaches telephone C via the junction X and portion of the modulator output is sent to the line via the junction Y and the high-pass filter. To prevent the system from "singing," as this action is usually termed, it is necessary at each terminal to isolate the input to the demodulator from the output of the modulator, and the input to the modulator from the output of the demodulator, that is, to keep separate the opposite directions of transmission.

6.2 There are two methods of achieving this separation -

(i) By using hybrid coils, and

(ii) By using what is known as a Four-Wire circuit or its equivalent.

/6.3

6.3 The hybrid coil is an application of the type of circuit used in local battery telephones to reduce sidetone. Fig. 8 shows a comparison of the two circuits, Fig. 8b showing merely the arrangements for isolating the output of the demodulator from the input to the modulator.

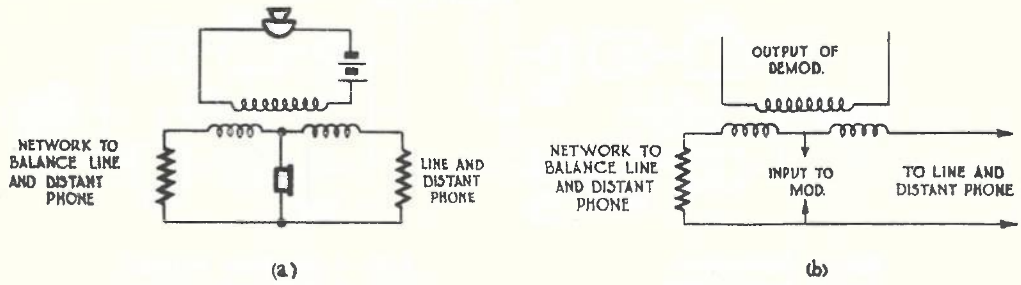


FIG. 8. PRINCIPLE OF HYBRID COIL.

6.4 Fig. 9 shows the actual arrangements of the hybrid coils in a typical system, one terminal only being shown.

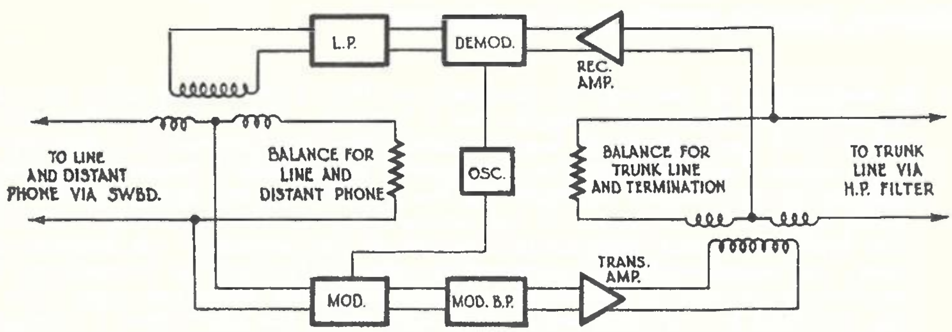


FIG. 9. TERMINAL OF A BALANCED CARRIER CHANNEL.

6.5 Whilst it is not difficult to maintain a balance between a network and a subscriber's line, as subscribers' lines are fairly short, it is very difficult to maintain a balance between a long trunk line and its network, particularly if the trunk line is aerial. The constants of a long aerial trunk line change continually due to climatic changes, so that maintaining a balance is almost impossible. For this reason, this system of separating the opposite directions of transmission is rarely used for carrier systems.

The advantage is that the same sideband of the same carrier frequency can be used for opposite directions of transmission, the separation being effected by the hybrid coils. This system is known as the Balanced System of Carrier Operation.

6.6 To eliminate the hybrid coil and the balance network on the trunk line side of a carrier terminal, four-wire working is used sometimes. Fig. 10 shows the arrangements at one terminal. Separate pairs are used for transmission in opposite directions, hence the name "Four-Wire" Circuit. Again, the same sideband of the same carrier frequency can be used for transmission in opposite directions. This system is extensively used on trunk cable routes.

/Fig. 10.

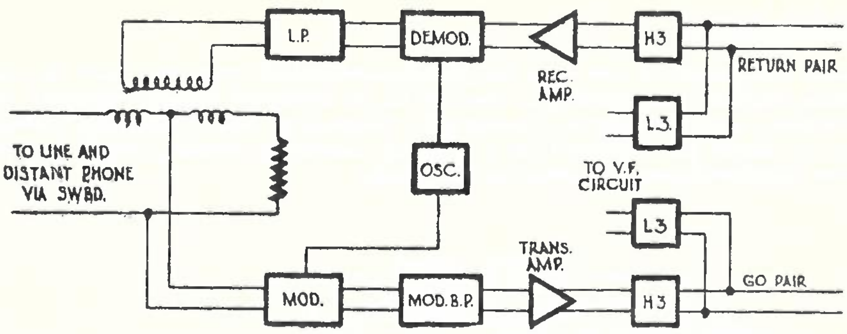


FIG. 10. TERMINAL OF A 4-WIRE CARRIER CHANNEL.

6.7 The equivalent of a four-wire circuit over a single pair of wires is obtained by using the same carrier frequency in opposite directions but transmitting the upper sideband in one direction and the lower sideband in the opposite direction. The band-pass filters preceding the demodulators and succeeding the modulators act as directional filters to separate opposite directions of transmission. Fig. 11 shows the arrangement, the carrier frequency used being 10 kc/s with the upper sideband transmitted in the C to D direction and the lower sideband in the D to C direction.

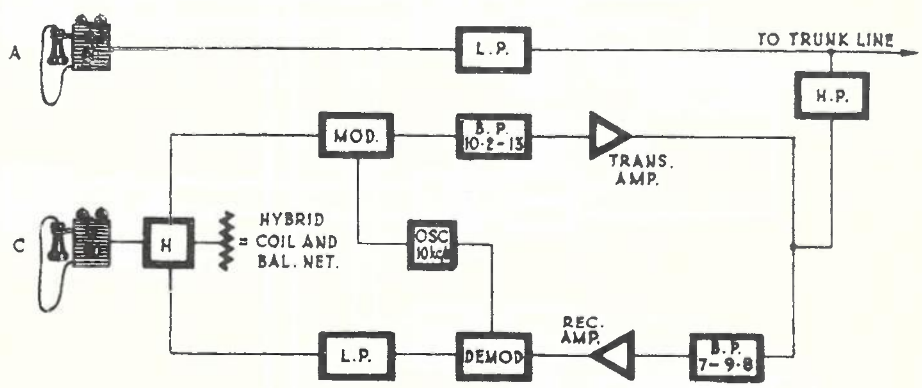


FIG. 11. ELIMINATION OF HYBRID COIL BY FREQUENCY DISCRIMINATION.

6.8 Alternatively, different carrier frequencies can be used in opposite directions of transmission. Here, the upper sideband of a 10 kc/s carrier is used for transmission in the C to D direction, and the upper sideband of a 15 kc/s carrier is used for transmission in the D to C direction.

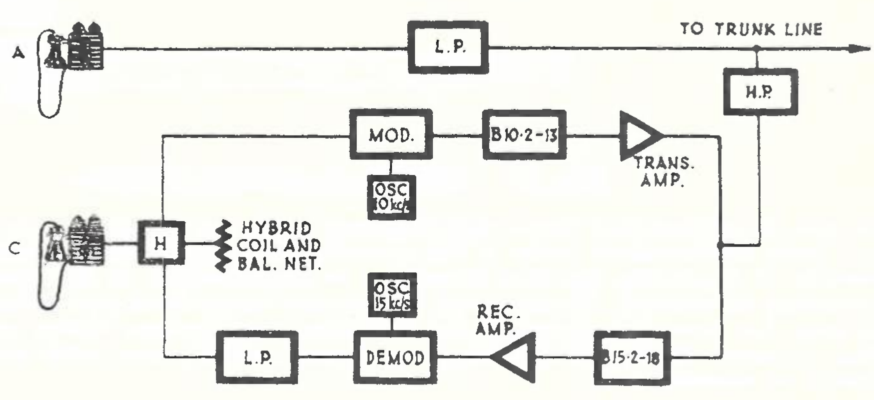


FIG. 12. ELIMINATION OF HYBRID COIL BY FREQUENCY DISCRIMINATION.

7. MULTI-CHANNEL SYSTEMS.

7.1 Carrier telephony is not limited to single-channel systems, that is, systems which enable another communication channel to be obtained from a physical circuit. In Australia, single-channel, 3-channel, 12-channel and 17-channel systems are in use. Fig. 13 shows one terminal of a typical 3-channel system, which is essentially three single-channel systems in parallel.

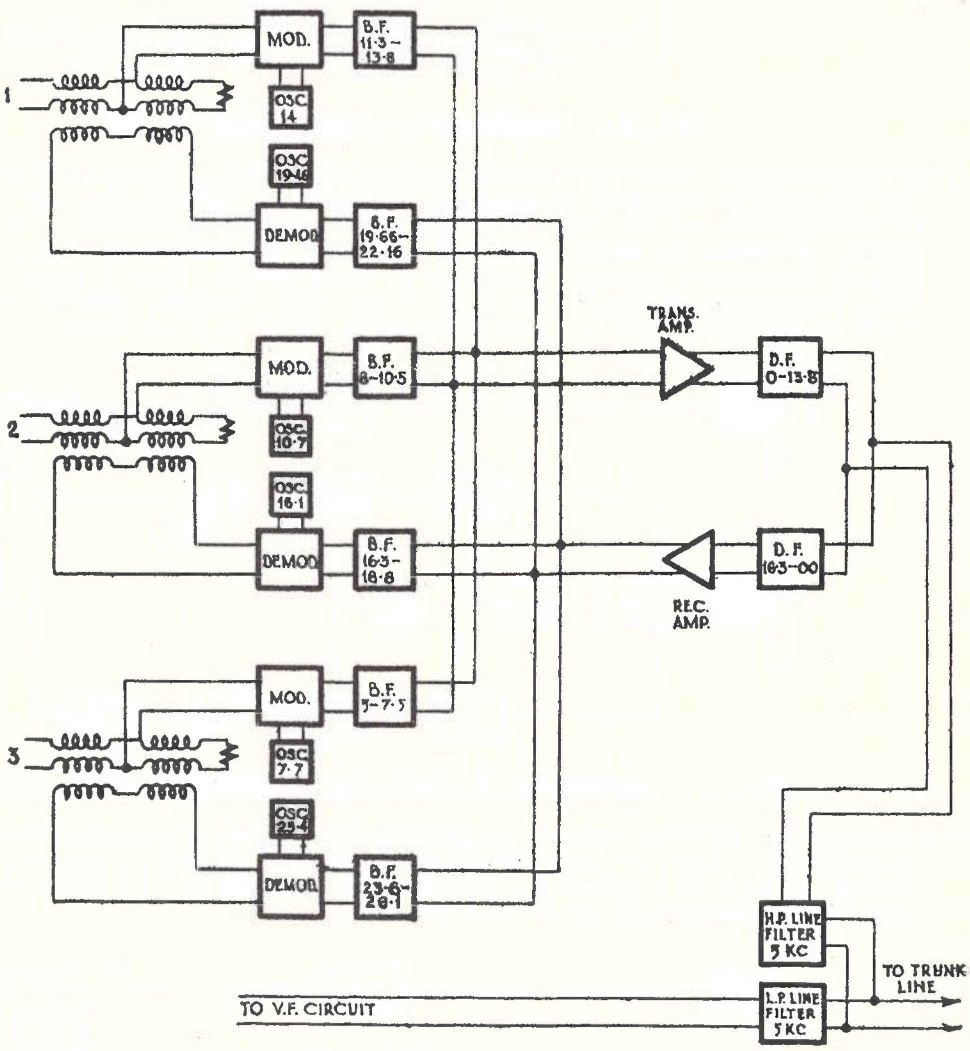


FIG. 13. ELEMENTS OF 3-CHANNEL TERMINAL.

It will be noticed that three high-frequency bands are transmitted in one direction and three low-frequency bands in the opposite direction. This simplifies the problem of separating opposite directions of transmission, as the transmitting and receiving directional filters which do this are merely band-pass filters designed to pass the three high-frequency and the three low-frequency bands, respectively. Fig. 13 shows the frequencies used in a typical system, from which this point should be apparent.

/8.

8. REPEATERS.

8.1 As the voice and sideband frequencies progress along the line, their power levels fall due to attenuation. This loss can be readily made up by amplification. It is not desirable that the signal power level be allowed to fall too low, as the signal level would eventually equal that of the line noise due to inductive disturbances, etc. Amplifiers, therefore, are provided at intervals along the line, such amplifiers being known as Repeaters. A function of a repeater, therefore, is to raise the signal power level on a line before it has fallen to a level comparable with that of the line noise.

8.2 Repeaters are merely amplifiers, and, as amplifiers are unidirectional in operation, two amplifiers are necessary in a repeater, one to amplify the power sent over the line in one direction and the other to amplify the power sent in the opposite direction. Fig. 14 shows the arrangement of a voice frequency repeater.

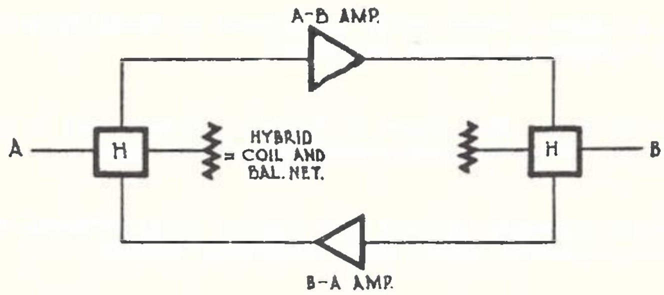


FIG. 14. V.F. REPEATER.

Hybrid coils separate the two directions of transmission for the same reason as in the carrier terminals previously described.

8.3 Where a carrier system is superimposed on a voice frequency circuit two repeaters are necessary, one for the voice frequencies and the other for the carrier frequencies. This arrangement is necessary as the carrier frequencies, being higher, suffer greater attenuation and therefore require more amplification than do the voice frequencies. Fig. 15 shows the arrangement, the H.P. and L.P. filters separating the voice and carrier frequencies, and the directional filters in the carrier repeater separating opposite directions of transmission.

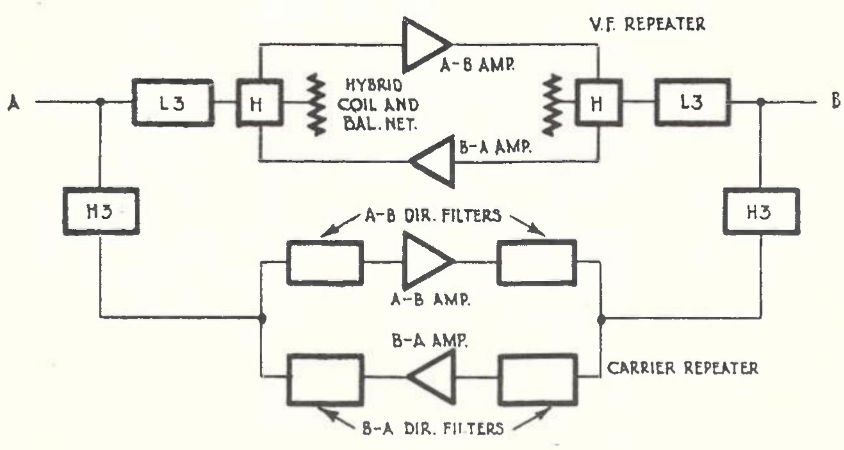


FIG. 15. V.F. AND CARRIER REPEATER.

9. CARRIER TELEGRAPH SYSTEMS.

9.1 The principle of carrier telegraphy is to interrupt an alternating current in much the same way as in a single current Morse Simplex System. By using alternating currents of different frequencies in the voice frequency band, a number of messages can be simultaneously sent over the same voice frequency or carrier channel, suitable filters at the distant terminal directing the signals of different frequency to their appropriate receiving circuits. In the receiving circuits, the voice frequency signal is rectified to produce a direct current suitable for operating a telegraph relay.

10. TEST QUESTIONS.

1. Explain, in general terms only, the purpose of "Modulation." What type of modulation is employed in Carrier Telephony?

2. What are filters? What types of filters are necessary in a single channel carrier system, and define the characteristics of each type?

3. Draw a block schematic circuit of a typical single-channel carrier system and explain the purpose of each item of equipment therein.

4. Explain why it is necessary to separate the opposite directions of transmission at a carrier terminal.

5. Explain how a hybrid coil separates the opposite directions of transmission through a V.F. repeater.

END OF PAPER.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

NETWORKS, ATTENUATORS, FILTERS AND EQUALISERS.

CONTENTS:

1. INTRODUCTION.

1.1 There is an almost infinite number of ways of connecting resistances, inductive reactances and capacitive reactances together to perform different functions, the different circuits so formed being called Networks. From the multiplicity of different possible combinations, a few types are encountered again and again in communication circuits so frequently as to warrant special treatment. It is the purpose of this Paper to study two of the most frequently encountered types and to deal with some of their applications.

1.2 The two types to be studied are called T and π networks from their configuration. These networks may be balanced or unbalanced, as shown in Fig. 1.

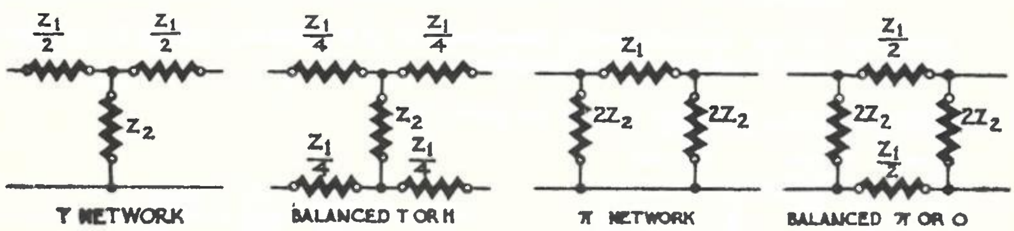


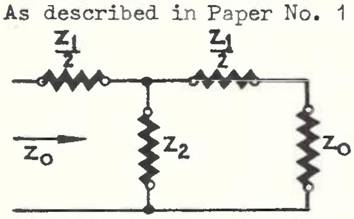
FIG. 1. T AND π SECTIONS.

2. CHARACTERISTIC IMPEDANCE OF NETWORKS.

2.1 If an infinite number of the single networks, or sections, shown in Fig. 1 were connected together and the input impedance measured or calculated, as was done in Paper No. 1, the same result would be obtained, that is, the infinite number of sections would exhibit a finite input impedance. This impedance is frequently referred to as the Iterative Impedance (to iterate means to repeat). In these books, this impedance is called the Characteristic Impedance and designated Z_o in order to be consistent with Paper No. 1 on Infinite Lines.

3. CHARACTERISTICS OF T SECTIONS.

3.1 As described in Paper No. 1, terminating a T network or section in its characteristic impedance makes the input impedance of the network equal to the characteristic impedance. This is shown in Fig. 2.



T SECTION TERMINATED IN Z_o.

FIG. 2.

3.2 Characteristic Impedance of T Sections. The characteristic impedance of such a network is:-

$$Z_o = Z_1Z_2 + \frac{{Z_1}^2}{4} \quad \ldots\ldots(1)$$

3.3 Attenuation Produced by T Sections. The attenuation of a T network can be calculated from the ratio of the input and output currents. These currents are shown in Fig. 3.

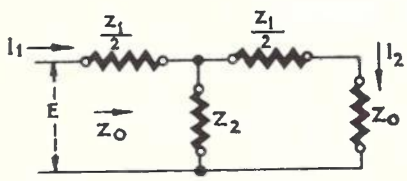


FIG. 3. CURRENTS IN T SECTION.

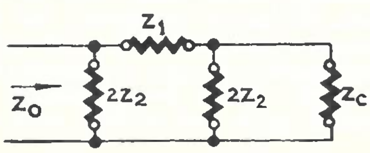
The ratio of input and output will be $\frac{I_1}{I_2}$, and is given by -

$$\frac{I_1}{I_2} = \frac{Z_2 + \frac{Z_1}{2} + Z_o}{Z_2} \quad \ldots\ldots(2)$$

$$\text{or } \frac{I_1}{I_2} = 1 + \frac{Z_1}{2Z_2} + \frac{Z_o}{Z_2} \quad \ldots\ldots(3)$$

4. CHARACTERISTICS OF π SECTIONS.

4.1 Characteristic Impedance of π Sections. In Fig. 4, Z_o is in parallel with $2Z_2$, this combination being in series with Z_1, and this combination, in turn, being in parallel with $2Z_2$. The characteristic impedance of such a network is given by -



π SECTION TERMINATED IN Z_o.

FIG. 4.

$$Z_o = \frac{Z_1Z_2}{\sqrt{Z_1Z_2 + \frac{{Z_1}^2}{4}}} \quad \ldots\ldots(4)$$

$$\text{or } Z_o = \frac{Z_1Z_2}{Z_{OT}} \quad \ldots\ldots(5)$$

where Z_{OT} is the characteristic impedance of a T section, having Z_1 as its series impedance and Z_2 as its shunt impedance.

4.2 Attenuation of π Network. This is given by -

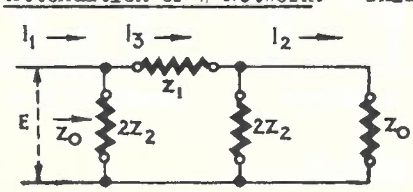


FIG. 5. CURRENTS IN π SECTION.

$$\frac{I_1}{I_2} = 1 + \frac{Z_o}{Z_2} + \frac{Z_1}{2Z_2} + \frac{Z_1Z_o}{4{Z_2}^2} \quad \ldots\ldots(6)$$

/4.3

4.3 T Sections are frequently called "mid-series" sections, and the impedance of such a section is referred to as the "mid-series impedance." π sections, on the other hand, are frequently called "mid-shunt" sections, and the impedance of such a section is referred to as the "mid-shunt impedance." The reason for these designations is shown in Fig. 6, which is termed a "ladder" network because of its configuration.

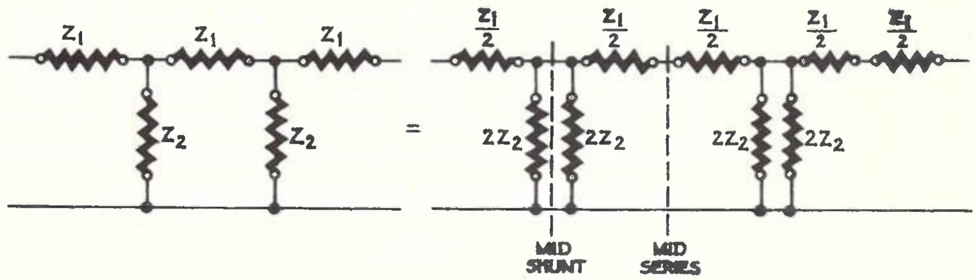


FIG. 6. MID-SERIES AND MID-SHUNT DERIVATION.

5. PADS AND ATTENUATORS.

5.1 Sometimes it is necessary to reduce the amount of power input to, or output from, a piece of equipment by a fixed amount. This is done by means of a suitable network having the desired fixed loss, such a network being termed a Pad.

5.2 Again, the same item of equipment, for example, an oscillator, may be used under circumstances which require the output power to be varied by known and variable amounts. This is done by means of a suitable network having the desired variable loss, such a network being termed an Attenuator.

5.3 Pads and Attenuators are usually of the "T" or "Balanced T" type, depending on whether strict balance is necessary or not. The series and shunt elements are purely resistive, as they usually have to offer a loss or attenuation which is the same at all frequencies. Further, the pads and attenuators usually operate between similar impedances, that is, their output and input impedances are the same, such networks being termed Symmetrical Networks. Where the input and output impedances of a network are unequal, the networks are termed Unsymmetrical Networks.

5.4 The problem of designing a symmetrical pad thus becomes one of working out values for the resistances in Figs. 2 and 4, so that -

$$\sqrt{R_1R_2 + \frac{R_1^2}{4}}$$

produces the required impedance between which the pad must work, together with the required amount of attenuation. In the above expression, R_1 and R_2 replace Z_1 and Z_2, respectively, of equation (1) because, as mentioned above, the pad elements are purely resistive. Tables are generally available where networks have to be designed frequently, such tables simplifying the calculations. Table 1 is included to indicate the accuracy required. By applying equations (1) or (4), it will be found that the characteristic impedances very closely approximate 600 ohms, whilst an application of equations (3) or (6) will show that the attenuation is correct.

/Table 1.

	T SECTION		π SECTION	
Loss in db.	Total Series Resist.	Shunt Resist.	Total Series Resist.	Shunt Resist.
1	69 ohms	5,208 ohms	68.6 ohms	10,440 ohms
2	137.6 ohms	2,582 ohms	139.4 ohms	5,232 ohms
3	205.4 ohms	1,703 ohms	212.5 ohms	3,505 ohms
4	271.6 ohms	1,249 ohms	287.5 ohms	2,651 ohms
5	336.2 ohms	987.6 ohms	364.5 ohms	2,141 ohms

VALUES OF RESISTANCES FOR 600 OHM ATTENUATORS.

TABLE 1.

5.5 The problem of attenuators is frequently complicated by the fact that the input and output impedances of the attenuator must remain unchanged as the values of the resistances in the attenuator change to produce the required amount of attenuation.

6. CHARACTERISTIC IMPEDANCE OF T NETWORKS OF PURE REACTANCES.

6.1 In order to understand the operation of filters, a knowledge of how networks having purely reactive elements behave is necessary.

6.2 As examples, some T networks of reactances will be examined. Figs. 7a and 7b show T sections of pure inductance and capacitance, respectively. The characteristic impedance of a T network of pure inductances is an inductive reactance at all frequencies, and the characteristic impedance of a T network of pure capacitances is a capacitive reactance at all frequencies.

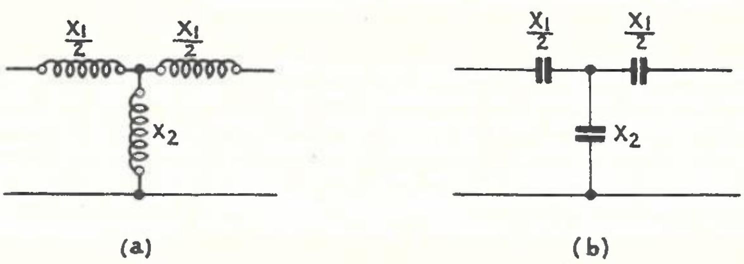


FIG. 7. T NETWORKS WITH SIMILAR REACTIVE ARMS.

6.3 Figs. 8a and 8b show two other arrangements of T sections, these having opposite types of reactances in the arms. The type of network shown in Fig. 8a has a characteristic impedance which is purely resistive at frequencies from zero up to that at which $X_2 = \frac{X_1}{4}$, and above that frequency the characteristic impedance is a pure inductive reactance. The network of Fig. 8b has a characteristic impedance which is a pure capacitive reactance between zero frequency and that at which $X_2 = \frac{X_1}{4}$. Above this frequency, the characteristic impedance is a pure resistance.

/Fig. 8.

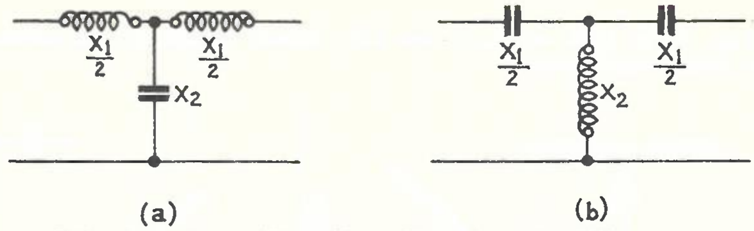


FIG. 8. T NETWORKS WITH DISSIMILAR REACTIVE ARMS.

6.4 It is not possible to explain by physical means, as was done for Wave Propagation in Paper No. 1 of this book, how a network consisting of pure reactances can behave as a pure resistance over a certain range of frequencies and as a pure reactance over other frequencies. The only advice that can be given to students who have difficulty in understanding the behaviour is that calculations indicate that the networks behave this way, and that practical measurements prove it.

7. ATTENUATION OF T NETWORKS OF PURE REACTANCES.

7.1 The networks dealt with in Fig. 7 have little or no application in these books - they were dealt with because they provided a simple introduction to the slightly more complex behaviour of Fig. 8. The networks of Fig. 8, however, are very important, as they form the basic types of low-pass and high-pass filter sections.

7.2 Whilst the two sections of Fig. 8 are behaving as resistances, they are capable of absorbing power from a generator connected to them, that is, the networks as a whole can take power from the generator. The individual reactances, however, still behave as pure reactances and, therefore, cannot dissipate any of the energy taken by the network as a whole from the generator. All of the power taken from the generator, therefore, is passed on to the termination connected to the other end of the sections. This means that the attenuation over the band of frequencies to which the networks offer a purely resistive characteristic impedance is zero. When the networks become purely reactive, however, they cannot absorb power from a generator connected to them, and the current and voltage will be 90° out of phase at all points. It would be physically possible for such a network to produce attenuation, that is, reduce the amplitudes of the current and voltage as they pass through them, because a decrease in voltage and current does not represent a dissipation of power when the current and voltage are 90° out of phase with each other.

7.3 From the general statement above, it should be clear that the attenuation produced by Figs. 8a and 8b is zero whilst the characteristic impedance is a pure resistance. The fact that attenuation is produced when the networks are reactive can be proved by using equation (3). A section, such as Fig. 8a, produces a phase shift of exactly 180° when the frequency is above the value at which $X_2 = \frac{X_1}{4}$.

8. PROTOTYPE L.P. AND H.P. FILTER SECTIONS.

8.1 Paragraphs 7.2 and 7.3 above indicate that a section of the type shown in Fig. 8a could be used as a low-pass filter because it produces zero attenuation between zero frequency and that at which $X_2 = \frac{X_1}{4}$ and attenuates all frequencies above this. Also, a section of the type shown in Fig. 8b could be used as a high-pass filter, because it produces zero attenuation at all frequencies above that at which $X_2 = \frac{X_1}{4}$ and attenuates all frequencies below this. These sections are called the basic or prototype filter sections because from these are developed the more complicated sections used in practical filters. The two sections, as they appear in Fig. 8, are not satisfactory for two reasons.

8.2 The first reason is that the attenuation does not rise sufficiently sharply beyond the frequency at which $Z_2 = \frac{Z_1}{4}$, called the "cut-off frequency" and designated fc. This

/is

is shown in Fig. 9 for both the L.P. and H.P. filter sections, the attenuation outside the pass band being calculated.

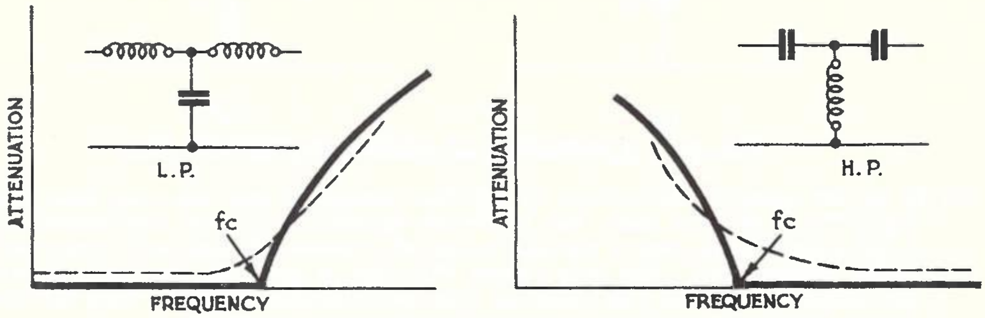


FIG. 9. ATTENUATION VERSUS FREQUENCY CURVES.

When resistance is present, as it always is because it is not possible to wind inductances without some resistance, the position is worsened because the presence of the resistance produces a gradual rather than a sharp cut-off, besides some attenuation in the pass band, as shown by the dotted lines of Fig. 9.

8.3 The second reason is that the characteristic impedance of such sections varies widely over the pass band. This can be shown as follows for the L.P. section.

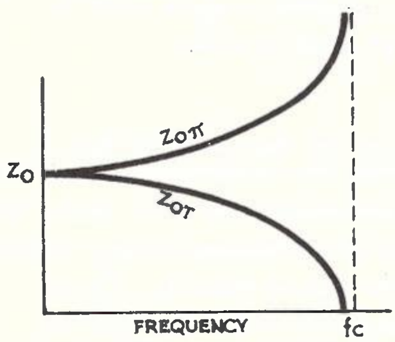
From equation (1) -

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

can be derived the equation for Fig. 8a -

$$Z_o = \sqrt{\frac{L}{C} - \frac{\omega^2 L^2}{4}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7)$$

An examination of equation (7) indicates that, when the frequency is zero, Z_o is $\sqrt{\frac{L}{C}}$ because $\frac{\omega^2 L^2}{4}$ is zero. As the frequency is increased, however, $\frac{\omega^2 L^2}{4}$ increases until, at the cut-off frequency, it equals $\frac{L}{C}$, so that over this band of frequencies, the pass band for Fig. 8a, the characteristic impedance varies between $\sqrt{\frac{L}{C}}$ and zero. This is shown in Fig. 10.



CHARACTERISTIC IMPEDANCE VERSUS FREQUENCY.

FIG. 10.

By the same reasoning, the H.P. section of Fig. 8b will exhibit a characteristic impedance which varies between $\sqrt{\frac{L}{C}}$ at infinite frequency and zero at the cut-off frequency. Such filter sections, therefore, could not be terminated in a single resistance as, if the resistance matched the characteristic impedance of the section at low frequencies, there would be severe reflection at high frequencies, and vice versa.

Included in Fig. 10 is the curve for the characteristic impedance for a π section low-pass filter over its pass band.

/9.

9. CUT-OFF FREQUENCIES FOR L.P. AND H.P. FILTERS.

9.1 The cut-off frequency, fc, occurs in both types when $X_2 = \frac{X_1}{4}$, as previously stated.

For the low-pass section -

$$X_1 = \omega L \text{ and } X_2 = \frac{1}{\omega C}$$

$$\therefore \text{ at fc, } \frac{1}{\omega C} = \frac{\omega L}{4}$$

$$\therefore \omega^2 LC = 4$$

$$\therefore \omega^2 = \frac{4}{LC}$$

$$\therefore 4\pi^2 fc^2 = \frac{4}{LC}$$

$$\therefore fc^2 = \frac{4}{4\pi^2 LC}$$

$$\therefore fc = \frac{1}{\pi\sqrt{LC}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(8)$$

For the high-pass section -

$$X_1 = \frac{1}{\omega C} \text{ and } X_2 = \omega L$$

$$\therefore \text{ at fc, } \omega L = \frac{\frac{1}{\omega C}}{4}$$

$$\therefore \omega L = \frac{1}{4\omega C}$$

$$\therefore \omega^2 = \frac{1}{4LC}$$

$$4\pi^2 fc^2 = \frac{1}{4LC}$$

$$\therefore fc^2 = \frac{1}{16\pi^2 LC}$$

$$\therefore fc = \frac{1}{4\pi\sqrt{LC}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(9)$$

Equations (8) and (9) are the cut-off frequencies for low-pass and high-pass filter prototype sections in terms of the inductance and capacitance used.

/10.

10. BAND-PASS FILTERS.

10.1 A band-pass filter is essentially a low-pass filter in series with a high-pass filter. For example, if it is desired to pass only the band 8-10 kc/s from a band containing all frequencies from zero to infinity, the low-pass filter would have a cut-off frequency of 10 kc/s and the high-pass filter a cut-off frequency of 8 kc/s. The two filters are shown in Fig. 11a and are combined together, as they are in practice, in Fig. 11b.

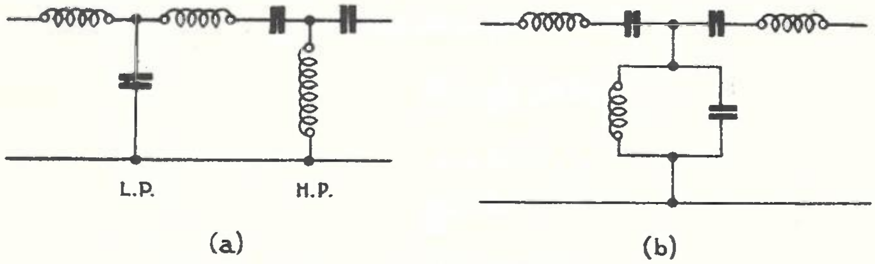


FIG. 11. BAND-PASS FILTER.

10.2 Fig. 12 shows the attenuation versus frequency characteristics of a band-pass filter of the type shown in Fig. 11b. The theoretical curve is shown in full line and the actual values shown dotted.

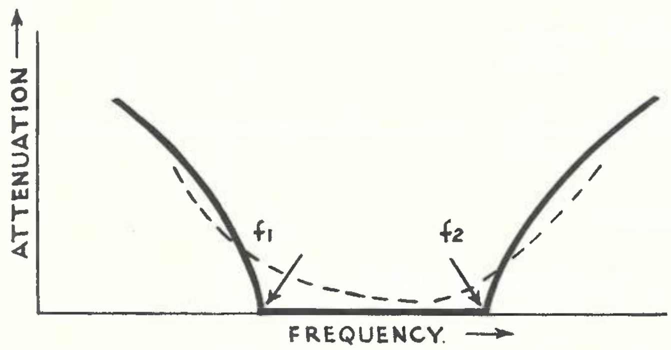


FIG. 12. ATTENUATION VERSUS FREQUENCY CHARACTERISTIC.

10.3 As it is impossible in practice to construct inductances and condensers which are entirely non-resistive, there is always a slight amount of attenuation in the pass-band of a filter. In general, this does not exceed 0.5 db, except in the neighbourhood of the cut-off frequencies, f_1 and f_2, where the effect of the resistance is to cause a rounding-off of the attenuation-frequency characteristic.

11. m-DERIVED FILTER SECTIONS.

11.1 A higher attenuation at frequencies just outside the pass-band, together with a sharper cut-off, can be obtained by the use of "m-derived" filter sections. Fig. 13 shows such a section for a low-pass filter.

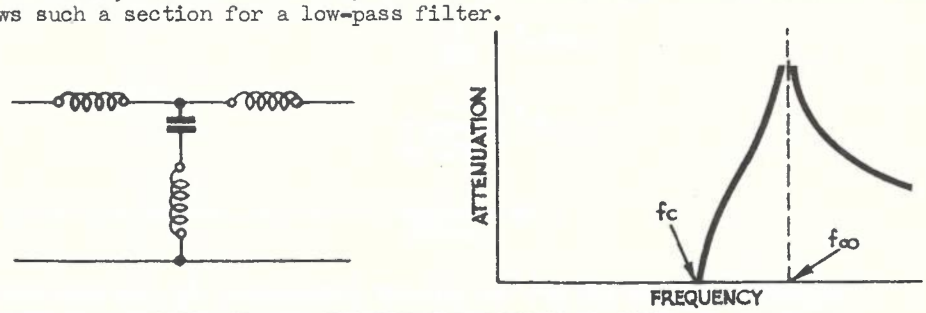


FIG. 13. m-DERIVED L.P. FILTER SECTION.

As shown in Fig. 13a, the shunt element is a series circuit and is designed to be resonant at a frequency just above the cut-off frequency. At this frequency, designated f_∞ in Fig. 13b, the shunt element is a short-circuit and the attenuation is infinite. However, past f_∞ the attenuation falls, so that, in practical filters, a number of such sections is used, each having a successively higher value of f_∞ and so providing the necessary high attenuation throughout the stop band.

/11.2

11.2 These sections are called m-derived sections, because of the relationship between the series elements of the prototype from which they are designed and the series elements in the derived type. The different sections in a filter must have the same characteristic impedance in order to prevent reflection between the sections. If the characteristic impedances are the same at all frequencies, then the different sections will have the same transmission bands as, in this band, and in this band alone, Z_0 is a pure resistance. The series elements of a derived type will, of course, differ in value from the prototype because of the inclusion of other components in the shunt element. Thus, in Fig. 14, Z_1 and Z_2 are the series and shunt elements, respectively, of the prototype, and $Z_1{}^1$ and $Z_2{}^1$ are the series and shunt elements, respectively, for a derived type.

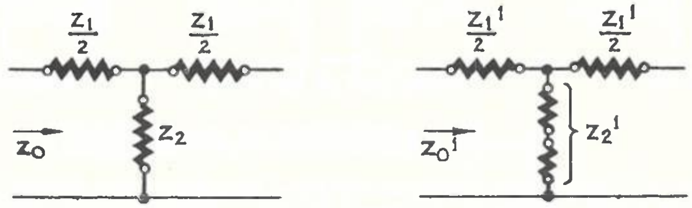


FIG. 14. PROTOTYPE AND m-DERIVED SECTIONS.

As mentioned above, the transmission bands match when the characteristic impedance match, the latter being necessary to prevent reflection. The relation between Z_1 and $Z_1{}^1$ has been designated -

$$Z_1{}^1 = mZ_1$$

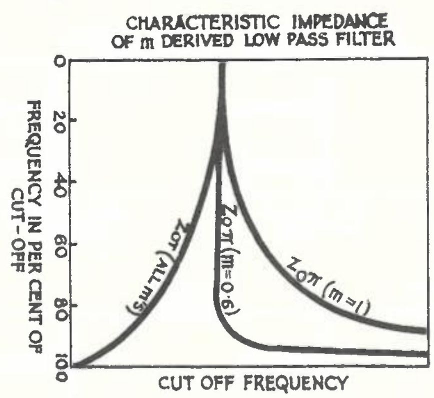
$$\text{Also } Z_0{}^1 = Z_0$$

11.3 It is possible to obtain a value for m for any value of fc and f∞ required (with which this Paper is not concerned) and, from a prototype, to design the required m-derived structures. This is a design problem and will not be attempted here.

12. PRACTICAL FILTERS.

12.1 A practical filter consists of prototype sections and m-derived sections designed to produce the required characteristic impedance (which is the impedance between which the filter works), together with a sharp cut-off and a high attenuation in the stop band and a low and even attenuation in the pass band.

12.2 The variation in characteristic impedance over the pass band has now to be corrected. This is done by taking advantage of the fact that the characteristic impedance of a section with an m of 0.6 is constant over practically the whole of the pass band. From equations (4) and (7), it is evident that the characteristic impedance of a prototype section L.P. filter is -



CHARACTERISTIC IMPEDANCE VERSUS FREQUENCY m-DERIVED SECTIONS.

FIG. 15.

$$\frac{\frac{L}{C}}{\sqrt{\frac{L}{C} - \frac{\omega^2 L^2}{4}}} \quad \text{........................(10)}$$

Equation (10) indicates that, at zero frequency, the characteristic impedance of a prototype π section of a low-pass filter is $\sqrt{\frac{L}{C}}$ as in the T prototype, but at the cut-off frequency it is infinity because the denominator of equation (10) is zero at that frequency. By working out the characteristic impedance at all frequencies in the pass band for all values of m for an m-derived π section, that is, by experimental mathematics, it is found that the characteristic impedance of an m-derived π section with an m of 0.6 is substantially constant over the pass band, as shown in Fig. 15.

/12.3

12.3 The procedure now is to design an m-derived T section with an m of 0.6, split the shunt element into two equal parts and, with these two half T sections, terminate the filter with them arranged as half π sections, as in Fig. 16.

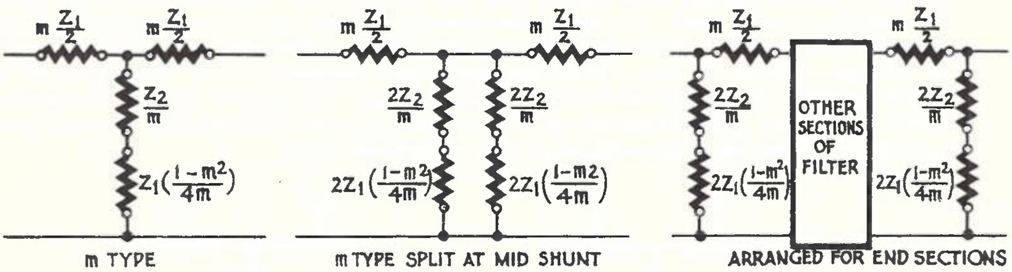


FIG. 16. ARRANGEMENT OF END SECTIONS.

Fig. 17 shows a 3 kilocycle low-pass filter, together with curves giving some idea of the purpose of each unit in a practical filter. Such filters are frequently called Composite Filters.

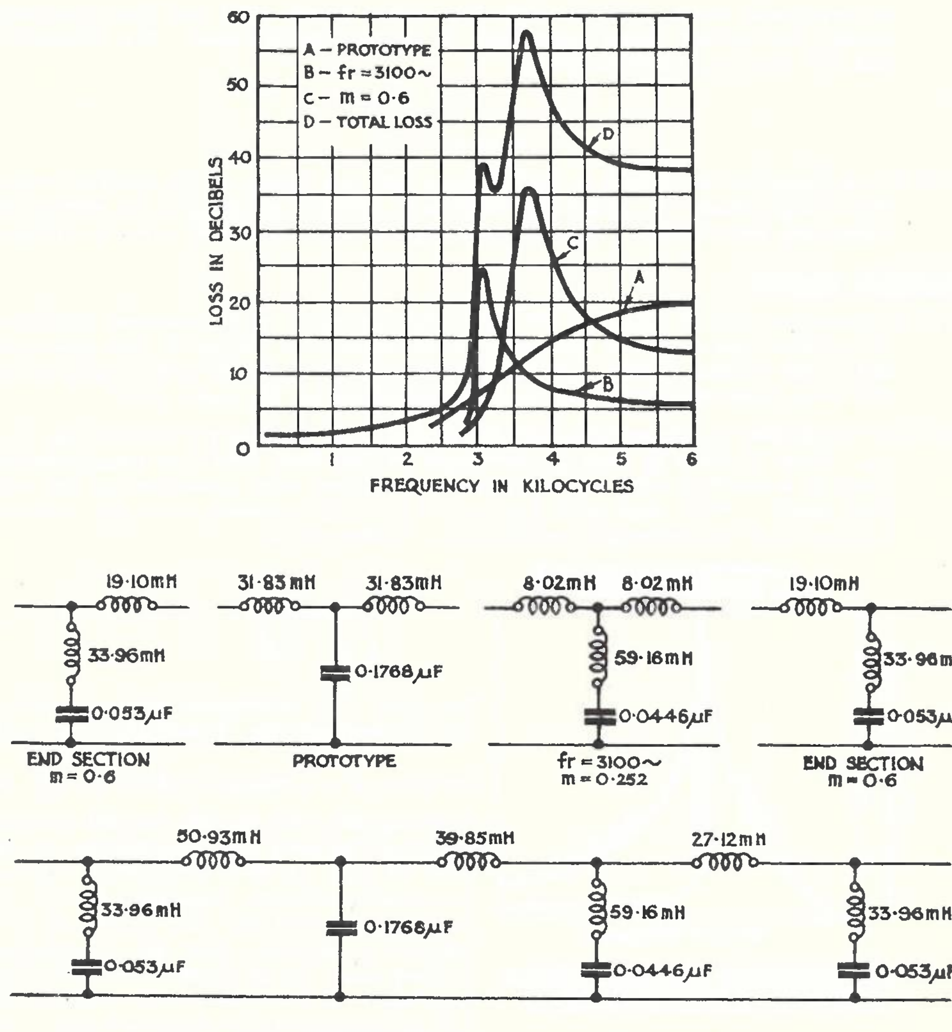


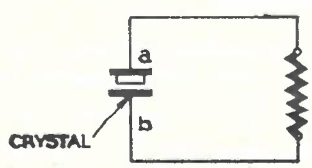
FIG. 17. TYPICAL LOW-PASS FILTER.

13. CRYSTAL FILTERS.

13.1 Certain materials, notably some species of quartz, exhibit the property of developing an e.m.f. across them when subject to pressure or tension. For example, when a mechanical pressure is applied to such a material, an e.m.f. is developed across two opposite faces of the material, which is in crystalline form. When the pressure is removed, the natural elasticity of the material allows it to resume its former state and the direction of the e.m.f. is reversed. This action is reversible in that when an alternating e.m.f. is applied to two opposite faces, slight changes of shape take place which cause the material to vibrate. As with all objects, a piece of such material will exhibit a natural frequency of vibration, that is, a frequency at which the applied alternating e.m.f. will cause the material to execute vibrations at a maximum amplitude.

13.2 As the material vibrates, therefore, alternating voltages are developed across opposite faces, and it would be expected that the larger the amplitude of vibration, the greater will be the amplitude of the voltage developed across it, reaching a maximum at the natural frequency of the specimen being used, and this is found to be so. Specimens of different size display different natural frequencies so that preparation of a specimen to display a particular natural frequency consists in cutting a section of a large specimen to the desired dimensions. In practice, the thickness of the specimen largely determines its natural frequency. The completed specimen is called a "crystal," and preparation of the crystal mainly consists of cutting a specimen from a larger one and grinding it to the size required to produce the required natural frequency of vibration.

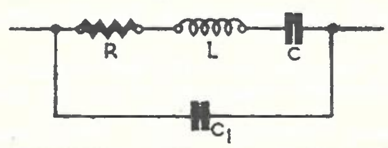
13.3 As discussed above, when a crystal vibrates, an alternating e.m.f. is developed across two of its opposite faces. The action is perhaps best explained electronically from Fig. 18. The crystal in Fig. 18 is caused to vibrate by means not shown, the opposite faces (across which the e.m.f. is developed) being connected to some circuit. During the half-cycle when side a is positive to side b, side b will exhibit a surplus of electrons that were transferred from side a (which exhibits a deficit) via the circuit connected to the crystal. During the other half-cycle, side a is negative and, therefore, displays a surplus of electrons collected from side b via the circuit. Thus, the alternating voltage developed across the crystal faces causes an alternating current to flow through the circuit. This alternating current will be a maximum at the resonant frequency of the crystal, that is, when the amplitude of vibration of the crystal and, therefore, the voltage developed across it is a maximum.



VIBRATING CRYSTAL IN CIRCUIT.

FIG. 18.

13.4 In this respect, the crystal behaves as a series resonant circuit, the equivalent circuit being shown in Fig. 19. L is the effective mechanical inductance due to the mass of the crystal, R is the effective resistance due to losses produced by intermolecular friction as it vibrates, and C is the effective mechanical capacity due to the elasticity of the crystal. C_1 is the capacity between the faces of the crystal due to its straight condenser action.



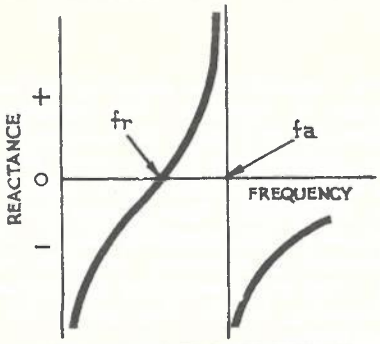
EQUIVALENT CIRCUIT OF CRYSTAL.

FIG. 19.

/13.5

/13.5

13.5 The reactance curve of a crystal is shown in Fig. 20. From Fig. 20 it will be seen that the crystal has a resonant frequency fr and an anti-resonant frequency fa. The resonant frequency is due to the series circuit alone, whilst the anti-resonant frequency is due to the shunting effect of C_1 in Fig. 19. The ratio of C_1 to C is a constant for any given material; for quartz it is 125 to 1. This gives an anti-resonant frequency which is 0.4% higher than the resonant frequency for quartz.



REACTANCE CURVES OF CRYSTAL.

FIG. 20.

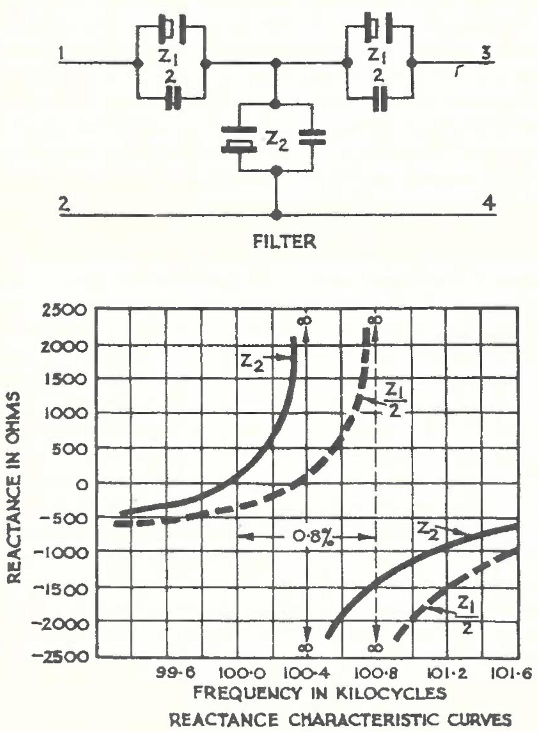
13.6 The desirable feature of crystals is their low equivalent resistance, being near enough to zero for most practical purposes. This makes crystals ideal for such uses as sharply tuned circuits and as reactive elements in filters where a sharp cut-off is required.

13.7 Fig. 21 shows a T section crystal filter with its associated reactance curve and attenuation curve.

The crystal elements are so selected that the resonant frequency of the series arm (zero reactance) coincides with the anti-resonant frequency of the shunt arm (infinite reactance). Under this condition, the network will have, from merely a general examination, maximum attenuation at the anti-resonant frequency of the series arm (infinite series reactance) and at the resonant frequency of the shunt arm (zero shunt reactance).

As in the preceding filters, the pass band will extend over the frequency band in which Z_1 and $Z_2 + \frac{Z_1}{4}$ are opposite types of reactance. The curve for $\frac{Z_1}{4} + Z_2$ can be obtained by adding half the values of the $\frac{Z_1}{2}$ curve of Fig. 21 to the Z_2 curve. The resultant curve will exhibit much the same resonant and the same anti-resonant frequencies as Z_2, and will exhibit a negative reactance below and a positive reactance above resonance. Similarly, the curve for Z_1 can be obtained by doubling all values of the $\frac{Z_1}{2}$ curve in Fig. 21. Such a curve will display the same resonant and anti-resonant frequencies as the $\frac{Z_1}{2}$ curve.

Thus, the pass band is near enough to between the resonant frequency of Z_2 and the anti-resonant frequency of $\frac{Z_1}{2}$. This means that the pass band of such a filter is only 0.8% of the mid-frequency, for example, in Fig. 21 the pass band is only 800 c/s at 100 kc/s, which is too narrow for a voice frequency channel. Where only a very narrow band is desired (much narrower than the 125 to 1 ratio of C_1 to C can provide), the ratio is increased by a condenser shunting the crystal, as in Fig. 21, which
/effectively



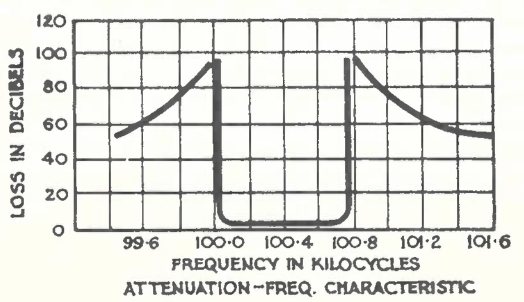


FIG. 21. T FILTER USING CRYSTALS.

effectively reduces the band-width passed.

13.8 The crystal filters used are usually of "lattice" formation, those dealt with up to the present being of "ladder" formation. (When T or π sections are connected one after the other, the structure is not unlike a ladder.) Fig. 22 shows a lattice filter, together with its reactance and attenuation curves.

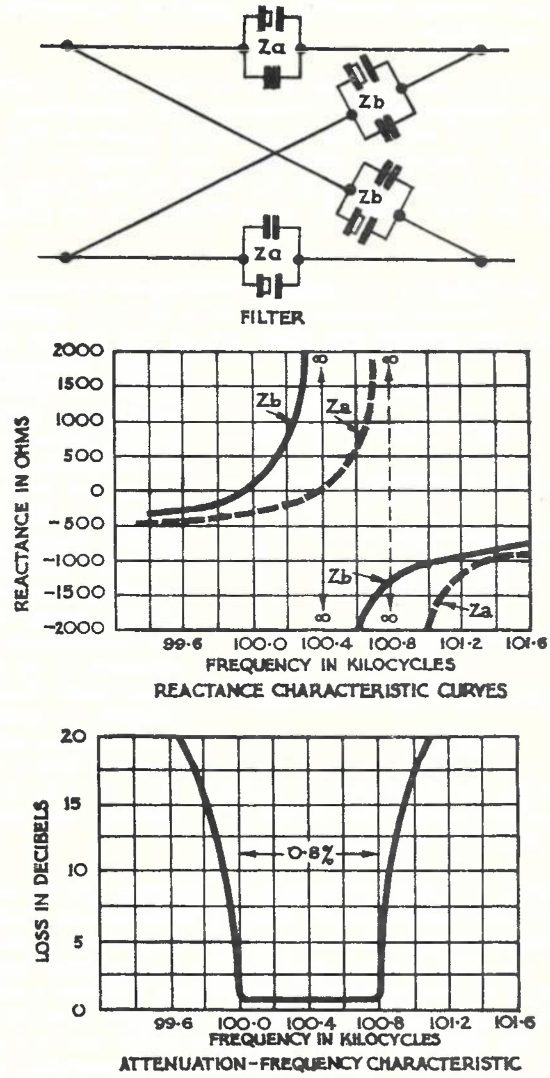


FIG. 22. CHARACTERISTICS OF LATTICE FILTER USING CRYSTALS.

The series arms are alike, as are the two shunt arms, but the series and shunt arms differ as will be seen from the reactance curves. The pass band extends over those frequencies at which the series and shunt arms are opposite types of reactance. Again, as in Fig. 21, the pass band is only 0.8% of the mid-frequency.

To produce the wider pass bands necessary in practice, an inductance is used in series with the crystals. Within certain limits, this series inductance does not introduce enough resistance to interfere with the sharpness of cut-off produced by the crystal, and, by a suitable adjustment of the shunt condenser across the crystals, pass bands of any desired width may be obtained within the limits permitted by the inductance.

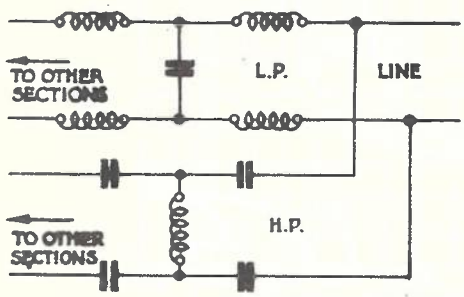
/14.

14. PARALLEL CONNECTION OF FILTERS.

14.1 It is frequently necessary to operate filters in parallel. In Fig. 13 of Paper No. 2, there are three examples of this practice. The L.P. and H.P. line filters are parallel connected on their line sides, the two directional filters are parallel connected on the H.P. line filter sides and the two groups of three band-pass filters in the channel circuits are parallel connected on the side of the appropriate amplifier.

14.2 Generally speaking, all filters have to work into a circuit that displays an impedance which is independent of frequency and, therefore, is resistive. As an example, an aerial line into which the line filter group of Fig. 13 in Paper No. 2 works is usually, for all practical purposes, near enough to a zero angle line. This means that the impedance looking into the line filter group from the line, that is, the impedance of the H.P. and L.P. line filters in parallel, should be a constant resistance.

14.3 As discussed previously, a constant resistance termination is provided for a filter by designing a derived section with an m of 0.6, splitting this section down its shunt element and terminating each end of the filter with the half sections so formed arranged as half sections. This effect is achieved in the parallel connection by using one of the T sections, either prototype or derived, in each filter to terminate the filters on the sides which are to be paralleled, as in Fig. 23. From Fig. 23 it will be seen that the first series capacity and shunt inductance of the H.P. filter provide a shunt path containing capacitance and inductance in series across the input to the filter and are equivalent, therefore, to the shunt element of an m-derived L.P. section. This means that merely connecting the two filters in parallel, as described above, provides an m-derived termination for the group. Where the components employed in the terminating sections do not provide an m of 0.6, an adjustment can be made to the first series inductance of the L.P. section and the first series capacitance of the H.P. section to achieve this.



L.P. AND H.P. FILTERS IN PARALLEL.

FIG. 23.

14.4 When paralleling a number of band-pass filters, it is necessary to ensure that the impedance of each filter rises sharply outside its pass band. This is done by using a suitable design of T section to terminate each filter at the end to be paralleled. By this means, each filter in its pass band will be shunted by the high impedance of the remaining paralleled filters outside their pass bands, and the characteristics of each filter in its pass band will be unaffected by the presence of the others. Where the pass bands are not spaced widely apart, it is necessary to go a step further, and this is generally the case. As before, the filters are connected in parallel, but there will now be a difference in that the impedances shunted across each filter by its neighbours will be lower owing to the narrower gaps between the pass bands.

The filters above and below will contribute reactive components which are opposite in sign and, therefore, will tend to correct one another. For example, above the pass band of filter 1 in Fig. 24, the reactive component is inductive, whilst below the pass band of filter 2 the reactive component is capacitive. Even though these reactances be low in value they will correct one another, as will the inductive reactance above the pass band of filter 2 and the capacitive reactance below the pass band of filter 3.

In Fig. 24, however, there is no compensating inductive reactance to correct the low capacitive reactance immediately below the pass band of filter 1, and no

/compensating

compensating capacitive reactance to correct the low inductive reactance immediately above the pass band of filter 3. It is often necessary, therefore, to connect an auxiliary network, called a "compensating network," across the whole filter group to correct these residual low reactances.

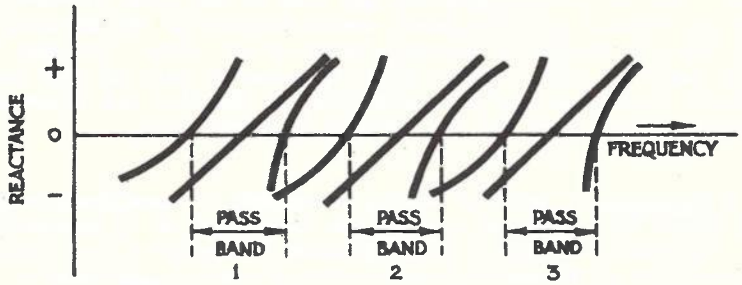


FIG. 24. B.P. FILTERS IN PARALLEL.

15. EQUALISERS.

15.1 As mentioned in Paper No. 1, one of the factors which tend to decrease the intelligibility of telephone speech is unequal attenuation of the currents of different frequencies as they are transmitted over the circuits. For example, the attenuation of an open wire circuit is greater for the higher frequencies than for the lower frequencies, and this difference in attenuation is directly proportional to the length of line. Therefore, when long circuits are employed, it is frequently necessary to employ attenuation equalisers to correct the unequal attenuation of the line. These equalisers are usually associated with the repeaters or amplifiers which are included in the circuit at various points, in order to provide a uniform level of current at the receiving end over the frequency range employed. The unequal attenuation of different frequencies is even more marked in unloaded cable circuits.

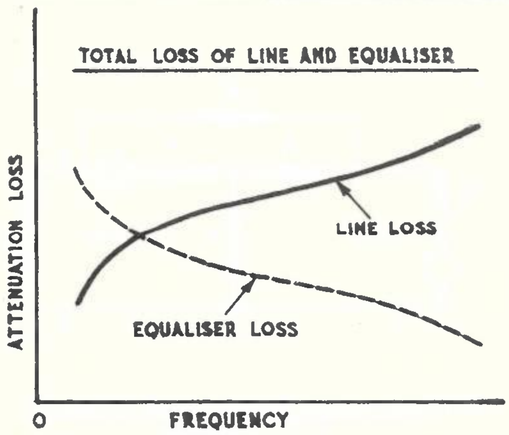


FIG. 25. PRINCIPLE OF EQUALISER.

15.2 Attenuation equalisers are networks consisting of inductances, condensers and resistances which are so proportioned and arranged that their attenuation-frequency characteristics are complementary to the line characteristics that produce the distortion. In brief, the total loss of the line plus the loss produced by the equaliser is the same for all frequencies within the band of frequencies concerned. This is shown in Fig. 25.

15.3 One of the simplest types of equalisers, shown schematically in Fig. 26, is bridged directly across the circuit to be corrected. The impedance of such a bridged equaliser must be low enough at certain frequencies to allow sufficient current to flow through it to produce the required losses at these frequencies. Accordingly, the equaliser circuit naturally changes the circuit impedance, particularly at the frequencies where the equaliser is to provide a substantial loss. This introduces an impedance irregularity of a sizeable value into the circuit. The use of bridged equalisers of this type, therefore, has definite limitations in practice.

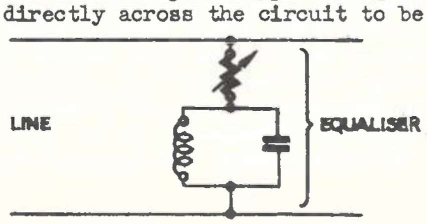
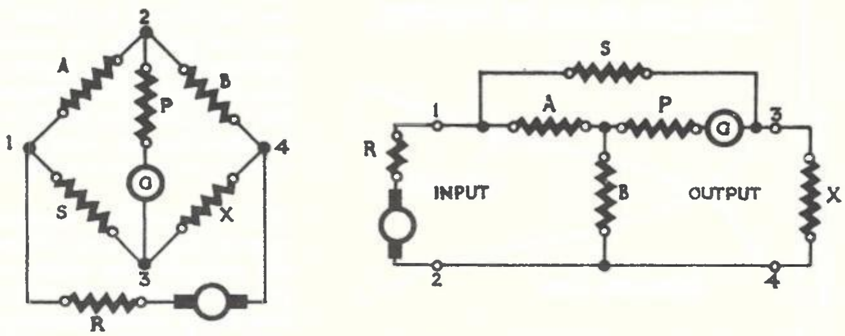


FIG. 26. SIMPLE BRIDGED EQUALISER.

15.4 In long circuits equipped with telephone repeaters, the desired equalising effects can be obtained without introducing an appreciable impedance irregularity by inserting equalising networks at the mid-point of the primary sides of the repeater input transformers. Instead of changing the net loss of the line, however, this arrangement changes the overall gain-frequency characteristic of the repeater to match reasonably closely the loss-frequency characteristic of the line. In other words, for the frequencies where the line loss is high the repeater gain is also high, and vice versa. The overall loss-frequency characteristic of the line and repeater together is then reasonably uniform over the transmitted frequency band.

15.5 Bridged T Equaliser. Both of the above methods of equalisation give satisfactory results where the amount of attenuation distortion to be corrected is relatively small. The use of either of these methods to correct a large attenuation distortion might result in an impedance irregularity of such a magnitude as to more than offset the benefits obtained by equalising. To equalise for these relatively large amounts of attenuation distortion, a somewhat more complex equalising network, in the form of a Bridged T structure, may be used. This equaliser is designed to have a constant impedance over the entire frequency band transmitted.

As the name implies, the Bridged T equaliser is built in the general form of a T network, but it has an additional impedance path bridged across the series elements. This latter path controls the loss of the equaliser. The elements of the Bridged T equaliser are connected in a Wheatstone bridge arrangement, and the principle of its operation may be best grasped by first referring to the ordinary Wheatstone bridge circuit shown in Fig. 27a.



(a) Wheatstone Bridge Circuit.

(b) Wheatstone Bridge in the Form of a "T" Network.

FIG. 27. BRIDGE PRINCIPLE APPLIED TO BRIDGED T EQUALISER.

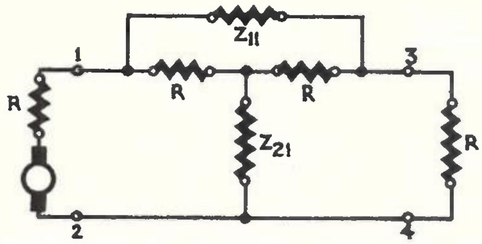
Now rearrange this bridge circuit in the form of a T network where the series elements are bridged by the impedance S, as shown in Fig. 27b. The T network proper is formed by A, P and B with S as the bridging impedance, while R and X now become the input and output impedances, respectively. Next, change the impedances R, A, P and X to resistances of equal value, which may then all be designated as R. For reasons to be explained later, the impedance S and B will also be redesignated as Z_{11} and Z_{21}, respectively. Then, as illustrated in Fig. 28, the bridge which was balanced when

$SB = XA$ is now balanced when -

$$Z_{11}Z_{21} = R \times R \text{ or } R^2.$$

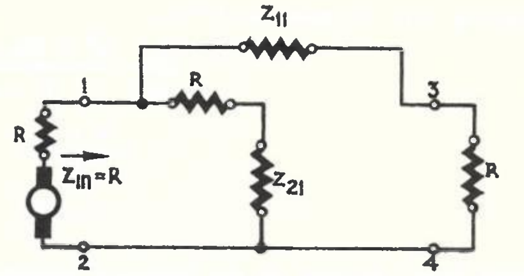
/When

When the bridge is balanced no current flows in the impedance, P (Fig. 27a), and, for purposes of analysis, the network may be simplified by removing the R resistance which replaced P in Fig. 28, producing the network of Fig. 29.



BRIDGED T EQUALISER.

FIG. 28.



REARRANGEMENT OF BRIDGED T EQUALISER.

FIG. 29.

Looking from the generator across terminals 1-2 of this network, two parallel paths can be seen which present an input impedance Z_{in} of -

$$Z_{in} = \frac{(R + Z_{11})\ (R + Z_{21})}{R + Z_{11} + R + Z_{21}}$$

$$= \frac{R^2 + RZ_{11} + RZ_{21} + Z_{11}Z_{21}}{2R + Z_{11} + R_{21}}$$

or, substituting R^2 for $Z_{11}Z_{21}$ -

$$Z_{in} = \frac{R(2R + Z_{11} + Z_{21})}{2R + Z_{11} + Z_{21}} = R$$

In other words, when the bridge is balanced ($Z_{11}Z_{21} = R^2$), the input impedance of the equalising network is a pure resistance, R. Moreover, since the T network is symmetrical, the same reasoning can be applied at the output terminals 3-4 and the impedance will also be found to be a pure resistance, R, for the balanced condition.

As in the case of any other circuit, the loss produced by this network may be determined by the ratio of the current, I_b (received in the output impedance before the network is inserted), to the current I_a (received after the network is inserted). Thus, the current, I_b, in the output before the network is inserted will be -

$$I_b = \frac{E}{R + R} \text{ or } \frac{E}{2R}$$

After inserting the network, the output of the generator will remain the same, because the impedance of the network as seen at terminals 1-2 is still R. As the input current divides into the two parallel paths (Fig. 29), the current, I_a, flowing in the output (terminals 3-4) is -

$$I_a = I_b \times \frac{R + Z_{21}}{R + Z_{21} + R + Z_{11}}$$

$$\text{Then } \frac{I_a}{I_b} = \frac{R + Z_{21}}{2R + Z_{11} + Z_{21}}$$

$$\text{or } \frac{I_b}{I_a} = \frac{2R + Z_{11} + Z_{21}}{R + Z_{21}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (11)$$

/Since

Since the balanced condition is being considered, where

$$Z_{11}Z_{21} = R^2, \text{ then } Z_{21} = \frac{R}{Z_{11}}$$

Substituting this in equation (11) -

$$\frac{I_b}{I_a} = \frac{R^2 + 2RZ_{11} + Z_{11}^2}{R^2 + RZ_{11}}$$

$$= \frac{R + Z_{11}}{R}$$

$$= 1 + \frac{Z_{11}}{R}$$

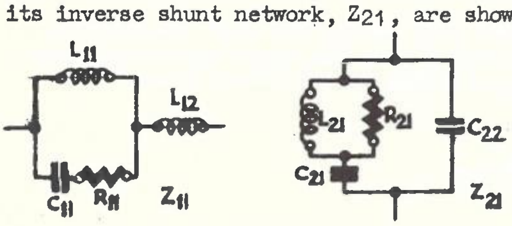
This shows that, as long as the balanced condition is maintained, the loss of the network is determined by Z_{11}. This is also apparent from an inspection of Fig. 29, because this impedance is in series with the receiving circuit and any value of loss may be secured without affecting the input or output impedances, providing the balanced condition is maintained.

To summarise, the Bridged T equaliser has a constant impedance, as seen from both terminals, equal to R when Z_{21} is the inverse of Z_{11} ($Z_{21}Z_{11} = R^2$), and its overall loss-frequency characteristic is determined by the bridged series impedance network, Z_{11}.

Both Z_{11} and Z_{21} represent generalised impedances, which may be resistances, capacitances, inductances or any combination of them. The one and only requirement is that established by the balanced condition ($Z_{11}Z_{21} = R^2$), which means there must always be an inverse relationship between Z_{11} and Z_{21}. If Z_{11} is a pure inductive reactance, then Z_{21} must be a capacitive reactance. On the other hand, if Z_{11} is a capacitance, Z_{21} must be an inductance, which is the reverse of the above case. If Z_{11} is a resistance, then Z_{21} will also be a resistance. When Z_{11} is a network, Z_{21} is a network with the same number of elements, but each element is the inverse of the corresponding element of Z_{11} as shown by the following table -

When Z_{11} is -	Z_{21} becomes -
Inductive reactance	Capacitive reactance
Capacitive reactance	Inductive reactance
Resistance	Resistance
Series inductance	Parallel capacitance
Series capacitance	Parallel inductance
Parallel resonance	Series resonance
Series resonance	Parallel resonance

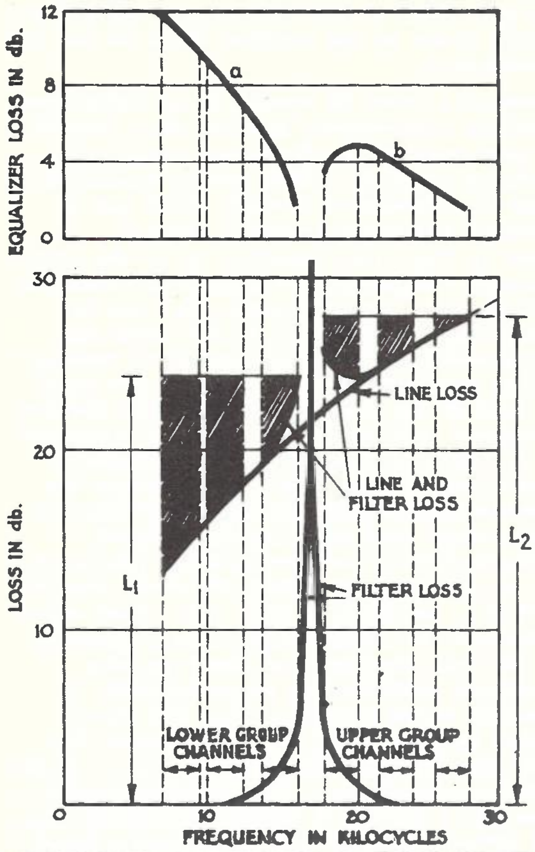
This inverse relationship is further shown in Fig. 30 where the series network, Z_{11}, and its inverse shunt network, Z_{21}, are shown at the left and right, respectively. Here, the advantages of using the two-digit subscript for Z become more evident. The first digit of the subscript indicates whether the element belongs to the series or shunt impedance, while the second digit designates the corresponding inverse elements of the two networks. Therefore, in Fig. 30, C_{21} is the inverse of L_{11}; C_{22} is the inverse of L_{12}; L_{21} is the inverse of C_{11}; and R_{21} is the inverse of R_{11}. In designing a Bridged T equaliser for a specific use, the attenuation-frequency characteristic of the Z_{11} network must be complementary to the attenuation-frequency characteristic of the circuit to be corrected. This is true, because the loss-frequency characteristic of the Bridged T equaliser is controlled by the series impedance network, Z_{11}.



INVERSE SERIES AND SHUNT NETWORKS.

FIG. 30.

As an example of the general problem involved in the design of the Z_{11} network, consider /a a representative application of its use in a 3-channel carrier system. In this, as in other carrier systems, separate frequency bands are employed for transmission in the two directions. For example, in one system, transmission in one direction occupies the frequency range from about 6 to 16 kc/s, while transmission in the opposite direction is in the range from about 18 to 28 kc/s. At the terminals and intermediate repeater points, the entire frequency band used in transmitting in each direction, which in the 3-channel systems includes three separate voice channels, is amplified by a single amplifier. The frequency bands transmitting in opposite directions are separated by means of so-called "directional filters."



ATTENUATION EQUALISATION FOR THREE-CHANNEL CARRIER SYSTEM.

FIG. 31.

The attenuation of the line facilities varies very considerably over the wide band of frequencies used. The directional filters also introduce appreciable distortion near their cut-off frequencies. In order to maintain uniform transmission, therefore, it is necessary to employ equalisers to counteract both of these attenuation distortion factors. This situation is shown in Fig. 31. Here, the losses produced by the line and the filters individually and their total loss are indicated by the curves so designated. (The frequency positions of the three voice channels in each direction of transmission are indicated by the vertical dashed lines.) The required loss-frequency characteristic of the equalisers is shown by the two upper curves a and b of Fig. 31, each of which is made complementary (inverse) to the total line and filter loss over the frequency band for its direction of transmission. By adding the losses of the line, filters and equaliser for each direction of transmission, the resultant loss-frequency characteristic becomes a straight horizontal line in each case. Because of the rising characteristic of the line, however, the total loss for the three lower voice channels, L_1, is less than that of the three higher voice channels, L_2. This difference is readily corrected by making the amplifier gains different for the two directions of transmission.

Now that the factors which give these equalisation curves (a and b of Fig. 31) and their particular characteristics have been noted, it is possible to analyse in a general way the equaliser design considerations for one curve - say curve a of Fig. 31. Clearly, the loss-frequency characteristic of the series impedance, Z_{11}, should conform as closely as practicable with the curve a of Fig. 31, or with the solid line curve a of Fig. 32, which is the same. As a first approach, a Z_{11} circuit made up of a single series condenser, as in a of Fig. 33, will give the general loss-frequency characteristic shown by curve b of Fig. 32. This, of course, is due to the fact that the current through a condenser increases with frequency; consequently, its loss decreases. However, it will be noted that curve b diverges widely from the desired characteristic curve a at the higher frequencies.

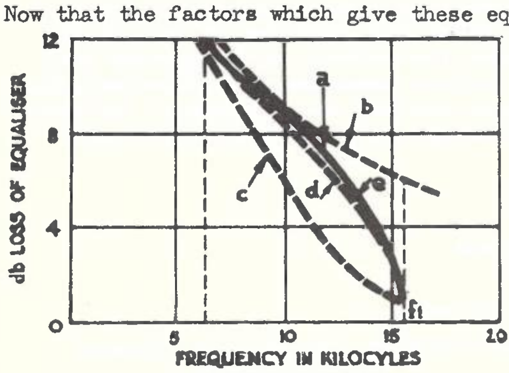


FIG. 32. EQUALISER CURVES.

To reduce the loss at f_1 and thus bring the curves closer together, an inductance, L_{11}, can be added in series with the condenser as in Fig. 33b. This inductance is of such a value that series resonance occurs at approximately the frequency f_1. This fails to solve the problem,

/however

however, because below the resonant frequency this series circuit produces a loss that increases with decreasing frequency, as shown by curve c of Fig. 32. Because of the inductance, the curve has now become too low over most of the frequency range but yet fairly close to the desired value at the two extremities.

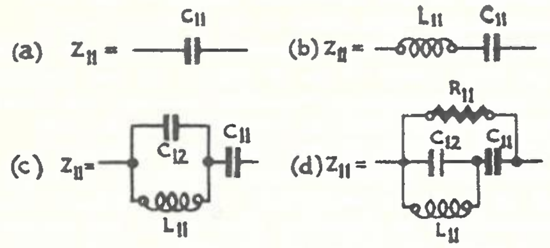


FIG. 33. EQUALISER ELEMENTS.

Apparently, what is needed is an inductance which is considerably smaller than that of L_1 over most of the frequency range but equal to it at f_1. This can be simulated by a parallel resonant circuit which has a resonant frequency somewhat above f_1, as shown in Fig. 33c, because, up to the anti-resonant frequency, the inductive reactance of a parallel resonant circuit increases with frequency. On this basis, L_{11} can be selected so that it is small enough to approximate the desired loss at the lower and mid frequencies. Then, by shunting a condenser around L_{11} (forming a parallel resonant circuit), the effective inductance of the parallel combination at f_1 can be made equal to that of L_{11} alone. In this way, the low impedance (and hence low loss) is preserved at f_1, and the loss is still increased at lower frequencies. The net effect is the characteristic shown by curve d of Fig. 32, which comes very close to the desired characteristic, but even greater precision can be obtained by adding the shunt resistance R_{11}, as shown by Fig. 33d. This shunt resistance R_{11} introduces a small increase in the loss over most of the frequency range and modifies the characteristic as shown by curve e of Fig. 32. The final series network, Z_{11}, and its inverse, Z_{21}, then take the form shown in Fig. 34. In the inverse network, Z_{21}, the shunting resistance, R_{11}, becomes a series resistance, R_{21}; the series condenser, C_{11}, becomes a shunt inductance, L_{21}; while the parallel resonant circuit, C_{12} and L_{11}, becomes a series resonant circuit, L_{22} and C_{21}. The degree of perfection with which a given loss-frequency characteristic can be matched by such an equaliser depends upon the number of coils, condensers or resistances it is considered economical to use.

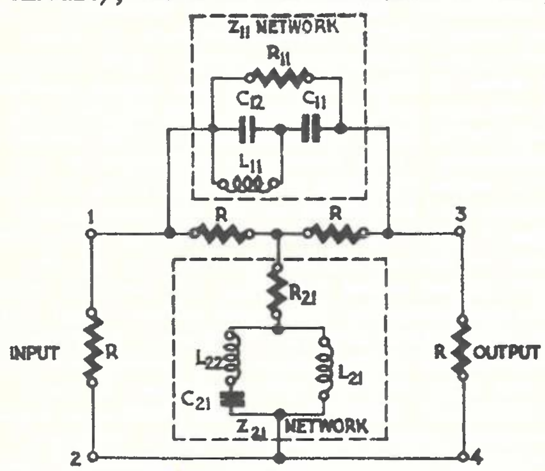


FIG. 34. EQUALISER FOR 3-CHANNEL SYSTEM.

Another general application of the Bridged T equaliser is in equalising lines for programme transmission. The loss on these circuits varies with frequency as on all circuits, and, in view of the wide voice-frequency band transmitted, it is apparent that attenuation equalisers must be used to provide a uniform loss-frequency characteristic over the frequency band transmitted. The principles involved are, of course, the same as those just considered, although the details of design may be somewhat different.

16. TEST QUESTIONS.

1. What is meant by -
   (a) a low-pass filter,
   (b) a high-pass filter, and
   (c) a band-pass filter?
2. What are the desirable characteristics of any filter?
3. What are the characteristics of prototype T and π Section L.P. and H.P. filters?
4. What is the purpose of m-derived filter sections in a composite filter?
5. How is an L.P. filter arranged to present a constant impedance over its pass band?
6. What is the purpose of a compensating filter in a 3-channel carrier system?
7. What is the purpose of an equaliser in a 3-channel carrier system?

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

LONG LINE EQUIPMENT I.

CROSSTALK, DERIVED CIRCUITS AND LOADING.

CONTENTS.

1. INTRODUCTION.

1.1 The intelligibility of telephone conversations depends not only on reducing the attenuation and distortion produced by the characteristics of telephone lines, as discussed in Paper No. 1, but also on the absence of noise and crosstalk introduced along a line from neighbouring circuits. This Paper will deal with the causes of such noise and crosstalk, and the line practices employed to reduce them.

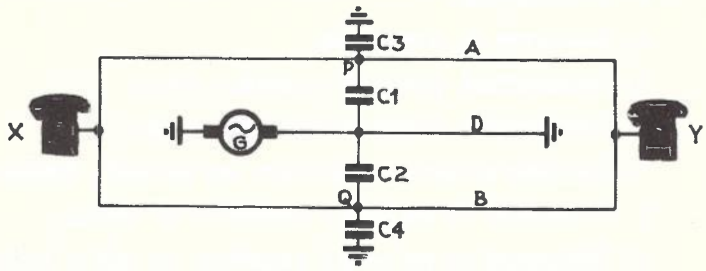
1.2 This Paper also deals with derived circuits and the practice of loading cables.

2. INDUCTIVE CO-ORDINATION.

2.1 If each telephone circuit were completely isolated from all other electrical circuits, no potentials, other than those deliberately introduced for the purposes of telephone transmission, would be present in any telephone circuit. In practice, however, telephone circuits are rarely entirely isolated, as they are in close proximity to other telephone circuits and to other electrical circuits, such as power lines. All electrical circuits set up fields which extend into space, and these fields cause interference in the form of noise and crosstalk in neighbouring circuits. The fields set up by an electrical circuit are electric and magnetic in nature, and, unless circuits (particularly those near to one another) are properly co-ordinated, these fields cause interference by induction. The proper co-ordination of the circuits to minimise interference has been called Inductive Co-ordination. Before proceeding with the methods used to minimise interference, a knowledge of how the two fields produce interference is necessary.

3. INTERFERENCE CAUSED BY ELECTRIC FIELD.

3.1 The potential difference between wires in neighbouring circuits and between those wires and the earth sets up an electric field, because of the dielectric properties of the insulating medium separating the wires. Thus, a number of capacities exist as shown in Fig. 1, which contains a single wire disturbing circuit D (to which an alternating voltage is applied from a generator G) and a metallic telephone circuit, the A and B sides of which connect two telephones, X and Y. If D is equidistant from the A and B sides of the telephone circuit, and the A and B sides are equidistant from the earth, then C1 = C2 and C3 = C4. Equal currents will flow from the generator G to earth via C1 and C3, and via C2 and C4, as well as to earth at the distant end of D, producing equal voltage drops across their reactances. Thus, point P will exhibit the same alternating potential as point Q, and no current from G will flow through the telephones X and Y, which connect these two points.

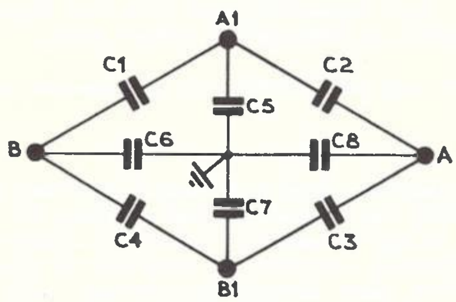


CAPACITANCES EXISTING BETWEEN NEIGHBOURING LINES.

FIG. 1.

3.2 If, now, D is not equidistant from the A and B sides of the telephone circuit and/or the A and B sides of the telephone circuit are not equidistant from the earth, then C1 ≠ C2 and/or C3 ≠ C4. Unequal currents, therefore, flow from the generator G to earth via C1 and C3, and via C2 and C4, producing unequal voltage drops across their reactances. Thus, an alternating difference in potential will exist between points P and Q, resulting in disturbing currents flowing through the telephones at X and Y, which connect these two points. If D is carrying a telephone conversation, the result is crosstalk which can be audible if the capacity unbalance is too large, whilst, if D is a power line, the result is noise, not only at the 50 c/s fundamental frequency in the power circuit, but also at harmonics of this frequency, as harmonics are almost invariably present in any power circuit. The same condition will exist if the insulation resistances between D and the A and B sides of the telephone circuit, or between the A and B sides of the telephone circuit and the earth, become unbalanced.

3.3 When the disturbing circuit is metallic, that is, two-wire, a complicated network of capacities exists which can be simplified into a bridge circuit, as shown in Fig. 2.



CAPACITY BRIDGE.

FIG. 2.

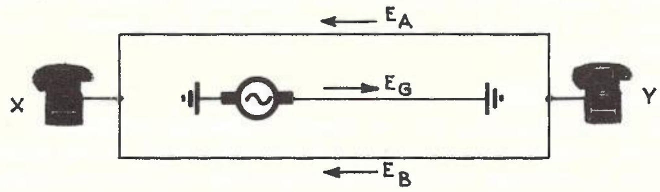
The bridge circuit is a cross section of two circuits, A and B being the two wires of one circuit and A1 and B1 the two wires of the other circuit.

If the four wires are symmetrically disposed, then each of the four capacities between the wires will be equal, these being C1, C2, C3 and C4 in Fig. 2. This, however, is not a balanced condition, as the capacities to earth provide parallel capacities across the wire to wire capacities. For example, C5 and C6 are in parallel with C1, C6 and C7 are in parallel with C4, and so on. Not only must the wire to wire capacities be equal, but also the wire to earth capacities, as in the single wire disturbing circuit case. Under this condition, wires A1 and B1 will act as null points of a balanced bridge when a voltage is applied across A and B, and wires A and B will act as the null points when a voltage is applied across A1 and B1, and no interference results.

/ 4.

4. INTERFERENCE CAUSED BY MAGNETIC FIELD.

4.1 The alternating current flowing through D of Fig. 1 produces an alternating flux which links the two sides of the telephone circuit, E_A of Fig. 3 being the voltage induced across the A side and E_B the voltage induced across the B side.



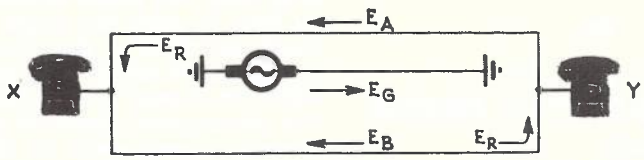
$E_A = E_B$

EQUAL VOLTAGES INDUCED BY MAGNETIC FIELD.

FIG. 3.

The directions indicated in Fig. 3 apply for one half-cycle, being reversed during the other half-cycle. When D is equidistant from A and B, then the amount of flux linking the A side equals that linking the B side, so that $E_A = E_B$, leaving no resultant voltage to send a disturbing current around the circuit and through the telephones at X and Y.

4.2 When D is not equidistant from A and B, a resultant voltage sends current through X and Y. The position is indicated in Fig. 4, where D is nearer to A than it is to B.



$E_A > E_B$

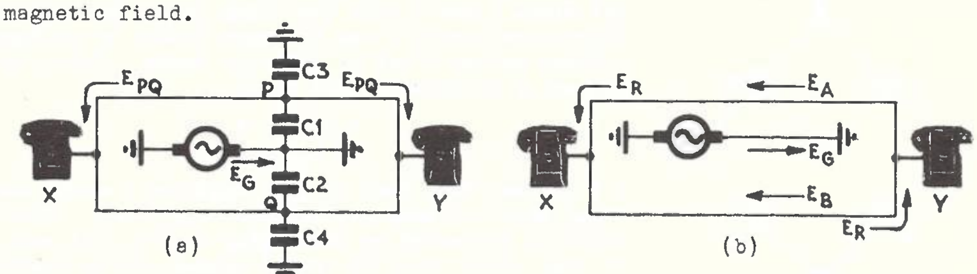
UNEQUAL VOLTAGES INDUCED BY MAGNETIC FIELD.

FIG. 4.

Here, the amount of flux linking A will be greater than that linking B, so that E_A is greater than E_B, leaving a resultant voltage E_R to send a disturbing current in the direction indicated. During the other half-cycle developed by G, all directions are reversed, so that a disturbing alternating current of the same frequency as that developed by G will pass through the telephones X and Y.

5. COMBINED EFFECTS OF ELECTRIC AND MAGNETIC INTERFERENCE.

5.1 Fig. 5 shows what can happen when the interferences caused by the electric and magnetic fields are considered together rather than separately. Here, again, D is nearer to A than to B, Fig. 5a indicating the effect of the electric field during the half-cycle of voltage developed by G, and Fig. 5b indicating the effect of the magnetic field.



COMBINED EFFECTS OF ELECTRIC AND MAGNETIC FIELDS.

FIG. 5.

As D is nearer to A than to B, C1 will be greater than C2, therefore the reactance of C1 will be smaller than that of C2. The voltage drop across C1 will be smaller, therefore, than that across C2, resulting in point P exhibiting a potential above, or positive to, that exhibited by point Q for the half-cycle indicated in Fig. 5. Combining Figs. 5a and 5b, it will be seen that the resultant voltage due to the magnetic field aids the voltage due to the electric field at X, whilst opposing it at Y. A little consideration will show that, during the other half-cycle, the same conditions exist but with all voltages reversed.

/ 5.2

5.2 Thus, cases can arise where one end of a circuit is noisy or where the crosstalk level is high, whilst the other end of the circuit is silent. Under such conditions, it is necessary to specify the end of a circuit to which a crosstalk level refers. In Fig. 5, the crosstalk at X is referred to as "near-end" crosstalk, because it is the end of the disturbed circuit nearest the end of the disturbing circuit to which the disturbing source of supply is connected. The crosstalk at Y is termed "far-end" crosstalk for the opposite reason.

6. PRINCIPLES OF CROSSTALK REDUCTION IN CABLES.

6.1 There are a number of ways of eliminating or, at least, reducing crosstalk. As metallic, that is, two-wire, circuits are almost exclusively used in telephone transmission, one method is to arrange the paralleling wires in such a configuration that the effect of the field of one pair will be the same at both wires of the other pair, and vice versa, thus leaving no resultant voltages to produce interference.

6.2 Fig. 6 shows two possible ways to effect such a non-inductive configuration.

(a) (b)

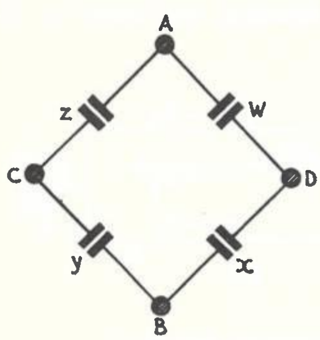
STAR QUAD (a) AND MULTIPLE TWIN (b) FORMATION.

FIG. 6.

It is not possible, because of practical difficulties, to arrange aerial circuits in this manner. Cables, however, can be manufactured with each two pairs arranged as in Fig. 6, each two pairs being known as "QUAD". When the quads are arranged as in Fig. 6a, the cable is termed a Star Quad cable. The arrangement shown in Fig. 6b is closely approximated by the Multiple Twin cable. The Star Quad cable is used almost exclusively, having superseded the Multiple Twin type about 1935.

6.3 In the manufacture of Star Quad cable, it is possible to restrict the capacity unbalances existing between the four wires of a quad to only a minimum value; manufacturing difficulties preclude the complete elimination of unbalance. It is, therefore, necessary to joint together the short lengths of cable which go to make up the whole length of a long cable, in such a manner that the over-all capacity unbalance is reduced to a predetermined minimum.

6.4 To illustrate the principle, Fig. 2 of this Paper is redrawn as Fig. 7. The capacities C5, C6, C7 and C8 of Fig. 2 are included in the capacities w, x, y and z of Fig. 7. This is possible because C5 and C6 shunt C1 in Fig. 2, C6 and C7 shunt C4, and so on. Wires A and B form one pair of the quad and wires C and D form the other pair. For any voltage introduced across A and B, there should be no resultant voltage across C and D, and vice versa. Thus, the following proportion must exist -



CAPACITY BRIDGE.

FIG. 7.

$$w : x :: z : y$$

$$\text{or } w.y = x.z$$

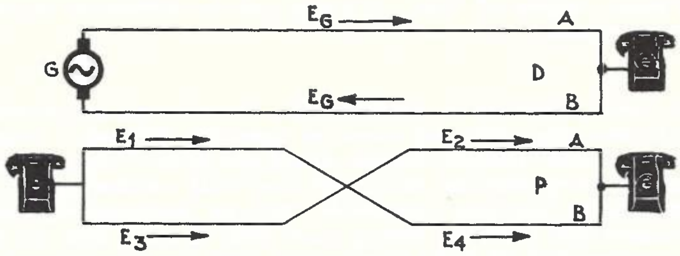
6.5 Assume, now, that in one length of a cable the capacities are measured and that $w.y > x.z$. This unbalance could be corrected by increasing x or z, and this is done in some cases (such as loaded cables, to be dealt with later). Where it is possible to connect together adjacent lengths of cable, the quad discussed above would be connected to a quad in the adjacent length whose measured capacities are such that $w.y < x.z$ by an amount approximately equal to that by which $w.y > x.z$ in the first quad. By providing balancing condensers over each predetermined length of cable in the case of loaded cables, the whole length of cable will exhibit zero unbalance when jointed right through. In the other case, the over-all unbalance is reduced to a minimum, which can be corrected at the end of predetermined long lengths by added capacity in the form of balancing condensers.

The above gives merely the principle used. A later Paper deals more fully with the subject and with the measuring technique employed.

/ 7.

7. PRINCIPLES OF CROSSTALK REDUCTION ON AERIAL LINES.

7.1 Because of practical difficulties, it is not possible to arrange aerial wires in the manner shown in Fig. 6. The scheme used on aerial circuits is one of transposition, and Fig. 8 shows the idea.



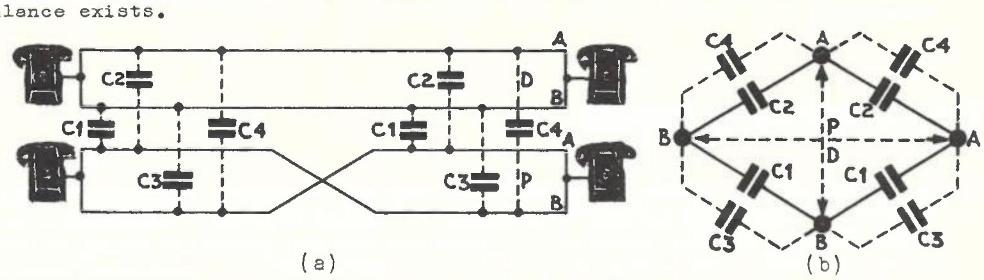
PRINCIPLE OF TRANSPOSITION.

FIG. 8.

A cross-over of the two sides of circuit P, which parallels another circuit D to which an alternating voltage is applied, is made half-way along the length of the line. The direction of the voltage acting in D at some instant is indicated.

7.2 Considering the magnetic interference first, the magnetic interference produced in P by the A side of D will oppose that produced by the B side, because the fields produced by the A and B sides of D oppose. However, as the B side of D is nearer to P than is the A side, the voltage induced across P by the B side of D will be greater than the voltage induced by the A side, producing resultant voltages across the two sides of P in the directions indicated. As the transposition is in the centre of P, then E1 = E2 and E3 = E4, and as E1 + E4 acts in opposition to E3 + E2, then there will be zero voltage acting around P and, therefore, no interference due to the magnetic field.

7.3 Turning now to the wire to wire capacities, the unbalancing of which causes the electric interference, an examination of Fig. 9 will show that a state of capacity balance exists.



CAPACITANCES PRESENT ALONG TRANSPOSED LINE.

FIG. 9.

The actual wire to wire capacities are shown in Fig. 9a whilst an equivalent network is shown in Fig. 9b. Equal capacities carry identical designations, for example, the capacity from one half of the B side of D to one half of the B side of P is designated C1, as is the equal capacity from the other half of the B side of D to the other half of the A side of P. From Fig. 9b it will be seen that a balanced bridge exists both for voltages applied across the A and B sides of D or the A and B sides of P.

7.4 By transposing D and leaving P wired straight through, a similar non-inductive stage will exist between the two circuits. A transposition at the same point in both

/ circuits,

circuits, however, will obviously have no effect in reducing interference.

7.5 Whilst a single transposition, as discussed, is effective in limiting crosstalk in a relatively short length of line, it would not be effective in the case of a long line for two reasons.

7.6 In the first place, because of attenuation, the voltage and current at the energised end of a line are many times as great as near the far end. Thus, the crosstalk voltages and currents induced on the energised side of the transposition will be greater than those on the far side, and they will neutralise in part only and not wholly. As regards near-end crosstalk, this is increased by the fact that the induced voltages and currents coming back from the far side of the transposition are necessarily attenuated to a greater degree than those coming back from the near side. On the other hand, far-end crosstalk is reduced because the slightly higher induced voltages and currents on the near side of the transposition are attenuated more in reaching the far end of the circuit than are those induced on the far side.

7.7 In the second case, the phase shift along the line will mean that the line may be one or a number of wavelengths long at higher frequencies. Thus, not only will the crosstalk voltages and currents induced along the line decrease along the line due to attenuation, but they will also change in magnitude and direction over the transposition sections due to the phase shift. Thus, if one transposition section has a maximum crosstalk voltage induced across it at some instant, that voltage cannot be neutralised by some other section across which the crosstalk voltage is perhaps zero or opposite in phase. It is necessary, therefore, that the transpositions be more frequent for higher frequencies, so that the crosstalk voltages and currents produced in one transposition section can be almost neutralised by approximately equal voltages and currents in the adjacent section.

7.8 In some cases (for example, 140 kc/s, the highest frequency allocation on Type J carrier telephone systems) transpositions as close as every second pole may be necessary.

8. EFFECT OF REFLECTION ON CROSSTALK.

8.1 The transposition scheme outlined above does not eliminate crosstalk - this can only be done by employing an infinite number of transpositions. Similarly, in cables, manufacturing and installation difficulties prevent perfectly balanced quads from being obtained. As described in Long Line Equipment III, unbalance measurements are made on cables when they are laid down and the unbalance correctives applied limit the unbalance, so that, as with a practical transposition scheme, the crosstalk is below the level of audibility when normal voltages are employed. As discussed previously, reflection produces waves whose amplitude is the vector sum of the reflected and incident waves. This means that reflection can increase crosstalk by increasing the amplitude of the voltages and currents in a circuit in which reflection takes place, that is, a circuit which is incorrectly terminated or has any impedance irregularities. Thus, crosstalk is an added reason why circuits should be correctly terminated and uniform in construction.

/ 9.

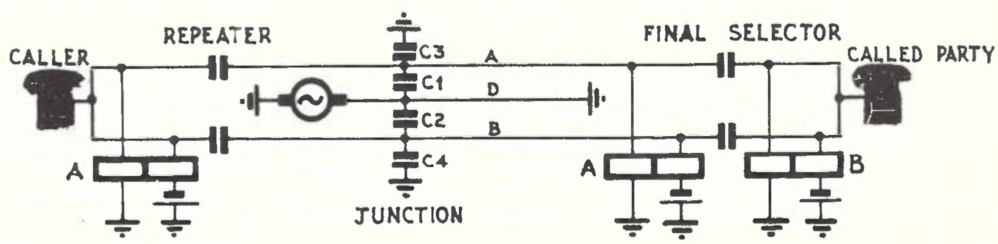
9. EFFECT OF BALANCE.

9.1 Whilst a circuit may be perfectly transposed, interference can still be produced if the linear impedance and metallic impedances to earth are not balanced. A metallic connection to earth exists because all C.B. manual and automatic telephones use an earthed battery for supplying transmitter battery feed current.

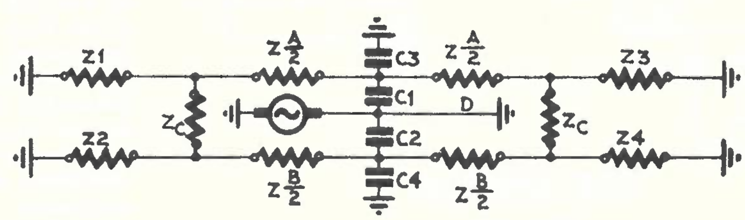
9.2 Fig. 10a shows the through connection between two subscribers connected to different automatic exchanges connected together via a junction, together with a disturbing circuit D equidistant from the A and B sides of the junction.

Fig. 10b is the equivalent circuit with Z1 and Z2 replacing the repeater A relay windings, Z3 and Z4 replacing the final selector A and D relay windings, Z_C replacing the calling and called parties' lines and telephones and Z_A and Z_B replacing the A and B sides of the junction respectively. Currents will now flow to earth via Z1, Z2, Z3, Z4, $Z\frac{A}{2}$ and $Z\frac{B}{2}$ from G, as well as via C3 and C4.

If Z1 $\neq$ Z2, Z3 $\neq$ Z4 and $Z\frac{A}{2} \neq Z\frac{B}{2}$, unequal currents will flow to earth through these impedances, producing unequal voltage drops across them which result in disturbing currents through the two telephones. Thus, accurate linear balance must be maintained, as well as a balance to earth.



(a)



(b)

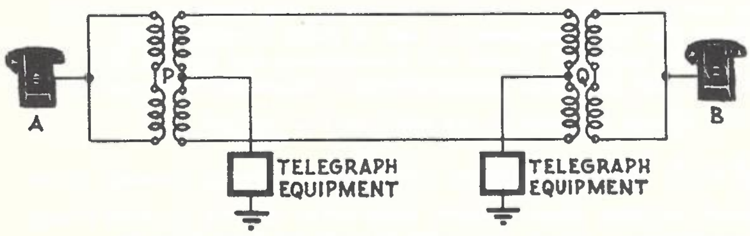
METALLIC IMPEDANCES TO EARTH ON TELEPHONE CONNECTION.

FIG. 10.

10. DERIVED CIRCUITS.

10.1 Where accurately balanced lines are available, additional channels can be derived without having to provide further line plant. Such circuits are termed "Cailho" and "Phantom" circuits. The term "Cailho" usually refers to a derived circuit using an earth return, and the term "Phantom" refers to a completely metallic derived circuit.

10.2 Cailho Circuits. Cailho circuits are generally telegraph circuits, the sensitivity of a telephone receiver precluding the use of earth return telephone circuits because of noise produced by the slight changes in potential which are continually taking place between different points on the earth's surface. These changes in potential are not great enough to produce enough current to operate relays but will make a telephone circuit extremely noisy. Also, it is not possible to transpose a single wire, so that crosstalk would be excessive between neighbouring single wire lines. Fig. 11 shows the principle of the cailho circuit to derive a telegraph channel from an existing physical telephone circuit.

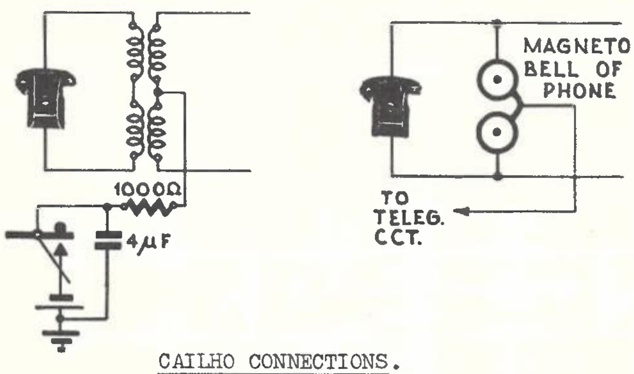


.CAILHO CIRCUIT.

FIG. 11.

No interference between the telephone and telegraph circuits will arise, provided the windings of the transformers are accurately differential and the two sides of the physical circuit are balanced as regards both linear impedance and impedance to earth. Under this condition, telegraph signals divide equally at the centre point of the line windings of the transformers to flow through the two halves in opposite directions, producing equal fluxes which neutralise to leave zero resultant flux.

Thus, no voltages can be induced across the transformer windings connected to the telephones by currents from the telegraph equipment passing through the line windings. If the transformer windings or the two sides of the line are unbalanced as regards either linear impedance or impedance to earth, the telegraph currents do not divide equally. Therefore, the two fluxes produced by the line windings of the transformers do not neutralise and interference arises between the circuits, because the resultant flux induces voltages across the windings of the transformers to which the telephones are connected. The unbalance may also cause the telephone circuit to interfere with the telegraph circuit. Under the balanced condition, point P will exhibit the same potential as point Q when A is speaking to B, or vice versa. In the unbalanced condition, however, point P will exhibit a different potential from point Q, so that, whilst speech will be practically unaffected because of the high impedance of the telegraph equipment, 16 cycle ringing current may interfere with the telegraph equipment.



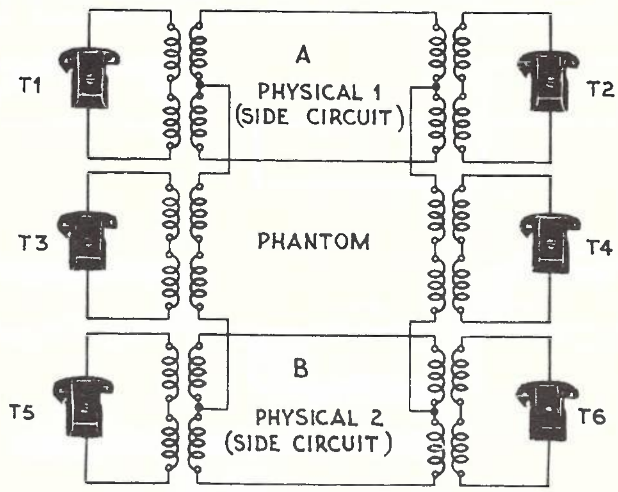
CAILHO CONNECTIONS.

FIG. 12.

In some cases, it is possible to utilise existing magneto bells on a trunk line in order to provide suitable centre points for connecting a cailho telegraph circuit. This is shown in Fig. 12, which contains both the transformer and bell connections to show the similarity.

10.3 Phantom Circuits. The principle used in the cailho circuit is used in the phantom circuit, except that the metallic return is supplied by another physical circuit as shown in Fig. 13. Here, again, the circuits must be accurately balanced as regards both linear impedance and impedance to earth, and the transformers must be accurately differential.

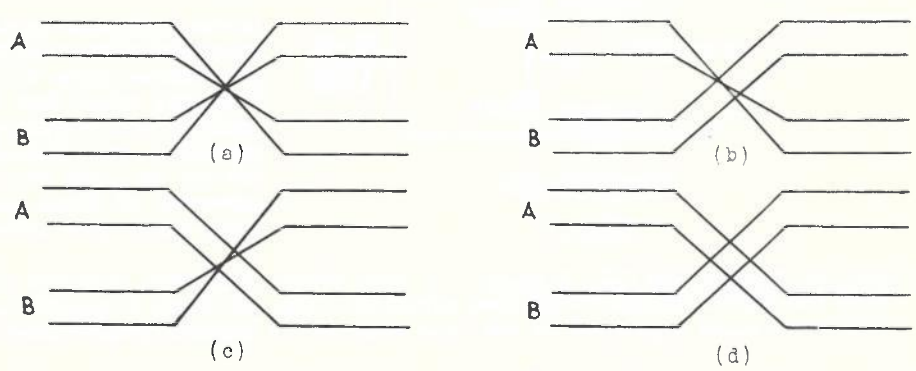
In Fig. 13, the two physical circuits A and B are usually referred to as "side" circuits. Whilst the transformers in the phantom circuit are not necessary for successful operation, they are usually included so that an unbalanced physical circuit will not upset the balance of the phantom when such a circuit is connected to the phantom.



PHANTOM CIRCUIT.

FIG. 13.

10.4 Phantom Transpositions. As each side circuit represents one side of a phantom circuit, it will be necessary to transpose the side circuits of a phantom as well as the two wires of each side circuit. Fig. 14 shows the four types of transpositions necessary to meet all conditions. Fig. 14a shows a transposition in the phantom as well as the side circuits; Fig. 14b shows a transposition in the phantom and the side circuit A; Fig. 14c shows a transposition in the phantom and side circuit B; and Fig. 14d shows a phantom transposition only.

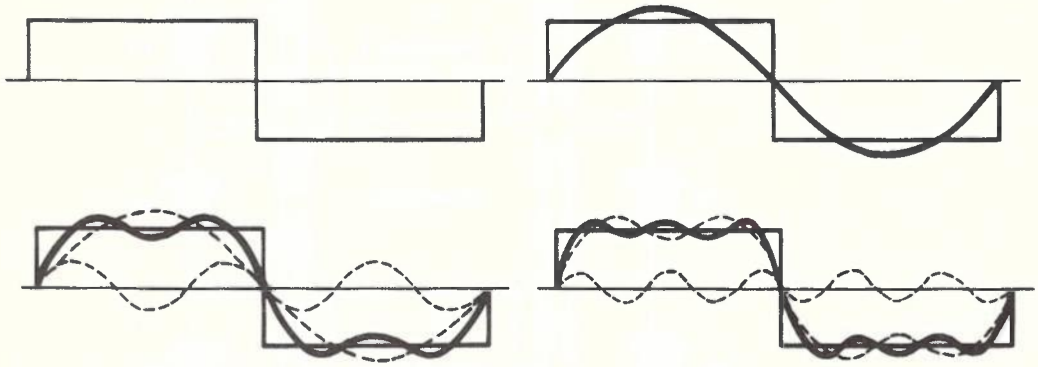


PHANTOM TRANSPOSITIONS.

FIG. 14.

10.5 Composite Circuits. Composite circuits are provided to derive two telegraph channels from a physical telephone circuit. The composite, or C.X., set uses crude low and high pass filters to separate the telegraph and telephone signals simultaneously sent over a physical telephone circuit. Discrimination on a frequency basis can be used, because any telegraph signal can be resolved into a fundamental frequency determined by the signalling speed plus a number of odd harmonics which, when added to the fundamental, produce the "square-topped" telegraph signal.

Fig. 15 shows the idea applied to a double current telegraph signal. In order to ensure reasonable signal formation, the third harmonic of the fundamental must be passed into the telegraph equipment so that the signalling speed over a C.X. circuit is fixed by the cut-off frequency of the low-pass filter used to pass the low frequency telegraph signals and block the high frequency telephone signals.

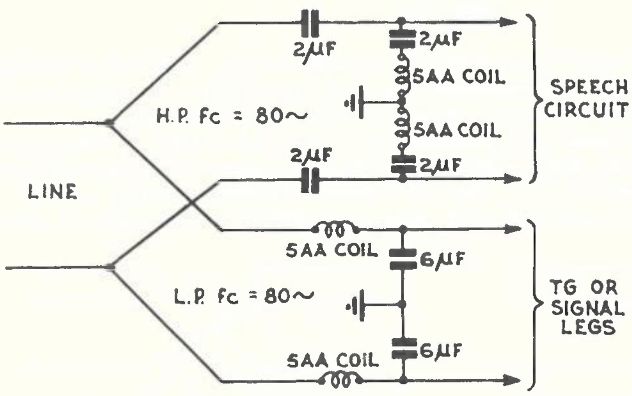


FORMATION OF SQUARE-TOPPED SIGNAL.

FIG. 15.

This cut-off frequency is about 80 c/s in a C.X. set, which fixes the upper limit at 75 c/s, producing a fundamental frequency of 25 c/s. Therefore, the signalling speed over the telegraph circuit is limited to 50 bauds. The high-pass filter, therefore, has a cut-off frequency of 80 c/s, in order to pass the higher frequency telephone signals and reject the lower frequency telegraph signals. Fig. 16 shows the arrangement of one terminal of a composited line.

As the high-pass filter in the telephone channel will not pass frequencies below 80 c/s, some frequency other than 16 c/s is required for signalling over such circuits. The frequency used is 135 c/s or 1,000 c/s, the operation of these ringing circuits being dealt with later.



COMPOSITE SET.

FIG. 16.

10.6 Combined Phantom and C.X. Circuit. In many cases, composite circuits are derived from the side circuits of phantom circuits. Fig. 17 shows a typical circuit, the C.X. sets being connected on the line side of the phantom transformers.

10.7 Fig. 18 shows, in pictorial form, the various methods of increasing the efficiency of line plant as far as telephone channels are concerned.

/ Fig. 17.

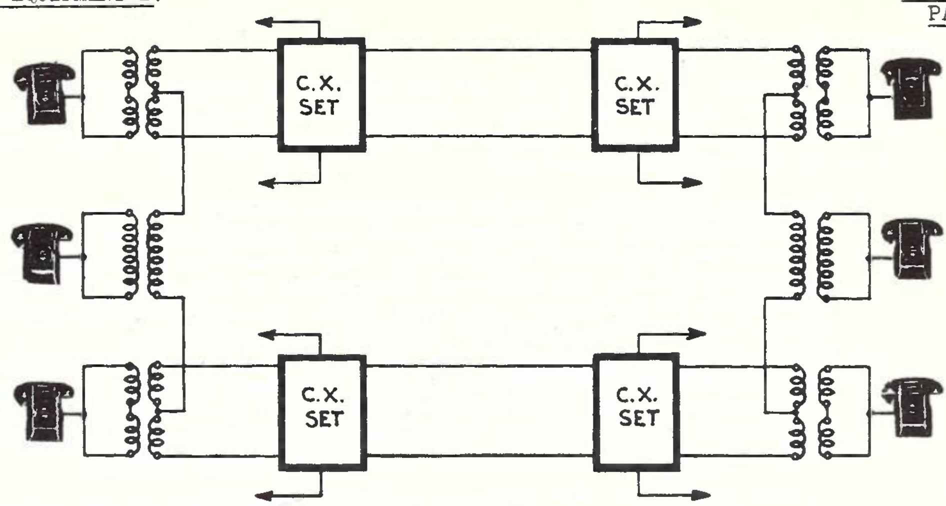


FIG. 17. PHANTOM AND COMPOSITE CIRCUITS COMBINED.

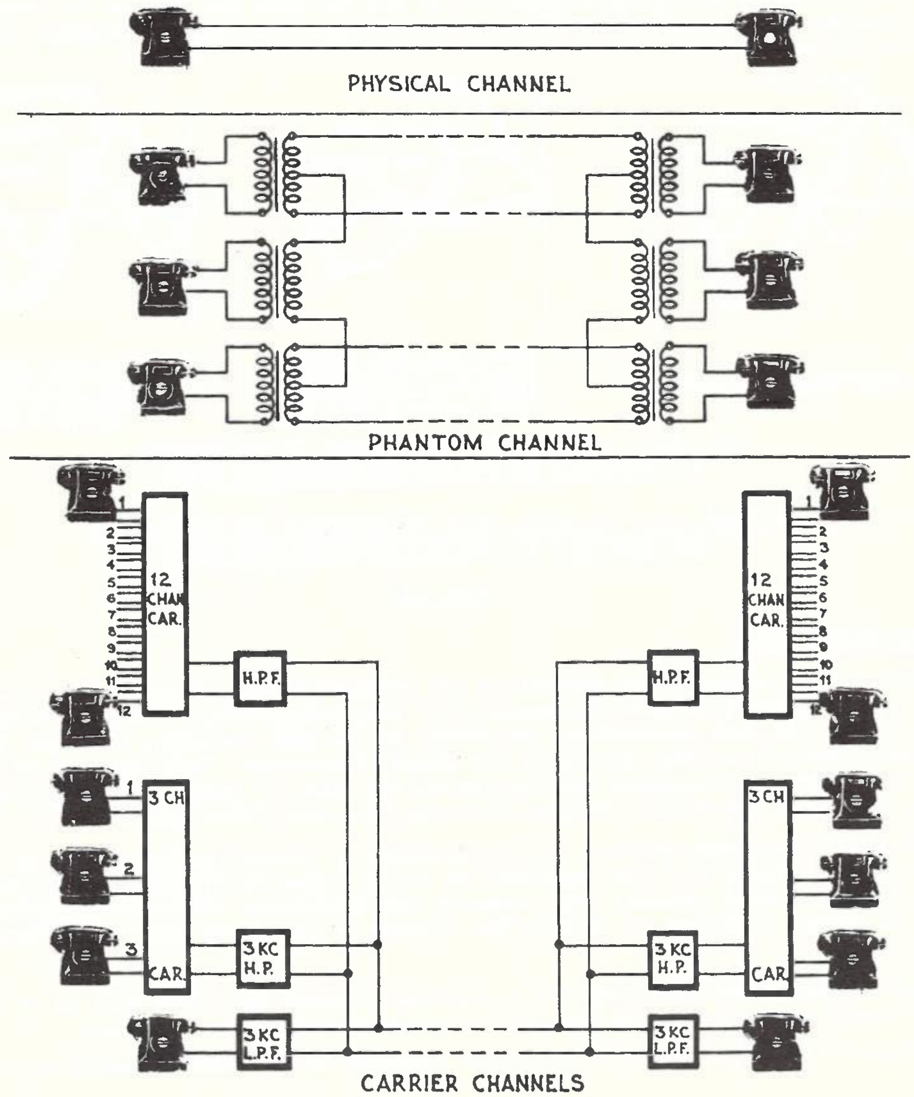


FIG. 18. TELEPHONE CHANNELS.

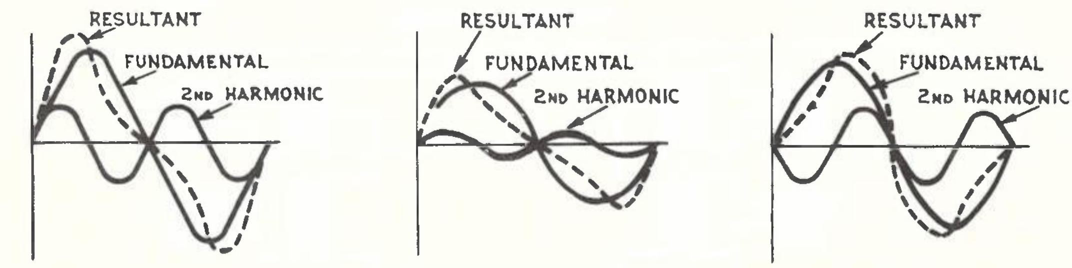
11. DISTORTION PRODUCED BY LINES.

11.1 When the various constants of a line are considered in their effect on a complex communication signal, two types of distortion are produced. By distortion is meant any alteration in the shape of a signal passing along a line. The two types of distortion produced are -

(i) Frequency Distortion. As the attenuation constant increases as the frequency rises, low frequencies will be transmitted with less attenuation than high frequencies. This means that the amplitude relations between the different frequencies at the output will differ from those at the input. In other words, the line discriminates in favour of low frequencies and against high frequencies.

(ii) Phase, or Delay, Distortion. The different frequencies travel along the line at different velocities, so that, although they may be applied to the input of a line simultaneously, they will not arrive at the output together. This means that the phase relations between the different frequencies at the output differ from those at the input.

11.2 Fig. 19 shows the effects of frequency and phase distortion on the shape of a signal containing a fundamental and a second harmonic. Phase distortion is not so serious as frequency distortion, as the ear does not seem to be sensitive to changes in the phasing of the different frequencies which go to make up a complex sound wave.



(a) Sending End of Line.

(b) Receiving End Showing Frequency Distortion With Respect to (a).

(c) Receiving End Showing Phase Distortion With Respect to (a).

(a) AND (b). RELATIVE AMPLITUDES OF FUNDAMENTAL AND HARMONIC AT SENDING AND RECEIVING ENDS.

(a) AND (c). RELATIVE PHASE RELATIONSHIP OF FUNDAMENTAL AND HARMONIC AT SENDING AND RECEIVING ENDS.

FIG. 19.

12. DISTORTIONLESS LINE.

12.1 Both frequency and phase distortion are produced by the reactive nature of a line. If a line could be made non-reactive at all frequencies, all frequencies would undergo the same attenuation and travel at the same velocity, that is, the line would produce neither frequency nor phase distortion. This would be so because, no matter what frequency was applied to the line, it would encounter a pure resistance, the effect of which does not vary with frequency.

12.2 To make a line non-reactive, it is necessary to neutralise the effect of the capacity inherent in the construction of all lines by the inductance, which is likewise always present. The capacity and inductance cannot be eliminated, so the method of attack is to make the effects of one neutralise the effects of the other.

12.3 The effect of the inductance in the series elements of a line will be to make the current through them lag the applied voltage by an angle θ so that $\tan\theta = \frac{\omega L}{R}$. The effect of the shunt capacity in the shunt elements will be to

/ make

make the current lead the voltage applied to them by an angle ϕ so that $\tan\phi = \omega CRS = \frac{\omega C}{G}$, G being the shunt conductance or leakance which is the reciprocal of the shunt or insulation resistance, RS. This is shown in Fig. 20.

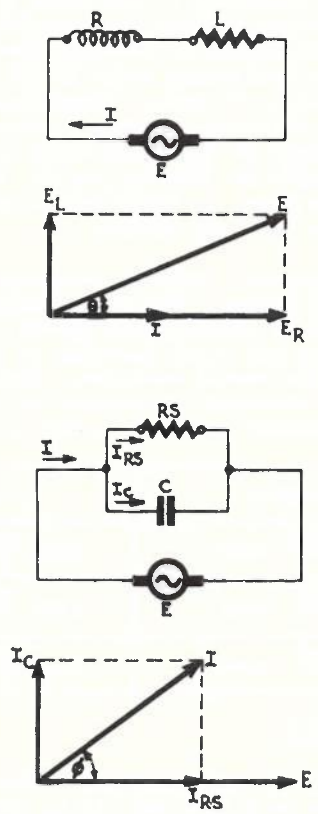
12.4 If $\theta = \phi$, there would be zero phase displacement and the line would be non-reactive. For θ to equal ϕ, $\tan\theta$ must equal $\tan\phi$, that is -

$$\frac{\omega L}{R} = \frac{\omega C}{G} \text{ or } R\omega C = G\omega L \text{ or } RC = LG \text{ or } \frac{R}{L} = \frac{G}{C}.$$

In all telephone transmission lines, $\frac{R}{L}$ is much greater than $\frac{G}{C}$, as the inductance for the lines is very small. For example, 40 lb. cable has the following primary constants -

$R = 44$ ohms per loop mile,
$L = 0.001$ henry per loop mile,
$G = 1 \times 10^{-6}$ mhos per loop mile, and
$C = 0.0625 \times 10^{-6}$ farad per loop mile,

so that $\frac{R}{L} = 44{,}000$ and $\frac{G}{C} = 16$.



$$\tan\theta = \frac{E_L}{E_R} = \frac{I\omega L}{IR} = \frac{\omega L}{R}$$

$$\tan\phi = \frac{I_C}{I_{RS}} = \frac{\frac{E}{\frac{1}{\omega C}}}{\frac{E}{RS}} = \frac{E\omega CRS}{E} = \omega CRS$$

$$RS = \frac{1}{G}$$

$$\therefore \tan\phi = \frac{\omega C}{G}$$

VECTOR DIAGRAMS OF SERIES AND SHUNT ELEMENTS OF LINE.

FIG. 20.

In order to approach this distortionless condition, that is, make $\frac{R}{L} = \frac{G}{C}$, it is necessary to modify artificially one of the primary constants. The most convenient one to manipulate is the inductance, and this is done by adding inductance to the line, the process being known as "loading".

/ 13.

13. LOADING.

13.1 There are two methods used in loading practice, "Continuous Loading" and "Lumped Loading".

13.2 Continuous Loading. This process involves wrapping the line in a tape of some magnetic material. This treatment is expensive, and the amount of inductance which can be economically provided is small.

13.3 Lumped Loading. This process involves introducing inductance coils at strictly equal intervals along the length of the line. Under these conditions, the performance of the line will be modified from that obtained when the inductance is evenly and continuously distributed. This is because a lumped loaded line constitutes a series of low-pass filter sections, having the lumped inductances of the loading coils as the series inductance and the coil capacity, together with the lines distributed capacity, as the shunt capacity. Loaded circuits, therefore, display a definite cut-off frequency, this being determined by the magnitude of the inductance used and the spacing of those inductances. Thus, the value of the inductance used, together with the spacings, will be determined by the frequencies to be sent over the circuit. Voice frequency circuits are loaded with 88 millihenry coils at 6,000 feet intervals. This produces a cut-off frequency of between 3.5 kc/s and 4 kc/s, depending on the gauge of the conductor used in the cable being loaded. These loading figures are unsuitable for circuits which are required to transmit carrier frequencies or to relay broadcast programmes. "Carrier Loading" employs 3.5 millihenry coils at 750 feet intervals, producing a cut-off frequency of approximately 54 kc/s. "Programme Loading" employs 14 millihenry coils at 3,000 feet intervals, giving a cut-off frequency of at least 12.5 kc/s, depending on the gauge of the cable being loaded. Aerial circuits are not loaded, as their phase angle is normally small. This is shown in Table 2 of Paper No. 1, which indicates that, in their normal condition, aerial circuits are much more nearly non-reactive than are cables.

13.4 Loading also raises the characteristic impedance. This is extremely useful where aerial and cable sections are connected in tandem. The characteristic impedances of cables are much lower than those of aerial lines, but a smooth, continuous circuit can be provided by suitably loading the cable sections to bring their characteristic impedances up to those of the aerial sections. Table 1 shows how V.F. loading increases the characteristic impedance of cables and reduces the phase angle, as discussed above. The frequency employed is 800 c/s.

Type of Cable.	Z_0 Unloaded.	Z_0 Loaded.
10 lb. S.Q. Cable	$366 \overline{\backslash 41^\circ 38'}$ ohms	$1085 \overline{\backslash 11^\circ 3'}$ ohms
20 lb. S.Q. Cable	$515 \overline{\backslash 43^\circ 16'}$ ohms	$1121 \overline{\backslash 5^\circ 40'}$ ohms
40 lb. S.Q. Cable	$683 \overline{\backslash 44^\circ 6'}$ ohms	$1113 \overline{\backslash 3^\circ}$ ohms

TABLE 1.

13.5 This increase in Z_0 also produces the highly desirable effect of decreasing the attenuation. If the same power is applied to a line and its Z_0 increased, the input current will decrease. As the power loss along the line is proportional to the square of the current flowing along it, and as Z_0 decreases this current, then the power loss will become smaller as Z_0 is increased, meaning that the attenuation is decreased. Loading for voice frequency purposes is used mainly for this reason. At carrier frequencies, the aim of loading is not so much to reduce the attenuation as to make the characteristic impedance independent of frequency, this being achieved because the line behaves largely as though non-

/ reactive.

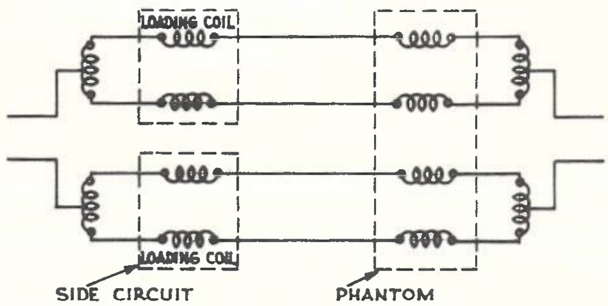
reactive. Table 2 shows how the attenuation constant is reduced by V.F. loading. The change in the phase constant is also included in Table 2. Loading will increase the phase constant, because increasing the series inductance will decrease the velocity of propagation and, therefore, increase the phase constant.

Type of Cable.	Attenuation Per Mile.		Phase Constant.	
	Unloaded	Loaded	Unloaded	Loaded
10 lb. S.Q. Cable	1.56 db	0.803 db	10°34'	33°28'
20 lb. S.Q. Cable	1.021 db	0.386 db	7°6'	26°48'
40 lb. S.Q. Cable	0.703 db	0.202 db	5°12'	26°36'

TABLE 2.

13.6 Loading Coils for Phantom Circuits. Coils for phantom loading usually have lower inductance values than side circuit coils, but they must have four windings.

Fig. 21 shows the connections of a loading point in a phantom group for the side circuits and the phantom.



LOADING COILS FOR PHANTOM CIRCUITS.

FIG. 21.

Phantom loading coils are connected in such a manner as to be non-inductive to the currents circulating in either side circuit, but inductive to the currents in the phantom circuit.

13.7 Half-Coil and Half-Section Terminations. When loading is introduced into a network, such as metropolitan junction network, attention must be given to the conditions possible when two loaded junctions are connected together.

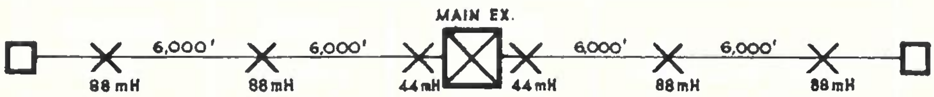
Assume it is decided to load a junction network with 88 millihenry coils at 6,000 feet. It is necessary at a certain point in the network (that is, the main exchange) to space the coils so that, when two junction circuits are connected together, the correct spacing is maintained. This can be done in two ways -

(i) At the main exchange, a half-coil (44 millihenrys) can be used to terminate each junction, whilst 6,000 feet away a full 88 millihenry coil is inserted. This will give conditions of 88 millihenrys at 6,000 feet uniformly, when two junctions are connected together. (Disadvantage. Purchase of special half-value loading coils is necessary.)

/ (ii)

(ii) Place the first coil in each junction only 3,000 feet from the main exchange. When two junctions are interconnected, 6,000 feet spacing is maintained. This method is generally preferred.

The two methods are shown in Fig. 22.



(a) Half-Coil Termination.

(b) Half-Section Termination.

SPACING OF LOADING COILS.

FIG. 22.

14. TEST QUESTIONS.

1. Explain briefly how crosstalk is produced between paralleling telephone lines.
2. Explain how Multiple Twin cable produces a non-inductive relation between the two pairs of a quad.
3. Explain why transpositions on aerial lines have to be closer at higher frequencies than at lower frequencies.
4. Discuss the conditions necessary for preventing interference between a phantom and its side circuits.
5. Describe, with sketches, the use and operation of a composite set.
6. What is meant by "lumped" loading? Discuss the advantages of loading cables.
7. What is meant by "half-coil" and "half-section" terminations?

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

THERMIONIC VALVES.

CONTENTS:

1. INTRODUCTION.

1.1 Thermionic valves are used in all amplifiers and oscillators, and as modulators and demodulators in many cases. By way of an introduction to the study of valves, it is convenient to think of a valve as a fully electrical application of the principle used in the carbon granule telephone transmitter.

1.2 In this type of transmitter the acoustical energy input is not converted into electrical energy - the acoustical energy input controls the amount of electrical energy passing through the transmitter circuit, that electrical energy being supplied from a direct current source. This direct current source can supply an amount of electrical energy far in excess of the amount of acoustical energy input to the transmitter, and this, together with the design of the transmitter, makes the controlled electrical output from a transmitter much greater than the acoustical input. In other words, the transmitter is an amplifier. An amplifier does not violate the energy conservation law as no conversion of energy takes place and an auxiliary supply of energy to be controlled must be provided. In a thermionic valve both controlling and controlled energies are electrical, thus making the device fully electrical and extremely useful for electrical amplification.

1.3 An amplifying device is an important unit in all oscillators, and the telephone transmitter can be used again to illustrate the principle here. The phenomenon of a telephone "howling" is quite familiar. This is caused by the transmitter picking up a sound and amplifying it, portion of that amplified sound being reproduced by the local receiver. If the phase shift in the electrical circuit between transmitter and receiver, plus that of the air path between receiver diaphragm and transmitter diaphragm, is 360^{o} or some integral multiple thereof, the sound produced by the receiver will reinforce the original sound, this action going on indefinitely. The original sound could be removed, as it frequently is, and the action would continue provided the amplification produced by the transmitter exceeds the losses produced around the electrical plus acoustical circuits. Under these circumstances, the circuit is said to be "oscillating." A thermionic valve would produce oscillations if, on an electrical signal being applied to the valve, some of the controlled output is returned to the input in phase with the original input, as in the telephone case.

1.4 The telephone transmitter has already been considered as a modulating device in Long Line Equipment I, Paper No. 2, and this description will not be enlarged on for the purposes of this introduction.

1.5 Before proceeding with a study of thermionic valves it is desirable to discuss briefly the electron theory of atomic structure, as it is on this theory that the operation of valves is based.

2. ELECTRON THEORY.

2.1 Molecules. A molecule is the smallest particle of a substance it is possible to have which retains the properties of the original substance as a whole. The molecules in every different kind of material, therefore, are obviously different from all other molecules, and it is the nature of these molecules which govern what any material really is. While there are countless different kinds of molecules, these molecules, in themselves, are compounded from smaller particles known as atoms.

2.2 Atoms. Atoms are made up of minute particles of positive and negative electricity known as Protons and Electrons. There are approximately 92 different kinds of atoms, each one being a different combination of these minute charged particles. The Proton, or positively charged particle, is the exact opposite of an Electron which consists of an equal amount of negative electricity. An atom consists of a central positive nucleus around which circulates a number of electrons in various orbits. An atom of hydrogen consists of a nucleus of one proton with one electron circulating around it. In all other atoms the nucleus consists of protons and electrons with the number of protons predominating, making the nucleus positive. Sufficient negative electrons to neutralise exactly the net positive charge of the nucleus rotate in various orbits around the nucleus to make the atom as a whole neutral, that is, neither positive or negative. Three types of atom are shown in Fig. 1.

/Fig. 1.

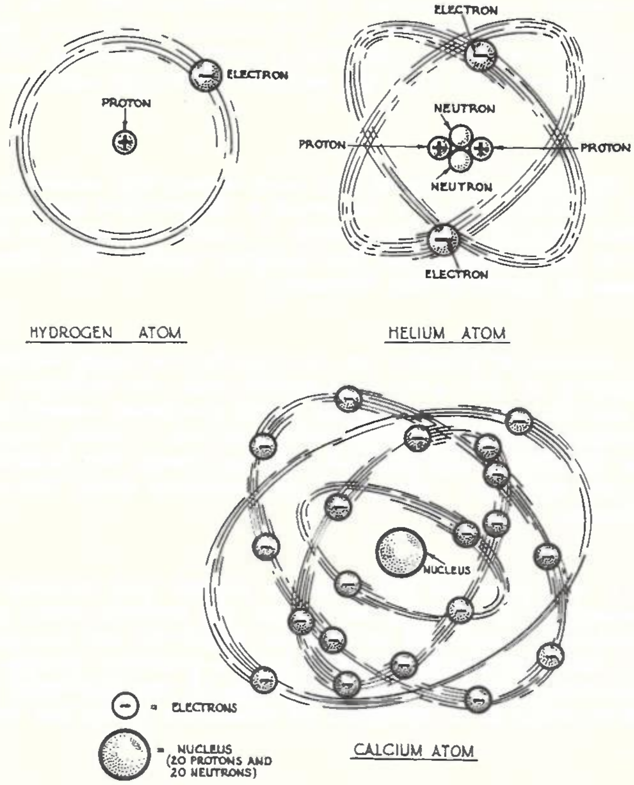


FIG. 1. ELECTRONS - BASIC UNITS OF ENERGY.

2.3 Thermionic Emission. In some kinds of atom some electrons describe much larger orbits than in other electrons - so much larger that such electrons are held to the nucleus by only a small attractive force. It is possible for these electrons to break free from one atom and attach themselves to another which is deficient of electrons at the time. These electrons are called "free" electrons, and atoms containing many free electrons are good conductors. Thus, the free electrons in a conductor are not quiescent but are continually in motion, just as molecules of gases are continually moving according to the Kinetic theory. Although the individual electrons are continually in motion, the average movement in any definite direction in the conductor is zero over a long time, for example, 1 second, which is long compared with, say, the time of 1 cycle at 10 kc/s. On an electromotive force being applied across the conductor the motion takes up a definite direction, the direction of movement being opposite to the conventional direction of current flow.

At normal temperatures there is an effect at the surface of the conductor which is analagous to surface tension in a liquid, and this effect prevents free electrons from escaping beyond the surface of the conductor. As the temperature is raised, however, a proportion of the electrons breaks through the surface of the conductor, just as an increase in the temperature of a liquid will cause evaporation. The breaking through of the electrons appears to be due to an increase in the Kinetic energy of the electrons with an increase in temperature, just as the Kinetic energy of molecules of a liquid increases with an increase in temperature.

/When

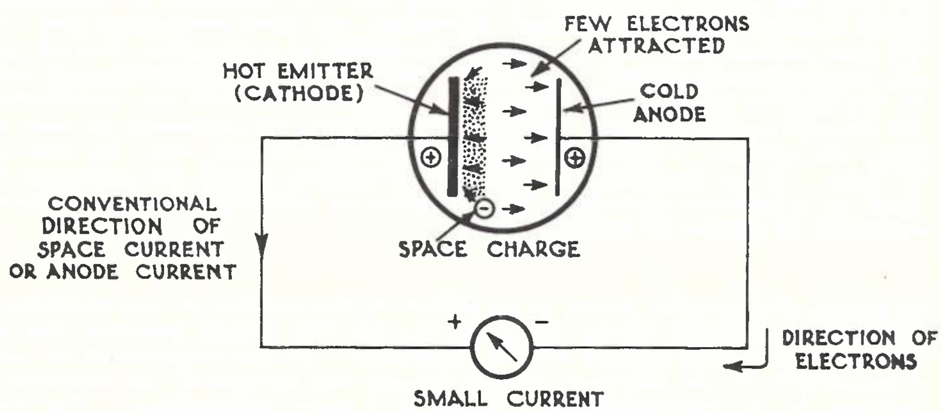
When the electrons escape from the hot conductor, this conductor will have lost a negative charge proportional to the number of electrons emitted. The conductor, therefore, will be positively charged with respect to the emitted electrons, and the decelerating force of the attraction between the positively charged conductor and the negatively charged electrons will gradually stop the emitted electrons and draw them back to the conductor again.

In the first valve developed, a filament of some conducting material capable of emitting a large number of electrons at a reasonable temperature was heated by passing a steady direct current through it. The number of electrons emitted was found to depend on the temperature of the filament and the material used. The electrode from which electrons are emitted is known as the "Cathode."

2.4 Space Charge. The cathode of a valve can be regarded as being surrounded by a cloud of electrons. The density of the cloud is greatest near the cathode, decreasing as the distance from the cathode increases because those electrons reaching some distance from the cathode repel other electrons subsequently emitted. This, together with the attractive force of the positively charged cathode, keeps the majority of the emitted electrons close to the cathode. The cloud of electrons about the cathode of a valve is called the Space Charge.

2.5 Space Current. If a cold piece of any metal is placed just outside the region of the space charge and connected by a conductor to the hot cathode then, since the cathode has been left with a positive charge due to loss of electrons and the piece of cold metal is connected to it, the cold metal will also have a positive charge. The cold electrode being positive, therefore, will attract to it some of the electrons (negative) from the space charge and there will be a movement of electrons along the conductor back to the hot cathode again.

The electrons of the space charge will either travel back to the cathode direct or via the cold electrode. The direction of travel is divided, depending on the distance the electrons have to travel through space and whether they are on the outside or inside of the space charge. Naturally, the further the cold electrode, or anode as this electrode is called, is placed into the space charge, the more electrons it will collect. Fig. 2 shows the electron movements which take place.



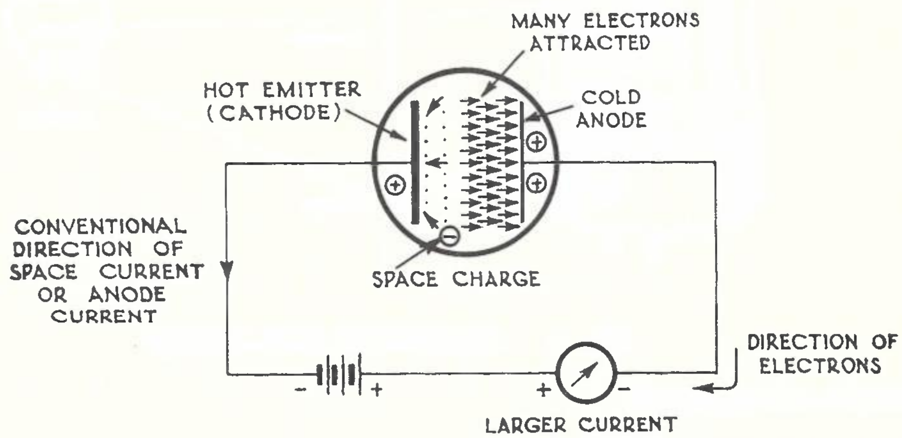
ELECTRON MOVEMENTS IN A VACUUM. NO ANODE POTENTIAL APPLIED.

FIG. 2.

The electrons passing back via the anode and its circuit constitute a small electric current called the space current or anode current.

It is important to note that the direction of the electron stream differs from the conventional direction of the flow of current, that is, from positive to negative in a conductor. When the terms positive and negative were first applied, means were not available for ascertaining in which direction current actually did flow. The terms, originally applied to the copper and zinc plates of a primary cell, were only arbitrary and could just as easily have been reversed originally.

2.6 Positive Anode. If a battery is connected between the anode and cathode with its positive pole to the anode to make it more positive, then the attraction of the anode becomes much greater. The current will now be increased due to more electrons being attracted by the higher positive potential. This is shown in Fig. 3.



ELECTRON MOVEMENTS IN A VACUUM WITH ANODE POTENTIAL APPLIED.

FIG. 3.

2.7 Emission Control. Electrons cannot only be drawn off the cathode and controlled by a suitable electric field but are capable of being moved at enormously high speeds and also of changing their speed and direction practically instantly. These are the properties which make the phenomenon of electron emission so valuable in practice.

/3.

3. TWO-ELECTRODE (OR DIODE) VALVE.

3.1 The type of valve shown in Fig. 3, and containing only two electrodes, is termed a "diode." The electrode which emits the electrons is called the "cathode," while that which attracts them by a positive potential is, as mentioned, called the "anode."

3.2 These electrodes are enclosed in a glass envelope which has been exhausted of all traces of air and exhibits the highest vacuum which it is practicable to obtain. The absence of air or any other gas allows the electrons to travel unimpeded from cathode to anode - collision with gas molecules would cause the electrons to lose much of their energy in the collision. Fig. 4 is an "exploded" diode from which the relations between the electrodes should be clear.

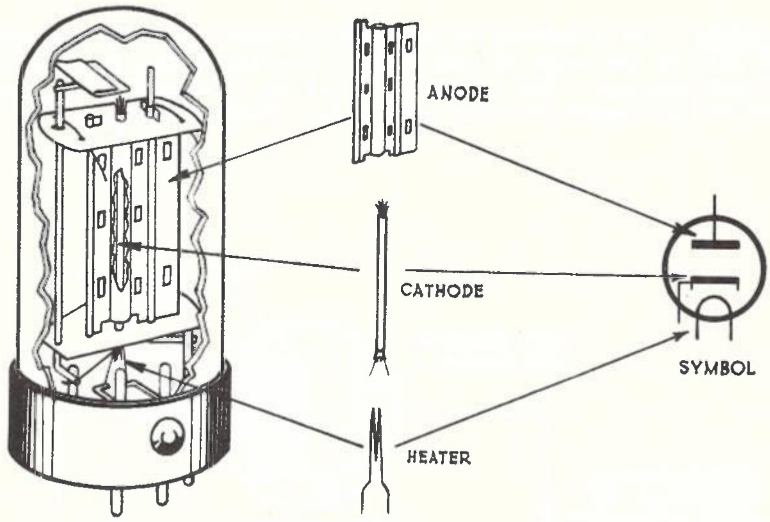


FIG. 4. TWO-ELECTRODE VALVE OR DIODE.

4. CONSTRUCTION OF THE CATHODE.

4.1 Two types of cathode are in use -

(i) Directly heated cathodes, and
(ii) Indirectly heated cathodes.

4.2 Both types are heated electrically, either by forming the cathode into a filament that is raised to the necessary temperature by the passage through it of a suitable current or by using a cylindrical cathode that is heated either by conduction or radiation from an internal heater consisting of an incandescent filament. Filament cathodes may be of the tungsten, thoriated tungsten, or oxide-coated type, while heater type cathodes always employ an oxide-coated emitter because of the impossibility of obtaining by indirect heating the high temperatures required by other emitters.

4.3 Directly heated tungsten filaments are made from the pure metal, and they must operate at very high temperatures for efficient electron emission, whilst thoriated tungsten filaments are made from tungsten impregnated with thorium. In valves with this latter construction, electron emission occurs at much lower temperatures than with pure tungsten filaments, and such valves are more economical, therefore, so far as filament power is concerned.

/In

In both directly and indirectly heated cathodes, oxides are usually applied as a coating on the surface of a suitable metal, such as a nickel or platinum alloy. This coating requires an even lower temperature for efficient electron emission, and is extremely economical as regards filament power. However, each of the above-mentioned cathode materials has special advantages which determine the choice for a particular application.

4.4 The indirectly heated cathode has many advantages. One advantage is that all points are at the same potential whereas a voltage drop takes place over directly heated or filament type cathodes. Another advantage is that no special circuit arrangements are necessary when alternating current or unsmoothed direct current is used to heat an indirectly heated cathode.

5. EFFECTS OF CATHODE TEMPERATURE AND ANODE VOLTAGE ON ANODE CURRENT.

5.1 In the diode only two means are available for varying the anode current, these being the variations in cathode temperature and in anode voltage. These two factors also cause anode current variations in the multi-electrode valves to be dealt with later, but in those valves other means are also available.

5.2 Electrons escaping from the cathode will be drawn to the anode by the attractive force of the positive potential on that electrode, and a continuous flow of electrons from cathode to anode will result. The velocity and number of electrons which cross from cathode to anode are determined by the cathode temperature and the potential of the anode with respect to the cathode. A milliammeter connected in the anode circuit, as shown in Fig. 5, will indicate the actual value of the anode current.

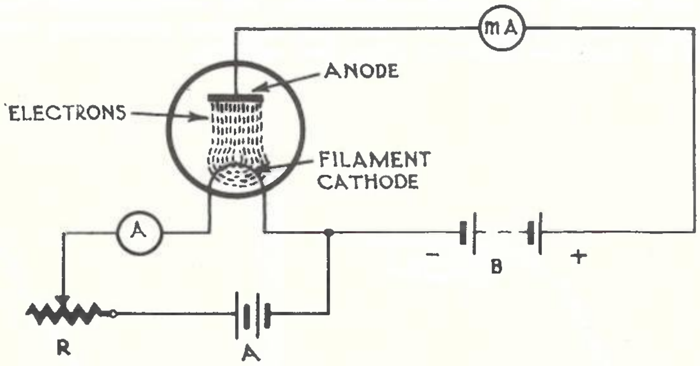
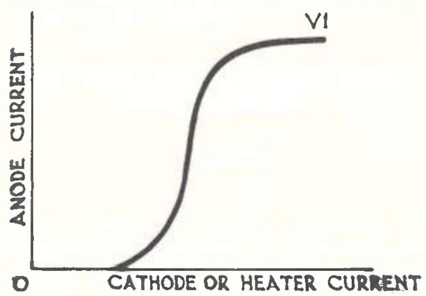


FIG. 5. CIRCUIT FOR SHOWING FACTORS AFFECTING ANODE CURRENT.

5.3 Cathode Temperature Effect. If the rheostat R in series with the cathode of Fig. 5 (or in series with the heater for an indirectly heated cathode) is adjusted so as to decrease the cathode or heater current, thereby lowering the temperature of the cathode, then the anode current as read on the milliammeter will decrease. Eventually, with the cathode at room temperature no anode current is indicated on the milliammeter, so that the space or anode current is zero. On the other hand, if the rheostat is adjusted to increase the cathode or heater current the temperature will also increase, thereby increasing the space or anode current. The general relationship between anode current and cathode or heater current for a representative valve is shown in Fig. 6.



CATHODE OR HEATER CURRENT VERSUS ANODE CURRENT.

FIG. 6.

There is a limit, however, to the increase in space current that can be obtained by increasing

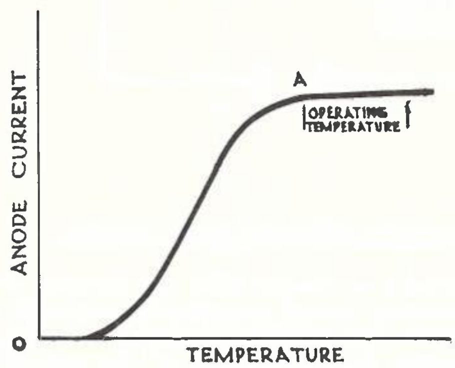
/the

the cathode temperature. This is due to the fact that electrons repel each other because they are all negatively charged and free electrons in the space surrounding the cathode tend to keep new electrons from leaving the cathode. In other words, the electrons themselves, when emitted, tend to counteract further emission of other electrons or to exert a repelling force on electrons within the cathode. This is called the "Space Charge" effect. When the cathode reaches a certain temperature there will be so many electrons in the surrounding space that their repelling effect prevents any further increase in the number of electrons leaving the cathode.

The space current then becomes constant regardless of further increase in temperature, as shown in Figs. 6 and 7.

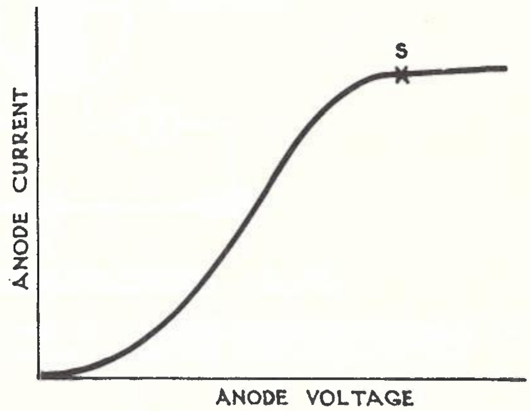
When conditions are as shown by point A in Fig. 7, the valve is said to have reached the temperature saturation point for the particular positive potential at which the anode is maintained. In practice, the operating temperature for the cathode is fixed at a value slightly higher than point A, so that a slight change in the filament battery voltage or heating current will not appreciably affect the performance of the valve.

5.4 Anode Voltage. Having considered the effects of changing the temperature of the cathode, next consider the effects of a change in anode voltage with the cathode at constant temperature. Fig. 8 shows that, as the anode potential is increased, thereby increasing its ability to attract electrons, the anode current increases to a more or less constant value - a state in which the electron attracting ability of the anode exceeds that of electron emission from the cathode.



CATHODE TEMPERATURE VERSUS ANODE CURRENT CHARACTERISTICS.

FIG. 7.



ANODE VOLTAGE VERSUS ANODE CURRENT CHARACTERISTIC.

FIG. 8.

The voltage at which the anode current becomes constant, point S in Fig. 8, is called voltage saturation point for the particular temperature at which the cathode is operated. In practice, the valve is generally operated well below the voltage saturation point.

If the anode potential is made negative with respect to the cathode, the space current will be reduced to zero. Anode current will flow, therefore, if the anode is positive

/with

with respect to the cathode and will flow only in the conventional direction through the valve from anode to cathode. This unidirectional conductivity makes the diode valve particularly adaptable as a rectification device for alternating currents.

5.5 Rectification. If an alternating e.m.f. is substituted for the anode battery and a load resistance connected in series, as shown in Fig. 9, anode current of varying amplitude will flow during each positive half-cycle of anode voltage (that is, with respect to the cathode) but will cease to flow during each negative half-cycle of anode voltage. In other words, there will be a continuous series of pulses of current flowing in the same direction as graphically shown. Consequently, the anode current flowing through the load resistance will be a unidirectional current pulsating at a frequency equal to that of the e.m.f. producing it. The diode valve has practical use, to some extent, in radio receivers, measuring equipment and power equipment because of its rectifying properties.

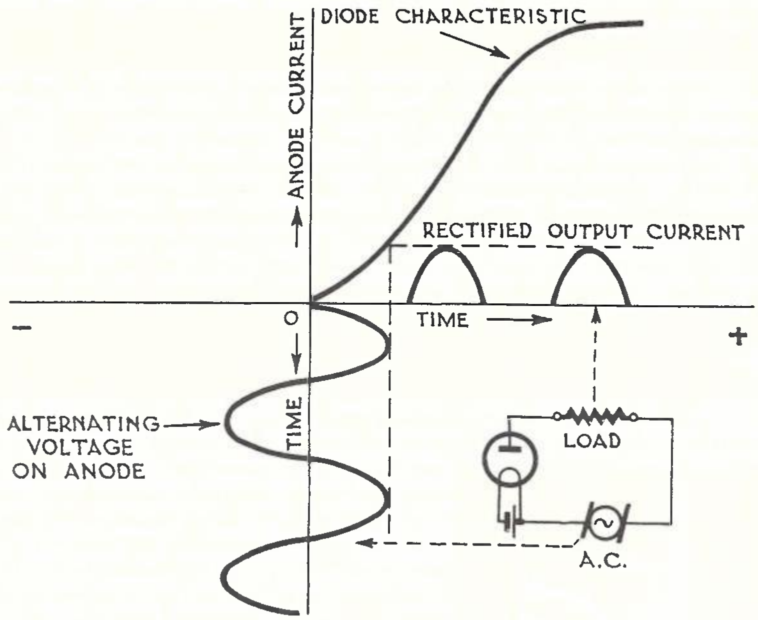


FIG. 9. SIMPLE DIODE RECTIFIER.

6. TRIODE VALVE.

6.1 The applications of a diode are limited, triodes, tetrodes, etc., having a far greater range of usefulness. If a third electrode is placed between the anode and cathode it will offer a means of controlling the flow of electrons between cathode and anode. In order to permit the passage of electrons between the two latter electrodes, the third electrode must be open in structure and, therefore, is of grid form. From its purpose and structure the third electrode is called the Control Grid. Fig. 10 shows the relations between the grid, cathode and anode of a triode valve, and also shows /clearly

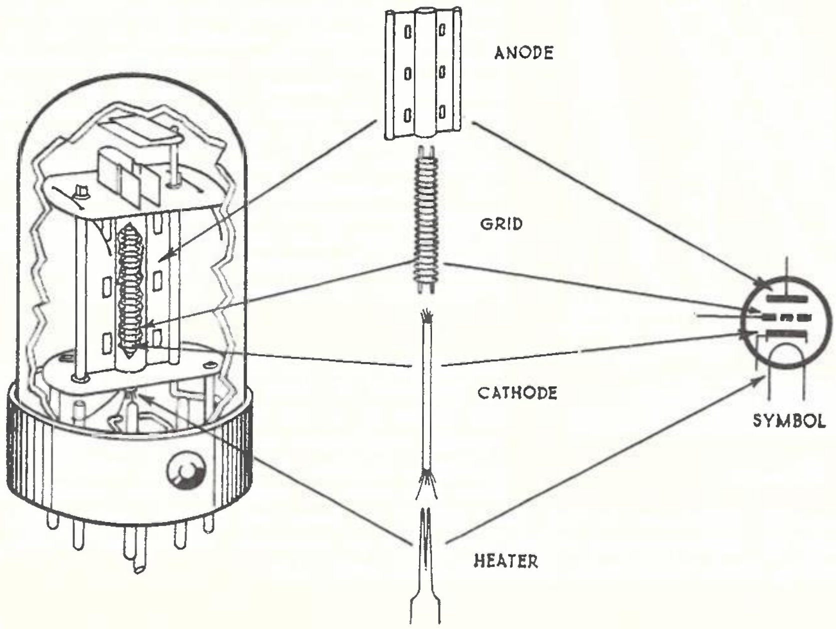


FIG. 10. CONSTRUCTION OF TRIODE VALVE.

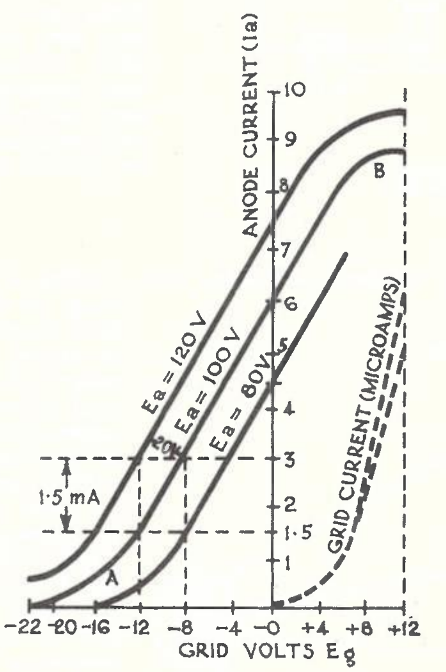
clearly the construction of those electrodes.

6.2 When the potential difference between grid and cathode is zero, the value of the anode current will be fixed by the anode voltage and cathode temperature. If a potential difference is applied between grid and cathode so that the grid is positive with respect to the cathode, then the anode current will increase. This is because the grid attracts electrons in the same manner as does the anode but, due to the open structure of the grid, the majority of these electrons pass through the meshes of the grid and on to the anode.

Increasing this potential difference increases the anode current but, as the grid is nearer the cathode than is the anode, small changes in grid voltage produce the same changes in anode current as do large changes in anode voltage. When the potential difference between grid and cathode makes the grid negative with respect to the cathode, this negative potential will tend to repel electrons emitted by the cathode and so reduce the anode current to a value below that fixed by the anode voltage and cathode temperature alone. If the grid is made sufficiently negative with respect to the cathode no anode current will flow, as the field due to the grid plus the space charge effect will neutralise the field due to the anode in the region close to the cathode.

7. CHARACTERISTIC CURVES.

7.1 Static Characteristic. If a fixed anode potential be applied to a triode valve, the cathode temperature kept constant and the grid voltage varied, a set of anode current readings corresponding to various grid voltages would be obtained. If these readings are plotted against the grid voltage to which they correspond, a curve called the Static or Mutual characteristic curve is obtained. This curve is obtained with no load impedance in the anode circuit of the valve. Fig. 11 shows three such curves, one for an anode potential of 100 volts, another for an anode potential of 120 volts, and another for an anode potential of 80 volts. Considering the curve E_a = 100 volts, the cut-off point occurs at -22 grid volts, that is, no anode current flows at this grid voltage. On reducing the grid volts to -12, the anode current, commencing to flow at about -20 grid volts, rises with increasing steepness. Between -12 and +6 grid volts, the change in anode current is almost directly proportional to the change in grid volts, and between these points the curve straightens. Increasing the grid volts beyond +6 causes the rate of increase of anode current to fall away with increasing rapidity until saturation is reached at +12 grid volts, after which no increase is possible.



CHARACTERISTIC CURVES.
ANODE CURRENT VERSUS GRID VOLTS.

FIG. 11.

Increasing the anode voltage will give curves to the left of that described, as shown by the curve E_a = 120 volts, that is, more anode current will flow for a given grid voltage. Included in Fig. 11 is a grid current curve. When the grid is positive with respect to the cathode it attracts electrons in the same manner as does the anode, and a small current measured in micro-amperes flows from grid to cathode, the value of the grid current increasing with increasing positive grid potential.

/7.2

7.2 Dynamic Characteristics. Static characteristics are obtained with no load impedance in the anode circuit and give no indication, therefore, of the performance of the valve under operating conditions, that is, when a load impedance is connected in the anode circuit. Referring to the circuit in Fig. 12, the anode voltage under load will be given by -

$$E_a = E_B - I_a R_L$$

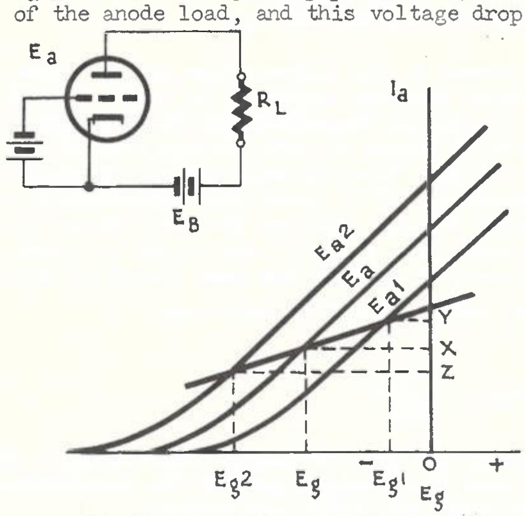
where E_a = anode voltage,

E_B = voltage of anode supply,

I_a = anode current, and

R_L = resistance of load in anode circuit.

$I_a R_L$ is the voltage drop produced by the anode current flowing through the resistance of the anode load, and this voltage drop will vary as I_a varies, thus causing E_a to vary. Suppose that the grid voltage is adjusted to a value E_g. The anode current I_a flowing through the load resistance R_L will cause the anode voltage to have a value E_a. This condition is shown in Fig. 12 on the curve drawn for a constant anode potential E_a, the anode current for E_g grid volts being OX. If the grid voltage is changed to E_g1, that is, the grid is made less negative with respect to the cathode, the anode current will increase and the voltage drop across R_L will correspondingly increase, causing the anode voltage to fall to E_a1. The value of anode current will now be OY on a curve drawn with a constant anode potential E_a1. Similarly, if the grid voltage is changed to E_g2, that is, made more negative, the anode current will decrease, the voltage drop across R_L will decrease and the anode voltage will rise to E_a2. The value of anode current will now be OZ on a curve drawn with a constant anode potential E_a2.



DERIVATION OF DYNAMIC CHARACTERISTIC.

FIG. 12.

It will be seen that the slope of the characteristic curve of a valve with a load connected in the anode circuit is less than when no load impedance is present. Increasing the resistance of the load will give a set of curves similar to those shown in Fig. 13.

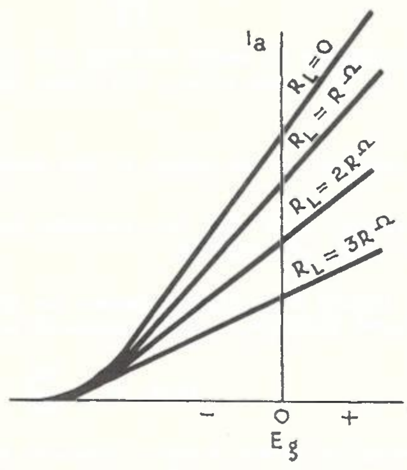


FIG. 13. DYNAMIC CHARACTERISTICS OF TRIODE.

These curves are known as Dynamic characteristics. It is of particular importance to note that as the load resistance increases the straight part of the characteristic curve increases, due to the straightening of the lower bend.

8. VALVE CONSTANTS.

8.1 It is useful to deduce certain information from the mutual characteristics of valves in order to determine the suitability of a valve for a certain purpose, for example, to determine whether a certain valve would be suitable for use as a voltage amplifier or as a power amplifier. This information is gained by expressing the relationship between changes in grid voltage, anode voltage and anode current. These relationships are termed the Valve Constants, of which there are three -

(i) Amplification Factor, designated μ.
(ii) Anode Impedance, designated r_a.
(iii) Mutual Conductance, designated g_m.

8.2 Amplification Factor. As the grid is nearer to the cathode than is the anode, a small variation in grid voltage will cause the same variation in anode current as a large variation in anode voltage. On the Mutual characteristic first studied (Fig. 11) with an anode potential of 120 volts, a change of grid voltage from -8 to -12 causes a decrease in anode current of 1.5 mA. With the grid voltage constant at -8 the anode voltage would have to be reduced to 100 to cause this reduction in anode current. Thus, a 20 volt change in anode voltage produces the same change in anode current as a 4 volt change in grid voltage. The ratio of the change in anode voltage to the change in grid voltage necessary to produce the same change in anode current is termed Amplification Factor, and is $\frac{20}{4} = 5$ in this case.

$$\text{Thus, } \mu = \frac{dE_a}{dE_g}$$

where dE_a and dE_g are the changes in anode and grid voltages, respectively. The amplification factor is constant only over the straight part of the characteristic curve, decreasing steadily as the upper and lower bends are entered.

8.3 Anode Impedance. As most valves work into impedance loads and, further, as they are equivalent in operation to alternating current generators, it is necessary for matching purposes that the impedance between anode and cathode be calculated and not the direct current resistance. As any impedance depends on the rate of change of current and voltage, the impedance between anode and cathode will be given by -

$$\frac{dE_a}{dI_a}$$

where dE_a is a small change in anode voltage and dI_a is the change in anode current which it produces. In the case taken, the anode impedance will be 20 volts/1.5 mA or 13,333-1/3 ohms.

$\frac{dE_a}{dI_a}$ is also the reciprocal of the slope of the anode characteristic shown in Fig. 8.

Thus, r_a will be constant only over the straight part of the curve, increasing as the upper and lower bends are entered.

8.4 Mutual Conductance. Mutual Conductance refers to the change in anode current caused by unit changes in grid voltage, and is measured in mhos.

$$\text{Thus, } g_m = \frac{dI_a}{dE_g}$$

where dI_a is the change in anode current produced by a change in grid volts of dE_g. In the case taken, g_m will be 1.5 mA/4 volts or 0.000375 mhos or 375 micromhos. As with the amplification factor, g_m decreases as the upper and lower bends are entered.

/8.5

8.5 Relationship between Valve Constants. The three valve constants are related to one another in the following manner -

$$\mu = \frac{dE_a}{dE_g} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

$$r_a = \frac{dE_a}{dI_a} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (2)$$

$$g_m = \frac{dI_a}{dE_g} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

Multiplying $\frac{dE_a}{dE_g}$ by $\frac{dI_a}{dI_a}$ or 1 in equation (1) gives -

$$\mu = \frac{dE_a}{dE_g} \times \frac{dI_a}{dI_a}$$

$$= \frac{dE_a}{dI_a} \times \frac{dI_a}{dE_g}$$

$$\text{but } \frac{dE_a}{dI_a} = r_a$$

$$\text{and } \frac{dI_a}{dE_g} = g_m$$

$$\therefore \mu = r_a\, g_m \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4)$$

$$r_a = \frac{\mu}{g_m} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (5)$$

$$\text{and } g_m = \frac{\mu}{r_a} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (6)$$

8.6 Significance of the Valve Constants. The significance of the Amplification Factor and Anode Impedance may be more readily appreciated if an actual case involving alternating voltage is worked out. This is done in Fig. 14, using the same valve as in Fig. 11.

Fig. 14a shows the valve with steady voltages of -8 volts and 100 volts applied across grid and cathode and anode and cathode, respectively. An alternating signal voltage of 4 volts is superimposed on the steady voltage across the grid and cathode giving rise to an alternating component of anode current of 1.5 mA. Referring to Fig. 10, the anode voltage varying between 80 and 120 also causes the anode current to vary between 1.5 and 4.5 mA. This means that when an alternating voltage of 4 volts is applied across grid and cathode it must produce an alternating voltage of 20 volts in the anode circuit in order to account for the 1.5 mA alternating component of anode current. Thus, the signal voltage is amplified by a factor $\frac{20}{4} = 5$ by the valve. This factor, 5, is the amplification factor, μ, derived previously.

An alternating voltage of 20 volts applied across a circuit and resulting in an alternating current of 1.5 mA means that the circuit has an impedance of $\frac{20\text{V}}{1.5\text{ mA}} = 13{,}333\text{-}1/3$ ohms.

/The

The only impedance in the anode circuits of Figs. 14a and 14b is that of the valve, so that 13,333-1/3 ohms is the impedance offered by the anode to cathode circuit inside a valve and is the anode impedance, r_a, derived previously. A valve, therefore, may be regarded as an alternating current generator developing a voltage of μE_s and having an internal impedance of r_a, E_s being the alternating signal voltage applied to the grid as shown in Fig. 14c.

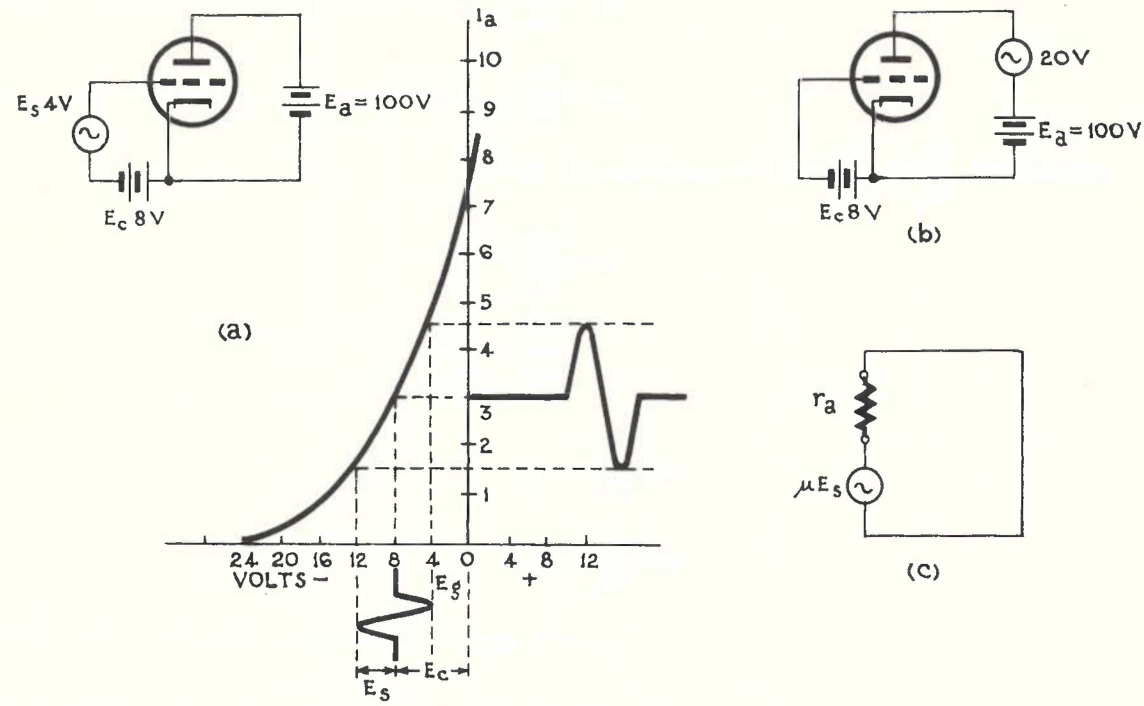


FIG. 14. ACTION OF VALVE WITH ALTERNATING VOLTAGE ON THE GRID.

9. SCREEN-GRID (OR TETRODE) VALVES.

9.1 Whilst the electronic action of a triode is unidirectional, capacitances are present which provide a path for energy to be fed back from the output circuit to the input. Referring to Fig. 15, it will be seen that the grid-anode capacity C_{ga} couples the input and output circuits together. At audio frequencies the coupling is not noticeable as the reactance of C_{ga} is high, but at radio and carrier frequencies the reactance becomes low enough for a transfer of energy between output and input circuits via C_{ga}, which produces undesirable effects.

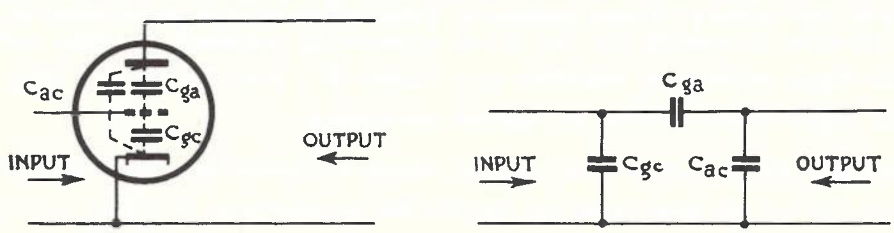
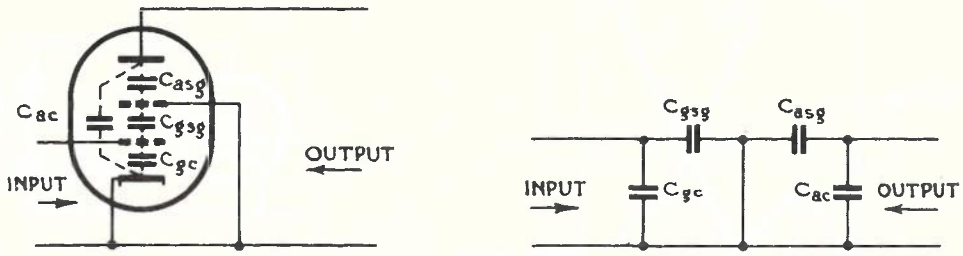


FIG. 15. TRIODE CAPACITIES AND EQUIVALENT NETWORK.

9.2 To eliminate this coupling, an electrostatic screen in the form of a grid is introduced between the control grid and anode. This extra electrode, termed the screen grid, is connected to the cathode, the effect being to split the grid-anode capacity into two series capacities with the input short-circuited from the output by the connection between screen grid and cathode. This is shown in Fig. 16, which also contains an exploded view of a tetrode.



Tetrode Capacities and Equivalent Network.

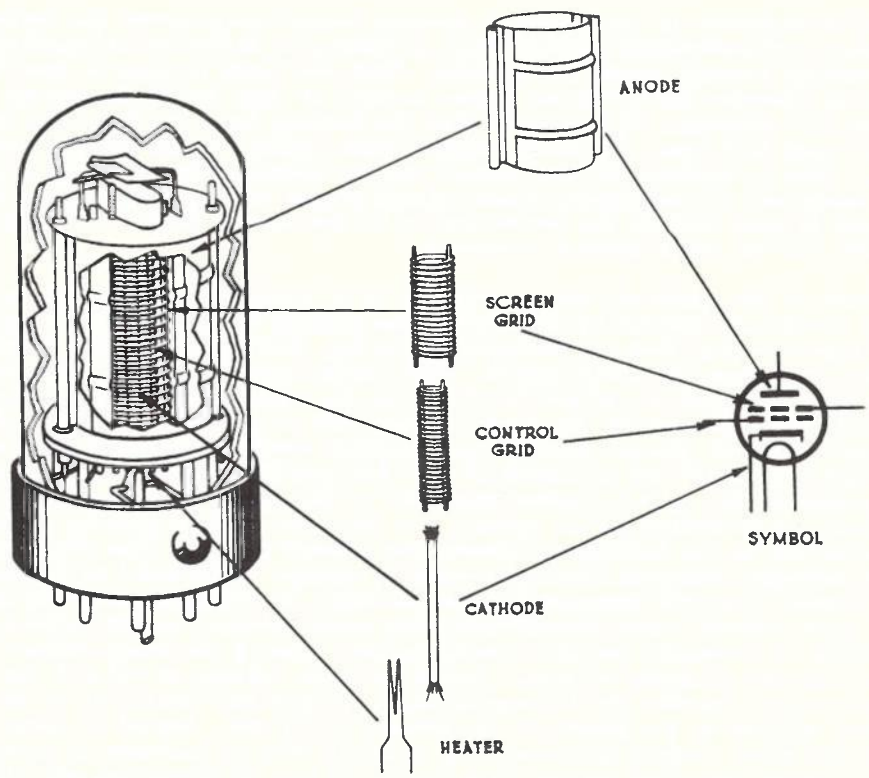
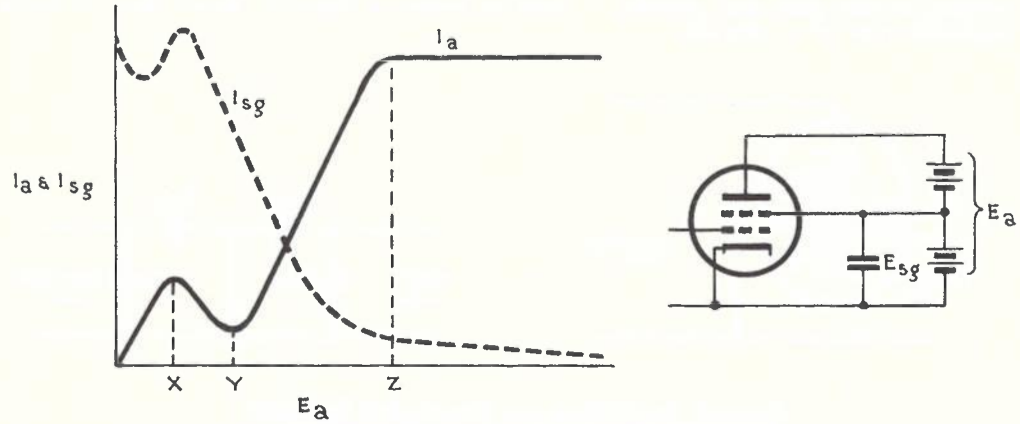


FIG. 16. SCREEN-GRID VALVE.

9.3 By operating the screen grid at a positive potential a little lower than that of the anode, the amplification factor can be made much higher than for an equivalent triode. The curves of Fig. 17 show the anode and screen-grid characteristic of a screen-grid valve with a fixed screen-grid voltage, together with a typical circuit.



ANODE AND SCREEN CHARACTERISTICS OF TETRODE.

FIG. 17.

A suitable condenser in the screen to cathode connection prevents the Screen Supply from being short-circuited, whilst still providing an effective alternating current short-circuit between screen and cathode. When the anode voltage is zero the anode current is likewise zero, but the screen current is high due to the screen being operated at a voltage a little lower than the working anode voltage. As the anode voltage rises from zero, the anode current rises rapidly - more rapidly than for a triode. This is because the high screen voltage attracts the electrons from the space charge and the increasing anode voltage causes an increasing number of the electrons to pass through the screen and on to the anode. Thus, the screen current decreases and the anode current increases.

The effect of the anode and screen voltages is to accelerate the electrons (called primary electrons) emitted by the cathode and passing to the anode, and to give them such a high velocity that, on striking the anode, they dislodge other electrons (called secondary electrons) by their impact. This latter effect is known as Secondary Emission. At low anode voltages the effect is slight as the velocity of the electrons is comparatively low. Increasing the anode voltage will increase the velocity of the primary electrons, and each primary electron may dislodge several secondary electrons. Some of the latter come under the influence of the high positive potential on the screen grid whose potential is, as yet, higher than that of the anode. Some secondary electrons, therefore, are attracted to the screen, and the total current passing between anode and cathode is due to the difference between the primary electron flow from cathode to anode and the secondary electron flow from anode to screen. The result is to decrease the anode current flow with increasing anode voltage. On the anode characteristics of Fig. 17 the anode current rises rapidly until the anode voltage reaches value X, beyond which the effects of secondary emission become marked, causing the anode current to decrease with an increase in anode voltage. Beyond an anode voltage of value Y the anode current increases again, because the increasing anode voltage can retain all of the secondary electrons emitted by the impact of the primary electrons. Beyond an anode voltage of value Z the anode current increases only slightly with an increase in anode voltage. This is because the anode voltage plus screen voltage is attracting all electrons from the space charge at anode voltage Z, and an increase of anode voltage beyond value Z merely draws a few electrons away from the screen. The screen current is the inverse of the anode current, that is, as the anode current rises the screen current falls, and as the anode current falls the screen current rises.

/10.

10. CONSTANTS OF SCREEN-GRID VALVES.

10.1 The three constants of screen-grid valves differ from those of an equivalent triode as follows -

Anode Impedance. Above anode voltage Z the anode current is practically independent of anode voltage, that is, large changes in anode voltage produce only small changes in anode current. Thus, $\frac{dE_a}{dI_a}$ is large, much larger than for an equivalent triode, that is, one with the same spacings between control grid and cathode and control grid and anode. This only applies when the anode voltage does not fall below value Z.

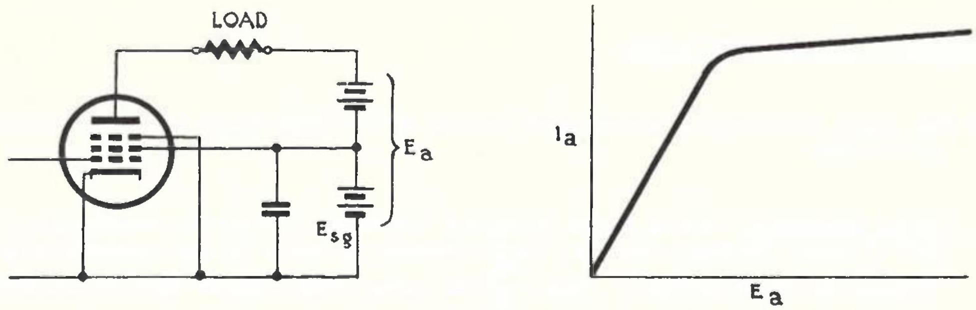
Mutual Conductance. Unit changes in grid voltage produce the same changes in anode current for screen-grid valves and their equivalent triode. Thus, the mutual conductance, g_m, for the two valves is the same.

Amplification Factor. As r_a is higher for a screen-grid valve than for its equivalent triode, and as g_m for the two valves is the same, then, as $\mu = r_a g_m$, the amplification factor for a screen-grid valve is much higher than for its equivalent triode. As this only applies provided the anode voltage does not fall below value Z, this increase in amplification factor is secured by having to apply a higher working anode voltage to a screen-grid valve than to its equivalent triode.

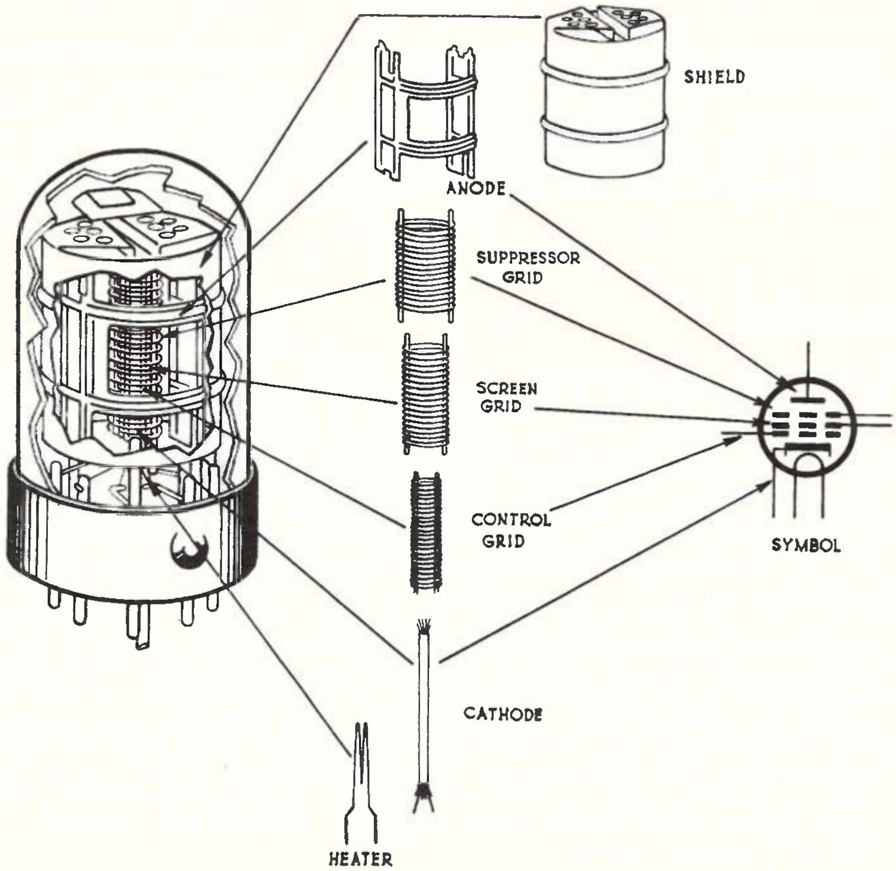
11. PENTODE VALVES.

11.1 If the anode current of a screen-grid valve could be made to continue to rise rapidly with increasing anode voltage instead of developing the "kink" due to secondary emission, the value of anode voltage beyond which the anode current becomes independent of anode voltage would be considerably lowered, so allowing a lower working anode voltage to be used. The pentode valve achieves this by suppressing the effects of secondary emission. Another grid is inserted between the anode and screen grid and is directly connected to the cathode, either internally or externally. As the cathode is negative with respect to the anode, the secondary electrons, on leaving the anode, encounter the negative field of this suppressor grid, as it is termed, and this field repels them back to the anode. Another way of looking at the function of the suppressor grid is to think of its effect on the velocity of the primary electrons. As primary electrons pass through the screen grid they will be decelerated by the negative voltage on the suppressor grid, so that they will not strike the anode at such a high velocity. Very few, if any, secondary electrons will be dislodged from the anode, those dislodged being repelled back to the anode as described. Fig. 18 shows how the elimination of the effects of secondary emission result in the continuous rapid rise of the anode current, resulting in a lower working anode voltage being required. Fig. 18 also contains an exploded view of a pentode valve.

/Fig. 18.



Anode Characteristic of Pentode.

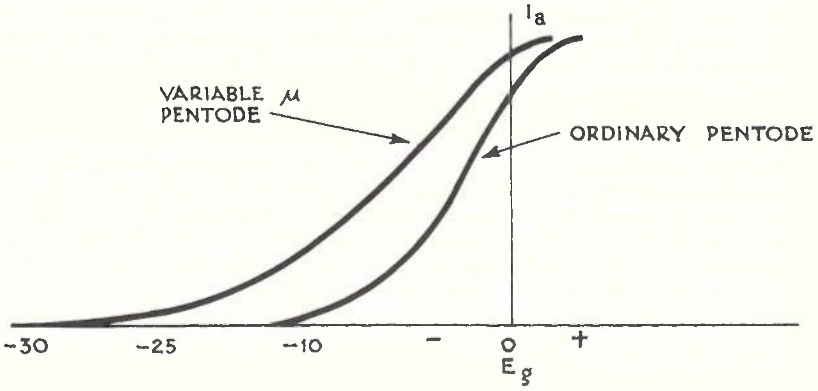


PENTODE VALVE.

FIG. 18.

11.2 Variable μ Pentodes. Variable μ valves (also termed remote cut-off valves and super control valves) are usually pentodes designed so as to cause the anode current to curve away very slowly to a not very well defined cut-off.

Fig. 19 compares the characteristic of an ordinary pentode with that of a variable μ pentode. The variable feature is obtained by using a control grid with a non-uniform grid mesh, so that the amplification factor is different for various operating points on the characteristic curve.



COMPARISON OF VARIABLE μ AND ORDINARY PENTODE CHARACTERISTIC.

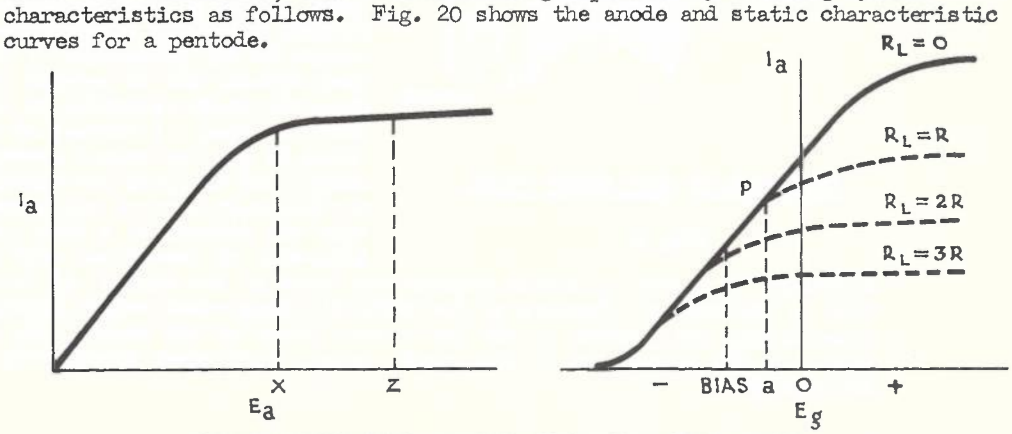
FIG. 19.

12. CONSTANTS OF PENTODE VALVES.

12.1 As in the screen-grid valve, g_m for a pentode valve is the same as for an equivalent triode, and r_a is higher over that portion of the anode current curve where the anode current is practically independent of anode voltage. The amplification factor, therefore, must be high as in the screen-grid valve ($\mu = r_a g_m$), but the higher μ is secured with a much lower working anode voltage than is the case with the screen-grid valve.

13. DYNAMIC CHARACTERISTICS OF SCREEN-GRID AND PENTODE VALVES.

13.1 When the dynamic characteristics of a triode valve were developed, it was pointed out that the extent of the lower bend curvature decreases as the impedance of the anode load increases. This applies to triodes alone and not to screen-grid and pentode valves. The anode current versus grid voltage characteristics of the latter valves develop an inflection point and become more curved when operating under load conditions, this behaviour being explained by deriving dynamic characteristics as follows. Fig. 20 shows the anode and static characteristic curves for a pentode.



ILLUSTRATING DYNAMIC CHARACTERISTICS OF PENTODES.
FIG. 20.

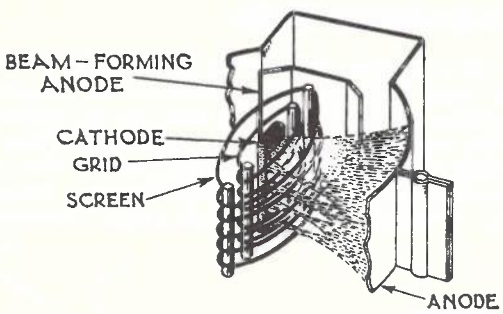
Assume, now, that a load impedance is connected in the anode circuit, the grid is biassed into the centre of the negative straight portion of the static characteristic and that the anode voltage is adjusted to value Z from which value the static characteristic was obtained. As the grid voltage is reduced towards zero the anode current increases, producing an increased voltage drop across the anode load and so decreasing the anode voltage. Until the anode voltage falls to value X, however, the anode current is practically independent of anode voltage and is dependent on grid voltage alone, so that the static and dynamic characteristics coincide. Assume, now, that when the grid voltage has fallen to value a, the current is such that the voltage drop it produces across the anode load reduces the anode voltage to value X.

On the grid voltage being reduced further towards zero, the anode current will tend to increase under the influence of the grid voltage change but will tend to decrease under the influence of the anode voltage change. Thus, the rise in anode current is now the difference between the rise due to the grid voltage change and the fall due to the anode voltage change. As the rise is steady and the fall increases with decreasing anode voltage, the dynamic characteristic curves away from the static characteristic at point P. Increasing the load impedance causes the inflection point to occur at lower grid voltages, as shown by the curves for $R_L = R, 2R$, etc.

13.2 As will be seen later, this effect of a load impedance on the characteristic curve of a valve is important because it determines to a great extent the magnitude of the load impedance which must be used to obtain the best operating conditions.

14. BEAM POWER VALVES.

14.1 A beam power valve is a tetrode in which use is made of directed electron beams to contribute substantially to its power handling capacity. This type of valve contains a cathode, a control grid, a screen and an anode. The electrodes are so spaced that secondary emission from the anode is suppressed by space charge effects between screen and anode. The space charge is produced by slowing up electrons travelling from a high potential screen to a lower potential anode. In this low velocity region the space charge produced is sufficient to repel secondary electrons emitted from the anode. Beam power valves of this design employ beam forming anodes at cathode potential to assist in the production of the desired beam effects and to prevent stray electrons from the anode returning to the screen outside the beam. A feature of a beam power valve is its low screen current. The screen and the grid are spiral wires wound so that each turn of the screen is shaded from the cathode by a grid turn. This alignment of screen and grid causes the electrons to travel in sheets between the turns of the screen, so that few of them flow to the screen. Because of the effective suppressor action provided by the space charge and because of the low current drawn by the screen, the beam power valve has the advantages of high power output, high power sensitivity and high efficiency.



STRUCTURE OF BEAM POWER VALVE.

FIG. 21.

The structure of a beam power valve is shown in Fig. 21.

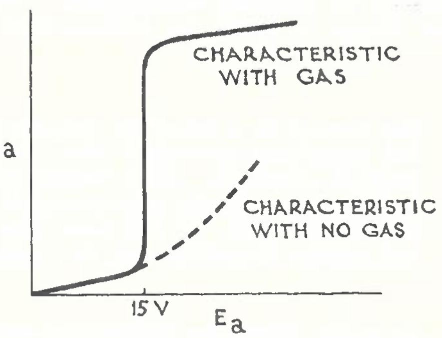
/15.

15. EFFECTS OF GAS.

15.1 As discussed earlier, when the cathode of a valve is heated electrons are emitted which, in the absence of any anode voltage, form a space charge about the cathode. The first electrons emitted can travel relatively large distances from the cathode as there is only the decelerating force of the now positively charged cathode to prevent them doing so. Electrons subsequently emitted are subject to two decelerating forces (that of the positively charged cathode and that of the negatively charged electrons initially emitted) which prevent the electrons from reaching any distance from the cathode. When an anode voltage is applied, electrons which pass to the anode to form the anode current must overcome these decelerating forces in order to pass to that anode, which means that a large part of the anode voltage is used to overcome the space charge effect. This accounts for the large voltage drop between the anode and cathode of high vacuum valves. By introducing gas into a diode at a suitable pressure, saturation current can be achieved with an internal drop in the valve of up to 20 volts.

The reason for this behaviour is that electrons travelling from cathode to anode may, upon striking a gas molecule, knock out one or more of its constituent electrons. This removes one or more negative charges from the gas molecule, leaving it positively charged or ionised. The process is known as ionisation by collision. The velocity needed by an electron to ionise a gas in this manner is usually expressed in terms of the potential difference through which the electron must fall in order to acquire the necessary velocity. This is, of course, the P.D. between anode and cathode, as the higher this P.D. the greater is the velocity of the electrons and, therefore, the greater the impact of their collisions with gas molecules. This P.D. is called the ionising potential of the gas, and those of most gases range from 10 to 25 volts. The positive ions, which result from the gas molecules losing electrons by collision, are repelled by the positive voltage on the anode and attracted towards the negative space charge. These positive ions, therefore, move towards the cathode where their presence neutralises the space charge effect, so permitting the anode to draw the electrons from the cathode as fast as they are emitted when the anode is only 15 to 20 volts positive with respect to the cathode.

The result is an anode characteristic of the type shown in Fig. 22.



ANODE CHARACTERISTIC OF GAS DIODE.

FIG. 22.

This principle is used in the Tungar rectifier, which uses Argon gas, and the Mercury vapour rectifier. The mercury vapour rectifier contains a small amount of metallic mercury which vaporises when its temperature is raised on the cathode being heated, so forming mercury vapour. The ionising potential of mercury vapour is about 10.5 volts, so that, when the P.D. between anode and cathode reaches this figure, the mercury vapour ionises. The positive ions then move towards the cathode, where they neutralise the space charge effect and cause the electrons in the space charge to be drawn to the anode as fast as they are emitted. Thus, saturation current is reached at the ionising potential of the mercury vapour, that is, about 10.5 volts.

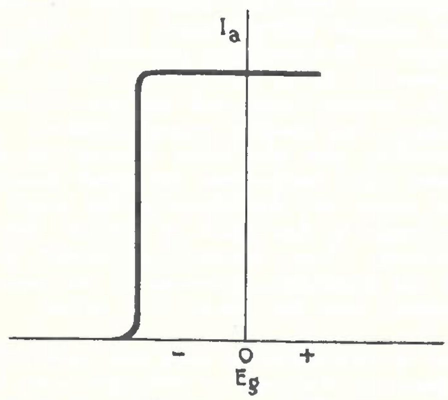
The mass of these positive ions produced by collision is very much greater than that of an electron - about 1,840 times as great in the case of hydrogen - so that, under normal conditions, ions will move very slowly. If they are produced in sufficient numbers and/or are subjected to a high P.D. between anode and cathode, ions will bombard the cathode in sufficient numbers and/or at a high enough velocity to

/eventually

eventually disintegrate the cathode. The amount of ionisation produced is controlled by the pressure of the mercury vapour which, in turn, is controlled by the temperature together with the amount of metallic mercury available for vaporisation. Whilst the P.D. between anode and cathode controls the amount of ionisation produced, perhaps its most important effect is on the velocity which these positive ions acquire. Too high a voltage between anode and cathode will give these ions a velocity high enough for even only a few of them to disintegrate the cathode by bombardment. This necessitates the strict precaution of always applying the cathode heating power at least 30 seconds before the anode supply voltage. If the anode voltage is applied before, or simultaneously with, the cathode heating power, then little mercury vapour will be available for ionisation whilst the valve is "warming up." This means that the valve will function largely as a high vacuum valve during this period, with a consequent high voltage drop across anode and cathode. This high voltage will give the few ions produced from the little mercury vapour available during the "warming up" period a velocity high enough to disintegrate the cathode by their bombardment.

Whilst these valves may be used in circuits rectifying thousands of volts, it should be remembered that nearly all of this voltage is dropped across the load, the arrangements being such that only the 10 to 15 volts required for ionising the gas are dropped across the valve. This makes these valves very efficient rectifiers.

15.2 Gas Triodes (Thyratrons). When gas having a suitable pressure is introduced into a triode (or any valve having a control grid) the control action exerted by the grid is changed in a very remarkable way. If the operation of the valve starts with a negative grid voltage considerably greater than cut-off and then gradually reduces this, it is found that at the point where the anode current just starts to flow if the valve contains no gas, the anode current suddenly rises from zero to a very high value and readily reaches the full emission of the cathode with anode voltages as low as 15 to 20 volts. After the flow of anode current has once been started, the control grid has no further effect and can be made much more negative than cut-off without altering the anode current appreciably. In order to stop the anode current, the anode voltage must be reduced to a very low value. This behaviour is caused by the electrons moving from the cathode to the anode, colliding with molecules of gas and producing positive ions by dislodging electrons from the gas molecules. These positive ions are attracted towards the negatively charged grid and towards the cathode, which is surrounded by the negative space charge. Thus, the normal control action of the negative voltage on the grid is neutralised and the space charge is likewise neutralised. Hence, once ionisation has started there is no space charge to limit the current flow and the control action of the grid has been lost.



CHARACTERISTIC OF THYRATRON.

FIG. 23.

The result is a relay or trigger device which has important practical uses, particularly in control circuits. This device takes practically no energy at the negative grid to initiate the action, and, at the same time, the resultant energy turned on can be large.

In cases where the ionisation and deionisation must be rapid, helium, argon or neon is employed because the positive ions of these gases are more mobile than those of mercury, which is used where the rapidity of ionisation and deionisation is not so important.

Fig. 23 shows a typical characteristic for a gas triode, the behaviour of which, under the control of grid voltage, is similar to the characteristic of the gas diode shown in Fig. 22.

/16.

16. VALVE FAULTS AND TESTING.

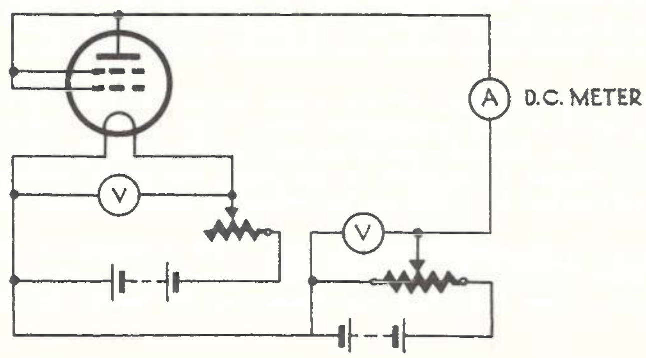
16.1 Typical valve faults are detailed hereunder -

(i) Low Cathode Activity, due to aging and deterioration of the cathode.
(ii) Low Insulation Resistance - a deterioration in the normal I.R. between electrodes with consequent change in performance characteristics.
(iii) Low Gain - a decrease in the original amplification factor.
(iv) Bright Spots. Some cathodes, due to manufacturing irregularities, exhibit excessive brightness over short portions of their length, which is indicative of excessive local heating. Failure of the cathode usually occurs at a bright spot.
(v) Poor Vacuum - sometimes caused by rough handling but usually due to faulty sealing or temperature stresses in the glass envelope. Indicated by failure of cathode to glow when carrying normal current and excessive heating of glass envelope.
(vi) Displaced Electrodes - Broken Supports. These may be caused by rough handling or temperature stresses and may cause distortion, low gain or noise.

16.2 Testing. The condition of valves governs the performance of the equipment in which they are employed. In order to determine that a valve is performing its functions correctly, certain tests are necessary. As the operating capabilities and design features of a valve are indicated by the valve's electrical characteristics, a valve is tested by measuring its characteristics and comparing them with representative values listed as standard for the type of valve concerned. It is usual to select one characteristic only to serve as an indication of the performance of a valve, and it is essential, therefore, that the characteristic selected be truly representative of the condition of a valve. The tests which fall within this category are -

(i) Emission Test, and
(ii) Mutual Conductance Test.

16.3 Emission Test. An emission test or cathode activity test is possibly the simplest method of indicating the condition of a valve. Since electron emission decreases as the valve wears out, low emission is indicative of the end of the useful life of a valve. An emission test is subject to the limitation that it tests a valve under static conditions and does not take into account the actual operation of the valve. A typical emission or cathode activity test circuit is shown in Fig. 24.



CATHODE ACTIVITY TEST CIRCUIT.

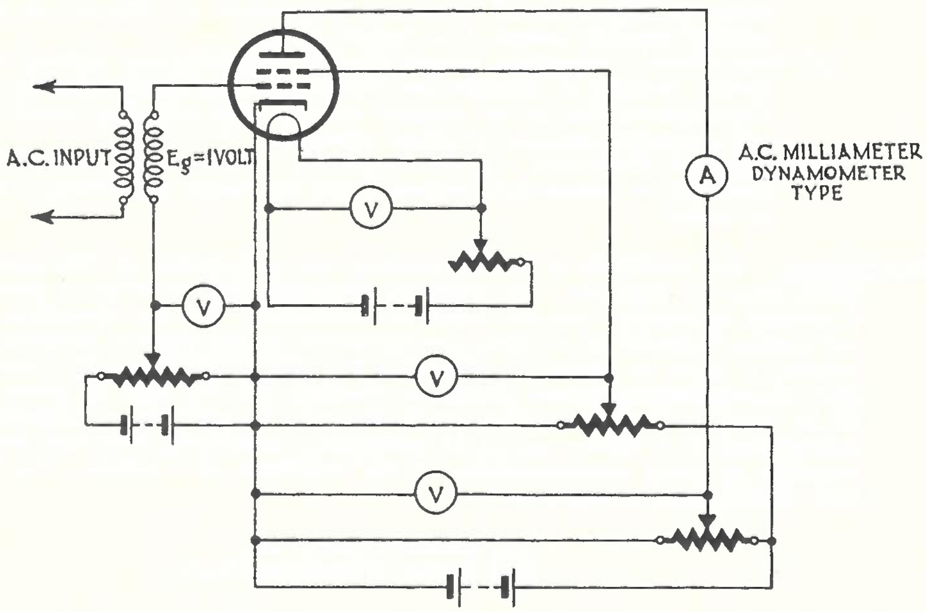
FIG. 24.

All the electrodes of the valve, except the cathode, are connected to the anode. The filament or heater is operated at rated voltage. After the valve has reached constant temperature, a low positive voltage is applied to the anode and the

/electronic

electronic emission is read on the meter. Readings which are well below the average for a particular valve type indicate that the total number of available electrons has been so reduced that the valve is no longer capable of functioning correctly.

16.4 Mutual Conductance Test. A mutual conductance test takes into account a fundamental operating principle of a valve. It follows that a mutual conductance test, when correctly performed, gives a more accurate indication of the performance of a valve than an emission test. A dynamic mutual conductance test circuit is shown in Fig. 25.



MUTUAL CONDUCTANCE TEST CIRCUIT.

FIG. 25.

The dynamic mutual conductance test is superior to a static mutual conductance test in that alternating current voltage is applied to the grid, and the valve, therefore, is tested under conditions which approximate actual operating conditions.

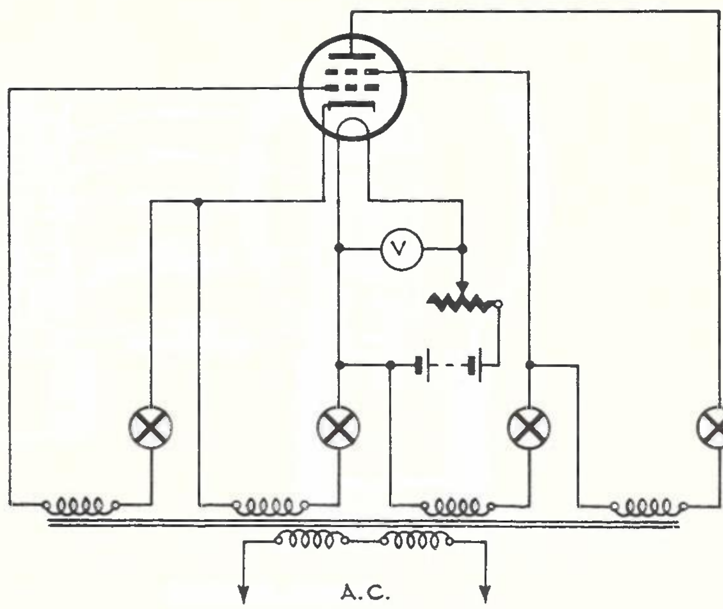
The alternating component of the anode current is read by means of an alternating current ammeter of the dynamometer type. The mutual conductance of the valve under test is equal to the A.C. anode current divided by the input signal voltage. If a one volt R.M.S. signal is applied to the grid, the anode current meter reading in milliamperes multiplied by 1,000 is the value of mutual conductance in micromhos.

16.5 Short-Circuit or Electrode Contact Test. The fundamental circuit of a short-circuit test is shown in Fig. 26. Whilst this circuit is suitable for tetrodes and triodes, valves having more than four electrodes can be tested by adding additional indicator lamps.

Voltages are applied between the various electrodes with lamps in series with the electrode leads. Any two shorted electrodes complete a circuit and light one or more lamps. Since two electrodes may be in high resistance contact, it is necessary that the indicating lamps operate on very low current. It is also necessary to maintain the filament or heater of the valve at correct operating temperature during the short-circuit test because short-circuits or contacts may only occur when the electrodes are heated.

/A

A valve tester cannot be regarded as the final arbiter of valve performance. An actual operating test in the equipment in which the valve is used will possibly give the best indication of the worth of a valve. On the other hand, a valve tester is a valuable guide to the serviceability of a valve and frequently results in the replacement of valves before they reach failure point and cause severe interruptions.



SHORT-CIRCUIT TEST.

FIG. 26.

17. VALVES - TYPES IN USE.

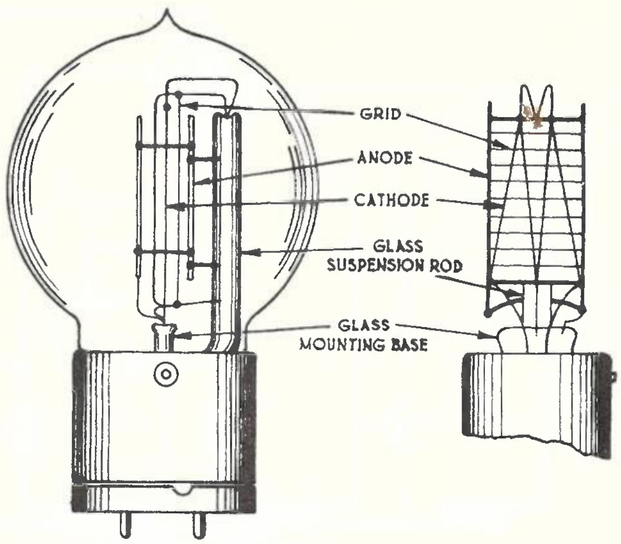
17.1 A wide range of valves is in use in Long Line Equipment. The present day tendency is to restrict the number of types as much as possible although, in many cases, valves having particular characteristics are required for specialised circuit applications. A large percentage of the applications required can be met by the use of two types - a high gain voltage amplifier and a power amplifier.

17.2 The valves in general use in all earlier long line equipment plant are the types 101, 102, 104 and V.T. 25. The first three are of the well-known dull emitter type. These valves have spherical glass bulbs containing vertical electrodes and an anode and grid on each side of the cathode.

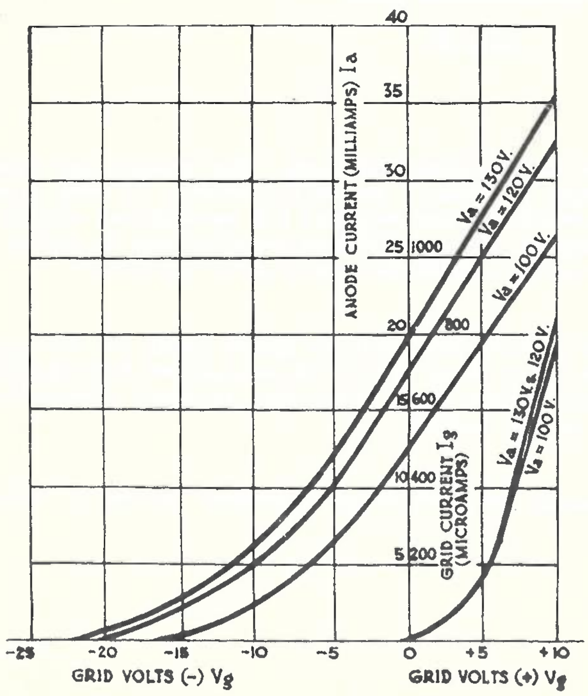
Valves 101, 102 and 104 are usually referred to as "1 ampere" valves by virtue of their normal cathode current.

/Valve

VALVE 101 is a general purpose type, which serves as an amplifier or a detector, and is capable of handling medium currents. This valve is used in telephone repeaters and carrier systems as an amplifier, and in telegraph carrier systems and measuring apparatus as oscillator, amplifier and detector. Valve 101 has a large cathode (of M shape) surrounded by a wide meshed grid, fairly close to which are two parallel plates of equal size to the grid forming the anode.

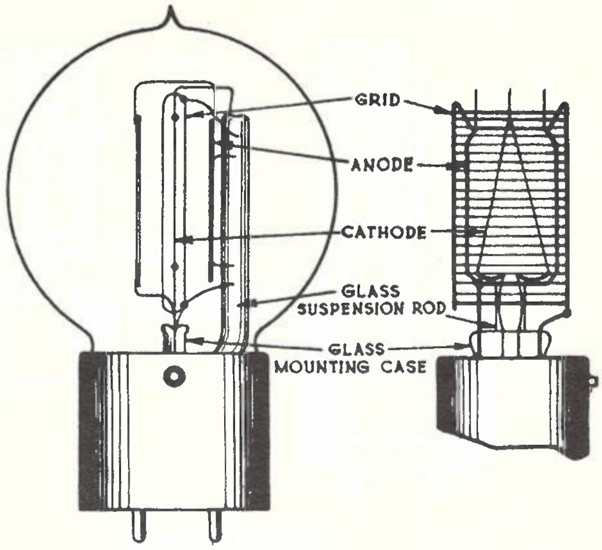


Element Assembly.
(With Front Anode and Grid Removed)

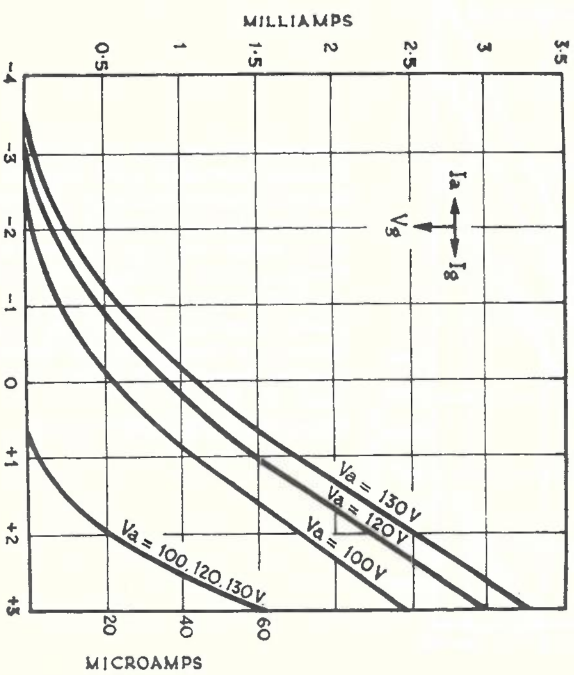


Cathode Voltage 4.4 volts
Cathode Current 0.97 ampere
Working Anode Voltage 130 volts
Working Anode Current 8 mA
Max. Anode Voltage 160 volts
Working Grid Voltage -9 volts
Amplification Factor 5.9
Output as Oscillator 1 watt
Output Power 0.059 watt
Anode Impedance 6,000 ohms
Gain 29.5 db
Expected Life 20,000 hours.

VALVE 102 is a voltage amplifier having a very large amplification factor. This valve, which can only handle small powers, invariably precedes a valve having larger output, such as Valve 101. Where a large power is not required Valve 102 can be used as a detector; for example, as in transmission measuring equipment. It is employed as a voltage amplifier in telegraph carrier systems. Valve 102 has a large, finely-meshed grid mounted close to a small cathode. Parallel plates forming the anode are provided but these are of much smaller dimensions than the grid, from which they are well separated.

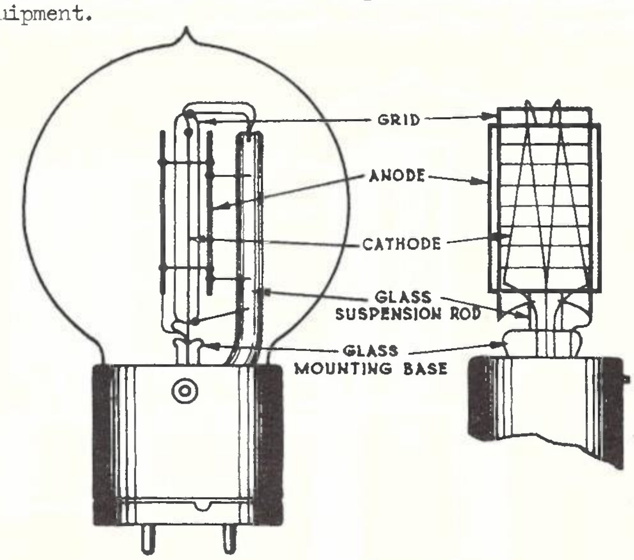


Element Assembly
(With Front Anode and Grid Removed)

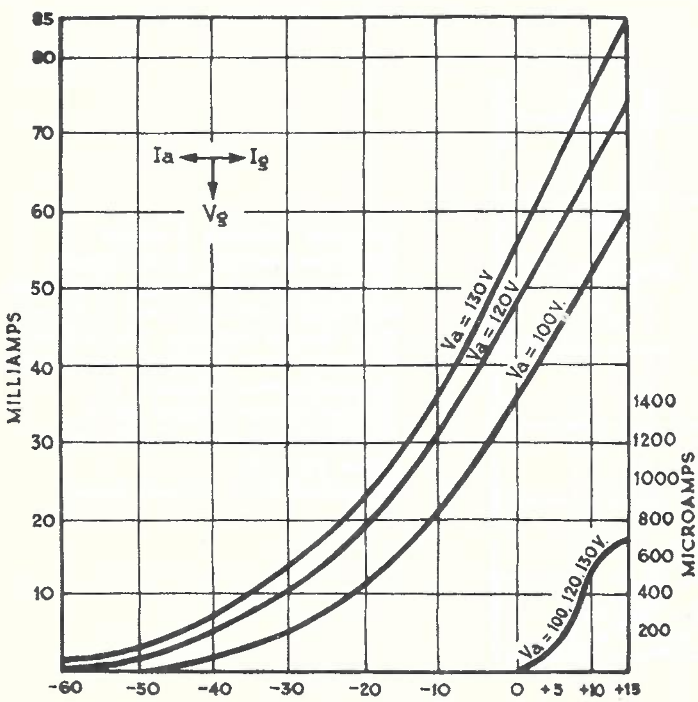


Cathode Voltage 2 volts
Cathode Current 0.97 ampere
Working Anode Voltage 130 volts
Working Anode Current 0.75 mA
Max. Anode Voltage 160 volts
Working Grid Voltage -1.5 volts
Amplification Factor 30
Output Power 0.0042 watt
Anode Impedance 60,000 ohms
Gain 33.5 db
Expected Life 20,000 hours.

VALVE 104 is a power valve used where high outputs into low impedance loads are required. Whilst generally similar in construction to Valve 101, the anodes of Valve 104 are closer to the cathode to give a low impedance. The amplification factor of Valve 104 is 2.4 compared with 6 of Valve 101 which usually precedes it in an amplifier circuit. Valve 104 is used mainly as an output valve in carrier amplifiers, as well as in measuring equipment.

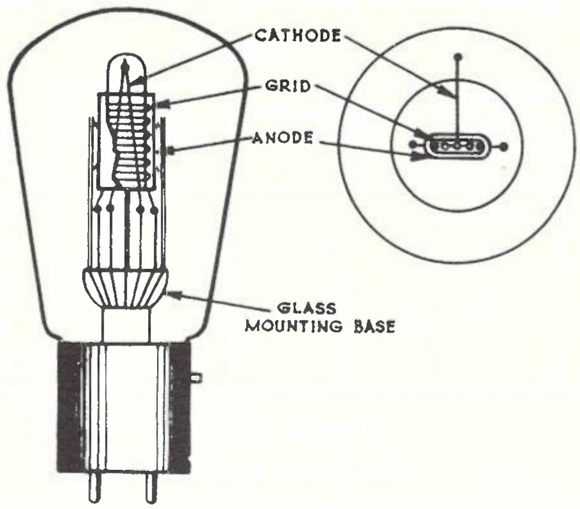


Element Assembly
(With Front Anode and Grid Removed)



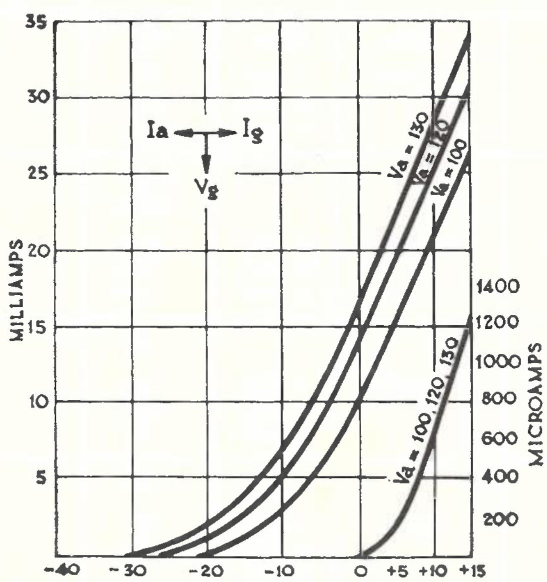
Cathode Voltage 4.4 volts
Cathode Current 0.97 ampere
Working Anode Voltage 130 volts
Working Anode Current 20 mA
Max. Anode Voltage 160 volts
Working Grid Voltage -20 volts
Amplification Factor 2.4
Anode Impedance at 20 mA 2,000 ohms
Output as Oscillator 2 watts
Output Power 0.17 watt
Gain 26 db
Expected Life 5,000 hours.

VALVE V.T. 25 conforms to a British Post Office specification and is standardised for use in G.E.C. type telephone repeaters and other English equipment. This valve is similar in performance to the S.T.C. Valve 101. The construction, however, is different. Instead of a spherical glass bulb it has a bulb resembling in shape an ordinary electric lamp. The cathode, like that of Valve 102, is V-shaped. Bent around the cathode, and in the same plane as it, is a wire loop support (resembling a hair-pin) and on this loop is wound a fine spiral of wire, the whole assembly forming a grid encasing the cathode. Outside this again is mounted a metal tube of elliptical section which forms the anode. Thus, the cathode is completely surrounded by grid and anode. Thoriated tungsten is used in the cathode, which has a fairly bright glow at normal current, as compared with that of any of the S.T.C. series which are dull emitters.



Element Assembly

(With Portion of Anode Broken Away to Show Grid and Cathode)



Cathode Voltage 4.5 volts
Cathode Current 0.8 ampere
Working Anode Voltage 150 volts
Working Anode Current 11 mA
Working Grid Voltage -8 volts
Amplification Factor 5.6
Output Power 0.06 watt
Anode Impedance 6,000 ohms
Approx. Gain 29 db
Expected Life 10,000 hours.

17.3 On later S.T.C. equipment, a new series was introduced which consumed only 1/4 ampere cathode current.

The 1/4 ampere valves are designated 4019, 4020, 4021 and 4022, respectively. The first three have similar characteristics to Valves 101, 102 and 104, respectively. Valve 4022 resembles Valve 101 but has a greater amplification. The following pages illustrate this series, and the particulars given thereon should be studied. It will be seen that the construction differs considerably from that of the earlier types. These valves more closely resemble Valve V.T. 25 in envelope shape and arrangement of electrodes. The latter, however, are set at an angle of approximately 45° to the axis.

The constants and typical working conditions of these four valves are as follows -

CONSTANTS.

Valve	Cathode Current	Nominal Cathode Voltage	Amplification Factor	Mutual Conductance	Capacities Grid Anode	Capacities Anode Cathode	Grid Cathode
	Ampere	Volts		mA/V	μμF	μμF	μμF
4019	0.25	4	* 7	*1.17	5.4	4.8	7.3
4020	0.25	2	*30	*0.6	6.3	4.4	6.2
4021	0.25	4	ø 6	ø3	9.1	4.6	8.0
4022	0.25	4	*11	*1.83	9.6	4.3	8.1

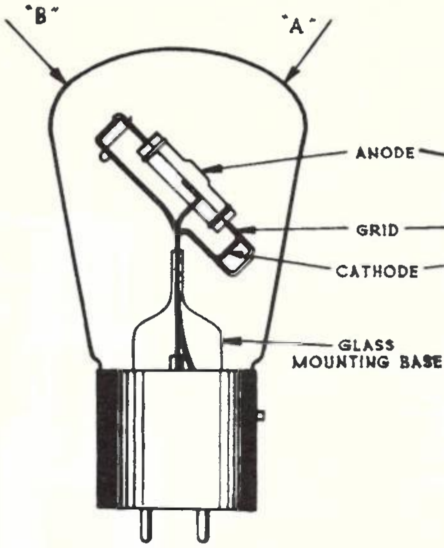
*At anode current of 0.8 mA.
øAt anode current of 23 mA.

TYPICAL WORKING CONDITIONS.

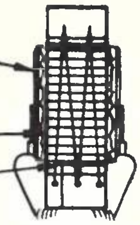
Valve	Anode Voltage	Grid Bias	Anode Current	Anode Resistance	Load	Output	Harmonic Content db below Fundamental
	Volts	Volts	mA	Ohms	Ohms	mW	db
4019	130	-8	8.1	5,100	5,100	79	25
					10,200	63	31
					15,300	50	34
						Peak Volts	
4020	130	-1.5	0.92	46,000	46,000	21.4	30
					92,000	25.4	35.3
					138,000	26.2	37
						mW	
4021	130	-8	22.5	2,050	2,050	140	27
					4,100	100	31.5
					6,150	87	35
						mW	
4022	130	-4.5	6.6	5,000	5,000	120	20
					10,000	79	28
					15,000	76	33

VALVE 4019. Wide Grid Mesh (18 turns), M shaped cathode close to grid, performs similar functions to Valve 101.

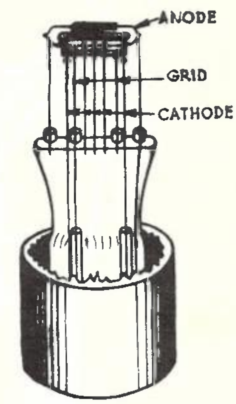
Use. Restricted at present to programme carrier systems, for example, in modulators and demodulators.



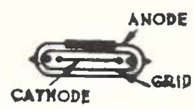
Element Assembly



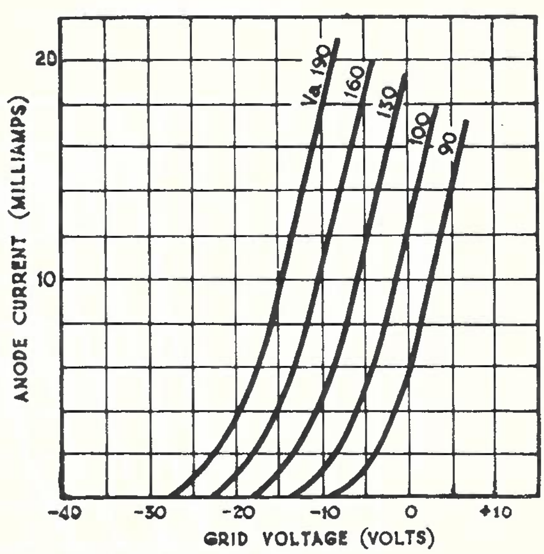
View From "A" (With Portion of Anode Removed to Show Grid and Cathode)



View From "B"



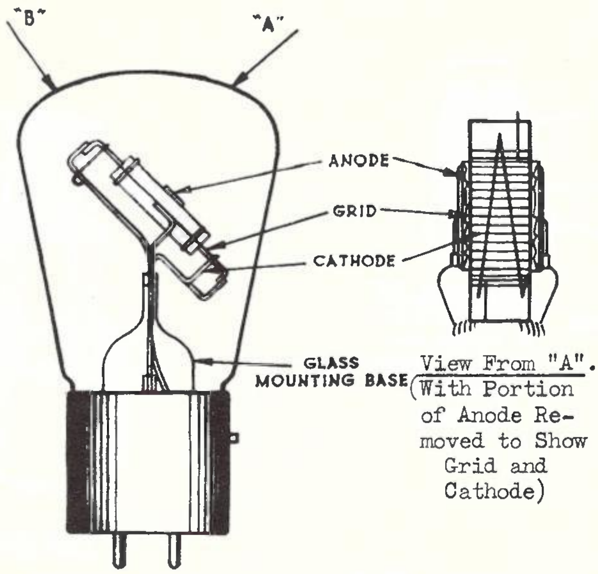
Electrode Spacing.



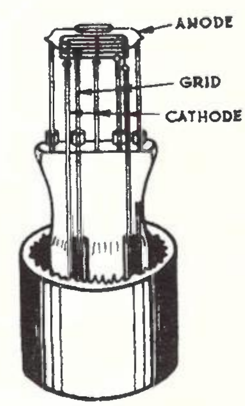
VALVE 4020.

Fine grid mesh (32 turns), V shaped cathode close to grid, plates spaced further apart, similar functions to Valve 102.

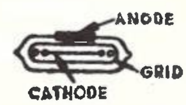
Use. Restricted at present to programme carrier systems as oscillators or amplifiers. Also in 1,000 cycle V.F. ringers.



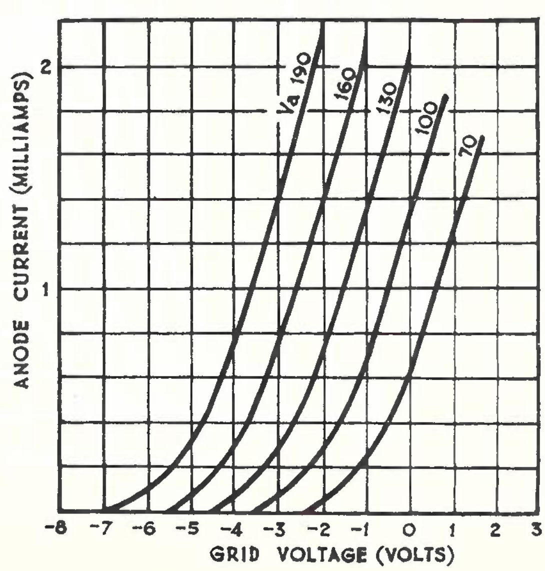
Element Assembly.



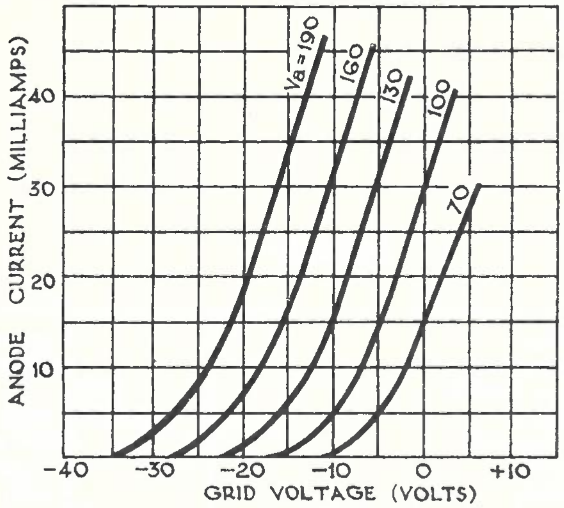
View From "B"

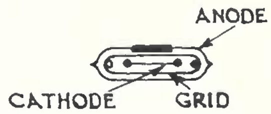


Electrode Spacing.



VALVE 4021. Medium grid mesh (24 turns), M shaped cathode, plates forming anode close together. Intended as a replacement for Valve 104. Application limited.





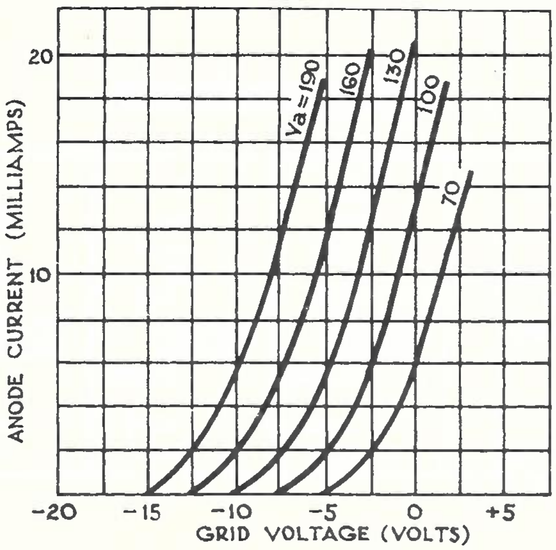
Electrode Spacing.

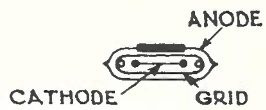
VALVE 4021 CURVES.

Element Assembly is similar to Valve 4019.
See Valve Sections Page 31.

VALVE 4022. Fine grid mesh (30 turns), M shaped cathode close to grid, plates forming anode close to grid. Similar functions to Valve 4019.

Use. Restricted at present to programme carrier systems where higher gain than offered by Valve 4019 is required, for example, pilot amplifier.





Electrode Spacing.

VALVE 4022 CURVES.

Element Assembly is similar to Valve 4019.
See Valve Sections Page 31.

17.4 **VALVE 4046A.** A type of pentode used in departmental equipment is the Valve 4046A, which is an indirectly heated valve. The constants and typical working conditions of the valve are -

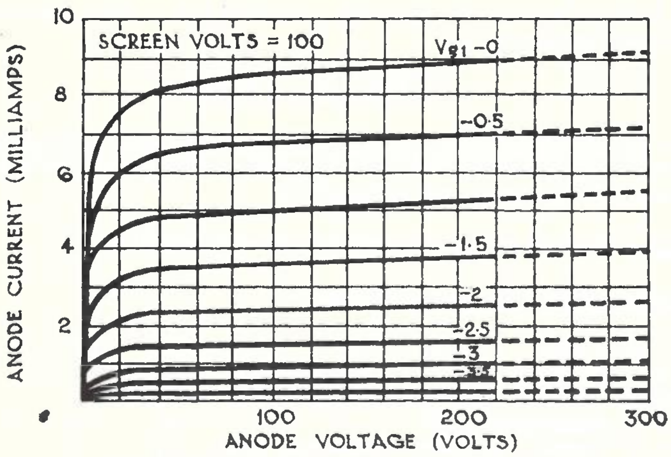
CONSTANTS.

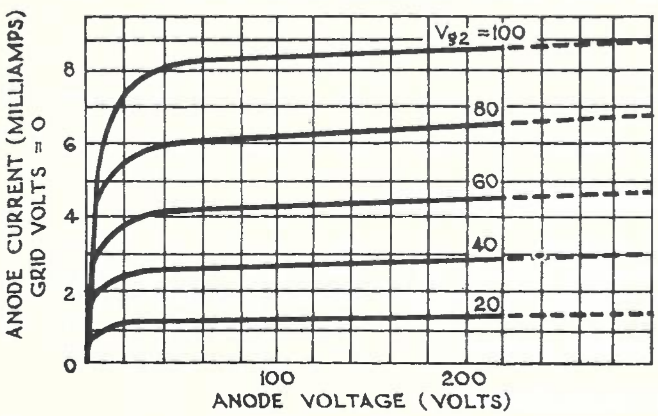
Heater Voltage	4 volts.
Nominal Heater Current	0.95 ampere.
* Amplification Factor	2,400
* Anode Impedance	800,000 ohms.
Grid-Anode Capacity	0.007 $\mu\mu F$.
Input Capacity	10.7 $\mu\mu F$.
Output Capacity	8 $\mu\mu F$.
Maximum Safe Direct Anode Voltage	250 volts.
Maximum Safe Direct Screen Voltage	100 volts.
* At $V_a = 200$, $V_{sg} = 100$, $V_{g1} = -1.5$.	

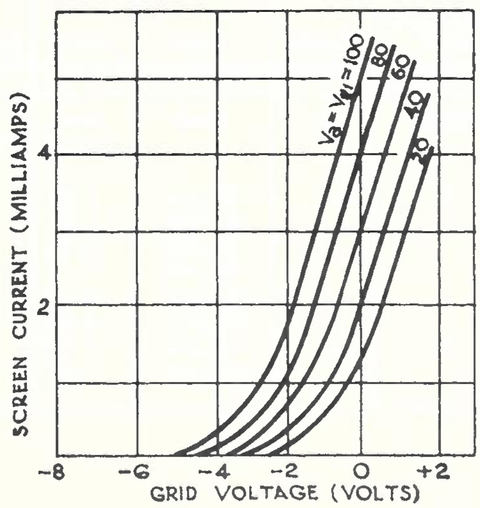
TYPICAL WORKING CONDITIONS.

	(1)	(2)	(3)
Anode Voltage	250	200	150 volts.
Grid Bias	-1.5	-1.5	-1.5 volts.
Screen-Grid Voltage	100	100	100 volts.
Anode Current	3.9	3.8	3.7 mA.
Anode Impedance	500,000	500,000	500,000 ohms.
Load	50,000	46,000	30,000 ohms.
Output	0.315	0.258	0.175 watt.
2nd Harmonic (percentage)	8.6	8.75	11.5 per cent.
2nd Harmonic (db)	21.3	20.7	19 db.

The characteristic curves of Valve 4046A are shown below.







CHARACTERISTIC CURVES OF VALVE 4046A.

17.5 Pentode valves of the type 328A (or 310A) and 329A (or 311A) were later employed in long line equipment. These valves (the former a high gain voltage amplifier and the latter a power amplifier) were specifically designed for carrier systems, and have an exceedingly long life under normal operating conditions. The characteristics of each valve are listed hereunder -

VALVE 328A.

Classification - Voltage-amplifier, suppressor-grid pentode with indirectly heated cathode. This valve is intended for use in audio, carrier and radio-frequency voltage amplifiers, oscillators and modulators.

Mounting Position - The valve may be mounted in any position.

Average Direct Inter-electrode Capacitances -

Control grid to anode	0.007 μμF
Suppressor grid to anode	12.5 μμF
Anode to heater, cathode and screen grid	3.2 μμF
Control grid to suppressor grid	1.3 μμF
Control grid to heater, cathode and screen grid	6.5 μμF
Suppressor grid to heater, cathode and screen grid (with close-fitting metal shield connected to cathode)... ...	14.5 μμF

Heater Rating -

Heater Voltage	7.5 volts, A.C. or D.C.
Nominal Heater Current	0.425 ampere

Limiting Conditions for Safe Operation -

Maximum anode voltage	250 volts
Maximum screen-grid voltage	180 volts
Maximum space current (screen-grid current plus anode current)	10 milliamperes
Maximum screen-grid current	2.5 milliamps.

Operating Conditions and Output -

Anode Voltage	Screen Grid Voltage	Control Grid Bias	Suppressor Grid Voltage	Anode Current	Load Resistance	Input Voltage	Output Voltage	Output Power	2nd Harmonic	3rd Harmonic
Volts	Volts	Volts	Volts	mA	Ohms	Peak Volts	Peak Volts	mW	db	db
135	135	-3	0	5.4	20,000	3.00		250	22	30
					60,000	1.60		130	26	28
					60,000	0.95		60	35	45
					60,000	1.15	100	-	33	39
					100,000	0.57	75	-	35	50
					100,000	0.40	50	-	40	55
180	135	-3	0	5.4	40,000	2.70	-	340	26	28
					100,000	1.50	175	-	26	30

VALVE 329A.

Classification - Low power, suppressor-grid pentode with indirectly heated cathode. This valve is intended primarily for use as an audio, carrier or radio-frequency power amplifier where power outputs of approximately 2 watts are required and where the anode voltage is not in excess of 250.

Mounting Position - The valve may be mounted in any position.

Average Direct Inter-Electrode Capacitances -

Control Grid to anode	0.07 μμF
Control Grid to heater, cathode and screen grid ...	9 μμF
Anode to heater, cathode and screen grid	12 μμF

Heater Rating -

Heater Voltage	7.5 volts, A.C. or D.C.
Nominal Heater Current	0.85 ampere

Limiting Conditions for Safe Operation -

Maximum direct anode voltage	250 volts
Maximum direct screen-grid voltage	180 volts
Maximum space current	60 milliamperes
Maximum direct screen-grid current	12 milliamperes

Operating Conditions and Output -

Anode Voltage	Amplification Factor	Anode Impedance	Trans-conductance	Anode current	Load Resistance	Input Voltage	Output Power	2nd Harmonic	3rd Harmonic
Volts		Ohms	Micro-ohms	mA	Ohms	Peak Volts	Watts	db	db
135	122	43,000	2,800	30	3,000	15	1.9	23	24
					3,500	15	2.0	27	21
					4,000	15	2.0	29	19
					6,000	15	1.9	20	18
180	146	50,000	2,900	31	3,000	15	2.5	18	30
					4,000	15	2.8	21	24
					7,000	15	2.5	23	18

17.6 In more recent years, the tendency has been to employ valves designed primarily for radio purposes. The two valves chosen for general use as a voltage amplifier and power amplifier are the 6SJ7 and 6V6, respectively.

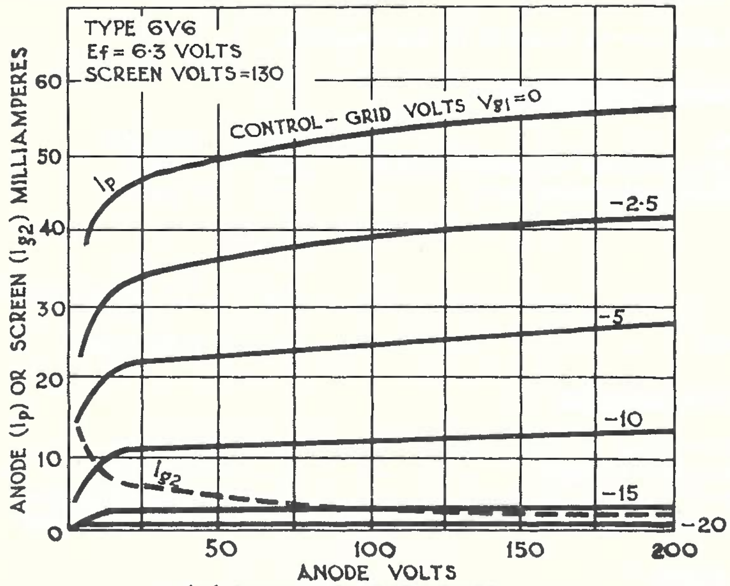
The reasons for employing this class of valve are as follows -

(i) The variation in the electrical characteristics of these types is sufficiently small to permit the use of these valves in carrier equipment.
(ii) Lower heater current than the standard types of pentode previously employed.
(iii) Readily obtainable (Australian manufacture).
(iv) Lower cost.

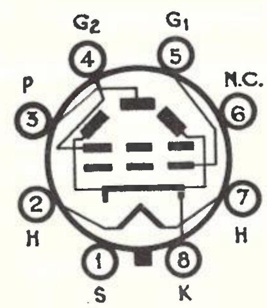
/17.7

17.7 **VALVE 6V6** (metal) or 6V6G (glass) is a power amplifier valve of the beam type, in which use is made of directed electron beams to contribute substantially to its power handling capability.

Valve 6V6 may be used as an audio, carrier or radio frequency power amplifier, and has characteristics as listed hereunder -



6V6 ANODE CHARACTERISTICS.



(The Octal Socket which accommodates the 6V6 Valve may be mounted to hold the Valve in any position.)

6V6 CONNECTIONS.

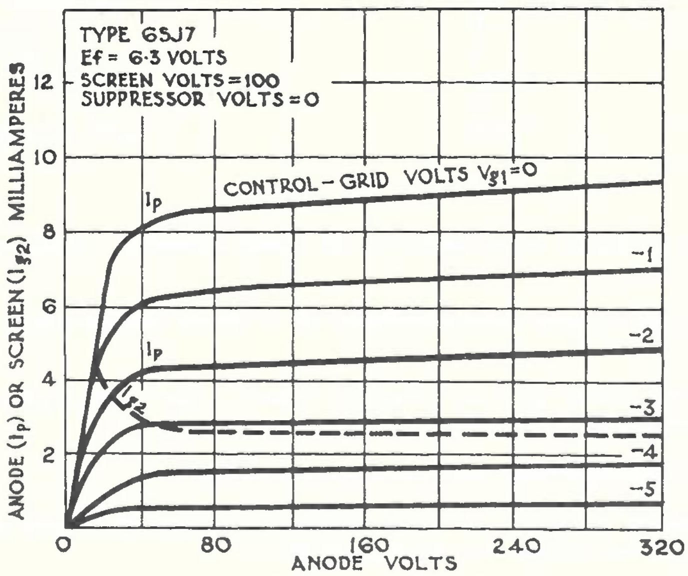
CHARACTERISTICS OF 6V6 VALVE AS A SINGLE-VALVE CLASS A AMPLIFIER.

Heater Voltage (A.C. or D.C.)			6.3 volts
Heater Current			0.45 ampere
Anode Voltage			315 max. volts
Screen Voltage			250 max. volts
Anode Dissipation			12 max. watts
Screen Dissipation			2 max. watts
Typical Operation:			
Anode Voltage	180	250	315 volts
Screen Voltage	180	250	225 volts
Grid Voltage	-8.5	-12.5	-13 volts
Peak A.F. Grid Voltage	8.5	12.5	13 volts
Zero-Signal Anode Current	29	45	34 milliamperes
Max.-Signal Anode Current	30	47	35 milliamperes
Zero-Signal Screen Current	3	4.5	2.2 milliamperes
Max.-Signal Screen Current	4	7	6 milliamperes
Anode Impedance	58,000	52,000	77,000 ohms
Transconductance	3,700	4,100	3,750 micromhos
Load Resistance	5,500	5,000	8,500 ohms
Total Harmonic Distortion	8	8	12 per cent.
Max.-Signal Power Output	2	4.5	5.5 watts

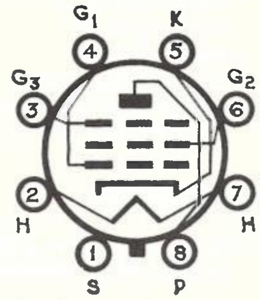
/17.8

17.8 **VALVE 6SJ7** is a voltage amplifier pentode of the metal type featuring single ended construction with interlead shielding. This shielding reduces the capacitance between leads in the glass of the stem and also between those pins that are diametrically opposite. In comparison with the grid cap type previously available, this valve offers the circuit advantages of more stable amplifier operation, greater uniformity of gain in amplifiers and higher gain.

The 6SJ7 Valve may be used in audio, carrier and radio frequency voltage amplifiers, oscillators and modulators, and has characteristics as listed hereunder -



6SJ7 ANODE CHARACTERISTICS.



(The Octal Socket which accommodates the 6SJ7 Valve may be mounted to hold the Valve in any position.)

6SJ7 CONNECTIONS.

CHARACTERISTICS OF 6SJ7 VALVE.

Heater Voltage (A.C. or D.C.)	6.3 volts
Heater Current	0.3 ampere
Pentode Connection (shell connected to cathode):	
Grid-Anode Capacitance	0.005 max. μμF
Input Capacitance	6 max. μμF
Output Capacitance	7 max. μμF
Triode Connection (shell connected to cathode, screen and suppressor connected to anode):	
Grid-Anode Capacitance	2.8 max. μμF
Grid-Cathode Capacitance	3.4 max. μμF
Anode-Cathode Capacitance	11 max. μμF

6SJ7 VALVE AS CLASS A AMPLIFIER - PENTODE CONNECTION.

Anode Voltage		300 max. volts
Screen Voltage (Grid No. 2)		125 max. volts
Screen Supply Voltage		300 max. volts
Grid Voltage (Grid No. 1)		0 min. volt
Anode Dissipation		2.5 max. watts
Screen Dissipation		0.3 max. watt
Typical Operation:		
Anode Voltage	100	250 volts
Screen Voltage	100	100 volts
Grid Voltage	-3	-3 volts
Suppressor	Connected to cathode at socket.	
Anode Current	2.9	3 milliamperes
Screen Current	0.9	0.8 milliampere
Anode Impedance	0.7	More than 1 megohm
Transconductance	1,575	1,650 micromhos
Grid Voltage (for cathode-current cut-off)	-9	-9 volts

6SJ7 VALVE AS CLASS A AMPLIFIER - TRIODE CONNECTION (SCREEN AND SUPPRESSOR TIED TO ANODE).

Anode Voltage		250 max. volts
Grid Voltage		0 min. volt
Anode Dissipation		2.5 max. watts
Typical Operation:		
Anode Voltage	180	250 volts
Grid Voltage	-6	-8.5 volts
Anode Current	6	9.2 milliamperes
Anode Resistance	8,250	7,600 ohms
Amplification Factor	19	19
Transconductance	2,300	2,500 micromhos

18. TEST QUESTIONS.

1. Explain how the grid of a triode valve controls the anode current. Illustrate answer with diagrams.

2. Explain why the static and dynamic characteristics of a triode differ and in what way they differ.

3. What is the object of the screen grid in a tetrode valve? How is this grid used to produce a high amplification factor?

4. What is the purpose of the suppressor grid in a pentode valve?

5. Explain why it is necessary to apply the anode voltage of a mercury vapour valve some time after the cathode voltage has been applied.

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

LONG LINE EQUIPMENT I.

AMPLIFIERS.

CONTENTS:

1. INTRODUCTION.

1.1 In the sections of Paper No. 5 dealing with valve constants and their significance, it was shown that when an alternating voltage of amplitude E_s is applied across the grid and cathode of a valve an alternating voltage of amplitude μE_s is developed in the anode circuit. This higher voltage is produced by the amplifying action of the valve. It was also shown that a valve may be regarded as a generator having an internal impedance of r_a and developing a voltage of μE_s. This equivalent circuit will be used in discussing the action of amplifiers throughout this Paper.

2. PHASE SHIFT THROUGH VALVES.

2.1 The alternating voltage developed in the anode circuit of a valve is 180° out of phase with the alternating signal voltage applied across grid and cathode, that is, a phase shift of 180° takes place between the grid and anode circuits of a valve. This can be readily seen from the dynamic characteristics of valves previously dealt with. As the alternating signal voltage on the grid of a valve drives the grid from some negative voltage towards zero the anode current increases, thus increasing the voltage drop across the load impedance in the anode circuit and so lowering the anode voltage. This is shown in Fig. 1, from which it will be seen that the alternating voltage acting in the anode circuit is 180° out of phase with the signal voltage at the grid.

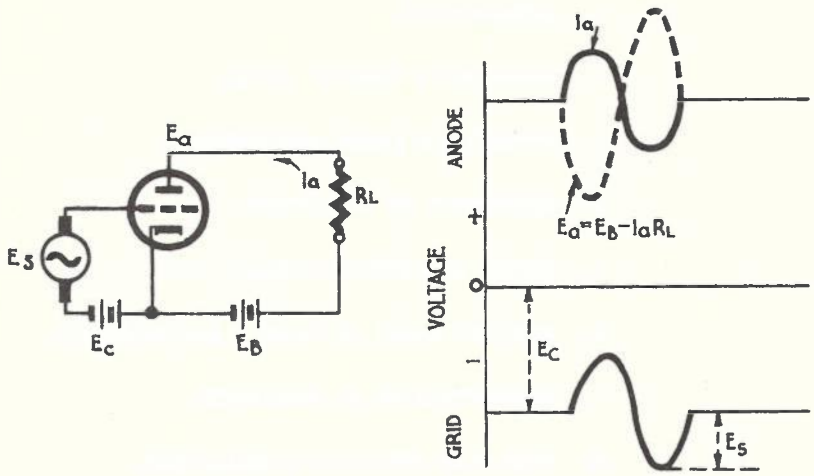


FIG. 1. ILLUSTRATION OF PHASE SHIFT THROUGH VALVES.

3. VOLTAGE AND POWER AMPLIFIERS.

3.1 Telephone receivers, loud-speakers, etc., are power operated devices in that the larger the amount of power supplied to them the louder will be the sound produced by them. It is not necessary to supply power to these devices at a fixed voltage, and, under this condition, a large amount of power will be transferred from a supply source to its load when the following conditions are fulfilled -

(i) The supply source should develop a high terminal voltage.
(ii) The supply source should have little impedance, so that little power is absorbed by this impedance.
(iii) The impedance of the load should equal that of the supply source for maximum power transfer.

3.2 When a valve is used as a supply source, it will be found that one valve cannot fulfil all of these conditions. Thus, for the alternating voltage developed in the anode circuit, μE_s, to be high, the amplification factor, μ, must be high. As $\mu = r_a g_m$, a high μ of necessity means a high r_a, which conflicts with the second condition that the supply source should have a low impedance. Also, all generators develop their maximum voltage on open circuit, that is, when there is very little or no current output to produce a voltage drop across the internal impedance of the generator, which drop, of course, lowers the terminal voltage. Thus, a high output voltage would be obtained when the load impedance is high, which conflicts with the third condition, that is, that the load and supply impedances should be matched. It will be apparent from these considerations that one valve cannot be used to amplify a signal voltage by a large amount and, at the same

/time

time, supply a large amount of power to a load. For this reason, an amplifier generally consists of a number of valves, or "stages" as the individual valves are called. These stages are connected in cascade, the first stages being voltage amplification stages and the last stage, which supplies power to the load, being a power amplifier. The voltage stages use valves with a high μ, and, therefore, a high r_a, which work into load impedances which are generally fairly high.

3.3 By using a number of such stages in cascade, almost any degree of voltage amplification can be obtained. The amplified voltage is then applied to the power stage, which consists of one or a number of valves with a low r_a, the output of which stage is matched with that of the load, usually by means of a transformer. Little voltage amplification is produced by this stage, as a valve with a low r_a will have a low μ. This is not important, however, as the preceding voltage stages have amplified the voltage to the value required. Fig. 2 indicates the layout of the amplifier.

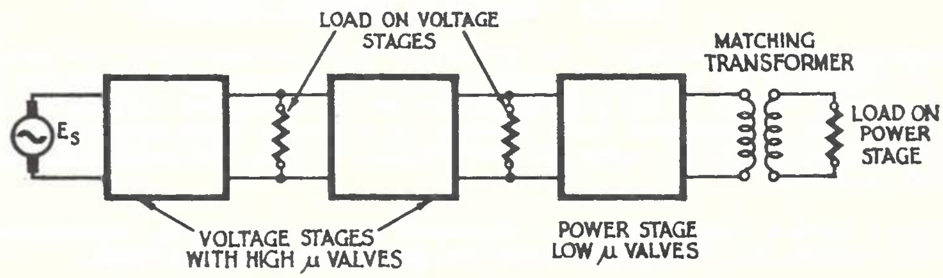


FIG. 2. MULTI-STAGE AMPLIFIER.

4. DISTORTION IN AMPLIFIERS.

4.1 An ideal amplifier produces an output which exactly duplicates the input in all respects except amplitude, in which respect the output is greater than the input. An actual amplifier can fall short of the ideal in three ways -

(i) By failing to amplify the different frequency components in the input voltage equally well.

(ii) By producing an output which does not bear the same proportion to the input over the whole of the signal.

(iii) By making the phase relations between the different frequencies in the output differ from those existing at the input.

These effects are referred to as Frequency, Harmonic or Non-linear, and Phase distortion, respectively. The effect is to change the shape of, or distort, the signal in its passage through the amplifier.

4.2 Frequency distortion is particularly important in amplifiers, and is more difficult to eliminate the wider the band of frequencies to be amplified. Frequency distortion is caused by the frequency characteristics of the input and output circuits associated with the valves. For example, if the anode circuit impedance is largely inductive, the value of this impedance at lower frequencies will be lower than at

/high

high frequencies. Thus, in Fig. 3, an input signal contains a fundamental and a harmonic. Due to the nature of the anode impedance, the harmonic is practically eliminated in the output of the amplifier and the signal is changed in shape or distorted in its passage through the amplifier.

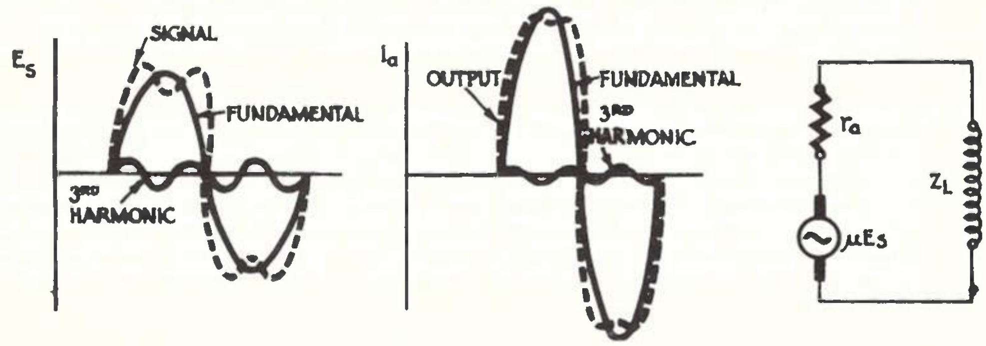


FIG. 3. EXAMPLE OF FREQUENCY DISTORTION.

4.3 Harmonic distortion is due to the production of new frequencies in the output which were not present at the input. The anode current wave-shape will be different from that of the signal voltage if the grid swings include the non-linear portions of the characteristic curve, because over those portions the change in anode current has a variable proportionality to the change in grid voltage, that is, the mutual conductance, g_m, is not constant.

Fig. 4 shows a typical case. A sinusoidal signal voltage, E_s, superimposed on a steady D.C. voltage, E_c, is applied across the grid and cathode of a valve.

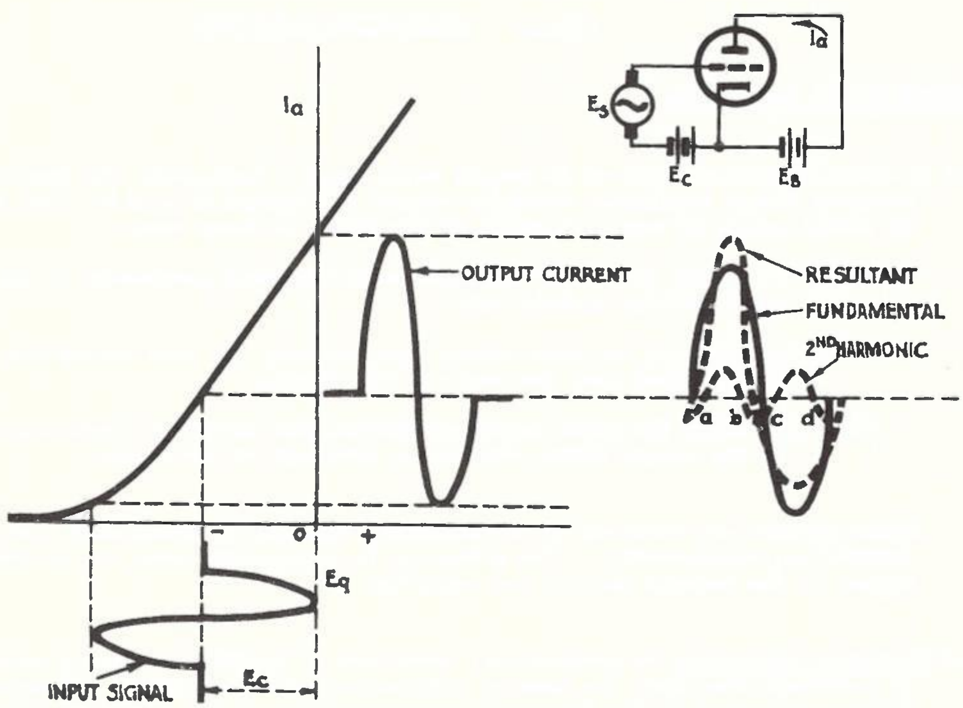


FIG. 4. NON-LINEAR DISTORTION PRODUCED BY VALVES.

This steady D.C. voltage, E_c, fixes the point of application of the signal voltage somewhere in the negative portion of the valve characteristic. The amplitude of the signal voltage is such that it drives the grid to zero during one half-cycle and into the lower bend during the other half-cycle. Under this condition, the two half-cycles of alternating current in the output will be of unequal amplitude, that is, the half-cycle above the zero or normal anode current value of Fig. 4 will be greater than the half-cycle below it. An analysis of such a wave-shape shows that it consists of a fundamental frequency of the signal frequency, together with harmonics which, when added to the fundamental, give the distorted wave-shape produced by the non-linearity of the characteristic curve. In Fig. 4, only the second harmonic is shown in the analysis - others are produced but their amplitude decreases with their order, that is, the third harmonic is of smaller amplitude than the second, and so on. In the analysis contained in Fig. 4, the second harmonic is in phase with the fundamental over the 180° of the second harmonic designated ab, so increasing the amplitude of the resultant, whilst over the 180° marked cd the second harmonic is 180° out of phase with the fundamental, so decreasing the amplitude of the resultant.

4.4 Phase distortion is, as mentioned, the change in the relative phase relations of the different frequency components in the signal voltage. This shift in phase of the different frequencies may occur without necessarily changing their amplitude relations, resulting in an output wave which is no longer identical in shape with that of the input signal.

This type of distortion is also caused by the characteristics of the input and output circuits and is accompanied by frequency distortion. Fig. 5 shows a case in which a valve, having an amplification factor μ and an anode impedance r_a, has for an anode load a pure inductance of L henrys.

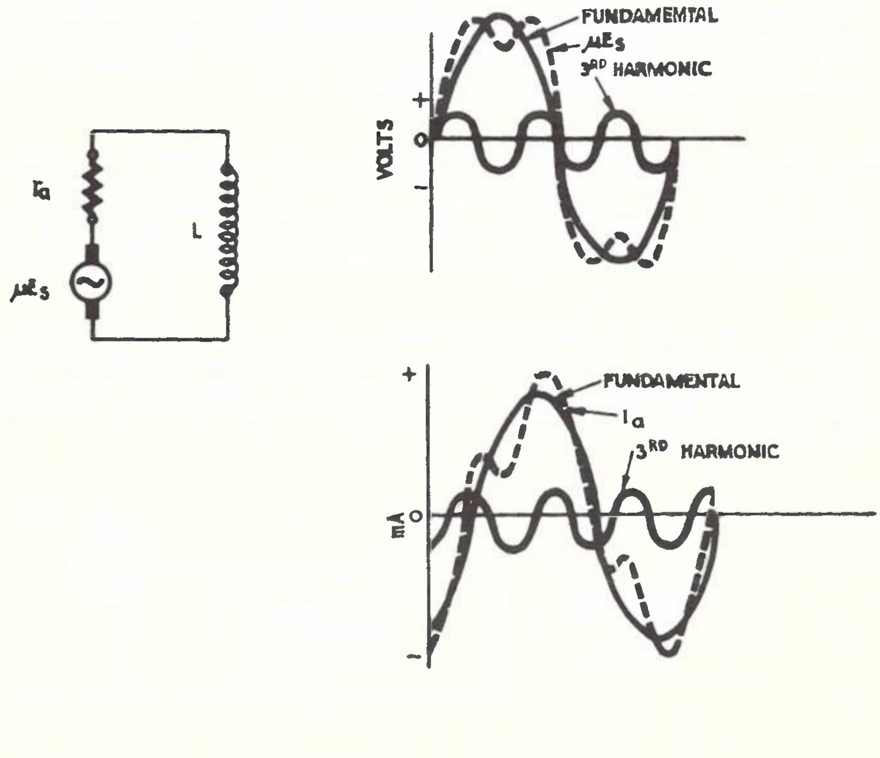


FIG. 5. EXAMPLE OF PHASE DISTORTION.

The signal voltage and, therefore, the alternating voltage acting across the anode circuit is assumed to be composed of a fundamental and a third harmonic. The fundamental component of the anode current will lag its applied voltage by an angle θ, so that -

$$\tan \theta = \frac{\omega L}{r_a} = \frac{2\pi f L}{r_a}$$

The third harmonic component of the anode current will lag its applied voltage by an angle ∅, so that -

$$\tan \emptyset = \frac{2\pi 3 f L}{r_a} = \frac{6\pi f L}{r_a}$$

where f, in each case, is the fundamental frequency.

Thus, the angles by which the fundamental and third harmonic components of the anode current are displaced from their respective voltages are different, and, further, the anode circuit will offer a higher impedance to the third harmonic than to the fundamental, causing frequency distortion as well as phase distortion.

5. EFFECT OF POSITIVE CONTROL GRID.

5.1 An examination of the anode current versus grid voltage curve of most valves will show that the straight or linear portion generally extends from some negative value to some positive value of grid voltage. These curves are usually drawn to show the relation between the voltage actually at the grid and the anode current. If the grid circuit contains impedance, as it does in most applications, the presence of this impedance will make the anode current curve for all positive values of grid voltage curved or non-linear as follows -

5.2 When the grid of a valve is driven positive with respect to the cathode it attracts electrons in the same manner as does the anode, and so grid current flows in the grid circuit, the amplitude of this grid current increasing as the grid is driven more positive. This grid current produces a voltage drop, which is 180° out of phase with the signal voltage, across the impedances in the grid circuit.

5.3 In Fig. 6, the grid is driven positive by means of battery E_c and the resultant grid current produces a voltage drop across the impedance in the grid circuit which is in opposition to, or 180° out of phase with, E_c. Thus, the voltage actually at the grid will be the difference between E_c and the voltage drop across the impedance of the grid circuit.

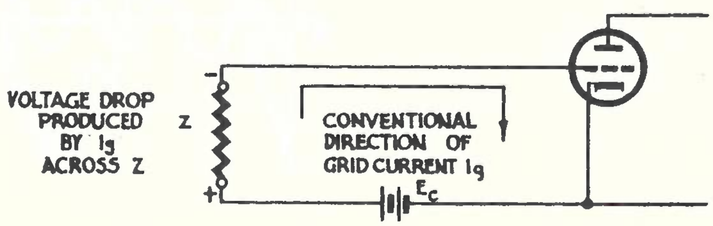


FIG. 6. VOLTAGE PRODUCED BY GRID CURRENT.

5.4 If an alternating voltage of an amplitude high enough to drive the grid positive is introduced into the grid circuit, as in Fig. 7, the amplitude of the voltage actually at the grid will be reduced, as shown, only over that portion of the signal voltage which drives the grid positive. This is because grid current flows only when the grid is driven positive and, therefore, the anti-phase voltage drop is likewise produced over the impedance of the grid circuit only when the grid is driven positive.

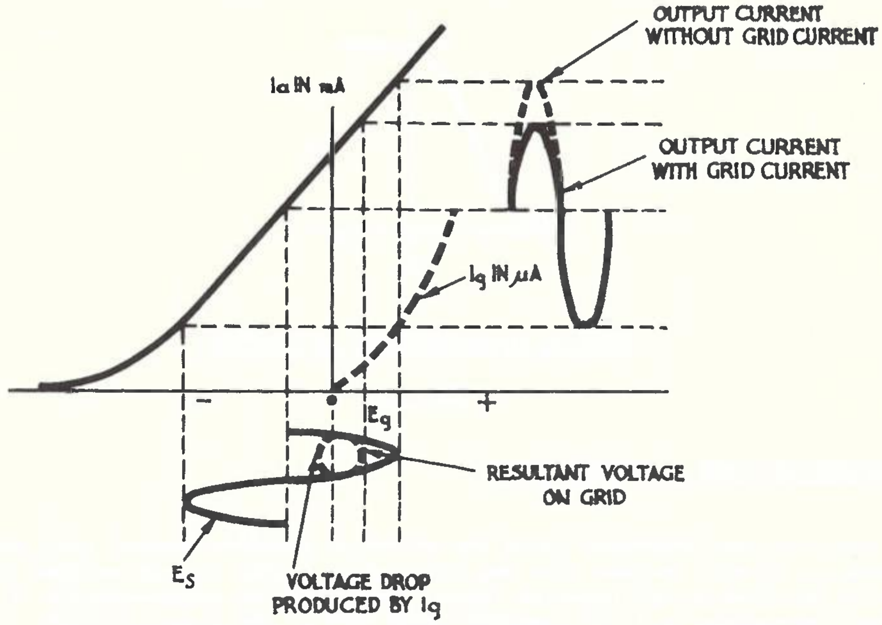


FIG. 7. EFFECT OF GRID CIRCUIT IMPEDANCE.

5.5 This means that the amplitude of those half-cycles of signal which drive the grid positive are reduced at the grid, whilst the other half-cycles are unaffected. The anode current wave-shape, therefore, will be distorted as in Fig. 7, which is the reverse of the anode current wave-shape of Fig. 4. As the distortion in Fig. 4 is caused by the curvature at the lower bend of the valve's characteristic, the distortion produced when the grid is driven positive could be regarded as being caused by curvature of the characteristic at positive grid voltages when impedance is present in the input circuit, as in Fig. 8.

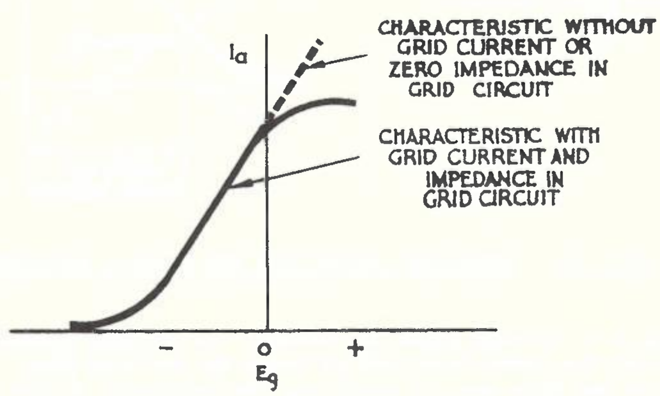
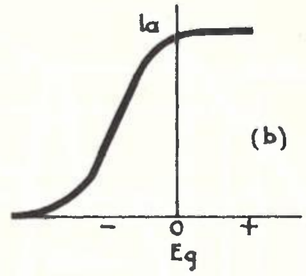


FIG. 8. CURVATURE OF CHARACTERISTIC DUE TO GRID CURRENT.

5.6 Thus, when a triode is driven to grid current a considerable amount of distortion is produced, whilst the normal dynamic characteristic of a pentode (see Fig. 9) will do likewise. It can be shown mathematically that this distortion is due mainly to the production of the third harmonic of the fundamental frequency.



DYNAMIC CHARACTERISTIC OF PENTODE.

FIG. 9.

6. DISTORTIONLESS OR CLASS A AMPLIFICATION.

6.1 From what has been discussed about the effect of positive control grid and lower bend curvature, it will be seen that the alternating signal voltage applied to a valve must be confined to the negative linear portion of the characteristic if the anode current wave-shape is to be identical with that of the signal voltage. This can only be achieved by the arrangements shown in Fig. 10.

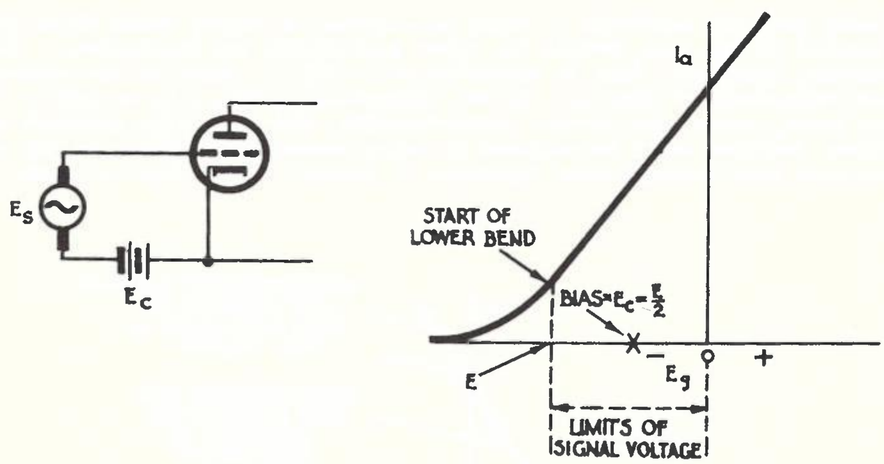


FIG. 10. OPERATING CONDITIONS FOR CLASS A AMPLIFIER.

E_c is a steady D.C. voltage equal to one-half the voltage between zero grid volts and the start of the lower bend, and is known as the Grid Bias. The signal is superimposed on the bias and, by confining the amplitude of the signal to that of the bias, the grid will never be driven positive or into the lower bend. By this means the anode current wave-shape will be identical with that of the signal voltage, that is, no distortion results. This type of amplification is referred to as Class A amplification.

/7.

7. CLASSIFICATION OF AMPLIFIERS.

7.1 Amplifiers are classified in ways descriptive of their character and properties.

7.2 The first classification is according to the frequency to be amplified, for example, audio frequency amplifiers to amplify frequencies from about 15 c/s to about 15,000 c/s, and radio frequency amplifiers to amplify frequencies above about 15,000 c/s.

7.3 Amplifiers are also divided into voltage and power amplifiers for the reasons discussed earlier in this Paper.

7.4 Amplifiers are also classified as to whether they handle a wide or narrow band of frequencies. A band of frequencies is considered wide or narrow in proportion to the ratio of the width of the band to the frequency at its centre. A band of frequencies lying between 100 and 5,000 c/s is said to represent a wide band, whilst the frequency band from 1,000,100 to 1,005,000 c/s, which extends over the same frequency range, is narrow. When substantially equal amplification is to be obtained over a band that is narrow according to this definition, tuned circuits can be associated with the amplifier. In the example used, the tuned circuits associated with the amplifier could be tuned to the frequency at the centre of the band, about 1,002,500 c/s, and the slightly "rounded off" response of a tuned circuit would pass frequencies lying 2,500 c/s on either side of this frequency, because such frequencies differ from the centre frequency by only about $\frac{1}{400}$ part of it. On the other hand, tuned circuits could not be associated with amplifying the wide band mentioned above. A circuit tuned to the centre frequency would be tuned to about 2,500 c/s and would have to pass frequencies twice this frequency and also $\frac{1}{25}$ of it, that is, 5,000 c/s and 100 c/s. Such a circuit is not a tuned circuit, as the simple series or parallel combination of inductance and capacity which forms tuned circuits will not pass frequencies that differ from their resonant frequency by such a large amount. Radio frequency amplifiers are tuned, because the high carrier frequencies employed in radio mean that the band handled by such amplifiers is narrow. Audio and carrier frequency amplifiers are wide band amplifiers and, therefore, untuned, because, although carrier frequencies extend into the radio portion of the frequency spectrum, for example, up to about 140 kc/s on open wire lines, the bands they handle are fairly wide according to the definition used.

7.5 Power amplifiers are also designated Class A, Class B or Class C according to their operating conditions.

The Class A amplifier has already been dealt with and is exclusively used for all amplifiers in carrier terminals, carrier repeaters and V.F. repeaters, because its operating conditions are such that the anode current wave-shape is identical with that of the applied signal voltage, that is, little or no distortion is produced. As an energy converting device, the efficiency of the Class A amplifier is very low. During a signal cycle, D.C. power is drawn from the anode supply source and is converted into A.C. power, this A.C. power being passed on to the load on the amplifier.

A low conversion efficiency cannot be tolerated in amplifiers handling large amounts of power; for example, power amplifiers in broadcast transmitters which sometimes have an A.C. power output of 100 kilowatts. To increase the conversion efficiency, Class B operation is used under some circumstances. Whilst Class B amplifiers have no applications in Carrier Telephone and Telegraph systems, some idea of their operation is necessary to understand Class C operation, which is used in many oscillators in the systems mentioned.

Class B amplifier is biased practically to anode current cut-off, so that anode current flows in pulses lasting for only 180° of each signal cycle. When used as an audio frequency amplifier two valves must be used, one to amplify one half-cycle and the other to amplify the other half-cycle.

Decreasing the angle of anode current flow apparently increases the efficiency because

/the

the efficiency of a Class B amplifier, through which anode current flows for only 180° of a signal cycle, is higher than that of a Class A amplifier through which anode current flows for the whole 360° of a signal cycle. In the Class C amplifier the angle of anode current flow is reduced to between 120° and 150° by biasing the grid beyond anode current cut-off. Class C operation cannot be applied to audio frequency amplifiers since using two valves, as in the Class B case, will produce incomplete half-cycles in the anode circuits.

8. GRID BIAS FOR CLASS A AMPLIFIERS.

8.1 There are many sources of grid bias, for example, dry cells, rectifier units, secondary cells, D.C. generators, etc., portion or all of the voltage they develop being connected between the grid and cathode of the valve to be biased, depending on the bias required and the voltage developed by the items mentioned.

8.2 Another method is to use the voltage drop over resistors in the cathode circuit when directly heated cathodes (filament type) are used. Fig. 11 shows how the bias for the two valves in a S.T.C. voice frequency repeater is obtained.

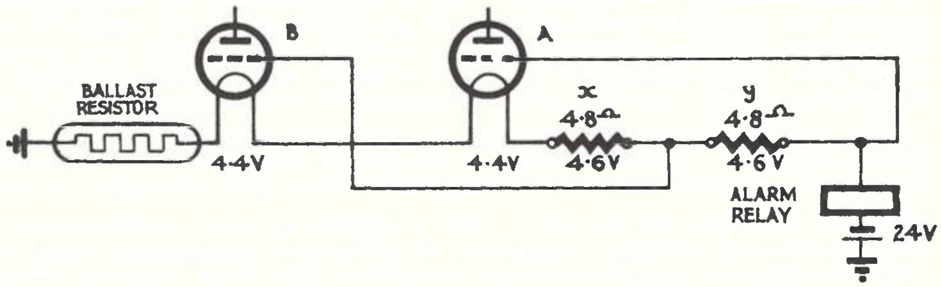


FIG. 11. BIAS ARRANGEMENTS IN S.T.C. VOICE FREQUENCY REPEATER.

The cathode current is 0.97 ampere, and the two valves require about -9 volts bias. Valve B has -9 volts bias, this being provided by the 4.4 volts drop across the cathode of A and the 4.6 volts drop across resistor x. Valve A has -9.2 volts bias, this being provided by the 4.6 volts drop across resistors x and y.

8.3 Perhaps the most widely used method of biasing a valve is by means of a cathode resistance. This type of circuit is suitable for either directly or indirectly heated cathodes.

Fig. 12 shows the arrangement. In Fig. 12 anode current flows in the conventional direction indicated, producing a voltage drop across the cathode resistor R_c in the direction shown.

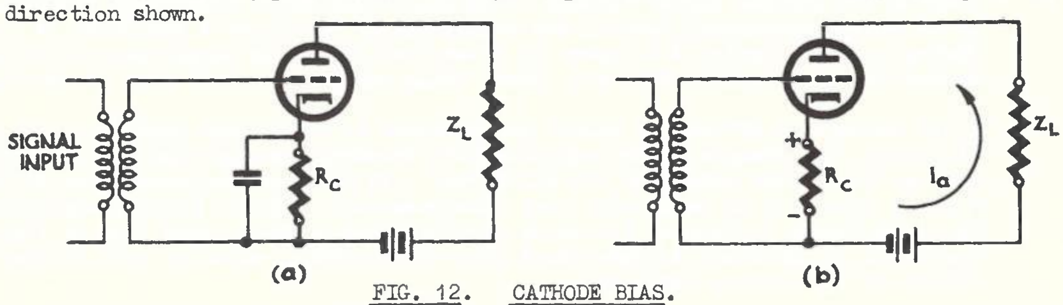


FIG. 12. CATHODE BIAS.

By choosing a suitable value of resistor, the voltage drop across it will equal the bias required and the grid will be biased negatively with respect to the cathode by the required amount. Under signal conditions the alternating component of the anode current flows through R_c as well as Z_L, so varying the voltage drop across R_c and varying the bias. This is prevented in Fig. 12 by connecting across R_c a condenser having a capacity large enough for it to short-circuit R_c effectively at the lowest frequency to be transmitted through the amplifier.

9. GRID BIAS FOR CLASS C AMPLIFIERS.

9.1 Class B amplifiers are used in circumstances such that the grid is driven positive during some signal cycles and not during others. On the other hand, Class C amplifiers are used in circumstances such that the grid is driven well positive during each signal cycle. The harmonics produced by this operation are rejected, of course, by the tuned load in the anode circuit. The grid current produced by driving the grid of a Class C amplifier positive is frequently used to produce the required bias. This grid current (like the anode current) flows in pulses and has an average value, as shown in Fig. 13.

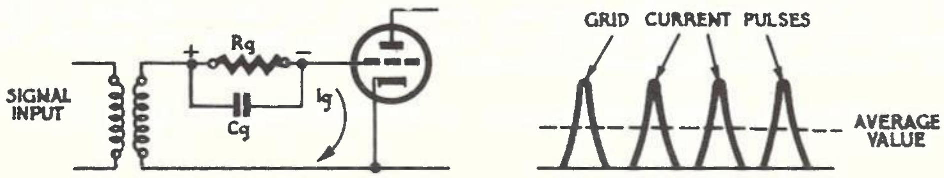


FIG. 13. GRID LEAK BIAS.

This average value of grid current passing through the grid resistor R_g in the conventional direction shown produces an average voltage drop across R_g in the direction necessary to bias the grid negatively by the required amount. The signal voltage, however, is acting into a finite impedance in that the input capacity of the amplifier valve will allow current to flow in the input circuit. This current will produce a voltage drop across R_g and lower the effective input to the valve. This is prevented by shunting R_g with a suitable condenser C_g, which effectively short-circuits R_g at the signal frequency. This resistance-condenser combination is called a "grid leak," and this type of bias is termed "grid-leak" bias.

10. VOLTAGE AMPLIFIERS - METHODS OF COUPLING.

10.1 As mentioned previously, by employing a number of amplifiers in cascade almost any degree of voltage amplification can be obtained. The method of applying the voltage output from one stage to the input of the succeeding stage is called the method or type of coupling. Before proceeding with a discussion on the different types of coupling circuits employed, it will be desirable to qualify the generalisation that voltage amplifiers work into load impedances which are fairly high, as stated in the section on Voltage and Power Amplifiers. The coupling circuit is actually the load impedance on a voltage stage, and its value should be such that maximum voltage amplification is obtained with little distortion. Thus, the impedance of the coupling circuit on a triode voltage amplifier will be higher than that on a pentode, because high load impedances on triodes decrease distortion and on pentodes increase distortion. In the circuits used, typical values will be included for both pentode and triode valves. It must also be pointed out that the coupling methods to be dealt with are for untuned amplifiers, that is, amplifiers for carrier and audio frequencies and not for radio frequencies.

10.2 Resistance Coupling. In this method, a high resistance called a coupling resistance, R_c in Fig. 14, is placed in the anode circuit across which the amplified voltage is developed.

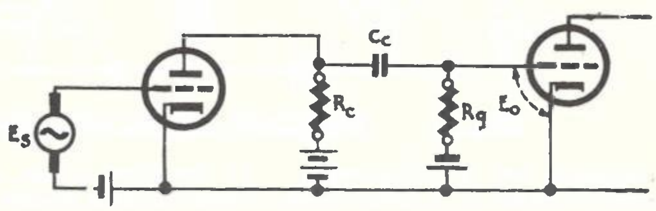


FIG. 14. RESISTANCE COUPLED AMPLIFIER.

The coupling condenser C_c prevents the D.C. voltage applied to the anode of the first stage from being applied to the grid of the valve in the succeeding stage. C_c should be large enough to offer a low reactance to the lowest frequency to be transmitted through the amplifier. Bias is applied to the grid of the second valve via a high resistance R_g. This resistance prevents the connection necessary for providing the bias from shunting the coupling resistance, so lowering the impedance of the load on the first stage with a consequent reduction in amplification.

Typical values for R_c, R_g and C_c, respectively, are 0.1 megohm, 1 megohm and 0.01 microfarad. This applies to both pentode and triode valves. A typical triode voltage amplifier, a 6SF5, has an r_a of 66,000 ohms, whilst a typical pentode, a 6J7, has an r_a of 1 megohm. The impedance provided by the above combination will be greater than the r_a of the triode, but much less than that of the pentode, for distortion considerations discussed previously.

The amplification varies with frequency in the manner shown in Fig. 15.

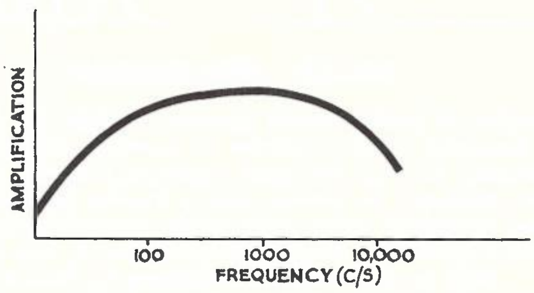


FIG. 15. AMPLIFICATION VERSUS FREQUENCY.
(RESISTANCE COUPLING)

This curve is obtained by applying a constant voltage to the input whilst varying the frequency from zero upwards. The output voltage, which is the voltage input to the next stage, is measured at all frequencies, and the ratio of E_o to E_s, plotted as a function of frequency and called a "frequency response" curve, will give the result shown in Fig. 15. The falling off at low frequencies is a result of the fact that the higher reactance, which the coupling condenser C_c offers to low frequencies, causes a considerable voltage drop to take place across it at those frequencies, resulting in a steadily decreasing voltage drop across R_g and, therefore, across the input to the second stage as the frequency is lowered. The reduction in amplification at high frequencies is caused by the valve and stray wiring capacities shunting the grid and coupling resistances. These capacities have a low enough reactance at high frequencies to lower the effective load impedance with a consequent reduction in the voltage developed across that load impedance.

10.3 Transformer Coupling. In a transformer coupled amplifier, the load impedance connected in the anode circuit of the valve is the primary winding of a transformer, the secondary voltage of which is applied across the grid and cathode of the succeeding valve as shown in Fig. 16.

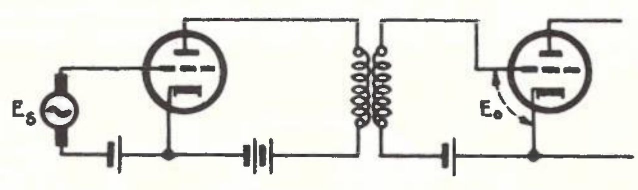


FIG. 16. TRANSFORMER COUPLED AMPLIFIER.

Transformer coupling has two advantages. In the first place, as a voltage step-up can be obtained by means of a transformer, fewer voltage stages are required to

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produce a given voltage amplification by using transformer coupling than by using resistance coupling. In the second place, as the primary winding of a transformer can be wound with little resistance there will be very little of the anode supply voltage dropped across it, so that the anode supply voltage can be lower than with resistance coupling. Fig. 17 shows the way in which the amplification will vary with frequency unless the transformer is very carefully designed.

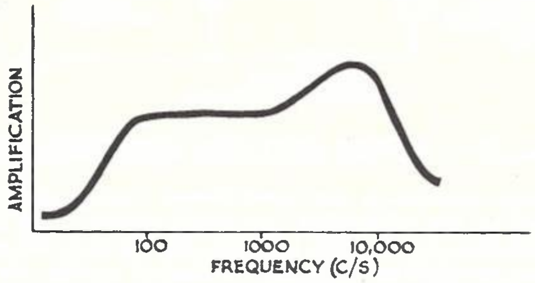

FIG. 17. AMPLIFICATION VERSUS FREQUENCY.
(TRANSFORMER COUPLING)

The loss in amplification at low frequencies results from the low reactance which the transformer primary has at low frequencies, whilst the falling off at high frequencies is a result of the leakage reactance and distributed turns capacity of the secondary winding. As these come into series resonance, the amplification may rise to a peak and then fall rapidly. The leakage reactance of a transformer takes into account those primary and secondary turns across which an e.m.f. of self-induction is developed but which are not responsible for an e.m.f. of mutual induction. For example, a transformer may have 100 primary turns, and across these 100 turns an e.m.f. of self-induction is produced. Only 90% of the primary flux cuts the secondary, so that the primary is the equivalent of a 100 turn primary having 90 turns, the whole of the flux from which cuts the secondary, in series with 10 turns, none of the flux from which cuts the secondary. The primary flux, however, links the whole 100 primary turns, so that these 10 turns have an inductive reactance which is called the Primary Leakage Reactance. In exactly the same way the secondary winding has a leakage reactance which, in this case, will be 10% of the total secondary turns, in series with those secondary turns which are effective in inducing a voltage across the primary.

These leakage reactances are shown in Fig. 18.

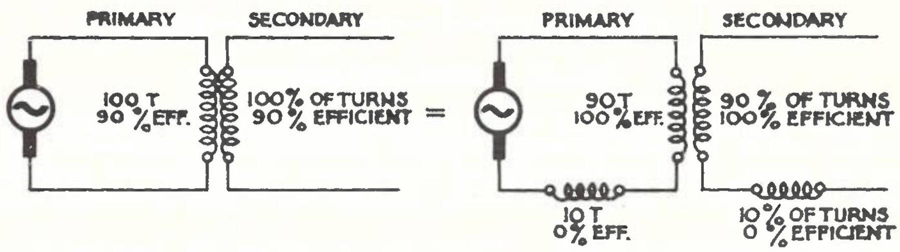

FIG. 18. TRANSFORMER LEAKAGE REACTANCES.

The distributed turns capacity takes into account the fact that the turns on a coil act as the plates of small condensers, with the insulation separating them acting as the dielectric. Thus, across the whole secondary winding, both effective and ineffective turns, is shunted a capacity which is the equivalent of this distributed capacity between the secondary turns. Across the effective turns is induced a voltage from

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the primary, so that these turns can be regarded as a generator, the whole producing an equivalent secondary circuit similar to Fig. 19.

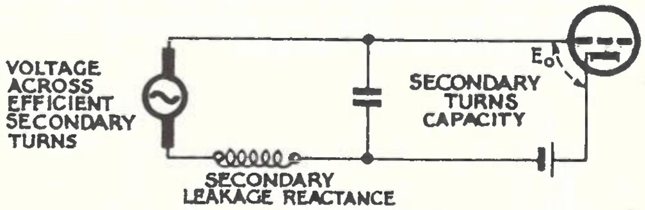

FIG. 19. VOLTAGE INPUT FROM TRANSFORMER COUPLING.

The secondary leakage reactance and distributed turns capacity form a series resonant circuit as the load on the effective secondary turns. At resonance, the voltage across either reactance in a series circuit can be many times the applied voltage and, as the input to the succeeding stage is connected across the capacity in the circuit of Fig. 19, a resonant rise in voltage amplification will result, producing the peak in the response curve of Fig. 17.

10.4 An improvement in the low frequency response can be obtained by increasing the inductance of the primary; for example, by using a large number of turns and a core of larger cross-sectional area. These, however, impair the high frequency response as the increase in the cross-sectional area of the core and the increased number of turns increase the distributed turns capacity of the secondary, so that resonance takes place at a lower frequency. The use of an Alloy core improves both the high and low frequency responses, as a high inductance can be obtained with few turns on a core of small cross-sectional area. The higher inductance improves the low frequency response, and the decreased leakage reactance and distributed turns capacity of the secondary, due to the smaller core and fewer turns, shift the resonant peak up to a much higher frequency. With correct design, this frequency is well above the highest frequency to be passed through the amplifier. The use of an alloy core generally means that the core will saturate with a low value of D.C. through the winding. This is overcome by using the arrangement shown in Fig. 20, and known as Parallel or Shunt Feed.

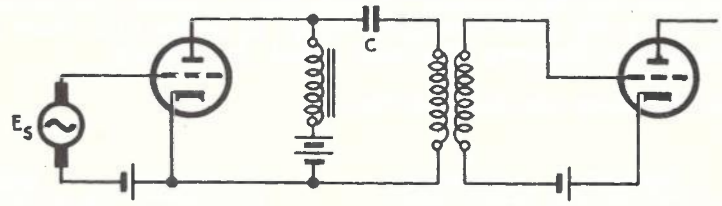

FIG. 20. SHUNT FEED ARRANGEMENT.

The anode voltage is supplied to the valve through a choke, the condenser C keeping the D.C. out of the primary winding whilst permitting the passage of the A.C. output from the stage. The difficulties of transformer design are now transferred to the choke, which must be just as carefully designed for a good frequency response.

10.5 Miscellaneous Coupling Methods. Whilst the resistance and transformer coupled circuits represent the most widely used coupling methods, others which are variations of the resistance method are sometimes used. For instance, in impedance coupling, a coupling choke L_c replaces the coupling resistance R_c of Fig. 14. Another circuit employs a coupling resistance R_c with a grid choke L_g replacing the grid resistor R_g of Fig. 14. Again, a choke may replace both coupling and grid resistors. As with transformer coupling, these chokes must be carefully designed for a good frequency response to be obtained.

11. GRID AND ANODE RETURN LEADS.

11.1 The use of alternating current or unsmoothed direct current for heating filament type cathodes always introduces a certain amount of hum into the anode circuit of such valves, even when the thermal capacity of such cathodes is high enough to prevent them from cooling appreciably between alternations. This hum is introduced because the return point becomes alternately more and less positive than the average of the filament as the heating voltage goes through its cycle, this being equivalent to applying voltages having the frequency of the heater voltage to both grid and anode. Fig. 21 shows the situation, from which it will be seen that the ends of the cathode become alternately positive and negative, so superimposing a voltage of this frequency on the anode and grid voltages.

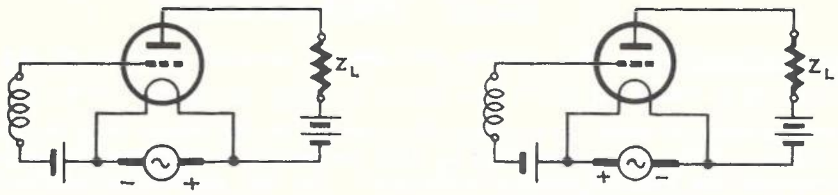

FIG. 21. FILAMENT TYPE CATHODE HEATED BY A.C.

11.2 To reduce this superimposed voltage, the grid and anode return leads are brought back to a point which, in effect, is the centre point of the cathode. By this means, the superimposed voltage is reduced by one-half, as the voltage at a point at the centre of the cathode will be only half that along its whole length. As the centre point of the cathode is inaccessible, the grid and anode return leads are usually brought to the centre point of a resistance connected across the cathode as in Fig. 22.

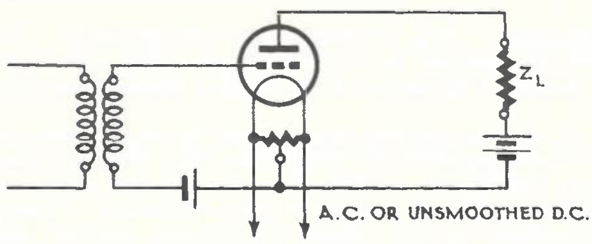

FIG. 22. METHOD OF OBTAINING CENTRE POINT OF CATHODE.

12. CLASS A POWER AMPLIFIERS.

12.1 As mentioned previously, all amplifiers used in carrier systems and repeaters are operated on a Class A basis. This is for considerations of fidelity, and also because the small amounts of power handled do not warrant the use of Class B or C operation. The valves used in the power stage generally have a low r_a, so that little of their developed power will be dissipated in that r_a, which means that such valves will have a low μ. However, this does not matter, as the preceding voltage stages have produced the necessary voltage amplification.

12.2 Maximum power transfer will take place when the load on the valve is equal in impedance to the r_a of the valve. However, as with the voltage amplifier, the power transfer must take place with little distortion. For this reason, some efficiency of power transfer is sacrificed. In practice, the load impedance on a triode is greater than its r_a, whilst that on a pentode is much less than its r_a. As examples, a 2A3 power triode has an r_a of 800 ohms and the recommended load impedance is 2,500 ohms, whilst a 2A5 power pentode has an r_a of 80,000 ohms and a recommended load impedance of 7,000 ohms. As the actual device to which the power is being supplied may have some other impedance value, this impedance value is transformed to the correct load impedance for the valve concerned by a matching transformer with a suitable turns ratio.

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12.3 The different power stage circuits will now be examined -

Single Power Valve. This circuit, Fig. 23, requires no explanation other than that the transformer transforms the load impedance Z_L to an impedance Z_p, equal to that recommended for the valve used.

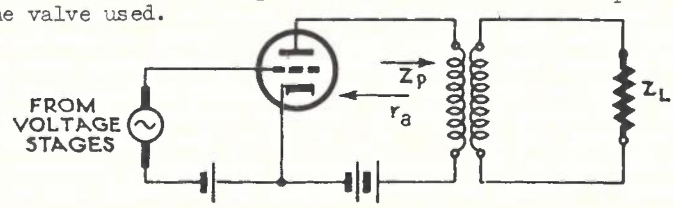


FIG. 23. SINGLE POWER VALVE.

Valves in Parallel. When valves are used in parallel, the r_a of the whole stage will be equal to $\frac{1}{n}$ times the r_a of one valve, n being the number of valves used in parallel. If two valves are used, as in Fig. 24, the r_a of the stage will be one-half the r_a of one valve. This will mean that Z_p will be half the recommended value for one valve to preserve the same impedance relations. As the impedances are halved, the power output from the stage is doubled.

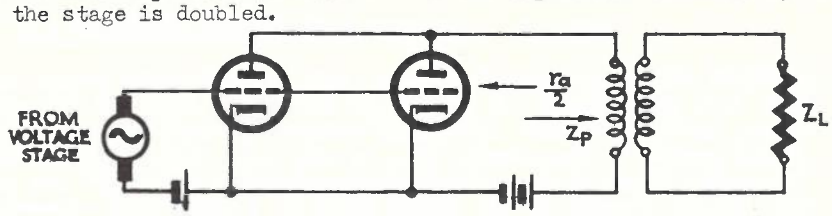


FIG. 24. VALVES IN PARALLEL.

Valves in Push-Pull. Fig. 25 shows this arrangement. In this circuit, the grids of the two valves are excited by equal voltages 180° out of phase, the outputs of the two valves being combined by means of an output transformer having a centre tap.

As the grids of the two valves are excited by equal voltages 180° out of phase during a signal, the anode current of valve A in Fig. 25 will increase, whilst that of valve B will decrease during one half-cycle; during the other half-cycle the anode current of A will decrease, whilst that of B will increase. Thus, during a signal, the two halves of the primary of the output transformer produce unequal fluxes, and the resultant flux, that is, the difference, induces across the secondary a voltage of the same frequency as the signal.

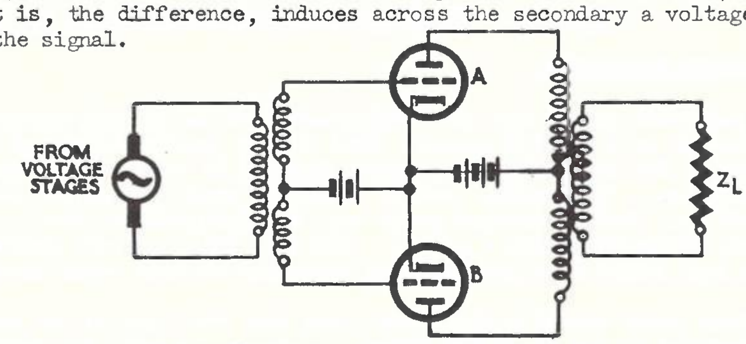


FIG. 25. VALVES IN PUSH-PULL.

The advantages of push-pull connection, assuming identical valves, are as follows -

(i) No D.C. saturation of the core of the output transformer. The D.C. in the two halves of the primary magnetise the core in opposite directions, so producing zero resultant magnetisation.

(ii) A.C. ripple voltages present in the source of anode power, for example, a rectifier unit, produce no hum voltages in the output because the hum currents flowing in the two halves of the primary balance each other.

/(iii)

(iii) There is less distortion for the same power output per valve or more power output per valve for the same distortion as a result of the cancellation of all even harmonics. This can be explained as follows -

If the grids of the two valves are driven into the lower bend during portion of the input signal cycle, harmonic distortion will be produced, as explained in paragraph 4.3.

Fig. 26 shows the effects when the grids of the two valves are excited by equal voltages, 180° out of phase. As previously explained, the individual output currents may be resolved into a fundamental frequency equal to the input signal frequency plus a second harmonic introduced by the valves.

Reference to Fig. 26 shows that the fundamental frequency currents in the output of each valve are 180° out of phase, that is, as they are increasing in Valve A they are simultaneously decreasing in Valve B, and vice versa. As mentioned earlier, these currents produce a resultant flux in the core which induces a voltage of the signal frequency across the secondary of the output transformer. The second harmonic current components, however, are in phase, increasing and decreasing simultaneously in the output of each valve. These equal in-phase currents will produce no resultant flux.

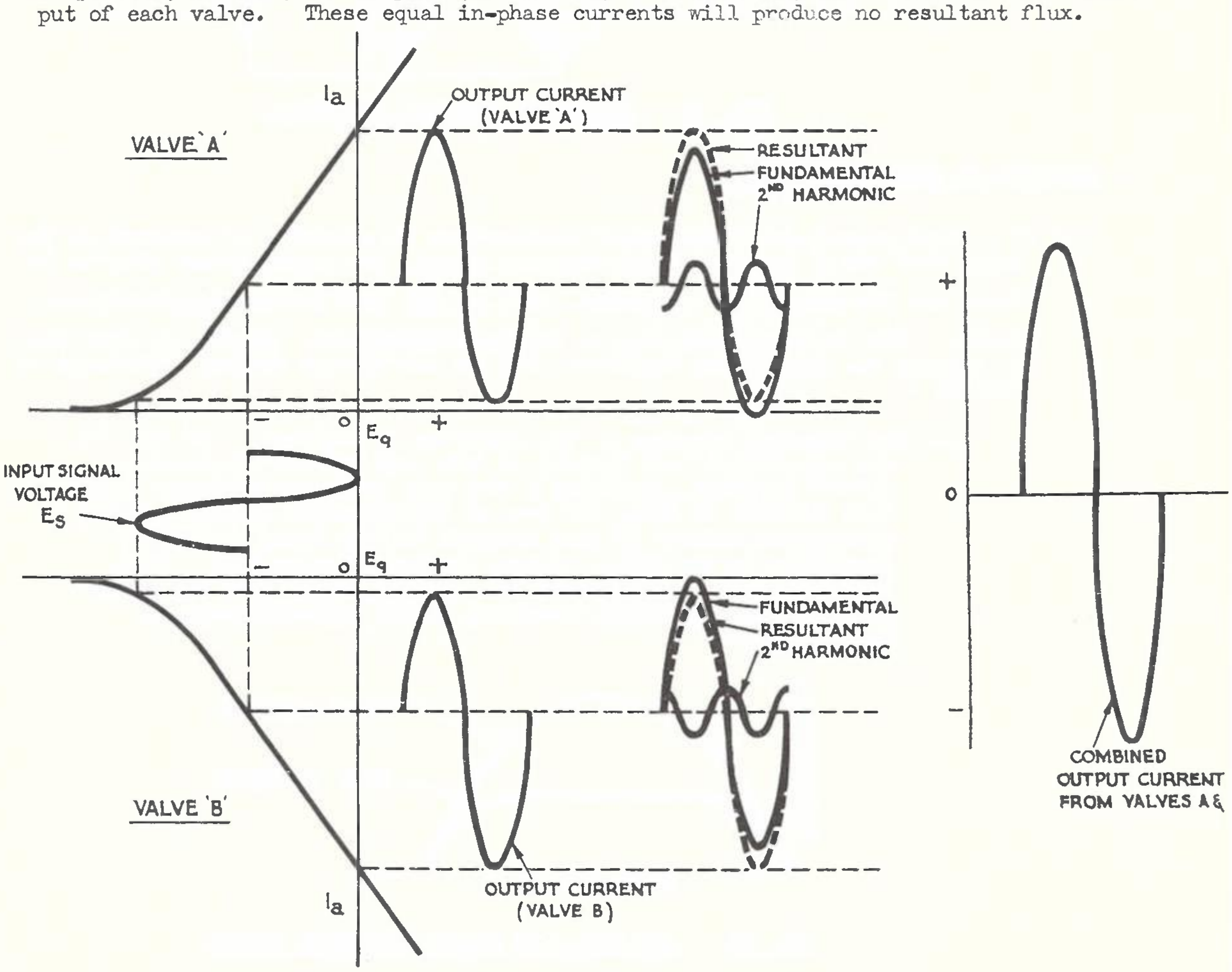


FIG. 26. DISTORTION IN PUSH-PULL AMPLIFIER.

Thus, the second harmonics, although present in the output of each valve, induce no voltage across the secondary winding of the output transformer and, therefore, do not appear in the load. This applies to all even harmonics and even order combinations. The push-pull connection, however, does not eliminate odd harmonics, as their phase relations /in in the two halves of the primary of the output transformer are not such as to produce zero resultant flux.

Valves in Parallel Push-Pull. In this arrangement, Fig. 27, a number of valves in parallel form each side of the push-pull circuit, giving a high output power with low distortion.

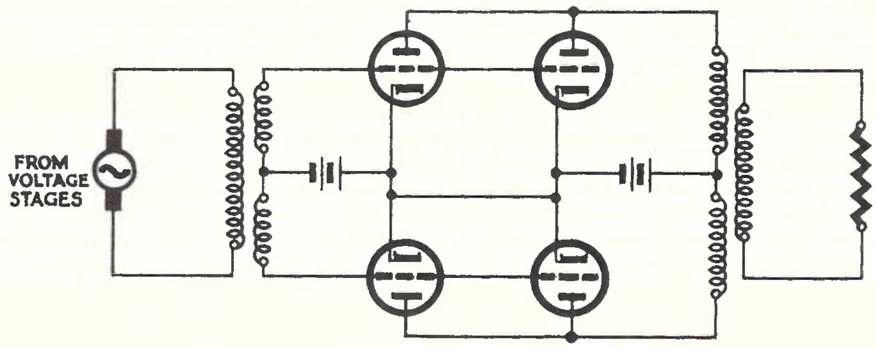


FIG. 27. VALVES IN PARALLEL PUSH-PULL.

13. NEGATIVE OR INVERSE FEEDBACK.

13.1 If a fixed percentage of the output voltage from an amplifier is returned to the input 180° out of phase with the exciting signal, the input voltage will be reduced and the gain of the amplifier, that is, $\frac{E_o}{E_s}$, will be correspondingly reduced. This is termed Negative or Inverse Feedback and, despite the fact that it reduces the gain of the amplifier, its use has certain decided advantages as follows -

(i) The frequency response of the amplifier is improved. In Fig. 28, the output falls off rapidly at both high and low frequencies without feedback, producing a very poor frequency response curve. When feedback is used, the loss of amplification at high and low frequencies results in less feedback voltage being applied in opposition to the signal voltage at those frequencies. Thus, the input voltage is lowered to a greater degree over the intermediate frequency range, resulting in the improved response curve shown.

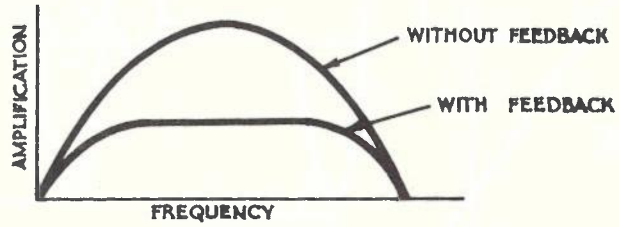


FIG. 28. AMPLIFICATION VERSUS FREQUENCY CURVES.

(ii) The gain of the amplifier is not affected by normal variations in the voltages of the various supply sources. If the amplifier tends to produce an increased or decreased output due to such variations, a check is provided by a corresponding increase or decrease in the feedback voltage.

/(iii)

(iii) A substantial reduction is effected in the harmonic distortion in the output. In the output of an amplifier, not only will the input signal frequencies be present but also their harmonics, these being produced by the non-linearity of the amplifier valve's characteristic. A percentage of these harmonic voltages is applied to the input, together with the originals, thus reducing them.

13.2 Practical Feedback Amplifiers. There are many ways of applying negative feedback to an amplifier, all of which may be reduced to one of two types of circuit, namely, Current feedback or Voltage feedback.

Current Feedback. Fig. 29 shows a current feedback circuit arranged in a resistance coupled stage. It will be remembered that the alternating voltage output from an amplifier valve is 180° out of phase with the signal voltage, that is, the alternating voltage acting in the anode circuit of an amplifier valve is 180° out of phase with the signal voltage. This means that the alternating voltage developed across a cathode bias resistor, with its by-pass condenser removed, will be 180° out of phase with the signal voltage, because the cathode resistor forms part of the anode circuit as well as part of the input circuit.

Fig. 29 uses portion of this cathode bias resistor across which to develop the feedback voltage. Portion, rather than all, of the bias resistor is used as, in general, a different value of resistor will be required for bias purposes from that required to provide the desired amount of feedback. This circuit, whilst reducing harmonic distortion in the output and so on, will not improve the frequency response. This is because, for typical values of R_c, C_c and R_g, the impedance on the stage to which feedback is applied is almost independent of frequency, so that the current output from E_s is likewise almost independent of frequency. As this current flows through R_{fb}, across which it develops the feedback voltage, this feedback voltage will likewise be independent of frequency. Thus, the feedback voltage will not rise and fall with frequency, as may the voltage across R_g which is the output voltage, so that no improvement in the frequency response results from the use of this circuit.

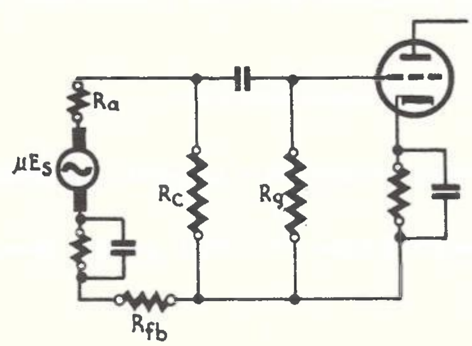


FIG. 29. CURRENT FEEDBACK CIRCUIT.

Voltage Feedback. This arrangement, shown in Fig. 30, is a real voltage feedback arrangement, in that $R + R_{fb}$ in series are across R_g, the resistor across which the output voltage is developed. Thus, the voltage across R_{fb} is a proportion of the output voltage, the proportion depending on the ratio of R_{fb} to R. This arrangement, therefore, will correct the frequency response as well as have the other advantages of negative feedback.

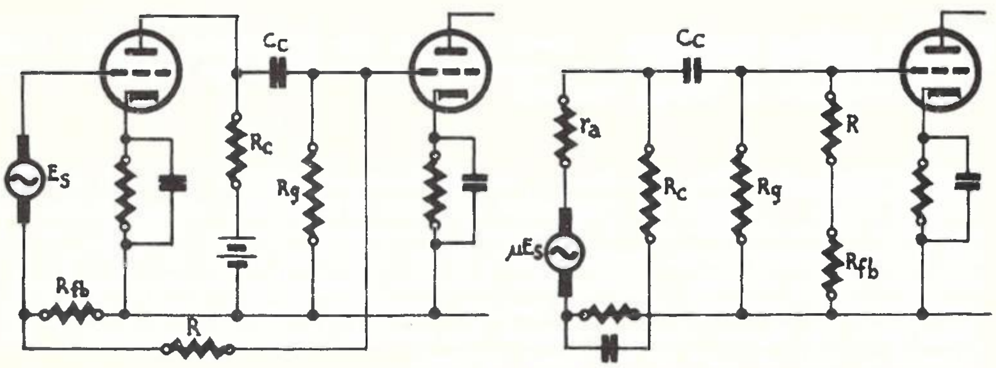


FIG. 30. VOLTAGE FEEDBACK CIRCUIT.

14. TEST QUESTIONS.

1. Why is it necessary to provide separate voltage and power stages in an amplifier?

2. Explain how cathode bias is applied to an amplifier.

3. Explain, using sketches, how the non-linear portions of a valve characteristic introduce distortion.

4. What are the operating conditions for Class A amplifiers and why are these operating conditions necessary?

5. List the advantages of a push-pull amplifier.

6. With the aid of diagrams, explain how a push-pull amplifier realises these advantages.

7. What is meant by "Negative Feedback?" What are the advantages of its use?

8. What is meant by -

(i) A resistance coupled amplifier, and
(ii) A transformer coupled amplifier?

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

LONG LINE EQUIPMENT I.

OSCILLATORS.

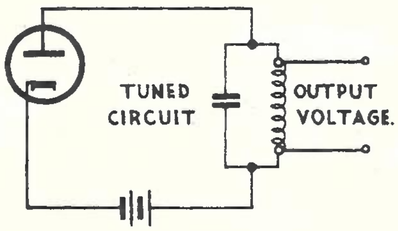
CONTENTS:

1. INTRODUCTION.

1.1 As has been mentioned previously, thermionic valves are used to supply the carrier frequencies required for carrier systems. To do this, thermionic valves have associated with them resonant circuits which may be either electrical, mechanical or electro-mechanical. Examples of the three types are inductance-capacitance combinations, tuning forks and piezo-electric crystals, respectively. Up to the present, the oscillators in Carrier Systems in use in Australia employ either inductance-capacitance combinations or tuning forks. Before discussing the various circuits which will be encountered in practice, the general conditions necessary for the production of alternating currents and voltages of constant amplitude in such circuits will be dealt with first.

2. PRINCIPLE OF OSCILLATOR.

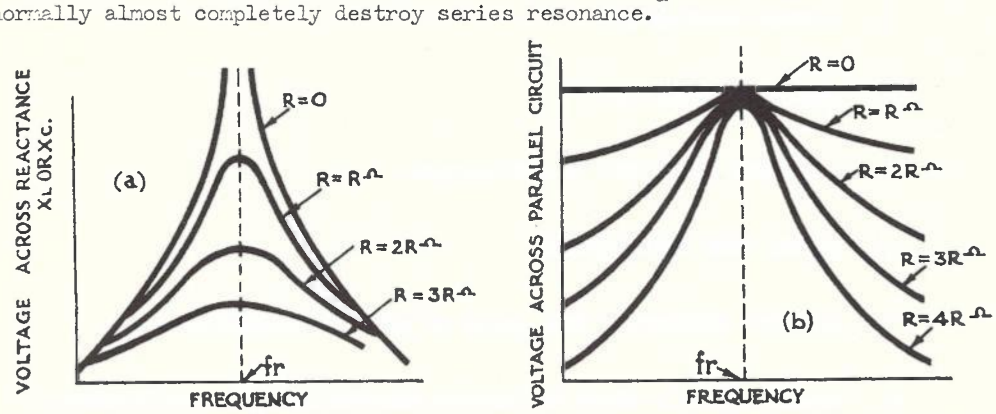
2.1 Up to the present the anode current flow through a valve has been regarded as perfectly regular - somewhat like a smooth, fluid flow. This is not a true conception, as the stream of electrons flowing from cathode to anode is made up of a series of particles rather than a continuous fluid flow. This means that the electrons strike the anode in a manner somewhat resembling hailstones striking a roof, this action producing minute irregularities in the anode current. This effect is known as "shot effect." The minute irregularities in the anode current contain all frequencies from zero to infinity so that, if it is desired to use one particular frequency, a circuit tuned to that frequency could be placed in the anode circuit and would develop maximum voltage at the resonant frequency. In other words, the tuned circuit would "pick off" the frequency required from all others present. Fig. 1 shows the idea.



TUNED CIRCUIT IN ANODE CIRCUIT.

FIG. 1.

2.2 A parallel circuit is used in preference to a series circuit because the effect of resistance in a series circuit is to reduce the selectivity, that is, the ability of a series circuit to pass its resonant frequency whilst rejecting others close to that frequency, by reducing the current at resonance. This, in turn, reduces the voltage across X_L or X_C, as shown in Fig. 2a. The majority of the resistance present in an oscillator circuit would be the r_a of the valve, which would normally almost completely destroy series resonance.



(a) Series Circuit. (b) Parallel Circuit.

SELECTIVITY CURVES.

Note. R = Resistance in series with each circuit.

FIG. 2.

2.3 On the other hand, the r_a of the valve in series with a parallel circuit would aid the selective action of that circuit. This would be because a parallel circuit draws minimum current at resonance, the current drawn from the supply increasing as the frequency departs from resonance. At resonance, therefore, minimum voltage drop takes place across the resistance in series with a parallel circuit, so that maximum voltage is applied across the latter. As the frequency departs from resonance the current increases, increasing the voltage drop across the series resistance and therefore decreasing the voltage applied across the parallel circuit, as shown in Fig. 2b.

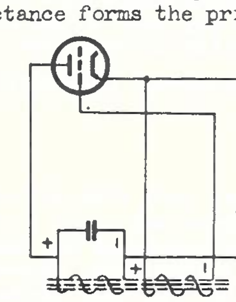
/Of

Of course, maximum current circulates in the parallel circuit at resonance, although minimum current is being drawn from the supply, whilst the circulating current decreases as the frequency departs from resonance. At resonance, the circuit is colloquially said to be "oscillating", and the tuned circuit is often spoken of as an "oscillatory" circuit.

2.4 Thus, an alternating current of the required frequency would flow in the tuned circuit, the amplitude of this current being many times the amplitude of that of the same frequency in the anode current irregularities. As an alternating current flows in the tuned circuit, an alternating voltage will be developed across it which could be amplified for useful purposes. The valve that produces the irregularities from which the required frequency is selected is almost exclusively used as the amplifier also. For the valve to do this, correct phase relations must be maintained in that the amplified oscillations must appear in the anode circuit in phase with the original oscillations. This means that the phase shift from the oscillatory circuit through the valve and back to the oscillatory circuit must be 0°, 360° or some integral multiple of 360°. As the valve is responsible for a phase shift of 180° between grid and anode, an additional 180° is required in the amplifying process. This is provided in a number of ways, the exact method depending on the type of oscillator circuit employed. Some of these circuits will now be dealt with.

3. TUNED ANODE OSCILLATOR.

3.1 The tuned anode circuit is shown in Fig. 3. As described above, a tuned anode circuit selects the frequency required from the anode circuit irregularities. The tuning inductance forms the primary winding of a transformer, the secondary winding of which is connected across grid and cathode. If both windings of the transformer are wound in the same direction their induced voltages during one half cycle of oscillatory current will be in the direction shown. When the secondary voltage is applied to the grid and cathode in the manner shown a phase shift of 180 is obtained, and this, together with the 180 phase shift through the valve enable the oscillations to appear in the anode circuit in phase with the originals. These will steadily build up in amplitude as successive cycles are developed and amplified, the limit being reached when the grid swings extend from anode current cut-off to saturation.



TUNED ANODE CIRCUIT.

FIG. 3.

3.2 From this circuit the phase relations necessary for the operation of most oscillators can be deduced. From Fig. 3 it will be seen that the oscillatory voltage across anode and cathode is always 180° out of phase with that across grid and cathode, these voltages being indicated for some instant. Any circuit arrangement which brings about this condition will oscillate.

3.3 Fig. 4 shows a practical tuned anode oscillator. Oscillators of this type supply the carrier frequencies for the S.T.C. Type C three-channel carrier telephone system.

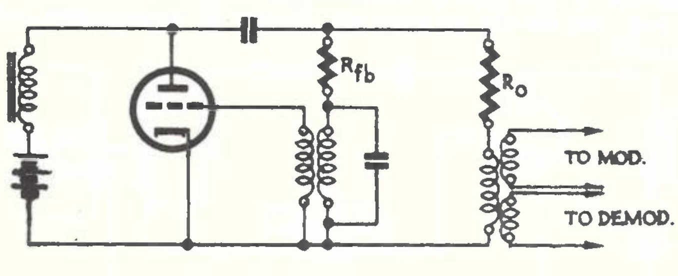
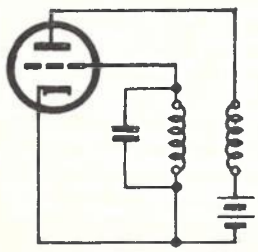


FIG. 4. PRACTICAL TUNED ANODE OSCILLATOR.

Shunt feed is used to prevent direct current from passing through the tuning inductances and transformer windings. The feedback and output resistors, R_{fb} and R_o, respectively, are proportioned so that, whilst enough of the output power is passed to the modulator and demodulator for their successful operation, the amount of the total output power applied to the oscillatory circuit is sufficient to produce the output power required for the above two purposes.

4. TUNED GRID OSCILLATOR.

4.1 In the tuned grid circuit, Fig. 5, the tuned circuit is in the grid circuit. When the bias is zero, as it is in this circuit and also in Figs. 3 and 4, grid current flows and the minute irregularities produced by the shot effect at the grid give rise to oscillations in the grid tuned circuit. As in Fig. 3, the phase relations between the anode and grid circuit oscillatory voltages are such as to maintain the oscillations.



TUNED GRID CIRCUIT.

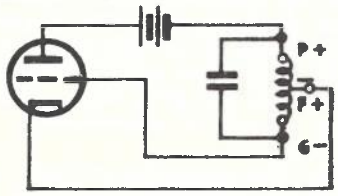
FIG. 5.

4.2 Another way of looking at the operation of the Tuned Grid Oscillator is to think of the anode current irregularities as inducing voltages of all frequencies across the grid tuning inductance. As the grid circuit is tuned to the frequency required from the oscillator, currents of this frequency will flow in the tuned grid circuit and voltages of this frequency will be developed across the grid tuned circuit for application to the grid for amplification. Here, again, the phase shift from the anode circuit, through the transformer and amplifying process and back to the anode circuit is 360°, so that the general condition discussed earlier is fulfilled.

4.3 This type of oscillator is used to develop the base frequency of 4 kc/s in 17-channel carrier telephone systems.

5. HARTLEY OSCILLATOR.

5.1 The Hartley oscillator is widely used, as only a single coil with an intermediate tap is required. The directions of the grid to cathode and anode to cathode oscillatory voltages necessary for oscillation were shown in Fig. 3, and these directions agree with those in Fig. 6. At some instant the oscillatory voltage acting across the whole of the tuned circuit is in the direction indicated, that is, point P is positive with respect to Point G. Point F, which is intermediate between points P and G, will be negative, therefore, with respect to point P but positive with respect to point G. Thus, the voltage across points P and F, that is, across anode and cathode, is opposite in direction to, or 180° out of phase with, the voltage across points G and F, that is, across grid and cathode, which agrees with Fig. 3.



HARTLEY OSCILLATOR.

FIG. 6.

5.2 Fig. 7 shows an application of the Hartley oscillator, the circuit being the oscillator for some issues of the Type F single-channel systems. Here, again, shunt feed is used to keep the direct current out of tuning inductances, etc. Grid leak bias is employed.

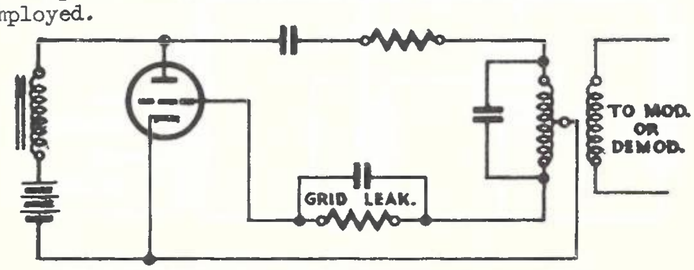


FIG. 7. PRACTICAL HARTLEY OSCILLATOR.

6. COLPITTS OSCILLATOR.

6.1 The Colpitts circuit is similar to the Hartley circuit except that, instead of tapping an intermediate point on the inductance, the tuning capacity is divided into two series capacities with the intermediate point connected to the cathode, as in Fig. 8.

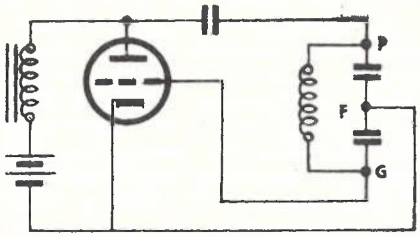


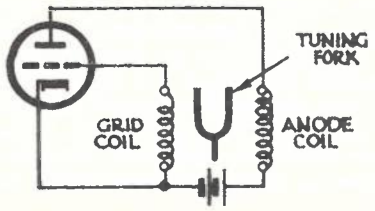
FIG. 8. COLPITTS CIRCUIT.

7. FREQUENCY STABILITY OF OSCILLATORS.

7.1 Where a high degree of frequency stability is required from an oscillator, mechanical or electro-mechanical oscillating systems are frequently used as their frequency stability is greater than that of electrical oscillatory circuits. For example, the natural frequency at which a tuning fork or piezo-electric crystal oscillates will exhibit practically no change over very wide ranges of temperature, whilst the natural or resonant frequency of an inductance-capacitance combination changes with temperature because temperature changes cause slight changes in the inductance and capacity of the combination. In 17-channel carrier telephone systems the tuned grid circuit of the oscillator developing the base frequency of 4 kc/s is placed in an oven, which is maintained at a constant temperature by thermostat control, so keeping the inductance and capacity and, therefore, the resonant frequency, of the combination constant. In Type J systems, and also Type K 12-channel carrier telephone systems, a tuning fork oscillator is used to develop a base frequency of 4 kc/s. Most of the carrier frequencies used in such systems are multiples of 4 kc/s, and these are provided by distorting the 4 kc/s output from the base frequency oscillator. The resultant distortion produces the required harmonics, or multiples, of the base frequency. This principle is also used in 17-channel systems. As mentioned above, the reason for using these mechanical or electro-mechanical oscillating systems is that their frequency stability, even without such arrangements as temperature-controlled ovens, is much greater than that of inductance-capacitance combinations.

8. TUNING FORK OSCILLATORS.

8.1 In the tuned anode oscillator circuit of Fig. 3 energy is transferred from the anode circuit to the grid circuit at one frequency alone, the selective action being accomplished by tuning the anode circuit to the frequency at which the energy transfer is to take place. In the tuning fork oscillator used in Type J and K systems, the tuning fork provides a resonant link between the anode and grid circuits via which the energy transfer takes place. As the fork prongs vibrate, they disturb the magnetic field produced by the anode current flowing through the anode coil at the natural frequency of vibration of the prongs, this disturbance inducing a voltage of the corresponding frequency across the grid coil, the turns of which are linked by the flux produced by the anode coil. This voltage is, of course, amplified by the valve. Fig. 9 shows the idea.



TUNING FORK OSCILLATOR.

FIG. 9.

/8.2

8.2 Because of friction losses the prongs of a tuning fork cease to vibrate a short time after having been set into vibration, usually by a slight blow. To keep the tuning fork vibrating and so sustain the energy transfer between anode and grid circuits, the oscillatory current produced by the amplified voltage across the grid coil flowing through the anode coil must produce a flux in a direction such as to aid the mechanical movement of the fork prongs. For example, when the prongs are moving outwards they induce a voltage across the grid coil in a certain direction. This voltage is amplified, and the effect of the resultant flux change in the anode coil, produced by the amplified anode current flowing through it, must be to aid the outward movement of the prongs. In other words, the anode coil must supply energy to the fork to maintain it in oscillation, and this energy must be supplied in phase with the original energy represented by the vibrations of the prongs. This brings one back to the original condition for generating sustained oscillations, that is, that the phase shift from the source of oscillation, through the amplifying process and back to the source again must be 0°, 360° or some multiple thereof. Just as the correct connection of the grid-cathode and grid-anode coils achieve this in the preceding circuits, so the same precautions will maintain the tuning fork in oscillation in this circuit. In some Type J and K systems, the selective action of the tuning fork is aided by tuning the anode and grid circuits by shunt condensers not shown in Fig. 9.

9. MAGNETIC GENERATION OF A GROUP OF CARRIER FREQUENCIES.

9.1 As mentioned previously, the carrier frequencies for Type J and Type K 12-channel systems are provided by distorting the output from a 4 kc/s oscillator, this distortion producing harmonics of 4 kc/s, many of which are used as carriers. In the systems mentioned, the harmonic producing device, together with its oscillator, is referred to as the "carrier frequency generator."

9.2 The essential circuit elements employed in the carrier frequency generator are shown in Fig. 10.

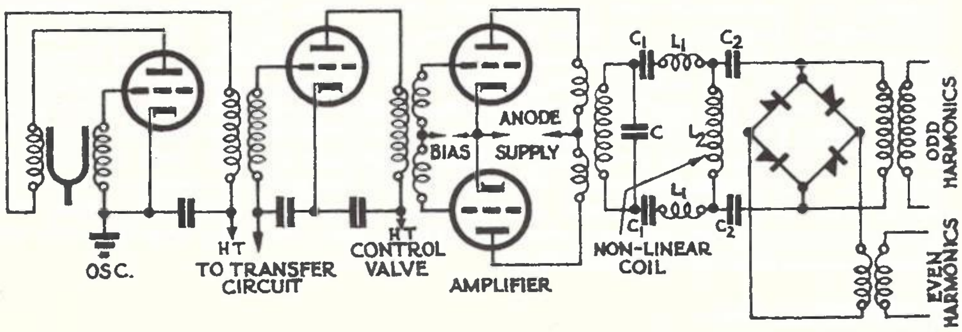


FIG. 10. MAIN ELEMENTS OF CARRIER FREQUENCY GENERATOR.

The generated frequency of the oscillator valve is controlled by a 4 kc/s tuning fork. This tuning fork is made of an alloy having a low temperature coefficient, and is operated in a vacuum in a sealed container. The stability of this tuning fork is such as to hold the frequency of oscillation accurate to within ± 1 cycle in one million. The oscillator output is amplified in two stages to a value of about 4 watts by the control valve and two power valves operating in a push-pull arrangement. The control valve also acts in conjunction with an auxiliary transfer circuit (not shown in Fig. 10) to put automatically into service an emergency oscillator in case of the failure of the regular circuit.

9.3 The secondary of the output transformer and condenser C are designed to be resonant at 4 kc/s, and so C practically shorts out any second harmonics developed in the amplifier valves. The series condensers C_1 and inductances L_1 are resonant at 4 kc/s, and thus favour transmission of a pure sine wave of 4 kc/s current to the bridged coil L_2. This latter coil, in conjunction with the condensers C_2, produces odd harmonics of the applied 4 kc/s frequency.

/9.4

6. COLPITTS OSCILLATOR.

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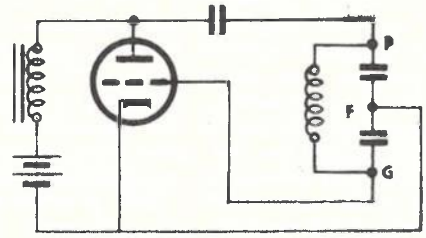

FIG. 8. COLPITTS CIRCUIT.

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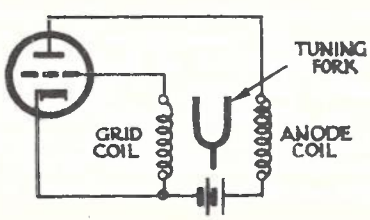


TUNING FORK OSCILLATOR.

FIG. 9.

/8.2

8.2 Because of friction losses the prongs of a tuning fork cease to vibrate a short time after having been set into vibration, usually by a slight blow. To keep the tuning fork vibrating and so sustain the energy transfer between anode and grid circuits, the oscillatory current produced by the amplified voltage across the grid coil flowing through the anode coil must produce a flux in a direction such as to aid the mechanical movement of the fork prongs. For example, when the prongs are moving outwards they induce a voltage across the grid coil in a certain direction. This voltage is amplified, and the effect of the resultant flux change in the anode coil, produced by the amplified anode current flowing through it, must be to aid the outward movement of the prongs. In other words, the anode coil must supply energy to the fork to maintain it in oscillation, and this energy must be supplied in phase with the original energy represented by the vibrations of the prongs. This brings one back to the original condition for generating sustained oscillations, that is, that the phase shift from the source of oscillation, through the amplifying process and back to the source again must be 0°, 360° or some multiple thereof. Just as the correct connection of the grid-cathode and grid-anode coils achieve this in the preceding circuits, so the same precautions will maintain the tuning fork in oscillation in this circuit. In some Type J and K systems, the selective action of the tuning fork is aided by tuning the anode and grid circuits by shunt condensers not shown in Fig. 9.

9. MAGNETIC GENERATION OF A GROUP OF CARRIER FREQUENCIES.

9.1 As mentioned previously, the carrier frequencies for Type J and Type K 12-channel systems are provided by distorting the output from a 4 kc/s oscillator, this distortion producing harmonics of 4 kc/s, many of which are used as carriers. In the systems mentioned, the harmonic producing device, together with its oscillator, is referred to as the "carrier frequency generator."

9.2 The essential circuit elements employed in the carrier frequency generator are shown in Fig. 10.

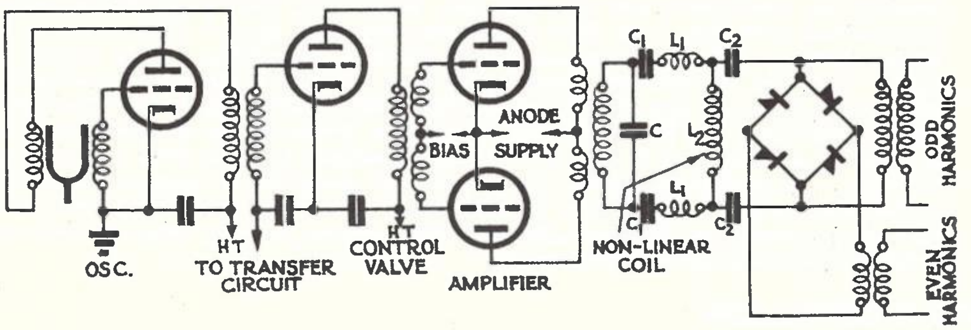


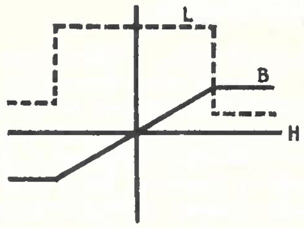
FIG. 10. MAIN ELEMENTS OF CARRIER FREQUENCY GENERATOR.

The generated frequency of the oscillator valve is controlled by a 4 kc/s tuning fork. This tuning fork is made of an alloy having a low temperature coefficient, and is operated in a vacuum in a sealed container. The stability of this tuning fork is such as to hold the frequency of oscillation accurate to within ± 1 cycle in one million. The oscillator output is amplified in two stages to a value of about 4 watts by the control valve and two power valves operating in a push-pull arrangement. The control valve also acts in conjunction with an auxiliary transfer circuit (not shown in Fig. 10) to put automatically into service an emergency oscillator in case of the failure of the regular circuit.

9.3 The secondary of the output transformer and condenser C are designed to be resonant at 4 kc/s, and so C practically shorts out any second harmonics developed in the amplifier valves. The series condensers C_1 and inductances L_1 are resonant at 4 kc/s, and thus favour transmission of a pure sine wave of 4 kc/s current to the bridged coil L_2. This latter coil, in conjunction with the condensers C_2, produces odd harmonics of the applied 4 kc/s frequency. /9.4

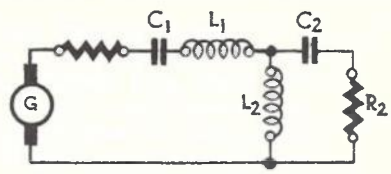
9.4 The action of the coil as a harmonic producer depends upon the fact that its core, made of permalloy, becomes magnetically saturated at relatively low current values. The coil is physically quite small, the core being a coil of permalloy ribbon. A B/H curve for the core is shown in Fig. 11, from which it will be noted that it saturates at very low field intensities and, therefore, with a comparatively low current in its winding. Since the inductance is proportional to the permeability, B/H, this means that the inductance has a high value at low current values but, outside of a very narrow range of current values, the inductance becomes nearly zero as the curve becomes approximately horizontal. This is also shown in the curves of Fig. 11.

9.5 With these facts in mind, the behaviour of the coil and its associated condensers can be analysed by using the simplified circuit of Fig. 12 where all of the circuit to the right of C_2 is represented by the resistance R_2.



CHARACTERISTICS OF NON-LINEAR COIL.

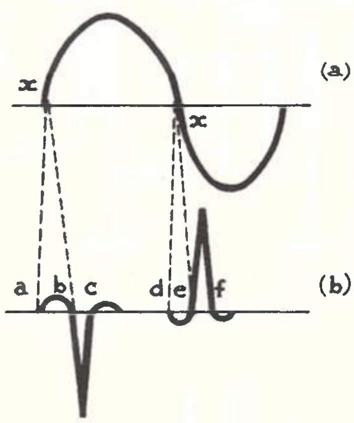
FIG. 11.



SIMPLIFIED CIRCUIT OF NON-LINEAR COIL.

FIG. 12.

One cycle of current set up by the 4 kc/s applied voltage is shown in Fig. 13a. As this current increases from zero the inductance of the coil bridged across the line will at first be high and, as a result, current will flow into the condenser and the load R_2. This current is shown by the small section ab of the curve of Fig. 13b. When the applied current reaches the critical value, x, however, the core of the coil becomes saturated and the inductance of the coil immediately decreases to zero. As the coil has a low resistance it then becomes effectively a short-circuit, and no additional current flows into R_2. On the contrary, the charged condenser C_2 discharges through the coil, causing the sharply peaked negative current surge shown in the section bc. For the remaining part of the positive pulse of the applied voltage the coil continues to act as a short-circuit, the condenser remains discharged and no current flows in the resistance R_2. When the applied current reverses in direction, however, the coil again presents a high inductance to the low values of negative current, and a small negative current, de, flows into the condenser C_2 and resistance R_2. Again, as soon as the coil becomes saturated the condenser discharges to cause the sharp positive peak of current ef. An analysis of the curious current wave of Fig. 13b would show that included are all of the odd harmonics of the applied 4 kc/s current and, further, that up to very high frequencies the amplitude of all of these harmonics is approximately the same. This is because the sharp peaking of the wave form produces harmonics at the expense of the amplitude of the fundamental, that is, the amplitude of the 4 kc/s fundamental decreases with the "peaking," so increasing the amplitude of the harmonics, the arrangement producing harmonics which do not vary greatly in amplitude.



CURRENT WAVE FORMS.

FIG. 13.

/9.6

9.6 As indicated by Fig. 10, these odd harmonics are separated for use in the various carrier channels by means of filters. Even harmonics are obtained by means of a full-wave rectifier bridged across the output of the non-linear coil. This rectifier rectifies about half of the output from the non-linear coil and, in the ideal case, the rectifier output appears as a full-wave rectified 4 kc/s. This can be analysed into a fundamental frequency of 8 kc/s plus all harmonics, both even and odd. Since even and odd harmonics of 8 kc/s are even harmonics of 4 kc/s, this circuit produces the even harmonics of the 4 kc/s fundamental, which are selected by a second group of filters.

9.7 The complete separation of odd and even harmonics by the method described tends to simplify the design of the selecting filters since it automatically separates any two frequencies in either of the output circuits by 8 kc/s.

10. GENERATION OF A GROUP OF CARRIER FREQUENCIES BY A MULTIVIBRATOR.

10.1 In 17-channel systems used on cable circuits in Australia, a base frequency of 4 kc/s is generated by a tuned grid oscillator. To preserve frequency stability, the tuned grid circuit is maintained at a constant temperature in a temperature-controlled oven. As in the Types J and K systems, all carrier frequencies are multiples, or harmonics, of 4 kc/s. In these 17-channel systems a sharply peaked wave form, rich in harmonics, is produced by passing the 4 kc/s output from the base frequency oscillator through a unit termed a "Multivibrator." The multivibrator is actually a number of stages in tandem coupled by resistance-condenser combinations which, together with the suitable adjustment of their electrode and input voltages, produce the peaked wave form required. Fig. 14 shows the elements of the circuit, together with the wave forms produced by the various stages.

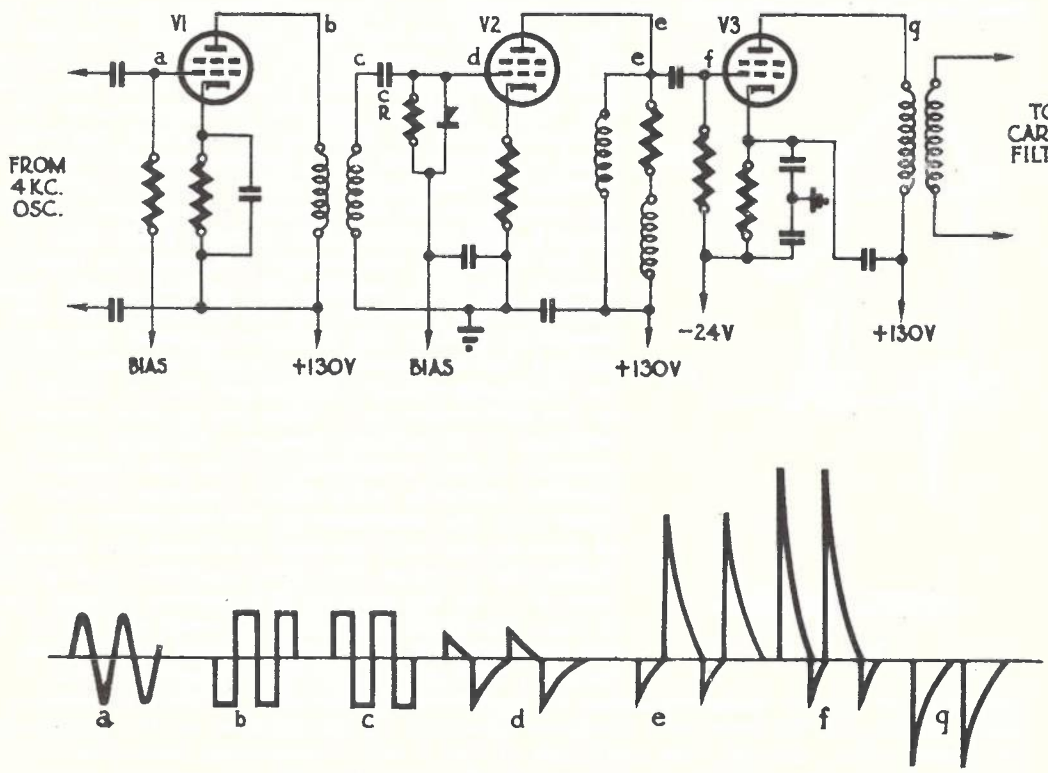


FIG. 14. MAIN ELEMENTS OF MULTIVIBRATOR CIRCUIT.

10.2 The circuit functions as follows -

The output from the 4 kc/s oscillator (not shown in Fig. 14 but similar to Fig. 5) is resistance coupled to V1, which is termed a "chopper" valve. The sinusoidal 4 kc/s voltage input to this valve, shown in Fig. 14a, drives the grid of this valve well beyond saturation and anode current cut-off, the electrode and input voltages being so arranged that saturation and cut-off are reached before the output voltage from the oscillator has risen very far in either direction. By this means, that portion of the sinusoidal output from the oscillator over which the rise and fall of voltage is least rapid is "chopped off," so that the output from the "chopper" valve is a square-topped wave (Fig. 14b) whose sides rise very steeply.

10.3 This "chopper" stage is transformer-coupled to what is termed a "clipper" stage, V2. A square-topped wave will be produced across the secondary of the coupling transformer (Fig. 14c), and the purpose of this "clipper" stage is to produce sharply peaked pulses from the square wave applied to it. Across the secondary of the coupling transformer is connected a resistance condenser combination, R-C, the resistance being shunted by a metal type rectifier. The purpose of the R-C combination is to produce a sharply peaked voltage input to V2, which is taken from across R, and the purpose of the rectifier is to limit the amplitude of the positive peaks.

The arrangement functions as follows -

At instant x of Fig. 15a, a voltage is suddenly induced across the secondary of the coupling transformer in the direction indicated in Fig. 15b. This voltage charges the condenser C, the charging current flowing in the direction indicated. A condenser charging current is initially high and gradually falls to zero, so that, as this current passes through the rectifier it will develop a voltage drop across the rectifier. Due to the low resistance of the rectifier, the amplitude of the voltage pulse developed across it will be small, as indicated by the positive pulse b of Fig. 15d.

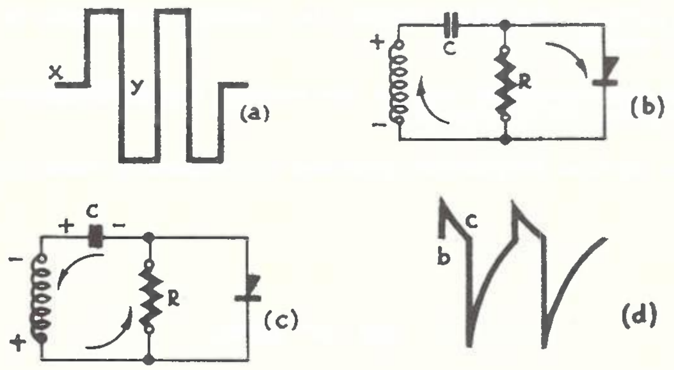


FIG. 15. ACTION OF R-C COMBINATION WITH RECTIFIER.

This voltage is maintained across the condenser until instant y, when it is removed to be replaced immediately by an equal voltage in the opposite direction. The position is now indicated in Fig. 15c, which indicates the directions of the voltages across the condenser and secondary winding of the coupling transformer.

The condenser will now discharge and recharge in the opposite direction, this discharge and charging current passing through the resistance R, as the rectifier is now blocking. This charge and discharge produces a current pulse of the same shape as the original, but the resultant voltage pulse is of higher amplitude as shown at c of Fig. 14d. This is because the reverse resistance of the rectifier is higher in value than the forward resistance, so that a higher amplitude voltage pulse will be produced across it. Thus, as each cycle of the square wave is applied to the R-C combination, the square wave half-cycles are "peaked" to appear as in Fig. 15d. As the input to the first "clipper" valve V2 is connected across R, Fig. 15d and also Fig. 14d

/represent

represent the shape of the voltage wave applied to it. This wave has a fundamental frequency of 4 kc/s and, due to the "peaking," is very rich in harmonics. V2 has about zero bias in its operating condition, so that the small amplitude positive pulses drive it towards saturation whilst the negative pulses drive it towards cut-off. This produces Fig. 14e in the output of V2, which is the same shape as Fig. 14d but inverted (because of the 180° phase shift in the valve) and amplified.

10.4 V2 is resistance coupled to another clipper stage, V3. The coupling condenser and grid resistor in the coupling circuit are proportioned to "peak" or "sharpen" the signal voltage applied to V3 in the same manner as did the R-C combination described above. V3 is biased beyond anode current cut-off so that it amplifies only the sharpest portions of the positive pulses of Fig. 14f, these being the output pulses from V2 "sharpened" by the coupling circuit. The output of V3 is shown in Fig. 14g which contains only the sharpest portions of the positive pulses of Fig. 14f inverted by the valve and amplified. Thus, V3 further sharpens the waveshape and half-wave rectifies it, the result being a fundamental frequency of 4 kc/s plus harmonics up to a very high order. The output of the multivibrator unit is passed to a number of filters which select the required carrier frequencies. Each carrier frequency is amplified by a separate carrier amplifier, these amplifiers providing enough power and a sufficiently low output impedance to supply all of the modulators and demodulators for 10 systems.

11. TEST QUESTIONS.

1. Why is a parallel resonant circuit preferred to a series resonant circuit in a tuned anode oscillator?

2. Describe briefly the conditions necessary for sustaining oscillations in a valve oscillator.

3. How does a Hartley oscillator fulfil the conditions outlined in the answer to Question 2 above?

4. Explain how the non-linear coil of the carrier frequency generator for 12-channel systems fulfils its functions.

5. How are the carrier frequencies produced in 17-channel cable carrier systems?

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

MODULATION AND DEMODULATION.

CONTENTS:

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1. INTRODUCTION.

1.1 In Paper No. 2 of this book it was shown that modulation is the process of translating the voice frequency band into a band of higher frequencies by means of suitable circuit elements, and that the type of modulation employed in carrier systems is known as Amplitude Modulation. Amplitude Modulation is so called because the different frequencies produced by the process, when added together, give the impression that the amplitude of the carrier current or voltage which has been used in the modulation process has been varied or "modulated" by the modulating voice frequency. Nevertheless, Amplitude Modulation is primarily a frequency translation process, and it is from this point of view that the subject will be attacked.

2. PRINCIPLE USED IN MODULATORS.

2.1 An example of a circuit element producing frequency translation has been dealt with already in Paper No. 6 of this book under the section on Amplitude Distortion. In that section it was shown that when an alternating voltage is applied to the non-linear, or curved, portion of the characteristic curve of a valve, the anode current contains not only the original frequency applied but also harmonics of that frequency which, in an amplifier, are distortion products. This is an example of frequency translation, as the applied voice frequency is translated upwards to frequencies equal to twice, three times, and so on, that applied to the amplifier.

2.2 As a modulator, the principle of applying the voice frequency band alone to a non-linear circuit is not suitable. If the voice frequency band from a telephone is applied to such a device, the frequencies present will be -

200 to 3,000 c/s - the original Voice Frequency band.
400 to 6,000 c/s - twice the original Voice Frequency band.
600 to 9,000 c/s - three times the original V.F. band, and so on.

Here, the band of voice frequencies is 2,800 c/s wide, whereas the first translation products are contained in a band 6,000 - 400 = 5,600 c/s wide. Besides this band being twice as wide as the original band, it is not possible to translate this band down to the original voice frequency band again.

2.3 The principle used is to apply the voice frequency band to a non-linear circuit element with a carrier frequency, the resultant output containing bands of frequencies which are of exactly the same width as the original voice frequency band which can be quite simply translated back to that band again.

2.4 If a carrier frequency voltage of 10 kc/s and voice frequency voltages in the band 200 to 3,000 c/s are simultaneously applied to the non-linear portion of a valve characteristic, or to any non-linear circuit element for that matter, the output current will contain the following frequencies -

f_v, this being the band 200 to 3,000 c/s in this case.

f_c, the carrier frequency of 10 kc/s.

$2f_v$, this being the band 400 to 6,000 c/s and containing second harmonics of the voice frequencies.

$2f_c$, this being 20 kc/s, the second harmonic of the carrier frequency.

$f_c + f_v$, this being the upper sideband, 10.2 to 13 kc/s in this case.

$f_c - f_v$, this being the lower sideband, 7 to 9.8 kc/s in this case.

2.5 This principle is used in the modulators of all carrier systems, so that it can be said that the process of modulation, as applied to carrier telephone systems, consists of the simultaneous application of the voice frequency band and a suitable carrier frequency to a non-linear circuit element. The action of this element produces new frequencies amongst which are the upper and lower sidebands, one of these bands being the frequency range to which it was originally desired to translate the voice frequency band.

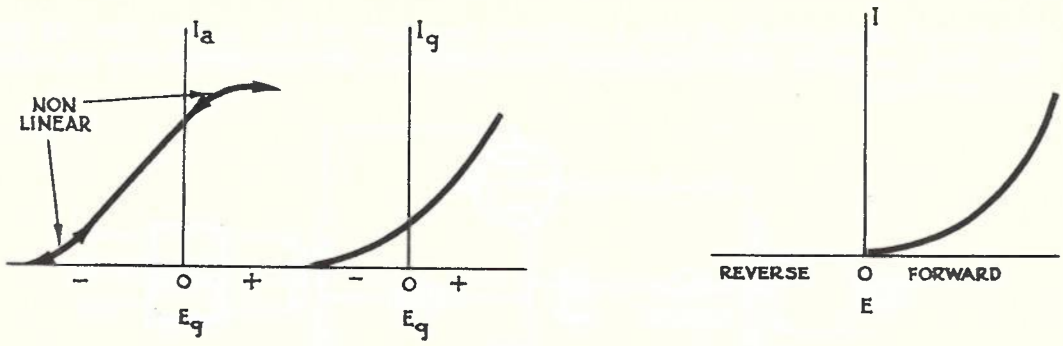
3. CIRCUIT ELEMENTS USED IN MODULATORS.

3.1 From what has been dealt with above it will be seen that the circuit elements employed in modulators must have a non-linear response. Suitable elements are thermionic valves (suitably operated) and metal type rectifiers.

3.2 Fig. 1 shows the characteristic curves of these elements.

From Fig. 1 it will be seen that the upper and lower bends of the anode current versus grid voltage curve could be used for modulation. In practice, only the lower bend is used. Similarly, the non-linear grid current versus grid voltage curve could be used as it is in the modulators of S.T.C. programme carrier systems. When metal rectifiers are used, as they are in all modern carrier systems, the non-linear characteristic of such rectifiers can be used. The modulation products from a metal rectifier modulator seem to be brought about by a number of different actions, so that metal rectifier modulators will be explained along different lines to valve modulators.

/Fig. 1.



(a) Valve Characteristics.

(b) Metal Rectifier Characteristic.

FIG. 1. CHARACTERISTIC CURVES.

4. MODULATORS EMPLOYING THERMIONIC VALVES.

4.1 The simplest form of valve modulator employs a single valve, as shown in Fig. 2.

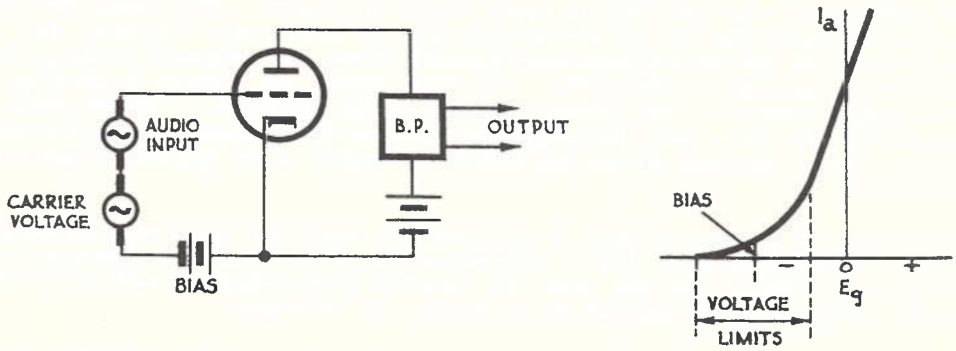


FIG. 2. SINGLE VALVE MODULATOR.

In these modulators, the grid is biased into the lower bend and the carrier and modulating voice voltages are applied in series across the grid and cathode. The alternating voltage applied to the grid, which is the vector sum of the carrier and audio voltages, should not drive the grid beyond anode current cut-off or into the linear part of the characteristic. This means that the anode current will contain the frequencies listed on page 2 of this Paper. To eliminate all products of modulation except the sideband required for transmission, a band-pass filter designed to pass only the required sideband frequencies and reject all others is connected in the output of the modulator.

4.2 In the circuit of Fig. 2, the carrier frequency is suppressed by the band-pass filter following the modulator. This places rather severe demands on the filter, which can be more easily understood by considering the frequencies listed on page 2. There it will be seen that the lowest frequency of the upper sideband and the highest frequency of the lower sideband lie only 200 c/s away from the carrier frequency. This means that the band-pass filter of Fig. 2 has to pass all the frequencies in the sideband selected for transmission and reject the carrier frequency, which is only 200 c/s outside of the pass band. The filter will have to be carefully designed to do this. In practice, the filter design is simplified by using a push-pull arrangement which eliminates the carrier in the windings of the output transformer, just as even harmonics are eliminated in the windings of /the

the output transformer of a push-pull amplifier.

The circuit arrangements of such a modulator are shown in Fig. 3, from which it will be seen that, as far as audio frequencies are concerned, the arrangement acts as a push-pull amplifier.

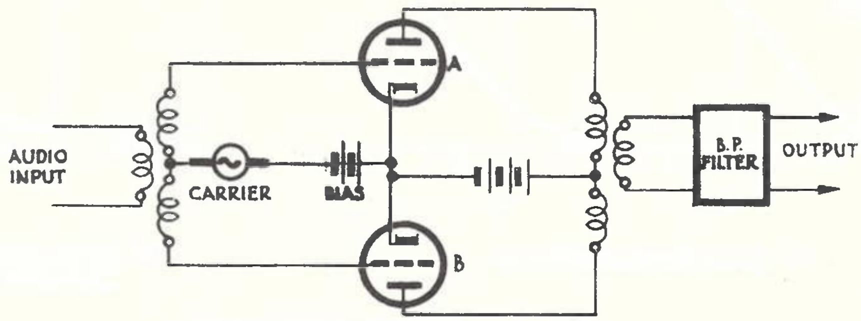


FIG. 3. BALANCED VALVE MODULATOR.

Insofar as the carrier frequency is concerned, this is applied to the grids of the two valves in phase so that the grids of these valves are excited by equal in-phase carrier voltages, causing the anode currents in the two primary windings of the output transformer to rise and fall by equal amounts in phase. This means that there will be no resultant flux in the transformer core at the carrier frequency and, therefore, no voltage of the carrier frequency induced across the secondary winding, so that no carrier is applied to the filter. In other words, the carrier is balanced out in the output circuit by the push-pull action of the modulator. As the grids are biased back into the lower bend, harmonics and sideband frequencies will be produced by each valve, these appearing across the secondary winding to be applied to the filter for selection.

The action is similar to that of the push-pull amplifier dealt with in Paper No. 6. Thus, the arrangement of Fig. 3 eliminates the carrier and even harmonics of the audio frequency, leaving only the audio and sideband frequencies and odd harmonics of the audio frequency to be applied to the filter for selection.

4.3 Modulators employing the grid current versus grid voltage characteristic operate with zero bias, as shown in Fig. 4.

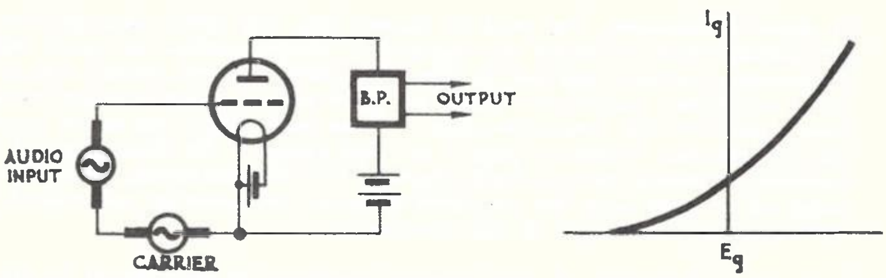


FIG. 4. GRID CURRENT MODULATOR.

Here the modulation takes place in the grid circuit instead of the anode circuit as in the modulators described above. The non-linearity of the grid current versus grid voltage characteristic produces harmonics and sum and difference products of the carrier and audio frequencies in the grid current. This current flowing through the impedance of the grid circuit produces a voltage drop across that impedance, so that voltages of the sideband frequencies are developed across that impedance and applied across the

/grid

grid and cathode of the valve with the carrier and audio voltages. These sideband voltages are amplified by the valve and appear in the anode circuit where the band-pass filter selects the sideband required for transmission. To eliminate the carrier a push-pull circuit of the type shown in Fig. 3 is used, except that no bias is required.

5. MODULATORS EMPLOYING METAL RECTIFIERS.

5.1 In all modern carrier systems, modulation is effected by employing metal rectifier units connected in special arrangements. Copper oxide metal rectifiers have, so far, been the most extensively employed, but some systems are equipped with selenium rectifiers. The operation of both types is similar, and the term "Metal Rectifiers" may refer to either.

5.2 The use of metal rectifiers as modulators dispenses with the necessity to use thermionic valves but, as the input levels it is permissible to apply to metal rectifier units are somewhat restricted, they must be followed by an amplifier. This, of course, must be a thermionic valve arrangement but, as valve modulators and demodulators usually employ an amplification stage, there is no economic disadvantage in the arrangement. The application of metal rectifiers has assisted in greatly changing the appearance of modern carrier systems as they result in considerable saving of space.

The main advantages of metal rectifier modulators and demodulators are -

(i) Greater stability and reliability.
(ii) Economy of space.
(iii) Reduced maintenance costs.
(iv) Where single sideband transmission is employed, a greater degree of carrier suppression is possible than can be obtained with modulator circuits employing thermionic valves.

5.3 There are many ways in which metal rectifiers can be connected together to act as a modulator, and many different circuit arrangements are used in practice. All arrangements used in practice suppress the carrier in the output, as do balanced valve modulators, leaving mainly sideband voltages to be applied to the equipment which follows them. The circuit arrangements used can be reduced to two main types, one of which produces sideband voltages in the output which are twice the amplitude of the sideband voltages produced in the output of the other type. For the purposes of these books the two types have been designated -

(i) Balanced Modulator.
(ii) Double Balanced Modulator.

It is not proposed to describe the operation of every modulator of each type but to describe the operation of one of each, leaving the student to apply the principles used in the examination done here to other modulator circuits as they are encountered in the Course.

5.4 Fig. 5a shows the circuit of a typical balanced modulator, whilst Figs. 5b to 5j illustrate the operating principles. Figs. 5b and 5c show the operation of the circuit when the carrier voltage alone is applied. The rectifiers are conducting to the positive half-cycles of carrier, as shown in Fig. 5b, and blocking to the negative half-cycles as shown in Fig. 5c. Thus, only positive half-cycles of carrier current flow through X and Y, the two primary windings of the output transformer. These positive half-cycles, shown in Fig. 5d, flow through X and Y simultaneously and in opposite directions, and are equal in amplitude. They produce equal and opposite fluxes and, therefore, zero flux, which means that no voltage of the carrier frequency appears across the secondary winding of the output transformer. Thus, the carrier is balanced out or suppressed in the output of this modulator, just as in the balanced valve modulator. By the same reasoning, the carrier voltage does not appear across the primary winding of the input transformer, although carrier current flows through the secondary windings, so

/that

that no carrier appears in the audio circuits.

During positive half-cycles of audio input, shown in Fig. 5e, the modulating audio voltage is impressed across MRA in the conducting direction and across MRB in the blocking direction. As the voltage across MRA is increased and that across MRB is decreased, the positive half-cycles of carrier current flowing through MRA and X will increase in amplitude, whilst those through MRB will decrease in amplitude with the audio voltage, as shown in Fig. 5g. During the negative half-cycles of audio voltage, Fig. 5f, the direction of the audio voltage is reversed and the amplitude of the positive half-cycles of carrier current flowing through MRA and X decreases, whilst the amplitude of those flowing through MRB and Y increases, as shown in Fig. 5g. The difference between the currents through X and Y will be responsible for inducing a voltage across the secondary winding of the output transformer, this difference being shown in Fig. 5h. This, then, will be the voltage induced across the secondary winding of the output transformer.

Some idea of the frequencies contained in Fig. 5h can be obtained by drawing a zero line through the points corresponding to one-half of the amplitude of each half-cycle contained in Fig. 5h, as is done there. When this zero line is straightened out, the result is Fig. 5j. This is not unlike the resultant which would be obtained by adding together the two sidebands of Fig. 5, Paper No. 2. In fact, a mathematical analysis or a practical experiment would show that the output of Fig. 5a above contains the upper and lower sidebands, the modulating voice frequency and harmonics introduced by the half-wave rectifying action of the rectifiers.

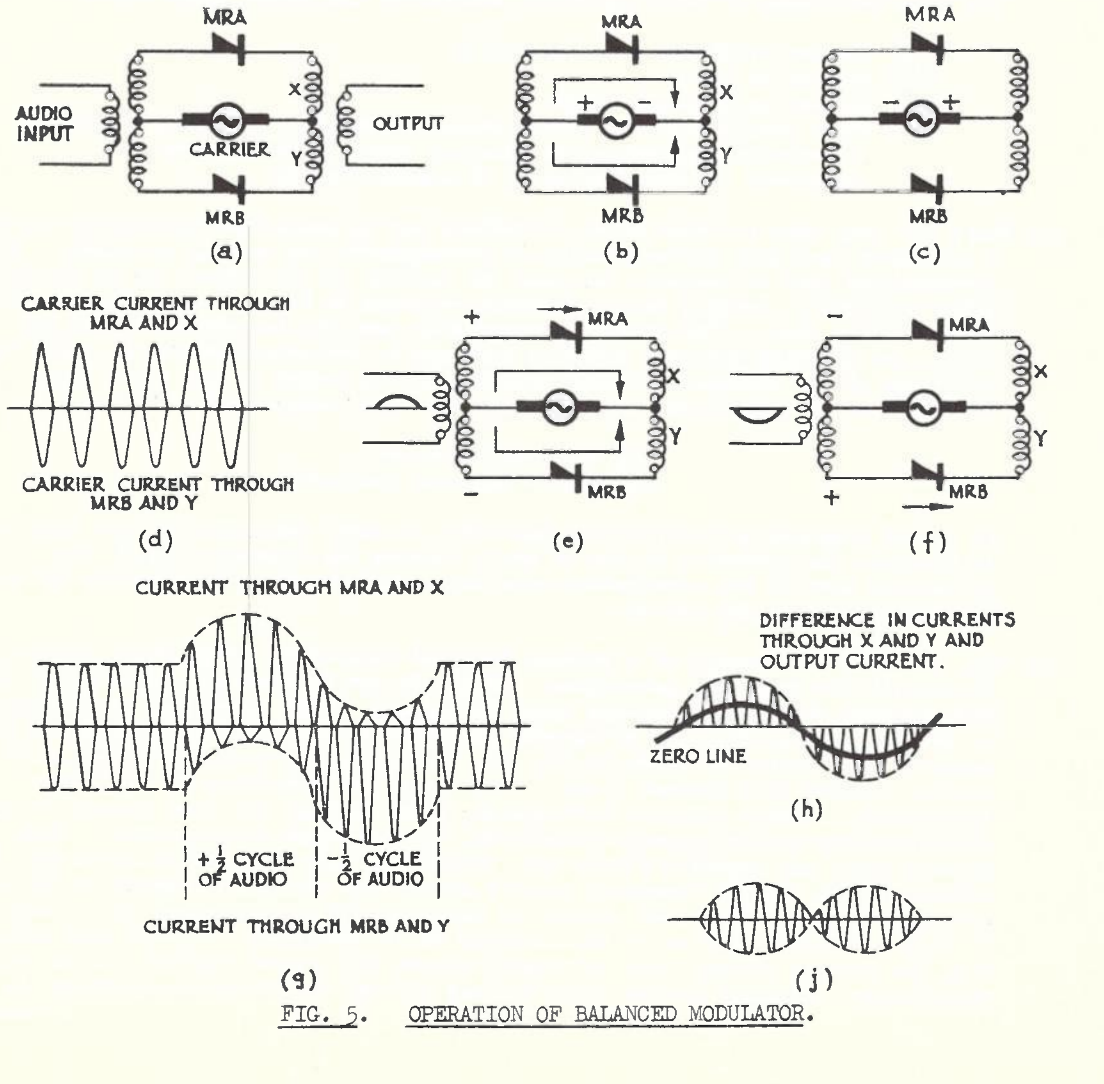


FIG. 5. OPERATION OF BALANCED MODULATOR.

5.5 Fig. 6 shows the circuit and operating principles of a double balanced modulator extensively used in carrier telephone systems.

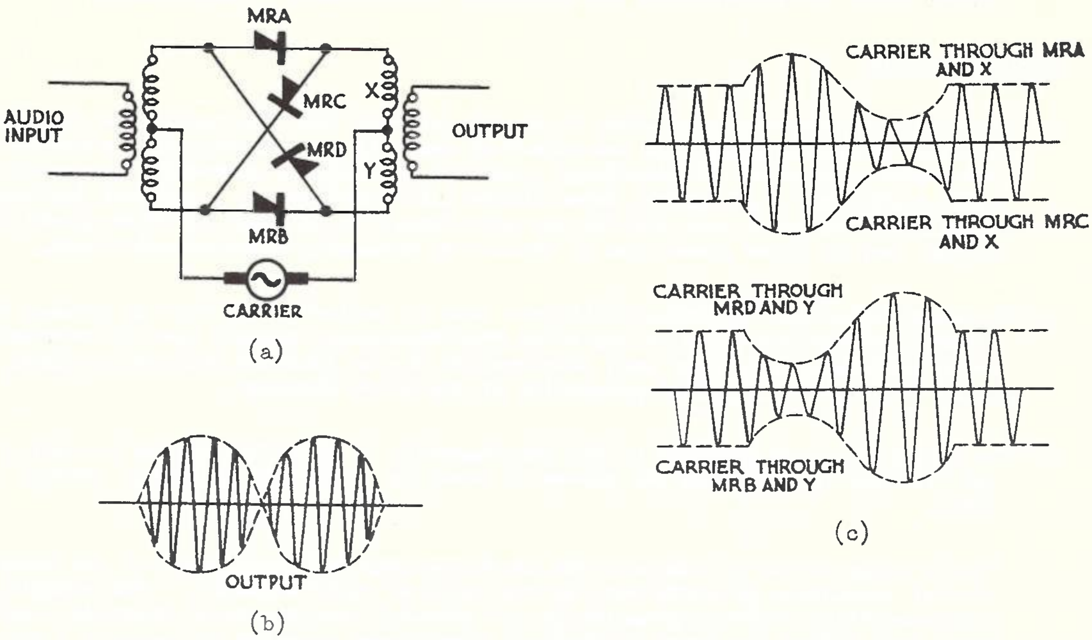


FIG. 6. OPERATION OF DOUBLE BALANCED MODULATOR.

A little consideration will show that rectifiers MRA and MRB produce exactly the same result as they did in Fig. 5, and that rectifier MRC behaves as does MRA for the positive half-cycles of the carrier, whilst MRD behaves as does MRB, the negative and positive directions being those indicated in Fig. 5. The current through each primary winding of the output transformer, therefore, will contain full cycles instead of half-cycles, as indicated in Fig. 6b. The difference between these currents, Fig. 6c, is responsible for the flux which induces the voltage across the secondary winding of the output transformer, and this voltage again contains mainly sidebands as shown in Fig. 6c. The difference between Figs. 5 and 6 is that the sideband voltage output from Fig. 6 is about double that output from Fig. 5, as can be seen from a comparison of Figs. 5h and 6c. Thus, the output of the double balanced modulator contains sideband frequencies mainly, together with other frequencies which can be readily filtered. It should be noted that the modulating frequency, as well as the carrier frequency, is suppressed.

5.6 It is important that the amplitude of the carrier voltage applied to a modulator of the metal rectifier type be always greater than that of the modulating audio voltage. The reason for this can be seen from Fig. 5. If the amplitude of the positive half-cycle of audio voltage in Fig. 5e is greater than that of the carrier voltage, rectifier MRB becomes blocking to the carrier and no carrier current flows through it and, therefore, none flows through winding Y. Besides unbalancing the modulator and permitting carrier to appear in the output, the modulation will be distorted and intelligibility impaired, if not ruined.

6. DEMODULATION.

6.1 As discussed in Paper No. 2, demodulation is a similar process to modulation except that in modulation an audio and a carrier frequency are used whilst in demodulation a sideband and a carrier frequency are used. Thus, the circuit arrangements discussed above are suitable for use as either modulators or demodulators.

7. INTERMODULATION.

7.1 Modulation of one wave by another is always produced when the waves pass together through non-linear apparatus the output of which is not strictly proportional to the input. Examples of non-linear apparatus are valves, networks incorporating metal rectifiers and magnetic cored chokes and transformers. Magnetic hysteresis modulation (being of poor quality and efficiency) is not used for intentional modulation, but it often gives rise to trouble by producing unwanted modulation.

7.2 In transmitting and receiving amplifiers used in multi-channel carrier systems where a number of conversations is amplified simultaneously, special care is necessary to avoid interference between each conversation due to unwanted modulation taking place as a result of the non-linearity of the valve elements.

7.3 The intermodulation produced by the non-linearity of amplifier valves was one of the main reasons for limiting the number of channels on carrier telephone systems to three.

When a large number of channels was simultaneously amplified, some of the inter-channel modulation products were in the range of other channels. For example, the intermodulation products produced by the sideband frequencies of channels 1 and 6 might fall within the bands of sideband frequencies occupied by channels 3 and 9, causing noise and crosstalk to be produced between the four channels. The advent of negative feedback with its reduction of non-linear distortion largely eliminated this inter-channel interference and made a greater number of channels possible in a single system.

8. VOLTAGE LIMITERS.

8.1 The characteristics of metal rectifiers used as modulators and demodulators require that the voice input voltage shall be low in comparison with the applied carrier voltage. It is usual, therefore, to precede metal rectifier modulators by a voltage limiter (sometimes termed "volume limiter") whose function is to keep the audio voltage input to the modulator from exceeding a predetermined value regardless of the level delivered by the subscriber.

One method of complying with these requirements is to make use of the characteristics of a special neon lamp which is connected across the secondary of the modulator input transformer.

At amplitudes below those at which limiting takes place the neon lamp produces negligible loss in the speech circuit. If the voltage during either a positive or a negative half-cycle of speech rises above the permissible limit the neon lamp "strikes" and produces a shunting effect across the input to the modulator, thus preventing overloading. The effect of this limitation of speech power on articulation is negligible, and the flashing over of the limiter cannot be detected by the subscriber.

/Another

Another method of achieving voltage limitation is the employment of metal rectifier elements themselves, and the circuit arrangements are shown in Fig. 7.

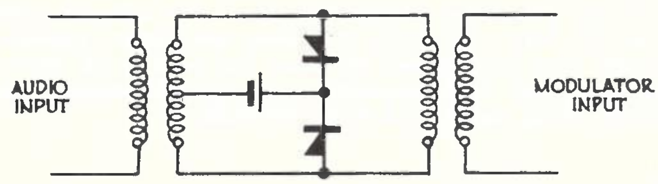


FIG. 7. VOLTAGE LIMITER.

The rectifier is normally biased to a potential which makes its impedance high, so that normally the shunt loss on transmission is negligible. If speech voltage (after transformation) exceeds the bias potential the impedance falls to a very low value and produces a shunting effect across the modulator input, thus preventing overloading.

The use of voltage limiters is not confined entirely to modulators. Voltage limiters can also be usefully employed in preventing overloading of amplifiers and oscillators.

9. BEAT FREQUENCY OSCILLATORS.

9.1 It is frequently necessary to have a variable frequency oscillator for testing purposes. For example, for voice frequency testing an oscillator is required which must be capable of developing any frequency in the voice frequency range 200 to 3,000 c/s. The oscillator used is usually of the "beat frequency" type and contains two oscillators, one developing a fixed frequency and the other developing a variable frequency. The two frequencies are applied to a modulator where sum and difference frequencies are produced. A filter then selects the difference frequency, the frequency of the fixed and variable frequency oscillators being such that this difference frequency is always the frequency required.

9.2 Fig. 8 illustrates the principle.

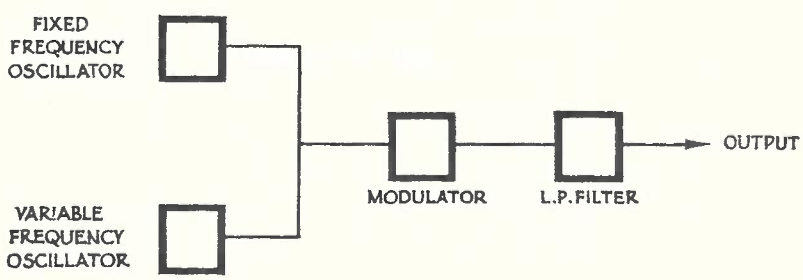


FIG. 8. PRINCIPLE OF BEAT FREQUENCY OSCILLATOR.

The frequency developed by the variable frequency oscillator is varied by having a variable condenser as the tuning condenser, the scale of that condenser being graduated in c/s or kc/s, depending on the range of the whole oscillator. By this means, if a frequency of 1,000 c/s is required the operator merely adjusts the tuning condenser of the variable frequency oscillator to 1,000 c/s, which fixes the resonant frequency of the tuned circuit in that oscillator 1,000 c/s above or below (generally below) that of the fixed frequency oscillator. On being applied to the modulator, these two frequencies will produce, amongst other frequencies, a difference frequency of 1,000 c/s, which is passed by the low-pass filter.

/9.3

9.3 The reason for employing this principle is that a single oscillator capable of being tuned over a wide range would require several sets of adjustable tuned circuits to cover the wide range. The design and operation of the oscillator are considerably simplified by the procedure outlined above. For example, with a fixed frequency oscillator of 650 kc/s a variable frequency oscillator capable of being tuned from 500 to 650 kc/s only is required to cover the band 0-150 kc/s. This represents a narrow band capable of being provided comparatively simply, as compared with a circuit capable of being tuned from 0 to 150 kc/s.

10. TEST QUESTIONS.

1. What is meant by modulation as applied to a carrier telephone system?

2. Explain, in general terms only, how modulation is achieved in carrier telephone systems.

3. What is the advantage of suppressing the carrier in a modulator? Describe how this is achieved in a valve modulator.

4. Explain, in general terms only and with the aid of sketches, the operation of a balanced modulator employing metal type rectifiers.

5. What is the difference between the output of a balanced modulator and a double balanced modulator (metal rectifier types)?

6. Why is a voltage limiter necessary when metal rectifiers are used as modulators? Explain how a limiter employing metal rectifiers functions.

7. Explain, with a block schematic circuit, the operation of a Beat Frequency Oscillator.

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

Engineering Branch,
Postmaster-General's Department,
Treasury Gardens,
Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

LONG LINE EQUIPMENT I.

LONG LINE EQUIPMENT POWER PLANT.

CONTENTS:

1. INTRODUCTION.

1.1 Basic differences exist in the design of power plant suitable to operate Long Line Equipment as compared with power plant associated with normal Telephone Equipment.

The power plant arrangements employed in telephone equipment practice are probably familiar, and a brief comparison of the different conditions of employment will serve to illustrate the nature of the problem affecting design and choice of plant suitable for long line equipment. In telephone equipment practice, continuous floating is adopted wherever practicable. The emergency condition of power mains failure is provided for by employing batteries of sufficient capacity to meet the load requirements for the busiest 12 hours' traffic. The exchange load varies considerably from hour to hour, and the permissible voltage variation is usually of the order of $\pm$ 6%. In large exchanges duplicate batteries are generally employed, and end cells or carbon pile regulators are employed to keep the voltage variations within permissible limits.

As in telephone equipment practice, so in long line equipment practice continuity of the power supplies is of primary importance, as a failure at any station on a main trunk route obviously affects all repeater or carrier circuits working through the station concerned.

A point of difference, as compared with exchange equipment, is that the long line equipment drains are, in general, reasonably constant over the full 24 hours, the only variations being those due to voice frequency telegraph circuits and the occasional use of testing equipment. Another point is that the permissible voltage variations of the supplies are more stringent than in the exchange practice, and it is generally desirable to keep them within approximately 1% of their normal value in order that variations in valve characteristics, which result from variation in supply volts, will not cause serious alteration in circuit performance.

2. BATTERY SUPPLIES.

2.1 Typical power supplies required for the operation of long line equipment are -

Battery	Voltage	Used For	Regulation
Filament or Heater Battery.	24 -	Filament heating and miscellaneous relay circuits.	Regulated to 24 ± 0.25 volts for filament circuits. Unregulated for miscellaneous circuits.
Normal anode or Plate Battery.	130 +	Anode potential	Regulated to 130 ± 2 volts.
Special Anode or Plate Battery.	200 +	Anode potential	Regulated to 200 ± 2.5 volts.
Telegraph Batteries.	(130 + (130 -	Send and receive loops telegraph systems.	130 ± 2 volts.
Alternative Telegraph Batteries.	(120 + (120 -	Send and receive loops telegraph systems.	120 ± 2 volts.

2.2 A 200 volt, or greater, anode potential is seldom used for carrier equipment. The question of the advantages to be obtained (particularly amplifier gains) by employing anode voltages in excess of the standard 130 volt potential is often debated, more particularly when commercial type radio valves are used. It is true that these valves are operated to greater advantage with anode voltages of approximately 200-250 volts. However, because of the large quantities of equipment designed for an anode voltage of 130 volts already installed, it is likely that this voltage will be used for some considerable time. Therefore, although some advantages are possible by use of higher anode voltages, the economic aspect justifies the retention of 130 volt standards. Special valves have been developed, therefore, to give satisfactory voltage amplification or adequate power handling capacity with low anode volts.

In equipment designed in recent years, some slight improvement in amplification and power handling capacity has been achieved by using 154 volts on the anode, this being obtained by returning the 130 volt battery to earth via the 24 volt filament battery. An example is the employment of the W.E. 310 and 311 A (regulated supplies) or their equivalent - W.E. 328 and 329 A (unregulated supplies) - as voltage and power amplifiers, respectively, in Type J, K, CS and CU carrier equipment.

/3.

3. TYPES OF POWER PLANT.

3.1 The general long line equipment power plant arrangements may be divided into two main classes -

(a) Where reliable commercial mains power supply (alternating current or direct current) is available.

(b) Where commercial mains power supply is not available.

3.2 (a) Where reliable commercial power supplies are available, it is usual to provide for the use of regulated float operation employing duplicate batteries. Such factors as reliability and constancy of commercial supplies, space available in existing buildings and maintenance requirements tend to affect the choice of plant adopted. The following details indicate the arrangements usually used -

Staffing Arrangements	Power Plant Installed	
	For Normal Working	For emergency (that is, failure of mains supply)
Continuously attended.	Duplicate motor generator sets and floating batteries.	Hand started petrol or diesel engine alternator to replace mains supply or, alternatively, engine arranged to drive the motor generator set for floating purposes.
Unattended at night and week-ends.	As above.	As above, but engine arranged to start automatically.
Unattended except for periodical maintenance visits.	Rectifiers and floating batteries.	Automatic change-over to separate batteries having sufficient capacity to maintain service for 24 hours.

Note. The introduction of selenium rectifiers, with resultant increased outputs consistent with reasonable size and cost of rectifier elements, is tending to offset the advantages of motor generator sets for continuously attended stations, the reduced maintenance being a decisive factor.

3.3 (b) When long line repeater stations are established in remote areas, a commercial mains supply is seldom available and the charge-discharge method of battery working is used. In general, such stations are attended on week days but not at nights or week-ends. Arrangements in general use are as follows -

/Staffing

Staffing Arrangements	Power Plant Installed.	
	For Normal Working	For Emergency (that is, failure of normal charging plant)
Unattended at nights and week-ends.	Hand started petrol or diesel engine direct coupled to 24 and 130 volt generators - duplicate batteries on charge-discharge operation.	Duplicate engine set or wind driven generators.
Unattended.	Automatically started petrol or diesel alternator set operating 24 and 130 volt rectifiers - duplicate batteries on charge-discharge operation.	Duplicate engine alternator sets automatically switched in circuit on failure of other set.

Note. Each of the two batteries for both 24 and 130 volt supplies has sufficient capacity for at least 24 hours' operation.

4. FLOATING BATTERIES.

4.1 As mentioned previously, when reliable commercial supply mains are available it is usual to provide duplicate batteries and employ a floating procedure. In such a procedure the charging source is continuously across the battery being used, so that this charging source (for example, motor-generator set or rectifier unit) supplies the power to the load, the battery acting mainly as a fully charged stand-by in the event of a mains supply failure.

4.2 The floating procedure replaced a charge-discharge procedure. In the charge-discharge scheme, duplicate batteries, designated No. 1 and No. 2, are provided. Whilst No. 1 battery, fully charged, provides the power for the load for a predetermined period (usually 24 hours), No. 2 battery, which has been discharged by supplying the load for the previous 24 hours, is being charged. The disadvantage of this scheme is that two batteries with a capacity sufficient to cover the requirements for a 24 hour period are necessary. Also, to charge either of these batteries in the usual period of 10 hours a motor generator set or rectifier unit with a large output is required, that is, the battery plant and its associated power supply is, in general, fairly large. At stations which are unattended over week-ends it is necessary for each battery to supply the load for 48 hours, thus further increasing the size of the battery and its associated power plant.

4.3 By adopting the floating procedure an extensive reduction in the size of the batteries and their associated power plant is brought about. When no emergency plant is available, the capacity and, therefore, the size of the batteries are reduced by half. This is because when either battery is in use the charging source is connected across it and that source, and not the battery, supplies the load. Thus, the full capacity of each battery is available in the event of an emergency. When emergency charging sources are available, the batteries have to supply the load only for the time between the supply mains failing and the emergency source being started.

The advantages of floating, as compared with a charge-discharge procedure, can be summarised as follows -

(i) Smaller size batteries with consequent reduction in initial cost and floor space.

(ii) Longer life of batteries owing to the elimination of strain on the plates which occur due to the plates changing shape over the charge-discharge cycle.

(iii) Reduction in sediment deposition for the same reason as in (ii) above.

(iv) Retention of the full rated capacity of the battery throughout its working life.

5. GENERATOR RIPPLE.

5.1 The introduction of regulated floating operation has created one difficulty which is not present with the charge-discharge method of working, namely, the introduction of audio frequency ripple from the generators or rectifiers to the equipment bus-bars. This condition is a particularly serious one so far as long line equipment in general, and filament supply in particular, are concerned. The filament battery is widely used for the derivation of grid bias, and any ripple superimposed on this supply, therefore, will be amplified and result in noisy circuits. The elimination of ripple, therefore, forms an important part of specifications for floating type generators and rectifiers. It is invariably necessary to employ smoothing circuits, consisting of choke coils and electrolytic condensers, in association with such charging equipment in order to reduce ripple voltages to the specified limits.

5.2 The permissible limits of ripple from generators are specified as follows -

(a) Noise measured across the generator terminals without smoothing equipment must not exceed that noise equivalent to 2 millivolts R.M.S. at 800 c/s per direct current volt generated at full load current. (The above values are specified in order to ensure that generators having reasonably small noise content are obtained.)

(b) Values covering the limits of noise not to be exceeded at the busbars in the long line equipment room are -

35 Volt Generators. When floating across the filament battery via the smoothing circuit, the noise measured at the busbars in the repeater room must not exceed that equivalent to 0.5 millivolt R.M.S. at 800 c/s at any value of current from quarter to full load.

180 Volt Generators. Similar to above, except that the noise measured must not exceed that equivalent to 5 millivolts (recently reduced from 7 millivolts) R.M.S. at 800 c/s.

5.3 It should be appreciated that the ripple superimposed upon the load at the equipment busbars is that developed across the battery, assuming the drop in the leads to be negligible between the power board and the load. The equivalent circuit is shown in Fig. 1.

/Fig. 1.

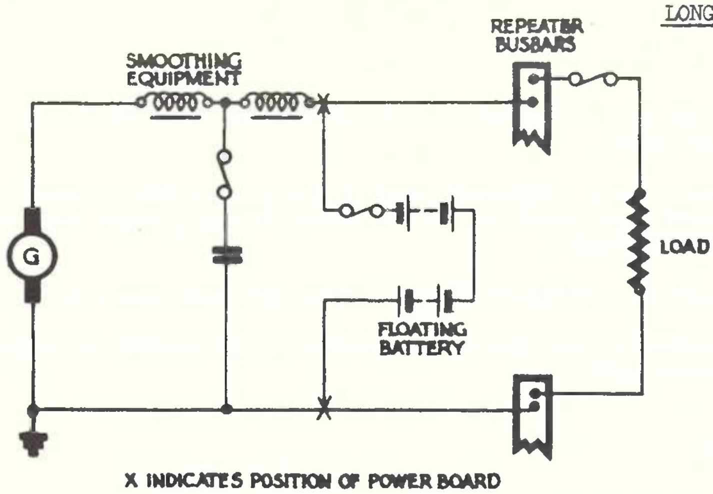


FIG. 1. POWER CIRCUIT FOR LONG LINE EQUIPMENT.

The impedance of the battery circuit must be kept as low as possible and, in view of this, the layout of the plant is arranged so that the shortest possible route is obtained for the cables between the power board and the battery. The impedance is reduced further by arranging each battery in two halves, thereby forming a "loop" with a resultant reduction of battery circuit inductance to a minimum. The reduction of inductance is important since it has been found by test that the alternating current impedance of a secondary cell battery has an inductive reactance which, at 800 c/s, is comparable to its effective resistance. This inductive reactance is between 10 to 20 microhenrys at 1,000 c/s measured with an Owen Bridge. The reactance of a 20 microhenry inductor at 800 c/s is 0.1 ohm, and this value may be taken as representing the impedance of a standard battery installation at this frequency.

5.4 Generators having a low frequency ripple are most likely to fulfil the required noise conditions. A low frequency ripple is also preferable to high frequency ripple due to the fact that the ear is considerably less sensitive to low frequency notes than to those of approximately 1,000 c/s of equal amplitude. This is illustrated by reference to the weighting curve (Fig. 2) in which all frequencies are weighted to the corresponding interference value at 800 c/s.

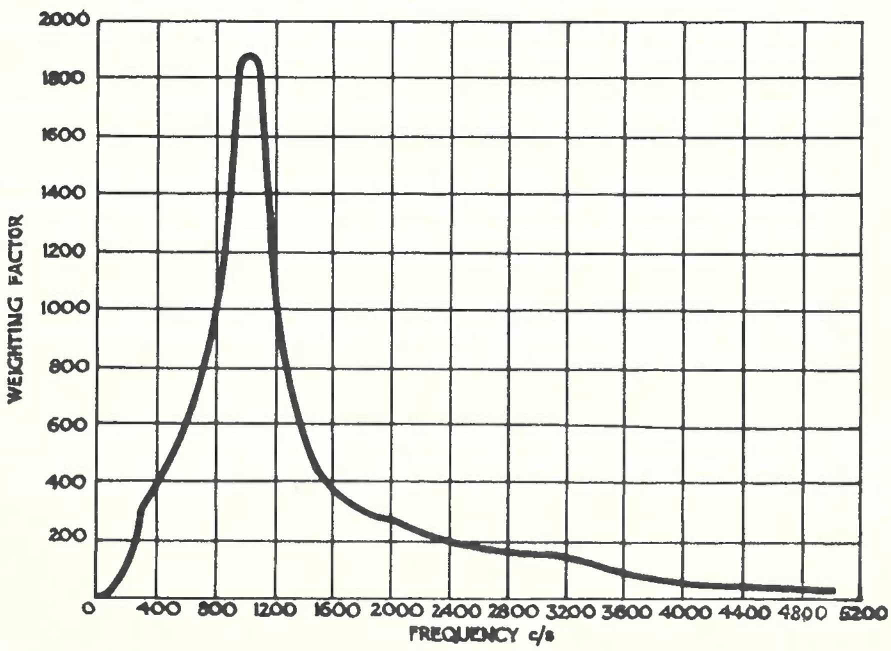


FIG. 2. WEIGHTING FACTOR.

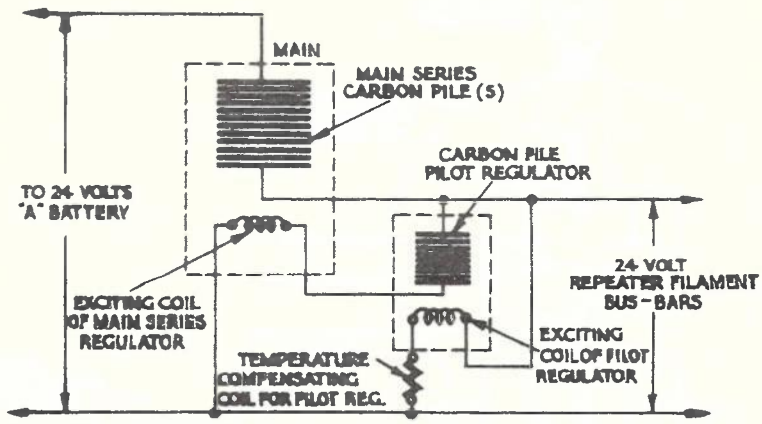
The "weighted" values indicate the amount of disturbance caused at different frequencies relative to that at 800 c/s for equal values of disturbing voltage, for example, the disturbance at 200 c/s would be only 0.1 of that at 800 c/s.

6. AUTOMATIC VOLTAGE REGULATION.

6.1 The most important requirement to ensure the success of a continuous float battery scheme is the application of a constant voltage from the floating generator or rectifier. This permits the battery to be kept in a satisfactory condition and, at the same time, provides a steady voltage for the discharge circuit. The usual hand regulation is inadequate to cope with momentary changes in output current, and some automatic device must be employed for the purpose.

Two types of automatic voltage regulators in use are Carbon Pile Voltage Regulators and Vibrating Contact Regulators. Other methods, such as Diverter Pole Generators and Contact Voltmeters operating regulating resistors via reversible motors, are under consideration. In addition, there are many types of Regulated Rectifiers available using "saturated choke coil phaseshift control" and, with the development of the Thyratron type of thermionic valve, many types of regulated rectifier circuits, making use of the special properties of this type of valve, have been developed.

6.2 Carbon Pile Automatic Voltage Regulator (see also Telephony V). The application of this type of regulator is shown schematically in Fig. 3.



DIAGRAMMATIC CIRCUIT OF USE OF CARBON PILE. AUTOMATIC VOLTAGE REGULATORS.

(Note. In the case of the pilot regulator, an increase of current through its exciting coil will cause a decompression of its carbon pile, with a resulting increase of resistance. The reverse is the case in the main regulator.)

FIG. 3.

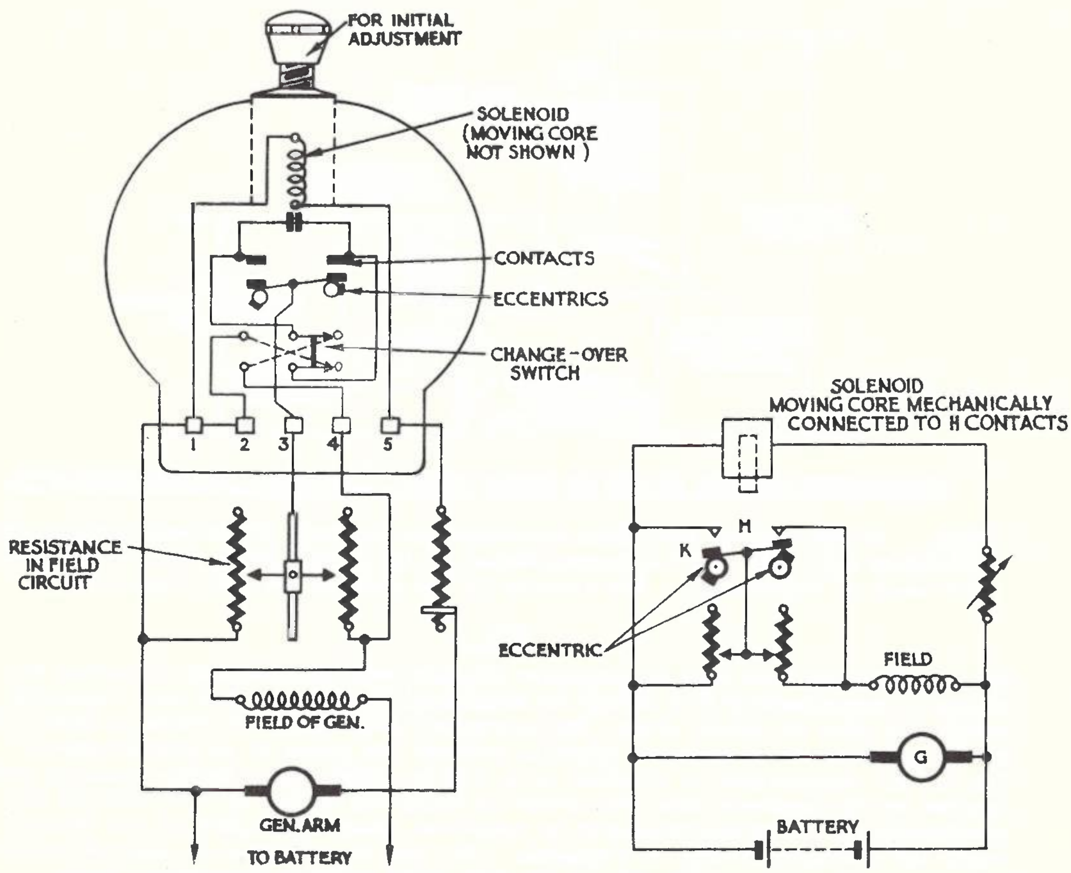
The operation of this type of regulator depends upon the property of carbon to change in resistance with a change in mechanical pressure. The main regulator consists of a pile of carbon discs, the mechanical compression of which is controlled by the position of an armature rotating between the poles of an electromagnet. The initial compression is obtained by a spring, which has a torgue opposing that exerted by the armature when the latter is being influenced by the field of the electromagnet. The electromagnet is connected across that part of the circuit where the constant voltage is required. If, due to variation in applied volts or change in load current, the voltage at this point alters, the correct field strength will no longer obtain and the armature will rotate with a resulting change in carbon pile resistance until once again the desired voltage exists. The armature now comes to rest, due to the state of equilibrium existing between the forces exerted on it. /Usually

Usually, as shown in the circuit two regulators are employed, the larger being actuated by the pilot. This is necessary when close limits of regulation are required as the pilot regulator of lighter construction is more sensitive.

The pile of the main regulator will carry 25 amperes, and three can be connected in parallel to give a maximum current carrying capacity of 75 amperes.

The overall regulation obtained is ± 0.25 volt for 24 volt supplies and ± 1.3 volts for 130 volt supplies.

6.3 Vibrating Contact Regulator. A regulator of this type is shown schematically in Figs. 4a and 4b. The contacts K and H are vibrated by the eccentrics which are belt driven from the generator shaft. The contacts alternately short-circuit and insert a resistance in the field circuit of the generator. This alternate shorting and inserting of the resistance produces an average value of field current which maintains the correct voltage. The relative position of the short-circuited contacts H and the resulting value of field current are controlled by a movable core of the solenoid, which is connected across the generator terminals. In the event of the generator voltage tending to rise, the core lifts the contacts H and causes a longer break period which results in a reduction of field current and corresponding restoration of voltage. The reverse operation occurs on decreased generator voltage. A screw cap provides initial adjustment of the solenoid core. A condenser prevents contact sparking. The changeover switch provides for disconnection of regulator and also allows for daily reversal of contact polarity to reduce contact wear.



(a) Diagrams of Connections.

(b) Simplified Circuit of Vibrating Contact Regulator.

FIG. 4. VIBRATING CONTACT REGULATOR.

This regulator allows a ± 0.5 volt variation in 24 volts and a ± 2 volt variation in 130 volts from a condition of no load to a condition of full load.

7. TELEGRAPH BATTERIES.

7.1 Where carrier telegraph equipment is installed and double current working is employed, it is necessary to provide for positive and negative current supply for the send and receive telegraph loops.

By using 130 volt positive and negative batteries for telegraph supplies it is practicable to employ the normal 130 volt positive anode battery for the "spacing" supply, the "marking" supplies being obtained by installing duplicate 130 volt negative batteries. Alternatively, separate positive and negative 120 volt batteries are installed in duplicate for the telegraph apparatus.

7.2 The size of the telegraph installation is usually the deciding factor in the selection of one of these methods. It is usual to employ the anode battery for positive telegraph battery drains in small installations, but in larger installations at intermediate stations 120 volt batteries are installed and utilised for normal telegraph equipment as well as loop supplies.

7.3 In main centres the loop supplies are often derived from 120 volt positive and negative direct current generators direct coupled to a common motor. It is necessary that the generator impedance be very low (less than 1 ohm) to avoid interference between circuits.

8. UNREGULATED EQUIPMENT.

8.1 Where automatic regulation equipment is not installed, the variations in supply volts to filament circuits are offset to some extent by installing ballast lamps in series with filament circuits.

9. WIND DRIVEN GENERATORS.

9.1 In remote locations, where conditions are suitable, wind driven generators have been used as a means of deriving 24 and 130 volt power supplies. The operation of the equipment depends on the presence of wind, but experience to date indicates that most locations in the Commonwealth are suitable in this respect, particularly as the latest designed plants commence charging at wind velocities of 5-6 miles per hour and charge at 50% rated output at wind velocities of 8-10 miles per hour. The mill is housed in a 40 ft. 4-posted V-shaped steel tower, separate towers being erected for 24 and 130 volt generators. Generator maximum capacities are 35 volt-60 amperes and 180 volt-12 amperes.

9.2 Equipment Features.

(1) Latest pattern aero foil all steel propeller blades carried on fixed central shaft of tempered spring steel on which they feather to the wind when governing.

(2) Governing effected by automatic device based on centrifugal force which varies the pitch of the propeller blades to the wind and ensures reasonably constant generator voltage.

(3) Tail fin - double fuselage type overcoming slip stream effect.

(4) Main drive by wedge shaped belts running over a plain face pulley. This form of drive reduces wear and overcomes alignment difficulties experienced with former oil immersed gear driven assemblies.

(5) Special expanding brake operated from ground capable of stopping mill in strongest gale.

/10.

10. METAL RECTIFIERS.

10.1 Metal rectifiers in general use consist of a disc of pure copper coated on one side with a thin layer of cuprous oxide. This layer is approximately one-thousandth of an inch thick and is deposited on the surface of the copper by a series of heat treatments. The oxidised disc has the peculiar property of conducting a current of electricity in the direction of oxide to copper, but not in the opposite direction. Perhaps it would be more correct to say that the resistance to a passage of current in the oxide to copper direction is very low but in the opposite direction is very high.

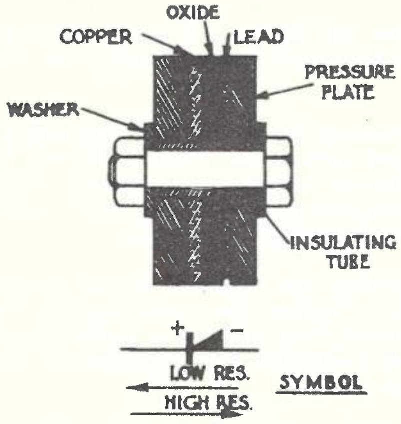


FIG. 5. COPPER-OXIDE RECTIFIER.

A single copper oxide element is shown in Fig. 5, together with the standard symbol.

This Paper will cover the power applications of the metal rectifier. Other applications are to be found in other Papers of Long Line Equipment I, II and III.

10.2 Fig. 6 shows how the resistance of a copper oxide element varies when negative and positive voltages are applied. The vertical ordinate of the right-hand curve is drawn on a scale approximately 200 times larger than the vertical ordinate on the left to illustrate more clearly the lower bend on the curve. It may be seen that to a pressure of 1 volt negative the resistance of the unit is approximately 4,000 ohms.

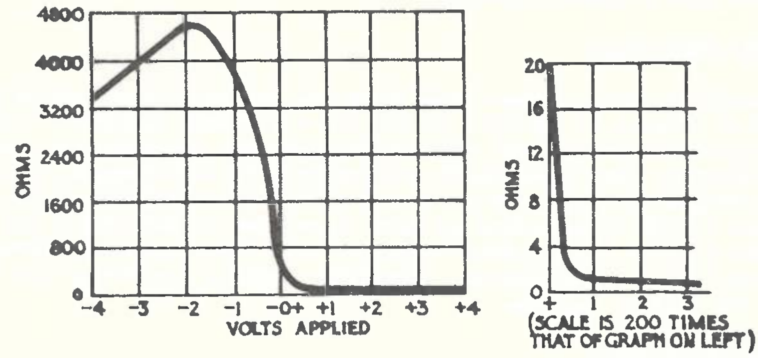


FIG. 6. VOLTAGE-RESISTANCE CHARACTERISTIC COPPER OXIDE UNIT.

10.3 Owing to the very light film of oxide and the difficulties of establishing close contact, the power carrying capacity of rectifier discs is not very great; therefore, except for very light loads, series parallel groups are used. Rectifier discs are connected in series where a high voltage is to be connected and in parallel to increase the current carrying capacity. In practice, it is not usual to allow an alternating voltage of more than 5 volts to be impressed across any single disc, so that a 24 volt application would require at least five discs in series. If commercial 3/4 ampere discs were used this combination would give 3/4 ampere from the five discs in series, and to give 3 amperes, for example, 20 discs would be required, five in series and four sets in parallel.

10.4 Large power ratings, therefore, will require a large number of copper oxide discs with consequent high cost. In some cases this difficulty is overcome, to some extent, by running the rectifiers at somewhat higher ratings than the conservative safe standard and cooling them during operation by means of a fan.

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10.5 The mechanical assembly of a copper oxide rectifier unit, as used for power rectification, is shown in Fig. 7.

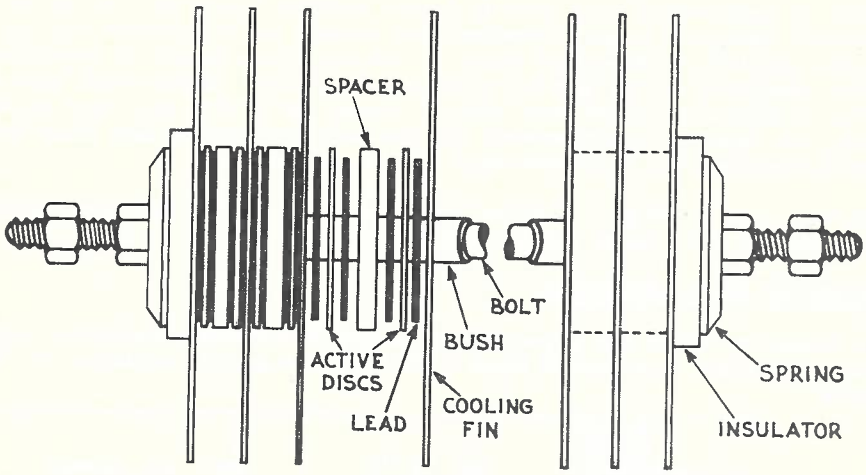


FIG. 7. MECHANICAL ASSEMBLY OF METAL RECTIFIER.

A number of oxidised copper discs are arranged on a spindle from which they are insulated by a non-conducting sleeve. Lead washers are inserted between discs to preserve intimate contact with the copper oxide, whilst metal spacing pieces and cooling fins are inserted alternately between discs to facilitate natural cooling of the whole unit. The whole assembly is clamped together by nuts and washers which must always be tight to ensure good contact. For heavier current requirements a copper oxide rectifier has been developed in plate form. This type is usually fan cooled and, under certain conditions, can be rated up to 20 amperes per plate.

10.6 Copper oxide rectifiers suffer a rapid decrease in the amount of current they pass in the first three or four months of use. This deterioration is called "aging." The decrease in initial output does not usually amount to more than 25%, however, and this is anticipated in the initial design. The amount of aging a rectifier undergoes is also due to the effects of temperature. For this reason, current drains in excess of the rated output should be avoided as excess current will raise the temperature, increase the aging and may destroy the rectifying action of the discs. The diameter of the elements governs the current rating and the number of plates governs the reverse voltage which can be applied.

10.7 The system of designation adopted consists of two numbers separated by a bar, the first indicating the number of groups in the unit and the second the number of plates in each group. This number is followed by a suffix letter which serves to indicate the type of disc. For example, a unit designated as 1/12A would mean one unit consisting of 12 "A" type discs. Normally, the ends are of opposite polarity but, in certain cases, this is departed from and the letter N or P is added to the group number to signify that the rectifier is assembled in such a manner that both ends are either negative or positive.

10.8 The unit largely used in circuits is the one designated 1/12A and, generally, this unit has a red disc at one end. This is the copper terminal (positive), and the negative battery must be connected to this terminal to obtain forward current. When carrying a forward current of 50 milliamperes, the resistance of this unit is approximately 40 ohms. With current in the opposite direction, the resistance is approximately 20,000 ohms. It should be remembered, however, that the resistance of the rectifier increases as the applied potential decreases.

/10.9

10.9 The maximum safe working reverse voltage recommended by the manufacturers is 5 volts per disc, but this figure applies only to its use as a rectifier of alternating current. Experience has indicated that where a unidirectional potential is applied for long periods a safe working figure is 3 volts per element, and circuit designers should remember that where a direct current voltage is continuously applied the number of discs in the rectifier should be numerically not less than one-third of the applied potential in volts. Metal rectifiers may be made from combinations other than that of copper and oxide, and, as a matter of interest, the junction point between any two different conductors gives some degree of rectification when an alternating current is applied.

Another type of metal rectifier frequently used is the Selenium Rectifier which consists of iron discs coated with selenium.

11. APPLICATIONS OF METAL RECTIFIERS.

11.1 The metal rectifier was primarily developed to convert alternating current to direct current, and Fig. 8 shows a circuit arrangement suitable for this conversion.

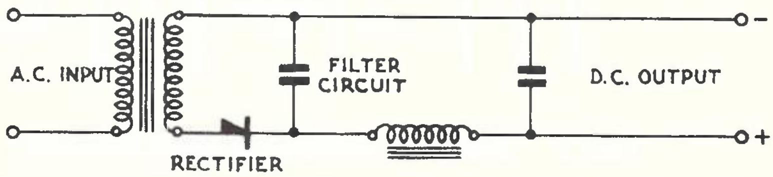


FIG. 8. SIMPLE CONVERTER USING HALF-WAVE RECTIFIER AND FILTER CIRCUIT.

Alternating current is connected to the primary winding of a transformer (see Fig. 8) and stepped up, via the secondary winding, to the voltage required to give the proper direct current output potential and, at the same time, compensate for losses in the converter. The transformer also performs the function of isolating the alternating current mains from the direct current side of the converter. The transformer output is passed through a rectifying unit which transforms the alternating current to pulsating direct current. The direct current thus produced would not be satisfactory owing to the low frequency hum present which must be eliminated by a filter circuit consisting of a series choke or inductance and a shunt capacity. The resultant current is approximately normal direct current.

Metal rectifiers can be used in circuits to provide either full-wave or half-wave rectification. As the term implies, the latter type (which is shown in Fig. 8) makes use of only one half of the alternating current cycle, the other half being lost. A condenser bridged across the output is discharged during the ineffective half-cycle, thus tending to maintain the direct current.

11.2 Full-wave rectification can be obtained by connecting a number of rectifiers in the form of a bridge as shown in Fig. 9. This method is generally adopted with metal rectifiers.

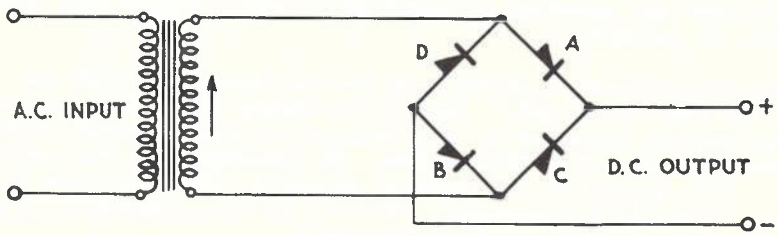
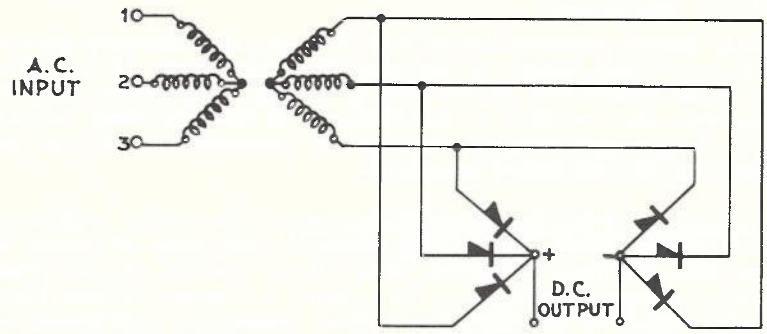


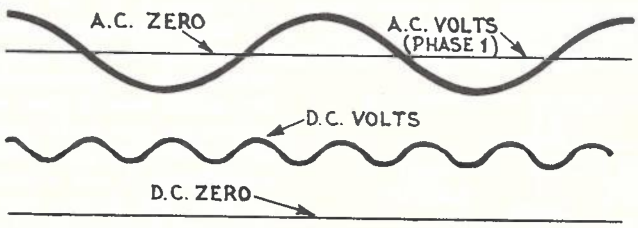
FIG. 9. FULL-WAVE METAL RECTIFIER (BRIDGE CONNECTION).

The action of the Bridge Rectifier may be seen from the diagram. A positive pulse from the line transformer in the direction of the arrow passes through rectifier A, through the load and back to the transformer via rectifier B. This pulse is in the copper to oxide direction in rectifiers C and D, which therefore block the current flow. On the next half-cycle, a pulse takes place in the opposite direction and readily passes through rectifier C (maintaining the correct polarity at the load), through the load as before and back through rectifier D. Rectifiers A and B are now blocking until the next reversal, when the original condition will be restored.

Where a large output is necessary from a rectifier and strict requirements of smoothing are specified, the best type of rectifier is the 3-phase full-wave rectifier using copper oxide discs as shown in Fig. 10.



(a) A Full-wave Bridge Connected Rectifier Three Phase Supply (Unsmoothed).



(b) Voltage Wave Form - Resistance Load.

FIG. 10.

11.3 Efficiency of the Metal Rectifier Converter. Assuming the efficiency of a metal rectifier converter to represent the ratio of the direct current watts output compared to the watts input measured on a wattmeter, the efficiency would be from 50% to 60% without smoothing. Whilst valve rectifiers give a higher efficiency than this, metal rectifiers have the advantage of low maintenance and no risk of glass breakages.

In Long Line Power supply electrical silence is important and, as more use is now being made of mains operated equipment, strict requirements for purity of output must be specified. In many cases, it is necessary to specify a total ripple voltage in the output of the rectifier of less than 0.1%. The tendency is to specify a certain total noise voltage to be measured with a special instrument (a Psophometer) which discriminates against different frequencies according to their disturbing effect upon the ear. Thus, a large voltage of a harmless frequency might be present in the output but the instrument is so arranged that this does not give a large deflection. On the other hand, disturbing frequencies, such as those in the vicinity of 800 or 1,000 c/s, are attenuated to a lesser degree.

/Each

Each particular arrangement of rectifier elements is accompanied by a characteristic ripple voltage which predominates in the output. For example, a single phase half-wave rectifier contains mostly 50 c/s ripple, whereas the greatest trouble in the case of a full-wave single phase arrangement comes from a frequency of 100 c/s. Similarly, in a 3-phase half-wave rectifier the predominant frequency is 150 c/s, and in a full-wave 3-phase rectifier 300 c/s. Multiples of these frequencies are also more noticeable.

The presence, for example, of a 50 c/s note in the output of the 3-phase rectifier would immediately indicate trouble either in the mains or the grouping of components. Unbalanced mains would allow a 50 c/s note from one phase to predominate in the output due to slightly higher amplitude than other phases, but the more likely cause would be due to bad screening or an ineffective arrangement of transformer cores to avoid inductive pick-ups. This trouble would be more noticeable where a separate transformer is used for each phase.

Where only partly smoothed rectifiers are used for charging batteries and where the currents are fairly heavy, it has been found necessary to arrange the battery leads and the cells themselves in the form of a loop - otherwise inductive coupling takes place between the charging of one battery and the discharge circuits of the other battery. This requirement is particularly necessary where high gain amplifiers are operated from the batteries concerned.

11.4 The Regulation of Metal Rectifier Converters. The output regulation of metal rectifiers is not of a high order, and a comparatively wide range of output voltage must be expected when the load is varied, say, from a quarter to full load. To overcome this difficulty, a common practice is to connect a bleeder ballast resistance permanently across the output of the rectifier. Some commercial rectifiers supplied with carrier systems include special carbon pile voltage regulating devices together with an automatic relay which brings in an artificial load should the normal drain be unexpectedly reduced.

11.5 Voltage Doubler Circuit. An interesting application of the metal rectifier is in the voltage doubler circuit shown in Fig. 11. This apparatus provides full-wave rectification as well as applying double the supply voltage to the output terminals. For each half-cycle of alternating current one of the two metal rectifiers is in the conducting direction and will permit its associated condenser to be charged. When under no-load conditions, condensers A and B charge during successive half-cycles, and double the supply voltage is obtained across the terminals. On a load being connected, during the half-cycle indicated by the full-line arrow, condenser B discharges through the load in series with the supply.

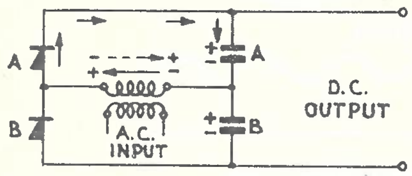
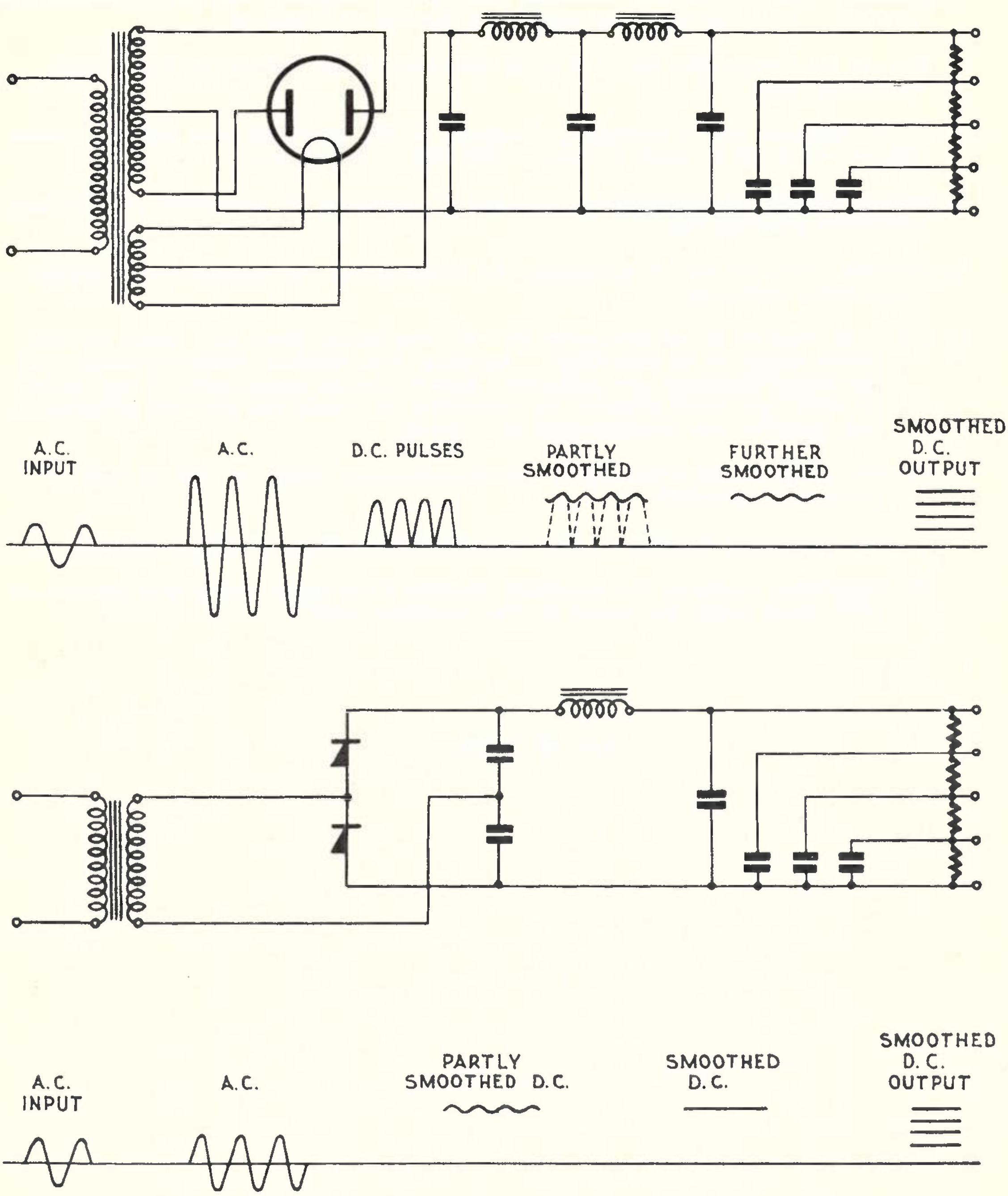


FIG. 11. VOLTAGE DOUBLER CIRCUIT.

As the potentials of both the supply and condenser B are in the same direction the sum of their voltages is applied to the load. During the next half-cycle condenser A discharges through the load in series with the supply while condenser B is being recharged. As the resistance of the load is increased from infinity, the rate of discharge of the condensers increases, and over each half-cycle the average voltage applied to the load becomes less. Consequently the use of the voltage doubler is limited to circuits requiring small currents, say, not greater than 150 to 200 milliamperes; because in order to draw greater currents without causing a large decrease in voltage, large capacity condensers would be necessary. As it is essential that they be of the paper or mica type they would have to be of considerable size to provide the required capacity. This scheme has the advantage that a high voltage is not impressed across the transformer winding.

A smooth output is obtained from the voltage doubler circuit and, for this reason, voltage doubler rectifiers are used to supply anode currents to valves used in amplifiers, studio equipment, etc. Fig. 12 gives a graphical representation of the ripple waves present in different parts of two types of rectifier. The top diagram is the conventional Full-Wave Thermionic Valve Rectifier, whilst the bottom diagram is a voltage doubler circuit. The relative amount of noise which would be present in the outputs of these rectifiers can be gauged from the amplitude shown on the wave diagram.

/ Fig. 12.



COMPARISON OF SMOOTHING REQUIREMENTS OF THERMIONIC VALVE RECTIFIER AND VOLTAGE DOUBLER.

FIG. 12.

12. TEST QUESTIONS.

1. Enumerate the advantages of floating batteries as compared with a charge-discharge routine. What difficulty is created by the use of floating?

2. What is the permissible limit of noise measured at the busbars in a long line equipment room?

3. Do you consider that a disturbing frequency of 200 c/s would have the same interference value as an equal voltage of 800 c/s? Give reason.

4. Explain briefly the fundamental principle of operation of the carbon pile voltage regulator shown in Fig. 3.

5. Describe, with the aid of a diagram, the construction and operation of a copper oxide metal rectifier.

6. (a) It is desired to use the power derived from single phase A.C. mains for the anode supply to an amplifier. Assuming that suitable metal rectifiers and smoothing equipment are available, draw a circuit arrangement of the manner in which you would arrange the components to obtain the greatest efficiency, and describe how rectification is achieved.

   (b) What would be the frequency of the predominating ripple voltage in this particular circuit?

   (c) What would you consider to be a reasonable percentage of ripple voltage?

   (d) Neglecting losses in the smoothing equipment and transformer, what percentage efficiency would you expect in the rectifier unit?

END OF PAPER.