COMMONWEALTH OF AUSTRALIA

ENGINEERING BRANCH,
POSTMASTER-GENERAL'S DEPT.,
TREASURY GARDENS,
MELBOURNE. C.2.

COURSE OF TECHNICAL INSTRUCTION

RADIO I.

(Reissued 1952.)

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COURSE OF TECHNICAL INSTRUCTION.

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1. PRINCIPLES ESSENTIAL TO RADIO.

1.1 Radio Communication is the art of sending information from one point to a distant point or points by means of free electric waves. The study of radio, therefore, involves the consideration of the Free Electric Waves from the following aspects -

(i) Basic Nature,
(ii) Production,
(iii) Propagation, and
(iv) Reception.

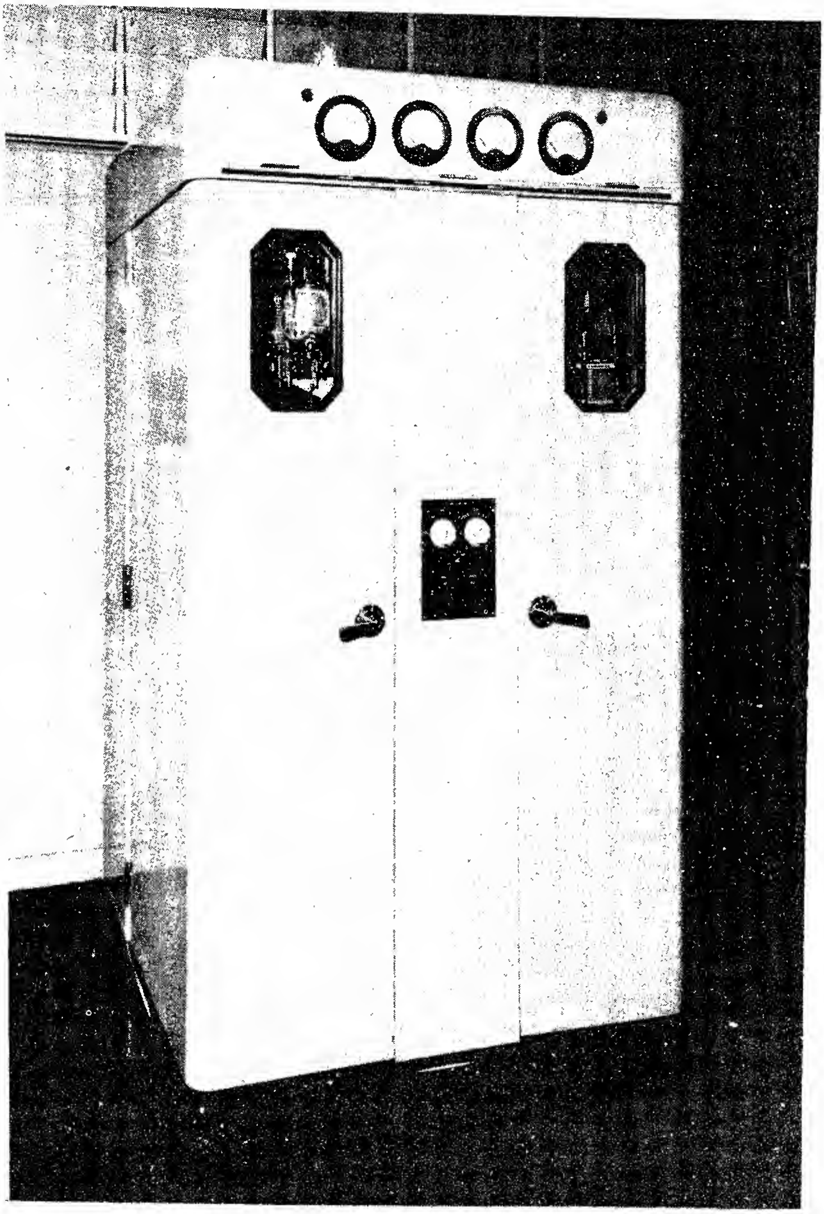
It is assumed that the reader has an understanding of the subject matter contained in Applied Electricity I, II and III. However, as a preliminary to the study of the nature, production, propagation and reception of Radio Waves, a short revision of the essential electric principles applicable to radio will be given here. Some of the symbols used in radio circuits are shown on pages 6 and 7.

1.2 Electron Theory. Much of the phenomena associated with radio communication may be explained in terms of electron behaviour or electron flow, and a brief review of the properties of electrons and atoms is desirable. The modern conception of electricity is based on the theory that all matter is built of atoms which, in turn, consist of various combinations of electric charges. An atom (being normally electrically neutral) is composed of equal positive and negative charges, the number of charges varying with the element concerned. Brief definitions of the basic units of interest are -

Electron. The Electron is the smallest particle whose mass and charge have been determined, and it is the particle that constitutes most of the flow of current encountered in thermionic valves. The Electron is a negatively charged particle.

Positron. The Positron is the particle having the same mass as an electron and carrying the same electric charge. The Positron is of opposite polarity from the electron, that is, positive.

Neutron. The Neutron is a particle without an electric charge but having a mass about 2,000 times greater than an electron or positron.



RADIO BROADCAST TRANSMITTER. (1 KILOWATT, MEDIUM WAVE).

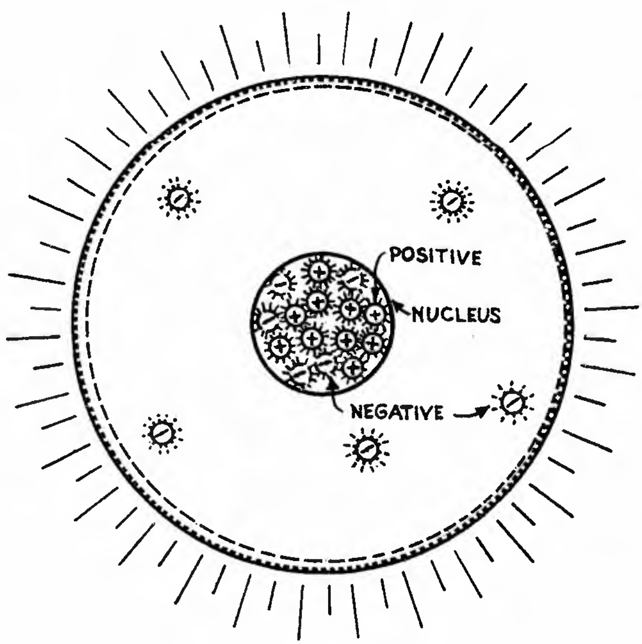
Proton. The Proton is a unit equivalent to a neutron associated with a positron, that is, it possesses the same magnitude of charge as an electron but has a mass about 2,000 times greater. In spite of the difference between their masses, equal numbers of protons and electrons are electrically neutral.

Ion. Another particle is called an Ion. An ion is usually created from a normal atom or molecule by the addition or removal of one or more electrons or protons from its structure. The mass of such ions is always much greater than that of the electron, and their charge is an integral multiple of the charge of the electron and is not restricted to negative sign.

Atoms are composed of various combinations of electrons and protons and, in order to obtain some idea of electronic behaviour, a sketch of the inner structure of an atom is shown in Fig. 1. The sign + indicates a unit charge of positive electricity, and the sign - indicates a unit charge of negative electricity. All the weight of the atom appears to reside in the central nucleus, which is made up of more positive charges than negative charges, so that a positive charge is always associated with a relatively large amount of matter. Electrons, on the other hand, are almost weightless and, therefore, can carry on a separate existance as disembodied electricity.

In ordinary uncharged matter, the electricity that electrons carry is neutralised by the net positive charge on the nucleus of the atom. Thus, although electrons are present in every substance, electric shocks are not necessarily experienced when the substance is touched by hand.

Fig. 1 is a diagrammatic sketch of the possible construction of an atom. This particular atom has a nucleus (of 10 protons and 5 electrons) and 5 planetary or free electrons. An atom of copper has a nucleus (of 64 protons and 35 electrons) and 29 planetary electrons.



ATOMS ARE MOSTLY SPACE!

Nucleus of this atom contains 10 protons (marked +) and 5 electrons (marked -), and there are 5 free electrons.

FIG. 1.

The atom may be likened to a solar system, the free electrons being the orbital planets and the nucleus being the Sun. As this planetary system is maintained in a state of equilibrium, there is an appreciable amount of energy expended in keeping the electrons bound to the nucleus. If, however, sufficient additional energy is imparted by some external means to the free electrons, they may be separated from their parent atoms to form a current flow. This additional energy may be imparted by several methods, four of which are -

(i) Chemical, as in the case of a battery.
(ii) Magnetic, as in the case of a dynamo.
(iii) Thermionic, as in the case of radio valves.
(iv) Light action, as in the case of photo-electric tubes.

Due to the reactions of the chemicals, etc., endeavouring to make up the initial deficiency of electrons, the current flow of electrons will continue as long as the influence of the external force is applied. This flow of electrons is termed a current. It is of interest to note that this conception of electron flow from negative to positive is at variance with the theory formulated by early scientists, who arbitrarily determined that current flow was from positive to negative. However, this distinction need not cause confusion if the latter point is remembered.

Further reference will be made to this subject when discussing emission.

(As a matter of interest, the effect of the "atomic bomb" depends on the release of this atomic energy by smashing the nucleus of an atom having a large number of orbital electrons, such as uranium.)

The velocities of the free electrons in a material depend on the temperature of the material. If an e.m.f. is applied to the two ends of a metallic conductor, these free electrons will drift towards the positive pole. This drift of electrons constitutes a current flow. The electrons collide with the metal atoms and impart energy to the atoms, causing them to vibrate more violently and thus heat the conductor. A particular electron need not pass completely from one end of the conductor to the other. It might displace an electron from an atom and the displaced electron, in turn, might displace an electron from another atom. Further reference will be made to this effect also when discussing emission.

1.3 In materials classed as insulators, it is assumed that only a comparatively few free electrons exist. The structure of the atoms of insulators is such that all electrons are securely bound to the nucleus, and the application of even a high voltage merely deforms the atom by stretching the bonds. An appreciable flow of current is not possible without an electron flow and as the electrons in an insulator are merely caused to change the path of their travels around the nucleus and do not move from their own planetary system, current will not readily flow through an insulator.

An electric current, therefore, consists of electrons in motion. Each electron has associated with it an electric field which radiates in all directions. This field is caused by the lines of force which also radiate from the electron in all directions. If an electron is set in motion, the associated lines of force move with it. Thus, an electric current has associated with it a moving electric field and, if the value or the direction of the current varies, the field will vary in strength and position according to the number and motion of the electrons constituting the current. If the current is alternating, the associated electric field will move to and fro in space, in accordance with the motion of the electrons.

An important secondary effect occurs when the current variation is rapid. Under these conditions, the lines of force cannot immediately follow (throughout their whole length) the motion of the electrons to which they belong, and a certain time lag elapses between the movement of the electron and the change in position of the associated lines of force. A mechanical analogy will illustrate this point. Consider a long rope fastened firmly at one end, and the other end held in the hand. If the rope is suddenly jerked, the portion of the rope which is held in the hand will follow the movement of the hand exactly.

A little way along, however, the jerk in the rope will take place a little after the motion of the hand, and the farther the distance away, the longer the time elapsing before that particular portion of the rope is affected. The ultimate effect is to produce a ripple which travels along the rope to the far end. This ripple occurs only when the rope is jerked. If the hand is moved slowly up and down, the rope is able to follow the motion completely throughout its whole length and no ripple will result.

There is another important difference between the two cases which should be noted. If the hand is moved slowly up and down, the whole rope moves and the actual movement grows less and less as the distance decreases, until, at a short distance away, there is no appreciable motion. If the rope is jerked, however, the ripple which travels along the rope is nearly as large at the far end as it was at the beginning.

A similar phenomenon takes place when an electron is set in motion. If the motion is slow, the electric field follows the electron faithfully and can only be detected at a short distance, the strength (according to the law of inverse squares) being inversely proportional to the square of the distance. The ordinary electric field of a condenser or coil falls off in this manner. When the motion is rapid, however, ripples are formed in the lines of force associated with electrons, and these ripples travel outwards for appreciable distances with little reduction in strength, the field strength being inversely proportional to the distance. As Wireless Telephony, Telegraphy and Television are dependent on these ripples or electromagnetic waves, the immediate concern of this book is how these ripples or electromagnetic waves may be produced and controlled. These first papers deal with this subject.

1.4 Alternating Currents. The principles of alternating current are dealt with in detail in Applied Electricity, which should be studied in conjunction with these papers. As Radio is a development of these principles, they must be understood if a proper appreciation of Radio is to be attained.

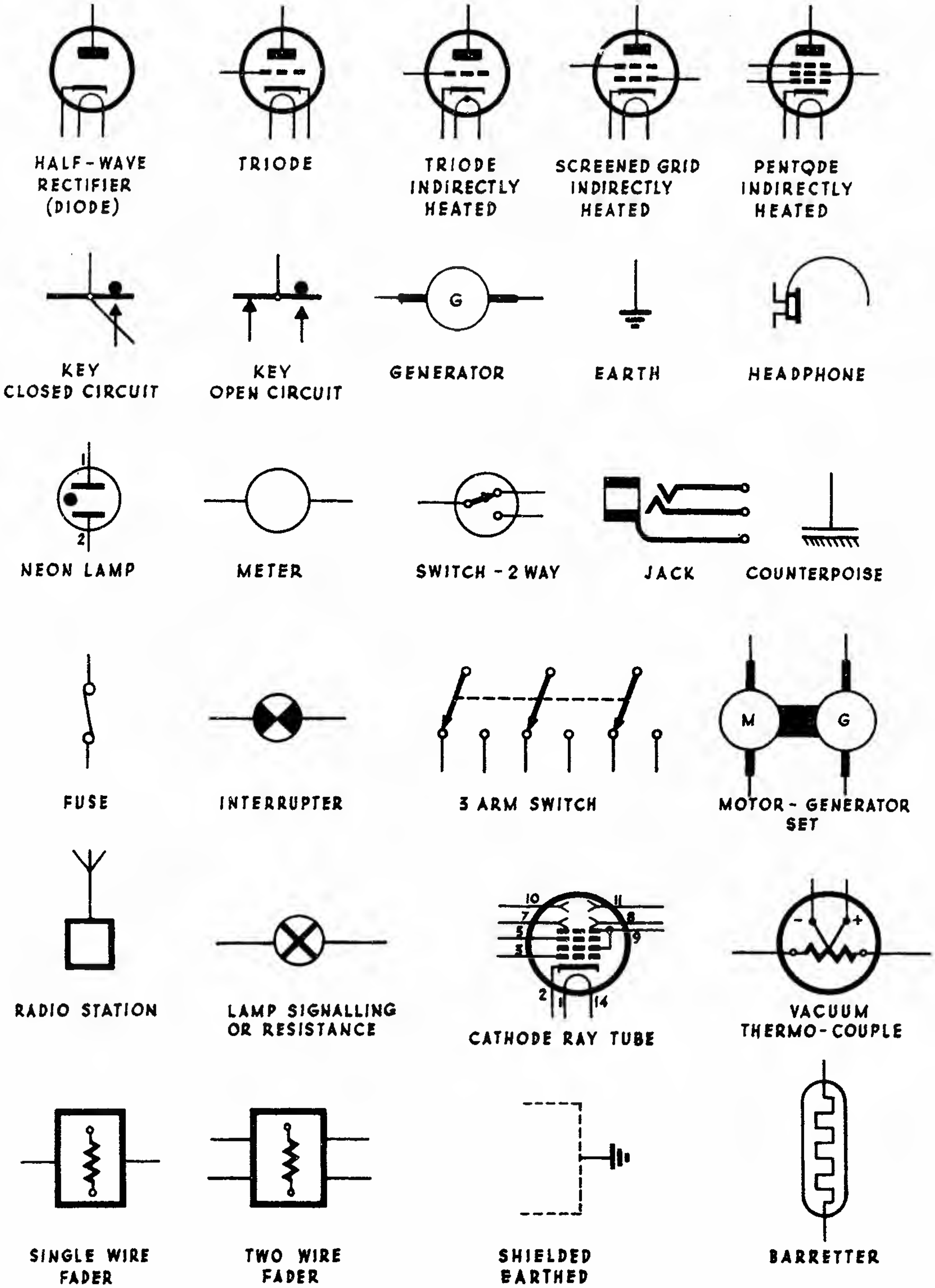
2. RADIO CIRCUIT ELEMENTS.

2.1 Inductance. Inductance is the property of a circuit by virtue of which self-induction occurs. This property is dependent on the fact that, if a current is passed through a wire, a magnetic field is set up round the wire. If the wire is made into a coil, the field becomes stronger. This field increases or decreases in direct proportion to the change in the current. The RATIO OF THE CHANGE IN FLUX to the CHANGE IN CURRENT has a constant value, known as the INDUCTANCE of the coil.

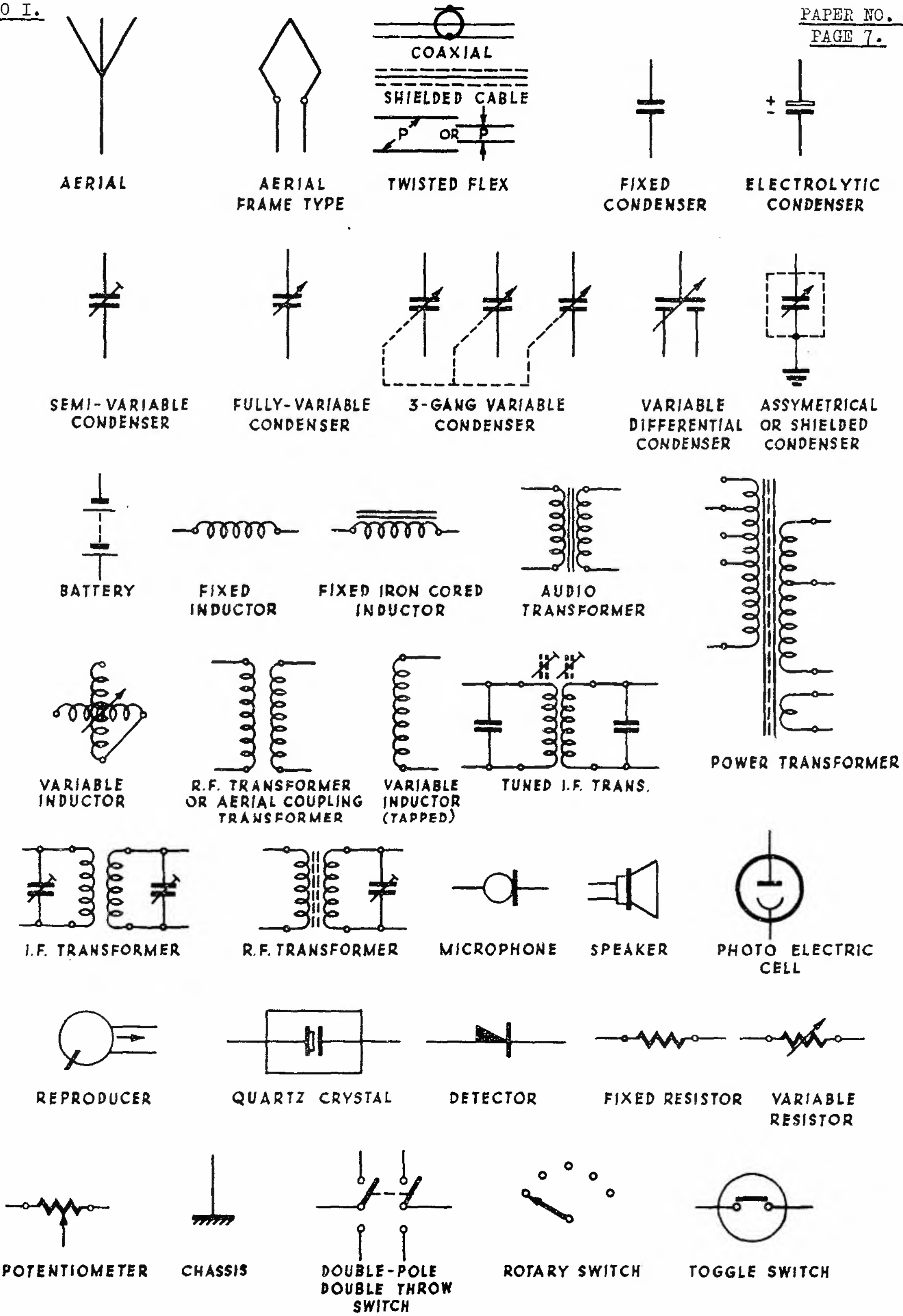
2.2 It should be stressed that, whenever a current passes through a coil, it causes a magnetic field around the coil; that the strength of the field varies as the current varies; and that the direction of the field is reversed if the direction of current flow is reversed. As explained in other books of the Course of Technical Instruction, the converse is also true, namely, that if a magnetic field passes through a coil, an electromotive force is induced in the coil; that if the applied field varies, the induced voltage varies; and that if the direction of the field is reversed, the direction of the current produced by the induced voltage is reversed. This phenomenon provides the explanation of many electrical effects. It serves, in the present instance, to give some understanding of that valuable property of coils - Self-Inductance. When an alternating current flows through a coil of many turns of wire, the field around the coil will increase and decrease, first in one direction and then in the other direction. The varying field around the coil, however, will induce a varying e.m.f. in the coil, and the current produced by this induced e.m.f. will always be in the opposite direction from the current originally passed through the wire. This self-induction of the coil, therefore, tends to prevent any change in the current flowing through it and limits the amount of alternating current flowing. This effect can be considered as electrical inertia.

Unit of Inductance. The unit of self-inductance is the Henry. A coil has a self-inductance of 1 henry when a rate of current change of 1 ampere per second causes an induced voltage of 1 volt. This basic unit is generally used with iron-core coils (as in power-supply filter circuits), but is too large for convenience in many radio applications.

/ Radio.



RADIO SYMBOLS.



RADIO SYMBOLS

Therefore, smaller units are also used. These units are the millihenry (mH), equal to one-thousandth henry, and the microhenry (μH), equal to one-millionth henry. Stated generally, the self-inductance of a coil is inversely proportional to the reluctance of its magnetic circuit and is proportional to the square of the number of turns. If the magnetic circuit is a closed iron core, the inductance value might be several thousand times what it would be for the same coil without the iron core, the reluctance being that much less than with an air-core. If the number of turns is doubled, the inductance would be four times as great.

Inductances in Series and in Parallel. Coils may be connected in series, in parallel or in series/parallel. If connected in series, the total inductance is increased (provided the magnetic fields do not link), just as the total resistance is increased when resistances are in series. With the same reservation, the total inductance of coils connected in parallel is reduced, just as the total resistance is reduced when resistors are connected in parallel. Correspondingly, coils may be connected in series/parallel combinations. The equations for inductances in series, in parallel and in series/parallel are the same as those for resistances with the proper inductance values substituted for resistance values.

Magnetic Energy Store (W). The tendency of coils to prevent change in current flow gives them the ability to store energy. This energy storage is proportional to the inductance of the coil and to the square of the current.

$$\text{Energy stored in coil} = \frac{LI^2}{2}$$

where the energy is in joules or watt seconds, L is the inductance in henrys and I is the current in amperes. This property is of particular importance in the filter systems used for radio transmitter and receiver power supplies.

Inductive Reactance (X_L). A coil tends to limit the amount of current which an alternating voltage causes to flow. An important fact is that a given coil with a fixed amount of inductance will retard the flow of a high frequency alternating current more than a low frequency current. The characteristic of a coil in retarding an alternating current flow depends on the inductance of the coil and on the frequency of the current. This combined effect of frequency and inductance in coils is termed reactance or inductive reactance (X_L).

The inductive reactance formula is -

$$X_L = 2\pi fL$$

where X_L = the inductive reactance in ohms,
f = the frequency in c/s, and
L = the inductance in henrys.

From this, it is evident that inductive reactance is directly proportional to frequency and also to the value of inductance.

Self-Inductance or Self-Induction. Self-Inductance is that property of electrical circuits which opposes a change in the current flowing. The opposition to the change in current manifests itself in the form of a back e.m.f. that is developed when the current is changed. This opposition is present only when the field is varying and is termed self-induction, because it is an e.m.f. induced in a conductor by its own moving magnetic field. A circuit is said to have an inductance of 1 henry if it develops a back e.m.f. of 1 volt when the current through it changes at the rate of 1 ampere per second.

Because of the manner in which the back e.m.f. in a circuit is related to the time rate of change of the magnetic field, inductance is related to the number of flux linkages per unit current in the circuit. A flux linkage represents one flux line encircling the circuit current once. Flux linkages may be fractional as well as integral. The relation between inductance, total flux linkages and current is given by -

/ Inductance

$$\text{Inductance (L, in henrys)} = \frac{\text{flux linkages}}{\text{current (amperes) producing the flux}} \times 10^{-8}$$

Mutual Inductance. In general, whenever two coils are near one another, a part of the flux produced by a current in one coil will pass through the other coil. If the current in the first coil changes, the flux through the second coil will change and an e.m.f. will be induced in the second coil. The magnitude of this induced e.m.f. will depend on the rate of change of current in the first coil, the number of turns of wire in each coil, the geometry of the coils and, perhaps, on other factors. Mutual inductance may be defined in terms of the number of flux linkages in the second coil per unit current in the first coil, or vice versa. In practical units the relation is -

$$M = \frac{\text{Flux linkages in the second coil produced by current in the first coil}}{\text{Current in the first coil}} \times 10^{-8}$$

and vice versa, where M is the mutual inductance in henrys.

Mutual inductance may also be defined as the voltage induced in the second circuit when the current in the first circuit is changing at unit rate.

Coefficient of Coupling. The maximum value of mutual inductance between two coils of inductance L_1 and L_2 is $\sqrt{L_1L_2}$, which occurs when all the flux of one coil links with all the turns of the other coil. The ratio of the mutual inductance actually present to the maximum possible value is called "the coefficient of coupling" (K), and is written as -

$$K = \frac{M}{\sqrt{L_1L_2}}$$

K is a dimensionless quantity having a maximum value of 1.

Combinations of Inductances involving Mutual Inductances. The total self-inductance of combinations of inductances depends upon the self- and mutual inductances involved. The total inductance resulting from two inductances in series is -

Aiding ... ... $L_1 + L_2 + 2M$

Opposing ... ... $L_1 + L_2 - 2M$

The total inductance resulting from two inductances in parallel is -

Aiding ... ... $\dfrac{L_1L_2 - M^2}{L_1 + L_2 - 2M}$

Opposing ... ... $\dfrac{L_1L_2 - M^2}{L_1 + L_2 + 2M}$

where L_1 and L_2 are the self-inductances of the two coils and M is the mutual inductance between them. In general, the total self-inductance of a number of inductances in series is the sum of the self-inductances of all the components plus the algebraic sum of the mutual inductances of each one of the component parts to all the other parts.

Condensers and Capacity. Condensers form as important a part as coils in radio circuits and, in fact, condensers and coils are generally associated. The condenser, as explained in Applied Electricity I, consists essentially of two or more metal plates separated by a thin layer of some insulating medium from a second similar plate or set of plates.

/ The

The insulating medium between the metal elements of the condenser is termed the dielectric. Unvarying direct current cannot flow through a condenser because of the insulation between the plates, but a steady voltage applied to the terminals of such a condenser will cause it to become charged. The effect, to return to a discussion of electrons, is simply that one element of the condenser is provided with an excess of electrons, thus becoming negatively charged, while the other plate suffers a deficiency of electrons and is, therefore, positively charged. If the charging voltage is removed and the two elements of the condenser are jointed with a conductor, a flow of electrons will take place from the negative to the positive plate. In other words, a current would flow.

Capacitance. If two similar parallel conductors, separated by an insulating medium, are each charged with the same quantity of electricity but of opposite sign, a difference of potential will be set up between the conductors. Such a combination is said to possess "capacitance," and is known as a condenser or capacitor. Capacitance may thus be defined as the ratio of the charge on one of the conductors to the difference of potential between them. Capacitance should not be taken to mean "the amount of electricity that a condenser can hold," but rather "the amount of electricity that a condenser requires to bring its two conductors to unit difference of potential." Capacity depends on the geometry of the conductors and the nature of the dielectric between them.

Capacity of a Condenser. The number of electrons necessary to establish a difference of potential between two plates is a measure of the capacity of the condenser thus formed. The defining relation for capacity may thus be written -

$$C = \frac{Q}{E}$$

where C is the capacity in farads, Q is the charge in coulombs and E is the potential in volts.

From this, $Q = CE$

Capacity may also be defined in terms of the energy stored in the electrostatic field about the condenser plates, thus -

$$W = \frac{CE^2}{2}$$

where W is the energy stored in watt-seconds (joules), C is the capacity in farads and E is the potential in volts. Capacity has the dimensions of length when expressed in electrostatic units, thus the capacities of similar condensers are directly proportional to their linear dimensions.

Farad. The unit of electrostatic capacity is called the "Farad." A condenser has a capacity of 1 farad when a charge of 1 coulomb will establish a potential difference of 1 volt between its plates. This unit is too large for practical use and submultiples of the farad, namely, microfarad (one-millionth of a farad) and micro-microfarad (one millionth of a microfarad), are generally used.

Dielectrics. The insulating medium that separates the plates of a condenser is known as the dielectric, and plays an important role in determining the characteristics of the condenser. The presence of a dielectric, other than a vacuum, raises the capacity of the condenser (in comparison to its capacity in the absence of the dielectric) by a factor known as the "Dielectric Constant" or "Specific Inductive Capacity" (symbol "k"). This characteristic determines the quantity of charge which a given separation and area of plates will accumulate for a given voltage. When the air dielectric in a variable condenser is replaced with a fluid dielectric, the maximum and minimum capacitance values of the condenser are multiplied by "k" and the "sparking" potential is increased.

Fluid dielectrics repair themselves after a breakdown unless an arc is maintained which carbonises the oil. Dry oil is a good dielectric with low losses. When solid dielectric is used, it should be borne in mind that dielectric strength (breakdown voltage) becomes lower as temperature rises. Breakdown is a function of time as well as voltage. A condenser, which will not break down under several thousand volts for a few seconds, may break down when connected to 2000 volts for some time.

Condenser Losses and their Representation. A perfect condenser, when discharged, gives up all the electrical energy that was supplied to it in charging. Actual condensers never realise this ideal perfectly but, rather, dissipate some of the energy delivered to them. Most of the loss in ordinary condensers occurs in the dielectric. Other ways by which energy can be lost in a condenser are from the resistance of the leads and the metal plates, from leakage resistance between the plates and as a result of corona.

The merit of a condenser from the point of view of freedom from losses is usually expressed in terms of the power factor or phase angle. The power factor represents the fraction of the input volt-amperes that is dissipated in the condenser, while the phase angle is the angle by which the current flowing into the condenser fails to be 90° out of phase with the applied voltage. When the losses are low, which is generally the case, the phase angle expressed in radians is equal to the power factor. The power factor (or phase angle) is a ratio representing the fraction of the input volt-amperes that is dissipated in the condenser. For a given type of condenser, the power factor tends to be independent of the applied voltage and of the condenser size, etc., and with ordinary dielectrics is substantially constant over wide frequency ranges.

The effect of the condenser losses on the circuit, in which the condenser is connected, can be obtained by replacing the actual condenser by the combination of a perfect condenser of the same capacity with a resistance in series (Fig. 2), or a resistance in parallel (Fig. 3). The value of the series (or shunt) resistance is so selected that the power factor of the combination of perfect condenser with its associated resistance is the same as the power factor of the actual condenser. The values of the resistances are -

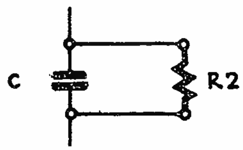
$$\text{Series resistance} = R_1 = \frac{\text{Power factor}}{2\pi fC}$$

$$\text{Shunt resistance} = R_2 = \frac{1}{(2\pi fC)\ (\text{power factor})}$$

where R is in ohms, f is in c/s and C is in farads.

EQUIVALENT CONDENSER LOSS.

FIG. 2.



ALTERNATIVE CONDENSER LOSS.

FIG. 3.

Capacitive Reactance (X_c). When a direct current voltage is applied to a condenser, it will cause a sudden charging current in one direction, but an alternating voltage will result in the condenser becoming charged first in one direction and then in the other. This rapidly changing charging current is actually the equivalent of an alternating current flow through the condenser. Many of the condensers in radio circuits are used to allow an alternating current to flow through some portion of the circuit but, at the same time, prevent the flow of any direct current.

Condensers, however, do not permit alternating currents to flow without offering some opposition, and the term "Capacitive Reactance" is used to describe this opposition. Condensers have a reactance which is inversely proportional to the capacitance and to the frequency of the applied voltage. The formula for capacitive reactance is -

$$X_c = \frac{1}{2\pi fC}$$

where X_c is the capacitive reactance in ohms, f is the frequency in c/s, and C is the condenser capacitance in farads.

Where the capacitance is in microfarads (μF), as it is in most practical cases, the formula becomes -

$$X_c = \frac{10^6}{2\pi fC}$$

Condensers in Series and Parallel. Condensers may be connected in series or in parallel. Connecting condensers in parallel makes the total capacitance greater. The equivalent capacity of condensers connected in parallel is the sum of the capacities of the several condensers so connected, thus -

$$C = C_1 + C_2 + C_3$$

The equivalent capacity of condensers connected in series is expressed by the following formula -

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

When two condensers are connected in series, the following expression can be used to determine the resultant capacity value -

$$C = \frac{C_1 C_2}{C_1 + C_2}$$

When the net capacitance of a series/parallel combination is to be found, the capacitance of the series groups can be determined separately and then added in parallel combination

Connecting condensers in series increases the breakdown voltage of the combination but decreases the capacity available.

Condensers of identical capacitance are most effectively connected in series for this purpose. Voltage tends to divide across series condensers in inverse proportion to the capacity, so that, if the condensers are of equal voltage rating, the smaller of two series condensers will break down first. Before selecting filter condensers, the operating conditions, voltage peaks and R.M.S. values should be carefully considered.

/ Energy

Energy Stored in Condensers (W). Magnetic energy, as explained previously, is stored in coils and, likewise, energy is stored in condensers. In the case of the coil, however, the amount of energy is associated with current value but, in the case of a condenser, the amount of energy is associated with the e.m.f. and is termed electrostatic energy. The amount of energy stored by a condenser is given by this equation -

$$\text{Energy stored in condenser} = \frac{CE^2}{2}$$

where the energy is in joules (or watt-seconds), C is the capacity in farads, and E is the e.m.f. in volts. When the capacitance is in microfarads, as is usual in practical cases, the equation is -

$$\text{Energy stored} = \frac{CE^2}{2 \times 10^6} \text{ joules.}$$

This energy storage relation for condensers, like the energy storage relation for coils, is of importance in filter circuits.

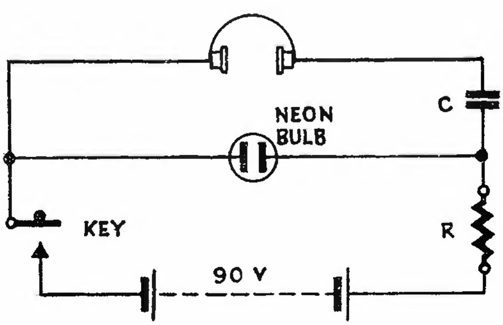
Resistance-Capacitance Time Constant (RC). If a charged condenser had infinite resistance between its plates, it would hold a charge indefinitely at its initial value. However, since this ideal condition does not apply to practical condensers, the charge is gradually lost. High class condensers have a very high "leakage resistance," however, and hold a charge for a long time.

In a circuit containing only capacitance and resistance, the time required for the potential difference between the charged plates of a condenser to fall to a definite percentage of its initial value is determined by the capacitance of the condenser and the value of the resistance. The relation is of practical importance in many radio circuit applications, such as the time delay with automatic volume control, resistance-capacitance filters, etc. For the voltage to fall to 37 per cent. (0.37) of its initial value -

$$t = RC$$

where t is the time in microseconds (millionths of a second), R is the resistance in ohms and C is the capacitance in microfarads. RC should be divided by 1 million to give the answer in seconds. This is called the time constant of the combination. The time required for the voltage to fall to one-tenth (10 per cent) of its initial value can be found by multiplying RC, as given above, by 2.4 when time constant, t (for 90 per cent. fall in voltage) $= 2.4 \frac{RC}{10^6}$, t being in seconds, R in ohms and C in µF.

A circuit diagram of an audio-frequency oscillator used for operating practice, which will serve to illustrate the effect of time constant on the output of a circuit, is shown in Fig. 4. The operation of the oscillator is such that, when the key is closed, condenser C charges slowly through the resistance R. When the condenser is charged, it discharges quickly across the neon bulb and, at the same time, causes a current to flow through the receivers. During the time the key is closed, there will be a series of charging and discharging of the condenser and a tone will be heard in the receivers. The frequency of the tone can be altered by changing the value of C or R, or both. For instance, if the value of R is reduced, the condenser will charged and discharge more quickly and the frequency will increase.



PRACTICE AUDIO OSCILLATOR.

FIG. 4.

/ If

If the value of R is increased, the rate of charge and discharge of C will be decreased, with a consequent reduction in frequency.

2.3 Impedance. The functioning of most radio equipment depends upon three important properties of electrical circuits and apparatus -

(i) Resistance, (ii) Inductance, (iii) Capacity.

Resistors, coils and condensers are usually built to have as much as possible of one of these properties with as little as possible of the other two. These "lumped" properties can then be utilised in a circuit to produce the required effect on the current and voltage distribution. However, in all coils and condensers there is a combination of all electrical properties, and not just the one property for which the apparatus is used. For this reason, most design work is somewhat of a compromise. Every coil and transformer winding has resistance and distributed capacity between the turns, in addition to inductance, and every condenser has some resistance and more or less inductance. Resistors, as another example, often have appreciable inductance and distributed capacity.

Applied Electricity III deals in detail with the varied combinations of these properties and a brief resumé only will be given here.

The importance of these characteristics is realised when an attempt is made to apply Ohm's Law to circuits in which alternating current flows. If inductances did not have any resistance, the current through the coil would be equal to the voltage divided by the reactance. However, the coil will have resistance, and this resistance will act with the reactance in limiting the current flow. The combined effect of the resistance and reactance is termed "impedance" in the case of both coils and condensers. The symbol for impedance is "Z" and it is computed from this formula -

$$Z = \sqrt{R^2 + X^2}$$

Where R is the resistance in ohms and X is the reactance of the coil in ohms. Z is also expressed in ohms. Ohm's Law for alternating current circuits then becomes -

$$I = \frac{E}{Z}; \quad Z = \frac{E}{I}; \quad E = IZ.$$

For a given coil and condenser, the inductive reactance increases with frequency, the capacitive reactance decreases with frequency, X_L is conventionally considered positive and X_c is negative. Thus, in a circuit containing both X_L and X_c, the net reactance will be the difference between that of the inductance and that of the condenser. In most cases, the resistance of a high class condenser is such that it is negligible compared with the reactance, and Z is not computed unless the conditions warrant this step.

Other formulae frequently used are -

Inductance and Resistance in Series.

$$Z = \sqrt{R^2 + X_L^2}$$

Capacity and Resistance in Series.

$$Z = \sqrt{R^2 + X_c^2}$$

Inductance, Capacity and Resistance in Series.

$$Z = \sqrt{R^2 + (X_L - X_c)^2}$$

/ Inductance

Inductance and Resistance in Parallel.

$$Z = \frac{\omega L R}{\sqrt{R^2 + \omega^2 L^2}}$$

Capacity and Resistance in Parallel.

$$Z = \frac{R}{\sqrt{1 + \omega^2 C^2 R^2}}$$

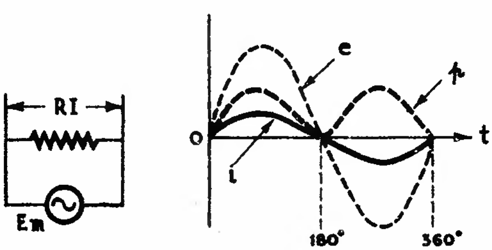
In the general parallel case, the inductance and the resistance (effective resistance of coil) are considered as being in series with the condenser shunted across the combination. The impedance in this case approximates -

$$Z = \frac{R + \omega^2 L^2}{(1 - \omega^2 LC)^2 + \omega^2 C^2 R^2}$$

Phase Relationships.

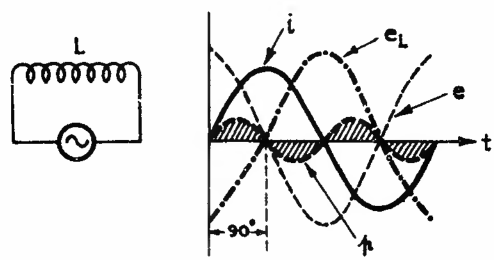
In Non-Reactive Circuits. When an alternating current voltage is applied to a non-inductive circuit, the current value is obtained by Ohm's Law ($E = RI$). At zero voltage, therefore, the current is zero, and, at maximum voltage, the current is also at a maximum. In other words, the current and the applied electromotive force are in phase. This condition is in Fig. 5, which shows curves for the instantaneous e.m.f. (e), current (i) and power (p).

In Inductive Circuits. In the case of a circuit containing pure inductance, it can be shown that the current and electromotive force are out of phase by 90°, with the current lagging the applied e.m.f. In other words, when the voltage across the inductance is a maximum and decreasing, the current through it is zero and starting to increase. The current leads the self-induced electromotive force by 90°. The lagging of the current is the result of the "inductance" or "self-induction" of the coil, which causes an e.m.f. (180° out of phase with the applied e.m.f.) to build up. Fig. 6 shows this effect, and it will be noticed that the current is a maximum when the two e.m.f's. are zero. (e_L = induced e.m.f.) The shaded portion of the figure is the instantaneous power, and it will be seen that, over one cycle, this is zero, that is, power cannot be dissipated by a pure inductance.



ALTERNATING CURRENT CIRCUIT
WITH RESISTANCE ONLY.

FIG. 5.



ALTERNATING CURRENT CIRCUIT
WITH INDUCTANCE ONLY.

FIG. 6.

In Capacitive Circuits. In this case (Fig. 7), the applied e.m.f. and consequent current are 90° out of phase. This time, however, the current leads the e.m.f. by 90°.

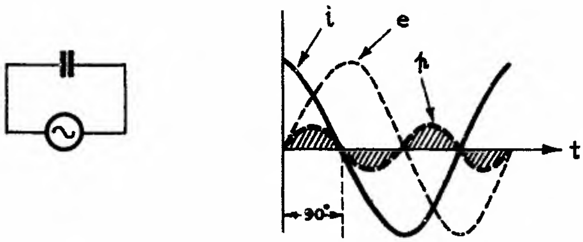


FIG. 7. ALTERNATING CURRENT CIRCUIT WITH CAPACITY ONLY.

Effect of Resistance. The effect of resistance on the phase angle in the last two cases is to reduce the angle to one less than 90°, depending on the magnitude of the resistance. It also reduces the value of current flowing for a given e.m.f. and causes a loss of power through dissipation equal to I^2R. Expressed mathematically, the angular difference may be obtained by any of the following relations -

$$\cos \theta = \frac{R}{Z}$$

$$\sin \theta = \frac{X}{Z}$$

$$\tan \theta = \frac{X}{R}$$

where θ = phase angle, R = resistance, X = reactance, and Z = impedance.

Power Factor. In the case of direct current circuits the current and voltage are always in phase and the power in the circuit is obtained from the product of the current and voltage. In the case of alternating current circuits, however, it has been mentioned that the current and voltage are not always in phase, and the power cannot be computed so simply, since allowance must be made for the fact that the maximum values of current and voltage are not always reached at the same instant. If the current and voltage are measured, the product represents the "apparent power" of the circuit. To obtain the true power in an alternating current circuit, this product is multiplied by the "power factor," which is a factor related to the phase angle. Numerically, the power factor is equal to $\frac{R}{Z}$, which, it will be recalled, is the cosine of the phase angle. Thus -

Power in alternating current circuit = $(E \times I \times \cos \theta)$ or $EI \cos \theta$.

Sine $\cos \theta$ is always less than 1, the power in an alternating current circuit is always less than the apparent power. The difference between the true and apparent power is known as "wattless power," as it is really ineffective in the circuit. While on the subject of alternating current circuits and power, mention must be made of effective average values, etc.

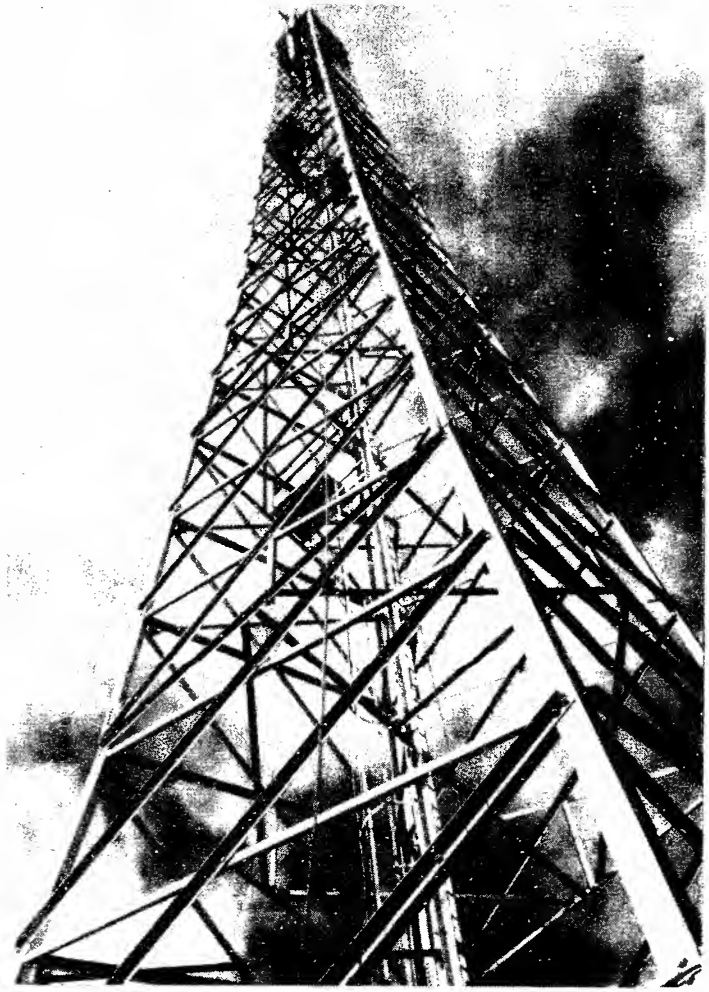
Average and Effective Values. The average value of an alternating current for a complete cycle is zero. Thus, the ordinary direct current instrument, which reads average values, will read zero when connected in an alternating current circuit. The usual meaning of average value as applied to alternating current is the average ordinate of a half-wave, and this can be shown to be equal to 0.637 of the maximum value. The average value of an alternating current or voltage is of minor importance, the Effective or Root Mean Square (R.M.S.) value being important.

THE EFFECTIVE VALUE OF AN ALTERNATING CURRENT IS THAT VALUE OF DIRECT CURRENT WHICH WILL PRODUCE THE SAME HEATING EFFECT.

This is also known as the R.M.S. value and is the square root of the mean value of the squares of the instantaneous values taken over one complete cycle. The heating effect would be the same as in direct current circuits, that is, proportional to the square of the current. It is of interest to note that, due to squaring, the effect of the negative half-cycles of current is considered, which is not the case with the true average value. The effective value is found from the maximum value as follows -

$$E_{eff} = 0.707 \text{ maximum } E \ (\text{or } I_{eff.} = 0.707 \text{ I max.})$$

/ SELF-SUPPORTING.



SELF-SUPPORTING RADIO MAST. JOLIMONT, VICTORIA.

The average value, however, is of importance when alternating current has been rectified to direct current, so the three relationships should be remembered. The relationships are -

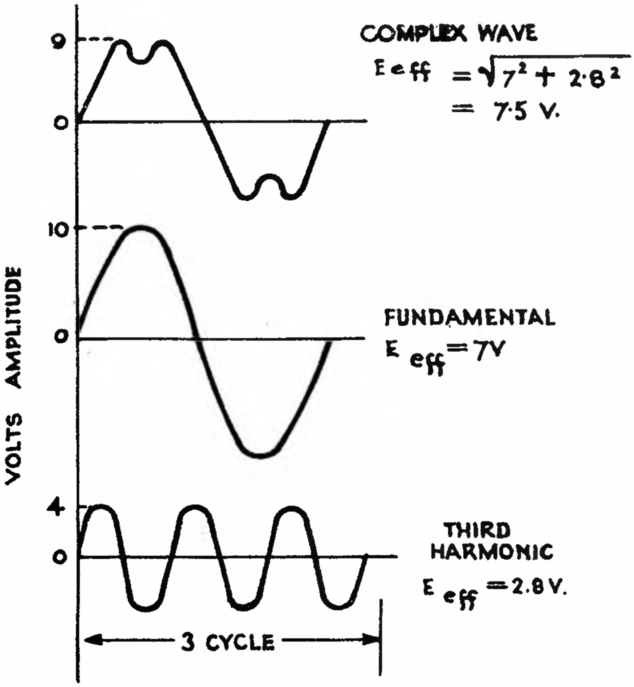
$$E_{max.} = (E_{R.M.S.} \times 1.414) = (E_{av.} \times 1.57)$$

$$E_{R.M.S.} = (E_{max.} \times 0.707) = (E_{av.} \times 1.11)$$

$$E_{av.} = (E_{max.} \times 0.636) = (E_{R.M.S.} \times 0.9)$$

The current relationships are the same as those given above for voltage. The usual alternating current ammeter or voltmeter gives a direct reading of the effective or R.M.S. value of current or voltage. A direct current ammeter in the anode circuit of a thermionic valve approximates the average value of rectified anode current. Maximum values can be measured by a peak thermionic valve voltmeter.

Complex Alternating Current Waves. Alternating currents, having the ideal sine wave form just described, are rarely found in radio circuits, although waves closely approximating the perfectly sinusoidal wave can be generated with laboratory-type equipment. Even the current in power mains is somewhat non-sinusoidal, although it can be considered sinusoidal for most purposes. In the usual case, such a current actually has components of two or more frequencies integrally related, as shown in Fig. 8. The lowest and principal frequency is the fundamental. The additional frequencies are whole-number multiples of the fundamental frequency (twice, three times, etc.) and are called harmonics. A harmonic of double frequency is the second harmonic, one of triple frequency is the third harmonic, etc. Although the wave resulting from the combination is non-sinusoidal, the wave form of each component taken separately has the sine wave form.



COMPLEX WAVE AND ITS SINE WAVE COMPONENTS.

FIG. 8.

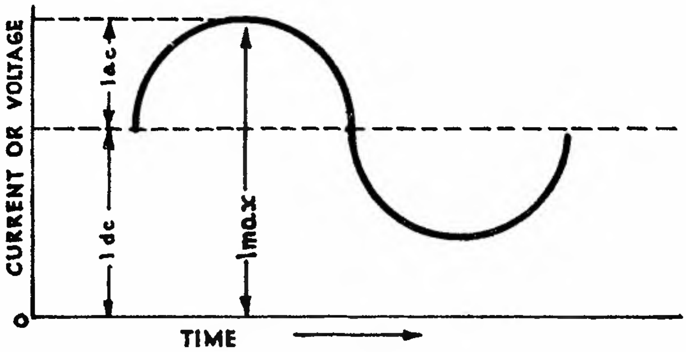
The effective value of the current or voltage for such a complex wave will not be the same as for a pure sine wave of the same maximum value. Instead, the effective value for the complex wave will be equal to the square root of the sum of the squares of the effective values of the individual frequency components, that is -

$$E = \sqrt{E_1^2 + E_2^2 + E_3^2}$$

where E is the effective value of the complex wave, and E_1, E_2, etc., are the effective values of the fundamental and harmonics. The same relation also applies where currents of different frequencies not harmonically related flow in the same circuit.

/ Combining

Combining Alternating and Direct Currents. There are many practical instances of simultaneous flow of alternating and direct currents in a circuit. When this occurs there is a pulsating current, and it is said that an alternating current is superimposed on a direct current. As shown in Fig. 9, the maximum value is equal to the direct current value plus the alternating current maximum, whilst the minimum value (on the negative alternating current cycle) is the difference between the direct current and the maximum alternating current values. If a direct current ammeter is used to measure the current, only the average or direct current component will be indicated. An alternating current meter, however, will show the effective value of the combination. This effective value is not the simple arithmetical sum of the effective value of the alternating current and the direct current, but is equal to the square root of the sum of the effective alternating current value squared and the direct current value squared, that is -



PULSATING CURRENT COMPOSED OF ALTERNATING CURRENT SUPERIMPOSED ON DIRECT CURRENT.

FIG. 9.

$$I = \sqrt{I_{ac}^{2} + I_{dc}^{2}}$$

where I_{ac} is the effective value of the alternating current component, I is the effective value of the combination and I_{dc} is the average (direct current) value of the combination. If the alternating current component is of sine wave form, the maximum value will be the effective value, as determined above, multiplied by 1.414. If the alternating current component is not sinusoidal, the maximum value will have a different ratio from the effective value, depending on its wave form, as discussed in the preceding section.

The Power from Pulsating Current. In a resistance circuit, the power developed by a pulsating current will be I^2R watts, I being the effective or R.M.S. value of the current and R the resistance of the circuit in ohms. In the special case of sine wave alternating current having maximum value equal to the direct current, which represents 100 per cent. modulation of the direct current by the alternating current, the effective value of the alternating current component is 0.707 (approximately 70 per cent.) of its maximum alternating current value and likewise of the direct current value. If the two maximum values are each 1 ampere -

$$I = \sqrt{1^2 + 0.707^2} \text{ or } \sqrt{1.5} \text{ which } = 1.226 \text{ amperes.}$$

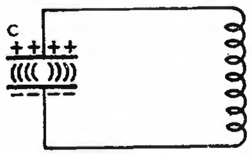
$$\text{and } P = I^2R \text{ or } 1.5R.$$

Hence, when sine wave alternating current is superimposed on direct current in a resistance circuit, the average power is increased 50 per cent. if the maximum value of the alternating current component is equal to the direct current component. If the alternating current is not sinusoidal, the power increase will be greater or less, depending on the alternating current wave form. This point is dealt with further in connection with speech modulation.

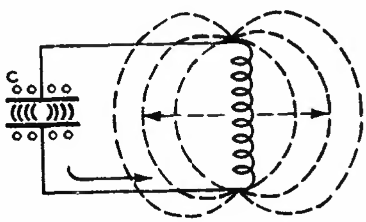
To summarise, a flow of electrons through a wire constitutes an electric current, and this current will, under certain conditions, give rise to electric and magnetic effects as changes in the current flow take place. The inclusion of an inductance or coil in an alternating or direct current circuit will tend to prevent any change in the current flowing in the circuit, because of the magnetic field around the inductance. This field varies in strength with each variation of current flow. The field around a coil can also link with the turns of a second coil, thus inducing voltages in the second coil - voltages which vary in accordance with the changes in the original current flow. Further, a condenser can be charged by an applied voltage, and the energy represented by this charge can cause a current to flow in any conductor which is connected across the condenser terminals. Lastly, in an alternating current circuit, inductance causes the current to lag the voltage, while capacity causes the current to lead the voltage.

An examination of the circuit diagram of almost any piece of radio equipment will reveal one or more combinations of coil and condenser, that is, of inductive reactance and capacitive reactance. In order to be able to read and understand radio circuits, therefore, it is necessary to understand the principles of inductance, capacity and resistance. The following paragraphs explain how these properties form the tuned circuit.

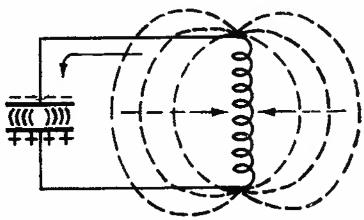
2.4 The Tuned Circuit, an Important Radio Element. Assume that a condenser C and coil L are connected, as shown in Fig. 10, to form a simple tuned or oscillatory circuit, and that the condenser is initially charged as indicated in (a), one plate having a surplus of electrons and, therefore, being negative, while the other plate, being correspondingly deficient in electrons, is positive. When the condenser plates are connected together through the coil L, there will be a flow of current as shown by the arrow in (b). The rate of flow of current will be retarded by the inductive reactance of the coil and the discharge of the condenser will not be instantaneous, even though the velocity of flow is constant. As the current continues to flow from the condenser into the coil, the energy initially stored in the condenser as an electrostatic field will be stored in the electromagnetic field of the coil. When most of the energy in the circuit has been stored in this field, the lines of force about the coil begin to collapse, and thus cause a continued flow of current through the circuit, the flow being in the same direction as the initial current. This operation again charges the condenser, but in opposite polarity to the initial charge. Then, when all the energy has been stored again in the condenser, the sequence is repeated in the opposite direction. This process is one of "Oscillation." Because an external driving source is not applied to the circuit after the initial condenser charge to sustain oscillation, these oscillations are known as "free oscillations." During one complete cycle, the energy is twice stored in the condenser and twice in the coil, and there is one reversal in the direction of current flow. This represents a complete cycle of alternating current. The process would continue indefinitely if the circuit possessed only inductance and capacitance, but all circuits contain some resistance. Therefore, during each cycle, a part of the energy will be dissipated in the resistance as heat. Each cycle will be of less amplitude than the preceding one, and the process will finally stop because there is no longer energy to sustain it, the waves being said to be damped. In practical oscillatory circuits, as will be shown later, this damping caused by resistance is overcome by continuously supplying energy from a generator to replace the energy dissipated.



(a) Electrostatic Energy Stored in a Condenser.



(b) Electromagnetic Energy Stored in Coil.



(c) Electromagnetic Field Collapses.

FIG. 10. BASIC TUNED CIRCUIT.

/ Natural

Natural Frequency of a Tuned Circuit. In such an oscillatory circuit, the larger the coil is made the greater will be its inductance and the longer will be the time required for the condenser to discharge. Likewise, the larger the condenser and the greater its capacitance, the longer it will take to charge or discharge. Since the velocity of the current flow is substantially constant, the circuit with the larger coil or condenser will take a longer period of time to complete a cycle of oscillation than a circuit where the inductance and capacitance are small, that is, the number of cycles per second will be greater as the inductance and capacitance values become smaller. Hence, the smaller the coil or condenser, or both, in the tuned circuit, the higher will be the frequency of oscillation.

The important practical aspect of all this is that, in a circuit containing capacitance, inductance and a low resistance value, the introduction of a pulse of electrical energy will cause an alternating current oscillation of a frequency determined mainly by the values of inductance and capacitance. This frequency is known as the "natural frequency" and is the frequency of free oscillation, that is, the frequency at which the circuit will oscillate, and is given by -

$$f = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$$

where L = inductance in henrys,
C = capacity in farads,
R = effective resistance of circuit, and
f = frequency in cycles per second.

The Aperiodic (or Non-Oscillatory) Circuit. If the quantity $\frac{R^2}{4L^2}$ in the formula for natural frequency is equal to or greater than $\frac{1}{LC}$, then free oscillations in the circuit are impossible, for when -

$$\frac{1}{LC} = \frac{R^2}{4L^2}$$

f will equal zero. The current in the circuit will not reverse its direction but will cease. The circuit is said to be in an "aperiodic" condition, that is, without period or oscillation. From the above, it may be seen that the value of resistance above which the circuit becomes aperiodic is given by -

$$R = 2\sqrt{\frac{L}{C}}$$

Forced Oscillation in a Tuned Circuit. For any combination of inductance and capacitance in a tuned circuit, there is one particular frequency at which maximum current will flow. It has been shown that the inductive reactance of the coil and capacitive reactance of the condenser are oppositely affected by frequency. Inductive reactance increases with frequency and capacitive reactance decreases as the frequency increases. In any combination of inductance and capacitance, if a constant alternating voltage is applied to the circuit, a constant alternating current is produced with an amplitude which does not vary, that is, oscillations will be forced in the circuit at the frequency of the applied voltage, and the oscillations produced are known as "forced oscillations." Energy losses in the circuit are supplied from the external source of power.

Natural and Resonant Frequencies Compared. It must be stressed that a circuit oscillates at its natural frequency only when its electrons are disturbed and left to return to rest. While such effects occur in radio, the natural frequency of free oscillation is usually of small importance, but the determination of the impedance value of various circuits to obtain maximum effect from the "driver" (generally a valve oscillator) is important.

/ In

In such cases, the oscillations are "forced" at a definite frequency whatever the constants of the circuit, and adjusting the circuit components for maximum current flow, or TUNING it, is simply arranging for the capacitive reaction to be equal and opposite to the inductive reaction, so that the circuit will be non-reactive or resonant.

While it is essential that the distinction between these two processes should be clear, it is interesting to note that the natural frequency of FREE oscillations is approximately equal to the resonant frequency of FORCED oscillations when the resistance is small, for if R is zero then -

$$f = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$$

becomes $f = \frac{1}{2\pi}\sqrt{\frac{1}{LC}}$

which simplifies to $\frac{1}{2\pi\sqrt{LC}}$

Note also that, while a circuit having a resistance equal to or greater than $2\sqrt{\frac{L}{C}}$ is APERIODIC or NON-OSCILLATORY as regards FREE oscillations, it still has a RESONANT frequency at which a driver will feed the maximum FORCED oscillations through it.

Calculation of Frequency of Circuit in Practical Units. In practice, the following formulae can be used to calculate either the natural or the resonant frequency of a radio circuit.

When L is in henrys and C is in farads -

$$f\ (c/s) = \frac{1}{2\pi\sqrt{LC}}$$

Or, when L is in μH and C is in $\mu\mu$F -

$$f\ (c/s) = \frac{10^9}{2\pi\sqrt{LC}}$$

and $f\ (kc/s) = \frac{10^6}{2\pi\sqrt{LC}}$

And, when L is in μH and C is in μF -

$$f\ (kc/s) = \frac{10^3}{2\pi\sqrt{LC}}$$

LC Constants. From this it may be seen that the product of L and C is a constant for a given frequency, and that the frequency of a resonant circuit varies inversely with the square root of the product of the inductance and capacitance. In other words, doubling both the capacitance and the inductance (giving a product of 4 times) would halve the frequency; or, reducing the capacitance by one-half and the inductance by one-half would double the frequency. Leaving the inductance fixed and reducing the capacitance to one-half would increase the frequency only 40 per cent. Reducing either the capacitance or the inductance to one-fourth and leaving the other fixed would double the frequency.

/ 3.

3. RESONANT CIRCUITS WHICH MAKE RADIO POSSIBLE.

3.1 In a circuit containing elements of inductance, capacitance and resistance in either series or parallel combination, a condition may be set up where the inductive and capacitive reactances are equal in magnitude. Since they are fundamentally 180° out of phase, this condition means that their reactive effects are equal and opposite and therefore cancel. Thus, a definition of Resonance may be given as "that condition in a circuit containing inductive reactance and capacitive reactance when the net reactance is zero". (In the parallel case there can be a slight departure from this, but, as normally applied to radio, this definition is satisfactory.)

It will be recalled that the reactive values are -

$$\text{Inductive reactance} = 2\pi fL \text{ or } \omega L \text{ ohms}$$

$$\text{Capacitive reactance} = \frac{1}{2\pi fC} \text{ or } \frac{1}{\omega C} \text{ ohms}$$

where f = frequency in c/s,
L = inductance in henrys,
C = capacity in farads, and
ω (omega) = $2\pi f$.

Thus, the resonant condition implies that -

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

$$\text{that is, } 2\pi fL = \frac{1}{2\pi fC}$$

Since the common variable is "f", it follows that there must be some value of "f" which will give the resonant condition. This value may be determined from -

$$\text{Resonant frequency } (f_r) = \frac{1}{2\pi\sqrt{LC}} \quad \text{(Very important formula!!)}$$

The frequency "f_r", at which the above condition is set up, is termed the "Resonant Frequency", but the behaviour of the two types of circuits (series and parallel) is entirely different when an alternating current at the resonant frequency is applied to them.

3.2 Series Resonant Circuit. Consider first the series circuit shown in Fig. 11. This circuit consists of a coil of inductance L, a resistance R and a condenser of capacity C connected in series with E, a source of alternating current potential. This combination gives a circuit in which the charge and discharge of a condenser and the rise and fall of a magnetic field are taking place simultaneously. It so happens that these two events tend to cancel the effects of each other. The energy required for the rising field is supplied at exactly the right instant by the discharging condenser, and the energy needed to recharge the condenser is supplied at exactly the right instant by the collapsing field.

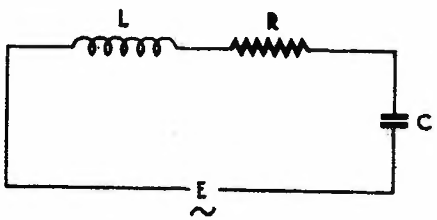


FIG. 11. SERIES CIRCUIT.

At low frequencies, the reactance value of C is large and the reactance value of L is small, and the magnetic field of L can only supply part of the energy needed for the condenser.

/ At

At high frequencies, the reactance value of L is large and the reactance value of C is small, and the discharge of C can only provide part of the energy needed to establish the field around L. There must be some intermediate frequency, however, at which these two reactance values are equal. At this particular frequency only, the charge on C and the field round L involve exactly equal amounts of energy.

Since L and C supply each other's needs exactly, the two jointly offer no reactance to the flow of current, which is therefore determined entirely by the value of R.

If alternating current at the resonant frequency is applied to the circuit of Fig. 11, the reactances cancel out, the circuit resistance becomes equal to R and the current will be a maximum limited only by R. R is the effective resistance of L plus any additional resistance in the circuit, the resistance of the condenser usually being negligible, that is -

$$I = \frac{E}{Z} = \frac{E}{R}$$

where E = applied voltage
R = impedance at resonance = effective resistance, and
I = current.

Since the individual reactances may be made relatively high with respect to the resistance, the relatively large resonant current will result in the building up of high voltages across these elements, but their joint effect in the circuit is nil due to their 180° phase separation. This effect is of great use in radio circuits, as a small applied voltage may be used to generate high voltages across tuned circuits.

Example. When 50 produces 5,000. An alternating current of 50 volts at a frequency of 159.2 kilocycles per second is applied to a circuit consisting of an inductance of 1,000 μH, a condenser of 0.001 μF and a resistance of 10 ohms, find -

(i) Current flowing.
(ii) Reactive voltages.

At resonance, $I = \frac{E}{R} = \frac{50}{10} = 5$ amperes.

$$X_L = \omega L = \frac{2 \times 3.14 \times 159{,}200 \times 1{,}000}{10^6} = 1{,}000 \text{ ohms}$$

Since $X_L = X_c$ at resonance

$X_c = 1{,}000$ ohms

Voltage across $X_L = I\omega L = 5 \times 1{,}000 = 5{,}000$ volts

and voltage across $X_c = 5{,}000$ volts.

Thus, an applied voltage of 50 volts results in reactive voltages of 5,000 volts at the resonant frequency!!!

A series circuit is often termed an "acceptor" circuit, because maximum current flo through the circuit when alternating current at the resonant frequency is applied.

3.3 Parallel Case. The parallel case is somewhat different, since there are three conditions which may be regarded as being resonant conditions -

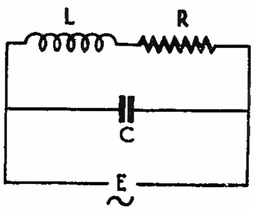
/ (i)

(i) Where $\omega L = \frac{I}{\omega C}$, that is, $f_r = \frac{1}{2\pi\sqrt{LC}}$

(ii) When power factor = 1, that is, R = Z.

(iii) When the reactive currents are equal, that is, $\frac{E}{\omega L} = \omega CE$.

In the series case these conditions occur simultaneously, but in the parallel case the resonant frequencies are slightly different. However, for radio purposes (i) is of major importance, therefore this case only will be considered.



PARALLEL CIRCUIT.

FIG. 12.

Fig. 12 shows the generally accepted condition of a parallel tuned circuit, the resistance and coil comprising one arm and the condenser the other arm.

When alternating current at the resonant frequency is applied to the circuit of Fig. 12, the impedance rises to a maximum and the current drawn from source is a minimum.

If the resistance is considered to be in the inductive arm (which is usually the case), it can be shown that the dynamic resistance of a parallel circuit at resonance is -

$$\text{Effective resistance} = \frac{L}{CR} \text{ ohms}$$

where L is in henrys, C is in farads and R is in ohms.

Example. Consider the values used in the series case in paragraph 3.2 -

$$\text{Effective resistance} = \frac{L}{CR} = \frac{1{,}000 \times 10^{-6}}{1{,}000 \times 10^{-12} \times 10}$$
$$= 100{,}000 \text{ ohms.}$$

Thus, in the series case R = 10 ohms, and in the parallel case R = 100,000 ohms

$$I_c = \text{current through condenser} = E\omega C = \frac{50 \times 10^6 \times 1{,}000}{10^{12}} = 50 \text{ mA.}$$

$$(\omega = 10^6 \text{ see series example})$$

$$I_L = \text{current through inductance} = \frac{E}{\omega L} = 50 \div 10^6 \times 1{,}000 \times 10^{-6}$$
$$= 50 \text{ mA.}$$

Thus, currents through reactances = circulating current = 50 mA.

$$\text{Current from source} = \frac{E}{\frac{L}{CR}} = \frac{50 \times 10^3}{10^5} \text{ mA} = \frac{50}{100} = 0.5 \text{ mA.}$$

Thus, a current of 0.5 mA from source maintains a circulating current of 50 mA through the reactances!!!

A parallel circuit may be termed a "rejector" circuit because it offers maximum impedance to currents at the resonant frequency.

/ 3.4

3.4 It will be convenient to compare mathematically the series and parallel resonant cases as they affect or are affected by the following -

(a) Impedance.
(b) Reactive Voltages.
(c) Reactive Currents.
(d) Supply Current.
(e) Effect of "Q".
(f) Effect of $\frac{L}{C}$ Ratio.
(g) Effect of Resistance.

SERIES CASE.	PARALLEL CASE.
(a) Impedance. $Z = \sqrt{R^2 + (X_L - X_c)^2}$ $Z = R$ at resonance.	$Z = \sqrt{\dfrac{R^2 + \omega^2 L^2}{(1 - \omega^2 LC)^2 + \omega^2 R^2}}$ $Z = \frac{L}{CR}$ at resonance. $Z = Q\omega L$
(b) Reactive Volts. $E_I = I\omega L$ $E_C = \frac{I}{\omega C}$ $E_I = E_C = QE_S$	$E_I = I\omega L$ $E_C = \frac{I}{\omega C}$
(c) Reactive Currents. $I = \frac{E}{Z} = \frac{E}{R}$	$\frac{E}{\omega L} = E\omega C$
(d) Supply Current. $I = \frac{E}{Z} = \frac{E}{R}$	$I = \frac{E}{Z} = \frac{ECR}{L}$ $I = QI_S$

(e) Effect of "Q". "Q" is a useful factor furnishing a measure of coil efficiency, and hence of tuned circuit selectivity. It is the ratio of reactance to effective resistance of the coil, that is -

$$Q = \frac{\omega L}{R} = \frac{\text{reactance}}{\text{resistance}}.$$

Therefore, the higher the Q, the greater the efficiency and selectivity of the coil and associated circuit. The effect of Q on the series and parallel circuits will be considered.

/ Series

Series Case. A sharp rise in voltage occurs at the resonant frequency across the reactive elements, the magnitude depending upon Q as follows -

Let E_C = voltage across the condenser,
E_I = voltage across the inductance,
E_S = applied voltage, and
I = current through circuit.

$$\text{Current at resonance } I = \frac{E_S}{R}$$

$$\text{Now } E_I = \omega LI = \omega L \frac{E_S}{R} = \frac{\omega L}{R} E_S = QE_S$$

so that the reactive voltage at resonance equals Q times the supply voltage, that is, $E_I = QE_S$. and so Q may be termed the "voltage amplification factor" of the coil or circuit.

Parallel Case. In the parallel case, at resonance -

$$Z = \frac{L}{CR}$$

$$\text{Multiply by } \omega = \frac{\omega L}{\omega CR} \text{ (this does not alter relative value).}$$

$$= \frac{\omega L}{R} \times \frac{I}{\omega C}$$

$$= Q \times \frac{I}{\omega C}$$

$$= Q\omega L.$$

Thus, the impedance of a parallel circuit at resonance is "Q" times the reactance of one branch of the circuit.

Another useful application of Q in the parallel case is when $\frac{I}{Q}$ is small compared to unity (that is, Q is large).

$$\text{In this instance, } E = IZ = IQ\omega L$$

$$\therefore I = \frac{E}{Q\omega L}$$

$$\text{but } \frac{E}{\omega L} = \text{circulating current.}$$

$$\text{thus } I = \frac{\text{circulating current}}{Q}$$

or circulating current = Q × supply current.

It will, therefore, be appreciated that the factor "Q" is of equal importance in series or parallel resonant circuits, hence the endeavour to keep resistance losses at a minimum in tuned circuits.

/ (f)

(f) Effect of $\frac{L}{C}$ Ratio.

Series Case. If the value $L \times C$ and the resistance Roof a series circuit are kept constant but the ratio $\frac{L}{C}$ is altered, the current at frequencies "off" resonance is decreased as this ratio increase.

Reverting to the expression for impedance -

$$Z = \sqrt{R^2 + (2\pi fL - \frac{1}{2\pi fC})^2}$$

it will be seen that an alteration to either L or C or both will affect the term in the bracket. If L is increased n times ($n > 1$), then C must be decreased n times, that is, divided by n to keep the ratio constant. The expression in the bracket then becomes -

$$2\pi fnL - \frac{1}{2\pi f\frac{C}{n}} = n\,(2\pi fL - \frac{1}{2\pi fC})$$

At resonance these reactances are still equal and cancel, but "off" resonance the bracket is n times the value before L and C were altered and, consequently, the impedance is higher and the "off" resonance current less.

$$\text{Suppose } R = 4,\ X_L - X_c = 5, \text{ and } n = 4$$

$$\text{then } Z_1 = \sqrt{4^2 + 5^2} = 6.4 \text{ ohms}$$

$$\text{and } Z_2 = \sqrt{4^2 + (n \times 5)^2} = 20.4 \text{ ohms.}$$

So that the net effect of increasing the $\frac{L}{C}$ ratio, while keeping the product constant, is to sharpen the tuning of the circuit.

Parallel Case. An increase in the ratio $\frac{C}{L}$ affects the current at non-resonance in a complicated manner. It is the effect at resonance, however, with which radio is concerned. In this case, the current is given by the expression -

$$I_S = \frac{ERC}{L} = ER \times \frac{C}{L}$$

thus an increase in the ratio $\frac{C}{L}$ increases the current taken from the supply, whereas in the series case the resonant current was unaffected.

(g) Effect of Resistance.

Series Case. An increase of R in the series circuit decreases the resonant current and flattens the tuning, since the current at resonance equals $\frac{E}{R}$ and Q is inversely proportional to R.

This effect is independent of the ratio $\frac{L}{C}$, excepting as the alteration of inductance affects the effective resistance. This, however, is usually negligible.

Parallel Case. An increase of R in the parallel case increases the current at resonance, since $I = \frac{ERC}{L}$. It also has a similar effect on the tuning and selectivity.

In both cases, the Q of the circuit is reduced.

/ Summarised

Summarised Comparison between Series and Parallel Circuits. The essential difference is that an acceptor or series circuit provides an easy path for currents at the resonant frequency and a more difficult path for all others, whereas a rejector or parallel circuit provides a difficult path for currents at the resonant frequency and an easier path for all others.

At resonance, current through an acceptor is inversely proportional to its resistance, while current through the rejector is directly proportional to its resistance.

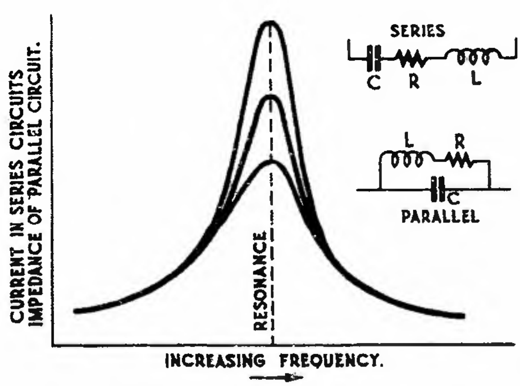
At resonance, current in the acceptor circuit is independent of the ratio of inductance to capacity; in the rejector case this is not so.

The resonant frequency for an acceptor circuit is independent of resistance. The resonant frequency for the rejector circuit is not independent of resistance but, for the small resistances usually encountered, it is substantially equal to that given by the series formula -

$$f = \frac{1}{2\pi \sqrt{LC}}$$

3.5 Resonance Curves.

Series Circuit. If an e.m.f. of constant amplitude but varying frequency is applied to a series circuit and a graph made of the currents at the various frequencies against the percentage variation of frequency above and below resonance, the curves shown in Fig. 13 will be obtained. These curves are known as series resonance curves, or resonance curves of the circuit. The "steepness" of the curves is a measure of the selectivity of the circuit. The effect of resistance on the circuit is shown by curves B and C.



SERIES AND PARALLEL RESONANT CIRCUIT CHARACTERISTICS.

FIG. 13.

Parallel Circuit. Instead of plotting current versus frequency in this case, it is more informative to plot impedance versus frequency. The curves obtained in this manner are similar in shape to the series curves, and enable the same set of curves to be used for both conditions by suitably designating the ordinate. One important difference in the parallel case, however, is that the flattening of the curves is the effect of varying the $\frac{L}{C}$ ratio, the result of this being the same as for variation of resistance in the series case.

Universal Resonance Curves. As mentioned above, the resonant rise of current in a series circuit is similar to the resonant rise of impedance in the parallel case, and, by suitably designating the ordinates, the same set of curves may be used for calculations in either case. A set of curves so drawn is known as "Universal Resonance Curves".

/ It

It is to be noted that the curves flatten out as the applied frequency nears the resonant frequency and the internal series resistance is increased, but the curves are of similar shape for all resistances at frequencies further removed from the resonant frequency. The sharpness of tuning or selectivity (ability to select one of a number of voltages of different frequencies) in such circuits is governed by the relative sharpness of the resonant curve near the resonant frequency and is extremely important in radio circuit design. Since the effective resistance is practically all in the coil, the condenser resistance being negligible, the efficiency of the coil is the important quality determining the "goodness" of a tuned circuit. A useful measure of coil efficiency, known as Q, was defined in paragraph 3.4.

The value of Q may be determined directly from the resonant curve of either a series resonant or parallel-resonant curve as shown in Fig. 14. Q is given by the ratio of the resonant frequency to the difference between the frequencies at which the series current (for the series-resonant circuit) or the parallel voltage(for the parallel-resonant circuit) becomes 70 per cent. of the maximum value. A Q of 100 would be considered good for the medium wave band, while the Q of short-wave coils may run into several hundred. It must be remembered, however, that Q represents a ratio so that the actual frequency width of the resonance curve would be proportionately greater for a high-frequency circuit than for a low-frequency circuit having the same Q.

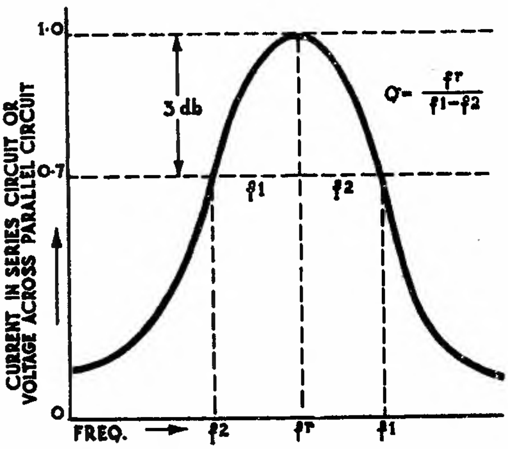
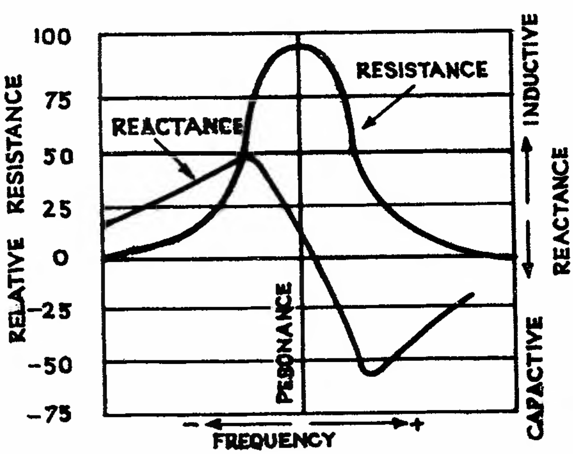


FIG. 14. RESONANT CURVE.

Parallel-Resonant Circuit Impedance. The parallel-resonant circuit offers pure resistance (its resonant impedance) between its terminals at resonant frequency, and becomes reactive for frequencies higher and lower. The manner in which this reactance varies with frequency is shown by the curve in Fig. 15. This figure also shows the parallel resistance component, which combines with the reactance to make up the impedance. The reactive nature of parallel impedance at frequencies off resonance is important in a number of practical applications of parallel-tuned circuits (in both transmitters and receivers). Note that the reactance component becomes practically equal to half the resistance component, capacitively above and inductively below resonance. The maximum value of parallel impedance which is obtained at resonance is proportional to the square of the inductance and inversely proportional to the series resistance. (This resistance should not be confused with the resistance component of parallel impedance which has just been mentioned.)



REACTANCE VARIATION WITH FREQUENCY VARIATION.
FIG. 15.

/ Resonant

$$\text{Resonant impedance} = \frac{(2\pi f_r L)^2}{R}$$

$$\text{and since } \frac{2\pi f_r L}{R} = Q$$

$$\text{Resonant impedance} = (2\pi f_r L)\ Q.$$

In other words, the impedance is equal to the inductive reactance of the coil (at resonant frequency) times the Q of the circuit. Hence, the voltage developed across the parallel resonant circuit will be proportional to its Q. For this reason, the Q of the circuit is not only a measure of the selectivity, but also of its gain or amplification, since the voltage developed across it is proportional to Z. Likewise, the Q of a circuit is related to the frequency stability of an oscillator in which it is used, the frequency stability being generally better as the circuit Q is higher. This is shown in practical applications described in subsequent Papers.

4. COUPLED CIRCUITS.

4.1 When two circuits are so arranged that energy can be transferred from one to the other, they are said to be coupled. Such coupling may be classified into "mutual" and "direct" coupling, according to the nature of the path connecting the one circuit to the other.

Mutual Coupling has already been encountered when dealing with mutual induction, where, owing to the proximity of the two inductances, the changing magnetic field, due to a changing current in one, sets up voltages across the other. Fig. 16 shows the most general form of mutual coupling, where the two coils L_1 and L_2 are coupled by mutual inductance M. The resistances R_1 and R_2 and capacities C_1 and C_2 represent the resistance and capacity usually associated with this type of circuit, and L_a and L_b represent the leakage inductances. The leakage inductances correspond to magnetic flux lines that do not produce linkages to any secondary circuit.

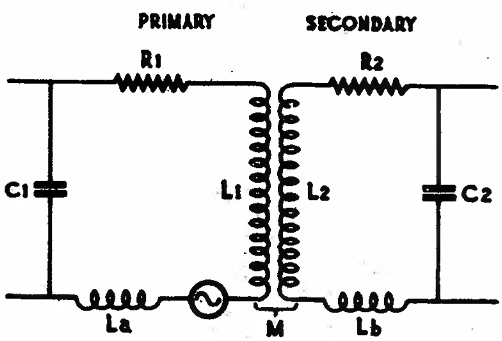


FIG. 16. MUTUAL COUPLING.

4.2 Coupling Coefficient (K). As stated earlier, the common property of two coils which gives transformer action is their mutual inductance (M). The value of mutual inductance is determined by the self-inductance of each of the two coils and their position with respect to each other. In practice, the coupling between two coils is given in terms of their coefficient of coupling, designated by "K". The coupling is maximum (unity of 100 per cent.) when all the flux produced by one coil links with all the turns of the other coil. With air-core coils in radio-frequency circuits, however, the coupling is much "looser" than this. It is generally expressed by the following relation -

$$K = \frac{M}{\sqrt{L_1 L_2}}$$

in which K is the coefficient of coupling (expressed either as a decimal part of 1 or, when multiplied by 100, as a percentage), M is the mutual inductance, L_1 is the self-inductance of one coil, and L_2 is the self-inductance of the other coil. M, L_1 and L_2 must be in the same units (henrys, millihenrys or microhenrys).

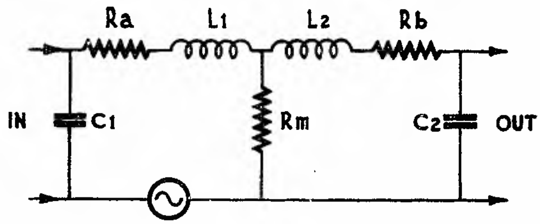
/ 4.3

4.3 Critical coupling is that which gives the maximum transfer of energy from the primary to the secondary. However, the sharpness of resonance for the combination is considerably lessened under this condition, the resonance curve usually having two peaks appreciably separated. For good selectivity, the coupling is, therefore, made considerably less than the critical value, even though this reduces the amplification or gain. With the coil combinations used in radio receivers, coupling of approximately K = 0.05 per cent. or less is represenative, whereas, for critical coupling, the coefficient might be 0.5 per cent. to 1.0 per cent. The value of the coefficient for critical coupling is also related to the respective Q's of the two coils -

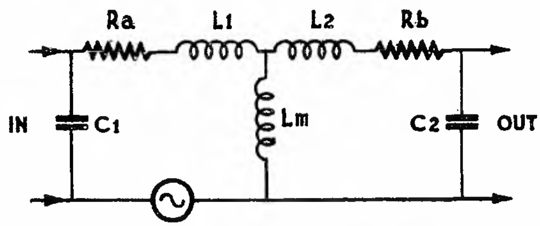
$$K_{crit} = \frac{1}{\sqrt{Q_p Q_s}}$$

where the two Q values are for the primary and secondary, respectively. For instance, if the primary and secondary Q's are equal, the value of critical K is the reciprocal of the Q for one coil - 0.01 or 1 per cent. where each has a Q of 100. Therefore, for the same value of self-inductance, K becomes smaller as Q becomes higher.

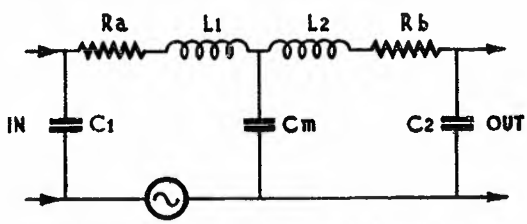
It should be remembered that, as previously mentioned, both single resonant circuits and coupled circuits are used in conjunction with other circuit elements. These other elements introduce resistance into the resonant circuits and modify the constants. In practice, it is seldom possible to precalculate the effect of such reactions, since the other quantities are usually unknown. In any case, it is usually necessary to arrive at "best conditions" by the practical process of adjustment. However, the foregoing general information is helpful in preliminary design or choice of tuned circuit combinations and in understanding why certain changes are likely to cause different behaviour in circuit performance.



Coupling Through Common Impedance Rm.



Coupling Through Common Impedance Lm.



Coupling Through Common Impedance Cm.

IMPEDANCE COUPLING CIRCUITS.

(Note. Capacitive coupling may be obtained by substituting condensers for L_1 and L_2.)

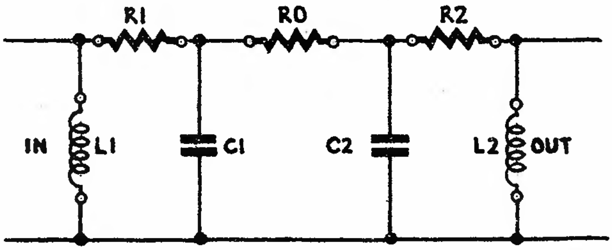
FIG. 17.

4.4 Direct Coupling may be of two types -

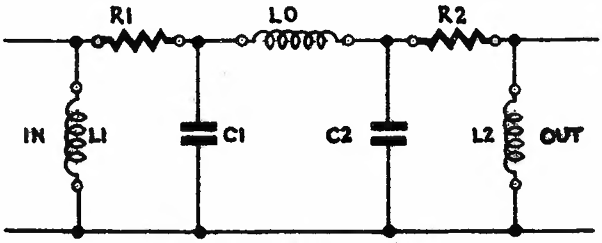
(1) The two circuits may contain a common impedance, the actual portion of the circuit which is common being a resistance, an inductance or a capacity. These couplings are shown in Fig. 17. Current circulating in one LC branch flows through the common element (that is, the inductance, resistance or capacity) and the voltage developed across this element causes current flow in the other CL branch.

/ (2)

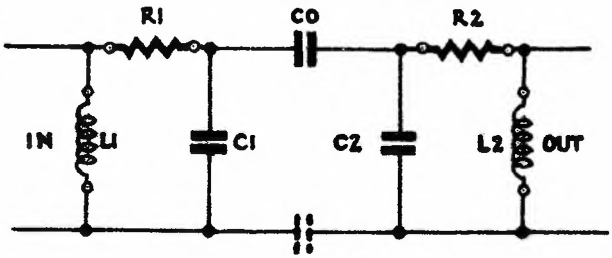
(2) The two circuits may be connected together through an impedance which does not form a part of either individual circuit. Fig. 18 shows three of this type. The two circuits $L_1R_1C_1$ and $L_2R_2C_2$ are connected by a resistance R_o, an inductance L_o or a capacity C_o, these impedances not forming part of the individual circuits. The third case shown is commonly seen either with or without the condenser shown dotted.



R_o = Coupling Resistance.



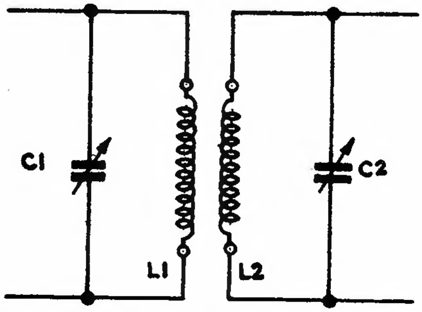
L_o = Coupling Impedance.



C_o = Coupling Impedance.

FIG. 18. COUPLING CIRCUITS.

The coupling most common in high frequency circuits, however, is that shown in Fig. 19. In such an arrangement, the coupling value may be changed by changing the number of active turns in either coil or by changing the relative position of the coils (distance or angle between them.) All of the above coupling schemes may be classified as either tight or loose. Coupling cannot, however, be measured simply in "inches" separation of coils. The separation between the coils (distance and angle between axes) and the inductance in each determine the "coefficient of coupling". Many turns in two coils very close together give a tight coupling and a large transfer of power. Few turns in two coils at right angles or far apart give a loose coupling with small actual energy transfer. More detailed descriptions of coupling methods will be found in the papers dealing with audio and radio frequency amplifiers.



HIGH FREQUENCY COUPLING CIRCUIT.

FIG. 19.

5. IMPEDANCE MATCHING.

5.1 It is a principle in telecommunication design that the maximum gross power of a generator, such as a thermionic valve, will be delivered to its load when the load resistance is equal to the internal resistance of the generator. In other words, maximum power will be taken from the generator when its resistance is matched by the load resistance. Although this particular statement is literally true, it might not describe the most desirable condition for loading a valve. For one thing, the efficiency would be only 50 per cent., half the power being consumed in the generator and half in the load. From the principle, however, a system of more or less standard practice in designing radio circuits has developed which comes under the broad heading of "impedance matching". The term means, generally, that the load impedance presented to the source is transformed to suit given requirements. This is accomplished by transformers and other coupling devices.

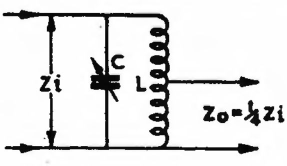
5.2 Iron core transformers are widely used for coupling between load and valve, in audio-frequency amplifiers for instance. In such cases, the value of proper load resistance (load impedance) for maximum undistorted power output will be given for the valve. This load resistance, it will be noted, is not the same as the rated anode resistance of the valve, which is equivalent to its internal resistance as a generator. A second figure will be given for the actual impedance of the load device to which the valve must supply undistorted power. The matching of this load to the given requirements of the valve is a function of the coupling transformer. This function is to make the actual impedance of the load device match the rated load impedance of the valve, so far as the valve is concerned. This requires that the transformer will have the proper ratio of secondary to primary turns. The turn ratio will be equal to the square root of the impedance ratio -

$$\frac{N_s}{N_p} = \sqrt{\frac{Z_s}{Z_a}}$$

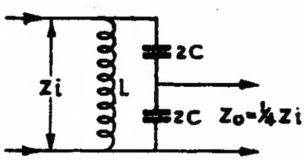
where N_s and N_p are the numbers of secondary and primary turns, Z_s is the impedance of the load device and Z_a is the rated load resistance of the valve.

Transformers are also used to provide proper impedance matching in radio-frequency circuits, although here the problem is not one of simply choosing a calculated turn ratio. The right condition is obtained by varying the mutual inductance between the output circuit and the load, that is, by adjustment of turns and distance between coils, as will be shown in later Papers.

5.3 Matching by Tapped Circuits. In addition to impedance matching by indirect inductive coupling with tuned circuits, frequent use is made of directly tapped resonant circuits. Two methods for parallel resonant circuits are shown in Fig. 20. In one case (a) the tapping is across part of the coil, while in the other case (b) the tapping is across one of two tuning condensers in series. In both cases, the impedance between the tapped points will be, to the total impedance, practically as the square of the reactance between the tap points is to the total reactance of the branch in which the tapping is done. If the coil is tapped in the centre, the reactance between the tap points will be one-half the total inductive reactance and the impedance between these points will be $(\frac{1}{2})^2$ or one-fourth the total parallel impedance of the circuit. Similar results are obtained if the tap is made across one of two equal capacitance condensers connected in series. If the condenser across which the tap is made has twice the capacitance of the other, however, the impedance Z_o will be one-ninth the total, since the reactance between the tap points would then be but a third - capacitive reactance decreasing as the capacitance is increased. These methods are, of course, equivalent to adjusting the coupling between the two circuits.



(a)

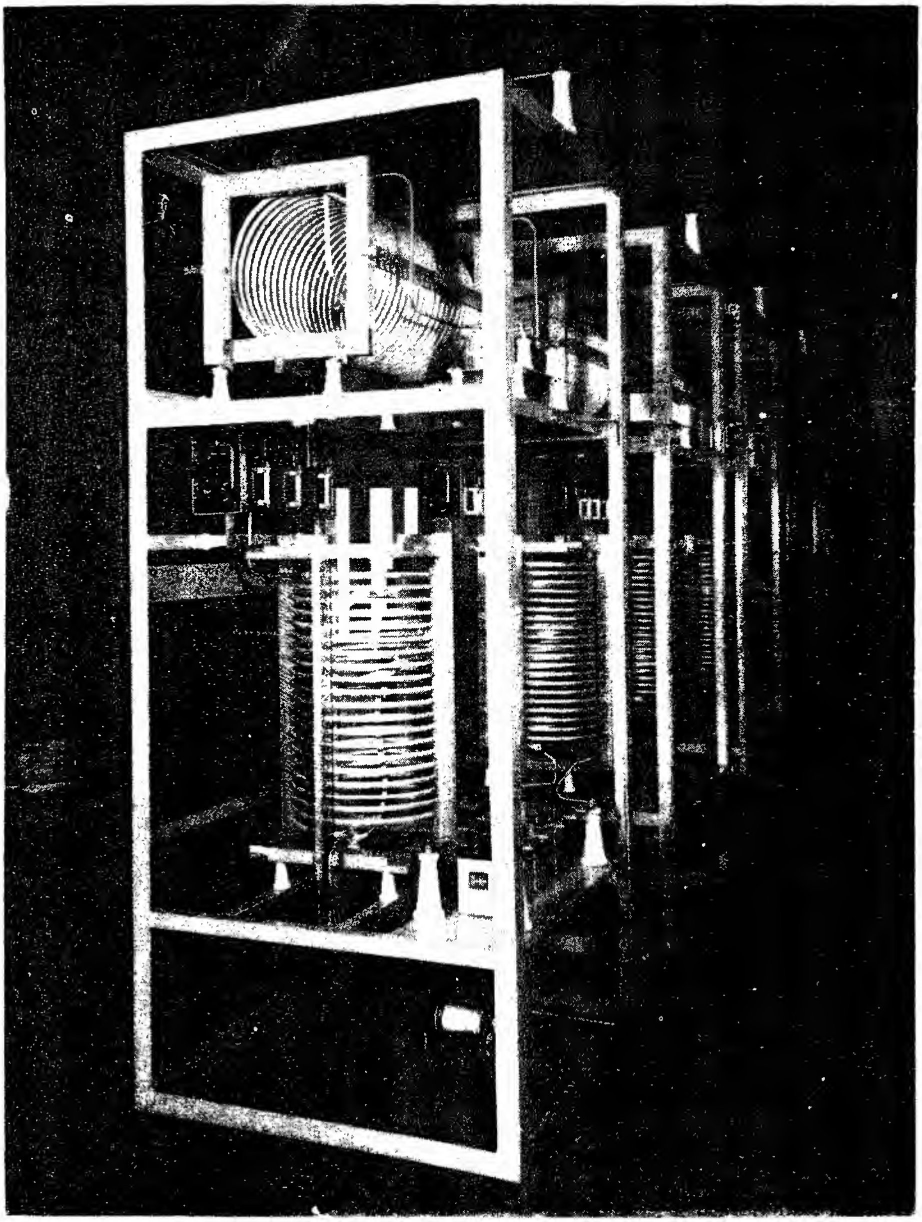


(b)

PARALLEL RESONANT CIRCUITS.

FIG. 20.

/ Aerial



AERIAL COUPLING UNITS IN COURSE OF MANUFACTURE.

5.4 Link Coupling. Another coupling arrangement used for impedance matching radio-frequency circuits is that known as "link coupling". It is used for transferring energy between two tuned circuits which are separated by space, so that there is not a direct mutual coupling between the two coils. Link coupling is especially helpful in minimising incidental capacitive coupling between the two circuits due to the distributed capacitance of the windings, thereby minimising the transfer of undesired harmonic components of the desired fundamental. Two typical versions of link coupling are shown in Fig. 21. Both versions represent an impedance step-down from one tuned circuit to the coupling line, and then an impedance step-up from the line to the other tuned circuit. The arrangement of Fig. 21a will be recognised as an adaptation of the impedance-tapping method previously shown in Fig. 20a. It is sometimes called auto-transformer link coupling, because the link turns are also included in the tuned-circuit turns. The arrangement of Fig. 21b differs only in that the link turns are separate and inductively coupled to the tuned-circuit turns. The latter system is somewhat more flexible in adjustment than the tapping method, since the coupling at either end of the line can be adjusted in small steps by moving the link turns with respect to the tuned-circuit coils.

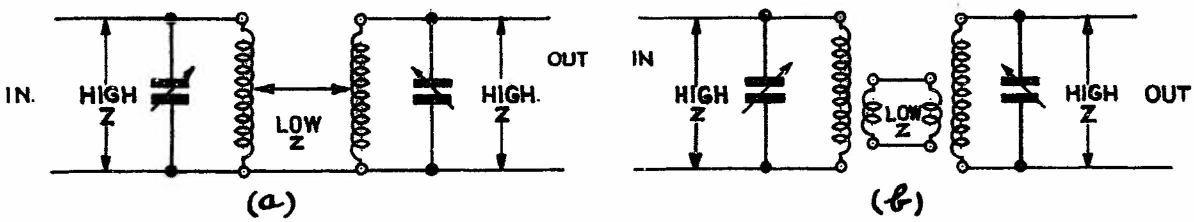


FIG. 21. LINK COUPLING.

6. FILTER CIRCUITS.

6.1 Although any resonant circuit is useful for selecting energy of a desired frequency and rejecting energy of undesired frequencies, certain combinations of circuit elements are better adapted to transmitting more or less uniformly over a band of frequencies, or to rejecting over a band of frequencies. Such rejecting action is known as "Attenuation" and such combinations are called "Filters". Filter combinations are basically of three types, as shown in the simple forms of Fig. 22.

6.2 Low Pass Filter. A "low pass" filter, as shown in Fig. 22a, is used to transmit energy below a given frequency limit and to attenuate energy of higher frequencies. Filters of this type are generally used with iron core coils or filter chokes in anode power supply systems for transmitters and receivers.

A combination of inductance and capacitance elements of the arrangement of Fig. 22a is known as a "π" or "pi" section, because its appearance resembles that of the Greek letter.

6.3 High Pass Filter. A section of the type shown in Fig. 22b is of opposite character to that shown in Fig. 22a, passing frequencies above a designated cut-off limit and attenuating lower frequencies and, therefore, being designated "high-pass". The section shown is known as a "T" section, because its form resembles that letter.

6.4 Band-Pass Filter. A type of filter for transmitting over a band of frequencies and attenuating outside this band is shown in Fig. 22c. A combination giving this action is termed a "band-pass" filter. The particular section shown will be recognised as having the same form as the indirect-capacitive coupling arrangement of Fig. 21. Similar performance is also obtainable with two tuned circuits inductively coupled. Therefore, such tuned transformers with proper coupling are used as band-pass filters, particularly in the intermediate-frequency circuits of superheterodyne receivers.

/ 6.5

6.5 Band Rejection Filter. A particular combination of series-resonant and parallel-resonant circuits intended to attenuate over a narrow band of frequencies and transmit at frequencies outside that band is shown in Fig. 22d. The series-resonant circuit would give a very low shunt path impedance at one particular frequency, while the parallel-resonant circuit in the series path would have high impedance at that frequency. Both circuits would, therefore, combine to reject or trap out energy over a narrow band of frequencies. Such action is used in wave traps.

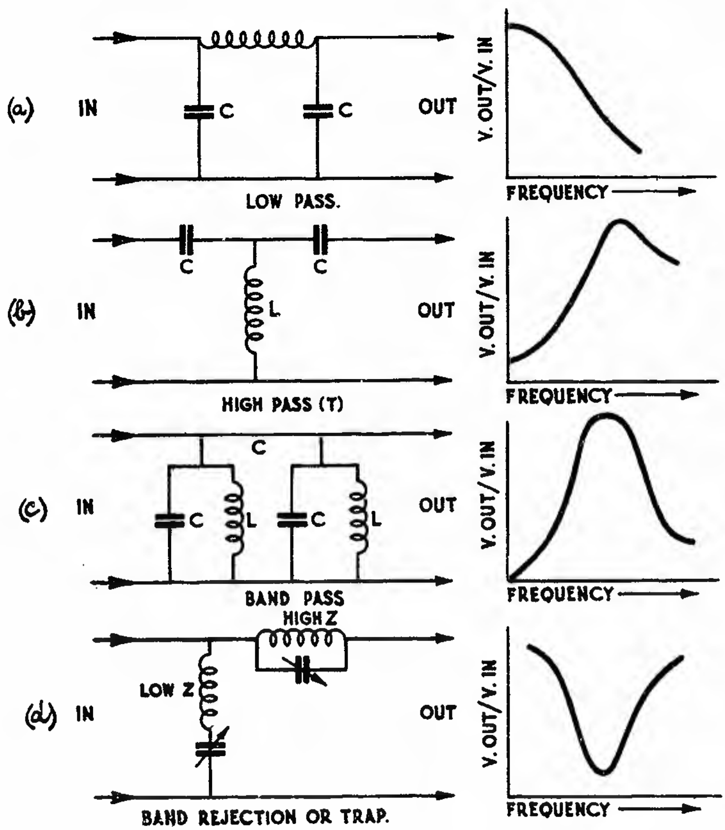


FIG. 22. CHARACTERISTICS OF FILTERS.

6.6 A given type of filtering action is increased by using more sections in cascade, or combined effects are obtained by combining different types of filter sections.

6.7 Bridge or Neutralising Circuits. Another special type of circuit, widely used in radio transmitters and to some extent in radio receivers, is the bridge circuit. Employing combinations of inductance and capacitance, it is used especially to neutralise the undesired coupling effect of a capacitance while permitting desired coupling. For instance, bridge combinations are generally used for neutralising the grid-anode capacitance of triode valves in transmitter radio-frequency amplifiers to prevent the feedback of energy from the anode to the grid circuit. Such bridge circuits are generally of the forms shown in Fig. 23.

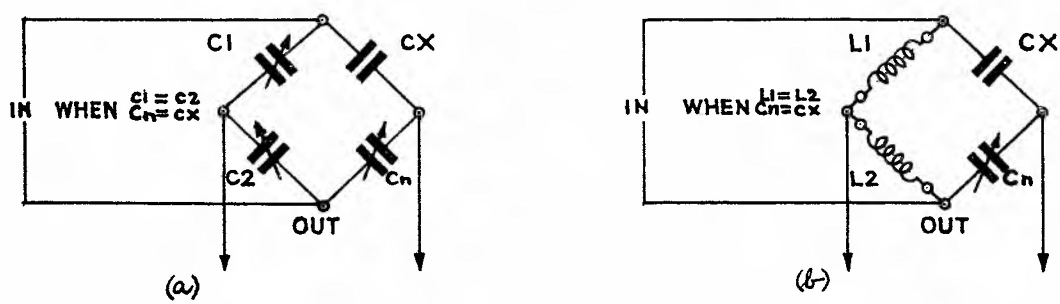


FIG. 23. BRIDGE CIRCUITS.

When a voltage is applied to one pair of terminals and the bridge is balanced, there will be no voltage across the other pair of terminals. In most practical cases, two arms of the bridge will be capacitances C_1 and C_2, as shown in Fig. 23a, or inductances L_1 and L_2 as shown in Fig. 23b. In both cases, C_x is the capacitance to be neutralised, while C_n is the capacitance adjusted to obtain the balance. With the capacitance arms of (a), balance will be obtained when -

$$C_n = \frac{C_2 C_x}{C_1}$$

and, with inductance arms of (b), balance will be obtained when -

$$C_n = \frac{L_1 C_x}{L_2}$$

When $C_1 = C_2$ in (a), or when $L_1 = L_2$ in (b), then $C_n = C_x$. This represents a desirable condition in practical neutralising circuits, because balance will be maintained over a wider frequency range of L_1, L_2 or C_1, C_2 tuning.

Bridge circuits are also used in resistance, inductance and capacitance measurement. Such bridges usually have calibrated resistances in two arms, and a calibrated resistance, inductance or capacitance in the "n" arm, the apparatus to be measured being connected in the "x" arm.

7. TEST QUESTIONS.

1. A current of 0.5 ampere at a frequency of 159,000 c/s is passed in series through a condenser of 1,000 μμF and an inductance of 800 microhenrys. Find the voltage across each and across the two together. Resistances may be neglected.

2. An alternating voltage of 100 volts (R.M.S.) at a frequency of 100,000 c/s is applied to a circuit consisting of a coil of 1,500 microhenrys inductance, a condenser of 0.007 μF capacitance and a resistance of 25 ohms all connected in series. Find the value of and the phase angle of the resulting current.

3. A condenser, resistance and inductance are connected in series across an alternating current supply of 100 volts, 500 c/s. If the capacitance of the condenser is 2 microfarads, the inductance 1 henry and the resistance 10 ohms, what is the current flowing in the circuit, and what is the voltage across each item?

4. An inductance and condenser are connected in series and are found to resonate at a frequency corresponding to a wavelength of 750 metres. A second condenser of 250 μμF is connected in parallel with the first condenser, and the wavelength is now found to be 1,000 metres. What is the value of the inductance and the capacity of the first condenser?

5. What is meant by resonance in a circuit? What is a resonance curve and how is it affected by resistance? A condenser of 1,000 μμF is connected in series with an inductance of 500 μH. What is the wavelength at which this circuit would resonate? If an additional condenser of 250 μμF is connected in series, what will be the wavelength of resonance?

6. A tuned circuit consists of a condenser of 0.0005 μF capacitance and an inductor of 1 mH. What is the wavelength at which it will resonate? What will be the effect on the wavelength of adding a condenser of 0.0003 μF in series with the original condenser and an inductor of 0.4 mH in parallel with the original inductor?

END OF PAPER.

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COURSE OF TECHNICAL INSTRUCTION.

RADIO APPARATUS.

CONTENTS:

1. INTRODUCTION.

1.1 The basic circuit theory of radio has been discussed, but, before studying the practical application of these principles, it is advisable to examine briefly the practical design of elements used in radio circuits. These elements are -

(i) Resistors,
(ii) Inductors,
(iii) Condensers, and
(iv) Dielectrics.

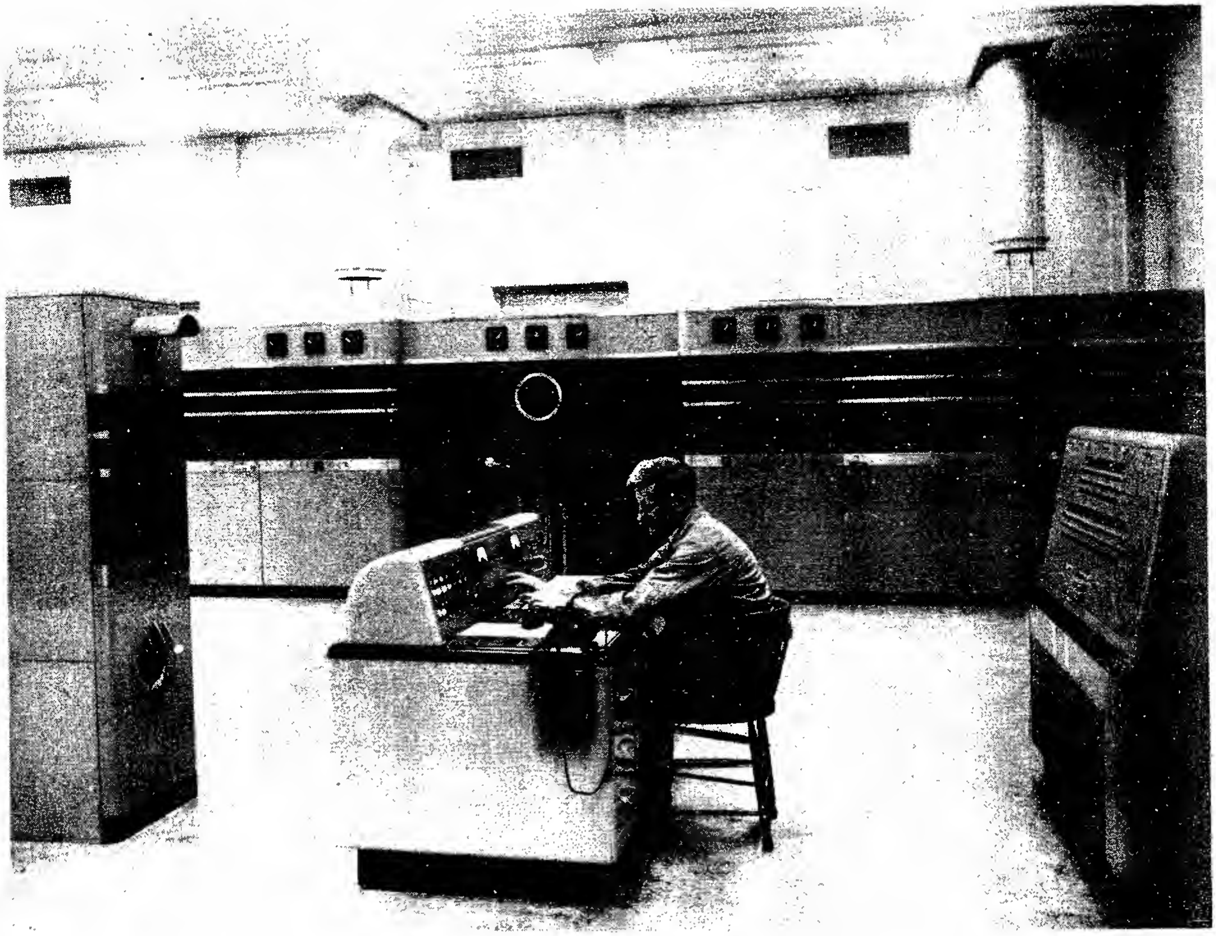
2. RESISTORS FOR RADIO.

2.1 A Resistor is a circuit element so designed that electrical resistance is its important property. A resistor consists essentially of an electrical conductor, usually of relatively high resistivity, formed in such a manner that a relatively high resistance is enclosed in a relatively small volume.

2.2 Physical and Electrical Conditions. Resistors find application in circuits for current and voltage adjustment, and may be required to carry large or very small currents which may be at audio, radio or very high frequencies or direct current. In such applications, there are two general considerations. First, the energy dissipation and, second, the accuracy of the resistor resistance value. The first of these factors must be considered in order that the resistor will not be physically damaged by overheating. Many communication circuits require that the value of a resistance should be constant and should not vary with temperature, humidity or frequency change, and also that it should not alter with age. The proper choice of resistance material will minimise the effects of humidity, temperature and time, but careful design is necessary to minimise the effect of frequency. This is further discussed under "Variation of Electrical Quantities with Frequency" in paragraphs 5.1-5.3.

2.3 Energy Dissipation, Temperature Rise. The electrical energy dissipated in a resistor of resistance R may be measured from the power, which equals I^2R, where I is the effective current. The energy dissipated is the product of power and time and equals I^2Rt, where t equals the time in seconds, the product being watt-seconds or joules. It is thus seen that the total electrical energy dissipated in the resistor is in the form of heat. The resultant temperature increase in the resistor may constitute a hazard to the resistor materials or a fire hazard to nearby materials. In order to safeguard against the various conditions, certain standards of temperature rise have been adopted.

The resistive conductors, which have general use in the construction of resistors, are usually alloys containing various proportions of nickel, chromium, iron, copper, manganese and zinc. The materials which serve as supports for the conductors are prepared in many forms and include mica, micarta, porcelain, pottery, lava asbestos, vitreous enamel and other compositions. The non-metallic resistor elements are usually formed with carbon or graphite, generally in the form of powder, and held together by suitable binders.



CONTROL DESK AT VLC TRANSMITTER.

Radio transmitters, such as VLC, depend to a large extent on the quality of the components, such as resistors, inductors and condensers.

2.4 Noise. Potentiometers (used as volume controls), fixed resistances, or wire-wound or composition forms of resistors may create microphonic noise when carrying current. This noise may be due to poor sliding contacts, loose connections, resistor material of poor composition, or to thermal agitation in the resistor material. Thermal agitation increases with the temperature. When resistors with these defects are used in circuits followed by high amplification, the resulting noise in the output circuit may seriously affect the amplified signal. Construction of resistors should be such that microphonic disturbances will not occur during normal operation.

/2.5

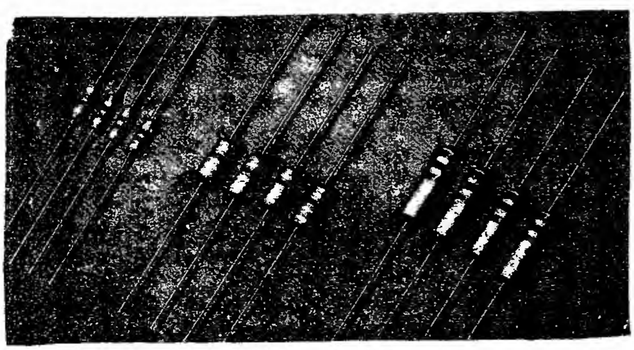
2.5 Types of Resistors.

Vitreous Enamelled Resistor. This type of unit has many uses in moderate and small power circuits, such as voltage dividers, loads, ballast and grid leaks in high power transmitters. Windings are generally of one of the nickel-chrome-iron alloys, such as nichrome, nilvar or manganin, wound on a ceramic tube with a winding pitch usually not less than twice the wire diameter. The resistive conductor, ceramic base and vitreous enamel are chosen on physical and electrical merits, for these units operate at high temperatures and must be physically and electrically sound. Typical values range from 1 to 100,000 ohms with ratings from 5 to 300 watts, sizes varying from 1/4 inch diameter by 1 inch long to 1-1/4 inch diameter by 8-1/2 inches long.

Wire-Wound Resistors. There are many uses for this resistor unit (which consists of a winding on a spool (from one to three inches long) of some insulator material) in both small power and high precision applications. The resistive conductor (nichrome, advance, manganin) may be wound on this spool and covered in several forms, dependent on the intended use. If considerable energy is to be dissipated, the winding may be placed on a small porcelain cylinder, with a brass core added if greater dissipation is required. Adjacent turns may be insulated from one another by the use of silk, cotton, mica, asbestos coverings or spacers, or the wires may be enamel or oxide covering. When very high dissipations are intended, the winding is usually covered with a refractory cement, otherwise the covering may be of paper or fibre. The resistance of precision resistors is specified within a tolerance of plus or minus 1 per cent. They are usually wound non-inductively, and typical uses are voltage multipliers in meters, unit resistors in attenuators and laboratory plug-in resistance boxes. Since the resistance in resistors for power use generally need not have a precise value, these are made with greater tolerance - plus or minus 5 per cent.

Wire-Wound Flat Type. Another form of wire-wound resistor having general use is that in which a bare winding is applied to an insulating card or strip of fibre. The single layer resistive conductor is wound in values of 10 to 3,000 ohms, which may be centre-tapped to a precision of 1 per cent. at voltages to about 20 volts and dissipations to about 3 watts. Typical applications are bias resistors, loading centre-tapping alternating current-heated valve filaments and telephone communication circuits. Another type commonly used as a loading unit, which may be used at high-frequencies as well, is constructed by weaving the resistive conductor into supporting strands of long-fibre asbestos thread. This produces a unit which is very light, rugged and well ventilated. These units may be combined to provide an adjustable load up to 7 kilowatts.

Composition of Carbon Resistors. There are numerous applications in communication circuits where non-metallic resistors are used. These resistors generally incorporate powdered carbon or graphite in their construction, although the grid lead resistor may be formed with a metallic film.



CARBON RESISTORS.

The fixed resistor of the composition type is made by combining in the right proportions -

(i) an inert material,
(ii) a synthetic resin bond, and
(iii) a material which is an electrical conductor.

Most composition resistors use for the inert material a sand, talc, asbestos pulp or some such substance as filler. The filler, with the appropriate quantity of powdered carbon or graphite, is bonded with a synthetic resin to give a resistor unit of the desired rating. Design requires that the unit be able to dissipate safely the energy liberated within it, and the resistance must not be

/unduly

unduly influenced by temperature, humidity, or voltage or frequency of the current through it, and must also maintain the value while aging. Typical applications of this type are grid bias resistors, anode load resistors, voltage dividers, voltage reduction, decoupling resistors, oscillation suppression, etc. Values range from ohms to megohms, with power ratings up to approximately 10 watts. These resistors are painted according to a colour code which is fairly generally adopted. Resistors are painted with three distinguishing colours -

(i) The body of the resistor indicates 1st digit of value.
(ii) The end colour indicates 2nd digit of value.
(iii) The dot or band indicates the number of ciphers following the first two digits.

The colour code for composition resistors is as follows -

0 Black	5 Green
1 Brown	6 Blue
2 Red	7 Violet
3 Orange	8 Gray
4 Yellow	9 White

For example -

Red body 2
Green end 5
Yellow dot 4 ciphers

= 250,000 ohms.

2.6 Non-Reactive Wire-Wound Resistors. Reactance effects associated with wire-wound resistors (referred to again later) can be minimised by special winding arrangements.

The inductance of a resistor is determined primarily by the number of turns of wire and the area enclosed by the individual turns. To keep the inductance low, each turn should enclose the minimum possible area and the wire should have as many ohms per foot of length as possible so that the length required to obtain the desired resistance will be small. In addition, it is desirable that adjacent turns carry current in opposite directions, so that the residual inductance of an individual turn is neutralised by the effect of adjacent turns. A low capacitive reactance associated with a resistor is obtained by arranging the winding in such a way that adjacent turns of wire have a low potential difference between them and are as far apart as possible. Methods that can be used to minimise the reactive effects associated with a resistor are shown in Fig. 1.

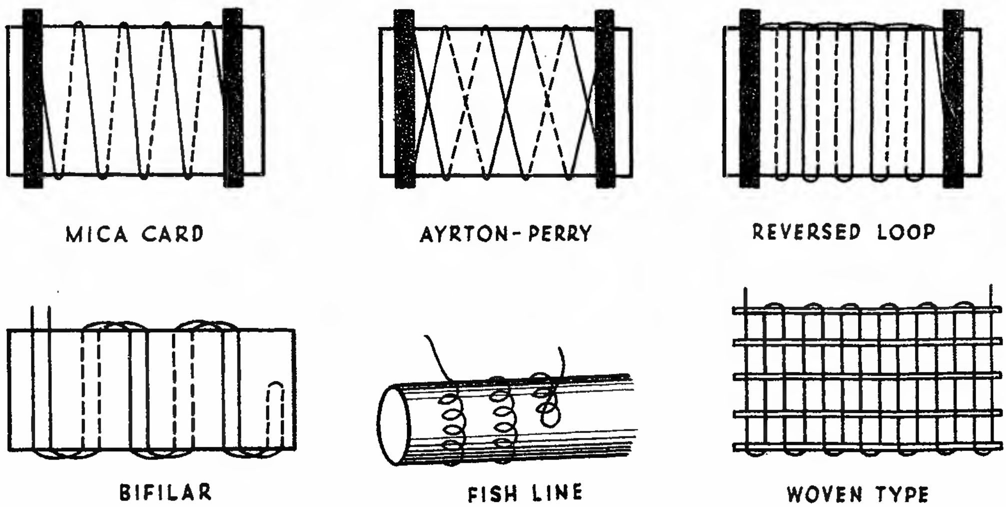


FIG. 1. SOME METHODS OF REDUCING THE INDUCTIVE AND/OR CAPACITIVE EFFECTS OF WIRE-WOUND RESISTORS.

The Mica-Card type uses a single-layer winding on a thin mica form provided with copper end strips to serve as terminals and reinforcing. A low inductance can be obtained by making the card very thin and using small gauge wire to give a high resistance per turn.

The Ayrton-Perry type of resistor is constructed by winding a spaced layer of insulated wire on a thin strip, after which a second wire is wound in the opposite direction between the turns of the first winding. The two windings are connected in parallel and, therefore, produce practically zero resultant magnetic effect. The distributed capacity is low because adjacent turns have very little potential difference between them.

The Reversed-Loop winding obtains low inductive effects by making a half hitch at the end of each turn and thus reversing the direction of the current in adjacent turns.

The Fish-Line type of resistor consists of a fine resistance wire wound over a silk cord that serves as a core, and the resulting "fish-line" is then space wound on a cylindrical form.

The Bifilar Winding has negligible inductance, but the capacity is relatively large because the beginning and the end of the resistance are close together.

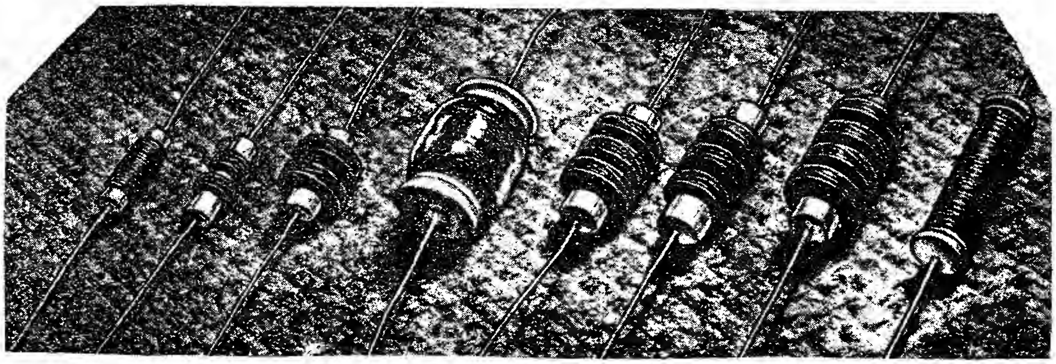
The Woven type of resistor has been referred to previously, but there is also a low rated type using silk thread in lieu of the asbestos fibre mentioned.

3. INDUCTORS FOR RADIO.

3.1 An inductor is a circuit element designed to possess the quality of inductance. There are two main types used in radio -

(i) Inductors designed to operate in low frequency circuits. These usually possess ferrous cores.

(ii) Inductors designed to operate at radio frequencies. These may have air cores or special low-loss powdered iron cores.



RADIO INDUCTORS.

3.2 Low Frequency Inductors. By the use of some ferrous core to form a complete magnetic path within a coil winding, the inductance can be increased from several hundred to several thousand times its air core value. The basic expression for the inductance of an iron core with a single-layer winding appears as -

$$L \text{ (microhenrys)} = 0.4\pi n^2 \mu A \ell \times 10^{-2}$$

where n = number of turns of winding,
A = cross-section of winding in centimetres,
ℓ = mean length of winding in centimetres, and
μ = effective permeability for conditions of use.

/This

This is a fundamental equation for any magnetic circuit and has general use in magnetic equations.

Losses in Ferrous Cored Inductors. Since the core of an inductor is an integral part of the device, energy losses which occur in this material are considered as energy losses in the inductor. The losses in the core, therefore, must be added to the loss due to the current flow in the inductor winding. This gives a total loss -

$$W\ (\text{total}) = W_e + W_h + I^2R.$$

where W_e = eddy current loss,
W_h = loss due to hysteresis, and
I = effective current flowing through the winding with an ohmic resistance R under a given set of conditions.

Due to the factors involved, the total loss will vary with frequency. When it is essential that an iron-core inductor has a more constant inductance, minimising the effects of direct current saturation and high alternating current magnetomotive forces, an air-gap is left in the magnetic circuit. This causes a reduction of the amount of self-induction possible with a given winding. The amount of decrease depends upon the comparative lengths of iron and air in the magnetic circuit and upon the degree of saturation of the iron.

Applications.

Filter Reactors. Ferrous-cored inductors find uses in communication circuits involving frequencies from 120 c/s to the broadcast frequency band. The major use for inductors at the lower end of the frequency spectrum is found in the power supply for communication equipment, where current at power frequencies must be rectified to provide continuous currents. The filtering systems of such supplies are composed of capacitors and reactors, the latter commonly known as "choke" coils. Designed with a wide range of direct current ratings, winding resistances and potential ratings, filter reactors are rated up to values of about 30 henrys.

Audio Chokes. Certain communication circuits require a means for the suppression of audio frequency currents, and this action is effected by the audio reactor or retard. Such coils may be constructed of laminated silicon steel or of a pressed composition core with iron or ferrous alloy powder. A wide variety of applications requires a wide variety of constructions, etc. Nominal inductance ratings extend up into hundreds of henrys.

3.3 Inductors for Radio Frequencies. The problem of the design of an inductor has many complications, and the final design will depend on many factors. A general requirement in the design of an inductor is that the largest inductance value be produced for a given length of wire. This part of the design includes the determination of the best shape of coil form to give the required inductance with a minimum resistance or a certain inductance value within a given space.

Best Coil Form. It can be shown that with a given length of wire, wound with a given pitch, the single-layer coil which has the maximum inductance value is so shaped that the ratio -

$$\frac{\text{Diameter}}{\text{Width of winding}} = 2.46 \text{ approximately.}$$

Precision design can be effected only when the inductor is to be used at low frequency. At radio frequency, the problem is mainly the reduction of resistance and distributed capacitance. A coil is designed for use in a certain range of frequencies and, generally, an attempt is made to construct a coil with a uniformly high value of Q in this range. The radio frequency resistance is reduced by a special construction of the winding conductor, such as hollow tubing, flat or edge wound conductor, and stranded cable such as "Litz."

Form of Winding. Coils may be wound in a great many forms, and each form will have some particular application and frequency range. The coil form most generally meeting the requirements of high Q, moderate size, and reasonably small distributed capacitance is the single-layer solenoid. Among other types are honeycomb, multi-layer, basket weave, bank-wound, etc.

As the operating frequency becomes higher, the design grows more difficult. Some typical formulae for single-layer solenoids are included as a matter of interest.

Some Useful Inductance Formulae.

Inductors. Single-Layer Solenoid.

L (microhenrys) = Fn^2d

where n = number of turns,
d = diameter of coil in inches, and
F = quantity dependent on the ratio of diameter to length of coil.

Some typical values of F are -

$\frac{d}{\ell}$	0.1	0.5	1.0	5.0	10.0	50	100
F	0.0025	0.01	0.018	0.043	0.05	0.075	0.088

when d and ℓ are in inches.

A simple approximate formula for the low-frequency inductance of a single-layer coil is -

$$L \text{ (microhenrys)} = \frac{r^2n^2}{9r + 10\ell}$$

where r = radius of coil in inches, and
ℓ = length of coil in inches.

The above formula is accurate to within 1 per cent. for $\ell > 0.8r$, that is, if the coil is not too short.

4. CONDENSERS FOR RADIO.

4.1 A Capacitor or Condenser consists essentially of two conductors (the surfaces of which are relatively close together) separated by an insulating medium called the dielectric. The dielectric is that part of the condenser which is mainly responsible for the properties of the condenser itself.

The main types of condensers are -

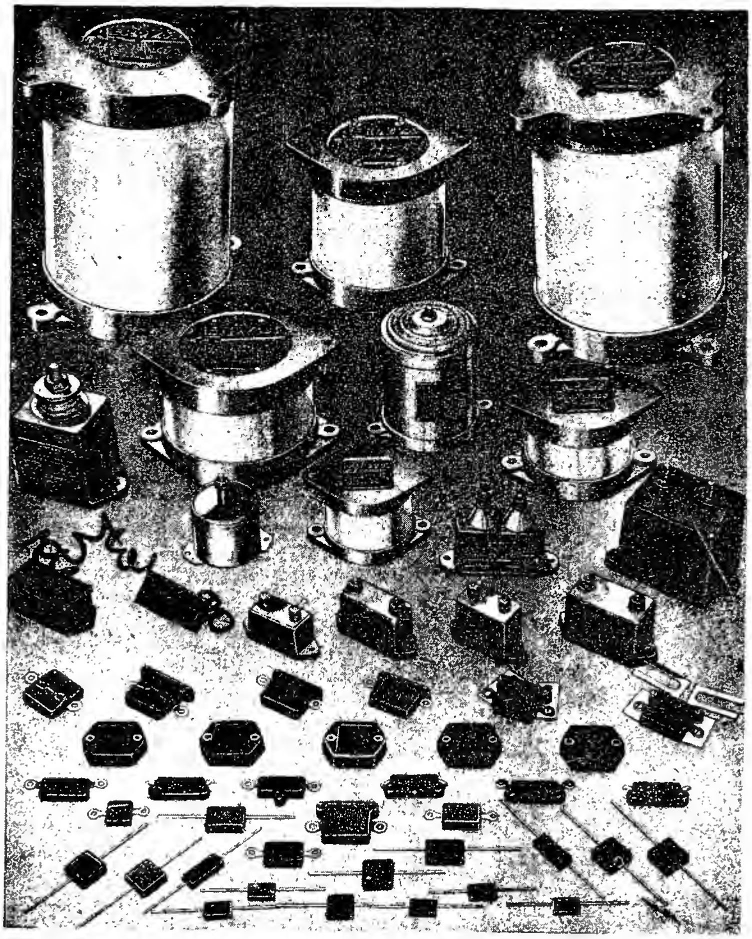
(i) Fixed Value Condensers,
(ii) Variable Condensers, and
(iii) Electrolytic Condensers.

4.2 Fixed Value Condensers. Fixed value condensers generally make use of some solid, such as mica, paper or glass, as a dielectric. The mica condenser is usually assembled by placing one or more sheets of mica between two sheets of metal foil and building up many small sections of this type into stacks. The impregnated paper condenser is usually built up of multiple layers of paper and foil, and these layers are formed into a roll. This permits the construction of a relatively large capacitance in a relatively small volume. Larger types suitable for use at relatively high voltages, such as are encountered in radio transmitters, may consist of solid metal plates immersed in oil, compressed air or some other insert gas. The essential advantage derived in use of these fluids, etc., is the increase in dielectric constant and dielectric strength as compared with normal pressure air. The self-healing characteristic is desirable as compared with the condition following breakdown of a solid dielectric condenser, which in the latter generally ends its useful life.

/Rating

Rating. At low frequencies, the volt-ampere rating of a solid dielectric condenser is limited by the dielectric strength of the insulation. At high frequencies, since the voltage drop through the condenser greatly diminishes, the safe heating of the dielectric limits the rating of the condenser. It must be recalled that solid dielectric materials, such as glass, paper, rubber, wax, etc., lose their insulating properties very rapidly as the temperature increases. The effect of energy loss on the insulation requires a variable volt-ampere rating for a condenser. Typical rating of a 0.001 μF mica condenser is shown below -

Frequency.	Safe Current.	Safe Voltage.	Safe kVA.
0.1 kc/s	1.9 amperes	3,000 volts	5.7
0.3 kc/s	5.65 amperes	3,000 volts	17
1.0 kc/s	11.2 amperes	1,780 volts	20
3.0 kc/s	11.4 amperes	605 volts	8.9
10.0 kc/s	11.2 amperes	178 volts	2.0



TYPICAL RADIO CONDENSERS.

Losses. The losses in solid dielectric condensers may be grouped under the following headings -

/(i)

(i) Loss due to leakage through or round the dielectric. Loss is appreciable when the insulation resistance of the dielectric is low.

(ii) Actual conductor resistance loss of leads, connections and of the plates themselves.

(iii) Loss due to dielectric absorption or hysteresis.

(iv) Corona loss due to discharge from plate edges.

It is difficult to determine separately the individual losses. Therefore, for any set of conditions, the losses are determined together, and the total is known as "equivalent series resistance." The value of this resistance, when multiplied by the square of the alternating current flowing in the condenser, represents the rate of energy consumption in the condenser, that is, its power loss.

4.3 Variable Condensers. Variable value condensers usually consist of two sets of metal plates, one set being mounted for rotation on a shaft. It is, of course, necessary to maintain the separation between the condenser plates, and this is generally done by the use of supporting and mounting insulation. Apart from this material, the dielectric between the plates is air. Variable air condensers have their greatest use in adjusting the electrical characteristics of circuits at radio and higher frequencies. For this, the units are constructed to have a maximum capacitance (plates completely overlapping) between 500 and 15 $\mu\mu F$. The latter sizes have a major application in the generation and reception of "short-wave" signals on frequencies higher than 3,000 kc/s. The variable air condenser, having very little loss in the dielectric and separating insulation, will show negligible change in capacitance with frequency. Accurate measurements of the capacitance values at audio frequencies may therefore be depended upon at radio frequencies with a high degree of precision. The variable air condenser, therefore, finds a singular application as a capacitance standard, and, for this use, units are made with capacitances up to 5,000 $\mu\mu F$.

Effect of Stray Capacitance. Generally, condensers of variable capacity have relatively small values, and in a circuit, particularly at high frequencies, the capacitance effects of the various other parts of a circuit will be appreciable as compared with the condenser unit. Every part of the apparatus has capacitance to other parts, and these small stray capacitances may have to be taken into account as well as the capacitance of the condenser which is intentionally inserted in the circuit. The stray capacitances are particularly objectionable because they vary when parts of the circuit or conductors nearby are moved, for this makes it difficult to keep the capacitance of the circuit constant. These effects may be minimised in practice as follows -

(i) by keeping the condenser a considerable distance away from conducting or dielectric masses;

(ii) by shielding the condenser; and

(iii) by using a condenser of sufficiently large capacitance so that stray capacitances are negligible in comparison.

A combination of (ii) and (iii) is the most satisfactory compromise for general practice.

Condensers having various shapes of plates designed for particular circuit conditions are met with. Typical of these condensers may be mentioned -

(i) Straight-line capacitance.
(ii) Straight-line frequency.
(iii) Straight-line wavelength.

Type (ii) is probably the most satisfactory for general radio use since the variation of the condenser setting is directly proportional to the frequency for which the total circuit is tuning. The importance of this type follows from certain considerations

/in

in radio communication which require the spacing of broadcasting stations from each other by equal increments of frequency. Thus, this type of condenser will tune for the various broadcasting stations at equally spaced points on the condenser dial. Fig. 2 shows capacitance versus scale reading for the abovementioned types.

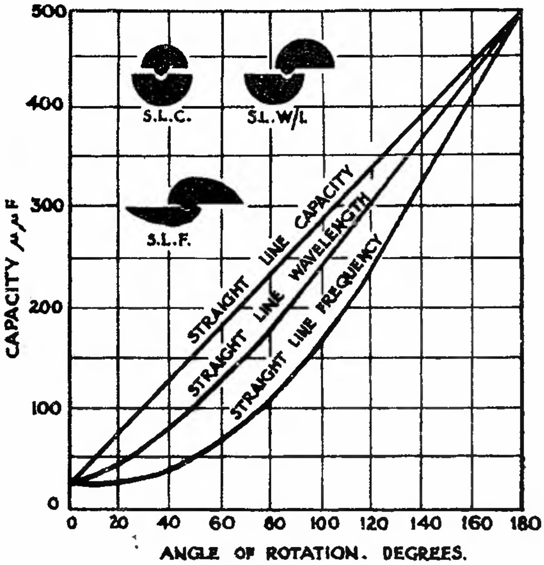


FIG. 2. CAPACITANCE VERSUS SCALE READINGS.

The ratio of maximum to minimum capacitance of the variable condenser governs the tuning range of the circuit if the coil inductance is fixed. Thus, if a 3 : 1 frequency range is to be covered, a 9 : 1 change in capacitance must be provided. Average values vary from about 10 : 1 to 25 : 1, the latter ratio applying to the larger sized condensers. The distributed capacitance of the coil and the input capacitance of the valve or other associated apparatus must be considered with the minimum condenser capacitance in computing the tuning range of the circuit.

4.4 Electrolytic Condensers. If two aluminium plates are immersed in a suitable electrolyte, such as a solution of ammonium borate or sodium phosphate, and connected to a source of direct current, a thin film of aluminium oxide forms on the positive plate which will gradually insulate the plate from the solution. The thickness of the film depends upon the voltage value used in its formation, higher voltages resulting in thicker films. Owing to the extremely small thickness of the film, a capacitance of from 0.1 to 1.5 μF per square inch of area is obtained, depending on the forming voltage. Very large values of capacitance, therefore, can be obtained with only a fraction of the space required by paper-dielectric condensers. The film gradually disintegrates if the impressed voltage is removed from the cell. The film is again formed when the voltage is reapplied, accompanied by a large leakage current which soon drops to the normal value of about 200 microamperes per μF for working voltages of about 450.

The film possesses three distinct characteristics -

(i) It will conduct more freely in one direction than the other.

(ii) It will break down as an insulator between the metallic electrode and the solution when voltages above a critical value are applied.

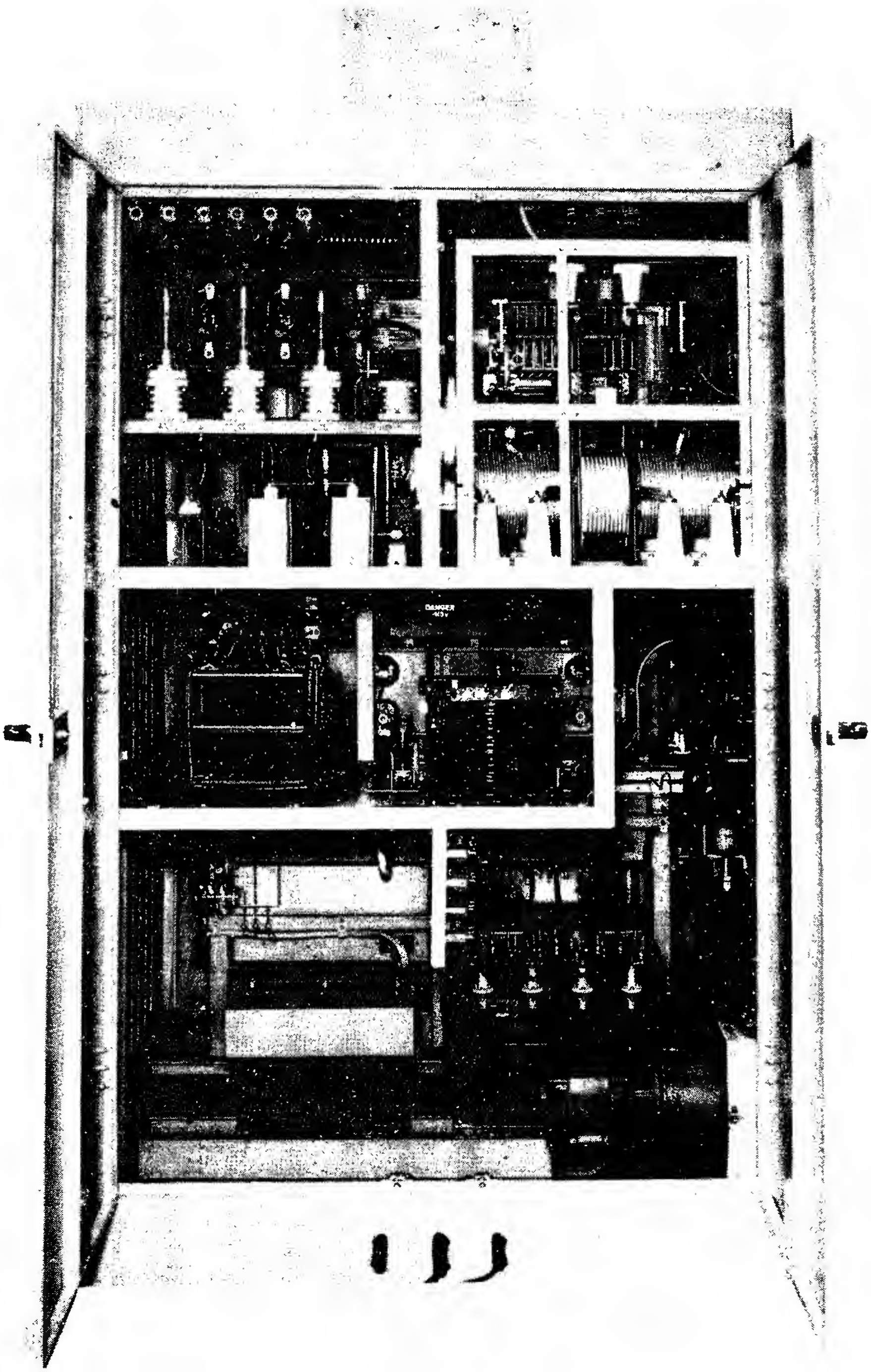
(iii) By reason of the extreme thinness of the film, it will hold a substantial charge at potentials below the breakdown voltage.

By reason of the film holding a charge for a given potential, the device assumes the important characteristic of an electrical condenser and, as such, finds considerable application as by-pass and filter condensers in radio-receivers in parts of the circuit where a direct current potential is present and where a moderate leakage current can be tolerated. Electrolytic condensers are also to be found in the anode power supplies of low power transmitters and power supply units in laboratories, broadcasting studios, etc.

Electrolytic condensers are characterised by -

Low cost per microfarad.
Very large capacity in proportion to volume.
High power factor.
Appreciable leakage.

/Broadcast



BROADCAST TRANSMITTER (500 WATT).

Note the various types of condensers, transformers and chokes used.

The characteristics, such as capacity, leakage resistance and power factor, depend greatly upon the applied voltage, temperature, previous history of the condenser, etc. The ordinary electrolytic condensers used in radio work are polarised, so that the voltage on the anode plate must never be allowed to become negative, even instantaneously. This limits the use of this type of condenser to circumstances where the alternating voltage present is less than the superimposed direct current voltage, that is, this type of condenser must not be used in circuits carrying alternating current only.

Types. There are two main types of electrolytic condenser, the "wet" and the "dry" types.

In the wet type, the anode consists of folded aluminium foil immersed in the electrolyte, which is held in an aluminium container. This container acts as the cathode electrode for the purpose of connection to outside circuits, and is usually lined inside with a perforated insulating material to prevent contact between the anode and the cathode. The anode is folded or spiralled so that it presents as large a surface to the electrolyte as possible.

In the dry type, the positive and negative electrodes consist of aluminium foil separated by paper or gauze that is saturated with an electrolyte, which is either a highly viscous liquid or a fudge-like solid. The entire assembly is wound in a roll and mounted in a waxed cardboard tube or box. The insulating film on the anode is formed chemically before assembly and, finally, the condenser is subjected to a re-forming process to repair any damage that may have occurred during assembly.

4.5 Dielectrics. The insulating medium that separates the plates of a condenser is known as the "dielectric," and plays an important part in determining the characteristics of the condenser. The presence of a dielectric (other than a vacuum) raises the capacity of the condenser, in comparison to its capacity with vacuum as the dielectric, by a factor known as the "Dielectric Constant" or "Specific Inductive Capacity." The dielectric constants of some of the more commonly used insulating materials are shown in Table 1 and, of these, plastic and ceramic types are perhaps the most important. Some of the plastic dielectrics, such as bakelite, cellulose acetate, vinylite, etc., have very poor power factors. Others, notably polystyrene, have an unusually low power factor which, in some cases, approaches that of quartz. The more important ceramic dielectrics include steatite, cordierite and titanium dioxide ceramics and mycalex.

The term Steatite refers to ceramic products that contain hydrous magnesium silicate (talc) and a predominant constituent of the fired body. Steatite materials combine low dielectric losses, high dielectric strength and high insulation resistance. They are particularly suitable for general insulation purposes at high frequencies and high temperatures. It is possible to employ dry pressing in forming the product so that parts can be manufactured in automatic presses to good accuracy and with economy.

Cordierite is a complex magnesium aluminium silicate. Ceramics based upon cordierite can be made that have an unusually low coefficient of thermal expansion, so that this material is particularly suitable for coil forms.

The Titanium Dioxide Ceramics are characterised by a high dielectric constant and low dielectric losses. This material finds its chief use as a dielectric of condensers, and gives high mechanical stability and a high capacity in a relatively small volume. The temperature coefficient of capacity of such condensers depends upon the mixture, and values can be obtained that range from substantially zero to relatively large negative values. This gives the possibility of providing a condenser that will correct for the positive coefficient of the remaining coils and condensers in a circuit.

Mycalex is made from ground mica and lead borate glass mixed and fired. Metal inserts can be moulded into mycalex pieces because of the relatively low firing temperature and the fact that there is no shrinkage. Mycalex can be machined, although with difficulty. It is used in radio transmitters for coil supports, control shafts, antenna insulators and in other places where an insulator must be employed that is capable of withstanding high voltage.

/Table 1.

PROPERTIES OF DIELECTRICS.

Dielectric	Dielectric Constant	Power Factor Per Cent. 60 c/s	1 kc/s	1 Mc/s	Specific Gravity	Machineability
Amber	2.9			0.2	1.1	Very Good
Casein, Moulded	6.4			6	1.33	Very Good
Cellulose Acetate	6-8	7		3-6	1.3	Very Good
Cordierite Ceramics	5-5.5		0.6-1	0.4-07	2.1	
Fibre	4-5	6-9	5	5	1.3	Very Good
Glass, Pyrex	4.5		0.5	0.2	2.25	Very Poor
Mica, Clear India	7-7.3	0.03	0.02	0.02	2.8	
Mycalex	6-8		0.6	0.3	3.5	Poor
Phenol, Pure	5	2		1	1.3	Very Good
Phenol, Cloth Base	5.6	5	5	5	1.38	Good
Porcelain, Dry Process	6.2-7.5	2	1	0.7	2.3	No
Quartz, Fused	4.2	0.03	0.03	0.03	2.21	Very Poor
Rubber, Hard	2.3	1	1	0.5-0.9	1.15	Fair
Steatite	6.1	1	0.4	0.3	2.5	No
Styrene, Polymerised	2.4-2.9	0.02	0.02	0.03		Good
Titanium Dioxide	90-170		0.1	0.06	4-5	No
Vinyl Resins Unfilled	4		1.4	1.7	1.35	Very Good

TABLE 1.

Dielectric Efficiency. Dielectric efficiency is the ratio of the energy output to the energy input, thereby taking account of any waste of energy that occurs during the charge and discharge of a condenser. The waste may be considered as being made up of the following losses -

(i) Conductor losses.

(ii) Chemical action which may take place if damp is present.

(iii) Leakage losses. These may be due to faulty insulation or to corona discharge from points and edges of condenser plates.

(iv) Dielectric absorption. When a condenser is charged, the initial rush of current is followed by a relatively small, more gradual current, which appears to "soak in" to the dielectric. Similarly, when discharged by short-circuiting, the initial heavy discharge, which should leave it practically uncharged, is an incomplete one. If the condenser is set aside with the short-circuit removed, and again short-circuited a few minutes later, a second discharge can be obtained. This is due to the charge absorbed by the dielectric, which gradually leaks out into the plates and so gives the condenser another charge. If the condenser is charged and discharged periodically, this absorption causes heat to be generated in the dielectric and can be looked upon as a resistance in series with the condenser. Absorption loss varies with frequency, becoming less as the frequency rises, due to the fact that there is less time available for the absorption to take place.

This last cause of loss of efficiency (absorption) also causes an apparent variation of dielectric constant with frequency since, if the condenser is unable to absorb its full charge each time, it is equivalent to a condenser having a smaller capacity or a different dielectric. When the dielectric is air or gas, absorption is negligible.

4.6 Some Useful Formulae for Condensers.

Simple Parallel Plate Condenser.

$$C = \frac{kA}{11.3d}\ \mu\mu F$$

where k = dielectric constant of separating medium,
A = area of each plate (square centimetres), and
d = distance apart (centimetres).

$$\text{or } C = 0.2244\ k\frac{A}{d}\ \mu\mu F$$

where A is in square inches and d is in inches.

Multiple Plate Condenser.

$$C = \frac{kA\ (n - 1)}{11.3d}\ \mu\mu F$$

where n = total number of plates,
A is in square centimetres, and
d is in centimetres.

5. VARIATION OF ELECTRICAL QUANTITIES WITH FREQUENCY.

5.1 It might be inferred from the discussion on resistors, inductors and capacitors that their values are constant with frequency, but this is not the case. The ohmic resistance of a circuit is greater to radio frequency than to direct current, and a coil which at first sight would seem an inductance may actually have capacitive reactance, or a resistance may be merely a low capacitive reactance. The three quantities will now be discussed from this aspect.

5.2 Resistance. The resistance value of a given resistor depends primarily upon the resistivity and physical dimensions of the wire (or strip) used for the winding. In practically all materials, the resistivity will change as the temperature varies. The temperature changes may be due to the absorption of energy from outside sources or from energy dissipation due to current flow in the resistor. In conductors of large cross-section or in ANY conductor carrying currents of high frequency, the current flows more readily near the skin or outer surface of the wire than in the high impedance inner filaments of the wire. This phenomenon is known as "skin effect."

When a current is flowing through such a conductor, the magnetic flux that results is in the form of concentric circles around the wire. Some of this flux exists within the conductor and, therefore, links with (that is, encircles) current near the centre

/of

of the conductor while not linking with current flowing near the surface. The result is that the inductance of the central part of the conductor is greater than the inductance of the part of the conductor near the surface because of the greater number of flux linkages existing in the central region. At radio frequencies, the reactance of this extra inductance is sufficiently great to affect seriously the flow of current, most of which flows along the surface of the conductor (where the impedance is low) rather than near the centre where the impedance is high. The centre part of the conductor, therefore, does not carry its share of the current, and the effective resistance is increased since, in effect, the useful cross-section of the wire is very greatly reduced. It can be shown that skin effect in a conductor depends, in addition to the thickness of the conductor, on the parameter -

$$\frac{\sqrt{2\mu f}}{\rho}$$

where μ is permeability of the material, f is frequency of the current and ρ is volume resistivity in microhm-centimetres.

Energy loss will occur in the insulation on the wire or in any material within range of the electric and magnetic fields set up by the wire. When this loss occurs in materials which usually form an integral part of the resistor, it may be attributed to the resistor and manifests itself as resistance in addition to the resistance of the winding. This loss may be either a dielectric loss in the insulating material or hysteresis and eddy current loss in the conducting material and, as these losses vary with resistor current and frequency, the total resistance of the resistor will not be constant. This effect, together with the action of inductance and capacitance, may cause considerable variation in the total impedance. Examples of this effect will now be given for several types of resistors.

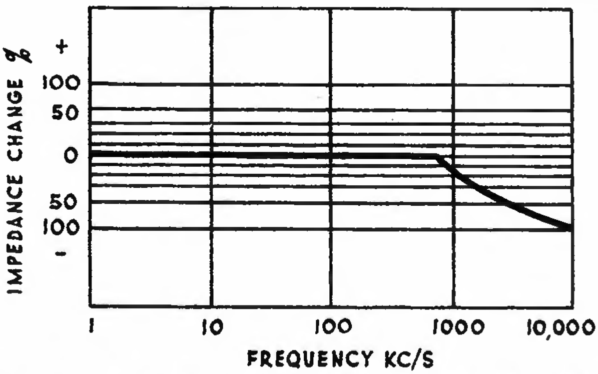


FIG. 3. HOW THE IMPEDANCE OF A PORCELAIN BASE, VITREOUS ENAMELLED RESISTOR (2850 OHMS) CHANGES WITH FREQUENCY.

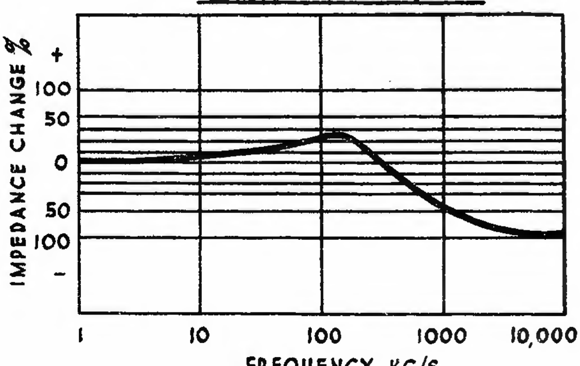


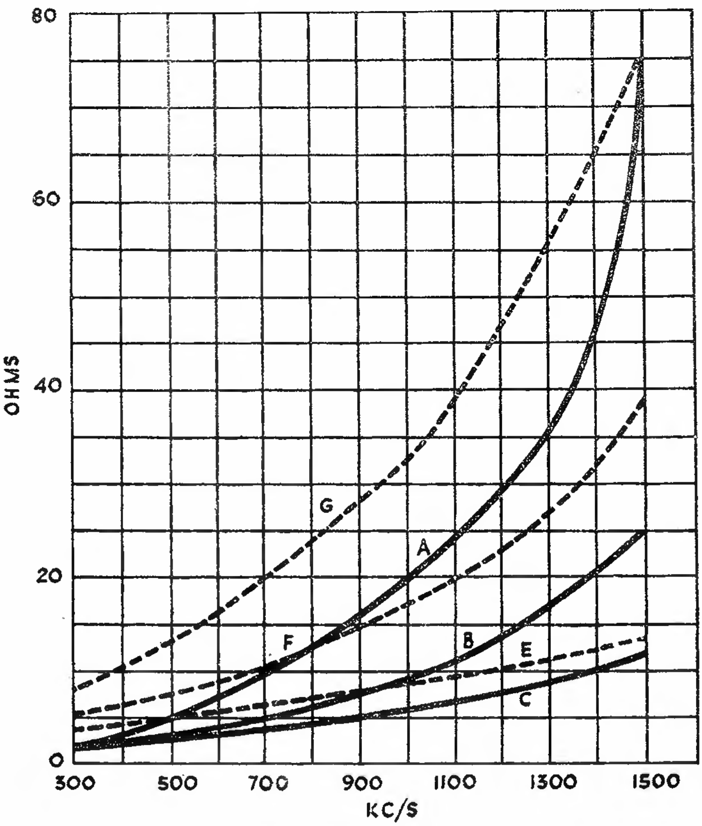
FIG. 4. HOW THE IMPEDANCE OF A WIRE-WOUND SPOOL TYPE PRECISION RESISTOR (10,000 OHMS) VARIES WITH IMPEDANCE.

Vitreous Enamelled Resistor. Fig. 3 shows a frequency characteristic curve of a porcelain wire-wound resistor, illustrating the effects of distributed capacitance and inherent inductance of the solenoid winding. For very high frequencies, the resistor impedance falls off rapidly as the distributed capacitance by-passes the current from the high-resistance winding.

Wire-Wound Resistors. Fig. 4 shows the frequency characteristic curve of a wire-wound precision type resistor. The effect of inductance and distributed capacitance is seen in the impedance variation which occurs at frequencies over 100 kc/s.

Composition Carbon Resistors. The inductance possessed by this type of resistor is that which would be found in a short straight conductor of the same size, and is of practically negligible value. The capacitance effect is that due to the electric influence of one connection terminal to the other. This is also of very small value and can usually be disregarded in the frequency bands up to about 30 Mc/s.

5.3 Inductors with Air Cores. Distributed inductance in an inductor is determined mainly by the spacing between the turns of the coil winding and particularly between those turns near the opposite ends of the winding. To diminish this distributed capacitance, the turns should be spaced and wound in such a manner that points at a great difference of potential are kept apart. From this consideration, it is seen that a longer coil has less capacitance than a shorter coil of the same inductance because, in the shorter coil, the ends are closer together. Multi-layer coils, in general, have greater distributed capacitance than single-layer coils. Of these, the closely wound layer winding has the most capacitance, the banked winding, a deep narrow winding or a spaced layer winding all having less than the close-wound layers. With the flow of direct current or low frequency current through an inductor winding, there is practically a uniform distribution of current in the winding conductors. With increasing frequency, however, there is a change in current distribution in each conductor and in the whole winding. This causes a change of the flux distribution through and about the conductors. As an indication of the effect on coils of currents in the radio broadcast range, some curves of resistance variation are given in Fig. 5 for an inductance value of 291 microhenrys at 1 kc/s on hard rubber formers.



RESISTANCE VARIATION FOR INDUCTORS AT RADIO RANGE FREQUENCIES. (COILS 291 μH AT 1,000 c/s.)

A is a Four Bank Coil (32/38 Litz).
B is a Two Bank Coil (32/38 Litz).
C is a Single Layer Coil (32/38 Litz).
E is a Single Layer Coil (28 A.W.G.).
F is a Two Bank Coil (28 A.W.G.).
G is a Four Bank Coil (28 A.W.G.).

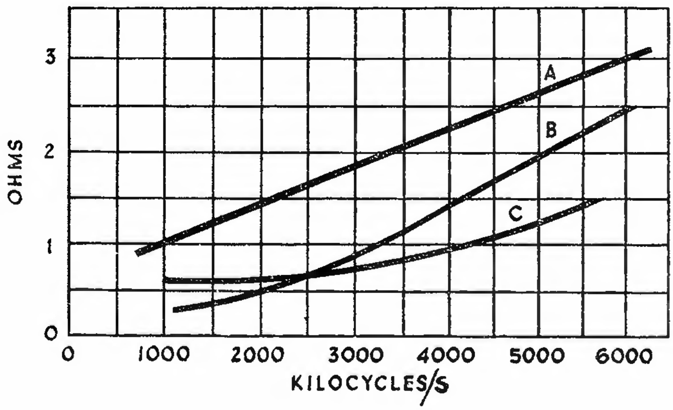
FIG. 5.

Fig. 6 shows the resistance variation of inductor windings between 1,000 and 6,000 kc/s. The coils were space-wound on a 4-1/8 inch former with 5 turns per inch, giving an inductance value of 10 μH at low frequencies. It will be noted that solid conductors give better performance above about 2,000 kc/s.

Cores for Inductances. The advantages of ferromagnetic cores in increasing and concentrating the flux linkage with a coil for a given current are well known. The difficulties found in their use with radio frequency currents arise principally from the loss of energy due to eddy currents in the core. Two effects, mainly, are produced -

(i) The apparent resistance of the coil is greatly increased due to power losses.

(ii) The flux produced by the eddy currents is in the opposite direction to that produced by the magnetising current in the coil. The resultant flux is much smaller than with a direct current of the same value, that is, the permeability seems lower.



RESISTANCE VARIATION OF INDUCTORS.

Each coil 4-1/8" diameter, 8 turns, 5 per inch.
L = 10μH at 1 kc/s.
A is 30 A.W.G., 0.89 ohms, D.C.
B is 96/38 Litz, 0.07 ohms, D.C.
C is Aluminium Ribbon, 0.44 ohms.

FIG. 6.

The eddy current losses increase with frequency, and so the inductance of the coil decreases with increasing frequency.

Various methods of powdering the material are sometimes used to improve the characteristics of iron cores.

Use of "Litz" Wire. One of the construction problems in the use of litzendraht wire for coil is that frequently strands of wire are broken or are not connected into the circuit. The effect on the coil characteristics is not serious when a few strands are disconnected, but the disconnection of over half the strands causes about 100% increase in the radio frequency resistance of a coil.

This effect is shown in Table 2 on a coil of 32/38 litz wire at a frequency of 750 kc/s.

INCREASE IN RESISTANCE OF LITZ WIRE DUE TO BROKEN STRANDS.

No. of Broken Strands	0	6	12	18	24	29	30	31
Resistance (Ohms)	3.1	3.4	4.4	6.1	9.5	21.7	42.4	51.6

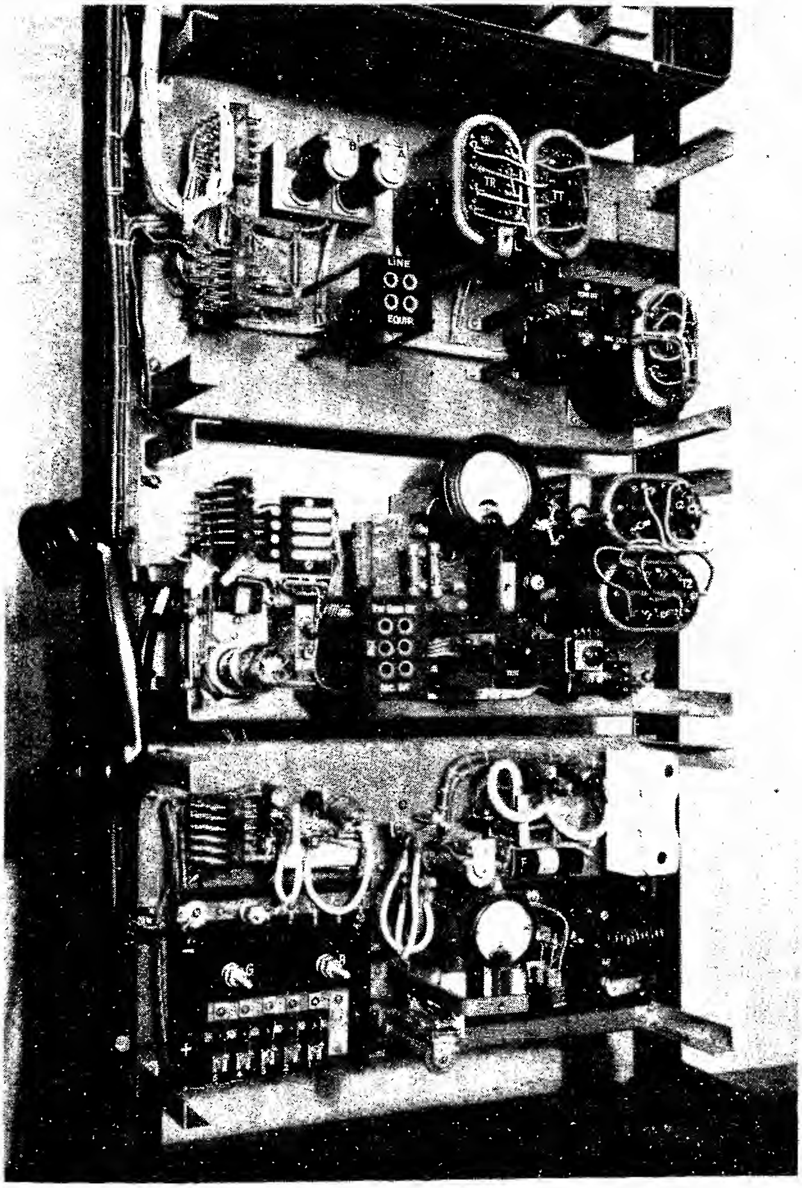
(The increases should be noted)

TABLE 2.

5.4 Capacitors or Condensers. The solid dielectric condensers which find their application in communication circuits generally have either mica or an impregnated paper as the dielectric. Because of the imperfections of dielectrics, there is a change in the condenser capacitance with frequency. The capacitance of a condenser decreases as the frequency is increased. However, the amount of change cannot be predicted accurately from a knowledge of the phase difference due to dielectric loss. This loss is generally the major cause for energy dissipation in the solid dielectric condenser, and it can be said only that the change in condenser capacitance is great when the dielectric dissipation is great.

When the leads inside the case of a condenser are long enough to have appreciable inductance, the capacitance measured at the terminals appears to be greater than it actually is, due to the lead-in inductance.

/Many



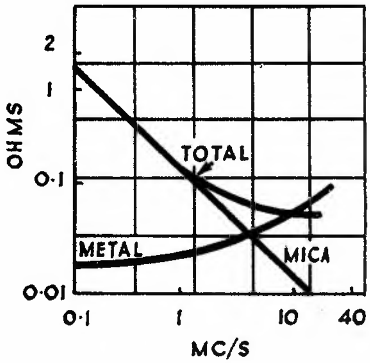
MANY TYPES OF RADIO APPARATUS MAKE UP THIS LOW POWER U.H.F. RADIO TELEPHONE UNIT.

How many different types can you identify?

Fig. 7 shows the variation of equivalent series resistance of a 400 μμF mica condenser with frequency, the condenser having a metal case or other high loss insulating material.

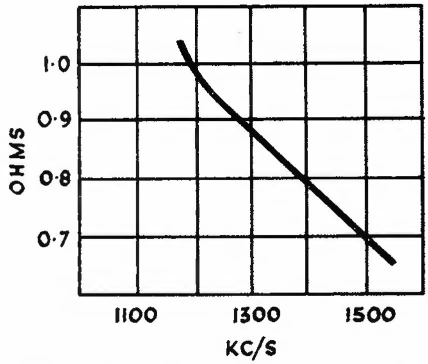
Fig. 8 shows the variation of equivalent series resistance of a variable air condenser with frequency with condenser set at 500 μμF.

/Fig. 7



VARIATION OF LOSS IN MICA CONDENSER (400 μμF IN METAL CASE) WITH FREQUENCY.

FIG. 7.



VARIATION OF EQUIVALENT SERIES RESISTANCE OF AIR CONDENSER WITH FREQUENCY.

FIG. 8.

The effect of dielectric losses, etc., in increasing the resistance of a capacitive circuit have already been discussed. In order to bring out the most important point about capacities at high frequencies, that is, the low reactance of even minute capacities and the eddy shunt paths they provide, Table 3, comparing the reactances of a condenser of 0.001 μF and an inductance of 100 microhenrys, is given -

Frequency	Reactance of Condenser 0.001 μF ($X_c = \frac{10^9}{2\pi f}$)	Reactance of Inductor 100 μH ($X_L = 2\pi f \times 10^{-4}$)
100 c/s	1.59 megohms	0.063 ohm
1 kc/s	159,000 ohms	0.63 ohm
1 Mc/s	159 ohms	630 ohms
100 Mc/s	1.59 ohms	63,000 ohms

TABLE 3.

6. SHIELDING AND SCREENING.

6.1 The successful operation of laboratory instruments, radio receivers, transmitters and studio equipment depends upon the effectiveness of the shielding between the various parts. Insufficient shielding will result in interference due to unwanted inter-coupling between units, wiring and components and to the effects of external fields. Shielding has to be provided against the effects of both electrostatic and electro-magnetic fields, and is particularly necessary in circuits carrying high frequency currents or very low level currents, such as microphone outputs.

6.2 Magnetic and Electrostatic Fields can be confined to restricted spaces, or can be prevented from entering a particular space, by the use of suitable shields. The most effective means of controlling magnetic fields at radio frequencies is by the use of shields consisting of a good conductor, such as copper or aluminium. Magnetic flux penetrates such a shield only with great difficulty because, as the flux cuts into the conducting material, it produces eddy currents that oppose the penetration. To be most effective, such shields must completely enclose the space to be protected, and all joints should be well lapped to minimise the resistance offered to the eddy currents. Where the shielding must be extremely effective, as

/in

in signal generators, joints should be soldered wherever possible and, in other cases, should be literally watertight. The thickness should be a number of times the skin depth of current penetration where the latter is given by the formula -

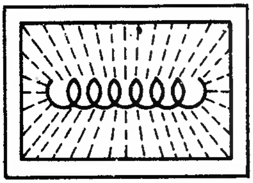
$$\text{Depth of penetration (centimetres)} = 5033\sqrt{\frac{\rho}{\mu f}}$$

where ρ = ohms per cubic centimetre,
f = frequency in cycles per second, and
μ = permeability of conductor.

6.3 Shielded Coils. One of the most important examples of a conducting shield is the copper or aluminium cover frequently placed around air-cored coils. Such shields serve to confine most of the magnetic field to the space within the cover. These covers also reduce the effective inductance of the coil by restricting the cross-sectional area through which the flux can pass, and increase the losses as a result of the energy dissipated in the shield by the eddy currents. However, if the clearance between the coil and its shield is at least equal to the radius of the coil, the reduction in inductance and the decrease in coil Q produced by a copper or aluminium shield will not be serious.

Design formulae are published in many books relating length and diameter of coils, shields and frequencies.

6.4 Magnetic Shields for Low Frequencies. Shields for unidirectional and low frequency magnetic fields are made of magnetic material, preferably having a high permeability. Such shields act as low reluctance paths for the flux, thereby diverting the flux from the space to be shielded. Fig. 9 represents a coil carrying current, the whole completely enclosed in a ferromagnetic shield box. Lines of force from the coil take the shortest path to the shield and travel through it, the latter acting as a magnetic short-circuit preventing the flux from spreading outside the screening box. This method is commonly seen in the audio frequency stages of radio receivers and elsewhere when the frequency is below about 15 kc/s. It is of prime importance to minimise voltages induced into audio circuits and equipment by the magnetic fields of power frequencies such as would be associated with power transformers and chokes. By placing transformers in ferromagnetic boxes or covers (as in Fig. 9) or in short cylinders of similar material with closed ends, a good degree of shielding is provided. The degree of protection obtained in this way depends on the size and thickness of the shield and upon whether the shield consists of a single layer or several concentric layers separated by air spaces.



FERROMAGNETIC SHIELD.

FIG. 9.

When the degree of shielding required is greater than that obtainable from a single shield, two or three concentric magnetic shields are employed, as shown in Fig. 10. When the total thickness of such a shield is greater than one-third the radius of the equivalent sphere, this subdivision of the magnetic material into concentric layers separated by air spaces gives more effective shielding than if the air spaces were filled solidly with magnetic material. The effectiveness of concentric magnetic shields to alternating currents can be greatly increased by placing copper shields in the air spaces between the magnetic shields. The behaviour in a typical case with and without copper shields is shown in Fig. 11. With direct current, the copper has no effect upon shielding.

The effectiveness of magnetic shields depends primarily upon the initial permeability of the metal used. Permalloy and similar magnetic materials having high initial permeability are accordingly superior to ordinary cast iron or silicon steel, particularly in the case of multilayer shields.

A magnetic shield enclosing an air-cored coil causes the inductance to increase as a result of the lowered reluctance offered to the flux paths by the high permeability material. The effective resistance of the shielded coil is also increased because of eddy current and hysteresis losses in the iron.

/Fig. 10.

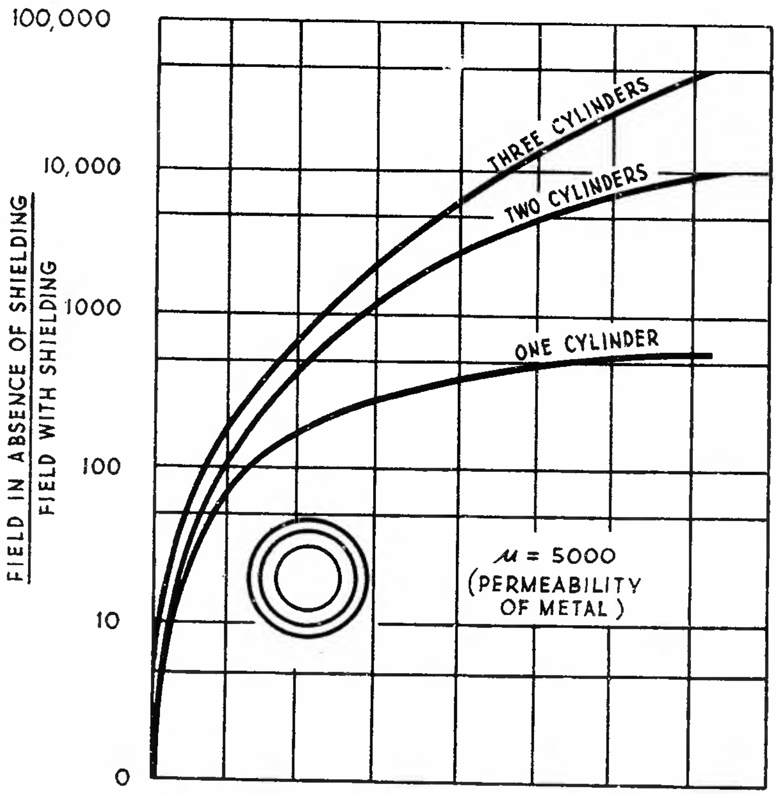


FIG. 10. EFFECT OF TREBLE SHIELDING.

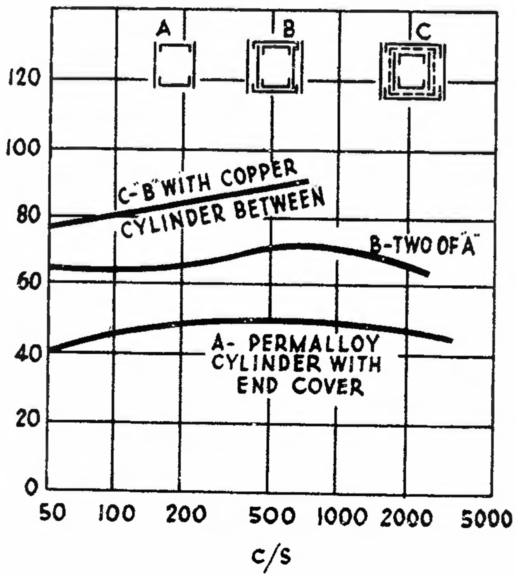
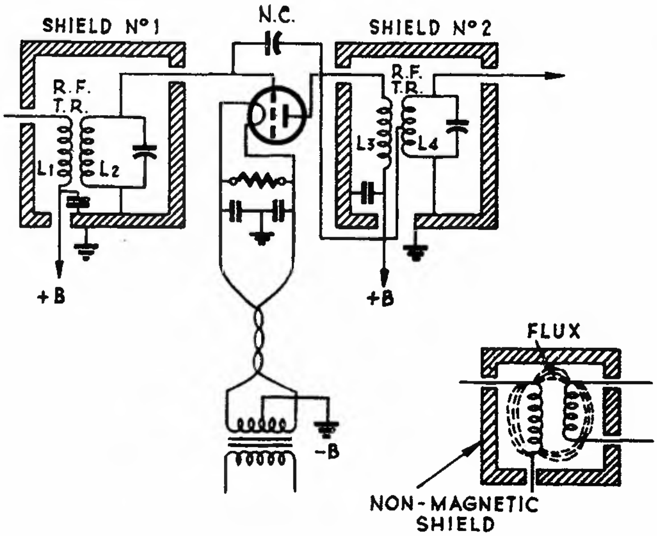


FIG. 11. EFFECTIVENESS OF SHIELDS.

6.5 Magnetic Shields for Radio Frequencies. At radio frequencies, it is found that more effective shielding is provided by employing a different principle. The ferromagnetic screening box is replaced by a non-magnetic metal of high conductivity, and shielding is achieved by eddy current action. Fig. 12 shows the arrangement of shielding in a radio frequency amplifier circuit. The magnetic fields that are produced around the radio frequency coils L1L2 and L3L4 when an alternating current signal is applied to the circuit are confined almost entirely to the enclosed compartments of the shields as shown. The action may be summarised as follows -



TYPICAL EXAMPLES OF NON-MAGNETIC SHIELDING.

FIG. 12.

When a varying magnetic field is moving about the radio frequency coils, the flux will tend to spread out in a more or less circular movement, in which the distance of its travel depends upon the intensity of the currents flowing in the respective coil circuits. If, therefore, the fields are strong enough and the two radio frequency transformers are in the magnetic axis of each other, instability and self-oscillation will result. If, however, the coils are enclosed in a metallic compartment of non-magnetic material, such as copper, aluminium or brass, and, assuming that the compartment forms a complete electrical circuit, the varying magnetic fields around the coils will tend to displace electrons in the shield at the particular frequency in which the magnetic field is varying. This action generates eddy currents which, in turn, produce magnetic fields and oppose the field created by the high frequency currents circulating in the coils. This opposing effect varies directly with frequency, so that, as the frequency increases, the shielding effect is greatly improved.

The efficiency of the shielding depends a great deal upon the high conductivity of the material used, so that it is essential to ensure that all ends of the compartment are thoroughly soldered to form a complete current path. Fig. 13 compares the magnetic and non-magnetic shields.

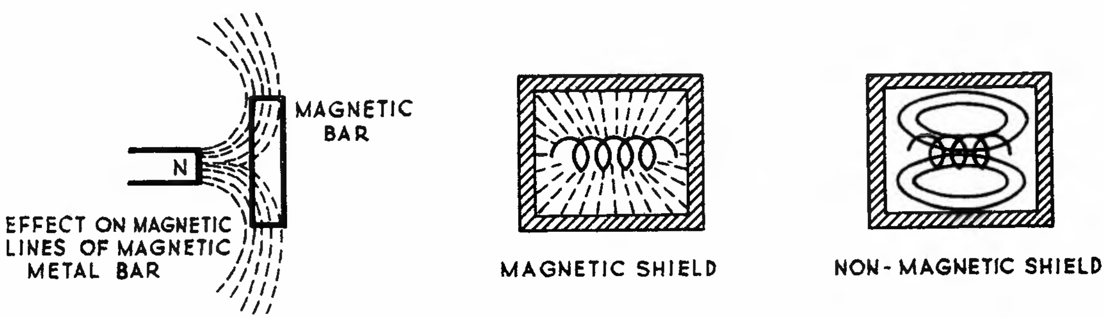


FIG. 13. MAGNETIC AND NON-MAGNETIC SHIELDS.

6.6 Electrostatic Shields. Any conductor can be used as an electrostatic shield. There are no particular requirements as to thickness or conductivity. It is necessary merely that the electrostatic fields terminate on the shield instead of passing through to the space being protected. To be most effective, electrostatic shields should be continuously closed surface, but even a screen will give fair shielding. The conducting and magnetic shields for magnetic flux discussed above, therefore, simultaneously provide electrostatic shielding along with the magnetic shielding.

Electrostatic Shielding without Magnetic Shielding. Electrostatic shielding can be obtained (without affecting magnetic flux) by surrounding the space to be protected with a conducting non-magnetic shield arranged to provide termination for electrostatic flux lines without providing closed loops of low resistance around which eddy currents can circulate. Such a shield is shown in Fig. 14. (This shield is also known as a "Faraday" screen or shield.)

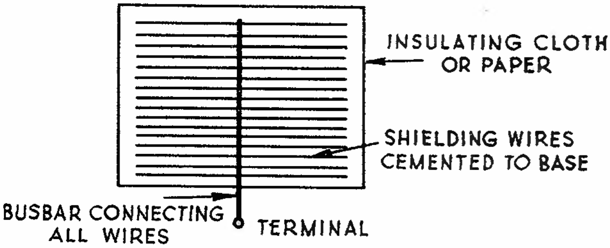


FIG. 14. ELECTROSTATIC (FARADAY) SCREEN.

Electrostatic shielding is frequently used in transformers to prevent electrostatic coupling between the various windings. Such a shield consists of a metal foil wrapped around the windings in order to enclose the windings completely, but with an insulated lapped joint (as shown in Fig. 15) so that the foil does not act as a short-circuited turn. Electrostatic shielding replaces direct capacity coupling between coils by capacity from the individual coil to the shield (that is, to the earth).

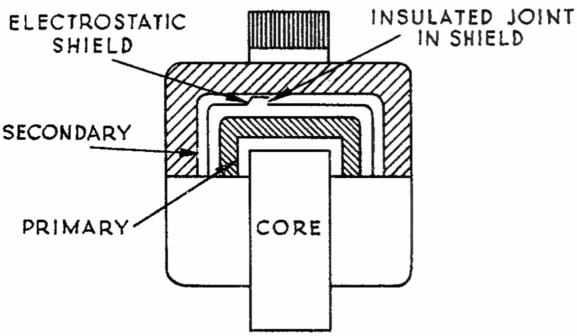
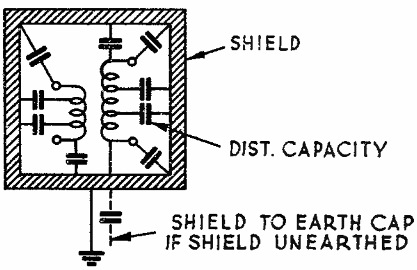


FIG. 15. BALANCED SECONDARY WINDING WITH ELECTROSTATIC SHIELD.

Double and treble shielding of this nature is sometimes required, as in very low level audio-circuits, for example, microphone amplifier input circuits.

6.7 Effects of Shielding. The inductance of a coil will decrease when enclosed in a non-magnetic shield, owing to the reduction of the magnetic flux around the coil. This decrease in the inductance takes place because the magnetic fields set up by the eddy currents in the shield circuit are in opposition to the coil flux, resulting in a reduction of the total lines of force. Hence, since the inductance of a coil depends upon the amount of energy that is stored in magnetic form, any decrease in the number of the magnetic lines results effectively in a decrease of the inductance.



ILLUSTRATING DISTRIBUTED CAPACITY OF A COIL IN A SHIELD.

FIG. 16.

The electrostatic capacity, however, increases when a coil is placed in a non-magnetic shield, owing to the relationships of the various parts of the coil to the shield (see Fig. 16). The capacity effect may greatly increase the effective resistance of the coil, particularly at high radio frequencies, since the condenser effect introduces additional energy dissipation or loss. These losses may be reduced by arranging that the shield be equivalent from all portions of the coil with a spacing at least equal to the radius of the coil. The shields must be efficiently earthed to avoid a "floating" effect of the shield above the earth potential.

/If

If the shielding is not earthed, an electrostatic field may exist between the shield and earth, owing to the potential difference between these points. If, therefore, any wires or circuit parts are situated in this field, serious instability may result owing to the effect of the field upon the associated circuits.

6.8 Miscellaneous Points to Reduce Instability Effects.

(i) Connect all common return leads to one point in as direct a manner as possible to reduce voltage drops, particularly as the frequency increases.

(ii) Use a busbar or copper-ribbon strip (not the chassis) for common earth returns.

(iii) Keep grid and anode leads as short as possible, so that magnetic or electrostatic fields may not exist between these leads with preceding or succeeding radio frequency stages.

(iv) Shield radio frequency valves, particularly high gain valves.

(v) Completely shield all high frequency units for maximum degree of efficiency and stability.

(vi) Shield all low frequency and power components with magnetic shields.

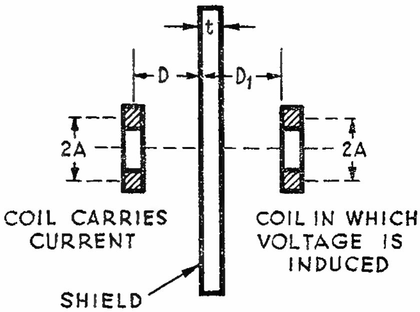


FIG. 17. PLANAR SHIELD.

6.9 Planar Shield. An important case of shielding exists where a flat sheet of conducting material is used to reduce the voltage induced in a second coil by current passing through a first coil. Fig. 17 shows a typical case where the two coils are coaxial and similar and the shield is a sheet placed at right angles to the axis. To obtain best results, the length and width of the shield should be considerably greater than 2A or D + D1, whichever is larger.

7. TEST QUESTIONS.

1. What methods are adopted in practice to reduce the effects of capacity and inductance in resistances used for high frequency measurements?

2. A condenser has five plates, each measuring 5 centimetres by 2 centimetres. These plates are separated by mica plates 0.1 millimeter thick and having a dielectric constant of 9. Calculate the capacitance in micro-microfarads.

3. A condenser of eleven rectangular plates, each 2 inches by 1 inch, separated by mica plates 4 mils thick, has a capacitance of 4,000 $\mu\mu$F. What is the dielectric constant of the mica?

4. A variable condenser has a maximum capacitance of 1,000 $\mu\mu$F and a minimum capacitance of 100 $\mu\mu$F. When in the maximum position, the condenser is charged to 1,000 volts. The charging supply is then disconnected and the condenser turned to the minimum position. What was the original energy and the final energy stored in the condenser? Explain the reason for any difference.

5. Describe the construction and explain the action of one type of electrolytic condenser. How does the working voltage affect the capacity? Why is a direct current potential necessary?

6. Two condensers of capacities 2 μF and 3 μF are in series and a voltage of 500 volts is applied. Find the equivalent capacity, the charge in each condenser and the division of the voltage between them.

7. The apparent resistance of conductors to the flow of high frequency alternating currents differs from that for direct current. What term is used to express this resistance to distinguish it from the direct current resistance? Explain why the apparent resistance varies with frequency.

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

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

THERMIONIC VALVES.

CONTENTS:

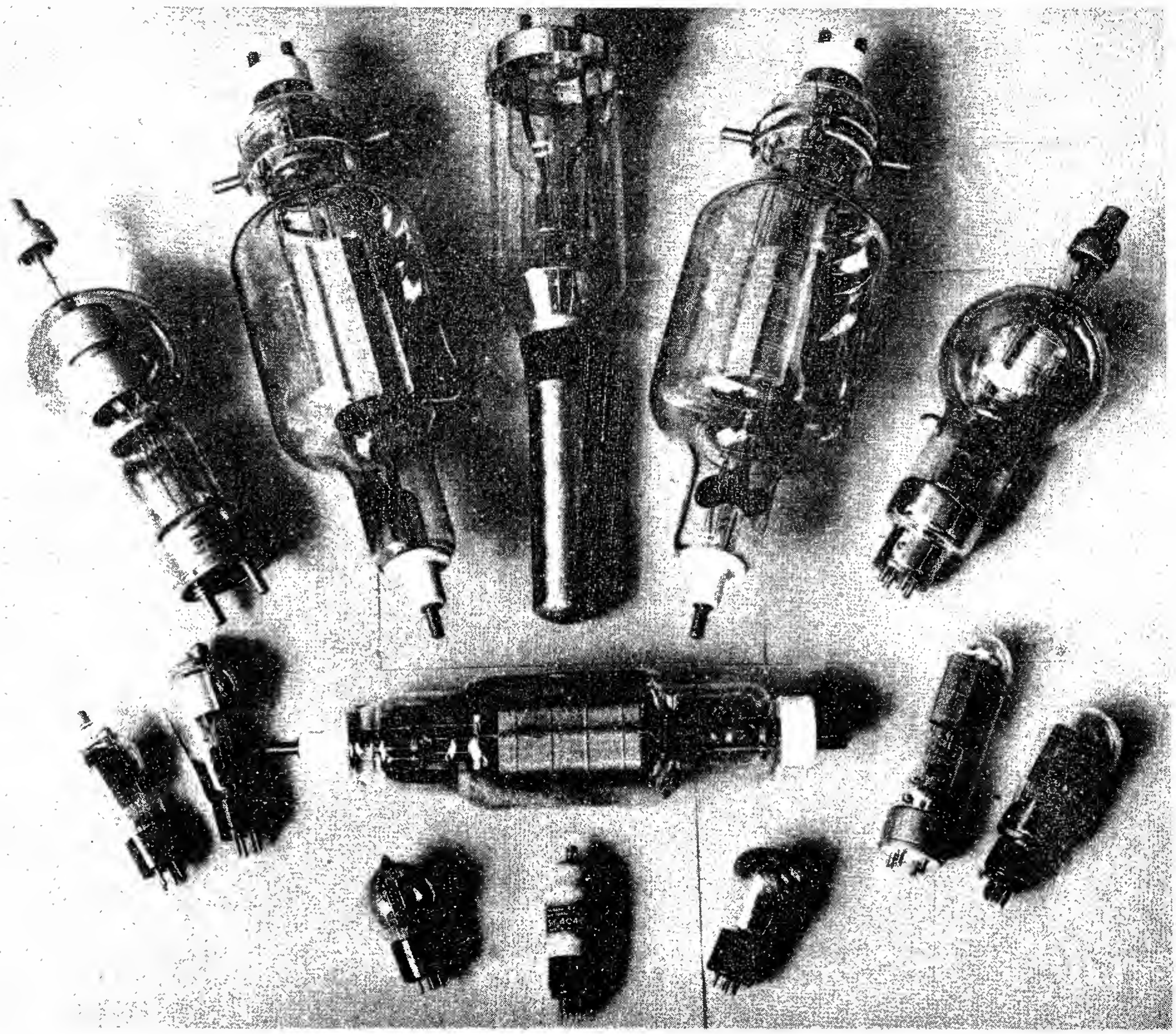
1. THERMIONIC EMISSION.

1.1 The basic theory of electron activity was referred to in Paper No. 1 and the applications of this theory to thermionic valves are dealt with in greater detail. There are several methods of producing electron emission, but Thermionic Emission is the immediate interest. As the name implies, thermionic emission is electronic emission due to heating of the emitting element, which is usually metallic.

1.2 In a conductor, a large number of free electrons move around with various velocities. As the temperature of the conductor increases, the average velocity of the free electrons increases. At the surface of the conductor, there exists a restraint on the movement of the electrons in the form of an electrostatic force. This restraint is analogous to surface tension in a liquid and tends to prevent 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 causes evaporation. The breaking through of the electrons appears to be due to an increase of the kinetic energy of the electrons - only those possessing a velocity greater than a certain minimum value succeed in escaping. The amount of work represented by this escape is known as the "thermionic work function," and represents the minimum amount of kinetic energy the electron must have in order to pass through the surface. The energy (in electron volts) that an electron must have before it can leave a metal is known as the "Work Function" of the metal. The emission of electrons thus depends on -

(i) The temperature of the emitting metal, and

(ii) The work function of the metal.



ALADDIN'S LAMPS OF RADIO.

The source of the electron emission is known as the "CATHODE" and, as stated later, cathodes may be directly or indirectly heated.

1.3 Cathodes. Most pure metals are capable of electron emission but, because of their relatively low emission rate, are seldom used. Tungsten, however, is an exception. Tantalum is the only metal other than tungsten with possibilities as a practical emitter. Although the melting temperature of tantalum is 3,300° K as against 3,655° K for tungsten, the function of tantalum is such that at any temperature less than 2,500° K the electron emission is at least 10 times that for tungsten. A disadvantage of tantalum is that it is easily contaminated by residual gases, which form oxides that greatly reduce emission. (K, that is, kelvin or absolute scale, is obtained by adding 273.1° to the reading of the centigrade scale.)

Tungsten Cathodes. Tungsten, although a relatively poor emitter, can be operated at very high temperatures. Tungsten is extensively used as the cathode of high power valves, because of its durability under the exacting conditions in such valves, particularly as regards the positive ion bombardment which is associated with high power triodes. This bombardment, in the case of coated cathodes, would result in the disintegration of the emissive coatings with the consequent loss of efficiency of the valve. Tungsten cathodes operate at temperatures between 2,400 and 2,500° K with currents ranging from milliamperes to hundreds of amperes.

Thoriated Tungsten Cathodes. There are many metals with a higher emission rate than tungsten which cannot be used because of low melting temperatures. It is possible, however, to place a thin layer of such a metal on tungsten, and thus take advantage of the high melting temperature of tungsten and the high emission rate of the other metals. Thorium on tungsten gives the highest emission of any combination of this character. The layers of thorium resulting from the formation procedure are molecular in thickness. Thoriated tungsten emitters are formed by using tungsten wire in which from 1 to 2 per cent. of thorium oxide has been added. The thoriated cathode is placed in valves in the usual way and then subjected to an activating process. Normal operating temperatures are from 1,800 to 2,000° K.

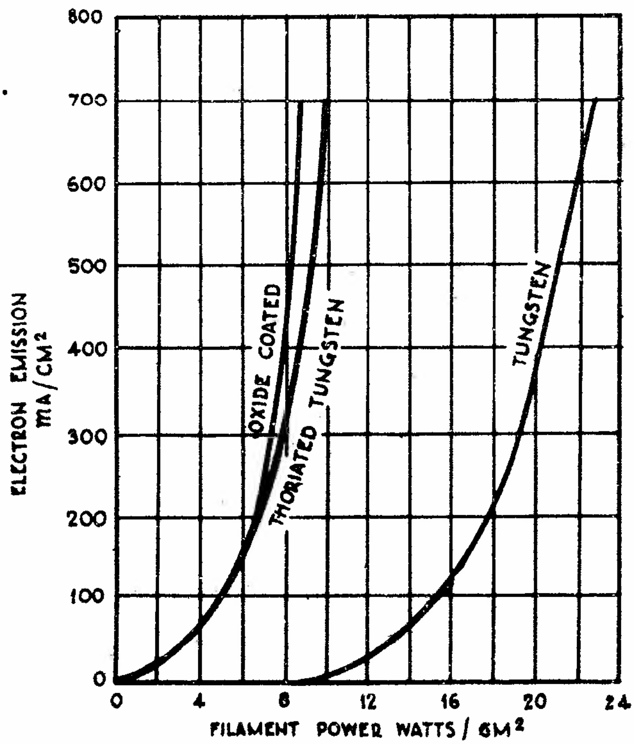


FIG. 1. EMISSION PROPERTIES.

Oxide Coated Cathodes. The most efficient type of emitter is the oxide coated cathode, which consists of a metallic base of some pure metal or alloy with a coating of the oxides of some of the alkaline earth metals. The base metals, known as cores, on which the oxide coating is formed may be of pure metal, such as platinum. Better emission results are, however, obtained from certain alloys, such as platinum-iridium, nickel-platinum and nickel-silicate. The best emission results from the use of an alloy of nickel, iron, cobalt and titanium, known as Konel metal. The best alkaline earths are barium, strontium and calcium, which are copious emitters of electrons at moderate red heat. The coating is applied to the core in the form of carbonates and subjected to an activation procedure.

Cathodes coated with these oxides have a very long life, a high emission rate at temperatures of about 1,100° K, and are economical in cathode power.

Comparison of Emitters. Fig. 1 shows the relative emission properties of some of the common types of emitters. / The

The emission in milliamperes per square centimetre is plotted against heating power in watts per square centimetre.

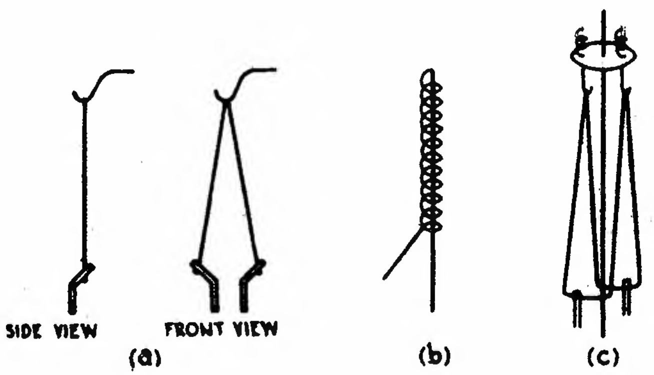
Practical Cathodes. Cathodes are in two main classes -

(i) Directly heated cathodes, and

(ii) Indirectly heated cathodes.

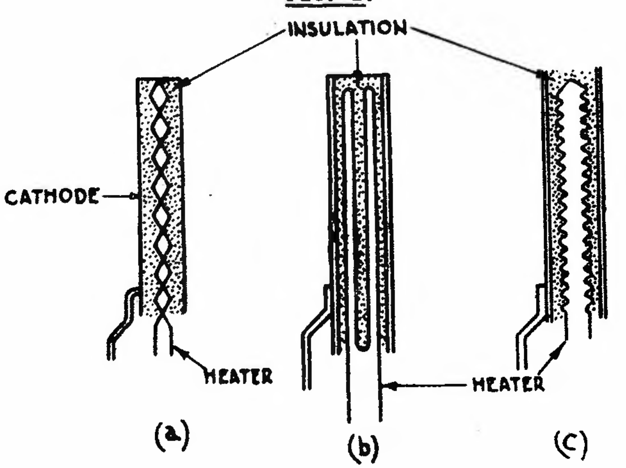
Directly Heated Cathodes. In these cathodes, the metal (tungsten, oxide-coated, etc.) is in the form of a filament suitably supported, and the heating current passes directly through the filament, which thus acts as the emitting source. In most cases, the heating source must be direct current, as alternating current would cause a noise in radio circuits. Some types of direct heated cathodes are shown in Fig. 2. These cathodes are arranged so that the temperature is as high as possible for a given input power. All the cathodes have spring suspensions that allow for the expansion of the filament wire.

Indirectly Heated Cathodes. In the directly heated cathode type valve, hum is caused when the cathode is heated by alternating current. The thermal capacity of the cathode is such, due to its small diameter, that it may cool appreciably between alternate peaks of current causing a rise and fall in emission at twice the supply frequency. (In high power valves with large diameter cathodes cooling between alternations is not appreciable.) Hum also results because the ends of the cathode are at different potentials, and thus one end acts as a grid to control the space charge at the other end.



THREE TYPES OF DIRECTLY HEATED CATHODES.

FIG. 2.



THREE TYPES OF INDIRECTLY HEATED CATHODES.

FIG. 3.

In the indirectly heated cathode type of valve, the functions of emission and heat supply are separated (see Fig. 3.). An indirectly heated cathode consists of a cylinder of thin sheet nickel with an emitting coating which contains the heater wire. The heater wire (generally called the heater) is invariably tungsten covered with some insulating coating, such as aluminium oxide.

The cathode is brought to emitting temperature by either conduction or radiation from the tungsten heater. Other insulations used are silica and porcelain.

Summary of Cathode Materials and Operating Characteristics.

(i) Tungsten Cathode. Used mainly in indirectly heated valves and high power triode valves. Currents range from milliampers to hundreds of amperes. (Average life 1,000-2,000 hours.)

/ (ii)

(ii) Thoriated Tungsten Cathode. Used extensively in low and medium power radio circuits with anode voltages of 500 to 5,000 volts. Cathode currents range from 0.1 to several amperes. (Average life 2,000-4,000 hours.)

(iii) Oxide-Coated Cathode. Cathode currents about 0.075 to 0.5 ampere. Not satisfactory where anode potential exceeds 500 to 2,000 volts, because of detrimental effects of positive ion bombardment. (Average life 5,000 hours.)

Oxide coated emitters have longer life and greater emission per watt of heater power than other types, and are used wherever possible and almost exclusively in radio receiving valves.

1.4 Physical Construction of Other Elements. Grids and anodes of valves are made of metals, such as nickel and molybdenum. In the case of anodes, it is essential to choose a metal with a high melting point (to withstand the heat generated) and which does not exude occluded gases that might impair the valve's operation by reducing the vacuum. Nickel, molybdenum, tantalum, zirconium, copper and carbon are typical metals used for anodes.

1.5 Abbreviations. The abbreviations frequently used in this Paper are -

Z_L = load impedance.

E_{ZL} = volts across load.

I_a = anode current.

r_a = anode impedance.

E_g = input signal to grid.

E_o = output voltage.

g_m = mutual conductance.

μ = amplification factor.

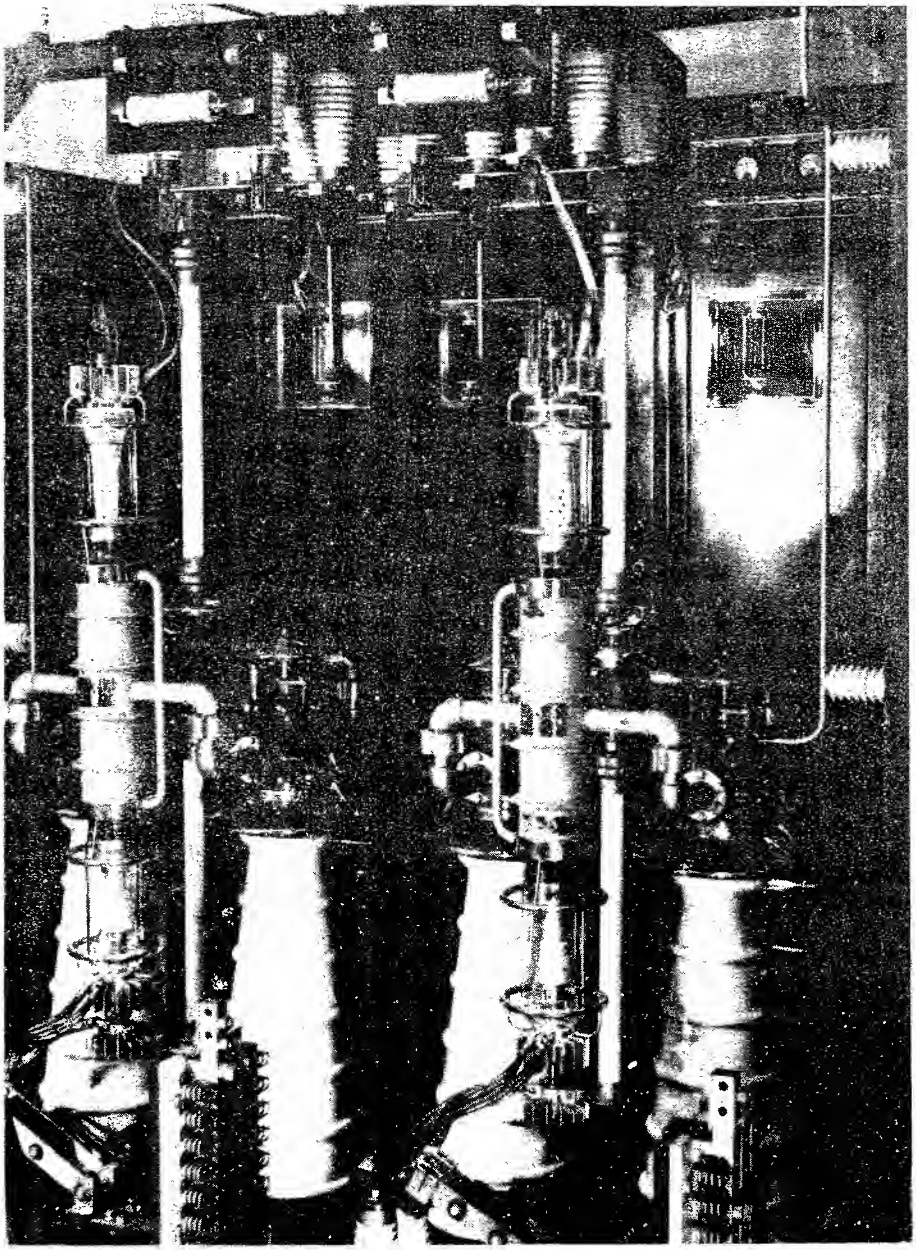
P_{max} = maximum power.

Other abbreviations are given throughout the Paper.

2. THERMIONIC VALVES.

2.1 The valve consists of a glass or metal container within which are two or more metallic parts known as the elements. After the elements have been placed within the valve, the air and gas are removed. During the removal process, the container is heated to drive the gas from it. Later, the elements are heated so that various gases in these metal elements may be removed. The modern valve is a high vacuum valve. When the exhaustion process is complete, the container is sealed and the remaining gas removed by the use of a "getter" (see Paragraph 10.5). Finally, the glass bulb or container is cemented to a suitable insulating base, in the bottom of which are four or more hollow brass rods which connect to the various valve elements and also make contact with the external circuits.

2.2 Basic Principles of Thermionic Valve Operation. It is recalled that electrons are negative elements and are thus attracted to a positively charged substance (that is, positively charged with respect to the emitter.) The resulting electron flow constitutes a current, and this current is maintained so long as the electron emission continues. An elementary valve thus consists of an emitter and a collector (that is a cathode and an anode). Such a valve is known as a "diode".



VALVES IN ACTION.

Clarity of reception in distant countries is determined by this unit - the modulator - of Australia's 100 kW short-wave transmitter VLA at Shepparton, Victoria. In the background are the self-locking safety doors.

Types of Valves. Valves may be broadly grouped, according to the number of elements they contain, as follows -

Diode, two elements, comprising the cathode and anode.

Triode, three elements, comprising the cathode, anode and one grid.

Tetrode, four elements, comprising the cathode, anode and two grids.

Pentode, five elements, comprising the cathode, anode and three grids.

Hexode, six elements, comprising the cathode, anode and four grids.

Heptode, seven elements, comprising the cathode, anode and five grids.

Octode, eight elements, comprising the cathode, anode and six grids.

There are also a number of compound types referred to later.

3. DIODE VALVES.

3.1 The Diode or Two-Electrode valve is the simplest type of valve and consists of an exhausted glass bulb containing two elements - an emitter or cathode and a collector or plate (anode). When this valve is connected, as shown in Fig. 4, the ammeter indicates a current flowing in the circuit. This current flows because the positive anode attracts the electrons emitted by the cathode, so that they move to the anode and thence flow around the external circuit through the meter back to the cathode. The amount of anode current flowing for a given cathode temperature depends on the value of the anode potential. At low potentials, the anode field is weak and only attracts a relatively small number of electrons which have reached the extremities of the space charge. As the anode potential is increased, however, the attractive power of the anode increases and more electrons are attracted to it, resulting in a higher value of current in the circuit. A point occurs at which the anode potential is attracting every electron emitted by the cathode, and an increase in anode potential does not bring about an increase in current. This effect is shown in Fig. 5, where the anode current is seen to rise rapidly until the curve tends to flatten out. This point is known as the saturation point of the valve at the particular cathode voltage and is shown as point "x" on the curves.

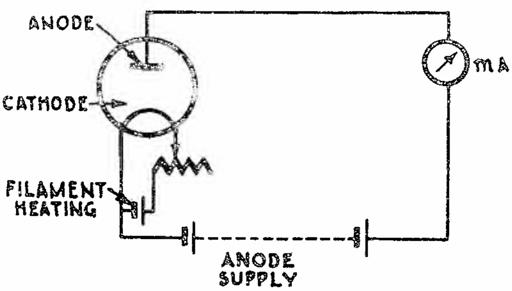


FIG. 4. DIODE VALVE.

When the cathode voltage is increased, an increase in emission results and the saturation point is extended to a higher anode voltage, curves B and C. There is a limit to the cathode potential which, if exceeded, results in the collapse of the cathode.

A curve showing how the current through any device increases as the voltage increases is called a "Characteristic Curve." In Fig. 5, the curves show the effect of anode voltage on anode current and are, therefore, known as "Anode Characteristic Curves."

/ Fig. 5

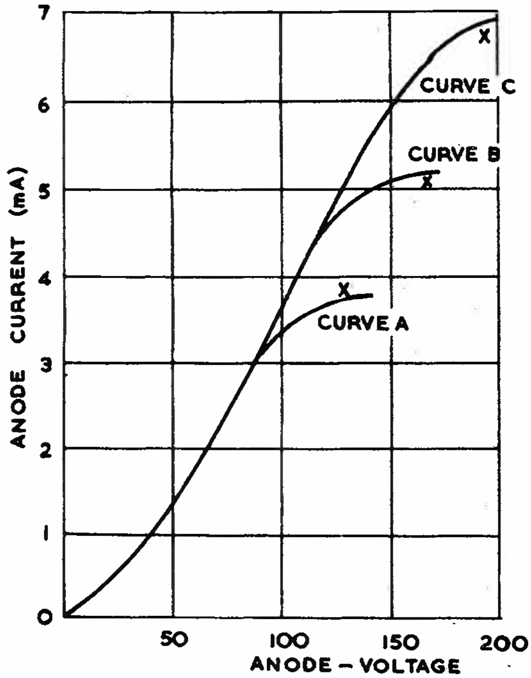


FIG. 5. ANODE CHARACTERISTIC CURVES.

Diodes are chiefly used as Detectors, Rectifiers in power converter circuits and Automatic Gain Control Voltage Rectifiers.

3.2 Space Charge. All electrons emitted by the cathode do not go to the anode, but many hover about the cathode in the form of a cloud. The density of this cloud depends to some extent on the anode potential, since for a given cathode potential the electron emission is constant. Only those electrons, however, which posses sufficient velocity to travel to the area where the influence of the anode potential extends, are attracted to the anode (which explains why low anode potentials result in low anode currents). Since the cathode is emitting electrons it must have a deficiency of negative charges, that is, it may be regarded as being positively charged with respect to the electrons and thus exerts an attractive force on them. This tends to keep a cloud hovering about the cathode, and this cloud, being negatively charged, tends to repel the electrons leaving the cathode, with the result that the actual current flow is limited by the repulsion between the electrons and not entirely by the emission capabilities of the cathode.

Variation of the anode potential varies the extent to which its influence penetrates the space charge field - the higher the potential the greater the penetration and the higher the resulting collection of electrons and current.

Modern design tendency is to manufacture valves with a surplus electron emission, and to make use of the space charge effect to obtain a limiting equilibrium condition which sets a limit to the current that can flow for a given potential difference between the electrodes. Under these conditions, the current is said to be "space-charge limited." In this case, the space-charge may be likened to an electron reservoir from which the anode draws electrons in accordance with its potential, the reservoir also having a cushioning effect on small variations of the filament supply potential.

/ 4.

4. TRIODE VALVES.

4.1 When a third electrode is placed between the anode and cathode of a diode valve, it offers a means of controlling the flow of electrons between the cathode and anode. To permit the passage of electrons between the two latter electrodes, the added electrode must be open in construction and is referred to as the Control Grid. As a positive charge applied to the anode attracts the electrons emitted by the cathode, a similar charge applied to the grid would cause the electrons to be attracted to the grid. However, due to the open structure of the grid, the majority of electrons pass through the grid and are attracted to the anode.

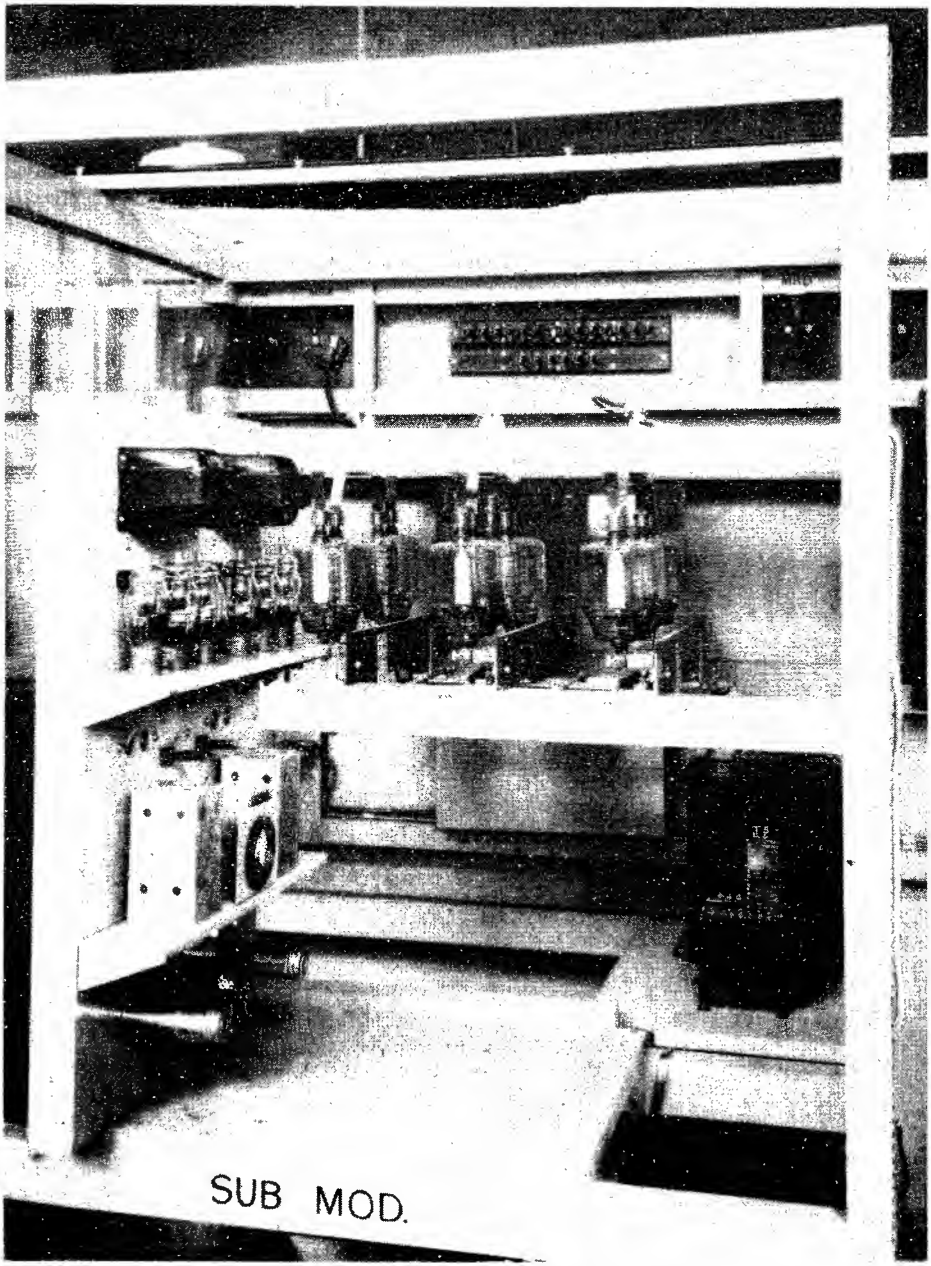
Similarly, a variation in grid potential would cause a variation in the number of electrons attracted, as in the case of anode potential variations. Owing to the fact that the grid is nearer to the cathode than the anode, the effect of slight changes of grid voltage produces the same effect on the electron flow as a large change in anode voltage.

When the grid potential is negative in respect to the cathode, this negative potential tends to repel electrons emitted by the cathode and so reduce the anode current flow below the normal value, this being the anode current flowing with zero grid potential. When the grid potential is made sufficiently negative, anode current does not flow, as the field due to the grid plus the space charge effect neutralises the field due to the anode in the region close to the cathode. The potential applied to the grid is usually termed the Grid Bias.

It was shown in the case of the two-electrode valve, that, when the electron flow to the anode is less than that corresponding to the total emission from the cathode, the current to the anode depends upon the electric force which is acting near the cathode, that is, the current is limited by space charge. Under these conditions, the presence of the grid when it is at cathode potential, results in a lower value of the anode current than would flow if the grid were absent. Suppose that a battery (usually called the C battery) is inserted in the lead connecting the cathode and grid, by means of which the grid can be made positive or negative with respect to the cathode by any desired voltage. When the grid is made somewhat positive, it exerts an attractive force upon the electrons near the cathode, and a greater number leaves the cathode and starts across the valve. In fact, the current flow increases until the space charge effect neutralises the force of the anode as well as the force of the grid. Most of this electron current flows through the spaces between the grid wires to the anode, so that, as a result of a positive voltage on the grid, the anode current increases. A negative voltage on the grid produces the opposite effect, for then the force exerted by the grid opposes the force of the anode and the anode current is reduced. In fact the grid voltage can readily be made negative enough to stop completely the current flow through the valve.

It is evident, therefore, that the grid acts as a control over the current flow to the anode in that, by variation of the voltage of the grid, the current to the anode can be increased or decreased. The electrons pass through the grid to the anode and, unless the grid voltage is made positive, very few hit the grid wires. The current to the grid is, therefore, practically zero and only a small current flows from the C battery, so that little power is required to regulate the voltage of the grid.

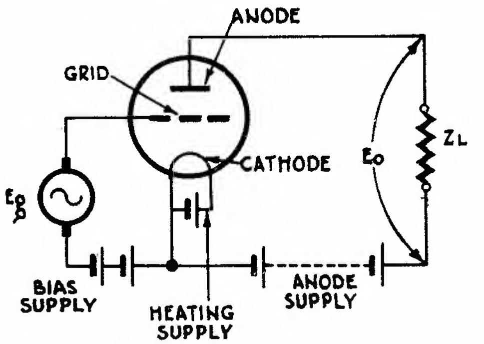
/ Modulator



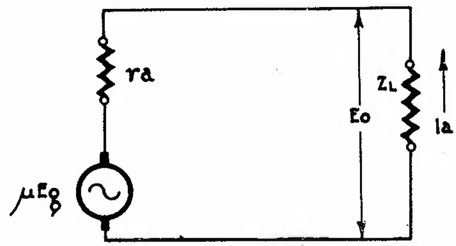
MODULATOR UNIT DURING CONSTRUCTION.

Lyndhurst Radio Station 5 kW Transmitter.

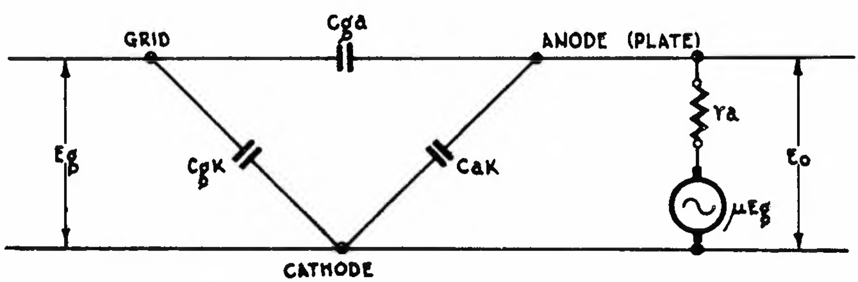
4.2 Fig. 6a is a schematic sketch of a triode, Fig. 6b shows the equivalent alternating current circuit with a resistance load, and Fig. 6c, shows the inter-electrode capacitances.



(a) Schematic Circuit.



(b) Equivalent A.C. Circuit.



(c) Inter-electrode Capacities.

ELEMENTS OF THE TRIODE VALVE.

FIG. 6.

These capacitances play an important part in the performance of triodes, and more information about these is given later in this Paper.

Fig. 6c indicates the following -

Grid-cathode capacitance (C_{gk}).

Grid-anode capacitance (C_{ga}).

Anode-cathode capacitance (C_{ak}).

The triode is studied in some detail, since, if the triode's principles are appreciated, the performance of other valves is more readily understood.

/ 4.3

4.3 Performance of Triode Valves. The performance of triodes is best studied by reference to certain parameters or constants and to what are termed "Characteristic Curves."

Parameters. The more important characteristics of a triode are expressed in terms of three parameters -

	Symbol
Amplification Factor	μ (Greek Letter Mu)
Anode (or Plate) Resistance	r_a
Mutual Conductance or Transconductance	g_m

Amplification Factor is a measure of the relative effectiveness of grid and anode voltages in producing electrostatic fields at the surface of the grid. The amplification factor depends on the spacing and size of the network of wires in the grid, that is, the closer the spacing the greater the screening effect of the grid on the electrostatic field of the anode. It also varies directly with the distances between the anode and the cathode and between the grid and the cathode. The nearer the grid is to the cathode, the smaller is the voltage which is needed to produce a field around the cathode equal to the field set up about it by the anode. Thus, a valve having a large amplification factor uses a fine grid mounted at a short distance from the cathode, as compared with the distance between the anode and the cathode.

The flow of electrons, therefore, is influenced by an electrostatic field which is the combination of the fields due to the anode and grid charges. The amplification factor is, therefore -

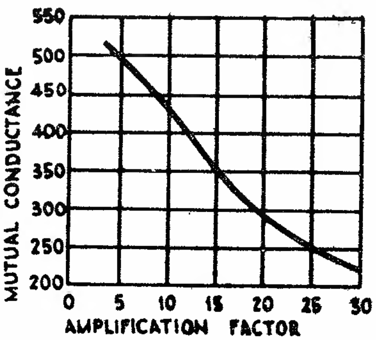
$$\mu = \frac{C_{gk}}{C_{ak}}$$

where C_{ak} is the electrostatic capacity of the anode and cathode,
and C_{gk} is the electrostatic capacity of the grid and the cathode.

This relation implies that the amplification factor is a constant and depends only on the valve structure and not on the operating voltages. It explains also why the curves showing the relation of grid voltage to anode current for various constant values of anode voltage are approximately parallel to each other. The amplification factor is not affected by those factors which influence anode resistance, such as electrode area and cathode condition, nor is it altered by changes in the applied voltage, except that at low anode voltages it may decrease slightly.

Actually, the amplification factor μ is not quite constant but varies with the grid and the anode voltages. The explanation of this effect is that a change in grid and anode voltages causes a change in the shape and location of the space charge. This, in turn, changes the effect of the grid and anode voltage on the flow of electrons.

The amplification factor μ is, however, practically constant in value over the straight portion of the characteristic curve. The value of the amplification factor of a valve expresses the relative effects of grid voltage and anode voltage on the anode current, and so determines the anode resistance of the valve. An increased amplification factor corresponds to an increased anode resistance, and vice versa. A change in the amplification factor also affects the Mutual Conductance (defined later) to some extent, even though the anode area, cathode length and other factors remain constant. A valve with a high amplification factor shows a lower mutual conductance than a valve of similar construction but with a lower amplification factor. This effect is shown in Fig. 7 for a number of valves with different amplification factors but using the cathode and anode construction of a UX-120 valve. Fig. 7 shows that a low value of the amplification factor is used in order to gain the advantage due to improved mutual conductance, provided that the impedance load can be adjusted to a suitable value.



RELATION OF AMPLIFICATION FACTOR TO MUTUAL CONDUCTANCE.

FIG. 7.

/ Such

Such conditions are conducive to maximum power output. For voltage amplification in circuits in which high anode resistance is not important, as in resistance - or impedance-coupled amplification, a high value of the amplification factor μ is desirable, because it allows an increase in voltage amplification to be obtained from each stage of the amplifier.

The Amplification Factor μ is a measure of the maximum voltage amplification obtainable from the valve alone. The grid-to-cathode voltage due to the reception of a radio signal appears in the anode circuit multiplied μ times. The voltage developed across a high-impedance load placed in the anode circuit is nearly equal to the value μE_g.

Amplification Factor is defined as the ratio of a small change in anode voltage to a small change in control-grid voltage under the conditions that the anode current remains unchanged and that all other electrode voltages are maintained constant.

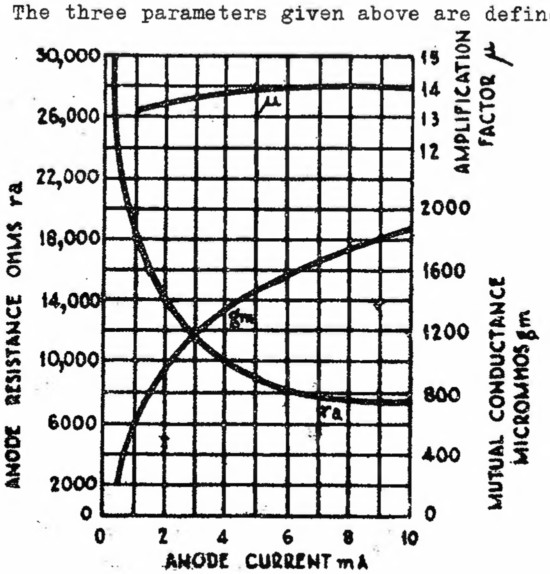
Anode (or Plate) Resistance. When this term is used, it normally refers to the dynamic or alternating current resistance of the valve, the direct current resistance being of little interest. The dynamic resistance of valves is important, since in operation they are equivalent to alternating current generators and, as such, must be matched into their load impedances. As any alternating current resistance depends on the rate of change of current and voltage, the dynamic resistance of a valve between anode and cathode is a function of the rate of change of anode current and anode voltage. Anode resistance, therefore, is regarded as the impedance offered to the passage of alternating current through the valve.

Anode Resistance is defined as the ratio of a small change in anode voltage to the small change in anode current produced, subject to the remaining electrode voltages being maintained constant. The anode resistance depends upon the dimensions and relative positions of cathode, anode and grid.

Mutual Conductance. One of the most useful parameters is the one which expresses the relationship between control-grid-voltage and anode current. Both the anode resistance and the amplification factor of a valve affect its performance as an amplifier. Mutual conductance takes both of these factors into consideration, which makes it useful in comparing the merits of valves. Mutual conductance is the ratio of the amplification factor to the anode resistance. It is commonly expressed in micromhos or milliamperes per volt, the latter term giving the change in anode current for one volt change of grid potential.

Mutual Conductance is defined as the ratio of a small change in anode current to the small change in grid voltage producing it.

The three parameters given above are defined as follows -



$$\mu = \frac{dE_a}{dE_g} \text{ when } I_a \text{ is constant,}$$

$$r_a = \frac{dE_a}{dI_a} \text{ when } E_g \text{ is constant, and}$$

$$g_m = \frac{dI_a}{dE_g} \text{ when } E_a \text{ is constant,}$$

where d = a small change,
E = voltage,
I = current,
a = anode, and
g = grid,

from which $g_m = \frac{\mu}{r_a}$

$\mu = g_m r_a$

and $r_a = \frac{\mu}{g_m}$

(Note. Fig. 8 shows how the above constants vary, μ being the only one reasonably constant.)

/ 4.4

FIG. 8. VARIATION OF CONSTANTS μ, g_m AND r_a.

4.4 Significance of the Valve Constants. The significance of the amplification factor and the anode resistance is more readily appreciated if an actual example involving alternating current voltage were solved. This is done in Fig. 9, using a valve having the anode characteristics shown. Fig. 9a shows the valve with steady voltages of -8 volts and 100 volts applied across grid and cathode, and anode and cathode respectively. Let an alternating current signal of 4 volts, E_g, be applied to the grid as shown.

This signal causes the grid voltage to vary between -4 and -12 volts. Consider the 100 volt anode curve. Without a signal (that is, the grid is at -8 volts potential), the anode current is 3mA. When the grid is at -4 volts, however, the anode current is 4.5 mA, and at -12 is 1.5 mA, thus the anode current is varying 1.5 mA above and below its normal value. This is equivalent to having 1.5 mA alternating current superimposed on the anode circuit.

Leave the grid-voltage at -8 volts and see what values of anode voltage give the same 1.5 mA variation of anode current.

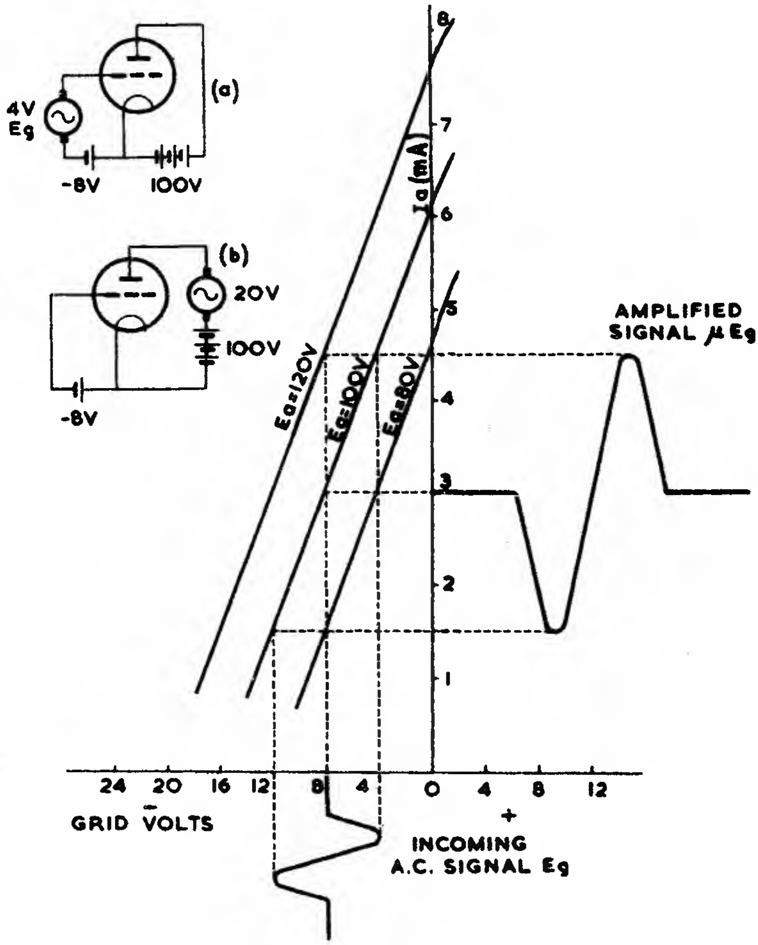


FIG. 9. SIGNIFICANCE OF THE AMPLIFICATION FACTOR.

It will be seen that the anode current is 4.5 mA when the anode voltage is 120 volts, and 1.5 mA when the anode voltage is 80 volts. Thus, the anode voltage must vary between 80 and 120 volts to produce the same effect on the anode current as a 4 volt signal on the grid. The anode voltage variation is equivalent to a 20 volt alternating current signal superimposed on the steady 100 volt anode potential (see Fig. 9b). Therefore, it may be said that a 4 volt signal on the grid produces a 20 volt signal in the anode circuit, or that the valve has an amplification factor of $\frac{20}{4} = 5$.

An alternating current 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 -

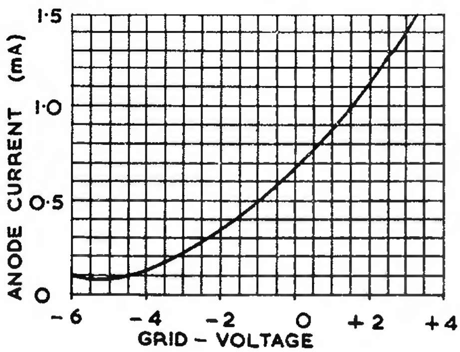
$$\frac{20 \text{ volts}}{1.5 \text{ mA}} = 13,333.33 \text{ ohms.}$$

The only impedance in the circuits shown is the impedance offered by the anode to cathode circuit inside a valve, and is the anode resistance, r_a, referred to previously. A valve may, therefore, be regarded as an alternating current generator developing a voltage of μE_g and having an internal impedance r_a, E_g being the alternating signal voltage applied to the grid.

4.5 Characteristic Curves. One of the more convenient methods of expressing valve characteristics is by "Characteristic Curves." These curves show the manner in which variation of potentials or current of one electrode affect the potentials or current on other electrodes. The most commonly used characteristic curves are -

Grid-voltage/anode current.
Grid-voltage/grid current.
Anode voltage/anode current.
Anode voltage/grid-voltage.

Grid-Voltage/Anode Current Characteristic Curve. The control which the grid exerts on the anode current is shown by the characteristic curve representing the relation between anode current and grid-voltage. This curve is shown in Fig. 10 for a type of tungsten cathode receiving valve. The anode voltage is 60 volts and the cathode current is 0.65 ampere. The current which heats the cathode also produces a drop in potential along it, so that, when one speaks of the grid or anode being so many volts positive with respect to the cathode, a particular point on the cathode must be specified. Usually, it is customary to take the terminal of the cathode, which is connected to the negative terminal of the cathode battery, as the point of zero potential, and the grid and anode voltages are then measured relative to this point. Thus, in Fig. 10, the grid and anode voltages are measured relative to the negative cathode terminal. The grid is at zero voltage when it is connected to the negative cathode terminal (that is, without C battery).



HOW GRID-VOLTAGE CONTROLS ANODE CURRENT.

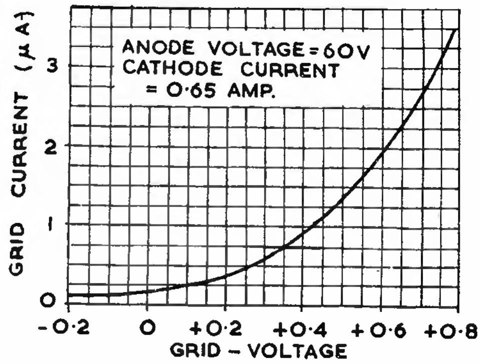
(Mutual Characteristic.)

FIG. 10.

The curve of Fig. 10, which shows how the grid-voltage controls the flow of current in the anode circuit, is of first importance in explaining how the thermionic valve operates, and will be referred to later. This curve is also known as the "Mutual Characteristic."

/ Grid

Grid-Voltage/Grid-Current Characteristic Curve. Another important curve shows the current which flows to the grid for different values of grid-voltage. Such a curve is given in Fig. 11 for the same valve as used for Fig. 10 at the same anode voltage and cathode current. As stated before, the current flow when the grid is negative is negligible, but the value of the current flow increases as the grid becomes positive, because some of the electrons on their way to the anode are attracted to the grid. The currents to the grid, as shown in Fig. 11, are small compared with the currents in the anode circuit. The grid currents are measured in micro-amperes (millionths of an ampere), while the anode currents are measured in milliamperes (thousandths of an ampere). The grid current is of importance in the action of the valve as a detector when a grid capacitor and grid leak are used, and also is of some importance in connection with amplifiers.



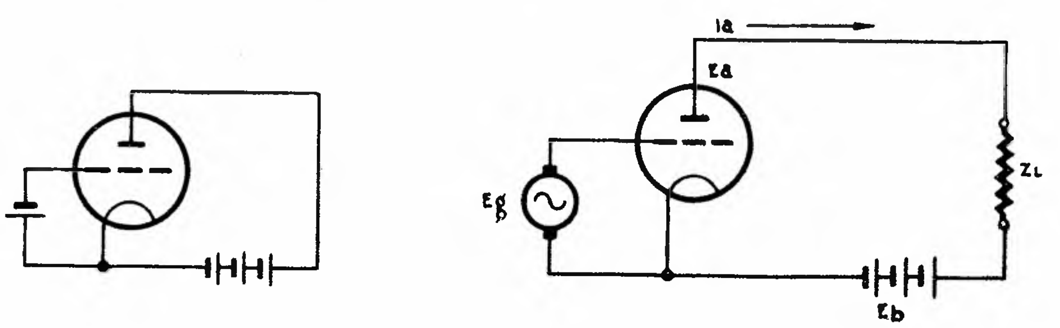
GRID-VOLTAGE/GRID-CURRENT CHARACTERISTIC CURVE.

FIG. 11.

Anode Voltage/Anode Current Characteristic Curve. If the anode voltage is varied and the corresponding anode current variations plotted against the anode volts, the resulting curve is termed the "Anode Characteristic." By repeating this test over a range of grid voltages, a set of curves is obtained which is commonly known as "the anode or plate family." These curves are of use in calculating power output, distortion, optimum load resistance, etc., as shown later.

Anode Voltage/Grid Voltage Characteristic Curve. A form of this curve, in which anode voltage is plotted against grid voltage each curve being for constant anode current, is known as the "Constant Current Characteristic" and is particularly useful in the case of radio frequency power amplifiers.

Static Characteristics. These curves are usually obtained by incremental variations of direct current potentials, and are termed "static" characteristics because the valve is not functioning in a normal manner, that is, with alternating current grid voltages. Also, there is no load in the anode circuit. Fig. 12a shows this condition. These curves serve, however, to indicate the suitability of valves for particular purposes.



(a) Static Conditions.

(b) Dynamic Conditions.

VALVE CHARACTERISTICS.

FIG. 12.

Dynamic Characteristics. These are measurements made with an alternating voltage on the control-grid and with various values of direct current potentials on the electrodes. They include the three constants previously mentioned, namely, μ, r_a and g_m, and, used in conjunction with load conditions, determine the operating characteristics of the valves. Fig. 12b shows this condition.

4.6 Application of Characteristic Curves.

Anode Family. Consider first a set of anode characteristic curves such as shown in Fig. 13. These are for a typical triode valve and are obtained as described above. From these curves, the amplification factor, anode resistance and mutual conductance may be readily calculated.

Amplification Factor. By definition -

$$\mu = \frac{dE_a}{dE_g} \text{ (with } I_a \text{ constant)}$$

where dE_a = a small change in anode volts, and

dE_g = a small change in grid volts.

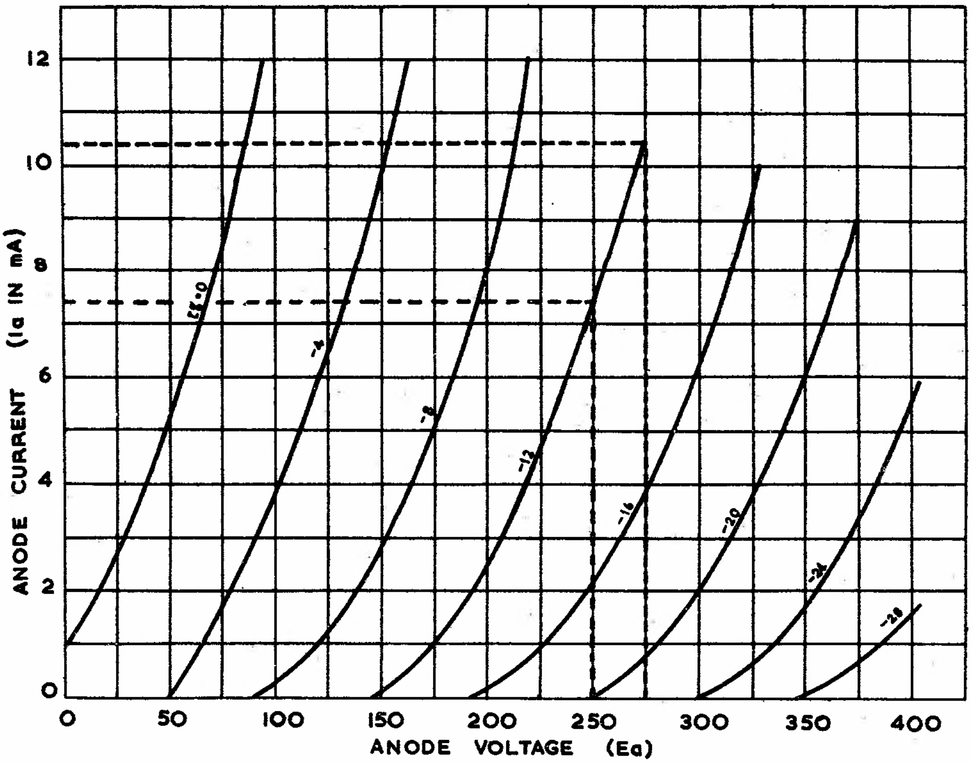


FIG. 13. ANODE CHARACTERISTICS OF TYPICAL TRIODE VALVE.

Referring to Fig. 13 -

Let E_a = 275 volts, and E_g = -12 volts, then I_a = 10.4 mA.

Change E_a to 250 volts, and it is then necessary to change the grid voltage to about -10.5 to restore the current to its original value of 10.4 mA. From these figures -

$$dE_a = 275 - 250 = 25$$

$$dE_g = 12 - 10.5 = 1.5$$

$$\text{and } \mu = \frac{25}{1.5} = 16.7.$$

Anode Resistance. By definition -

$$r_a = \frac{dE_a}{dI_a} \text{ (with } E_g \text{ constant)}$$

Let E_a be 275 volts, E_g be -12 volts, and I_a be 10.4 mA as before, and change E_a to 250 volts, then with E_g at -12, I_a has changed to 7.4 mA.

$$dE_a = 275 - 250 = 25 \text{ volts}$$

$$dI_a = 10.4 - 7.4 = 3 \text{ mA}$$

$$r_a = \frac{25}{0.003} = 8{,}333 \text{ ohms.}$$

Mutual Conductance.

$$g_m = \frac{\mu}{r_a}$$

$$= \frac{16.7}{8333} \text{ mho}$$

$$= \frac{16.7 \times 10^6}{8333} \text{ micromhos}$$

$$= \underline{2{,}000 \text{ micromhos.}}$$

4.7 Mutual Family. The anode characteristics are replotted to give the mutual characteristics shown in Fig. 14. The information they furnish is the same, but, as mentioned before, these curves are often of more use than the anode family since they relate input voltage to output current. It is of interest to calculate the three constants using a different set of values, and then compare the results.

Amplification Factor.

Let E_a = 250, and E_g = -8 then I_a = 17 mA.

also let E_a = 200, and E_g = -5,

then, if I_a is constant (17 mA) -

$$\mu = \frac{250 - 200}{8 - 5} = \frac{50}{3} = \underline{16.7}$$

/ Fig. 14.

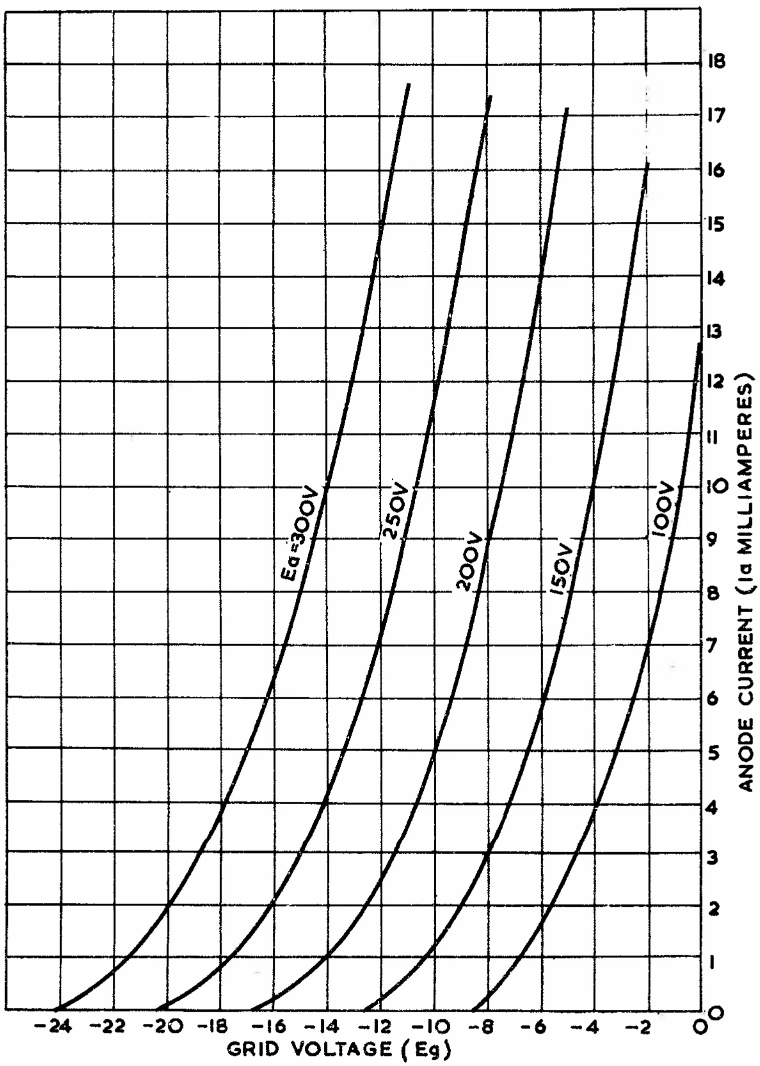


FIG. 14. MUTUAL CHARACTERISTICS FOR TYPICAL TRIODE VALVE.

Anode Resistance.

$$r_a = \frac{dE_a}{dI_a} \text{ (with } E_g \text{ constant)}$$

With $E_g = -14$, and $E_a = 300$, then $I_a = 10$ mA

and with $E_a = 250$, then $I_a = 4.1$ mA

$$\text{Thus } r_a = \frac{300 - 250}{0.010 - 0.0041} = \frac{50}{0.0059} = 8{,}500 \text{ ohms.}$$

/ Mutual

Mutual Conductance.

Let $E_a = 300$, and $E_g = -13$, then $I_a = 12.1$ mA.

Change E_g to -14, and $I_a = 10.0$ mA.

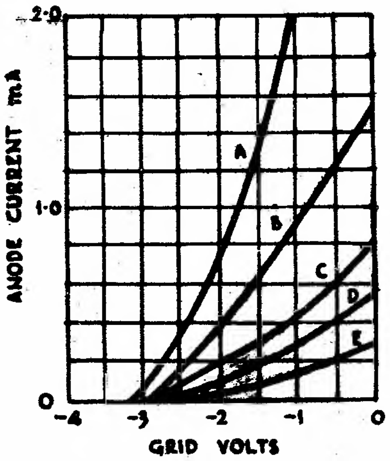
$$\text{Thus } g_m = \frac{12.1 - 10}{14 - 13}$$
$$= \frac{2.1 \text{ mA}}{1 \text{ volt}}$$
$$= 2{,}100 \text{ micromhos or } 2.1 \text{ mA per volt.}$$

4.8 The results of the two sets of methods are compared as follows -

Anode Curves.	Mutual Curves.
$\mu = 16.7$	$\mu = 16.7$
$r_a = 8{,}333$ ohms	$r_a = 8{,}500$ ohms
$g_m = 2{,}000$ μ mhos	$g_m = 2{,}100$ μ mhos

The results are in close agreement, since these three "constants" are not really constant.

4.9 Load Line. The characteristic curves referred to in the previous paragraphs are known as "Static" curves, but the addition of a load line or lines converts them to "Dynamic Curves," and they are then used to calculate the performance of the valve under operating conditions. The load line is added to either the anode or mutual characteristics, the anode family being more generally used since it involves less work. The load inserted in the anode circuit is in the form of a resistance, choke, transformer or tuned circuit, etc., and modifies the characteristic curve, as shown in Fig. 15, which shows the effect of different values of load impedance on the mutual characteristic of a triode. Although really belonging to amplifier discussion, the use of the load line is shown here, since it is associated with characteristic curves.



EFFECT OF LOAD RESISTANCE ON MUTUAL CURVE OF TYPE 2A6 VALVE.

Curve A. $Z_L = 0$
Curve B. $Z_L = 100{,}000$ ohms.
Curve C. $Z_L = 250{,}000$ ohms.
Curve D. $Z_L = 500{,}000$ ohms.
Curve E. $Z_L = 1$ megohm.

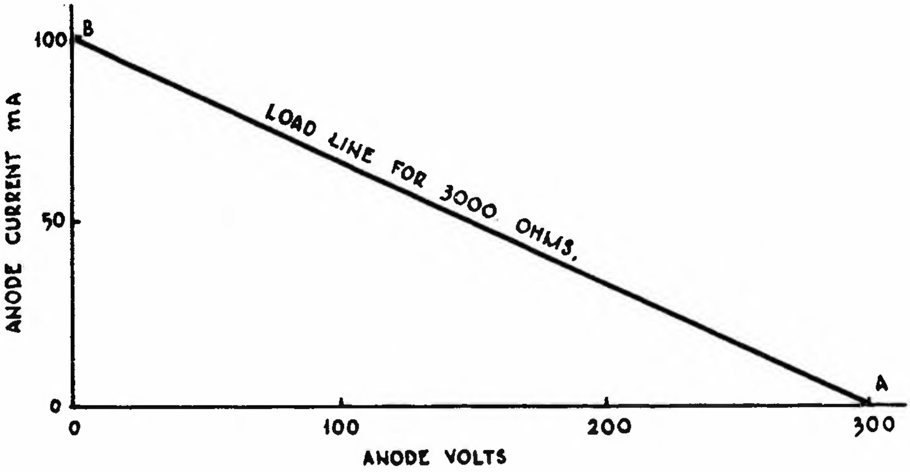
FIG. 15.

Construction of the Load Line. The load line is drawn independently of the curves, it being only necessary to know the anode supply voltage and the load resistance. Fig. 12b shows a valve with an external load Z_L. The limiting conditions of operation may be considered as being -

(i) Zero current through the load impedance.
(ii) Maximum current through the load impedance.

Condition (i) occurs when the grid voltage is such as to reduce the anode current to zero, the anode voltage then being a maximum. Condition (ii) occurs when the voltage drop across the load is equal to the supply voltage, that is, leaving the anode at virtually zero potential. / Let

Let the anode voltage be 300 volts and the load 3,000 ohms, as in Fig. 16. Point A is at the intersection of 300 volts and zero current, that is, at A. Point B occurs when 300 volts is dissipated across the load, that is, when the current equals $\frac{300}{3,000}$ = 100 mA. Join points A and B, and this is the load line for the assumed conditions.



CONSTRUCTION OF LOAD LINE (1st METHOD).

FIG. 16.

This method, however, does not make use of the additive effect in the anode circuit of a voltage applied to the grid. A voltage E_g applied to the grid produces approximately μE_g volts in the anode circuit, and this is considered as an additional anode potential superimposed upon the steady supply voltage. By consideration of this point, the output power from the valve is increased appreciably.

The other method of determining the load line is somewhat different, and is now described as applied to a triode Class A amplifier. It is necessary to know the anode supply voltage and the amplification factor of the valve, and, having these and the anode family curves, the following is determined -

Power output.
Second harmonic distortion.
Optimum load resistance.
Correct grid bias voltage.

To determine the load line by this method, let the two points determining the slope of the line be termed Y and Z, as shown in Fig. 17.

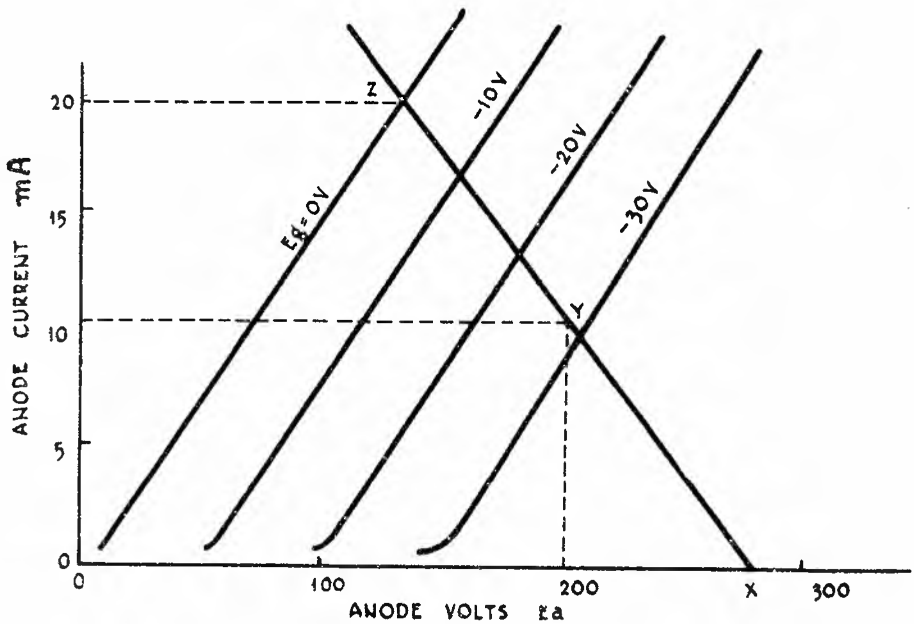
In Fig. 17, Y = zero-signal bias point = $-0.68 \frac{E_a}{\mu}$

where E_a = supply voltage,

and μ = the amplification factor.

Z is a point on the zero-voltage direct current bias curve ($E_g = 0$), and is determined by the maximum-signal anode current which is twice the value of the anode current at point Y.

/ Fig. 17.



CONSTRUCTION OF LOAD LINE (2nd METHOD).

FIG. 17.

Example.

Let E_a = 200 volts and μ = 5

then $Y = \frac{0.68 \times 200}{5} = -27.2$ volts

that is, point Y is at the intersection of the 200 volt ordinate and the -27.2 grid-voltage curve.

The anode current at this point is 10 mA, therefore, point Z is at the intersection of $2 \times 10 = 20$ mA and zero-bias curve.

Join YZ and extend it to X. XYZ is then the load line for the assumed operating conditions.

4.10 Examples on the Use of the Load Line. In calculating performance from the load line, it is always assumed that the peak alternating grid voltage is sufficient to -

(i) Swing the grid from zero-signal bias to zero bias on the positive half-cycles, and

(ii) Swing the grid to a value equal to twice the zero-signal bias voltage on the negative half-cycles.

/ The

The following formulae (where I is in amperes) are then applicable -

$$\text{Power Output } P = \frac{(E_{max} - E_{min})\ (I_{max} - I_{min})}{8} \text{ watts}$$

(See Section 11 of this Paper for derivation of this formula.)

$$\left.\begin{array}{r}\text{Percentage Second} \\ \text{Harmonic Distortion}\end{array}\right) = \frac{\frac{I_{max} + I_{min}}{2} - I_o}{I_{max} - I_{min}} \times 100$$

$$\left.\begin{array}{r}\text{Optimum Load} \\ \text{Resistance}\end{array}\right) = \frac{E_{max} - E_{min}}{I_{max} - I_{min}} \text{ ohms}$$

$$\text{Grid Bias Voltage} = \frac{0.68 \times E_a}{\mu}$$

where E_{max} = anode voltage at peak of negative swing,
E_{min} = anode voltage at peak of positive swing,
I_{max} = anode current at peak of positive swing,
I_{min} = anode current at peak of negative swing, and
I_o = anode current at zero-signal bias point.

4.11 The following examples of thermionic valve calculations indicate the method of using Characteristic Curves.

Example 1. Fig. 18 shows a set of Anode Characteristics for a triode having an amplification factor of 7.

$$\text{Let } E_a = 300 \text{ volts}$$
$$\text{then Point } Y = \frac{0.68 \times 300}{7} = -29 \text{ volts}$$

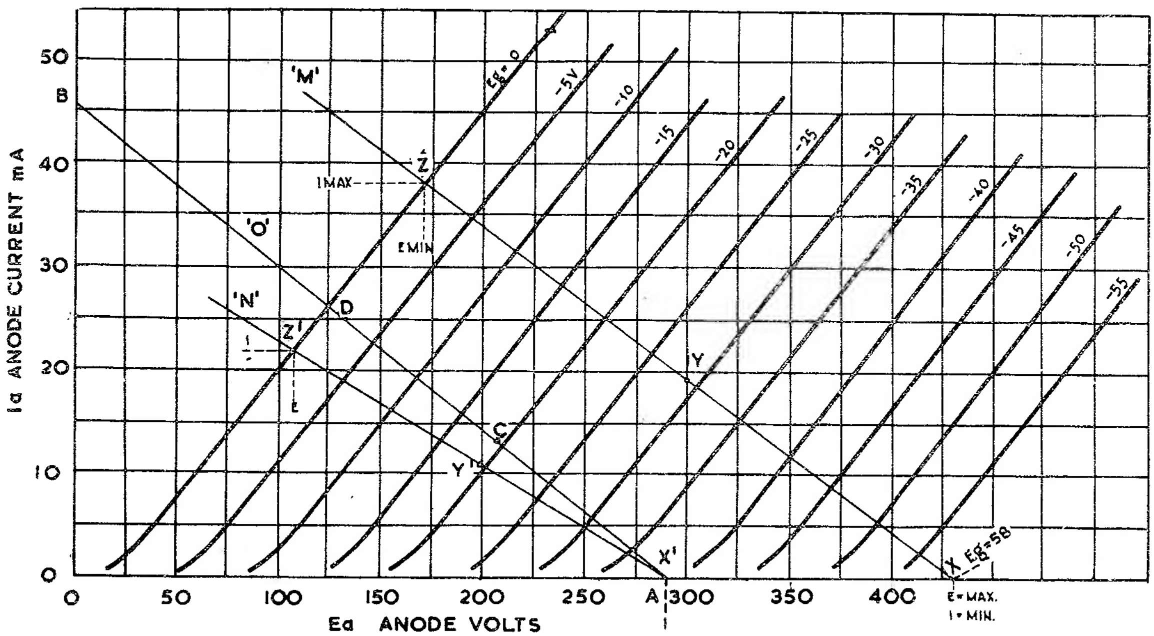


FIG. 18. ILLUSTRATING THE USE OF LOAD LINE. ANODE CHARACTERISTICS. TRIODE μ = 7.

Plot Y at the intersection of the 300 volt ordinate and -

$$E_g = -29 \text{ volts.}$$

The anode current at Y is 19 mA (I_o).

$2(I_o) = 38$ mA, thus point Z is at the intersection of $E_g = 0$ and $I_a = 38$ mA. Draw the load line ZYX. Line "M".

Let a 29 volt peak signal be applied to the valve at Y. The positive excursions of this signal drive the operating point to Z ($E_g = 0$) and the negative excursions to X.

Point Z is the location of minimum anode voltage and maximum anode current. Point X is the location of maximum anode voltage and minimum anode current.

At Z, E_a is 172 volts, and $I_a = 38$ mA.
At X, E_a is 460 volts, $I_a = 0$ mA, and $I_o = 19$ mA.

Substituting in formula -

$$\text{Power Output} = \frac{(460 - 172)(\frac{38 - 0}{1,000})}{8} = \frac{288 \times \frac{38}{1,000}}{8} = 1.37 \text{ watts.}$$

$$\left.\begin{array}{r}\text{2nd Harmonic}\\ \text{Distortion}\end{array}\right) = \frac{\frac{38 + 0}{2} - 19}{38} \times 100 = 0 \text{ per cent.}$$

$$\left.\begin{array}{r}\text{Optimum Load}\\ \text{Resistance}\end{array}\right) = \frac{460 - 172}{0.038 - 0} = \frac{288}{0.038} = 7,500 \text{ ohms (approximately).}$$

<u>Example 2.</u>

Let $E_a = 200$ volts (see Fig. 18)
then Y' = -19.4 volts
$I_o' = 11$ mA
and Z' = 22 mA

Draw load line ZYX. Line "N". Apply 19.4 volts peak signal and the following values are obtained -

$E_{max} = 290$ volts, $E_{min} = 108$ volts, $I_{max} = 22$ mA,
$I_{min} = 0$, $I_o = 11.0$ mA.

From these values -

Power Output = 500 mW,
Distortion = 0 per cent., and
Load Resistance = 8,270 ohms.

<u>Example 3.</u> Try an example using the first method (as shown in Fig. 16) of drawing the load line. Let $E_a = 300$ volts and $Z_L = 6,800$ ohms (similar values to Example 1).

Point A is at the intersection of 300 volts and zero current ordinates. Point B = $\frac{300 \times 1,000}{6,800} = 45.3$ mA. Join AB. Note that point A is at approximately -38 volts, thus the maximum peak signal that may be applied is 19 volts. With a signal of this value, the operating point is at C. (Line "O".) D = Point of E_{min} and I_{max}. /The

The following values are read from the graph -

$$E_{max} = 300 \text{ volts}, \quad E_{min} = 125 \text{ volts}, \quad I_{max} = 26 \text{ mA},$$
$$I_{min} = 0 \qquad I_o = 13.2 \text{ mA}.$$

From which -

Power Output = 572 mW, and
Distortion (2nd Harmonic) = 0.77 per cent.

Comparing Examples 1 and 3, it is seen that the output power is considerably reduced by using the first method of drawing the load line. In fact, the output with the first method (Example 3) is not much greater than that obtained by using the second method with an anode voltage of 200 volts (Example 2).

Example 4. Fig. 19 shows the anode characteristics of a type 45 triode. With the aid of this figure, calculate the power output, 2nd harmonic distortion, optimum load resistance and correct grid bias for anode supply voltages of -

(i) 280 volts,
(ii) 200 volts, and
(iii) Using the first method with a load resistance of 4,000 ohms and an anode supply of 280 volts.

The results should be as follows -

(i) 2.38 W.	5.7 per cent.	3,900 ohms.	-54 volts.
(ii) 760 mW.	2.73 per cent.	6,800 ohms.	-39 volts.
(iii) 1.1 W.	10.7 per cent.	(with 43 volt peak signal.)	

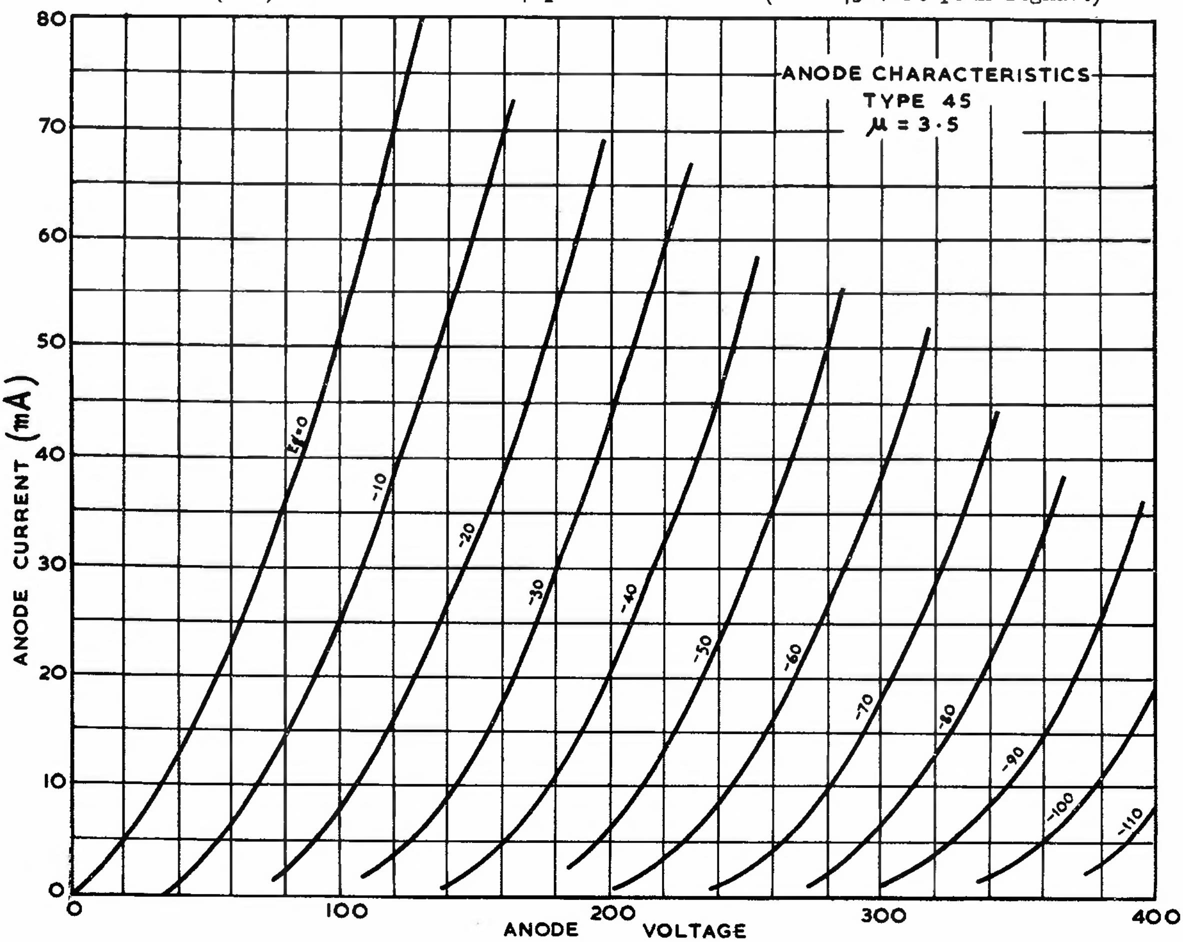
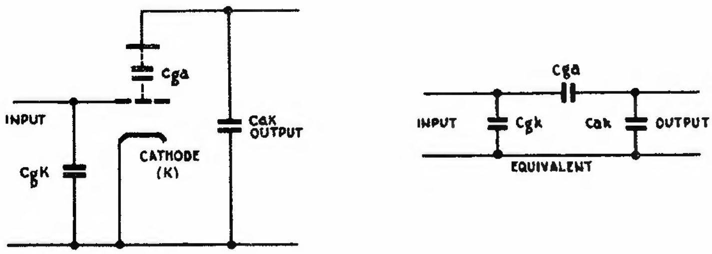


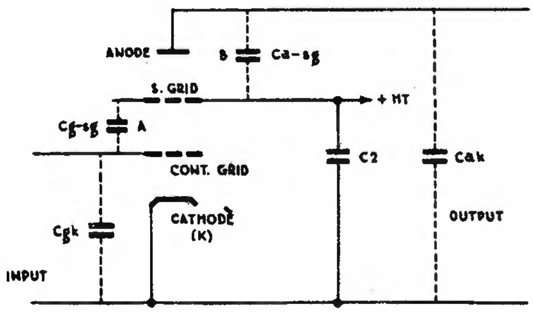
FIG. 19. ANODE CHARACTERISTICS TYPE 45 TRIODE.

5. SCREEN-GRID VALVES.

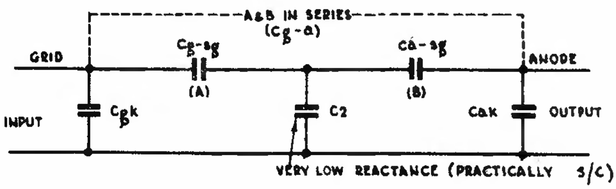
5.1 In Fig. 6c, reference was made to the inter-electrode capacitances associated with a triode, and these are shown in Fig. 20a. The most important capacitance is that between the anode and grid, since this, in effect, couples the output and input circuits together, giving rise to oscillation and stability and preventing full advantage being taken of the amplification properties of a triode. At audio frequencies, the coupling is not noticeable, as the reactance of C_{ga} is high, but, at radio frequencies, the reactance becomes low enough for a transfer of energy between output and input circuits via this capacity, producing the undesirable effects just mentioned.



(a) Triode Valve.



(b) Screen-Grid Valve.



(c) Equivalent Screen-Grid Circuit.

INTER-ELECTRODE CAPACITIES OF VALVES.

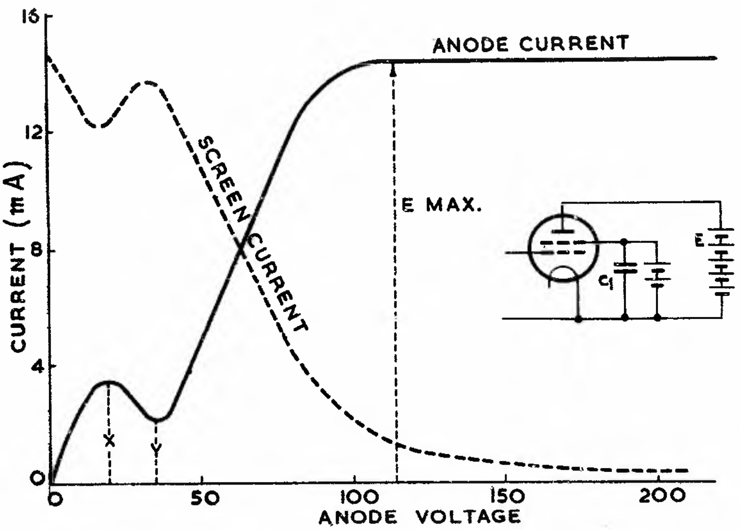
FIG. 20.

5.2 This trouble is minimised by incorporating some form of neutralising as is done in high power transmitting valves, but, for low power circuits, the use of a Screen-Grid Valve obviates the necessity for neutralising. In these valves, 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 effectively connected to the cathode via the condenser C2 in Fig. 20b. The effect of this grid is to split the grid-anode capacity into two series capacities with the input short-circuited from the output by the connection between the screen-grid and the cathode. This is shown in Fig. 20c. In normal operation, the screen-grid is given a positive potential equal to about two-thirds of the anode voltage, and, under these conditions, the amplification factor can be made much higher than for an equivalent triode. The following table will serve to compare a typical screen-grid valve and a typical triode valve.

	Triode Valve	Screen-Grid Valve.
Anode-Grid Capacity	2-8 $\mu\mu$ F	0.001-0.02 $\mu\mu$ F
Anode Resistance	20,000-50,000 ohms	0.2-1 megohm
Amplification Factor	15-50	100-1,500
Mutual Conductance	0.9-3.5 mA/V	0.5-4.0 mA/V

The result of inserting the screen-grid is effectively to screen the control grid from the anode, so that changes in anode potential have negligible effect on the electrostatic field around the control grid. Being closer to the space charge area than the anode, the screen-grid exerts an attraction for the electrons which, however, when they reach the vicinity of the screen-grid, come under the stronger influence of the anode and most of them pass through the screen-grid mesh and travel on to the anode. In other words, the anode, being shielded by the screen-grid from the other electrodes, has little effect in withdrawing electrons from the space charge area around the cathode, and the screen-grid virtually functions as the anode of a triode. Therefore, as long as the anode voltage is higher than the screen-grid voltage, anode current in a screen-grid valve depends mainly on the screen-grid voltage and very little on the anode voltage.

The curves of Fig. 21 show the anode and screen-grid characteristics of a screen-grid valve with a fixed screen-grid voltage, together with a typical circuit.



ANODE AND SCREEN-GRID CHARACTERISTICS OF SCREEN-GRID VALVE.

FIG. 21.

A suitable capacitor in the screen to cathode connection prevents the screen supply from being short-circuited, whilst still providing an effective alternating current short circuit. 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 - since the high screen voltage attracts electrons from the space charge and the increasing anode voltage causes an increasing number of 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 emitted by the cathode and passing to the anode (primary emission) and to give them such a high velocity that, on striking the anode, they dislodge other electrons by their impact. These are termed secondary electrons and this effect is known as "secondary emission." At low anode voltages, the effect is slight, since the velocity of the electrons is comparatively low. Increasing the anode voltage increases the velocity of the primary electrons and the amount of secondary emission. Some of the secondary electrons are attracted to the screen which, as yet, has a higher potential than the anode, the result being an increase in screen-grid current and a decrease in anode current. On the anode characteristics of Fig. 21, the anode current rises rapidly until the anode voltage reaches a value X. Beyond this value, 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 E_{max}, 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 E_{max}, and an increase of anode voltage beyond E_{max} merely draws a few electrons away from the screen.

/ The

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. This point on the characteristic curve is commonly known as the "dip." When the valve is operated near this point, this dip produces serious anode distortion and resultant harmonic generation.

5.3 Constants of Screen-Grid Valves. The three constants of screen-grid valves differ from those of an equivalent triode as follows -

Anode Resistance. Above anode voltage E, 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, a valve 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 E.

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.

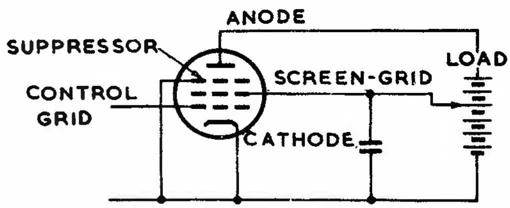
Amplification Factor. Since r_a is higher for the screen-grid valve than for its equivalent triode, and since g_m for the two is the same, the μ of a screen-grid valve is much higher from the relationship $\mu = r_a g_m$. This increase in amplification factor only applies provided the anode voltage does not fall below E, and is thus secured at the expense of providing a higher anode voltage for the screen-grid valve than for its equivalent triode.

6. PENTODE VALVES.

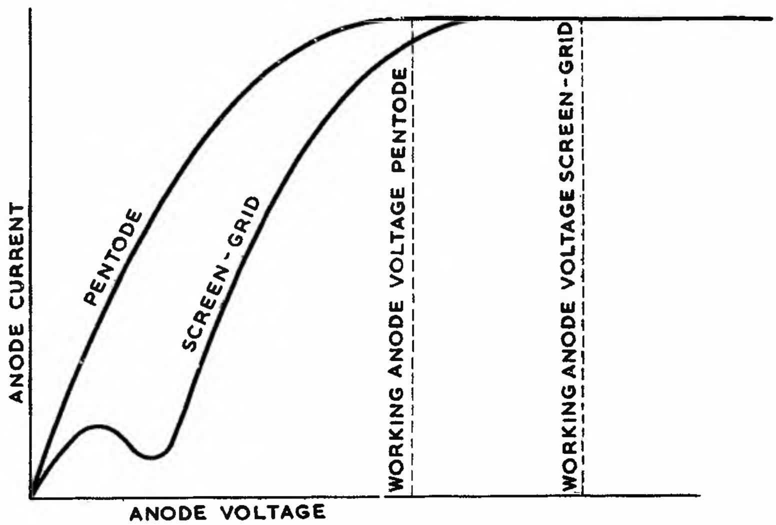
6.1 Pentode valves (see Fig. 22a) contain a cathode, an anode and three grids, and were developed to remove the "dip" from the anode current curve of the screen-grid valve which was referred to in the previous paragraphs.

The additional grid is introduced between the screen-grid and anode and is termed the "Suppressor Grid." By keeping this grid at cathode potential, it is negative with respect to anode and thus intercepts secondary electrons emitted by the anode and repels them from the screen causing them to return to the anode. A pentode is shown schematically in Fig. 22a, and the anode characteristics of a pentode and a screen-grid are shown in Fig. 22b to bring out the effect of the suppressor grid.

The pentode has practically replaced the tetrode in modern radio receivers. This valve is regarded as a constant current generator over the usual operating range, since the anode current is essentially independent of anode voltage.



(a) Circuit.



(b) Anode Characteristics of Pentode and Screen-Grid Valves.

FIG. 22. PENTODE VALVE.

6.2 Variable Mu Pentode. A glance at the mutual characteristics of an ordinary pentode shows that the anode current cuts off sharply at the lower end of the curve (Curve A, Fig. 23).

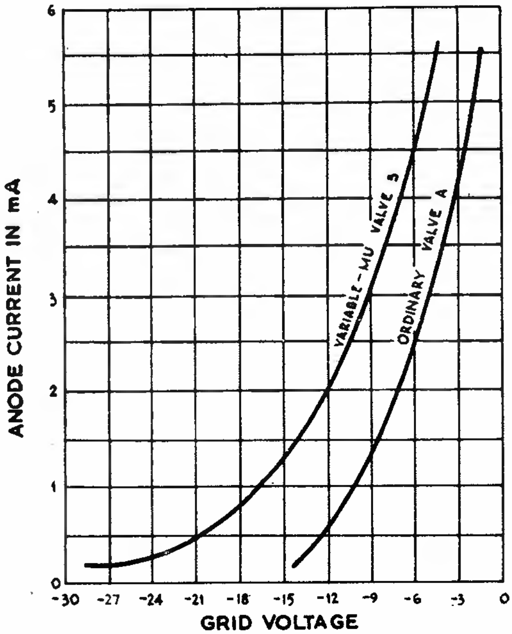


FIG. 23. MUTUAL CHARACTERISTICS PENTODE VALVE.

This cut-off prevents the satisfactory use of Automatic Gain Control and also introduces distortion with large grid swings. The requirement of Automatic Gain Control is that the valve bias (E_g) may be varied over a wide range without introducing distortion, the variation of bias producing a corresponding variation of μ and thus effectively controlling the amplification of signals. Curve B in Fig. 23 shows the variable mu characteristic. The difference on the lower position of the curve is evident. The variable mu characteristic is obtained by variation of the design of the control grid structure, such as tapering the over-all diameter, variation of spacing between the turns, etc. This type of valve also assists in reducing the effect of cross-modulation in a radio-receiver. The electrodes of a pentode valve may be connected in several ways to give special characteristics, such as those shown in Fig. 24.

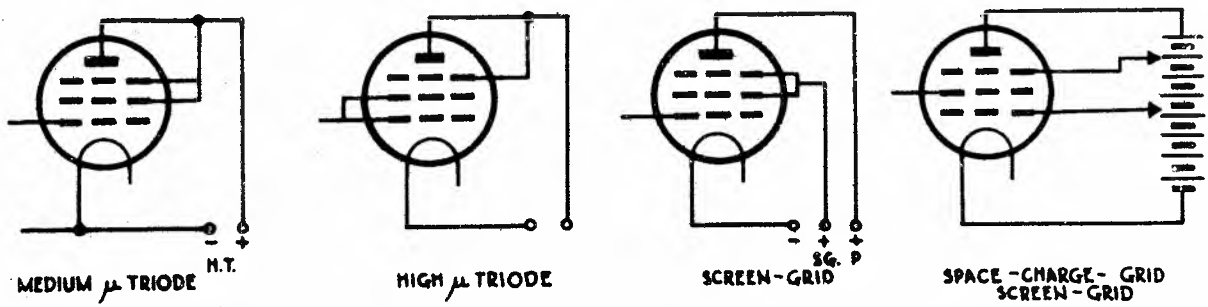


FIG. 24. PENTODE VALVE ARRANGED IN VARIOUS WAYS.

Pentode-type screen-grid valves are used as radio-frequency voltage amplifiers and as audio-frequency voltage amplifiers to give high voltage gain per stage, since the pentode resembles the tetrode in having a high amplification factor. Pentode valves also are suitable as audio-frequency power amplifiers, having greater anode efficiency than triodes and requiring less grid swing for maximum output. The latter quality can be indicated in another way by saying that the "power sensitivity" (ratio of power output to grid swing causing it) is higher. In audio power pentodes, the function of the screen-grid is chiefly that of accelerating the electron flow rather than shielding, so that the grid is often called the accelerator grid. In radio-frequency voltage amplifiers, the suppressor grid, in eliminating the secondary emission, makes it possible to operate the valve with the anode voltage as low as the screen voltage, which cannot be done with tetrodes.

As audio-frequency power amplifiers, pentodes have inherently greater distortion (principally odd-harmonic distortion) than triodes. The output rating usually is based on a total distortion of 10 per cent.

7. BEAM POWER VALVES.

7.1 The conventional power pentode, when operated as a Class A push-pull amplifier, has a rather large third-harmonic content in the output. The large third-harmonic component in the output is caused by a non-uniformity in the electron density between the screen and the anode. This condition, in turn, is due to an insufficient density of electrons in the stream to the anode. To remedy these conditions, the so-called "Beam-Power Valve" was designed.

7.2 In the beam-power valve, a cross-section of which is shown in Fig. 25, non-uniform density of electron stream and insufficiency of electrons are corrected by beam forming plates, which concentrate the electron stream into a well-defined beam as shown in Fig. 26. In the tetrode type of beam-power valve, the beam forming plates function in place of an actual suppressor to repel secondary emission. This they do by virtue of their physical shape, which causes their bent edges to coincide with that portion of the beam wherein it is desired that the space charge shall be intensive enough to prevent stray secondary electrons from returning to the anode. Hence, a suppressor grid is not needed. In the pentode type of beam power valve, an actual suppressor grid is used to repel secondary emission.

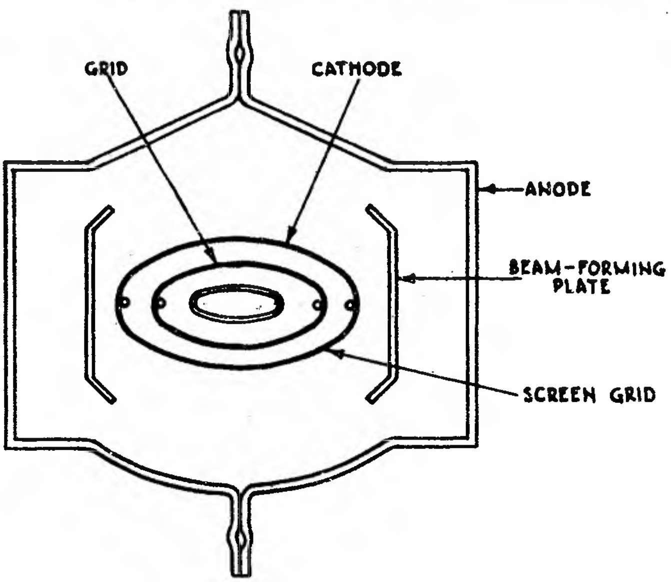
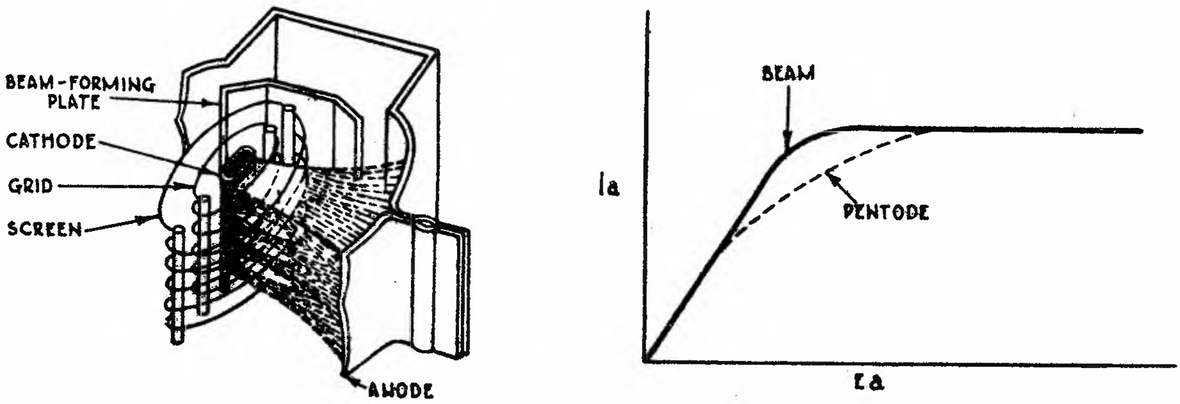


FIG. 25. CROSS-SECTION OF BEAM-POWER VALVE.

7.3 Over-all efficiency of this valve is increased by a careful placement of the electrodes within the valve, so that the beam-forming plates and the grids will operate on certain critical portions of the electron stream. Typical examples of the tetrode-type beam-power valve are the 6L6 and the 6L6G. Beam-power valves using an actual suppressor are the 6V6 and the 6V6G. The tetrode type of beam-power valve has found some application also as a crystal oscillator in transmitter circuits. A feature of the 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 the screen and the grid causes the electrons to travel in sheets between the turns of the screen, so that very few of them flow to the screen. Because of the effective suppressor action provided by 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.

/ Fig.

Fig. 26 shows the structure of a beam-power valve employing space-charge suppression and how the electrons are confined to beams. The high-density space-charge region is indicated by the heavily dashed lines in the beam.



(a) Construction.

(b) Difference in Characteristics of Beam and Pentode Valves.

FIG. 26. BEAM-POWER VALVE.

Note that the edges of the beam forming plates coincide with the dashed portion of the beam, and thus extend the space-charge potential region beyond the beam boundaries to prevent stray electrons from returning to the screen outside the beam.

8. VALVE INTER-ELECTRODE CAPACITY.

8.1 Inter-electrode capacities were referred to when discussing the screen-grid valve, and this section gives more information on this valve characteristic.

8.2 Effects of Inter-electrode Capacity. The elements of a valve form an electrostatic system, each element acting as one plate of a small capacitor. As stated previously, these capacitors are -

Grid-to-cathode capacity,
Grid-to-anode capacity, and
Capacity of the grid-to-anode and grid-to-cathode connected together.

The total capacity of a valve includes the capacity of the electrodes of the valve and the lead-in wires, and the capacity of the base. The capacity between the grid and the cathode and between the anode and the cathode may be about 5 $\mu\mu$F. The capacity between the grid and anode is larger, being, for example, approximately 8 $\mu\mu$F in the Type 210 valve.

The inter-electrode capacities of a valve, as measured when the elements are free, are not the same as when the elements are connected. Thus, the "direct" capacity between the grid and anode is increased by the mutual capacity from grid to cathode and from cathode to anode. The direct capacity between the grid and anode of a Type 01A valve, when the cathode has been removed, averages 6 $\mu\mu$F, while the capacity as measured between these two elements in a complete valve is 8.1 $\mu\mu$F. The effective value of this capacity is further increased by the capacity of the wiring of the valve socket, the valve base, and also by the amplification action of the valve.

8.3 Input and Output Circuits with Electrode Capacities. When a valve is in use its input circuit is considered to be from the grid to the cathode, and its output circuit from the anode to the cathode through a battery and some external load. Thus, the capacity of the input circuit may be considered as that of a capacitor, which has the grid for one plate and the anode and cathode connected together for the other plate. If an alternating voltage is applied to the grid-to-cathode circuit of a valve, an alternating current will flow in the grid circuit because of the grid-to-cathode capacity. Whether the cathode is heated or not, this grid voltage will cause a current in the anode circuit due to electrostatic induction through the capacity from the grid to the anode.

/ While

While the grid-to-cathode capacity and the anode-to-cathode capacity do not affect the performance of a valve at audio frequencies and have little effect at radio frequencies, the grid-to-anode capacity has a very marked effect on a radio frequency amplifier. As far as the valve itself is concerned, the capacities between the elements of a valve introduce a reactance effect.

When amplification is given in terms of the applied grid voltage, the cathode-to-anode capacity has only a small effect, so that the amplification is not affected by frequency for values up to several thousand kc/s. Usually, however, the amplification is given as the ratio of the output power to the input power, and the effect of the reactance due to electrode capacities depends on the kind of circuit that is used. If the reactance of the output circuit has the effect of capacity, or if the output circuit consists of a resistance, the input resistance is positive. Under such conditions, power is taken by the valve from the input circuit. The value of this power which is used is so small at ordinary frequencies that it may be neglected. At high frequencies, power is not taken by the grid circuit, but the electrode capacities offer a path to the input current and thus reduce the amplification.

The increase in effective inter-electrode capacity may become so large under certain load conditions as to affect the performance of the valve at high audio frequencies. Thus, in a resistance-coupled amplifier, the effective capacity reaches a value of 250 to 300 $\mu\mu$F, which is high enough to cause a decrease in amplification at frequencies over 5,000 c/s.

8.4 In general, the effect of inter-electrode capacity is to produce a coupling between the input and output circuits. Consequently, the valve does not have a true unilateral or single-direction characteristic curve. The extent of the coupling depends on the circuit constants. This kind of coupling may cause a feedback of energy to the input circuit or, with certain circuit adjustments, an absorption of energy from the input circuit. The effect of inter-electrode capacity is to reduce amplification at high frequencies.

9. RELATION BETWEEN VALVE CONSTANTS AND STRUCTURE.

9.1 The two main factors in the design of thermionic valves are the Amplification Factor and the Anode Resistance. The amplification factor (μ) depends upon the mechanical construction of the valve. In practice, it is found that μ is not quite constant but decreases slightly at low voltages, although the variation is not appreciable within the operating range. Fig. 8 (which appears early in this Paper) shows how the constants vary under different operating conditions.

The anode resistance (r_a) is inversely proportional to the surface areas of the anode and the cathode. It depends also upon the operating voltages. The value of r_a is further affected by μ, which, as stated previously, depends almost entirely on the structure of the grid and its position with relation to the anode.

An amplifying valve gives best operation when its anode resistance is equal to the impedance with which the valve operates. In cases where this is not possible, the total anode resistance may be reduced by operating the valves in parallel, or, by the use of an "output transformer", the anode resistance of a valve may be matched to the impedance of the device with which it operates.

The mutual conductance g_m, being equal to $\frac{\mu}{r_a}$, depends on the factors which determine these terms. In some types of valves, it is necessary to make this ratio $(\frac{\mu}{r_a})$ as large as possible. Then, for a given value of μ, r_a must be as small as possible. To make μ large and r_a small, therefore, the grid must be close to the cathode. When a valve is to be used as a detector, it should have a low internal resistance which changes suddenly within narrow limits when the grid voltage is varied. Since the amplification factor depends on the ratio of the change in the anode voltage to the change in the grid voltage to produce the same change in anode current, the maximum action is obtained, when, for a given change of the grid voltage, the necessary change of the anode voltage to provide the same current is a maximum. Thus, in a detector valve, the resistance must drop suddenly from a maximum to a minimum for a small change in grid voltage. The nearer the grid is to the cathode and the farther the grid is from the anode, the better are the detecting and voltage amplifying qualities of a valve.

/ 10.

10. MISCELLANEOUS TYPES OF VALVES.

10.1 Valves with special application are described in the appropriate sections, and a reference only can be made to some of the large number of compound or multi-purpose valves developed for special conditions.

10.2 Multi-Purpose Types. A great many types of valves have been developed to do special work in receiving circuits. Among the simplest of these are full-wave rectifiers, combining two separate diodes of the power type in one bulb, and twin-triodes, consisting of two triodes in one bulb, for Class B audio amplification. To add the function of diode detection and automatic volume control (described later) to that of amplification, a number of types are made in which two small diode anodes are placed near the cathode but not in the amplifier-portion structure. These types are known as duplex- diode triodes or duplex-diode pentodes, depending upon the type of amplifier section incorporated.

Another type is the Pentagrid Converter, a special valve working as both oscillator and first detector in super-heterodyne receivers. There are five grids between cathode and anode in the pentagrid converter. The two inner grids serve as the control grid and anode of a small oscillator triode, while the fourth grid is the detector control grid. The third and fifth grids are oonnected together to form a screen-grid, which shields the detector control grid from all other valve elements. The pentagrid converter eliminates the need for special coupling between the oscillator and detector circuits.

Another type of valve consists of a triode and pentode in one bulb, for use in cases where the oscillator and first detector are separately coupled. Still another type is a pentode with a separate grid for connection to an external oscillator circuit. This "injection" grid provides a means for introducing the oscillator voltage into the detector circuit by electronic means.

10.3 Power Valves - Special Considerations. Valves intended for use as power amplifiers must be capable of dissipating anode and grid losses in proportion to the desired power. Likewise, the electron emission from the cathode and the value of voltage that can be applied to the anode with safety must be in proportion to the power output that the valve is to develop. Small power valves are often merely large receiving valves constructed according to the standard receiving-valve technique, involving oxide-coated cathodes, "getters" to obtain and maintain vacuum, etc. Such valves are not robust, however, and the maximum power that can be developed with this type of construction is limited.

Large Valves. Larger valves up to dissipations of several hundred watts, are normally still enclosed in a glass envelope, but are constructed along lines entirely different from those of receiving valves. The anodes are commonly molybdenum, although tantalum carbon is sometimes used. Grids are usually molybdenum with tungsten and tantalum as alternatives. Cathodes are sometimes oxide-coated, but more frequently are thoriated tungsten carbonised for long life. The exhaust procedure ensures that a high degree of vacuum can be maintained without a getter. The various electrodes inside the valve are heated to high temperatures during exhaust, and the glass envelope is simultaneously baked at a temperature just below the softening point, and this process is continued until all the occluded gases have been removed. It is possible in this way to obtain a degree of vacuum such that a satisfactory life can be obtained with thoriated tungsten cathodes even at high anode voltages, particularly if tantalum is used for the anode.

Air and Water Cooling. In valves where the anode dissipation is of the order of kilowatts, the anodes of the valves are in the form of copper cylinders, which are part of the envelope and are cooled by means of circulating water, or by being soldered to a radiating structure with many fins which are cooled by means of an air-blast. Analysis and experience indicate that water and forced-air cooling are about equally effective when used with thermionic valves having external anodes.

/ Decision

Decision as to which type of cooling is preferable depends upon the individual factors involved in the particular case. Thus, air-cooled valves behave better in operation than water-cooled valves with respect to "flash-back". With forced-air cooling, the anodes tend to run hotter, but the glass ends run cooler than with water cooling, which has some advantages and some disadvantages. With forced-air cooling, the anode capacity to earth is increased by the large cooling structure - something to be avoided at very high frequencies. The most effective way to employ water-cooling is to arrange the water jacket so that the water passes over the anodes in a very thin stream of high velocity. In this way, any steam bubbles that may be formed on the anode tend to be scraped away by the velocity of the cooling water. The tendency for such bubbles to form limits the amount of heat that can be dissipated by a water-cooled anode, since, if the tendency of the bubble to stick to the copper is not over-come by the flow of cooling water, then that particular spot on the anode ceases to be cooled by direct contact with the water, and may readily become hot enough to cause failure of the valve.

In valves with external anodes that are water cooled, there is a tendency for the grid to produce a focusing effect on the electrons, which causes high spots to be produced on the anode at the point where the electrons concentrate. This is much more troublesome with water cooling than with forced-air cooling. With water cooling, these high spots tend to cause steam bubbles to form on the surface of the anode.

The cathodes of valves employing water or forced-air cooling are always tungsten. This is because such high power valves are operated at very high anode voltages, and it is not practicable to exhaust copper-anode valves as thoroughly as glass-envelope valves, where the anodes can be brought to incandescent heat during the exhaust process. Thoriated-tungsten cathodes accordingly do not behave satisfactorily, and pure tungsten must be employed. The thoriated-tungsten and oxide-coated cathodes should be operated at the rated voltage to obtain maximum life. With tungsten cathodes, however, decreasing the cathode voltage will increase the useful life at the expense of reduced electron emission (and hence lowered peak anode current). Varying the voltage applied to a tungsten cathode by 5 per cent. from normal will double or halve the life as the case may be, while a 10 per cent. variation affects the life by a factor of four. A study of the optimum diameter of a tungsten cathode, if the cost of power and valve replacement are taken into account, indicates that the optimum diameter is one that would correspond to a life of approximately 10,000 hours.

Arc-Back. In water-cooled or forced-air cooled valves operating at high anode voltages (such as 15 to 20 kilovolts), trouble is frequently encountered from arc-back or "flash-arcs" in the valve. These are ordinary high current arcs that suddenly form between cathode and anode, without apparent reason, and short-circuit the anode supply. After such an arc has been broken by opening of the circuit breakers and the anode voltage has been reapplied, the valve may continue to operate for many hours or days without any further trouble, while in other cases another flash-arc may occur almost immediately. A valve that has a tendency to arc back will generally improve in this respect if it is put through a conditioning process. This conditioning process involves applying an alternating potential of about 25,000 volts between anode and cathode of the valve with the cathode unheated, with suitable relays and current limiting devices in the circuit to prevent damage from arcs. This results in the production of many flash-arcs in the valve, which gradually clean up the gas that is causing the trouble and bring the valve to a satisfactory operating condition.

Operation. Power valves are commonly operated so that the grid becomes positive during part of the cycle. This results in grid current and causes power dissipation at the grid of the valve, which is sometimes the limiting factor in valve operation. Consequently, the grids of power valves often operate at relatively high temperatures, and such materials as molybdenum, tungsten or tantalum are accordingly generally used. The fraction of the primary electrons intercepted by the grid depends upon the grid potential relative to the anode potential and upon the grid structure. In ordinary triodes with equal grid and anode voltages, the effective grid area, insofar as intercepting the flow of primary electrons is concerned, is between 120 and 180 per cent. of the actual grid area.

The grid heating that takes place is determined by the number of primary electrons intercepted by the grid and by the grid voltage. The actual D.C. grid current, as measured by a meter, may differ from the number of primary electrons received by the grid as a result

/ of

of secondary emission, causing the grid to lose secondary electrons at the same time that it receives primary electrons. The amount of current thus lost through secondary emission will be affected by the electrode potentials, by the grid temperature and by the character of the grid surface. In the case of thoriated-tungsten and oxide-coated cathodes, the secondary emissions may, under some conditions, become quite large as a result of cathode material that has been deposited upon the grid.

10.4 Receiving Valves. Receiving valves may be divided into two main classes -

(i) The glass envelope valves, and
(ii) The so-called "metal" valves.

The glass envelope valves, as the name implies, have a glass envelope, and the electrodes are mounted on leads running through either a stem or a glass cup to which the bulb is sealed after the electrodes are mounted and adjusted. The metal-envelope valve employs a metal shell in place of the glass bulb, with all the leads (except perhaps a control-grid lead) brought through a glass button or glass beads in the base of the valve.

The cathodes of receiving valves (either filaments or heaters) are generally of the oxide-coated type. The grid and anode electrodes and side rods are usually constructed from nickel, and a "getter" is used to obtain the desired degree of vacuum. The exhaust, and also many of the other assembly operations, are performed on automatic machines.

10.5 "Getters." A high vacuum is obtained in receiving valves by means of a "getter", which is volatilised inside the valve for the purpose of removing residual gas by either chemical or mechanical action. Magnesium is widely used as a getter, but other materials, such as barium berylliate, zirconium, phosphorus, etc., can also be employed as well as various mixtures. A getter should be initially inert, but should be of such character that it can be made highly active by some process, such as a chemical change or vaporisation of material at room temperature. Some getters also act as "keepers", in that they not only remove what gas is present in the valve at the time of flashing, but also combine with any gas that may subsequently be liberated within the valve during normal operation. Getters cannot be used in valves where the anode dissipation is large, because the large amount of heat liberated in such valves would vaporise the getter and destroy the vacuum.

Certain metals have the property of absorbing gases when raised to a high temperature. For example, tantalum when very hot, will absorb gases, which makes it a desirable metal to use in power valves for anodes and grids. Zirconium is also a useful metal in this regard and, at 1,400 degrees, will absorb copious quantities of such gases as oxygen, nitrogen, carbon monoxide and carbon dioxide.

A small supply of the "getter" is assembled on a small metal plate attached to one of the element supports. After the valve is sealed, the "getter" is volatilised by heating the valve with a high frequency coil, and usually condenses in a silvery film on the inside of the container. The life of a valve depends to a considerable extent on the efficiency of the action of the getter.

11. POWER OUTPUT FORMULA.

11.1 The Power Output formula given in paragraph 4.10 is derived as follows -

Fig. 27 shows the fluctuations of anode voltage E_a about a mean steady state value E_{mean}. It is seen that -

The peak amplitude $A = \frac{X}{2} = \frac{1}{2} E_{max} - E_{min}$.

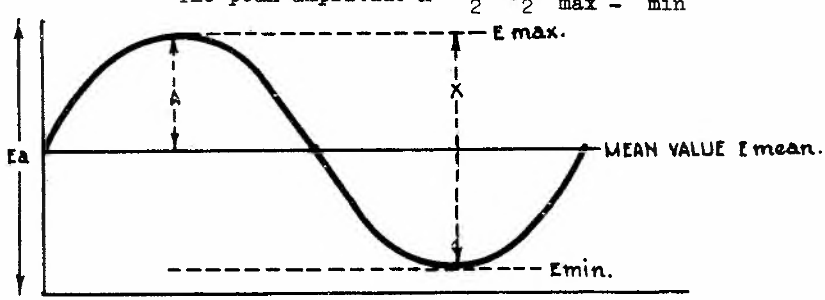


FIG. 27.

Similarly, Fig. 28 shows the current variations, and -

$$\text{The peak amplitude } B = \frac{Y}{2} = \frac{I_{max} - I_{min.}}{2}$$

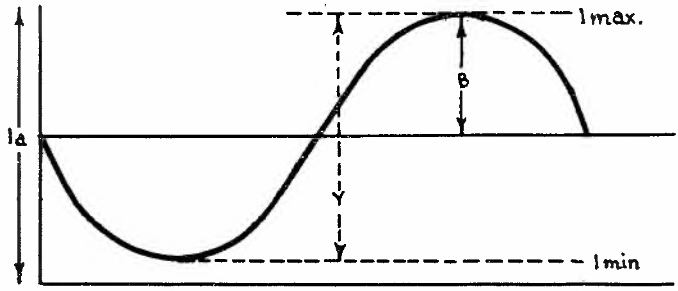


FIG. 28.

Since these are peak amplitudes, they must be multiplied by $\frac{1}{\sqrt{2}}$ (0.707) to obtain the effective values. Thus -

$$E_{eff} = \frac{1}{\sqrt{2}} \left(\frac{E_{max} - E_{min}}{2}\right)$$

$$\text{and } I_{eff} = \frac{1}{\sqrt{2}} \left(\frac{I_{max} - I_{min}}{2}\right)$$

The power output is the product of these effective values and is equal to -

$$P = \frac{1}{\sqrt{2}} \left(\frac{E_{max} - E_{min}}{2}\right) \times \frac{1}{\sqrt{2}} \left(\frac{I_{max} - I_{min}}{2}\right)$$

$$= \left(\frac{1}{\sqrt{2}}\right)^2 \left(\frac{E_{max} - E_{min}}{2}\right) \left(\frac{I_{max} - I_{min}}{2}\right)$$

$$= \frac{(E_{max} - E_{min})(I_{max} - I_{min})}{8}$$

12. TEST QUESTIONS.

1. (i) Explain the terms "Thermionic Emission," "Work Function," and "Indirectly-heated Cathode."
   (ii) Name two materials commonly used for cathodes.

2. Describe the construction and, briefly, the operation of a simple triode.

3. What is a load line and what is its function?

4. Define the following -

   (i) Amplification Factor,
   (ii) Mutual Conductance, and
   (iii) Anode Resistance.

5. Explain the effect of the "space charge" on the operation of a triode valve.

6. (i) What is the normal function of the "screen-grid" in a tetrode?
   (ii) A certain disadvantage of a tetrode led to the introduction of a "suppressor-grid." What was this disadvantage?

7. What advantages does a beam valve possess as compared with an ordinary pentode?

END OF PAPER.

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COURSE OF TECHNICAL INSTRUCTION.

THERMIONIC VALVE AMPLIFIERS.

CONTENTS.

1. INTRODUCTION.

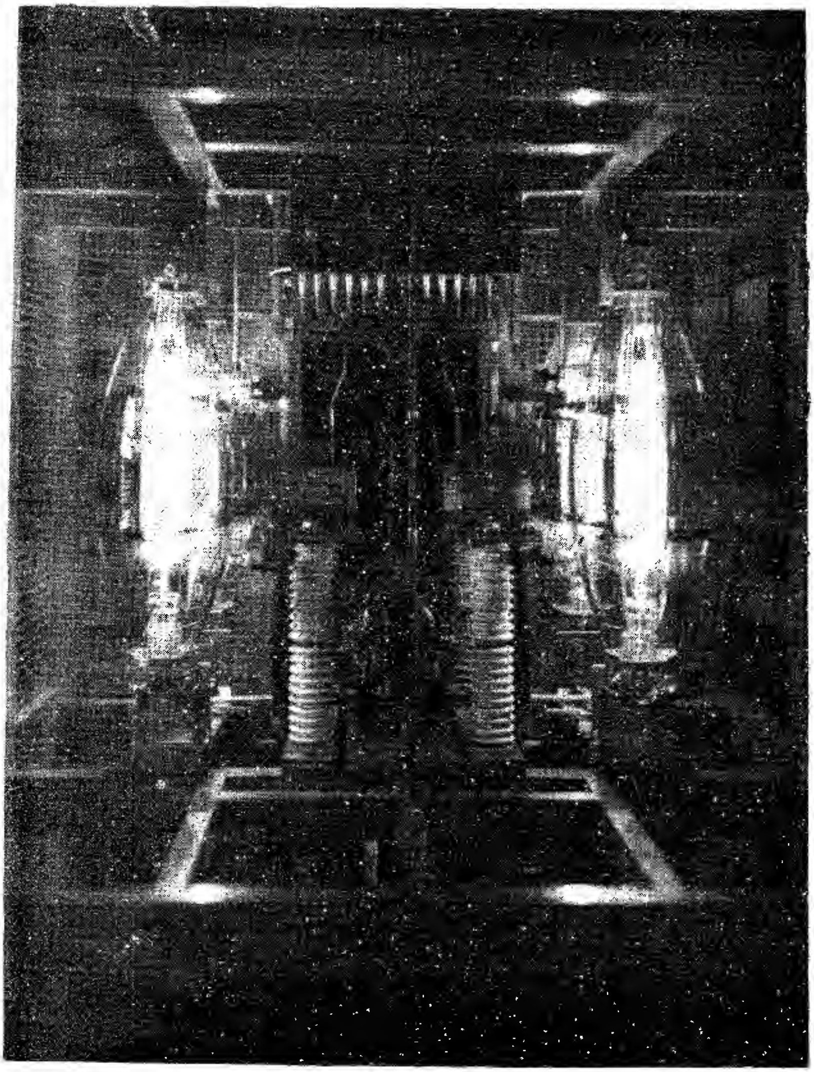
1.1 Radio Transmission and Radio Reception are concerned with the building up to high values of minute voltages and currents generated at various sources, such as microphones, oscillators and aerials. In order to accomplish this effectively, Amplifiers are required which possess no inertia, amplify without distortion, operate economically and require a minimum of maintenance and adjustment.

The Thermionic Valve Amplifier possesses most of these features. This amplifier may be regarded as a pure resistance device, hence the only frequency discrimination involved (except for frequencies in the region of 100 Mc/s or higher) is that due to the associated circuits. By using a number of amplifiers in series (or cascade), signals may be amplified many millions of times.

A Thermionic Valve Amplifier may thus be defined as a device for increasing the energy from a source without appreciably altering the quality. This Paper will only deal with the performance of valve amplifiers when operating in circuits associated with alternating currents and voltages. The frequencies at which these circuits function may vary from relatively low audio frequencies (below 15,000 c/s) to relatively high radio frequencies (extending beyond 330 Mc/s). A radio signal is usually a combination of both audio and radio frequencies, and the valves are required to amplify and rectify these currents.

These alternating voltages to be amplified are applied to the grid of a valve and act in addition to the steady anode battery voltage. Since the current which flows through the valve is carried by very light and easily influenced electrons, changes in the grid voltage can take place very rapidly, and the anode current will respond instantly. Even if the frequency of the alternating voltage applied to the grid is 50,000,000 c/s, the anode current can follow this frequency readily, so that valves can be used to amplify and detect at almost any frequency.

At present, these actions are treated in an elementary manner, in order to give a general idea of the functioning of the valve. The matter is treated in greater detail later.



POWER AMPLIFIER AT "VLR" SHORT-WAVE STATION.

Abbreviations. The abbreviations frequently used in this Paper are -

Z_L = load impedance.

E_{ZL} = volts across load.

I_a = anode current.

r_a = anode impedance.

E_g = input signal to grid.

E_o = output voltage.

g_m = mutual conductance.

μ = amplification factor.

P_{max} = maximum power.

Other abbreviations are given throughout the Paper.

1.2 Elementary Amplifier. The following is an elementary discussion of the action of the valve as an amplifier with a reference to input and output power, and may be omitted by those familiar with these principles.

For this example, assume a valve having the grid-volt/anode-current curve shown in Fig. 1a and the grid-volt/grid-current curve of Fig. 1c. Let point A be chosen as the operating point (zero bias), and let an alternating voltage of 2 volts (R.M.S.) be applied to the grid under these conditions. The applied signal causes the grid voltage to vary between V and W, and this, in turn, causes the anode current to vary between points B and C on the curve. It is noted that the output signal is a replica of the input signal, that is, it is undistorted due to operation being on the straight portion of the curve.

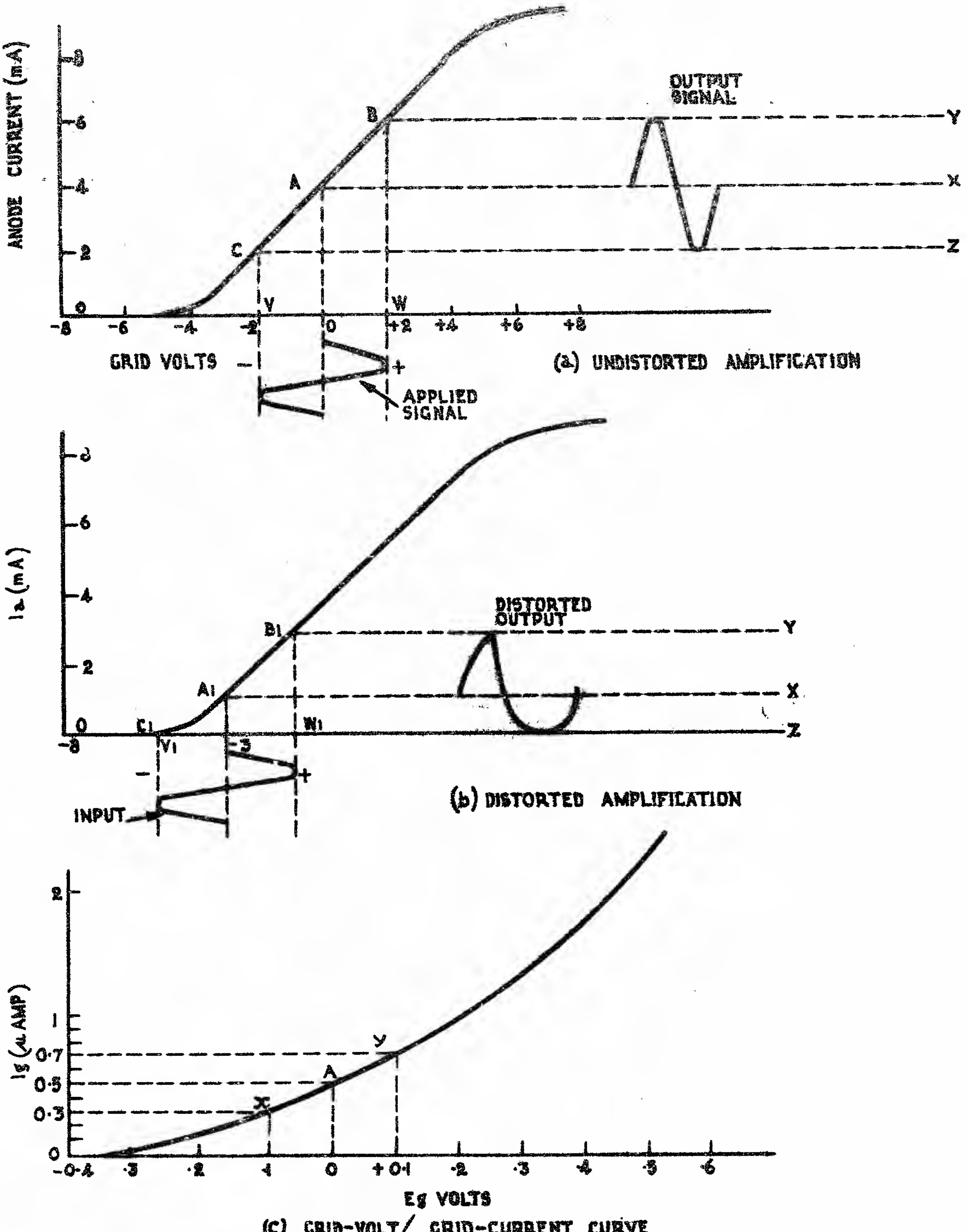
Now, as in Fig. 1b, change the operating point to A' (by increasing the bias to -3 volts). The applied signal now causes the grid voltage to vary between V' and W' and the anode current to vary between B' and C'. The output signal form is not a replica of the input signal, that is, distortion is present. This distortion is the result of operating too near the bottom of the curve, where bending occurs. In other words, the valve is operated so that the negative swing of the grid voltage causes the anode current to be cut off for a portion of the cycle.

In Fig. 1a, the positive half-cycle causes I_a to rise from 4 mA to 6 mA, an increase of 2 mA, and the negative half-cycle causes I_a to drop to 2 mA, a decrease of 2 mA. Thus, the variation is $\pm$ 2 mA and is symmetrical about the normal value of 4 mA. The alternating component thus has an R.M.S. value of 2 mA (that is, 1.414 mA effective).

In Fig. 1b, the positive half-cycle causes the anode current I_a to rise from 1 mA to 3 mA, an increase of 2 mA, but, on the negative half-cycle, the current cannot drop beyond zero current, which makes the decrease only 1 mA. Thus, there is a 2 mA increase on one half-cycle and a 1 mA decrease on the other half-cycle, and the variation is no longer symmetrical about the normal value and distortion occurs.

If the steady voltages acting on the valve are so chosen that the operating point is on a straight portion of the grid voltage-anode current curve, then variations in the grid voltage will produce similar variations in the anode current. This is true whether the variations of grid voltage are simple sinusoidal variations produced by an alternator, as shown, or of a complex form, such as the speech currents. (There is a second limitation, namely, that as soon as the grid voltage becomes sufficiently positive a grid current flows. This absorbs power from the input circuit, reduces the effective amplification and introduces distortion.) This explanation illustrates the use of a valve as an amplifier.

/ FIG. 1.



CURVES OF AMPLIFICATION.

FIG. 1.

The varying current of Fig. 1a is the same as would result if an alternating current and a direct current were both flowing in the anode circuit. The anode current can, therefore, be considered to have two components - one an alternating component and the other a direct component. The latter is the average value of the varying current. In Fig. 1a, the average current is 4.0 mA, which is the same value as the steady current flowing when an alternating voltage is not impressed upon the grid. The alternating component has an amplitude of 2 mA corresponding to an effective current of 1.414 mA. Because of the distorted form of the wave in Fig. 1b, however, it is evident that the average value of the current is greater when the alternating voltage is applied than it would be if no such voltage were acting on the valve. Since direct current instruments respond to the average value of the current flowing, a direct current ammeter in the anode circuit of the valve, under the conditions assumed for Fig. 1a, would show no change in reading when the alternating voltage is applied to the valve. The ammeter would, however, show an increased reading under the conditions of Fig. 1b upon the application of the alternating voltage. This change in average value of the current is of vital importance in the operation of the valve as a detector, and, in this use, the distortion produced by operating the valve on a bending portion of the characteristic serves a useful purpose.

1.3 Input and Output Power of Valves. The grid signal voltage assumed above (2 volts) is somewhat higher than usually encountered, so, for this example, a signal value of 0.1 volt will be assumed. Let this signal be applied to the valve. Refer now to Fig. 1c.

Input Power. The grid voltage will vary between points x and y on the curve, resulting in the following values of E_g and I_g -

Positive half-cycle E_g = 0.1 volt and I_g = 0.7 microampere

Negative half-cycle E_g = 0.1 volt and I_g = 0.3 microampere

Normal E_g = 0.0 volt and I_g = 0.5 microampere

The alternating component of the grid-current has an amplitude of 0.2 microampere (R.M.S.) which equals -

$$0.2 \times 0.707 = 0.1414 \text{ microampere (effective).}$$

The effective value of the grid voltage equals -

$$0.1 \times 0.707 = 0.0707 \text{ volt.}$$

The power taken from the source of grid voltage is equal to the product of the effective values of voltage and current -

$$\begin{aligned} W &= E \times I. \\ &= 0.1414 \times 10^{-6} \times 0.0707 \\ &= 1 \times 10^{-8} \text{ watts.} \end{aligned}$$

Output Power. In the previous example (paragraph 1.2), the effective value of the alternating component of the anode current was 1.414 mA. Since the signal voltage now being considered is 1/20th of that used above, the effective anode current will be 1.414 ÷ 20 which equals 0.0707 mA or 7.07×10^{-5} amperes. The value of the load resistance for this valve under the assumed conditions is approximately 10,000 ohms.

$$\begin{aligned} \text{Power output} &= I^2R \\ &= 7.07 \times 10^{-5} \times 7.07 \times 10^{-5} \times 10{,}000 \\ &= 50 \times 10^{-6} \text{ watts (approximately).} \end{aligned}$$

/ Thus,

Thus, the ratio of output power to input power is -

$$\frac{50 \times 10^{-6}}{1 \times 10^{-8}} = 5{,}000 : 1.$$

The output power is, therefore, 5,000 times as great as the input power. There is no violation of the law of conservation of energy in this result, for the output power is supplied by the anode battery, the grid of the valve acting merely as a control over the current being drawn from this battery. It is now justifiable to say that the valve is acting as an amplifier, because it has been shown that -

(i) The output current is of the same frequency and wave form as the input voltage, and

(ii) The output power exceeds the input power.

It is evident from the shape of the grid current-grid voltage characteristic that the current in the grid circuit can have a distorted wave form relative to the wave form of the applied voltage, in a similar manner to the case discussed above in connection with the anode current. Similarly, also, when the operation takes place over a curved portion of the characteristic, the average value of the grid current will be different, when the alternating voltage is impressed, from that flowing when the voltage is not impressed. This action is also utilised for the purpose of detection in circuits in which a grid condenser and grid leak are employed.

1.4 An amplifier (to expand the definition given earlier) is a device in which an input (voltage, current or power) is used to control a local source of power, so as to produce an output (voltage, current or power) which is greater than and bears a definite relationship to the input.

A thermionic valve amplifier, therefore, is one that employs thermionic valves to effect the control of power from the local source.

1.5 Amplifiers may be classified in a variety of ways descriptive of their character, properties, frequency range, etc. It is convenient in these notes to discuss amplifiers as follows -

(i) Summary of common types of amplifiers.
(ii) Amplifiers for audio frequencies.
(iii) Amplifiers for radio frequencies.

2. TYPES OF AMPLIFIERS.

2.1 A summary of the common classification types of amplifiers is as follows -

(i) According to predetermined operating characteristics -

(a) Class A amplifiers.
(b) Class B amplifiers.
(c) Class C amplifiers.

(ii) According to output requirements -

(a) Voltage amplifiers.
(b) Power amplifiers.

(iii) According to the method of coupling the valve into associated circuits -

(a) Transformer coupling.
(b) Resistance coupling.
(c) Impedance coupling.
(d) Direct coupling.

Reference is also made to amplifiers as being "Narrow-Band" or "Wide-Band" amplifiers, depending on the width of the band of frequencies which will be satisfactorily amplified. For example, an amplifier, such as the Intermediate Frequency amplifier of a radio receiver for normal broadcasting channels, might amplify a bandwidth of 15 kc/s, whereas a receiver for frequency-modulated signals might be required to amplify a band of 200 kc/s. The former would be termed a "narrow-band" amplifier and the latter a "wide-band" amplifier.

Before discussing amplifiers in more detail, reference is made to the more important characteristics which should be designed for or guarded against.

2.2 Amplifier Distortion. An amplifier should not introduce distortion into the signal passing through it, that is, the output wave should be a replica of the input wave. In order to ensure this, three conditions must be satisfied -

(i) The output must contain only the frequencies contained in the input.

(ii) The output must contain all the frequencies contained in the input, and the relative amplitudes of the various components must be the same as in the input.

(iii) If any component of the output is shifted in phase relative to the corresponding component of the input, all components must be shifted by the same number of electrical degrees of the fundamental cycle.

Failure to satisfy these three conditions results in three corresponding types of distortion -

(i) Amplitude or non-linear distortion.
(ii) Frequency distortion (or frequency discrimination).
(iii) Phase distortion.

Amplitude Distortion. Amplitude distortion may be defined as the generation in an amplifier of frequencies not present in the input signal, and is usually associated with a non-linear relation between output and input amplitudes. In thermionic valve amplifiers, amplitude distortion is the result of curvature of the dynamic characteristics of the valves. Amplitude distortion can be minimised by proper choice of valves, load impedances and operating voltages, and by avoiding too high an input voltage. Overloading is the very noticeable amplitude distortion that occurs when the input voltage is so large that the normal range of operation on the dynamic curve of the valve is exceeded.

Frequency Distortion. Frequency distortion is the variation of amplification or sensitivity with frequency of the impressed signal. Frequency distortion results from the dependence of circuit and inter-electrode impedances upon frequency, and can be minimised by proper design of input, output and interstage coupling circuits.

Phase Distortion. Phase distortion is the shifting of the phase of the output voltage or current of an amplifier relative to the input voltage or current by an amount that is not proportional to frequency. Phase distortion is due to the same causes as frequency distortion and may be minimised by the same methods.

2.3 Amplifier Types. The amplifier types referred to in the preceding general summary will now be treated in more detail.

(i) Class A Amplifiers. A Class A amplifier is one in which the grid bias voltage permits a steady anode current flow of such a value that the anode current varies directly with the grid voltage for the complete cycle of 360 electrical degrees. In other words, the valve is operated about the centre point of its characteristic E_g/I_a curve (see Fig. 1a). The operating conditions are so adjusted that the resulting output voltage, for an ideal Class A amplifier, is an exact reproduction of the grid voltage.

/ The

The characteristics of the Class A amplifier are -

Relatively low power output.
Relatively low anode power efficiency (maximum theoretical efficiency 50 per cent., average efficiency 25 per cent. to 30 per cent.).
Low distortion.
Anode dissipation is maximum at zero output.
Average value of anode current does not change during the cycle, so that the input anode power is constant.

It is usual, in practice, to operate a Class A amplifier in such a way that the grid potential is never positive at any point in its cycle; this is not, however, essential to the definition of a Class A amplifier.

In the majority of cases, where a small triode is operated at a low anode voltage, a typical set of operating conditions would be as follows -

Grid bias voltage equal to two-thirds of cut-off value (or anode supply voltage divided by 1-1/2 times the amplification factor).
Anode load impedance equal to double the anode slope resistance of the valve.
Alternating grid-voltage; peak value equal to the grid bias voltage.

Class A amplifiers may be found in the stages preceding the power stages in public-address systems, in audio systems, modulators, radio-frequency and intermediate-frequency amplifiers in receivers, etc.

(ii) Class B Amplifiers. A Class B amplifier is one in which the grid bias is approximately equal to the cut-off value, so that anode current flows only during the positive half-cycle of the alternating grid voltage, that is, anode current flows for approximately 180 electrical degrees.

The characteristics of the Class B amplifier are -

Comparatively high power output.
Relatively high anode efficiency (theoretical 78.5 per cent, practical about 65 per cent. at full output).
High distortion.
The anode dissipation is a minimum and fairly low at zero signal, increasing rapidly to approximately constant value at about 25 per cent. full output. The power input to the anode increases, however, with signal and power output until the peak output is reached. Therefore, the anode current is a variable, and the anode supply voltage requires good regulation.

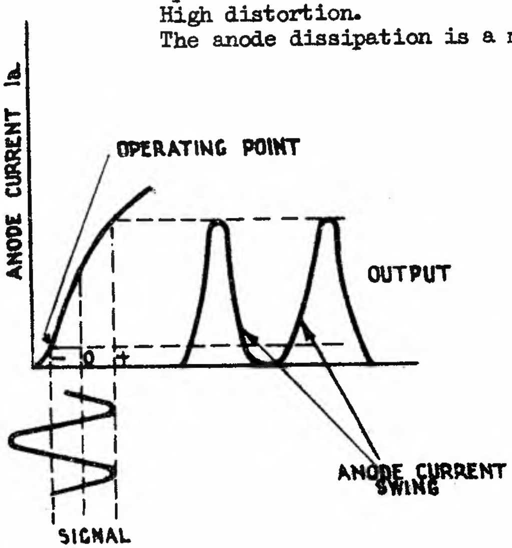


FIG. 2. CLASS B AMPLIFIER OPERATION.

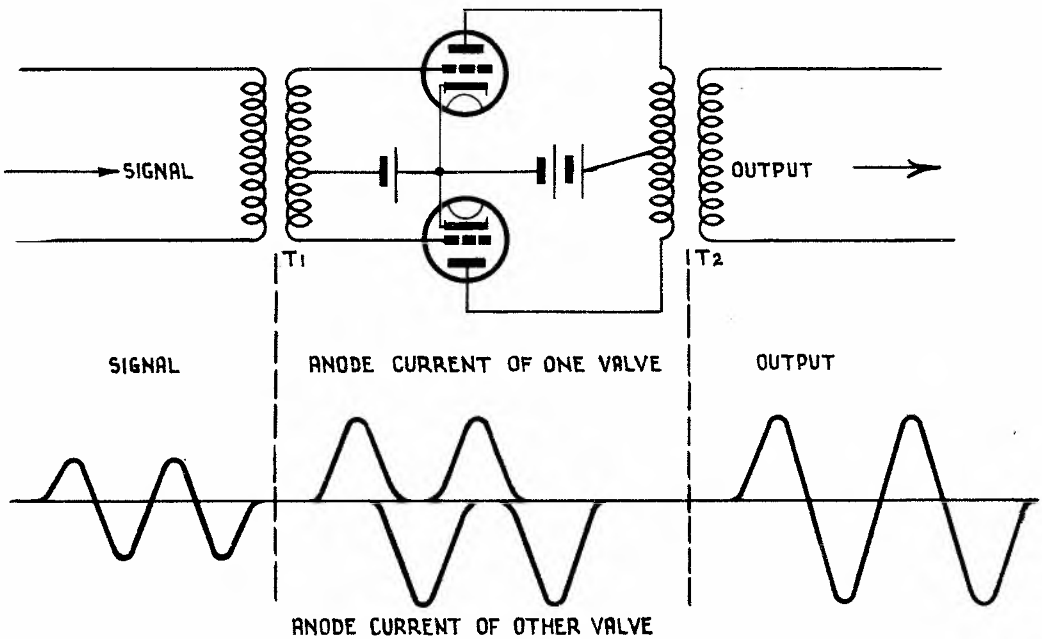
Fig. 2 shows Class B amplifier operation. Such an amplifier can be used in two ways -

(a) As an Audio Frequency Amplifier. For this purpose, it is necessary to operate two valves (or groups of valves in parallel) in a push-pull circuit. The negative half of the input wave for one valve is identical / with

with the positive half for the other valve, except for polarity. Thus, one half of the input wave is handled by each valve, the complete wave form being then reconstructed in the output transformer.

(b) To amplify a Modulated Radio Frequency Wave. The modulation wave form appears substantially without distortion on the positive halves of the radio frequency wave when the negative halves are suppressed. Hence, this function can be carried out by a single-ended amplifier. The radio frequency wave form is reconstructed by the resonant circuit which, in this case, always forms part of the radio frequency load circuit.

Class B Audio Amplifiers. For audio-frequency amplification, two valves must be used to permit Class B operation. Fig. 3 shows that, although the anode current pulses are of the same shape as the positive signal swing, considerable distortion at audio frequencies would be introduced if only one-half of each cycle were present in the output. For this reason, a second valve, working alternately with the first, must be included in the amplifier circuit, so that both halves of the cycle will be present in the output. A typical method of arranging the valves and circuit so that this end is achieved is shown in Fig. 3.



AUDIO AMPLIFIER (CLASS B).

FIG. 3.

The signal is fed to a transformer T_1. The secondary winding of T_1 is divided into two equal parts, with the valve grids connected to the outer terminals and the grid bias to the centre. A transformer T_2 with a similarly divided primary is connected to the anodes of the valves, the anode voltage being connected at the centre-tap. When the signal swing in the upper half of T_1 is positive, valve No. 1 draws anode current while valve No. 2 is idle; when the lower half of T_1 becomes positive, valve No. 2 draws anode current while valve No. 1 is idle. The corresponding voltages induced in the halves of the primary of T_2 combine in the secondary to produce an amplified reproduction of the signal wave-shape with negligible distortion. The Class B amplifier is

/ capable

capable of delivering much more power output for a given valve size than is obtainable from a Class A amplifier. In contrast to the Class A amplifier with its steady anode current, the average anode current drawn by the Class B audio amplifier is proportional to the amplitude of the exciting voltage. For reasons to be discussed in a later Paper in connection with the design of Class B modulators, valves most suitable for Class B audio service are generally those with high μ's.

Class B Radio Frequency Amplifiers. Class B radio frequency amplifiers are used as linear amplifiers to raise the output power level in radio telephone transmitters after modulation has taken place. A linear amplifier is a radio frequency power amplifier adjusted so that the voltage developed across the load is proportional to the exciting voltage applied to the grid of the valve. For this service, it is essential that the output power be proportional to the square of the excitation voltage, which varies at an audio-frequency rate. The valve can be driven into the upper-bend region of its characteristic, giving some flattening of the anode current pulse at the top, but, since the distortion is only present in the radio-frequency wave and not in the audio-frequency modulation, it can be filtered out in the resonant anode circuit.

In radio transmitters, Class B radio frequency amplifiers are often used where a fairly high power gain is required, even though it is not essential that the amplification be linear. With the bias set to cut-off, the excitation requirements are not as severe as with the high-efficiency Class C amplifier to be discussed later.

(iii) Class C Amplifiers. A Class C amplifier is an amplifier in which the grid-bias voltage is appreciably higher than the bias required for anode-current cut off, and the anode current flows for a period less than 180 electrical degrees during the half-cycle when the grid-swing is positive with respect to the bias voltage. The grid-swing is usually to the point of saturation, in which case the R.M.S. anode current is proportional to anode voltage and is not proportional to the grid voltage.

Alternatively, a Class C amplifier is one operated so that the alternating component of the anode current is directly proportional to the anode voltage. The output power is, therefore, proportional to the square of the anode voltage. An amplifier so operated is capable of being modulated linearly by anode voltage variations, as described in Paper No. 10. Other characteristics inherent to Class C operation are high anode efficiency, high power output and a relatively low power-amplification ratio.

The grid bias for a Class C amplifier is ordinarily set at approximately twice the value required for anode current cut-off without grid excitation. As a result, anode current flows during only a fraction of the positive excitation cycle. The exciting signal should be of sufficient amplitude to drive the anode current to the saturation point, as shown in Fig. 4. Since the grid must be driven far into the positive region to cause saturation, considerable numbers of electrons are attracted to the grid at the peak of the cycle, robbing the anode of some that it would normally attract. This action causes the drop at the upper bend of the characteristic, and also causes the anode current pulse to be indented at the top as shown in Fig. 4.

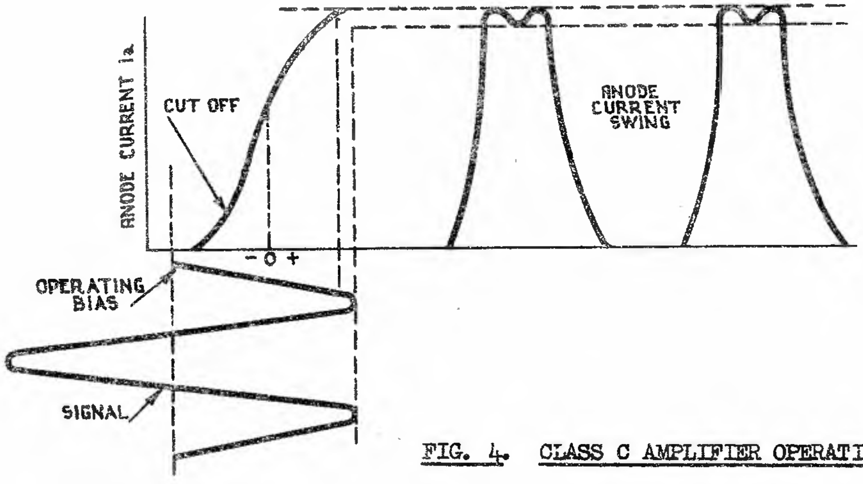


FIG. 4. CLASS C AMPLIFIER OPERATION.

The anode current wave form is not a replica of the grid voltage wave form or even of the positive half of it. Consequently, the Class C amplifier has no application in the amplification of audio frequency or modulated radio frequency grid voltages. Class C amplifiers are used with advantage for the amplification of radio frequency grid voltages of constant amplitude. The wave form is restored by a resonant circuit which is always associated with the load. By proper adjustment, including the use of a grid bias of at least double the cut-off value, the amplifier can also be made to operate with constant efficiency over a wide range of anode supply voltages, and hence can be used to convert a modulated direct current anode supply to a modulated radio frequency output.

Combinations of A and B Amplifiers. Since the basic A, B and C Class amplifiers represent three distinct steps in the operation of thermionic valves, it naturally becomes possible to adopt a set of operating conditions which uses two of the classifications although not adhering strictly to either. Such "midway" methods of operation can be classified as "AB" and "BC", but only the "AB" type of operation is in general use. The Class AB amplifier is a push-pull amplifier in which each valve operates for more than half but less than the whole cycle of the exciting voltage cycle. The amplifier bias is set so that the valves draw more anode current than in Class B operation, but less than for Class A operation. The anode current of the amplifier varies with the signal voltage, but not to the extent of the Class B operation. The Class AB amplifier is also occasionally called Class A Prime.

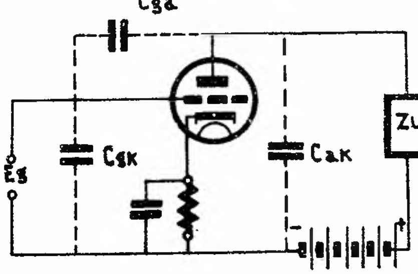
The efficiency and output of the Class AB amplifier are between those obtainable with Class A or Class B operation. Class AB amplifiers tend to operate as Class A amplifiers with low signal voltages and as Class B with high signal voltages, thus overcoming the chief objection in Class B operation - the distortion present with low-input-signal voltages. The Class AB amplifier is widely used where it is necessary to obtain a power output of considerable magnitude with a minimum of distortion.

2.4 Classification According to Output Requirements.

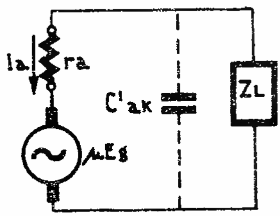
Voltage Amplifiers. A voltage amplifier may be defined as one which provides a greatly magnified reproduction of the input signal without regard to the power delivered to the load. Voltage Amplification is the ratio of the signal voltage available at the output terminals of the amplifier to the signal voltage impressed at the input terminals. The symbol generally used is μ.

Voltage amplifiers are usually operated in such a manner that amplitude distortion is small. Much can be learned regarding their performance, therefore, by taking into account only the fundamental components of anode current and making use of the equivalent-anode-circuit theorem. The results obtained are closely verified by laboratory experiments. When it is necessary to determine the harmonic content or to compute the amplification more accurately, the graphical methods associated with the use of a load line may be employed (see Paper No. 3).

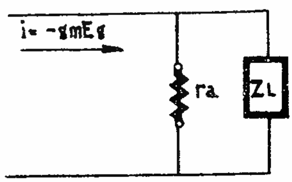
The simplest form of thermionic valve amplifier consists of a single valve with an impedance in the anode circuit, which is shown in Fig. 5a, and the simplified equivalent circuit is shown in Fig. 5b.



(a) Basic Amplifier Circuit.



(b) Simplified Equivalent Circuit.



(c) Equivalent for Pentodes.

FIG. 5. THERMIONIC VALVE AMPLIFIER AND EQUIVALENT CIRCUITS.

In the simplified circuit, the valve is replaced by a voltage generator delivering μE_g volts through the anode impedance of the valve and the load impedance Z_L in series. The grid-cathode capacitance C_{gk} and the anode-cathode capacitance C_{ak} are replaced by the capacitance $C^1_{ak} = (C_{gk} + C_{ak})$, and may be neglected at low frequencies because the reactance is high. Under this assumption, the voltage amplification is -

$$\mu^1 = \frac{E_{ZL}}{E_g} = \frac{I_a Z_L}{E_g} = \frac{\mu Z_L}{r_a + Z_L}$$

(See also Section 6.)

It will be seen from this that the voltage amplification approaches the amplification factor μ when the load impedance Z_L becomes large in comparison with the anode resistance r_a. At high frequencies, where the reactance of C^1_{ak} is comparable with the anode resistance, the effective load impedance consists of the parallel combination of Z_L and the reactance of C^1_{ak}. This impedance causes the voltage amplification to fall off at high frequencies. The voltage amplification that can be realised in practice with triodes approximates 80 per cent. of the amplification factor of the valve. The type of equivalent circuit referred to above is not very convenient when dealing with pentode valves, and another type is often used to simplify the analysis of pentode amplifiers. This circuit is known as the constant-current form, and is obtained by considering the valve as generating a current $-g_m E_g$ that flows through the impedance formed by the anode-resistance of the valve in parallel with the load resistance (Fig. 5c).

This circuit expresses the same relation as the constant-voltage form and leads to the same result as far as the load is concerned. The constant-current circuit is the most convenient to use in practical calculations when the anode resistance of the valve is much higher than the load resistance, as is the case with pentode and beam valves.

The constant-voltage circuit is most convenient when the anode resistance is of the same order of magnitude or less than the load resistance, as is commonly the case with amplifiers using triode valves.

Power Amplifiers. A power amplifier is one designed to deliver a relatively large amount of power to a load (for example, a loud-speaker in the case of an audio amplifier, or an aerial in the case of a radio-frequency amplifier).

Generally speaking, the last stage of any amplifier, whether audio or radio, is a power amplifier, since power is required for the operation of sound-reproducing devices and for the excitation of aerials. Amplifier stages preceding the last stage may be either voltage or power amplifiers, depending upon the purpose for which the equipment is designed. In audio circuits, the power valve or output valve in the last stage usually is especially designed to deliver a considerable amount of audio power, while requiring but negligible power from the input or exciting signal. The power amplification - ratio of output power to power supplied to the grid circuit - is consequently very high. Such valves generally require a large grid voltage swing for full power output, however, so that the voltage amplification - ratio of output voltage to signal voltage - is very small. Triode audio power amplifiers of this type often will give a power amplification ratio almost infinite, while the voltage amplification ratio may be less than 3 to 1. To obtain the voltage swing required for the grid circuit of such a power valve, it becomes necessary to use voltage amplifiers employing valves of high μ, which greatly increases the amplitude of the signal. Although such valves are capable of relatively high voltage output, the power obtainable is small. Voltage amplifiers are used in the radio-frequency stages of receivers as well as in audio amplifiers.

As explained earlier, the portion of the valve characteristic which can be utilised for distortionless amplification is limited. In radio-frequency circuits, where the input and output circuits are resonant, harmonic distortion of the radio-frequency wave form often can be neglected, since most of the harmonics so generated are filtered out in the tuned circuits with the result that the whole valve characteristic can be used. This leads to increased efficiency and higher power output for a given valve capacity.

/ To

To obtain high efficiency in the anode circuit, it is necessary that the grid be driven positive during part of the exciting signal cycle. During the time that the grid is positive with respect to the cathode, electrons are attracted to the grid and a flow of grid current results. This action, in turn, requires that the source of the exciting signal be capable of supplying the power. For this reason, it is usually found that all the amplifier stages in a transmitter, where high efficiency and maximum power output are desirable, are power stages. The voltage amplification in such a case is secondary.

In voltage amplifiers, the output voltage is proportional to the input voltage, but, in a power amplifier, the output power is proportional to the square of the input voltage, consequently greater input voltages are required for effective operation. If power is to be developed in the output circuit of a valve, the load resistance or impedance must not be too high. Thus, if a low resistance load circuit is required to absorb power, a valve having a low anode impedance must be used, as the maximum power will be transferred to the load impedance only when its ohmic resistance is equal to that of the valve impedance.

$$\text{Maximum power output } (P_0) = \frac{\mu^2 \times E_g^{\,2} \times Z_L}{(r_a + Z_L)^2} \text{ watts.}$$

The maximum possible undistorted power output may be obtained, however, when the load impedance is twice the anode impedance. This allows the introduction of a minimum amount of second harmonic distortion, which does not noticeably impair the quality.

For this case -

$$P_{max} = \frac{\mu^2 \times E_g^{\,2}}{8 \times r_a} \text{ watts.}$$

2.5 Classification According to Coupling Methods.

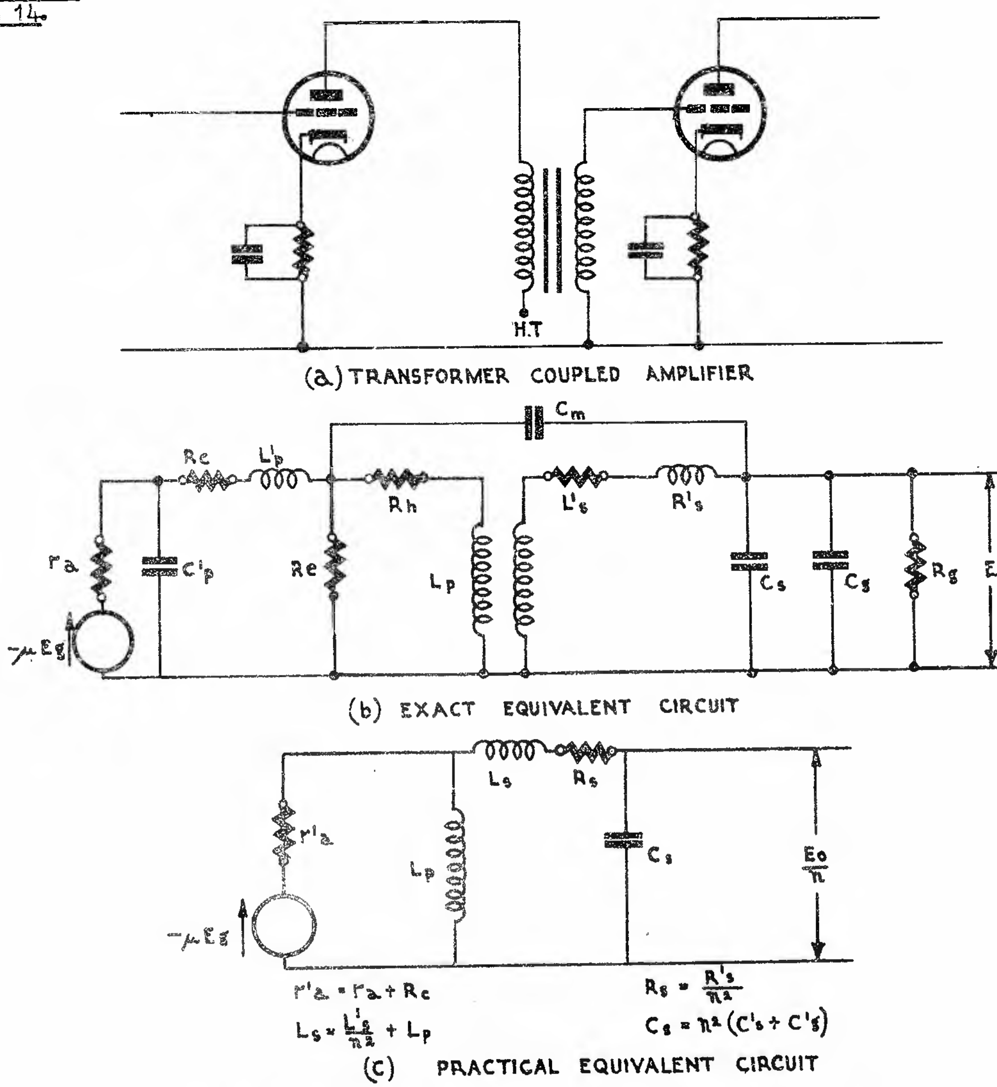
Transformer Coupling. In the transformer-coupled amplifier, the load impedance in the anode circuit of the valve is supplied by a step-up transformer (unless special conditions require a 1:1 or step-down ratio) that delivers its secondary voltage to the grid of another amplifier valve as shown in Fig. 6a. The amplifier valve exciting the transformer is ordinarily a triode having an anode resistance of approximately 10,000 ohms and a direct current anode current of a few milliamperes. Transformer coupled amplifiers are used primarily for supplying the exciting voltage of power valves. Compared with a resistance coupled amplifier, a transformer coupling has a low resistance in the grid of the succeeding valve. A transformer coupling will also permit a higher voltage to be applied to the anode of the valve, which will give a higher undistorted output. Transformer coupled amplifiers are more expensive than resistance coupled amplifiers, and provide less gain and reduced frequency response.

The more important applications of transformer coupled amplifiers are in the excitation of a push-pull amplifier from a single-ended amplifier stage, and in cases where the direct current resistance in the grid-circuit of the driven valve must be low because of grid current.

Due to the characteristics of the coupling transformer, the exact equivalent circuit is rather complicated. Fig. 6b shows the exact equivalent of Fig. 6a, and Fig. 6c shows the practical equivalent circuit.

In Fig. 7, the simplified equivalent circuits are shown for low, medium and high frequencies, since this approach lessens the complexity of calculating performance.

/ FIG. 6.



TRANSFORMER COUPLED AMPLIFIER AND EQUIVALENT CIRCUITS.

R_c = D.C. resistance of primary.

L_p' = Primary leakage inductance.

R_e = Resistance representing eddy current loss.

R_h = Resistance representing hysteresis loss.

L_p = Incremental primary inductance.

R_s' = Resistance of secondary winding.

L_s' = Secondary leakage inductance.

L_s = Total leakage inductance.

C_s' = Secondary distributed capacity.

C_p' = Primary distributed capacity.

C_m = Capacity between primary and secondary windings of transformer.

C_g' = Input capacity of valve to which E_o is delivered.

n = Ratio of secondary to primary turns.

FIG. 6.

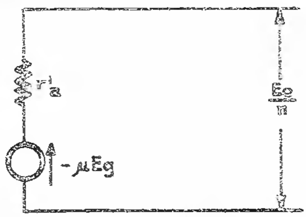
At low frequencies (Fig. 7b), the reactance of the leakage inductance and secondary capacitance can be neglected, but L_p must be taken into account. The shunting effect of the primary inductance L_p causes a dropping off at the low frequencies. At high frequencies (Fig. 7c), the primary inductive reactance has little effect, but the total stray capacitance C_s across the secondary has considerable influence. The reactance of C_s must be high at the highest frequency to be amplified.

For the medium frequencies, leakage and primary inductance can be neglected, and the amplification approximates the product of the turns ratio and the amplification factor, that is -

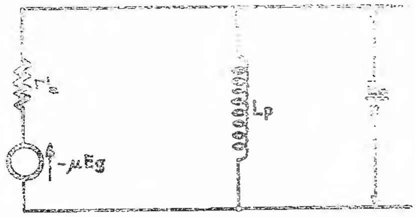
$$\text{Amplification at medium frequencies} = \frac{E_0}{E_g} = \mu n.$$

Transformers for Amplifiers. It is appropriate at this point to refer briefly to the main features of a transformer suitable for use in a transformer coupled amplifier. The correct design is not a simple matter, as a compromise must be made between conflicting requirements. A satisfactory response at low frequencies, for example, requires a high primary inductance, and high frequencies require low leakage inductance and small distributed capacitance. These latter two effects may be minimised by proper methods of winding and careful choice of shape and arrangement of coils, but, with any type of winding, the increase of distributed capacitance with number of turns affects the high-frequency response adversely. Since a high transformation ratio requires either a small number of primary turns or a large number of secondary turns, or both, it follows that the increase in voltage amplification, which results from the use of a high ratio, can be obtained only at the expense of poor response at low or high frequencies, or both.

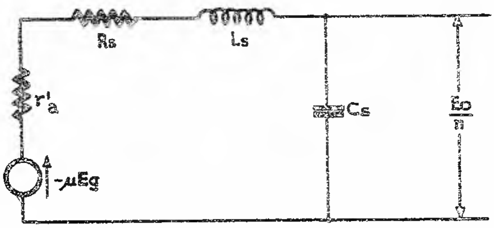
The common audio transformer employs a silicon steel core with a small air-gap, such as to give maximum primary inductance with the normal direct current anode current of the valve passing through the primary. The primary and secondary are layer-wound (secondary on the outside) with a turns ratio of about 2:1. The use of modern high permeability core materials, such as permalloy, increases the frequency range for a given size, or permits the design of a smaller transformer for a given range.



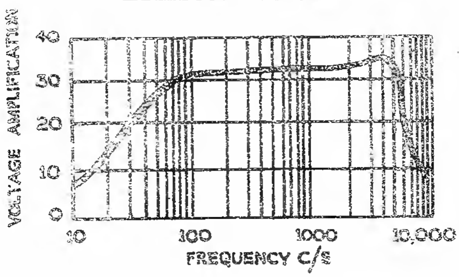
(a) Equivalent accurate for Medium Frequencies.



(b) Equivalent accurate for Low Frequencies.



(c) Equivalent for High Frequencies.



(d) Typical Frequency Response of a Transformer-Coupled Amplifier.

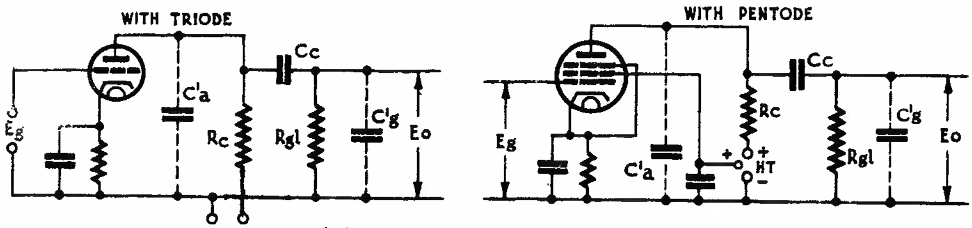
TRANSFORMER COUPLED AMPLIFIERS.

(EQUIVALENT CIRCUITS AND FREQUENCY RESPONSE.)

FIG. 7.

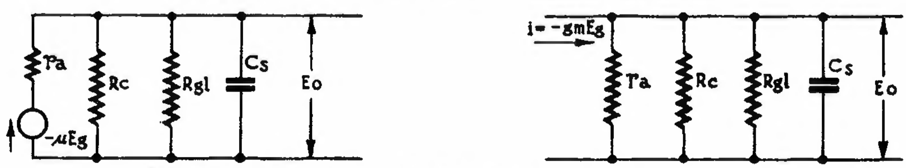
Resistance Coupled Amplifiers. In Voltage Amplifiers, the object is to obtain from the amplifier output as much voltage as possible to be applied to the grid of the succeeding amplifier valve. One way of doing this is to place in the amplifier anode circuit a high resistance load called the coupling resistance.

The circuits of practical resistance coupled amplifiers, using triode and pentode valves, are shown in Fig. 8a, in which R_c is the resistance load across which the amplified voltage is developed. The grid leak resistance and the coupling condenser are for the purpose of preventing the direct current voltage applied to the anode of the amplifier valve from also being applied to the grid of the valve to which the amplified voltage is delivered. The coupling condenser should be large enough to offer a low reactance to the frequencies to be amplified, while the grid leak should have a very high resistance in order that the shunting effect of the grid leak and coupling condenser upon the coupling resistance may be small. The most important property of the resistance coupled amplifier is the way in which the amplification varies with frequency.

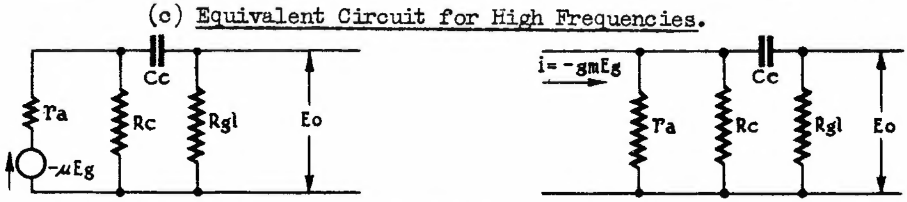


(a) Actual Circuits.

(b) Equivalent Circuit for Medium Frequencies.



(c) Equivalent Circuit for High Frequencies.



(d) Equivalent Circuit for Low Frequencies.

RESISTANCE COUPLED AMPLIFIERS.

e_o = Output voltage.

R_c = Coupling resistance.

$R_{g\ell}$ = Grid leak resistance.

C_c = Coupling condenser.

C_s = Total shunt capacity $(C_a' + C_g')$.

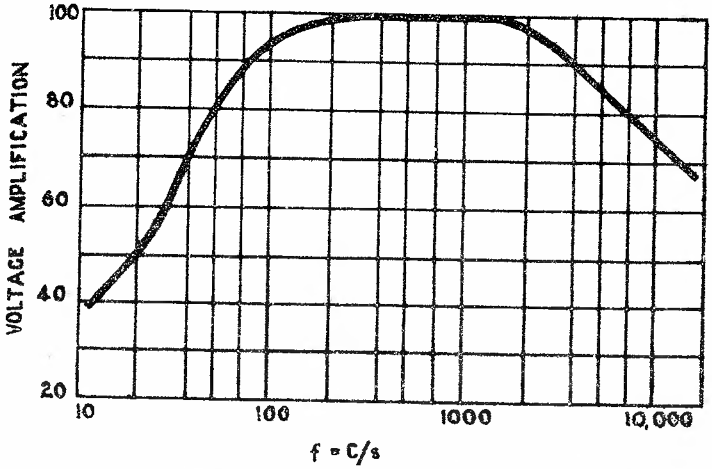
C_a' = Anode-cathode interelectrode capacitance plus capacitance of wiring to left of C_c.

C_g' = Input capacitance of valve plus capacity of wiring to right of C_c.

The constant-voltage equivalent circuit is used for the triode case, and the constant-current equivalent circuit is used for the pentode case as discussed previously.

FIG. 8.

Such a characteristic is shown in Fig. 9 for a representative case, and has, as its distinguishing feature, an amplification that is substantially constant over a wide frequency range but which drops off at both very low and very high frequencies.



TYPICAL RESPONSE OF RESISTANCE COUPLED AMPLIFIER.

FIG. 9.

The falling off at very low frequencies is a result of the fact that the high reactance, which the coupling condenser C_c offers to low frequencies, absorbs some of the low-frequency voltage that would otherwise be developed across the grid leak. The reduction in amplification at high frequencies is caused by the valve and stray capacities, which shunt the coupling and grid leak resistances and which have low enough reactance at high frequencies to lower the effective load impedance with a consequent reduction in the voltage developed at the output.

Fig. 8 shows the circuits of resistance coupled amplifiers, together with equivalent circuits and simplifications of the equivalent circuits. From these equivalent circuits, the following information is obtained -

For constant voltage generator (for triodes) -

Amplification in middle range of frequencies $= \dfrac{E_o}{E_g} = \mu \dfrac{R_L}{R_L + r_a}$

For constant current generator (for pentodes) -

Amplification in middle range of frequencies $= \dfrac{E_o}{E_g} = g_m R_{eq}$

Where R_L = The equivalent load resistance formed by the coupling resistance and the grid leak in parallel -

$$R_L = \frac{R_c R_{g\ell}}{R_c + R_{g\ell}}$$

and R_{eq} = The equivalent resistance formed by the anode resistance of the valve, the grid leak resistance and the coupling resistance, all in parallel. Generally, r_a is usually much higher than R_c and R_g, then -

$$R_{eq} = R_L = \frac{R_c R_{g\ell}}{R_c + R_{g\ell}}$$

At high frequencies, it is necessary to take into account the effect of the capacities shunting the coupling and grid leak resistances.

$$\frac{\text{Actual amplification at high frequencies}}{\text{Amplification in middle range}} = \frac{1}{\sqrt{1 + \left(\frac{R_{eq}}{X_s}\right)^2}}$$

Where $X_s = \frac{1}{2\pi f C_c}$ = Reactance of total shunting capacity C_s,

and R_{eq} = Resistance formed by anode, coupling and grid leak resistances all in parallel.

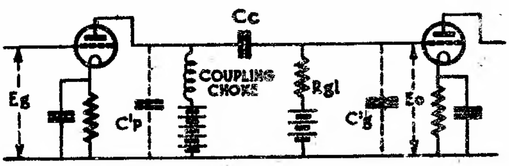
The extent to which the amplification falls off at high frequencies is, therefore, determined by the ratio which the reactance of the shunting capacity C_s bears to the equivalent resistance obtained by combining the anode resistance, coupling resistance and grid leak resistance in parallel.

At low frequencies, the shunting capacity C_s has such a high reactance as to be equivalent to an open circuit, but the reactance of the coupling condenser C_c becomes sufficient to cause a falling off in the amplification.

$$\frac{\text{Amplification at low frequencies}}{\text{Amplification in middle range}} = \frac{1}{\sqrt{1 + \left(\frac{X_c}{R}\right)^2}}$$

Where $X_c = \frac{1}{2\pi f C_c}$ = Reactance of coupling condenser C_c

and $R = R_{g\ell} + \frac{R_c\, r_a}{R_c + r_a}$ = Resistance formed by grid leak in series with the combination of anode and coupling resistances in parallel.

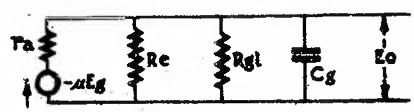


CHOKE COUPLED AMPLIFIER

EQUIVALENT FOR LOW FREQUENCIES

EQUIVALENT FOR MIDDLE RANGE

$$\frac{E_o}{E_g} = \mu \frac{R_1}{R_1 + R_a}$$



EQUIVALENT FOR HIGH FREQUENCIES

CHOKE COUPLED AMPLIFIER.

L = Inductance of choke.

$R_{g\ell}$ = Grid leak resistance.

R_e = Equivalent resistance of "L" due to eddy currents.

$$R_1 = \frac{R_e\, R_{g\ell}}{R_e + R_{g\ell}}$$

FIG. 10.

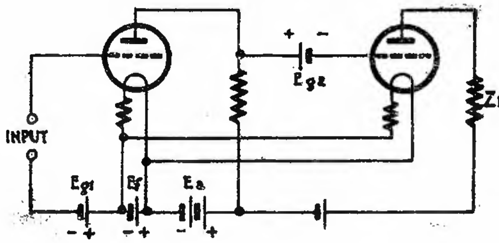
The extent to which the amplification falls off at low frequencies is, therefore, determined by the ratio of the reactance of the coupling condenser to the equivalent resistance obtained by combining the grid leak in series with the parallel combination of coupling resistance and anode resistance.

Impedance or Choke Coupling. This coupling method is of value when the anode supply voltage is not high, as for instance, the 130 volt anode supply battery of carrier equipment. This low voltage is not high enough to permit the use of resistance capacity coupling due to the voltage drop in the load resistance, particularly if a decoupling resistance is required.

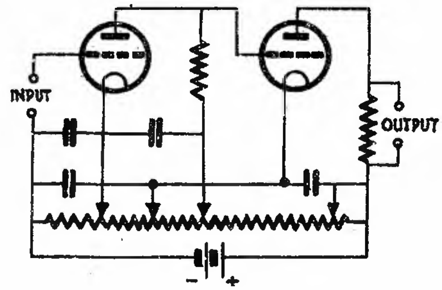
Fig. 10 shows the Choke Coupled Amplifier with its equivalent circuits. At low frequencies, the amplification is determined by the inductance of the choke, the reactance of which must be large for the lowest frequency to be amplified. The coupling condenser is large, so that the reactance is small at the lowest frequency. At high frequencies, the shunt capacities at the grid of the succeeding valve must be small, so that the reactance is high compared with the combined direct current resistances of the choke and grid leak in parallel.

/ Direct

Direct-Coupled Amplifier. A direct-coupled amplifier is one in which the anode of a given stage is connected to the grid of the next stage, either directly or through a biasing battery. Basic circuits of two-stage amplifiers with direct coupling are shown in Figs. 11a and 11b.



(a) With Batteries.



(b) With Voltage Divider.

DIRECT COUPLED AMPLIFIERS.

FIG. 11.

The biasing battery adjacent to the grid of the second valve in Fig. 11a is required to make the grid of the second valve negative with respect to its cathode. The several sources of voltage of Fig. 11a may be replaced by a single voltage source and voltage divider as shown in Fig. 11b. The function of the by-pass condensers, whose reactances are much smaller than the resistances they shunt, is to prevent the application to the grids of signal voltages caused by the flow of alternating anode currents through the voltage divider.

Direct-coupled amplifiers respond down to zero frequency, that is, they amplify changes of direct voltage. Although the small frequency distortion and the response at zero frequency are the advantages of this amplifier, the response to changes of steady voltage makes it difficult to use more than two stages without special design, such as the use of inverse feedback.

3. MISCELLANEOUS AMPLIFIER NOTES.

3.1 Amplifiers for Audio Frequencies and Some Relevant Data. Of the amplifiers just described, those suitable for use at audio-frequencies are -

Class A,
Class B } Push-pull only, and
Class AB }
Direct-coupled (limited application).

Class A may be used with resistance, impedance or transformer coupling.

Class B and AB may be used with either resistance or transformer coupling, though an output transformer is essential.

3.2 Some Advantages of Class B Push-pull Amplifier.

(i) Distortion is reduced by the balancing-out of the second-harmonics generated due to Class B operation.

(ii) The power output is greater than twice that of a single valve for the same distortion content.

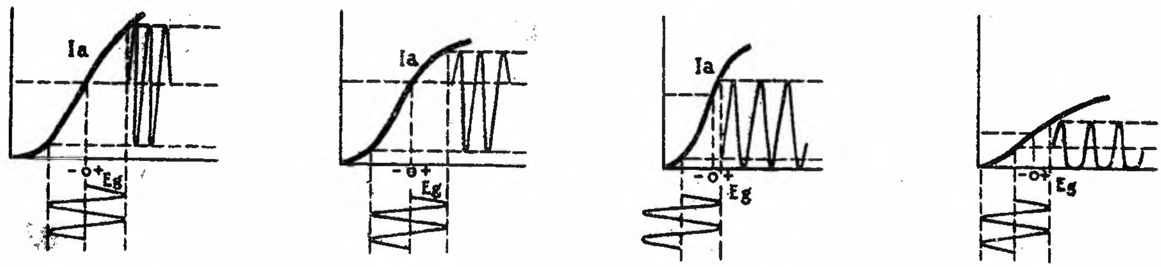
(iii) Owing to the fact that the direct current anode current components are always in opposite directions, the resulting magnetic fields cancel. This is advantageous in the case of the output transformer, because it prevents magnetic saturation of the iron core by the normal direct current anode current components. It also tends to simplify the design and, therefore, the cost of this transformer.

(iv) The possibility of a hum frequency due to the ripples developed in the power supply system is practically eliminated, since these ripples are in phase in the respective halves of the primary winding of the output transformer and will tend to cancel out.

3.3 Amplifier Distortion. Definitions of amplifier distortion were given at the beginning of this Paper, but a slight amplification of those remarks seems desirable. Some causes of amplifier distortion are -

Excessive grid swing.
Improper grid bias.
Improper anode voltage.
Improper balancing in push-pull circuit.
Core saturation of audio-frequency transformer.
Improper load-impedance matching.
Improper reflected impedance from secondary circuits.
Improper filtering.
Self-oscillation.

Any one of these conditions may cause an alteration of the anode-curve characteristic and produce an unsymmetrical wave-shape. Consider the anode-current characteristic shown correctly biased in Fig. 12a. If the bias is excessive, the result is distortion, as seen from Fig. 12b. If the bias is correct but the signal voltage is excessive, Fig. 12c results. If the load impedance is too low, the anode-current diagram resembles Fig. 12d.



(a) Correct Operation.

(b). Excessive Bias.

(c) Excessive Signal Voltage (Overload).

(d) Mismatched Load Impedance.

ILLUSTRATING DISTORTION IN AMPLIFIERS.

FIG. 12.

The total distortion in any amplifier is practically entirely dependent on the magnitude of the second and third harmonic components.

The presence of excessive distortion, due to harmonics, may be practically determined by inserting a milliammeter in the anode circuit. Since the generation of harmonics develops a condition resembling rectification, the milliammeter will rise or decrease about its normal reading while signal voltages are being received.

3.4 Power Output and Harmonic Calculations. These calculations have already been given for Class A, but are repeated for comparison with Class B.

/ Calculation

Calculation for-	Class A (Resistance Load)	Class B (Push-pull)
Power Output	$\dfrac{(E_{max} - E_{min})\,(I_{max} - I_{min})}{8}$	$\dfrac{I_{max}\,(E_a - E_{min})}{2}$
Anode Efficiency	$\dfrac{(E_{max} - E_{min})\,(I_{max} - I_{min})}{8\,E_a I_a}$	$\dfrac{\pi}{4}\,\left(1 - \dfrac{E_{min}}{E_a}\right)$
2nd Harmonic Distortion	$\dfrac{I_{max} + I_{min} - 2\,I_a}{4}$	
Load Resistance	$\dfrac{E_{max} - E_{min}}{I_{max} - I_{min}}$	$4\,\dfrac{E_a - E_{min}}{I_{max}}$ (Anode to Anode)

Where E_{max} = Maximum instantaneous value of anode voltage,

E_{min} = Minimum instantaneous value of anode voltage,

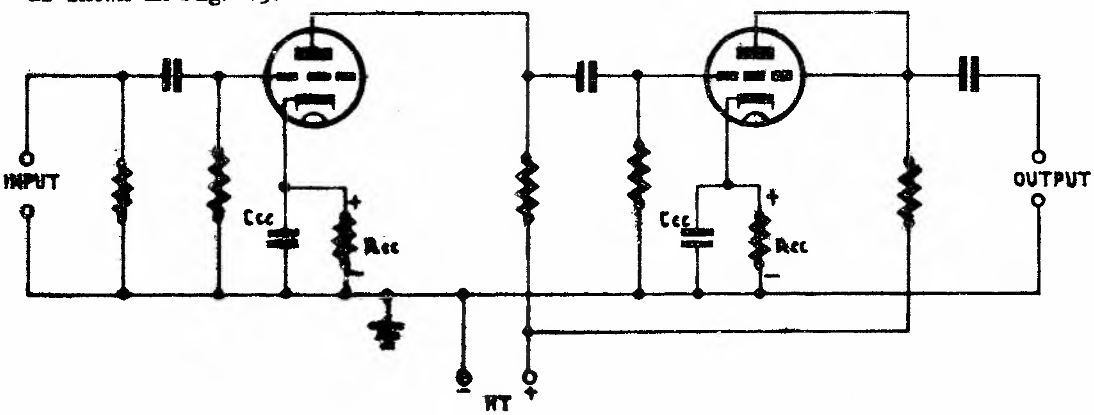
I_{max} = Peak anode current,

I_{min} = Minimum anode current, and

E_a = Anode supply voltage.

Note. The efficiency of a power amplifier may be defined as the ratio of power output that can be developed with a moderate distortion to the direct current power supplied to the anode.

3.5 Use of Cathode Resistors to Provide Bias. It is unnecessary to use a separate voltage source to supply grid-bias voltages. The normal steady components of anode and grid currents may be used to cause a voltage drop across a resistor in the cathode circuit as shown in Fig. 13.



R_{cc} = Bias Resistor. C_{co} = Decoupling or By-pass Condenser.

FIG. 13. AMPLIFIER WITH SELF-BIAS.

The steady anode currents passing through the bias resistors R_{cc} cause a steady voltage drop, which is of such polarity as to make the grid negative with respect to the cathode. The correct value of biasing resistance for any stage is equal to the required bias divided by the sum of the static operating screen and anode currents.

For example, let bias required be -6 volts, I_a = 4 mA and I_{sg} = 2 mA, then -

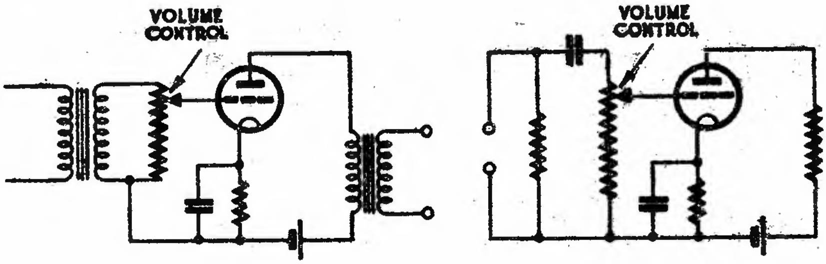
$$R_{cc} = \frac{6}{(4 + 2)\ \text{mA}} = \frac{6}{0.006} = 1{,}000 \text{ ohms.}$$

(**Note.** Milliamperes must be reduced to amperes.)

If the resistance alone is used, the signal voltage produced across this resistance is also applied to the grid, and, being in opposite phase to the input voltage, reduces the amplification. (This is actually a form of negative feedback, discussed later.)

This effect may be prevented by shunting the resistor with a by-pass condenser, C_{cc}, whose reactance is small at the signal frequency. To ensure that the amplification does not fall off at low frequencies, this condenser must have sufficiently high capacitance, so that the alternating voltage across R_{cc} is negligible at the lowest frequency to be amplified. This by-pass condenser also serves to reduce hum by preventing ripple voltage (which appears across R_{cc} due to poor B supply filtering) from being applied to the grid.

3.6 Amplitude (Volume) Control. The most common method of controlling the amplification of audio-frequency amplifiers is by means of a voltage divider, which varies the signal input voltage to one of the valves. The control must be located at a point of sufficiently low power-level in the amplifier that there is no possibility of stages being overloaded before the control can reduce the volume, and should not adversely affect the frequency response of the amplifier. Two typical methods are shown in Fig. 14, one for a transformer coupled amplifier and one for a resistance coupled amplifier. The potentiometers should be of 500,000 ohms resistance or higher.

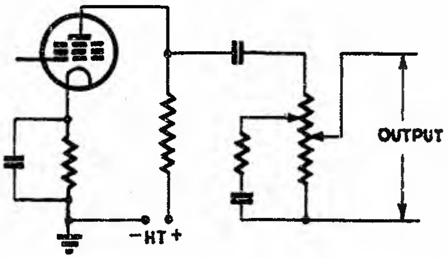


VOLUME CONTROLS.

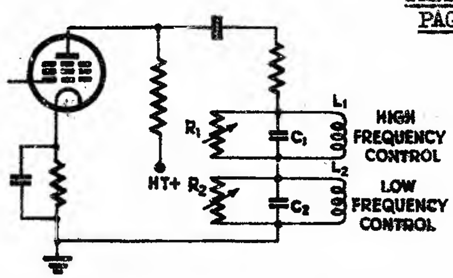
FIG. 14.

3.7 Tone Control. Variation in the acoustical properties of different rooms, non-uniform response of loud-speakers, individual preferences and the fact that the sensitivity of the ear falls off at low and high frequencies often make it desirable or necessary to change the frequency characteristics of audio frequency amplifiers. Sometimes tone control is combined with the manual gain control in such a manner that the amplification is reduced more at the middle of the audio range than at the upper and lower ends. This combination tends to correct the apparent loss of low and high tones at low sound level caused by the non-uniform response of the ear. Typical tone control circuits are shown in Fig. 15.

/ FIG. 15.



(a) Tone-compensated volume control.

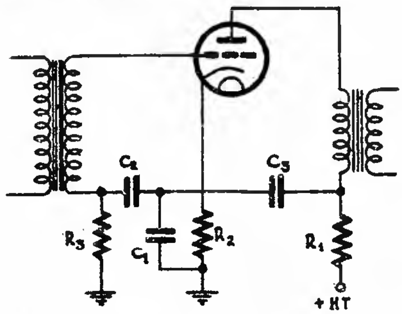


(b) Combined high and low frequency control.

FIG. 15. TONE CONTROLS.

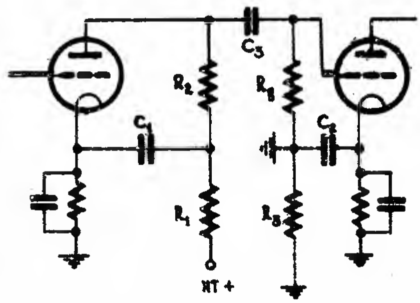
3.8 Decoupling or By-Passing. Another important consideration in the design of all types of amplifying systems, including both radio and audio frequency amplifiers, is the proper isolation between coupled stages to prevent interstage coupling and reactions. This is particularly important in high gain amplifiers, where the possibility of common coupling (common impedance) is prevalent, owing to the use of a single power supply for all stages. Common coupling may produce serious distortion in amplifiers, owing to the circuit reactions causing variations of grid voltages at different frequencies, thereby distorting the signal wave-shape. In many cases, the reactions may be great enough to produce whistling and howling, particularly in high gain amplifiers.

A common source of feedback in audio-frequency amplifiers is poor B supply regulation. Because of variation of voltage drop through the rectifier valves and filter, the terminal voltage of a B supply varies with the current drain. This type of feedback causes low-frequency oscillation, aptly termed "motor-boating". Typical decoupling circuits are shown in Fig. 16.



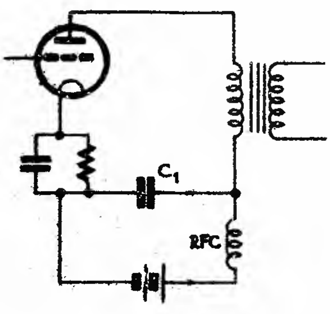
R_3 C_2 = Grid Filter Circuit
R_1 C_3 = Anode Filter Circuit

(a) For Transformer Coupled.

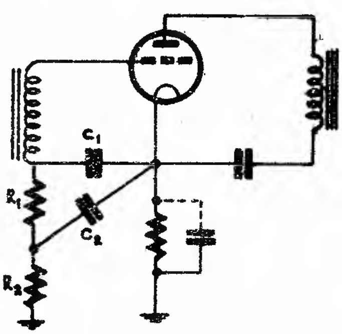


R_1 C_1 = Anode Filter Circuit
C_2 = Grid Filter Circuit (R_3 if Required)

(b) For Resistance Coupled.



(c) Choke Decoupling.



(d) Filtering in "Input".

FIG. 16. DECOUPLING CIRCUITS.

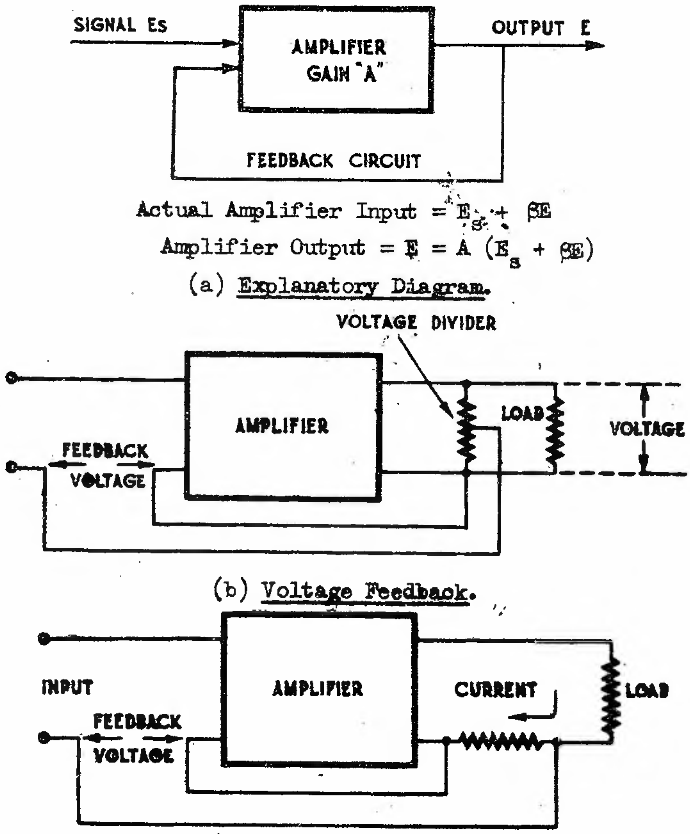
Chokes may be used as in Fig. 16c, particularly in radio frequency amplifiers. Chokes have the additional advantage of not appreciably reducing the operating plate voltage.

Fig. 16d shows the adaptation of filtering to the INPUT circuit of a high-gain amplifier. This point, of course, is the most critical in regard to hum. Suppose C_1 has a reactance of 1,000 ohms at 100 c/s (ripple voltage), then the first filter C_1R_1 will reduce this ripple voltage to 1/1,000th of its original value. The e.m.f. across C_1 is impressed on the second similar filter C_2R_2 and is further reduced by the same proportion; so the final reduction is 1,000,000 : 1. The complete stabilisation that may be obtained in amplifiers using this method has resulted in its wide adoption among radio engineers.

3.9 <u>Feedback Amplifiers.</u> Much improvement in the operating characteristics of an amplifier may be obtained by feeding back a portion of the output voltage to the input. An amplifier so designed is termed a "feedback" amplifier. If this feedback is in the proper phase relative to the input signal, the amplified feedback will tend to cancel the output noise and distortion. This advantage is obtained at the expense of reduced signal output, and so the feedback is said to be "inverse" or "negative". By proper design, it is possible to reduce the noise and distortion generated in the amplifier, to make the amplification substantially independent of electrode voltages and valve constants and to reduce greatly the phase and frequency distortion. The operation may be understood by reference to Fig. 17a. Here A represents an amplifier which has a gain A, when used as an ordinary amplifier, and which is supplied with a signal voltage E_s. Feedback is introduced by superimposing on the amplifier input a fraction β of the output voltage E_o, so that the actual input consists of a signal E_s plus the feedback voltage βE.

The effective gain of the amplifier is then equal to $\frac{A}{1 - A\beta}$.

(Note. When feedback is negative, β is negative.)



Actual Amplifier Input $= E_s + \beta E$

Amplifier Output $= E = A\,(E_s + \beta E)$

(a) Explanatory Diagram.

(b) Voltage Feedback.

(c) Current Feedback.

FIG. 17. FEEDBACK AMPLIFICATION.

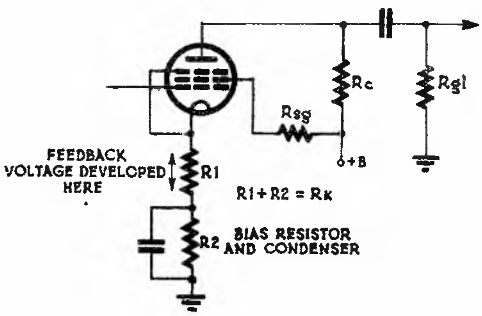
"Negative voltage feedback" occurs when the voltage fed back is proportional to the output voltage across the load, and provides a reduction of the effective internal resistance of the amplifier.

"Negative current feedback" occurs when the voltage fed back is proportional to the current through the output load, and provides an increase in the effective resistance of the amplifier. Fig. 17b shows these points.

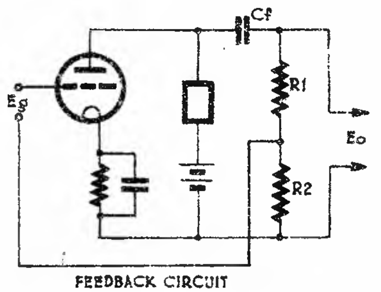
Figs. 17b, 17c and 18 show some methods of obtaining feedback. In Fig. 18a, the method is termed "current" feedback, in Figs. 18b and 18c it is termed "voltage" feedback and Fig. 18d shows the operation applied to a push-pull amplifier.

When negative feedback is applied to include the output stage, then -

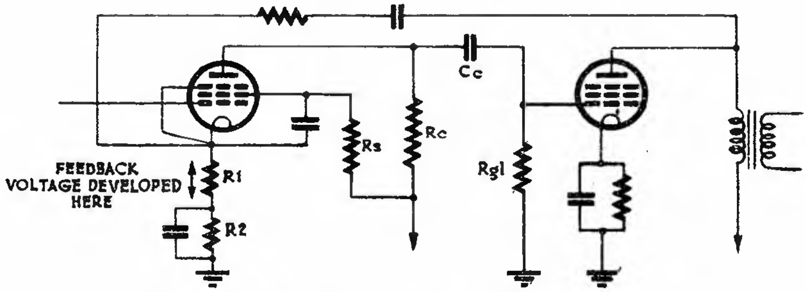
> Voltage feedback reduces the output impedance, and
> Current feedback increases the output impedance.



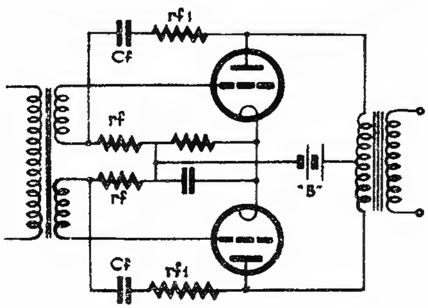
(a) CURRENT FEEDBACK
Rk NORMAL CATHODE RESISTOR VALUE



(b) VOLTAGE FEEDBACK
R1 AND R2 CONSTITUTE A VOLTAGE DIVIDER AND FEEDBACK PROPORTIONAL TO R1/R2



(c) VOLTAGE FEEDBACK OVER TWO STAGES



(d) PUSH PULL STAGE
rf = FEEDBACK RESISTANCES
Cf = ISOLATING CONDENSERS

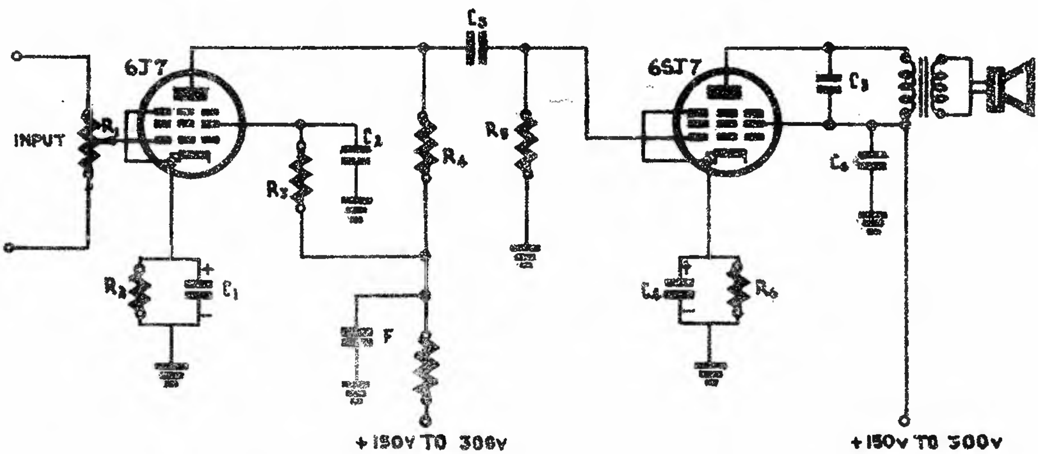
FIG. 18. FEEDBACK AMPLIFIERS.

Some of the advantages of feedback amplifiers are -

Reduction of amplitude distortion.
Reduction of noise.
Reduction of phase and frequency distortion.
Increase of stability.
Reduction of sensitivity or variation of amplification with input voltage.
Reduction of loud-speaker resonance.
Increased damping of loud-speaker transients.
A modification of effective internal resistance of the amplifier.

4. TYPICAL AUDIO AMPLIFIER CIRCUITS.

4.1 The circuits of several types of audio-frequency amplifiers are shown in Figs. 19, 20, 21 and 22.



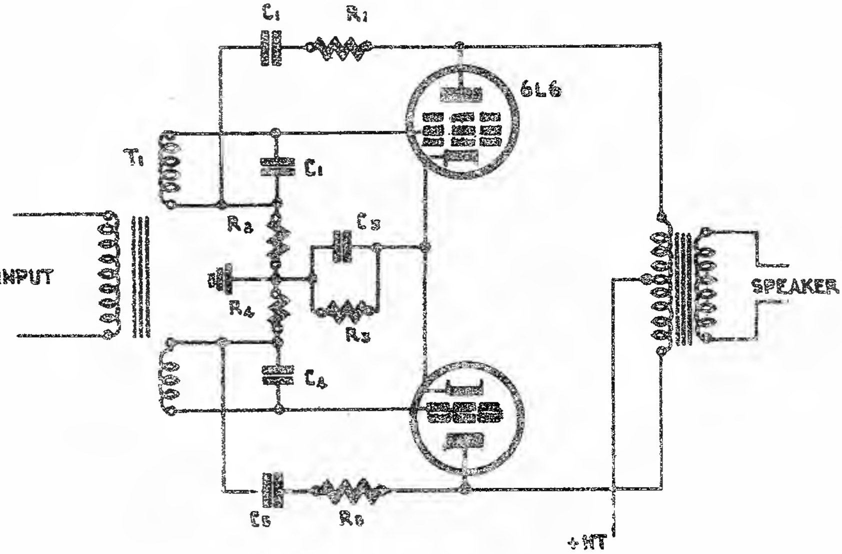
NON-MOTOR-BOATING AMPLIFIER (VOLTAGE GAIN = 9,000).

R_1 = Volume control potentiometer.
R_2 = 600 ohms bias resistor.
R_3 = 500,000 ohms.
R_4 = 100,000 ohms.
R_5 = 500,000 ohms.
R_6 = 600 ohms.

C_1 = 8 µF electrolytic 25 volts.
C_2 = 0.06 µF.
C_3 = 0.006 µF.
C_4 = 8 µF electrolytic 25 volts.
C_5 = 0.002 µF.
C_6 = 8 µF.

F = Decoupling filter as required.

FIG. 19.

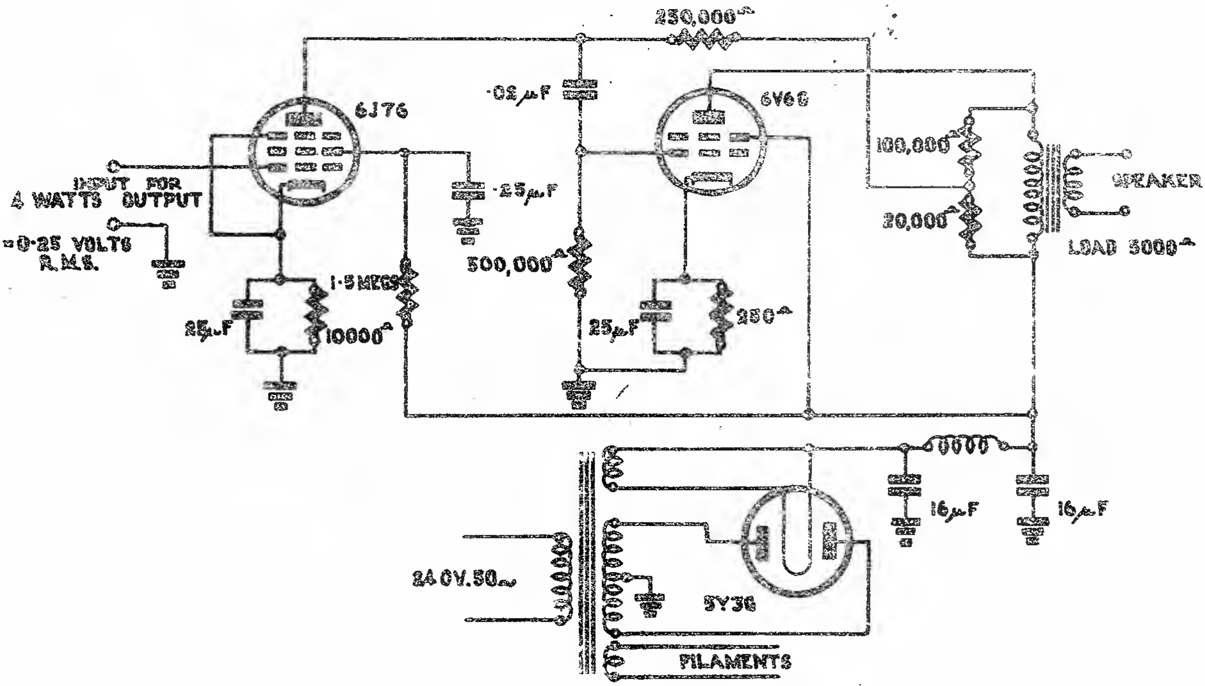


R_1 = 50,000 ohms.
R_2 = 5,000 ohms.
R_3 = 200 ohms.
R_4 = 5,000 ohms.
R_5 = 50,000 ohms.

C_1 = 0.1 µF paper.
C_2 = 0.00001 to 0.005 as required for filtering.
C_3 = 25 µF 25 volts as required for filtering.
C_4 = 0.00001 to 0.005 as required for filtering.
C_5 = 0.1 µF paper.

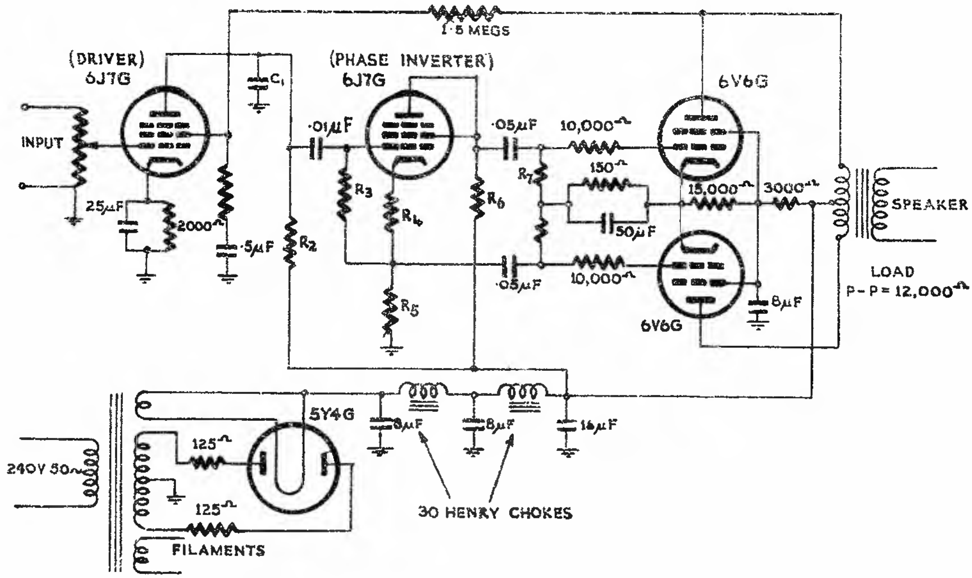
CLASS AB1, INVERSE FEEDBACK, POWER AMPLIFIER.

FIG. 20.



TYPICAL 4 WATT AUDIO AMPLIFIER WITH VOLTAGE FEEDBACK.

FIG. 21.



13 WATT AMPLIFIER PUSH-PULL, CLASS A1,
WITH NEGATIVE FEEDBACK OVER THREE STAGES.

R_1 = 30,000 ohms.	R_5 = 50,000 ohms.
R_2 = 250,000 ohms.	R_6 = 50,000 ohms.
R_3 = 1 megohm.	R_7 = 500,000 ohms.
R_4 = 5,000 ohms.	R_8 = 500,000 ohms.

C_1 = approximately 100 μμF.

FIG. 22.

5. RADIO FREQUENCY AMPLIFIERS.

5.1 Amplifiers for use in radio frequency circuits are usually associated with tuned circuits, and may be any of the three types Class A, Class B or Class C.

In practice, Class A amplifiers generally are found in crystal-oscillator circuits and in the radio-frequency and intermediate-frequency amplifier stages of radio receivers.

Class B amplifiers generally are found in buffer stages and following the modulated-amplifier stage in transmitters.

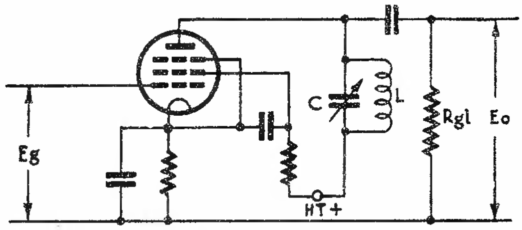
Class C amplifiers generally are found as modulators, where anode modulation is performed, and in power amplifier stages where the amplification of unmodulated carrier-wave is carried out. Class C amplifiers are also to be found in most types of oscillator circuits.

5.2 Tuned Voltage Amplifiers. A tuned voltage-amplifier is a Class A voltage amplifier in which the load impedance is supplied by a resonant circuit. Some typical circuits are shown in Fig. 23; these are used to amplify signal frequency voltages at radio frequencies in radio-receivers, including both radio frequency and intermediate frequency circuits. The important characteristics are -

(i) Amplification at resonance.
(ii) Variation of amplification with frequency in the immediate vicinity of resonance (see earlier notes on resonance curves).
(iii) The way the amplification changes as the resonant frequency of the tuned circuits is varied, for example, the way the amplification varies as a receiver is tuned over its tuning range.

/ Fig. 23

Fig. 23 also shows the approximate amplification formula at resonance of the circuits shown.



g_m = Mutual Conductance of Valve.

$\omega = 2\pi f$.

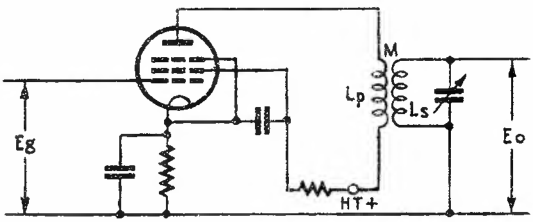
L = Inductance of Coil.

Q = "Q" of Coil.

$R_{g\ell}$ = Grid Leak Resistance.

Approximate Amplification At Resonance = $g_m\ \omega LQ$.

(a) Direct Coupling (Also Called Tuned Anode).



M = Mutual Coupling Between L_s and L_p.

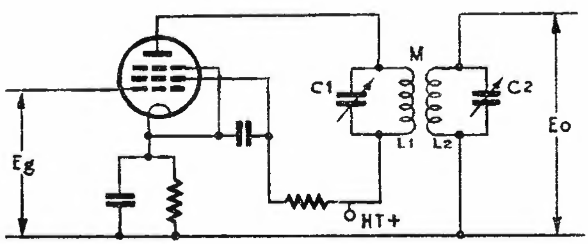
Q = "Q" of Resonant Circuit.

L_p = Inductance of Primary.

L_s = Inductance of Secondary.

Approximate Amplification At Resonance = $g_m\ \omega MQ$.

(b) Transformer Coupling.



L_1 = Inductance of Primary.

L_2 = Inductance of Secondary.

K = Coefficient of Coupling.

Q_1 = Effective Q of Primary.

Q_2 = Effective Q of Secondary.

ω_o = Twice Resonant Frequency.

Approximate Amplification when Primary and Secondary are Resonant at

$$\text{Same Frequency} = g_m K \frac{\omega_o \sqrt{L_1 L_2}}{K^2 + \frac{1}{Q_1 Q_2}}$$

(c) Band-Pass Coupling.

FIG. 23. TUNED VOLTAGE RADIO FREQUENCY AMPLIFIERS.

5.3 Class B Amplifiers. Class B amplifiers have been defined as those which operate with a bias approximately equal to cut-off, and in which the power output is proportional to the square of the grid-excitation voltage. Class B amplifiers are used in radio telephone transmitters to amplify the modulated carrier, and also in buffer stages immediately following an oscillator. It was stated that, with audio-frequency amplification, a Class B amplifier requires two valves in push-pull to avoid distortion. With radio frequency amplifiers, however, a single valve operating Class B will not distort the modulated envelope of the voltage applied to the grid. The tuned output-circuit restores symmetricality to the envelope, providing the dynamic characteristic of the tube is linear over the operating range.

/ Harmonics

Harmonics of the carrier frequency are present in the anode-current waves, but the use of two valves in push-pull remedy this and greatly reduce the burden imposed upon the subsequent tuned circuit of filtering out these undesired harmonics. These harmonics, if allowed to reach the aerial, would be radiated and would cause interference with other stations.

The instantaneous peak output of the valve at 100 per cent. modulation (see later) is four times the unmodulated output. The continuous power output with this degree of modulation is 1.5 times the output at zero modulation.

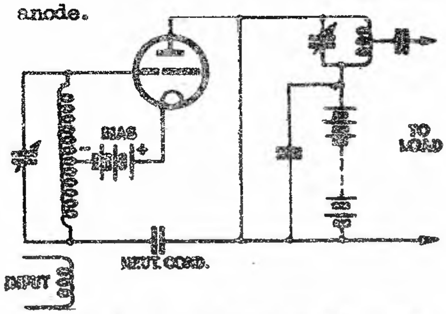
Anode Efficiency - 70 per cent. at 100 per cent. modulation.
33 per cent. unmodulated.

This type of amplifier is often referred to as a "Class B Linear Amplifier" or "Linear Amplifier."

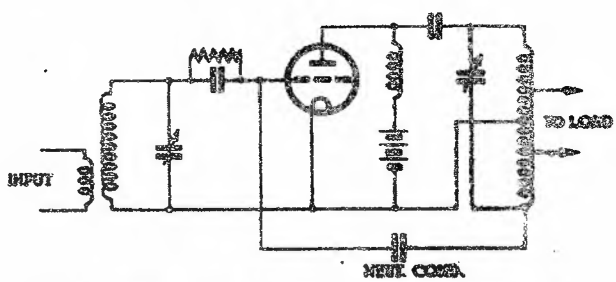
5.4 Class C Amplifiers. Class C amplifiers are used in radio-transmitters in the stages preceding the modulated stage. These amplifiers cannot ordinarily be used to amplify a modulated signal, since they are biased beyond cut-off and would cause serious distortion. Class C amplifiers are characterised by high anode efficiency, and are used primarily for the production of large quantities of power at a single frequency. Some typical circuit arrangements for Class C amplifiers are shown in Fig. 24. These circuits differ from each other in aspects such as methods of coupling, neutralisation and bias methods.

Triode valves are generally used for Class C high-power amplifiers, although pentodes and screen-grid valves have the advantage of not requiring neutralisation. These latter valves are not yet available in water-cooled types. The efficiency is high in a Class C amplifier, because current is permitted to flow only when most of the anode-supply voltage is used as voltage drop across the tuned load circuit, and only a small fraction is wasted in heating the anode.

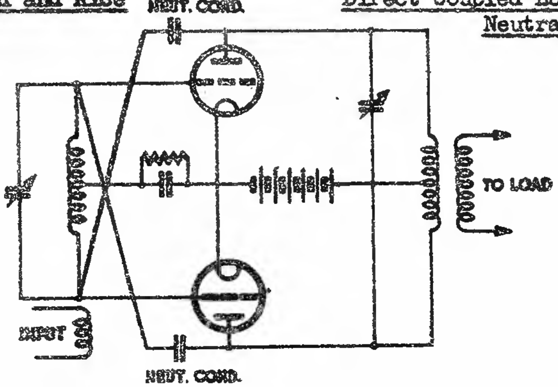
In Class A amplifiers, there is dissipation of power by the anode all the time, but, in Class C amplifiers, there is dissipation of power by the anode only a small part of the time. Thus, the Class C amplifiers may be driven to higher outputs without damaging the anode.



(a) Direct Coupling with Fixed Bias, Capacitively Coupled Load and Rice Neutralisation.



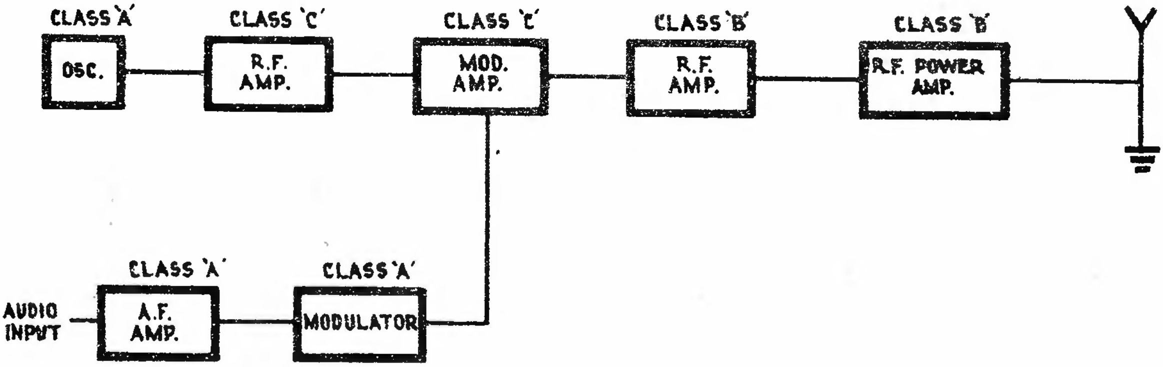
(b) Shunt-Feed Coupling, Grid-Leak Bias, Direct Coupled Load and Hazeltine Neutralisation.



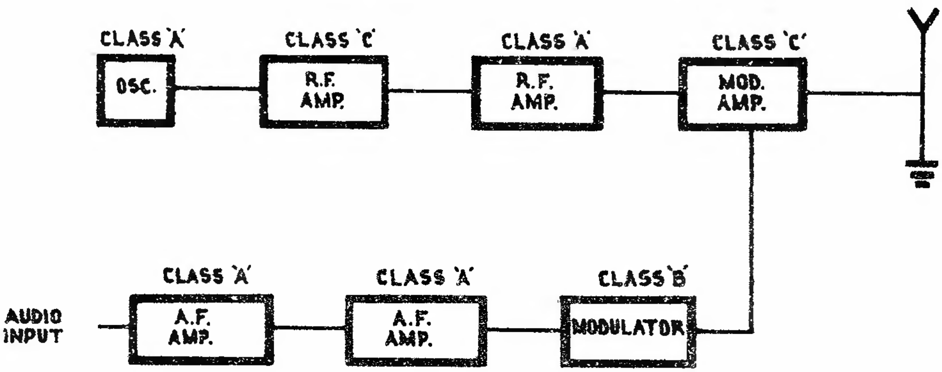
(c) Neutralised Push-Pull Amplifier with Grid-Leak Bias and Inductively Coupled Load.

FIG. 24. CLASS C RADIO FREQUENCY AMPLIFIERS.

5.5 Fig. 25 shows a radio transmitter showing examples of the use of Class A, Class B and Class C amplifiers.



(a) Low-Level Modulated Radio Transmitter.



(b) High-Level Modulated Transmitter.

TYPICAL RADIO-TRANSMITTERS IN BLOCK DIAGRAM FORM SHOWING THE POSITION OF B AND C CLASS AMPLIFIERS.

FIG. 25.

5.6 Neutralisation of Amplifiers. A discussion of radio-frequency amplifiers is not complete without reference to neutralising.

Since most high-power valves in transmitters are triodes, the grid-anode capacity of the valve elements (inter-electrode capacity) tends to provide a coupling between the input and output circuits. Since these circuits are usually tuned to the same frequency, this coupling is often sufficient to make the valve oscillate, and thus leads to instability, generation of harmonics, etc.

Neutralising may be regarded as a form of inverse feedback, in which sufficient out-of-phase voltage is fed back from the output to the input circuit to balance the voltage generated across the grid-anode capacity. There are two conditions to provide for -

Single-valve amplifiers, and
Push-pull amplifiers.

Single-valve Amplifiers. There are three neutralising circuits in general use for this case -

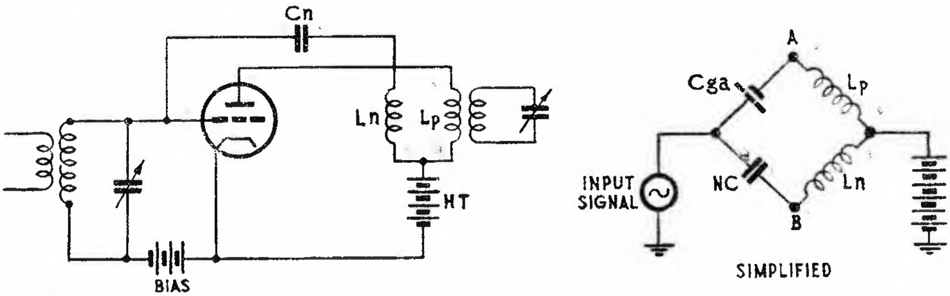
Neutrodyne or Hazeltine Method,
Rice Method, and
Coil Neutralisation.

/ Circuits

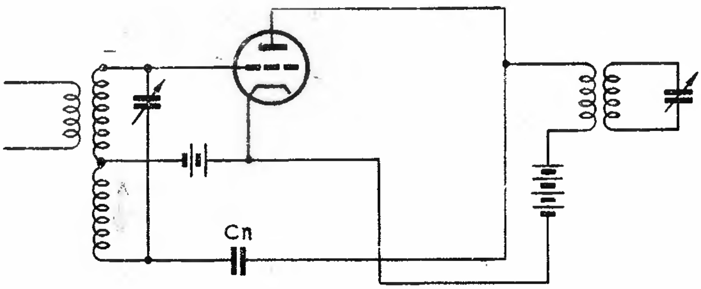
Circuits of these methods are shown in Fig. 26.

In the Neutrodyne system, the coil L_n is closely coupled to the primary and so polarised that it applies a voltage to the neutralising condenser C_n, which is of opposite phase from the alternating current voltage between anode and cathode. The circuit may be re-drawn in the form of a bridge (Fig. 26a, Simplified Sketch), where L_p and L_n are the tuning and neutralising coils, and C_{ga} and NC are the grid-anode and neutralising condenser capacities respectively. It can be seen that, provided L_p and L_n are equal when NC is adjusted to equal the grid-anode capacity, the circuit is balanced and the potentials at A and B are always equal and opposite.

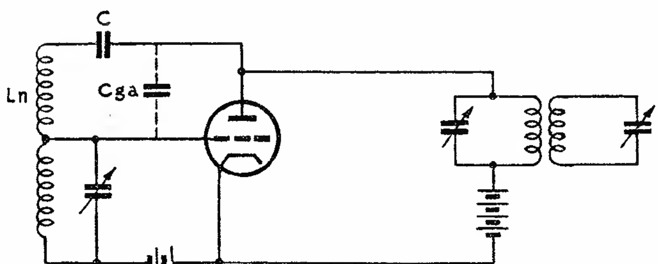
If the neutralising coil L_n is closely coupled to its primary and the leads have negligible inductive reactance, the neutralisation is substantially independent of frequency over a wide frequency band.



(a) Hazeltine or Neutrodyne Method.



(b) Rice Method.



(c) Coil Method.

NEUTRALISING SINGLE VALVE RADIO FREQUENCY AMPLIFIERS.

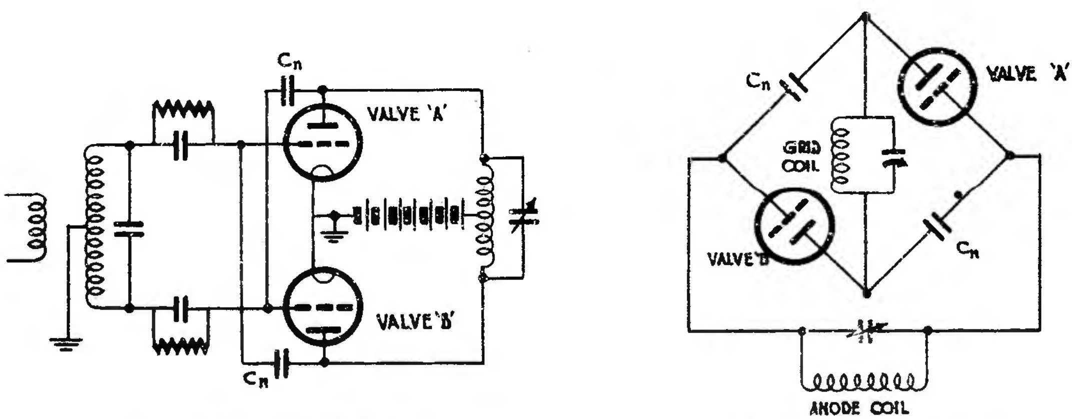
FIG. 26.

In the Rice circuit, the neutralising capacity C_n is adjusted so that the current through it neutralises the effect of the current through the grid-anode capacity, as far as the tuned circuit associated with the grid of the tube is concerned. The bridge diagram of the neutrodyne circuit is also applicable here, except that the grid-coil is balanced instead of the anode coil. This form of neutralisation is theoretically independent of frequency, just as is the neutrodyne type, but has the disadvantage that only half of the signal voltage developed across the input circuit is applied to the grid of the valve, and that neither side of the tuning condenser in the input circuit can be grounded.

In Coil Neutralisation, the neutralising inductance L_n is resonated with the grid-anode capacity C_{ga} at the frequency for which the neutralisation is to be effective. The coil is adjusted until it resonates with the grid-anode capacity of the valve, and thus forms a parallel-tuned circuit which offers a high impedance at the resonant frequency. The condenser C is merely a blocking condenser to prevent the short-circuiting of the anode supply voltage by the coil L_n.

Coil neutralisation is simpler than other types, but is effective for only one frequency, which limits its convenient application to broadcast transmitters operating at a single frequency.

Push-pull Amplifiers. Fig. 27a shows the push-pull or cross-neutralising circuit. This does not require the addition of special circuits other than the neutralising condensers, and is a form of neutrodyne that takes advantage of the fact that the voltages on the two sides of a push-pull amplifier are of opposite polarity, thus giving the phase relations required for neutralising. Fig. 27b shows that the push-pull circuit forms an easily balanced bridge, and, when balance is correct, there is no transfer of voltage from the grid circuit to the anode circuit.



(a) Circuit of Amplifier. (b) Diagrammatic Sketch of (a).

PUSH-PULL NEUTRALISATION.

FIG. 27.

5.7 **Parasitic Oscillations.** Often it is found that, even though a circuit is perfectly balanced, when the anode potential is applied, violent oscillations occur or voltage unbalances are present along earth leads and grid or anode leads.

These oscillations are generally high in frequency and occur through the valves and neutralising leads in a parallel path. Special suppressors are required to eliminate this form of oscillation - generally a radio frequency choke in parallel with a resistance right at the grid of the valve and in series with the grid lead.

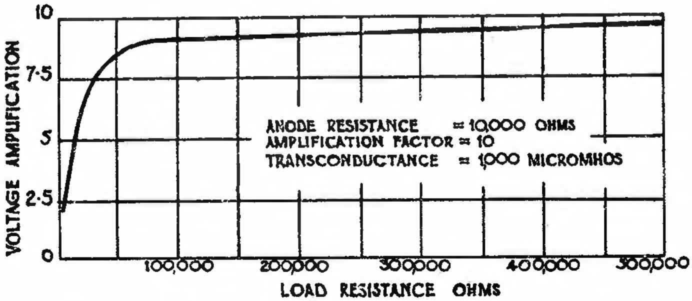
Oscillations of this type are more prevalent when valves are operated in parallel, and it is generally essential to provide non-inductive resistors of low resistances in each grid lead.

/ In

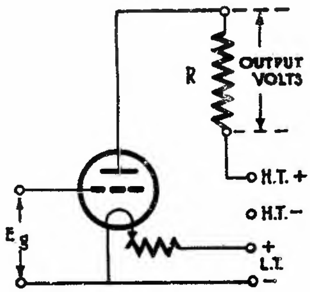
In the case of high gain audio amplifiers or in feedback circuits, it is often desirable or necessary to earth the anodes or grids of the stages at radio frequencies. Non-inductive condensers of 0.001 μF can be used for this purpose without affecting the circuit at audio frequencies.

6. EXPLANATION OF VOLTAGE GAIN CALCULATION.

6.1 The Voltage Amplification Factor, or Gain, of a stage may be defined as the ratio of output voltage variation across load impedance to the input voltage variation between grid and cathode of valve. This gain is not equal to the amplification factor of the valve alone, but can be made to approach it. Fig. 28a shows graphically how the gain approaches the amplification factor of the valve as the load is increased.



(a) Voltage Amplification versus Load Resistance for Theoretical Triode.



(b) Circuit.

VOLTAGE AMPLIFICATION FACTOR.

FIG. 28.

It is first necessary to determine the change in anode current I_a for simultaneous changes in grid volts E_g and anode volts E_a. The steady value of the anode current I_a is determined by the steady values of grid voltage E_g and anode voltage E_a. Suppose now that the grid voltage is changed by an amount e_g. If the anode voltage is kept constant, the corresponding change in anode current i_a will be given by -

$$i_a = g_m e_g$$

provided the change is confined to the straight portion of the appropriate mutual characteristic. It has been seen, however, that unless special arrangements are made to keep it constant, the anode voltage is affected by the change in grid voltage owing to the different current flowing in the external circuit. This change in anode voltage, in turn, affects the net change in anode current. If the anode voltage from the supply changes by an amount e_a, the change in i_a from this cause is connected with the change in E_a by the relation -

$$i_a = \frac{e_a}{r_a}$$

The total change in i_a due to simultaneous changes in E_g and E_a is, therefore, given by the sum of these two effects -

$$i_a = g_m e_g + \frac{e_a}{r_a}$$

Considering Fig. 28b, apply an alternating voltage variation of R.M.S. value E_g between grid and cathode of a valve as shown. As a result, oscillator variations will take place in the anode current flowing, but it cannot be said that these are given by -

$$i_e = g_m \times E_a$$

where g_m is the slope of the "static" mutual characteristic.

VARIATIONS IN ANODE CURRENT CAUSE VARIATIONS IN THE VOLTAGE BETWEEN ANODE AND CATHODE.

If the steady anode current is i_o and the voltage of the high tension battery is E_o, the actual potential difference between the anode and cathode is $(E_o - i_oR)$. Any change in anode current i_a results in a change in anode potential given by $E_a = -i_aR$. The general formula for the determination of i_a must, therefore, be used and combined with the last result to give the actual value of i_a. Thus, the two equations following are both applicable to the circuits of a Class A amplifier -

$$i_a = g_m E_g + \frac{E_a}{r_a}$$

$$E_a = -i_a R$$

Solving this by substituting for E_a in the first equation -

$$i_a = g_m E_g - i_a \times \frac{R}{r_a}$$

$$i_a \left(1 + \frac{R}{r_a}\right) = g_m E_g$$

$$i_a = \frac{g_m E_g}{1 + \frac{R}{r_a}} \qquad \text{.............................(1)}$$

Hence, the voltage "amplification factor", which is $\frac{\text{output voltage}}{\text{input voltage}}$, is given by -

$$\frac{i_a R}{E_g} = \frac{g_m R}{1 + \frac{R}{r_a}} = \frac{g_m R r_a}{R + r_a} = \mu \frac{R}{R + r_a} \qquad \text{..................(2)}$$

Equation (1) is an expression connecting the anode current variation with the grid voltage variation. Instead of the variation being g_m, it is now -

$$g_m - m = \frac{g_m}{1 + \frac{R}{r_a}} = \frac{i_a}{E_g}$$

Equation (1) might also be retained -

$$i_a = \frac{g_m r_a E_g}{R + r_a} = \frac{\mu E_g}{R + r_a}$$

The voltage amplification may also be conveniently found from the following formulae -

The voltage amplification equals -

(i) $\frac{\text{Amplification factor} \times \text{load resistance}}{\text{Load resistance + anode resistance}} = \frac{\mu R}{R + r_a}$

(ii) $\frac{\text{Transconductance in micromhos} \times r_a \times R}{10^6 \times (R + r_a)}$

/ It

It is shown in the first formula that the gain actually obtainable from a valve is less than the amplification factor, but approaches this value as the ratio of the load resistance anode resistance is increased.

7. TEST QUESTIONS.

1. What is the main difference between radio frequency and audio frequency amplifiers as regards input and output circuits?

2. What advantages accrue from the use of negative feedback? What is meant by -

   (i) Current feedback, and
   (ii) Voltage feedback?

3. Show by means of a sketch how distortion may be introduced by incorrect biasing of a valve in an amplifier.

4. Why is decoupling necessary in high-gain amplifiers? Illustrate your answer by a simple sketch of a two-stage amplifier.

5. Name three types of distortion which can occur in an amplifier and refer to several likely causes of their occurrence.

6. What is meant by Class A, Class B and Class C amplifiers? Which type would you use in the following stages of a radio transmitter?

   (i) Frequency multiplier stage.
   (ii) Unmodulated radio frequency amplifiers.
   (iii) Amplifier following the modulator.
   (iv) For best efficiency in a Heising type modulator.
   (v) In the speech amplifier units up to about 500 watts.

7. What is the purpose of neutralising an amplifier? Sketch a typical circuit associated with a push-pull or balanced amplifier.

8. Illustrate three common methods of coupling audio frequency amplifier stages, and briefly compare the merits of any two of them.

END OF PAPER.

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COURSE OF TECHNICAL INSTRUCTION.

ACOUSTICS, MICROPHONES AND LOUD-SPEAKERS.

CONTENTS:

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1. INTRODUCTION.

1.1 Radio Broadcasting is concerned with the translation of sound waves into electrical currents and radio waves, and converting these radio waves back into sound waves possessing all the characteristics of the original sound source. A brief review of the principles of wave motion and sound will first be given as an introduction to microphones and loud-speakers - the mouths and ears of radio!

RADIO AUSTRALIA SPEAKS TO THE WORLD!

1.2 Sound Waves. Sound, as explained in other books, is a sensation appreciated through the mechanism of the ears, when a sudden compression is produced in the atmosphere. This compression may be violent, irregular and not repeated, in which case it is recognised as a noise. Compressions may be created, for example, by the explosion of a bomb or gun, or by the sudden beating of the air by lightning, causing thunder.

If the compression is not violent and regularly repeated, it is recognised as a musical sound. If the compression is irregular and yet prolonged, it may be a sound called a speech sound. Sound waves, in general, find their origin in some sort of mechanical vibration, caused by impacts (hammering, drums), friction (violin, scraping) and other forms of energy dissipation in mechanical arrangements.

Musical sounds exhibit three qualities -

(i) Pitch or frequency,
(ii) Loudness or intensity, and
(iii) Quality or timbre.

1.3 Pitch or Frequency. Pitch is the number of compressions which strike the ear per second, and is the important characteristic that distinguishes musical sounds from the heterogeneous mixture of frequencies called noise. A male voice is of a lower pitch than a female voice, and a small bell has a higher pitch than a large bell. The upper limit of audible pitch for most people is approximately 16,000 c/s. Frequencies lower than about 32 c/s are felt rather than heard.

1.4 Loudness or Intensity. The intensity of a sound is really a measure of its energy, or the amplitude of its vibration. Intensity of sound is measured on an arbitrary scale, in which two sounds that stand to each other in energy ratio as 1 to 10 are said to differ by 1 bel (see Paper No. 1). Taking the threshold of hearing (an individually variable quantity) as having an intensity of 1 db, the approximate levels of some sources of sound are given in the following table -

Threshold of hearing	Unity (1 db)
A quiet whisper at 5 ft.	+ 10 db
Ordinary conversation at 3 ft.	+ 50 db
Congested city traffic	+ 75 db
Pneumatic riveting hammers	+ 90-100 db

There are many instances, such as frequency measurements and calibrating oscillators, where the frequencies of two alternating currents are varied until what is known as "zero beat" condition is obtained. Similarly, the production of two or more sounds of different frequencies results in a sound of periodically changing amplitude and hence intensity - the phenomenon of "beats". If the beat frequency is small, the individual pulsations are registered as such by the ear (say to about 30 c/s). If the beat frequency is within the audible range above about 30 c/s, a musical tone can be distinguished.

Fig. 1 shows the difference between a beat tone and a simple tone of the same frequency. If the tones producing the beat frequency are separated by 500 cycles and the simple tone has a frequency of 500 cycles, the effect on the ear will be similar.

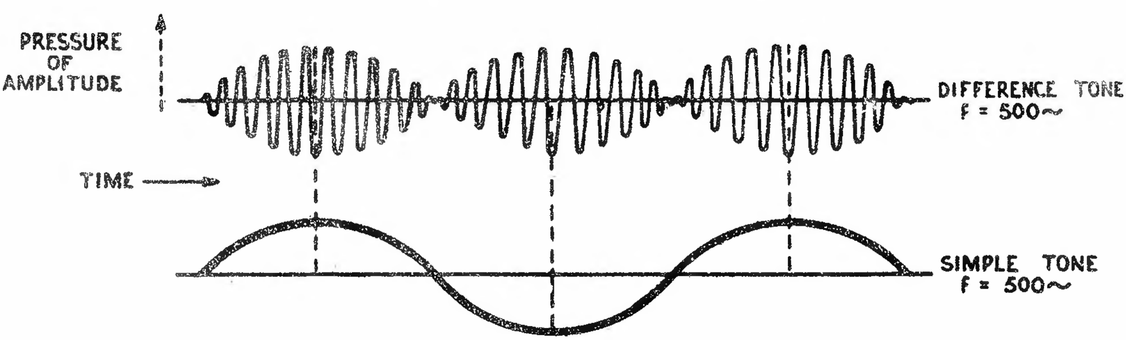


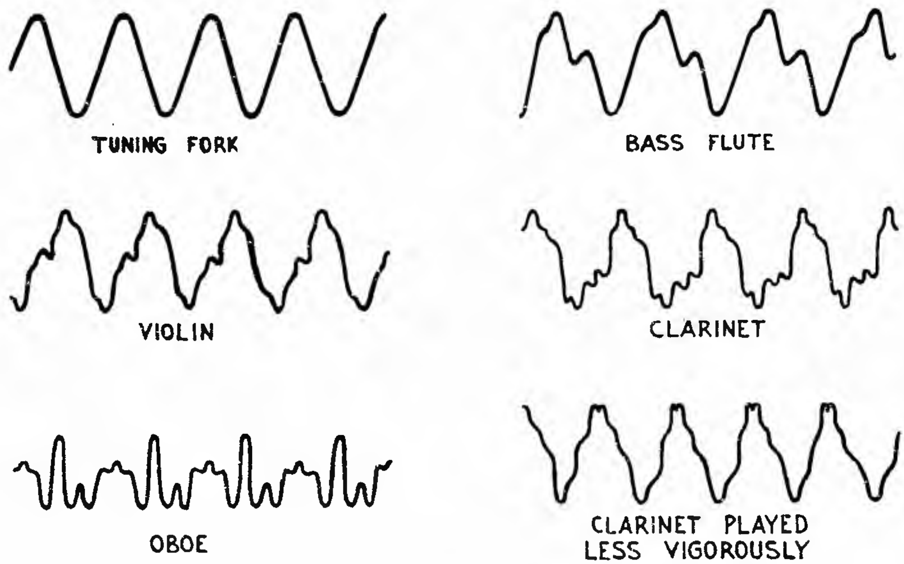
FIG. 1. BEAT TONE AND SIMPLE TONE OF SAME FREQUENCY.

1.5 Quality or Timbre. Quality is that characteristic of a sound which enables it to be distinguished from the same frequency sound (note) played on a violin, piano, trumpet, saxophone or made by the voice. This characteristic depends on the complexity of the vibration producing the sound or, in other words, the number of overtones or harmonics present.

/ For

For example, a note of 300 c/s is harmonically related to notes of 600, 900, 1,200 1,500, 1,800 etc., c/s, and it is the intensity and number of these harmonics present that determines the quality of the sound. Generally speaking, the harmonics do not have an amplitude as great as the fundamental, but the note can be rich in harmonics so that a high percentage of the total sound energy is contained in them.

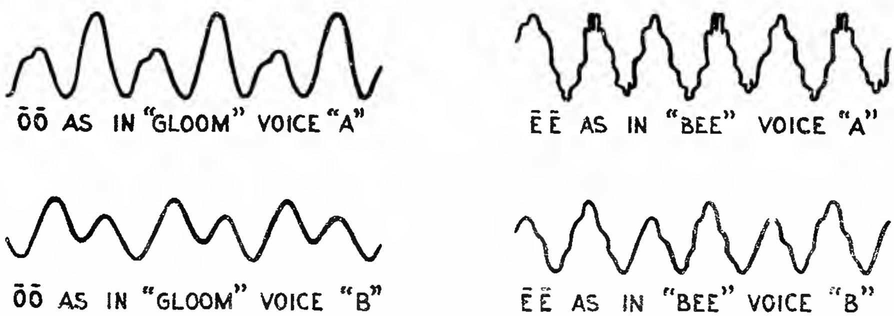
Fig. 2 shows the wave forms of the musical sound Middle C (262 c/s) as originated by several different instruments.



INSTRUMENT SOUND WAVES (MIDDLE C, 262 c/s).

FIG. 2.

Fig. 3 shows the difference between two baritone voices intoning the same sound.



SOUND WAVES (VOICE).

TWO BARITONE VOICES SHOWING INDIVIDUAL DIFFERENCES.

FIG. 3.

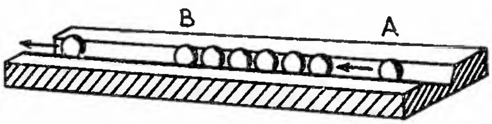
Since these differences in quality are brought about by the harmonic content of a sound, it will be apparent why so much thought has been given to designing microphones, loud-speakers, amplifiers, etc., capable of reproducing these harmonics.

1.6 Propagation of Sound. Sound is propagated through space by wave motion in the atmosphere or other medium. A movement of the sounding body causes compression or rarefaction in the surrounding medium. The compressed or rarefied atmosphere is elastic and possesses inertia, so that the particles comprising it, after having been displaced, return to and beyond their normal positions, causing the adjacent regions to be rarefied or compressed. These regions, in turn, act in a similar manner, transmitting vibratory energy to regions of the medium which are still more distant from the source of sound. This process continues until all of the energy is dissipated. The particles of the medium vibrate back and forth, but do not otherwise progress, although the vibratory motion and the energy of vibration are propagated through the medium.

1.7 Wave Motion. Energy of vibration or wave motion may be propagated through a medium in two ways. In some cases, as in a liquid or gas, the wave motion is parallel to the direction of propagation of the energy. In solids, the wave motion may be either parallel or perpendicular to the direction of energy propagation. These wave motions are respectively termed -

(i) Longitudinal wave motion.
(ii) Transverse wave motion.

1.8 Longitudinal Wave Motion. The propagation of sound waves through the air is entirely by longitudinal wave motion and may be appreciated by reference to Fig. 4.



ANALOGY OF LONGITUDINAL WAVE MOTION.

FIG. 4.

If a ball (say a billiard ball) is driven against the row at A, one ball will fly off the other end at B. The reason is that, when the ball is struck, it is momentarily compressed. In recovering its shape, it compresses the ball in front and so on along the row. The last but one drives off the last ball. Thus, a wave of compression travels down the row. If two balls, instead of one, are used at A, then two balls will fly off at B since two compression waves, one following the other, travel down the row.

A compression wave in air resembles this action. If vertical sections, of the air are likened to billiard balls then, if the first section is suddenly driven forward, the air in the neighbourhood of A is compressed. This air, in recovering, compresses the air in front of it, and so on. If, on the other hand, the first section had been pulled away, a lowering of pressure would have been produced and, in a similar manner, a rarefaction would have been produced.

It will be seen, therefore, that, if a piston is moving forwards and backwards, compressions and rarefactions are set up and the rate of motion determines the frequency, while the total movement of the piston determines the amplitude.

It can be shown that a longitudinal wave can be represented diagrammatically by means of a sine wave.

1.9 Transverse Wave. The most familiar case of a transverse wave is that which travels along a rope or string if one end is suddenly jerked sideways. If the rope is not stretched too tightly, the wave can be seen travelling along it. If the end of the rope is given a backwards and forwards motion, a succession of waves will travel along it.

A simple harmonic motion results, which can be represented as a sine wave as shown in Fig. 5, where the distance A to E is known as the wavelength. Wavelength is usually designated by a Greek character λ (lambda).

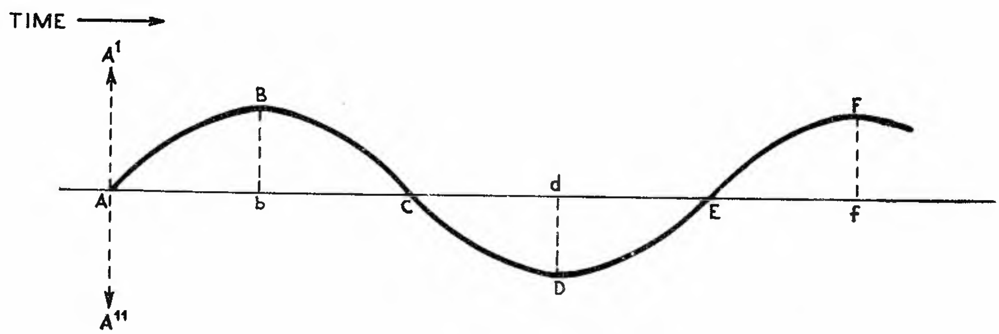


FIG. 5. TRANSVERSE WAVE.

1.10 Wavelength, Velocity and Frequency. From the instant at which the string commences to move until the jerk ceases, a finite time elapses. During this time, the wave motion will have travelled down the string away from the source of motion. Referring to Fig. 5, if A is the source of wave motion and A' and A" the sideways limits of the motion, then, during the time it takes to go from A to A' back through A to A" and back to A, the string will have described the double-arc A B C D E. The distance AE is called the "wavelength", the arc A B C D E a "complete cycle of motion", and the number of times A-E occurs in unit time (second) is termed the "frequency". It is apparent that the distance A-E will vary according to the number of cycles occurring in a second, since it is a function of time. Sound travels at about 1,100 ft. per second in air, so that, with a frequency of one c/s, the distance A-E would be 1,100 ft. and would be covered in one second. Similarly, if the frequency were 10 c/s, the distance A-E would be covered 10 times in one second, or the distance A-E would be $\frac{1,100 \text{ ft.}}{10} = 110$ ft.

Thus, Wavelength, Velocity and Frequency are related, and may be summarised -

$$\lambda = \frac{v}{f}$$

$$v = f \times \lambda$$

$$f = \frac{v}{\lambda}$$

where λ = Wavelength in same units as v,

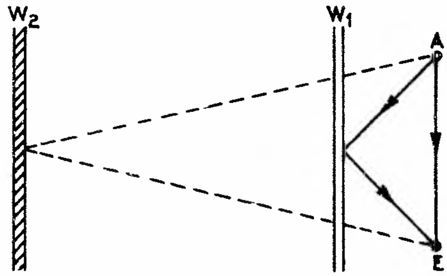
v = Velocity of sound (1,100 ft. per second in air),

and f = Frequency in c/s.

1.11 Reflection of Sound Waves. Where a compression meets a rigid wall, it can, in recovering, compress only the air behind it. The direction of a compression wave is, therefore, reversed on meeting a rigid obstacle. This process is known as reflection, and occurs whenever there is a discontinuity in the medium. Rarefactions behave similarly.

/ 1.12

1.12 Echoes. The reflection from a plain wall will give rise to a sound image. Thus, an observer at E (Fig. 6) receives sound waves directly from A, and also waves which have travelled from A to the wall and to E. When the wall is close to A and E, as for W_1, due to the small time interval between the two sets of waves, they cannot be distinguished from one another. In an ordinary room with plain walls, floor and ceiling, the only effect of the reflection is to increase the intensity of sound observed. Anyone who compares the difficulty of speaking or singing in the open air to that in a small room will appreciate this fact.



REFLECTIONS OF SOUND WAVES.

FIG. 6.

If, however, the distance from A to the wall is increased, as in W_2, so that the time of arrival of the direct and reflected waves is about 1/30th of a second, the sound, if of a sudden nature like a clap of the hands, will appear to be drawn out, that is, sustained. If the time of arrival is about 1/20th of a second, two sounds will be heard, and, at greater distances, two distinct sounds will be easily heard.

From the above, it will be seen that, depending on the distance which, in turn, controls the time of arrival of the direct and indirect waves, three conditions can be obtained -

(i) Sound re-enforcement (less than 1/30th of a second).
(ii) Interference (about 1/30th of a second).
(iii) Echo (about 1/20th of a second or greater).

1.13 Acoustics. The acoustic quality of nearly all sound waves from radio, telephone, motion pictures, gramophone recordings and public address systems is largely influenced by the acoustic properties of the rooms in which the sound to be transmitted or recorded is generated, and in which this sound is later reproduced. The term "live" and "dead" are often descriptively used to indicate the acoustic "feeling" of a room. A "dead" room is one possessing little or no reflection from the walls or ceiling, and singing or speaking in such a room is avoided, where possible, by artists or orators. On the other hand, a "live" room possesses good reflection characteristics. These two characteristics may exist in only a particular section of the room, and in various degrees of intensity. Thus, there are gradations between a very live room and a very dead one, depending on its "reverberation" characteristics, that is, upon its ability to reflect sound waves.

Thus, reverberation may be defined as the persistence of sound due to repeated reflections. As mentioned previously, sound energy gradually fades away at a rate depending on the surrounding media. This has been termed "the rate of decay" of sound energy, and may be defined as the time rate at which the sound energy is decreasing at a given point and at a given time. The practical unit is the decibel per second.

Reverberation Time. To enable comparisons of the characteristics of different rooms and types of sound insulation, etc., to be made, a unit called "reverberation time" has been adopted.

This unit may be defined as follows: The "reverberation time" for a given frequency is the time required for the average sound energy, initially at a steady state, to decrease to one-millionth (that is, 60 db) of its initial value after the source is stopped. The time is expressed in seconds.

/ Absorption

Absorption Coefficient. There is an ideal time lag (reverberation time) and volume proportion between the original and reflected sound striking the microphone. If this is achieved, the resultant sound will be natural in quality, neither too dead nor too alive.

The quantity relation between the original and reflected sound depends, among other things, upon the efficiency of the reflecting surface. Different materials possess reflecting properties in varying degrees, dependent upon their structure and surface. The value of the absorption coefficient increases with increasing frequency of the sound waves. The ideal number of fixed absorption units present in a studio or hall depends upon a number of performers present and the instruments involved. These characteristics will now be dealt with in a little greater detail.

1.14 Requirements for Good Acoustics. In order that a room may have good acoustics, it is necessary that -

(i) The sound be sufficiently loud in the room.
(ii) The room be free from noise whether of internal or external origin.
(iii) The room be free from echoes or interfering reflections.
(iv) The reflecting boundaries, that is, the walls, floor and ceiling of the room be so disposed as to provide a nearly uniform distribution of sound energy through-out the room.
(v) The room be free from undesirable resonance.
(vi) The reverberation in the room be sufficiently reduced to avoid excessive overlapping or commingling of excessive sounds of speech or music, but that the reverberation be sufficient at all frequencies to give a pleasing effect to either speech or music as judged by the average listener.

It will be seen from the above that the requirements for good acoustics are many, and no attempt will be made here to examine all of them. Instead, reverberation and noise will be considered.

The time required for the intensity of the sound to be reduced a specified amount will depend upon -

(a) The number of reflections which occur per unit time, and
(b) The amount of sound energy which is absorbed at each reflection.

If the room is a large one, there will be only a few reflections per second and, in addition, if but a little sound energy is absorbed at each reflection, it will require a relatively long time for the intensity of ordinary sound to be reduced appreciably. Such a room will be excessively reverberant. On the other hand, if the room is small and the boundaries highly absorptive, the room will be free from reverberation. It will be seen, therefore, that the reverberation of a room can be controlled by its dimensions and by the material used at the boundaries.

REHEARSAL FOR "RADIO-REEL" FROM RADIO-AUSTRALIA'S STUDIO.

(Note microphones, acoustical treatment of room and the producer of the programme.)

The reverberation of a room is not the same for all frequencies, and it is possible by the correct selection of acoustic material to control the reverberation to conform to set requirements.

/ This

This is necessary, as it has been determined that it is desirable to have the lower and higher frequencies more reverberant than the middle frequencies centred about 500 c/s. Reverberation time is illustrated in the form of graphs as shown in Fig. 7.

These graphs show the desirable reverberation time at different frequencies for speech and music in broadcast studios.

For the acoustic treatment of studios, sound absorptive materials are used, generally on the walls, and, of the many types available, "rock-wall" behind perforated metal sheet, such as zinc, is used extensively. Recently, the zinc sheet has been displaced by perforated ply wood. The latter has the advantage of a better appearance. "Rock-wool", as produced by one manufacturer, is basalt rock, which, when in the molten state is forced through fine nozzles to give a thread-like result.

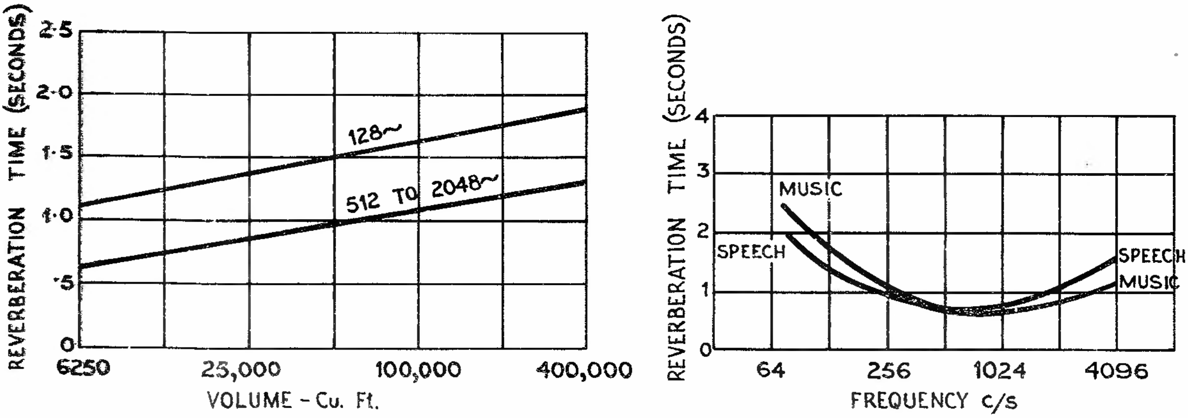


FIG. 7. REVERBERATION TIME FOR STUDIOS.

To achieve some of the other requirements of good acoustics, it is usual to resort to such devices as avoiding the use of panel walls and not having the ceiling parallel with the floor. In addition, the walls and ceiling may have curved sections in the convex manner.

Considering the question of noise in studios, the noise from external sources is of major importance, and the principal means whereby noise enters a studio are as follows -

(a) By means of openings, as windows, cracks around doors, ventilating ducts, or any other opening which will admit a free flow of air.
(b) By means of the refraction or transmission through partitions. This is analogous to the refraction and transmission of light from air to water, or between any two media.
(c) By means of the conduction of sound through solids. For example, "impact sounds", such as footfalls, hammering on walls or floors, or the moving of furniture on hardwood floors, are conducted through the dense and rigid structural members of a building.
(d) By means of the diaphragm action of walls, which communicate sound from one side of a partition to the other side.

Of the above, the refraction or transmission of sound from one media to another can be ignored, as the effect is negligible.

The following methods are used to reduce noise from external sources -

(a) Special attention to ventilating systems, such as quiet running machinery and special ducts.
(b) Build the studio so that it is a complete room in itself and resting on pads of rubber, cork or felt. This will reduce the conduction of sound through solids.
(c) Use a sound lock at the studio door.
(d) Use special windows.
(e) Use materials of high absorptivity for the external walls.
(f) Avoid the use of pipes between the studio and adjacent structures. This refers to heating pipes and conduits for power wiring.

/ 2.

2. MICROPHONES.

2.1 A microphone is a device which changes acoustical or sound energy into electrical energy, at the same time preserving all the frequency characteristics of the sound energy. If the changes in the electrical circuit follow faithfully the sound impulses, the microphone is said to have perfect fidelity.

Microphones must be able to fulfil the following general conditions -

(i) Reproduce with perfect fidelity the sound generated by a symphonic orchestra, which contains practically all the frequencies likely to be encountered in music. This range is approximately from 30 to 16,000 c/s.

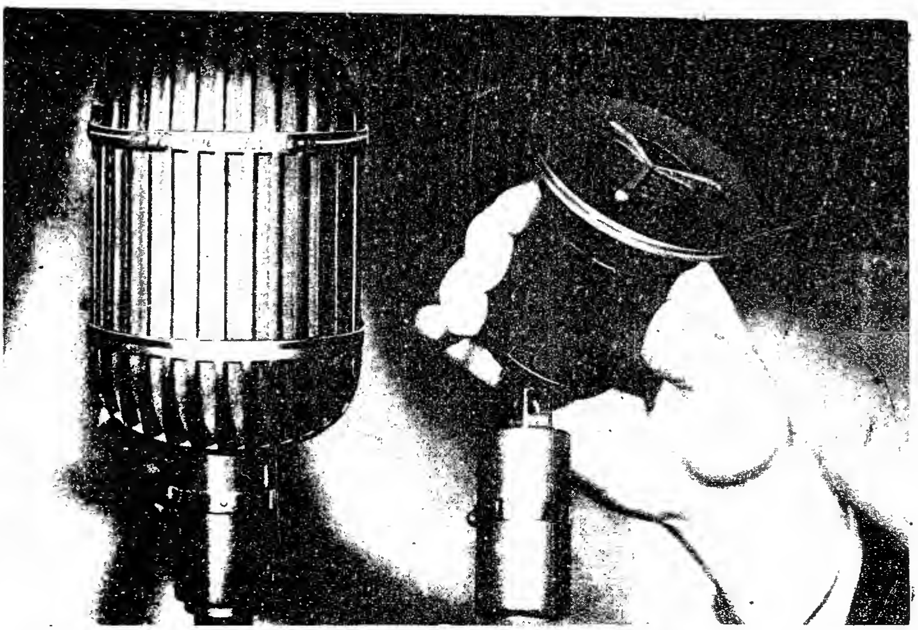
(ii) Frequencies not present at the source should not be added by the microphone (that is, it should not introduce distortion).

(iii) The microphone must be capable of reproducing great variations in sound pressure without distortion. The variation in volume, which the microphone is required to handle, may be approximately 60 db.

(iv) Any noise, such as hiss, generated by the microphone should be at least 40 db below the weakest signal encountered.

(v) The microphone should possess an impedance that is readily matched to a connecting cable and/or amplifier.

(vi) The microphone should possess high sensitivity, that is, the power output into its electrical load must be as great as possible for a given sound intensity.



MODERN MICROPHONES.

CARDIOID MICROPHONE (LEFT) AND A DYNAMIC MICROPHONE.

2.2 Types of Microphones. Microphones naturally fall into types descriptive of their basic electrical design. The most generally used are -

(i) Carbon microphone.
(ii) Moving coil microphone.
(iii) Velocity or ribbon type.
(iv) Cardioid types.
(v) Piezo-electric or crystal types.
(vi) Condenser type.

Although types (ii) to (iv) inclusive are generally used in broadcasting, the other three types will be discussed also since they are frequently used.

2.3 Carbon Microphone. The Carbon Microphone is the simplest type of microphone and is the basic "transmitter" of telephone systems (see Fig. 8). The carbon microphone depends for its operation primarily upon the variation in resistance of carbon granules in contact. The resistance of the carbon varies directly with the pressure exerted upon it. The granules are behind and in contact with a diaphragm, which vibrates according to the air pressure variations resulting from sound waves. The resistance of the carbon granules varies accordingly and controls the magnitude of a current passing through the carbon. The output of a typical carbon microphone is not of a very high quality, due to mechanical resonances which can occur; it is also liable to "pack" and cease functioning.

Most carbon microphones have a relatively high output but generate a steady hiss, owing to minute variations in the contact resistance of the carbon granules themselves. Some carbon microphones are also sensitive to vibration and cannot readily be handled in use.

Some improvement in characteristics was obtained by using two carbon granule chambers with a tightly stretched diaphragm between them. This type reduced the mechanical resonances and also the output, which was, however, somewhat improved by the push-pull action. Fig. 8a is a cross-section of a typical Double-Button type microphone, and Fig. 8b is the associated circuit. As mentioned before, this type is seldom used, but the description is included for its historic developmental interest. The output of this microphone averages about -45 db.

Fig. 8 includes a modern version of this type of microphone as developed for services, such as police, ambulance, forest protection.

2.4 Moving Coil Microphone (Magnetic). The moving coil or dynamic microphone (see Fig. 9) consists of a powerful cobalt-steel magnet, which provides a strong magnetic field across a narrow annular air gap. In this gap, with a clearance of a few thousandths of an inch, is a circular coil of aluminium ribbon wound on edge. This coil is attached to a light but stiff duralumin diaphragm, so designed as to vibrate as a whole up to a high frequency. The diaphragm has a dome-shaped centre portion, which extends to the inner edge of the moving coil. This stiffens the centre of the diaphragm, so that it vibrates substantially as a plunger throughout the frequency range. This type of microphone is a pressure operated device.

The electrical output of the moving coil microphone is low, and is approximately -60 db (reference 1 milliwatt) on average speech.

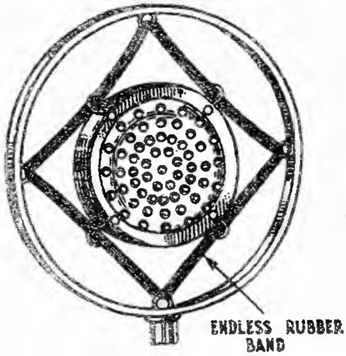
This microphone is of low impedance, 50 ohms, and must be used with a pre-amplifier, except in connection with outside broadcast amplifiers.

Generally speaking, the frequency response is uniform from 35 to 10,000 c/s. However, this is determined by the angle of incidence of the sound in relation to the microphone. A typical response curve is given in Fig. 9a.

Fig. 9b shows the essential parts of the moving coil microphone. Acoustic compensating devices are used to give a satisfactory response. Fig. 9 includes some typical designs of this type of microphone.

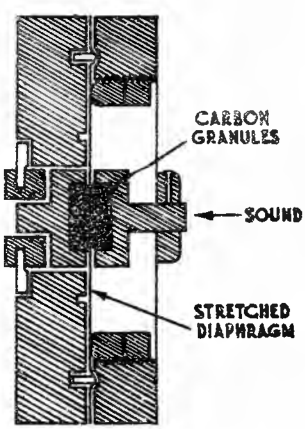
As the coil moves with the diaphragm in proportional response to the sound waves, it cuts the magnetic lines of force between the poles of the magnet. Thus, there is induced into the coil a current which, in frequency and amplitude, is a faithful replica of the diaphragm movement. The current produced in the coil is the electrical output of the microphone.

/ Fig. 8.

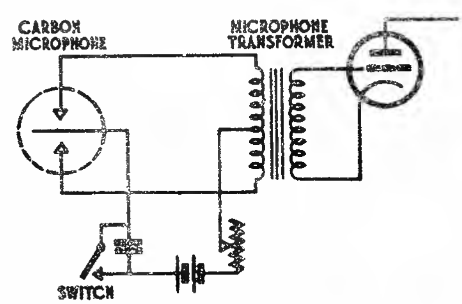


EARLY TYPE

MODERN TYPE



(a) CROSS SECTION

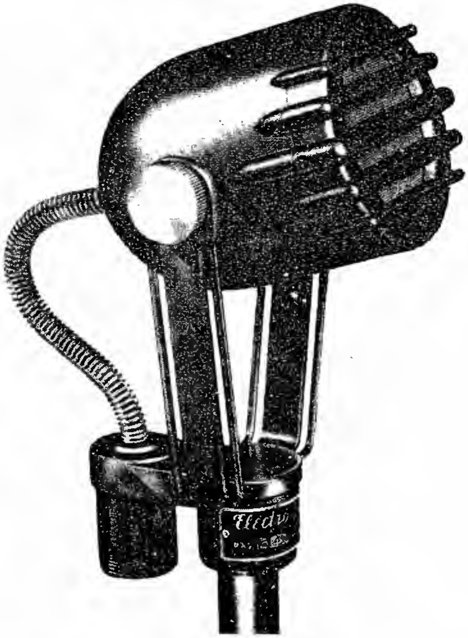


(b) ASSOCIATED CIRCUIT

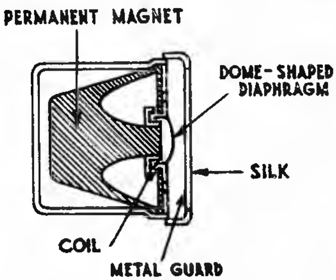
CARBON MICROPHONE.

FIG. 8.

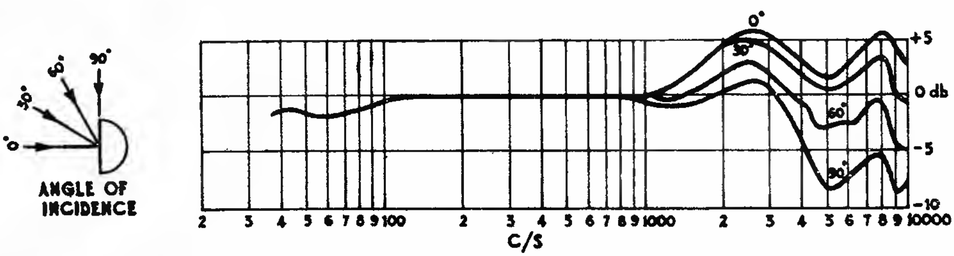
EARLY TYPE



MODERN TYPE



(b) ESSENTIAL FEATURES



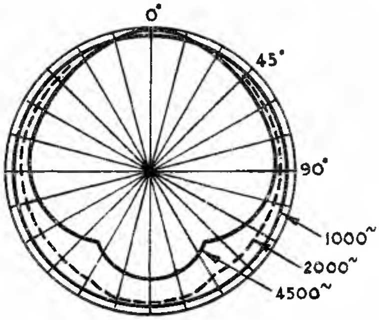
(a) TYPICAL FREQUENCY RESPONSE

DYNAMIC (MOVING COIL) MICROPHONE.

FIG. 9.

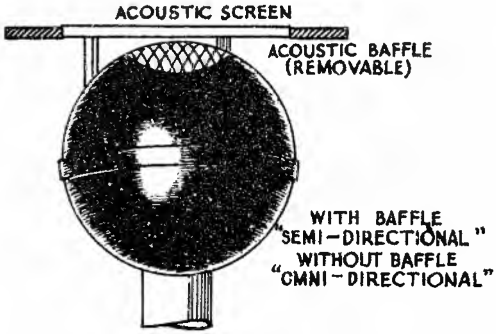
Moving coil microphones are considered to be of the semi-directional type, and a typical graph of the direction characteristics is given in Fig. 10. However, if moving coil microphones are used in the horizontal plane, they become non-directional in the horizontal plane.

Another form of moving coil microphone, sometimes known as the "Billiard Ball" type, is of the same basic design, but is smaller and is of spherical shape. This microphone has the advantage that it can be used without being too conspicuous, and is the type normally used on announcers' desks. An outline of this type is shown in Fig. 11.



DIRECTIONAL CHARACTERISTICS
MOVING COIL MICROPHONE.
"SEMI-DIRECTIONAL".

FIG. 10.



MOVING COIL MICROPHONE
"BILLIARD BALL" OUTLINE.

FIG. 11.

Moving coil microphones are robust, and one type is frequently used for outside broadcast work, chiefly for commentaries.

Velocity or Ribbon Type Microphones. The ribbon microphone (see Fig. 12) is a special type of moving coil microphone in which the moving coil consists of a flat piece of aluminium alloy, which is acted upon directly by the sound waves and which has a resonant frequency below the audible range. The construction of such a microphone is shown in Fig. 12a.

The ribbon is suspended between the pole pieces of the permanent magnet in such a way that the magnetic field cuts the edge of the ribbon. By allowing very small clearances, the ribbon is free to move back and forth between the pole pieces without touching them. The ribbon is actuated by sound waves moving past it at right angles to its broad surface. Such sounds form a pressure area on one side of the ribbon base and a partial vacuum on the other side, the ribbon being forced into the vacuum. When a plane wave passes by a ribbon microphone, the resulting force acting on the ribbon is proportional to the difference in sound pressure on the front and back of the ribbon. The resulting force exerted against the ribbon is proportional to the frequency and to the pressure gradient or particle velocity of the sound wave. As a result, this type of microphone is commonly called a "velocity" microphone.

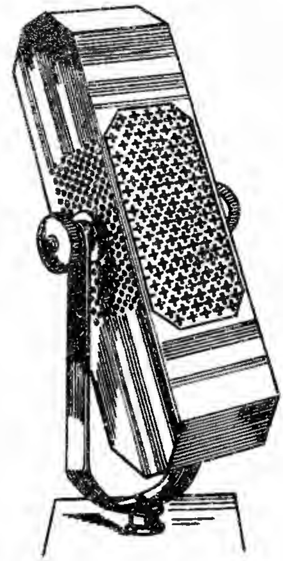
The velocity microphone has a directional characteristic known as bi-directional, as shown in Fig. 12b. This is because sound waves arriving from the side strike both the front and back of the ribbon at the same instant, and so produce no resultant force. This pattern can be used to advantage to minimise undesired reverberation and noises. This microphone is particularly useful for interviews and play production.

/ The

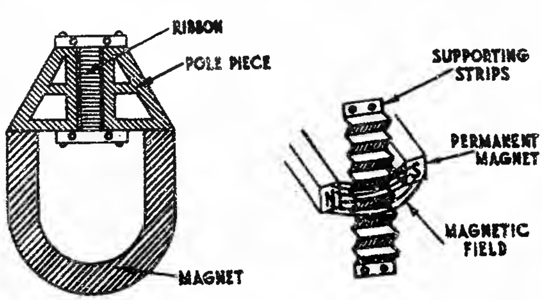
The velocity microphone is not as robust as the moving coil types, and an accentuation of the lower frequencies will occur if the microphone is too close to the broadcasting artist. It also needs protection from sudden winds and draughts.

Due to the fact that the impedance of the ribbon is only a fraction of an ohm, a coupling transformer is fitted as an integral part of the complete microphone. The secondary impedance for departmental purposes is 50 ohms. The electrical output of the velocity microphone is lower then for the moving coil types, averaging about -100 db.

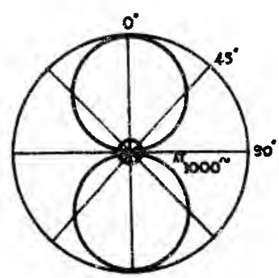
MODERN TYPE



EARLY TYPE



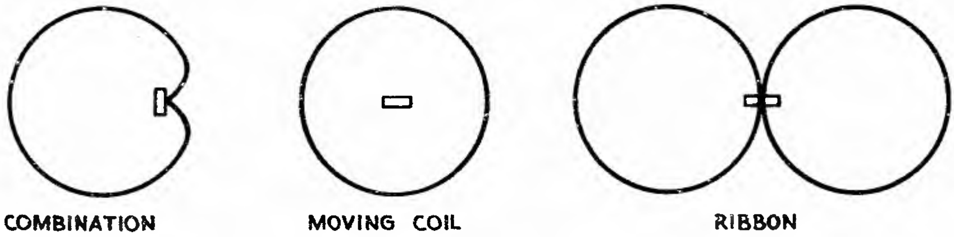
(a) CONSTRUCTION



(b) DIRECTIONAL CHARACTERISTICS

FIG. 12. VELOCITY OR RIBBON MICROPHONES.

Cardioid Microphone. Basically, the Cardioid Microphone consists of a moving coil and a ribbon type microphone in a common housing, so that, by means of a switch, any one of three directional characteristics can be obtained. These characteristics are shown in Fig. 13.



UNIDRECTIONAL. NON-DIRECTIONAL. BI-DIRECTIONAL.

FIG. 13. THREE POSITIONS OF A CARDIOID MICROPHONE.

In the first and second positions, either of the two microphones can be used alone. In the third position, the two microphones are electrically connected and the directional characteristic is of heart-like shape, hence the word "cardioid".

With the moving coil type of microphone, the diaphragm always moves in the same direction irrespective of the direction of the sound. This applies even if the sound is from the rear of the microphone. However, with a ribbon microphone, the ribbon reverses direction if the sound is from the rear instead of the front and, by combining the two microphones electrically together, the directional characteristic is cardioid in shape.

2.5 Piezo-Electric or Crystal Microphone. A crystal microphone is one in which the electrical output results from the stress of a crystal having piezo-electric properties. Such a crystal, when properly cut, if subject to mechanical stress, shows charges of opposite polarity upon its faces. The magnitude of the charges is dependent upon the displacement of the piezo-electric body. Rochelle salts have shown a marked superiority over other crystals and are used extensively for this type of microphone. Although these salts are soluble in solution, suitable protection can be obtained against moisture by enclosing the crystal in waterproof waxes and paper. The construction of a typical crystal microphone, using a number of crystal cells arranged to aid each other and improve the over-all sensitivity, is shown in Fig. 14.

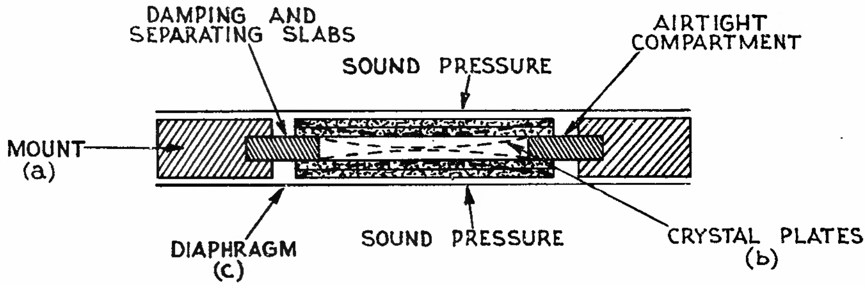
This Crystal Microphone consists of (a) a rectangular frame of bakelite in each side of which is supported (b) a thin Rochelle salts crystal unit. The space between the crystals and the frame is sealed by (c) a flexible air-tight ring, which permits the crystal to vibrate freely with variations of air pressure due to sound waves. When the entire unit is subjected to sound variations, the two sides of the crystal unit vibrate in phase with each other, which produces a resultant in the generation of an e.m.f. across them in direct proportions to the sound pressure, while, if the unit is subjected to mechanical vibration or shock, the voltages generated by both sides are out of phase with each other. Owing to the fact that this microphone is purely a pressure operated device and therefore non-directive, there is no cut-off effect at the low frequencies, and the response is flat over the audio frequency range.

The output level of typical units varies between -65 and -75 VU approximately.

The ruggedness, uniform frequency response and non-directivity in a horizontal plane make this an excellent microphone for broadcast purposes, although special design is needed to render it insensitive to atmospheric conditions. The horizontally non-directive properties, possessed by no other type, are often an especial convenience in studio set-ups, enabling performers to be comfortably grouped around the microphone instead of being clustered in front of the microphone to avoid distortion due to loss of the higher frequencies.

/ Fig. 14.

(a) Microphone.



(b) Construction.

FIG. 14. TYPICAL CRYSTAL MICROPHONE.

ADVANTAGES OF CRYSTAL MICROPHONE.	DISADVANTAGES OF CRYSTAL MICROPHONE.
Light Weight.	High impedance makes matching difficult with degrading response.
Ruggedness.	Unstable with temperature and humidity unless specially designed.
Ease of maintenance.	
No battery required.	
Non-directional.	
Not acoustically overloadable.	
No inherent background noise.	

2.6 Condenser Microphone. A condenser microphone depends for its operation upon variations of capacitance. Although not used often, this type is of interest in illustrating the many designs that have been tried in an effort to produce a microphone of high quality. In the condenser microphone, the diaphragm acts as one plate of a condenser. The diaphragm, by its movement, alters the charge and, in so doing, produces a voltage drop across a suitable resistor in series with it and the charging source. The voltage thus produced will be a true replica of the diaphragm movement in amplitude and frequency. Because of the very small movement of the diaphragm under ordinary sound pressures and the low capacity of the unit, very high values of resistance and charging voltage must be employed to obtain a reasonable output.

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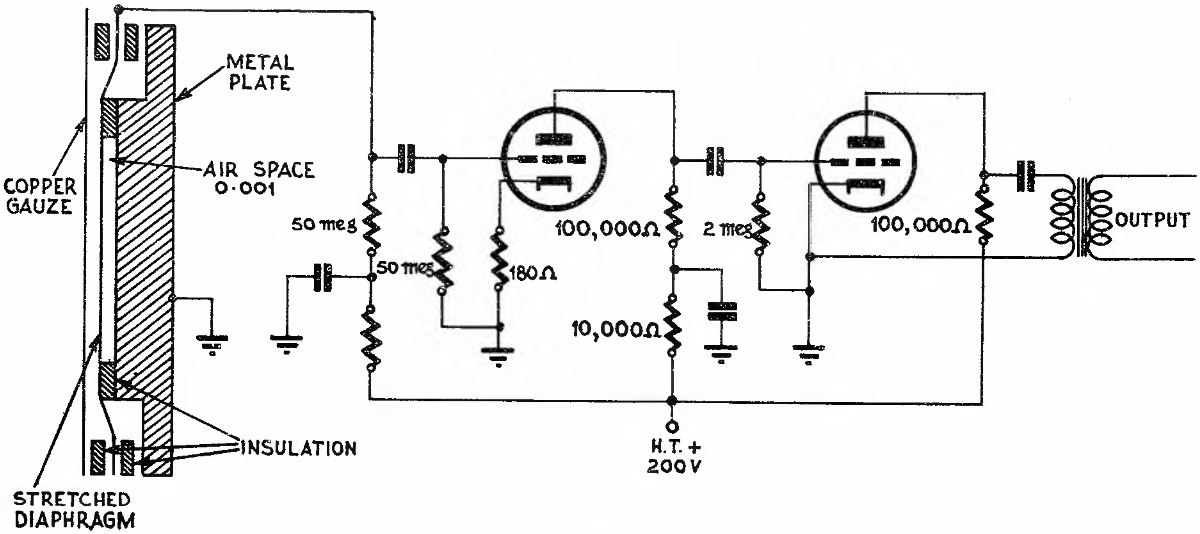
The high impedance and low output of the unit make it impossible to use this microphone with long leads, and a two or three stage amplifier is required. This amplifier is usually fitted with the microphone itself, the two becoming a self-contained unit called a condenser microphone.

The condenser microphone has had wide application because of its reasonably good frequency response, freedom from background noise and low transient distortion.

The disadvantage of the condenser microphone are lack of portability, comparative insensitiveness, the necessity for an amplifier in close proximity, its susceptibility to humidity and dampness effects and lack of ruggedness.

Thus, the condenser microphone has gradually been superseded by other types possessing greater sensitivity, ruggedness and other desirable qualities.

Fig. 15 shows a typical unit, the microphone output being approximately -95 db.



CONDENSER MICROPHONE UNIT.

FIG. 15.

2.7 Microphone Placement. The effective acoustics for any particular studio are very much influenced by the type of microphone used and its placement.

In determining the proper placement of a microphone, it is necessary to take into consideration the directional properties of the particular microphone being used as regards both electrical output and frequency response characteristics. Each type of microphone has a directional characteristic peculiar to its type, and both the horizontal and the vertical plane directional pattern of the microphone must be considered in connection with its application and placement. Some microphones are unidirectional, other bidirectional, and still others non-directional in the horizontal plane. These properties are extremely important and useful for discriminating against undesired sources of sound, and for obtaining a desired relation between sounds from different sources.

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It is evident then that a simple way of instantly changing effective studio acoustics is by using microphones having different directional characteristics. The use of different microphones, plus the use of absorbent and reflecting materials, enables an adjustment of studio acoustics to be made over a reasonable range of conditions.

The use of more than one microphone for the pick-up of a given performer or groups of performers working as a unit (orchestra, chorus, etc.) is generally avoided as far as is practicable. Serious frequency and delay distortion is likely to result if more than one instrument is employed for the pick-up, because, under these circumstances, each microphone will be a different distance away from the sound source. As a result, the sound waves do not reach each microphone at the same instant.

Not only does the type of microphone determine the results obtained, but also the distance of the microphone from the artist and the acoustics of the particular studio in use. If the microphone is close to the artist, the reverberation will not be great, but, as the distance is increased, so the reverberation is increased.

In the case of the studio, particularly the type with a "dead end" and a "live end", it is possible by both the placement of the microphone and the position of the artist to obtain varying results. Likewise, in the case of orchestra and chorus type of programmes, the actual arrangement of the musical instruments or the artists has a large bearing on the final result.

It can be stated then that the results obtained from the output of a studio are determined by the following factors -

(i) The directional characteristics of the microphone used.
(ii) The distance between the performer and the microphone.
(iii) The effective acoustics of the studio, particularly reverberation time.
(iv) The arrangement and position of the performers.

Even with the above aids, compromises must be made in setting up artists before the microphone. Instruments having different frequency ranges will have differing reflection characteristics, and also possibly different sound energy output, for example, a double bass and a flute will differ greatly.

Sufficient has been said in the above notes to indicate that studio design, microphone placement and acoustic treatment form quite a branch of radio engineering on their own. The importance of this has been well realised by the introduction of production engineers and technical courses dealing with the subject.

3. LOUD-SPEAKERS OR REPRODUCERS.

3.1 A loud-speaker is a device designed to radiate acoustic energy into a room or open air. The device is furnished with electrical energy, and the transformation to acoustic energy is usually accomplished by causing a surface or diaphragm in contact with air to move, thus setting the adjacent air in motion.

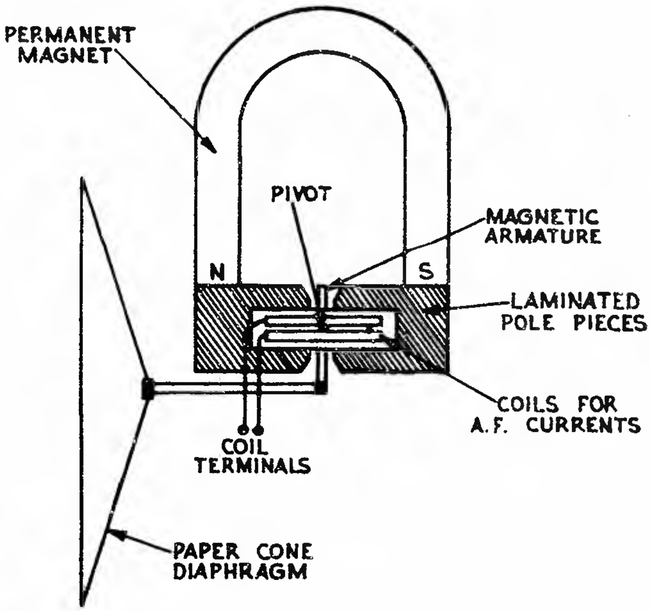
3.2 Diaphragms. In order to radiate large amounts of power into the open air, considerable motion of the diaphragm is required. This requires that the diaphragms be supported at their large ends by flexible "surrounds" or annular members, which have mechanical resistance as well as some stiffness. In large diaphragm speakers, the driving coil or rod must also be supported by a flexible centring member or "spider". This is usually made of a material such as bakelite, fibre or paper, which adds almost a pure stiffness to the system.

3.3 Types of Loud-Speakers. (i) Moving-Iron Type. Early types of loud-speakers were of the "moving iron" type, in which the diaphragm was attached to a small, light soft iron armature suspended in the field of a permanent magnet. The flux of the permanent magnet was caused to vary by the electrical currents from the driving source, and thus the armature vibrated in accordance with these variations and transmitted them to the diaphragm and the air.

/ These

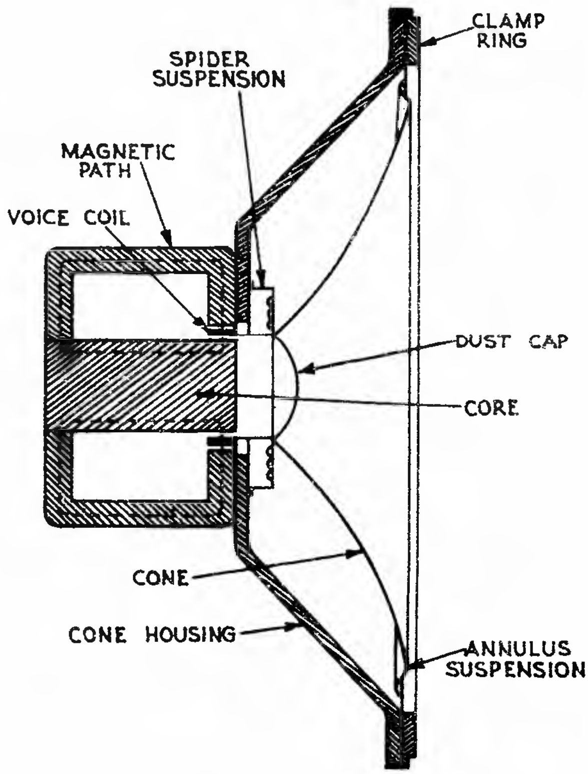
These speakers usually gave poor reproduction of low frequencies and possessed a high impedance varying with frequency, which made correct impedance matching impossible. However, their performance compares favourably with cheap speakers of the moving-coil type, and they still have a range of usefulness. Fig. 16 shows the usual form of this speaker. Currents through audio current coil cause variations in the magnetic flux in the field of the magnet N-S. These variations cause the magnetic armature to vibrate, thus transferring the vibrations through the driving rod and pin to the cone.

(ii) Moving-Coil Type. The loud-speakers in most general use are of the moving-coil or dynamic type, as shown in Fig. 17. This type consists of a paper cone, to the apex of which is fastened a coil (commonly called the "voice coil") located in a strong magnetic field and carrying the audio-frequency currents to be transformed into sound waves. In such an arrangement, the action of the magnetic field on the coil current produces a mechanical force that vibrates the paper cone and causes radiation of sound waves. The cone is supported in a metal frame around its outer edge, while the coil is held in position and supported by means of a flexible spider or a corrugated disc where dust proofing is essential. The entire coil is, therefore, free to move as a unit and, under ordinary conditions, is proportional in such a manner as to have a resonant frequency at the lower end of the frequency range to be produced.



MOVING-IRON SPEAKER.

FIG. 16.



MOVING-COIL LOUD-SPEAKER.

FIG. 17.

Fig. 18 shows the impedance characteristics of a typical moving-coil loud-speaker. The cone must be mounted in a baffle or in a box, in order to prevent the radiation from the front and back sides from cancelling at low frequencies.

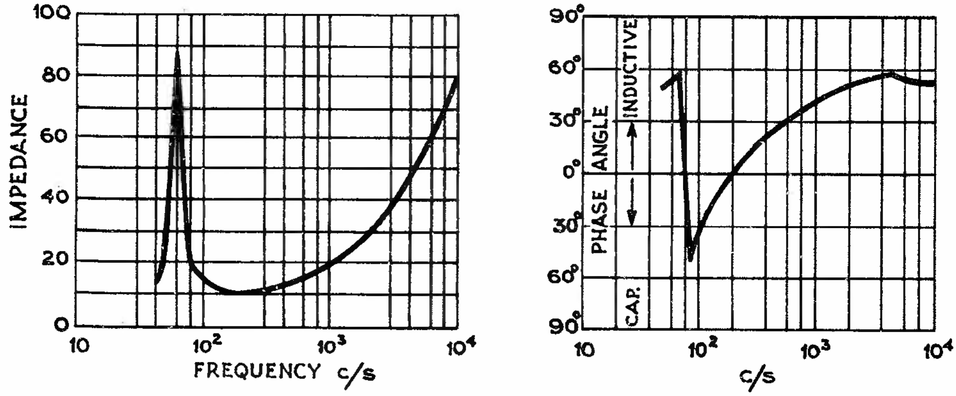
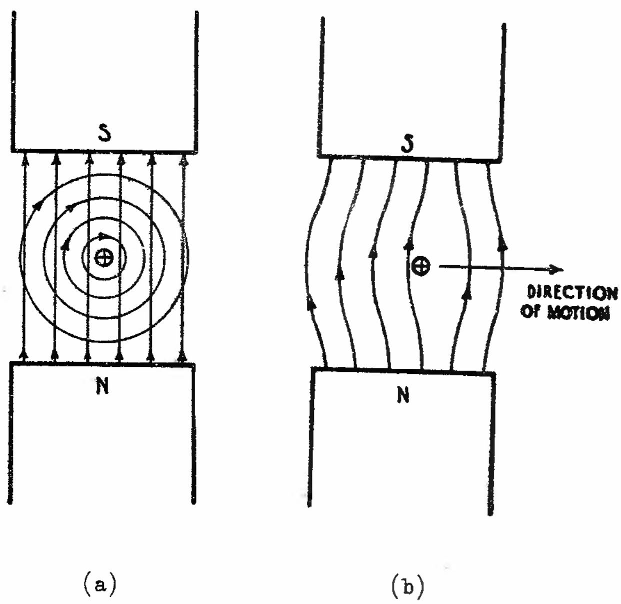


FIG. 18. MOVING-COIL LOUD-SPEAKER IMPEDANCE.

Operation. The operation of a moving-coil loud-speaker can be understood by considering the basic theory of its operation. Assume a straight conductor carrying current is placed between two large plane pole faces of a permanent magnet. The component fluxes are as shown in Fig. 19a. The flux due to the magnet is uniform in the gap between the poles, as shown by the equally-spaced parallel straight lines. The conductor lies in a plane parallel to the pole faces, and current is flowing through it into the paper. The lines of flux are concentric circles.



ILLUSTRATING THE DISTRIBUTION OF FLUX LINES IN THE MAGNETIC FIELD OF A MOVING-COIL SPEAKER.

FIG. 19.

The resultant distribution of flux lines is shown in Fig. 19b. On one side of the conductor, the component flux lines run in the same direction, and so the field is strengthened as shown by the packing of lines. The resultant field is weak on the other side, and the lines are comparatively far apart. This condition forces the conductor to the right in a plane parallel to the pole faces.

If the direction of the current in the conductor is reversed, it will be seen that the conductor will move in the opposite direction, that is, to the left. The force obtained in a moving coil is given by -

$$\frac{\beta \text{ I L}}{10} \text{ dynes}$$

where β = the flux density due to the magnet in lines per square centimetre,
I = the current in amperes flowing in the conductor, and
L = the length in centimetres of the conductor which is the region of uniform flux density β.

It will be apparent that, as the conductor moves in the magnetic field, it is cutting lines of force and will then follow Lenz's Law -

> "The direction of the induced e.m.f. produced by the motion of a conductor in a magnetic field is such that, if induced current could flow, it would produce a force opposing the motion."

This law is important as it explains the behaviour of the cone in respect to transient response, for, if a large induced current could flow, a large force would oppose the motion and the cone would lose its momentum quickly when the voice currents cease to flow.

The permanent magnet type of loud-speaker is preferred for two reasons -

(i) Power is not required to excite the field coil.
(ii) Greater efficiency is obtained. Field excitation cannot give the same flux density in the gap as a modern permanent magnet.

3.4 Loud-Speaker Damping. A loud-speaker possesses mass, and, therefore, inertia. (The effect of inertia is to tend to retard the stopping or starting of a mass.) In a loud-speaker vibrating under constantly changing frequency and amplitude conditions, the effect of inertia is to tend to prolong the cone excursions of, say, a certain frequency. As a result, transients may not be reproduced effectively with a consequent loss of quality of reproduction. This effect is minimised by "damping"

Every loud-speaker has a certain amount of internal damping, due to its construction, but in most cases, the damping is insufficient to give good reproduction of transients. The anode resistance of the output valve, as reflected through the coupling transformer, acts as an additional shunt damping resistance. Since the audio output voltage and the damping resistance are both passed through the same transformer, the effect may be considered as on the primary. Only the ratio of the load resistance (R_L) to the effective anode resistance (R_o) need be considered, therefore, and this ratio (R_L/R_o) is called the "damping factor". Triodes have good damping factors, but pentodes and beam power tetrodes, due to their high plate resistance, have very poor damping factors. The application of negative (voltage) feedback will reduce the anode resistance of a pentode or beam tetrode, and, if sufficient feedback is applied, the damping factor may be made even greater than with triodes, but this is not always desirable.

The damping, at frequencies at which the impedance of the loud-speaker rises above its impedance at 400 c/s, is greater than that at 400 c/s, since R_L is greater. This is particularly beneficial in improving the reproduction at the bass resonant frequency.

The cone produces the sound waves and, at low frequencies, it acts approximately as a piston or diaphragm having a diameter equal to the diameter of the cone. At frequencies above 1,000 c/s, the cone ceases to act as a simple piston and, instead, the cone vibrations are in the form of waves travelling outward from the apex, with the result that the centre part of the cone vibrates much more intensely than the outer edges. By using a hard cone material or corrugating the cone, the high frequency response can be maintained.

At low frequencies, the sound waves have a wide angle of distribution, and it is necessary that the cone be mounted in some form of baffle in order that the waves radiated from the two sides of the cone, which are produced with a phase difference of 180°, will not cancel each other.

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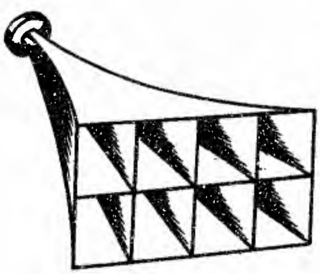
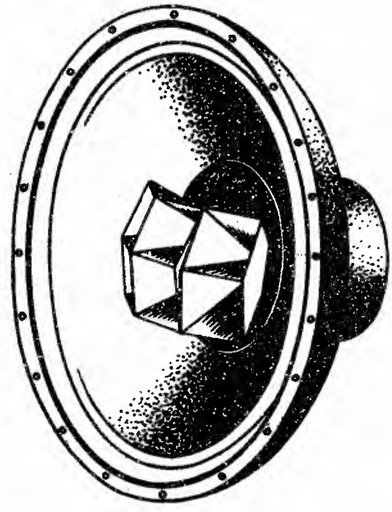
The baffle diameter should be approximately a half wavelength at the lowest frequency for which little or no loss in the sound output is desired, and, for best results, should have an irregular outline, in order to eliminate the possibility of destructive interference between front and back radiation at certain critical frequencies. In the usual radio receiver, the cabinet supplies the baffle and, where good response down to low frequencies is desired, the cabinet must be of ample size.

At high frequencies, the sound waves are projected in the form of a beam, which results in an unequal distribution of sound between the lower and higher frequencies. To reduce this effect, the multi-cellular horn has been developed. When fitted to a moving-coil unit having an aluminium diaphragm somewhat like that of the moving-coil microphone, a satisfactory distribution of the higher frequencies is obtained. For the lower frequencies, a separate moving-coil speaker of the horn loaded type is used. A dividing network is required to divide the upper from the lower frequencies. The network divides the frequencies somewhere in the range 200 to 1,200 c/s, depending on the type and design of the high frequency loudspeaker. Fig. 20a shows a multi-cellular high frequency unit, and Fig. 20b a combination high-quality unit.



(a) Multi-Cellular Horn.



(b) Combination High Quality Unit.

LOUD-SPEAKERS.

FIG. 20.

3.5 Horn Loaded Loud-Speakers. The horn is essentially an acoustic coupling device, which transforms energy at a high pressure and low velocity to energy at a low pressure and high velocity. A horn can, therefore, be used to increase the load that the air produces upon the driving mechanism of a loud-speaker, and thereby makes it possible to obtain a better match between a relatively heavy driving mechanism and the fluid air. It can be said that the horn is an acoustic impedance matching device.

In Fig. 20, a typical horn speaker is shown using a moving-coil driving unit and an exponential multi-cellular horn. Sound waves produced at the throat by the diaphragm vibrates travel along the horn, expanding in an orderly manner until large enough to transfer their energy to space without undue disturbance. In an exponential horn, the taper is such that the cross-sectional area is proportional to an exponent of the distance along the horn.

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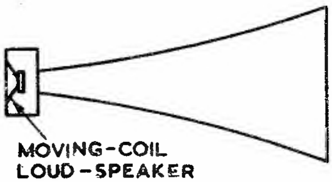
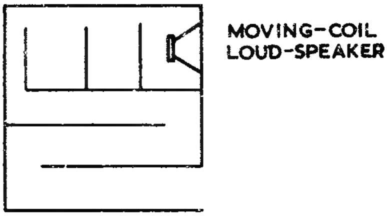
The exponential horn can be used to load the back part of the cone of the ordinary moving-coil loud-speaker as shown in Fig. 21a.



(a) Exponential Horn.



(b) Folded Exponential Horn.

FIG. 21. HORN LOADED MOVING-COIL LOUD-SPEAKERS.

With this arrangement, the lower frequencies up to 250 c/s are radiated by the front of the cone directly to the air, while the radiation from the rear of the cone is made through the horn. The frequencies above 250 c/s are radiated from the front of the cone only. The horn includes an acoustic low pass filter to ensure this condition.

Fig. 21b shows the practical form of this arrangement.

The horn is made up of rectangular sections and built into a cabinet, so that the mouth opening is in the same plane and direction as the front of the cone of the loud-speaker. Thus, the advantages of the horn loading are obtained in a cabinet suitable for broadcasting use.

3.6 Advantages and Disadvantages of the Baffle Mounted Moving-Coil Loud-Speaker and the Horn Type.

Baffle Mounted Type.

Disadvantages.

(i) Low efficiency, 10 to 15 per cent.
(ii) Back radiation.
(iii) Poor frequency response, due chiefly to the resonance of the cone and poor efficiency at the lowest frequencies.
(iv) Poor impedance conditions adversely affecting the output stage of an amplifier.

Horn Loaded Type.

Advantages.

(i) Higher efficiency, 25 to 30 per cent.
(ii) No back radiation.
(iii) Good impedance conditions.
(iv) Improved frequency response.

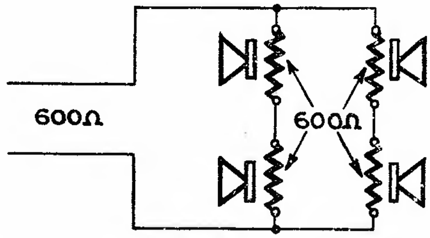
Disadvantages.

Size and shape. If of the folded type, shape is satisfactory. The size depends chiefly on the lowest frequency to be produced. The folded cabinet used for the national broadcasting system has a low frequency cut-off of about 70 c/s.

From the above, it will be seen that the horn loaded loud-speaker is the better, and the arrangement which would give excellent results is a system which employs two speakers, one to handle the lower frequencies and of the horn loaded type, while the other is of the multi-cellular horn type to handle the high frequencies. This arrangement has the added advantage that distortion of the loud-speaker system, due to intermodulation and frequency modulation, is reduced to a very low value.

/ 3.7

3.7 When operating more than one loud speaker from one output; care should be taken not to introduce impedance mismatches. For example, suppose it was desired to connect four 600 ohm speakers to a 600 ohm outlet. If the speakers were all placed in series, the impedance would be 2,400 ohms, and, if they were all in parallel the impedance would be 150 ohms. Both are obvious mismatches, but if placed in a series-parallel combination, an impedance match may be obtained as shown in Fig. 22.



OPERATING FOUR LOUD-SPEAKERS FROM ONE OUTPUT.
FIG. 22.

3.8 Typical types of loud-speaker cabinets are shown in Fig. 23.

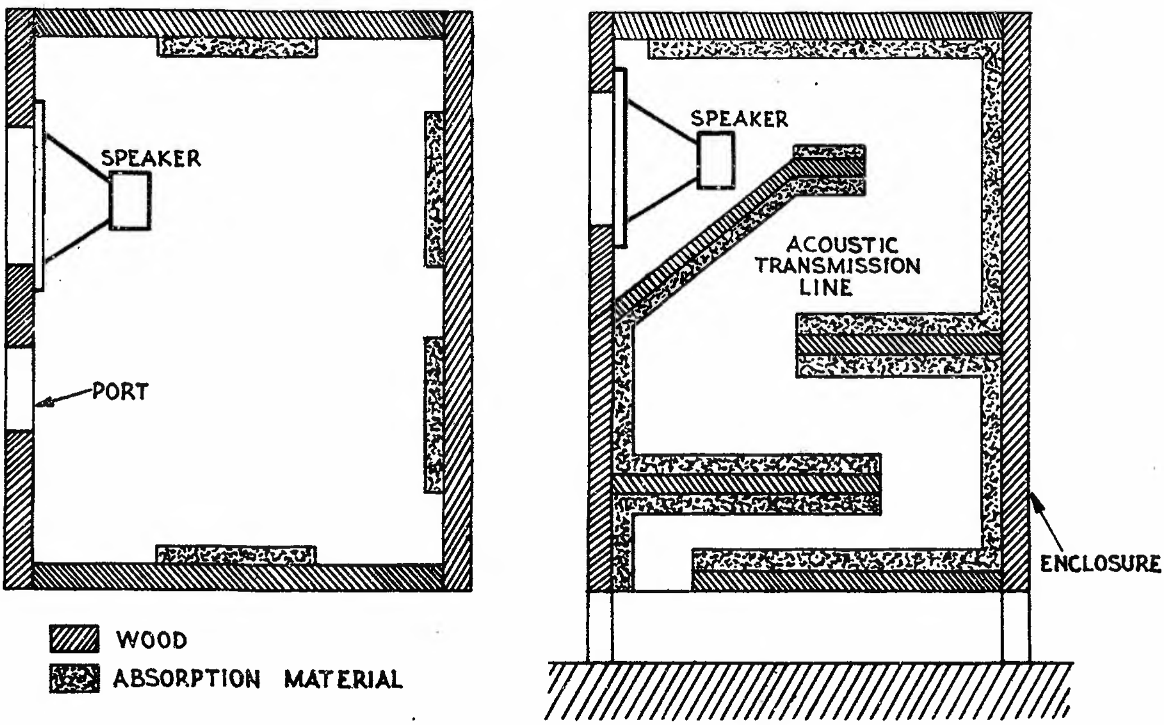


FIG. 23. TYPICAL LOUD-SPEAKER CABINETS.

4. TEST QUESTIONS.

1. (i) Define the terms "Wavelength", "Frequency" and "Reverberation Time"
   (ii) What are the requirements that give a room good acoustic characteristics?

2. (i) Explain the terms "Unidirectional", "Bi-directional", and "Cardioid" as applied to directional microphones.
   (ii) State some advantages of the cardioid type of microphone.

3. Explain the operation of a moving-coil loud-speaker and illustrate with sketches.

4. (i) Name and define three distinguishing characteristics of sound.
   (ii) What would be the main difference between a sound wave of similar frequency originating from a tuning form and from a violin?

5. (i) What is a microphone and what desirable characteristics should one possess for broadcasting use?
   (ii) Sketch and briefly explain the action of a microphone other than a carbon type.

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

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COURSE OF TECHNICAL INSTRUCTION.

RECORDING AND REPRODUCING.
ATTENUATORS AND FADERS.

CONTENTS.

1. INTRODUCTION.

1.1 Sound-recording and reproducing apparatus has many applications in radio. Artists and orchestras may be recorded at one place and heard, by means of these recordings, in many parts of the world. Another useful application is the recording of special features, sporting fixtures and educational talks. Items may be recorded as convenient and broadcast at suitable times; recordings may be sent to other stations in the network or exchanged with organisations in other countries.

Recording facilities are available in several forms, three of these are -

(i) Mechanical recording (sound on disc).

(ii) Magnetic recording (sound on tape or wire).

(iii) Optical recording (sound on film).

Each of these types will be described in this Paper.

1.2 The requirements for good recording and reproduction of sound are that -

(i) The system (from the point where the sound reaches the pick-up device to the point where it is reproduced as sound) must have a linear relationship between its input and its output.

(ii) The system should have a uniform Response versus Frequency Characteristic.

(iii) A linear relationship should exist between its phase shift and the frequency impressed on the system.

2. MECHANICAL RECORDING.

2.1 Mechanical recording is done by cutting a shallow undulating groove in a relatively soft disc or record. This groove is cut by a steel or sapphire stylus actuated by an electromechanical recording instrument.

Record materials are in two broad classes, wax and lacquer. Wax, the softer of the two, is generally used for making "pressed" or commercial disc records, but is not suitable for "instantaneous playback" discs.

The introduction of "instantaneous playback lacquer" coated discs gave great impetus to disc recording for broadcasting. Recordings of this type are usually made on cellulose nitrate lacquer which has been applied to a disc of aluminium or glass. (The term "lacquer" is used here for mixtures having about the same degree of density and firmness, usually with cellulose-nitrate as a basic ingredient plus resins, oils, lacquers, glycerine, paint products and a volatile solvent.) The lacquer is soft but rapidly hardens after being cut; it is hard enough to be played back immediately after cutting, hence the name. These discs do not keep indefinitely without hardening, and must be carefully stored in hermetically sealed tins to prevent drying out before being cut.

STOREHOUSE OF SOUNDS!

Two general methods of disc recording are in use -

(i) Lateral recording.
(ii) Vertical recording.

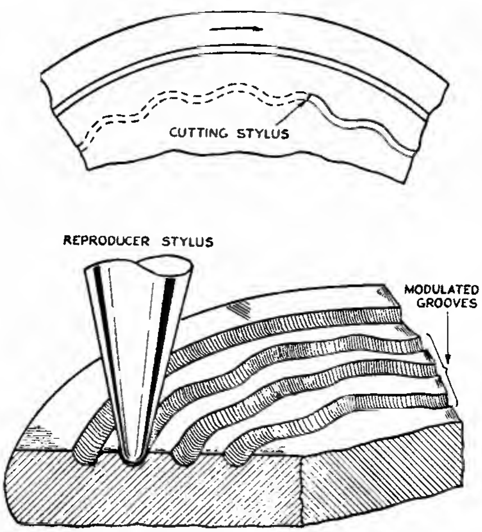
2.2 Lateral Recording. In lateral recording, the frequency and amplitude variations of the sounds to be recorded are translated into horizontal undulations of the cutting stylus, resulting in a groove in the record which varies from side to side about a mean position. See Fig.1 which shows the action of the cutting stylus and a reproducer stylus in a modulated groove.

The more commonly used method of lateral disc recording is known as engraving, and the recording needle or stylus is set almost at right angles to the record surface. The stylus cuts the soft material, just as a machinist's lathe cuts revolving material with a cutting tool.

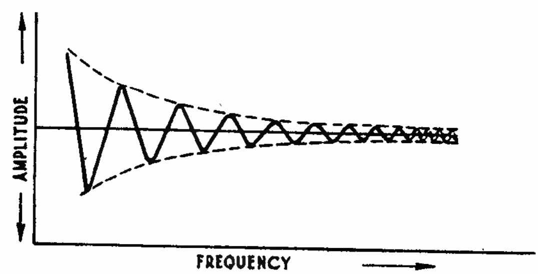
Two terms, Constant Velocity and Constant Amplitude, will be explained as a preliminary to a description of lateral recording.

Constant Velocity. The velocity of lateral recording is the speed of travel of the stylus from side to side. Constant velocity is where the peak lateral velocity of the cutting stylus, which is attained as it crosses the centre of the groove, that is, zero axis, remains constant for all frequencies for constant sound pressure at the recording microphone. As the frequency is increased, the stylus must travel across the groove many more times in unit time, and, if this lateral velocity is to remain constant, the excursion distance each side of the zero axis will diminish. This is shown graphically in Fig. 2a. Thus, the amplitude of the groove is inversely proportional to frequency. If the amplitude of the record groove for the sine wave from an audio frequency oscillator is examined, it will be found that at the lower frequencies a point is reached in the recording where the grooves run into one another. The lower the frequency, the worse the effect. For this reason, constant amplitude instead of constant velocity is used for frequencies below 250 c/s.

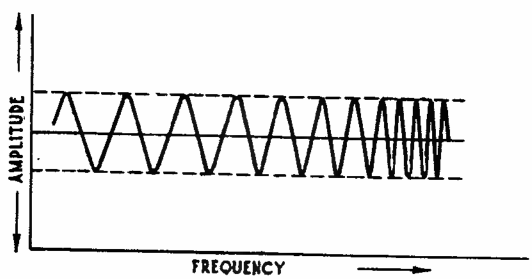
Constant Amplitude. With constant amplitude recording, the amplitude is constant for constant sound pressure at all input frequencies as is shown in Fig. 2b. It will be seen from Fig. 2b that the peak lateral velocity becomes very high at high frequencies.



LATERAL TYPE RECORD GROOVE.
FIG. 1.



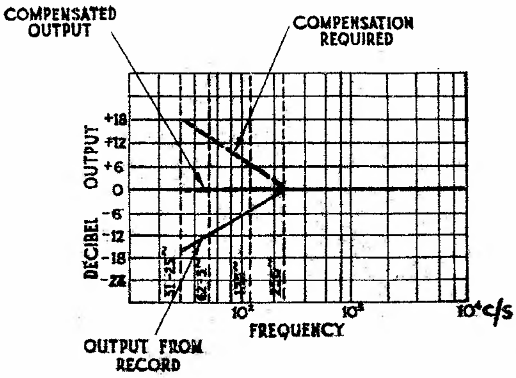
(a) Constant Velocity.



(b) Constant Amplitude.

RECORDING CHARACTERISTICS.
FIG. 2.

Practical Consideration. A cutting head for recording sounds may be either of the magnetic type or crystal type, but the former only will be described since the application of the latter, insofar as cutting is concerned, is restricted. Theoretically, pick-ups and cutting heads are interchangeable, but, practically, this is not always the case. Basically, a magnetic recording head records a frequency response which has a constant velocity characteristic, but, to avoid grooves running into one another, this is changed to constant amplitude below 250 c/s. This involves attenuation of frequencies below this, and the rate of attenuation is 6 db per octave, that is, 6 db at 125 c/s, 12 db at 62.5 c/s, and so on. Actually, 50 c/s is the lowest frequency considered. Due to frequency response limitations of equipment, the highest frequency used is about 6,000 c/s. Fig. 3 shows this point.



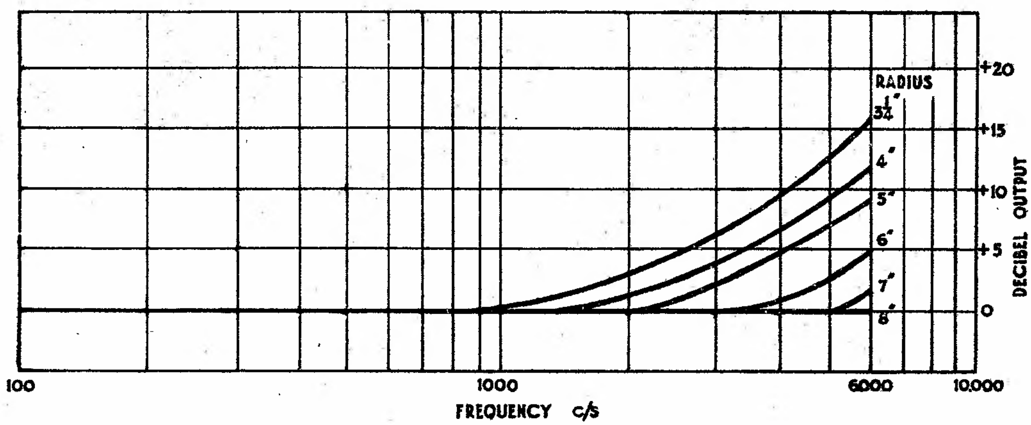
IDEAL RECORDING CHARACTERISTIC.

FIG. 3.

Recording Speeds. The two standard recording speeds are 78 r.p.m. and 33-1/3 r.p.m. The former speed is used for discs up to 12 inches in diameter, while, for the latter speed, 16 inch discs are used. At 78 r.p.m. a maximum playing time of 4 minutes is obtainable, and, at 33-1/3 r.p.m. a maximum playing time of 15 minutes is possible. These times refer to one side of the disc only.

When a disc is recorded at 33-1/3 r.p.m., the velocity of the groove towards the inside is so low that the higher frequencies are attenuated appreciably. To minimise this effect, the minimum diameter of the innermost groove is not less than 6-1/2 inches, and an automatic equaliser is used which accentuates the higher frequencies to the correct amount throughout the recording. This accentuation amounts to 16 db at 5,000 c/s for the innermost grooves and, at the outside, there is no accentuation. This equalising is possible, as the maximum power in speech and music is present at about 400 c/s, but is down about 20 db at 5,000 c/s.

Fig. 4 shows the equalising applied at several radial distances.



AUTOMATIC EQUALISER-FREQUENCY RESPONSE.

FIG. 4.

Signal/Noise Ratio. An important factor in recording system design is its signal-to-noise ratio, which may be defined as the ratio of the loudest tone the system can reproduce without objectionable non-linear distortion to the background noise of the system. It is in many ways the most difficult characteristic of design considerations, and its importance may be deduced from the continual investigations that are being carried out to improve this ratio.

An associated characteristic is the "dynamic range" of a system, which may be considered as the ratio of the loudest sound which can be reproduced (for a given distortion) to the sound which is just distinguishable over the background noise of the system. The dynamic range is also dependent upon the character of the sound and the band-width of the system.

The dynamic range of an orchestra may be of the order of 70 db. This is too great a range to be handled generally and modifications such as "compression" of loud passages, and "expansion" of soft passages are used to reduce the range. This can be done without seriously impairing the reproduction.

A large percentage of the surface noise from a record is of a high-frequency nature, and some systems of recording accentuate the higher frequencies during the recording and attenuate them during replay to improve the signal to noise ratio. For example, if the higher frequencies are accentuated 10 db, the noise will remain at the same level. During replay, the high frequency end, including the surface noise, is attenuated 10 db.

Recording (or Cutter) Head. The audio frequency voltages of the item to be recorded are connected to the recording head. These voltages operate through an electrical circuit which controls the movement of a special cutting stylus which is turn causes a groove to be cut in a prepared disc. The electrical circuit may be one of three types -

Crystal or piezo-electric.

Moving coil.

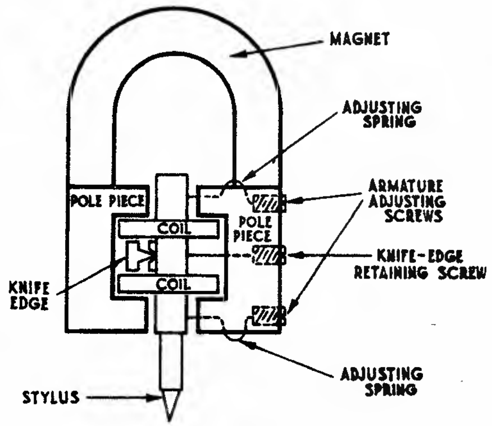
Moving iron.

Some of the requirements of a cutter head are conflicting. For example, it should be small but robust, and reasonably independent of changes of temperature and ageing; it should possess good sensitivity and yet accommodate changes of load on the cutter point that occur with different disc materials and different cutting speeds. No one type meets all the requirements satisfactorily, but compromises enable reasonable performance to be obtained.

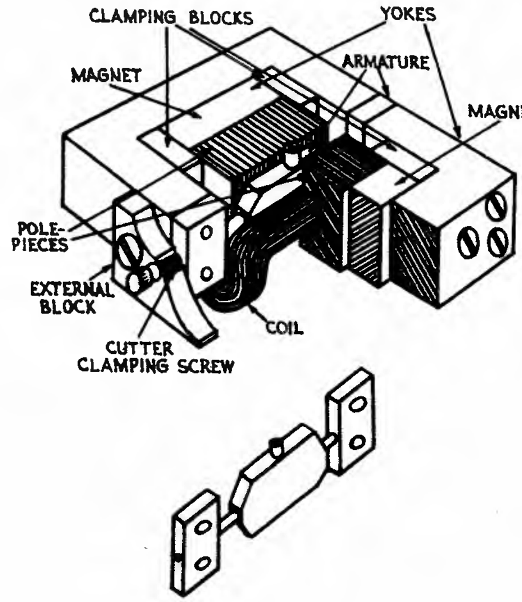
Crystal Type. The Crystal Type has some advantages but is seldom used because it is not robust.

Moving Coil Type. The Moving Coil Type has good frequency response, stability and low distortion, but does not readily lend itself to small compact design. A useful design employing a balanced armature is shown in Fig. 5a. In this type the "constant amplitude" characteristic is obtained in the head itself; the high-frequency response is controlled by a synthetic rubber-like material known as "viscaloid." There are three adjusting springs; of these the centre spring (see Fig. 5a) is the retaining spring and the two outer springs are used to centre the armature. The pole pieces are laminated and an extension (not shown in Fig. 5a) fitted at right angles to the armature is centred in a soft damping medium "viscaloid."

Moving Iron Type. A Moving Iron Type is shown in Fig. 5b. The ends of the armature are clamped between two U-shaped steel yokes within which lie two Ticonal magnets and the laminated pole pieces with their brass clamping blocks. The metal at the bottom of the deep slots at either end of the armature is shaped to form torsion bars which support the active centre portion of the armature in the gap between the pole pieces. To avoid trouble with non-torsional deflections these torsion bars, both of which are hollow, are made as short as possible, consistent with reasonable stress at maximum deflection. Through the centre of one of them passes the long plain shank of the cutter clamping screw and, as the thread is carried in an external block, it is easy to remove the broken parts if this screw should be broken by over-tightening. The shank has a high torsional compliance so that the presence of the clamping screw does not appreciably add to the mechanical impedance of the armature.



(a) Balanced Armature Type.



(b) Moving Iron Type.

RECORDING HEADS.

FIG. 5.

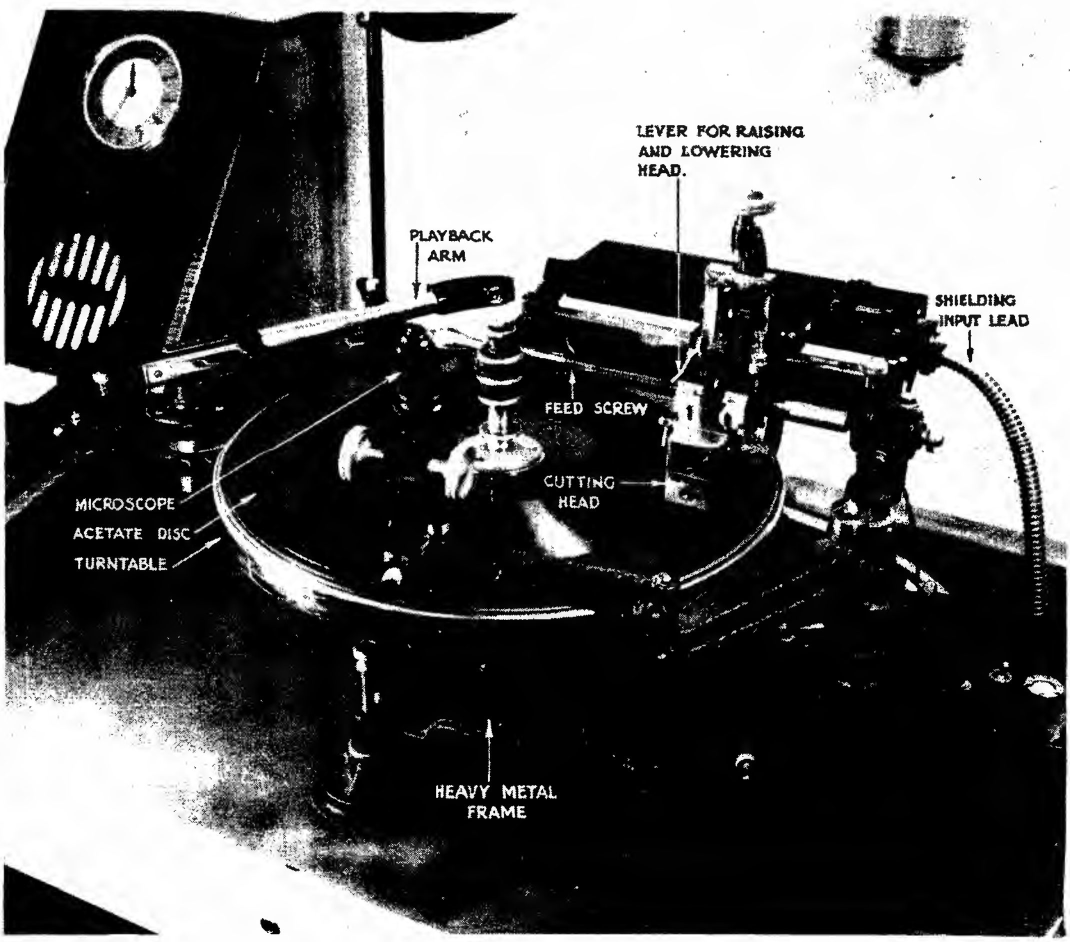
The coil is in slots in the faces of the pole pieces, and within the main, or driving, winding is a second coil which is connected to an intermediate stage of the recording amplifier so as to provide negative feedback. The voltage across this coil is proportional to the flux linked, and the amplifier, therefore, is controlled to compensate for distortion of the flux wave form. On initial assembly the air-gaps between the armature and the pole pieces are adjusted to 0.005 in., and no further adjustments are required or provided. It is necessary only to add the damping, which is done by inserting a film of oil or grease into the bottom air-gaps.

Recorder Unit. The recorder unit in general use (see Page 7) consists of a heavy cast-iron frame carrying a synchronous alternating current motor, a heavy cast-iron turn-table and an overhead mechanism carrying the cutting head with its carriage and lead screw. The motor shaft carries a steel pulley which drives to the inside edge of the turn-table through a rubber idler wheel. Two sizes of idler wheels are used, one for each speed. The motor is suspended on rubber, and, with the rubber idlers and heavy turn-table, vibration is negligible.

/ The

The cutting stylus may be a steel needle or a jewelled needle, such as a sapphire. For high quality recording, a sapphire is almost invariably used; it requires careful handling due to its inherent brittleness.

The heavy turn-table acts as a flywheel and assists in maintaining a constant speed when cutting. The cutting head is moved across the disc by a feed mechanism which is operated by worm gears driven by the rotating turn-table. The weight of the cutting head on the disc is approximately 3 oz. (representing several tons to the square inch at the point of the stylus) and is adjustable. Upon this adjustment depends the depth of the groove, and the relative widths of groove and wall. A microscope is provided for inspecting the cutting stylus and grooves. Audio power required for the cutting head varies from 400 mW to 2 W, according to design and manufacture. The impedance of the cutting heads is typical of voice-coils, that is, 2 to 15 ohms, other types may be 500 ohms.



DISC RECORDER UNIT.

2.3 Making a Recording. A summary of recording operations:

(i) The disc is wiped clean and placed on the turn-table.

(ii) A sapphire cutter is inserted in the cutting head.

(iii) The overhead mechanism is lowered and the cutting head angles checked.

(iv) A test cut without modulation is made; this cut is examined with a microscope for depth and to see that the sapphire is in good order.

(v) The modulation level is checked on the volume indicator and if satisfactory the recording proceeds. A sash brush is handy to keep the swarf clear of the cutter.
When the programme is finished a groove is cut without modulation. A repeat groove is also cut, that is, the cutting head is disengaged from the lead screw and a continuous groove is cut; this prevents the pick-up head from leaving the record when replaying.

(vi) At the conclusion of the recording, the disc is examined with the microscope and if satisfactory, the disc preservative is applied. This liquid hardens the surface and also applies a wax.

(vii) The disc is now replayed to test that the technical standard is satisfactory.

2.4 Other Points of Disc Recording Interest.

(i) A good needle makes a quiet cut and a dull needle makes a noisy cut, that is, a record with background noise.

(ii) The needle should be inserted with its flat face on the cutting side, that is, so that it is cutting against the rotation of the disc. The angle of the needle should be just off the vertical by a few degrees and inclined in such a direction that the record rotation will not cause it to dig into the record.

(iii) The balance weight on the cutting needle should be so adjusted that, when the needle is cutting at normal speed, it produces a thread about the thickness of a coarse human hair.

(iv) The input level should always be checked against the manufacturer's recommendation to avoid overload and underload.

(v) The cutting process should be practically inaudible. A needle that cuts with a hissing sound will produce a high background noise. The causes of a noisy cut may be -

(a) used or faulty needle,
(b) old record stock,
(c) insufficient input level.

(vi) The shavings from most records are highly inflammable, and care should be taken to avoid danger of fire when cutting discs. It is a good plan to place cuttings in a metal container as cutting proceeds or to put these into water.

(vii) "Wows," or a rhythmic change in amplitude in disc recording will be caused if the speed of the groove is changed in a rhythmic manner during recording or play-back. Wows are usually due to a slowing down or speeding up of the groove, but may be due to any of a multitude of causes. Typical of these causes, wow may be due to dynamic unbalance of the table, off-centring of the record blank, dragging of the record on loud passages, or out-of-roundness of the rubber idler or driving wheel, owing to its being left in prolonged contact with the turn-table at one point during periods of idleness.

(viii) A method of observing the frequency/amplitude response of a lateral disc is to hold the record at an angle to a point source of light. A pattern will be seen on the disc having a width related to the amplitude of the groove. The picture on Page 2 gives an example of this method of observing the frequency/amplitude response of a lateral disc. It is, of course, necessary in the case of a test record that the input level to the recording head be constant over the test frequency range.

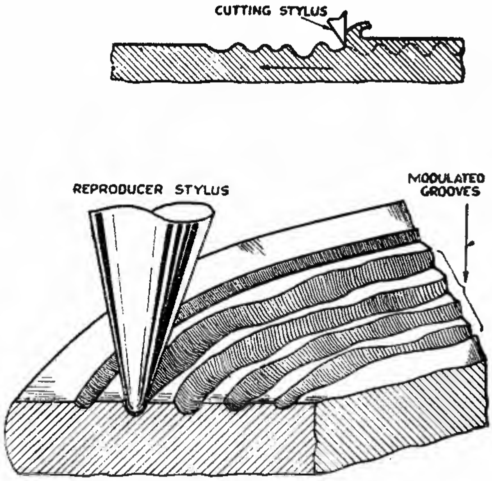
/ 2.5

2.5 Vertical Recording (Hill and Dale). In vertical recording, the cutting stylus moves in a direction vertical to the plane of the record as the record revolves. This results in a groove of varying depth. Some advantages attributed to this method are -

(i) Closer groove spacing, giving greater playing time per square inch of record surface.

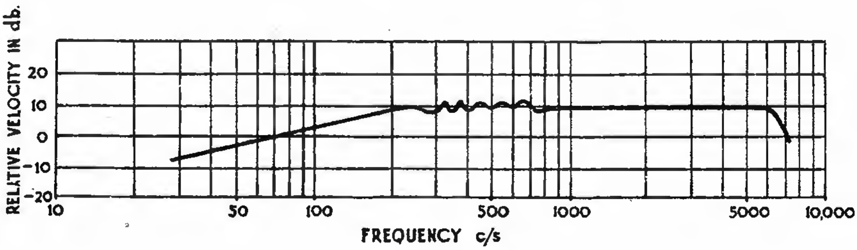
(ii) Greater positiveness of the driving stylus (since it is held in contact with the record groove) as contrasted with the slippage of the needle across the bottom of the groove in the lateral type.

The general cutting procedure is similar to the lateral, excepting that the stylus moves up and down instead of laterally across the record. This system has only a limited application and will not be described further. Fig. 6 shows this method of recording and reproducing, and Figs. 7a and 7b compare the frequency response characteristics of vertical and lateral recording systems.

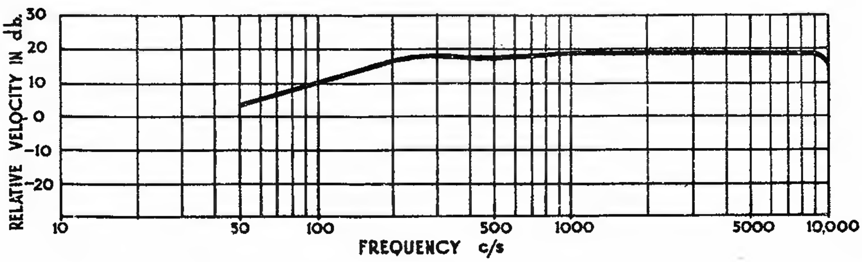


VERTICAL (HILL AND DALE) RECORD GROOVE.

FIG. 6.



(a) Lateral Recorder.



(b) Vertical Recorder.

RECORDING CHARACTERISTICS.

FREQUENCY RESPONSE.

FIG. 7.

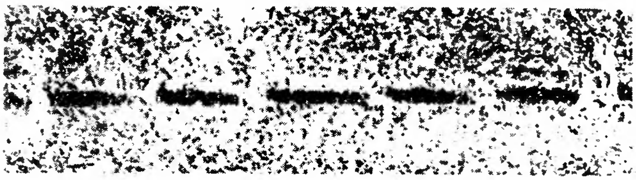
3. MAGNETIC RECORDING.

3.1 When a recording of an event, such as a sporting result or resume, is required, a Magnetic Recorder is frequently used. Magnetic recording differs from both the mechanical and optical methods in that the medium suffers no permanent change. The process consists in magnetising the medium to varying degrees along the sound track, and these variations can be removed at any time by entirely demagnetising or magnetically saturating the medium. The variations along the sound track take the form of variations in the remanent magnetisation of the magnetic recording medium, these variations being directly proportional to the instantaneous value of the signal strength corresponding to that period.

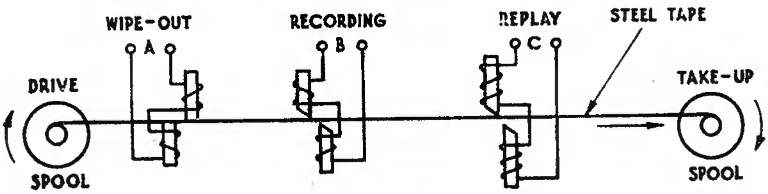
The recording medium can be any material capable of being permanently magnetised to a varying degree. Typical of the mediums used are steel wire, steel tape, coated paper tape, coated plastic tape, coated paper discs and the like.

In operation, the medium is drawn past magnetic cores surrounded by coils through which the audio currents are passed. These coils magnetise the medium longitudinally in a fashion varying with the characteristics of the audio currents. The medium requires no processing and may be played many hundreds of times without deteriorating; after being demagnetised it may be used again.

3.2 Fig. 8a is a picture of length of magnetised steel tape and Fig. 8b the elements of a magnetic recording system.



(a) Magnetisation of Tape.



(b) Diagrammatic Sketch.

TAPE RECORDER.

FIG. 8.

Although in the following description the polarising current is referred to as "D.C." a number of modern units use H.F. currents in lieu of D.C. The use of H.F. currents for erasure stabilises the magnetic medium and results in a lower background noise than the D.C. erasure. The general description which follows is however applicable to both cases.

In operation (see Fig. 8b), the tape first passes through the wipe-out head (A). This removes the previous recordings. The wipe-out head has about 10 to 20 mA D.C. flowing through it; this gives a magnetising force sufficient to bring the tape to magnetic saturation.

After leaving "A," the tape remains uniformly magnetised and passes into the recording head "B." This head carries a small polarising current flowing in the

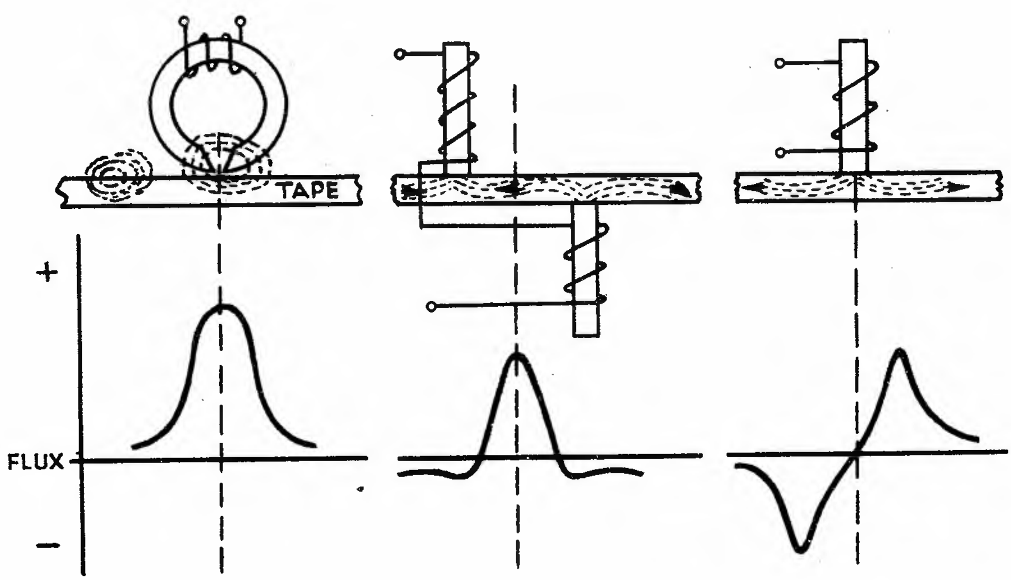
/opposite

opposite direction to that of the wipe-out head. (This may be compared with the bias which selects the correct operating point for an amplifier valve.) This small current (2-3 milliamperes) sets the tape magnetically on the straight part of its magnetising curve. Superimposed upon this direct current are the alternating speech currents from the recording amplifier. The speech currents will alternately oppose and assist the polarising current, producing a decrease or increase of magnetising force on the tape. Since the tape is set on the straight part of the β/H magnetic curve where linear conditions can be obtained, this will result in a corresponding increase or decrease of magnetic flux. The alternating speech currents will thus be represented on the tape by a variation in the intensity of magnetisation along its length.

The replay head "C" is similar to the recording head, but has no polarising current. The windings are connected in series aiding, and the tape, by virtue of the changes in the state of magnetisation along it, causes changes of flux in the magnetic circuit of the coils as it passes between them. This induces a voltage into the coil windings, which corresponds both in frequency and amplitude to the input signal voltage to the recording head. This output is fed into the amplifying systems as required. As in the case of discs, equalisers can be used to improve the noise and response characteristics.

3.3 The following additional details of tape recording systems are of interest. Reference should be made to Applied Electricity II to refresh the memory regarding hysteresis loop, remanence, β/H curves etc., which are fundamental to magnetic recording.

Recording Heads. Fig. 9 shows three different types of recording heads and of these the ring head is used by many manufacturers. Most systems use longitudinal magnetisation of the tape (see Fig. 8a) and this component is shown in the figure. The ring head gives the most sharply defined field, whilst as shown by the curves the double head has a negative field component which could have undesirable effects.

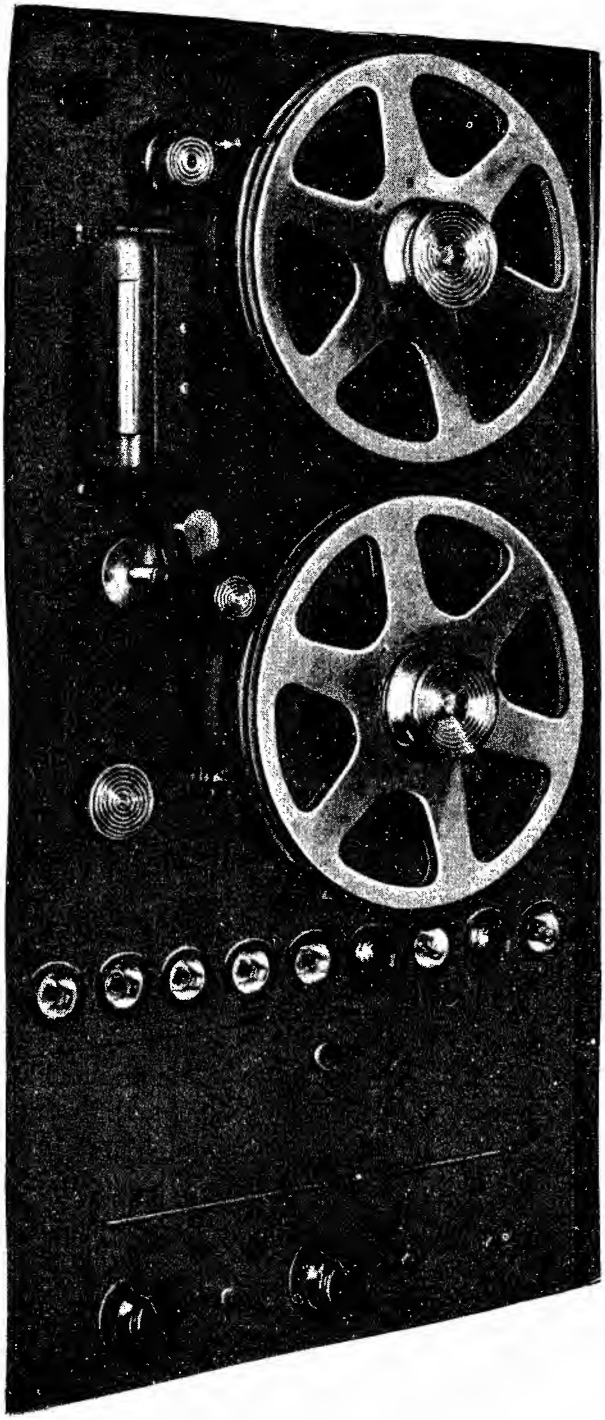


(a) Ring Head. (b) Double Pole Piece. (c) Single Pole Piece.

RECORDING HEADS.

Note. For explanatory purposes the tape thickness has been increased.

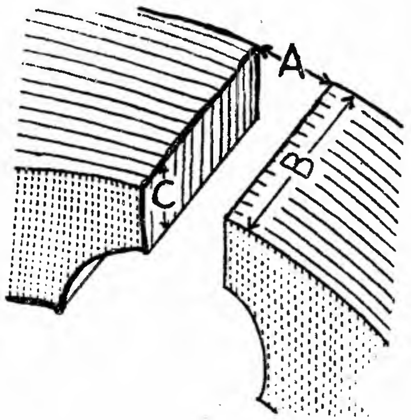
FIG. 9.



TAPE RECORDER.

RACK MOUNTED, WITH PUSH BUTTON OPERATION. RESPONSE 50-15,000 c/s.

Gap. In the ring head the gap provides for the external flux field necessary for magnetising the tape. It should be as small as possible and in practice varies from about 0.005" to 0.001". Fig. 10 shows a typical design, and the designations generally adopted. The exact effect of the gap in reference to type of bias, recording and replaying, is beyond the scope of these notes. The gap in a replay head should be small; the tendency is to use the one head for both recording and replaying with a gap of 0.001".



GAP OF MAGNETIC HEAD.

A = Width
B = Length
C = Depth

FIG. 10.

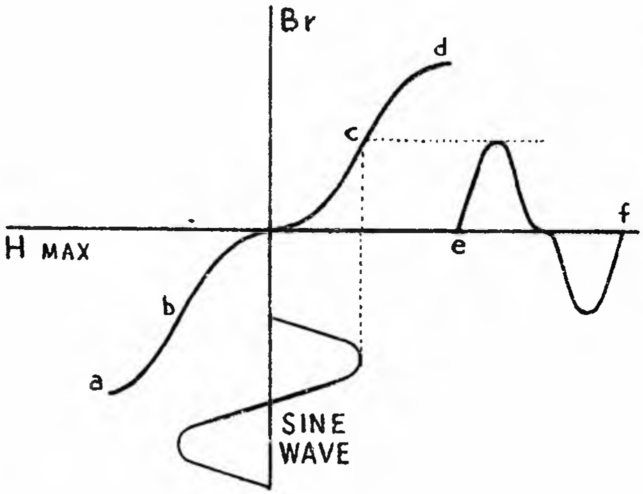
Wow. Magnetic recording must be free from discernible frequency wow and amplitude wow. Frequency wow is the result of variations in speed of the tape. A flywheel is usually coupled to the capstan to help in maintaining a uniform speed.

Amplitude wow is more likely to occur with magnetic recording than with ordinary disc recording. Such periodic or random variation of loudness is due to inequalities in diameter of wire or uneven density and thickness of magnetic particles in coated tape, and variation in mechanical contact between tape and replay head.

The ear is most sensitive to wow which occurs between 1 and 8 c/s; the effect is more noticeable on high notes than on low frequencies.

Magnetisation of Heads. The core of the head is usually assembled in two halves, with a separate winding on each, so that when brought together two gaps are formed. The rear gap serves to reduce remanence in the core, and to ensure that it does not become magnetically saturated. The coils are wound to reduce hum pick-up in the head from the motor, mains transformer or associated parts.

In spite of precautions, the heads may become partially magnetised during use, with an increase in noise level.



REMANENCE CURVE OF MAGNETIC MATERIAL, WITH RESULT ON PURE SINE WAVE INPUT.

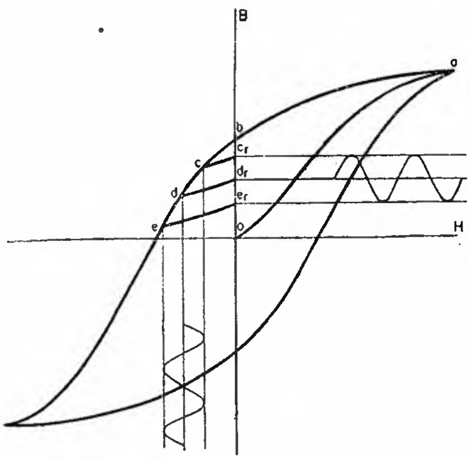
FIG. 11.

Bias. It is essential to have some form of bias in order to limit the recording to a linear portion of the hysteresis loop. Without bias, reproduction of speech or music is badly distorted. Reference to Fig. 11 will show this. It will be seen that the wave form is distorted by the non-linearity of the remanence curve.

One of the earliest methods of overcoming this distortion was the application of D.C. bias to bring the recording operation to a linear section of the curve around point c in Fig. 11. By limiting the level of modulation, an output of good quality could be obtained.

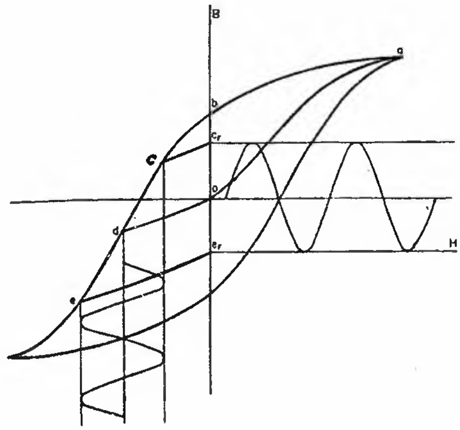
Improved methods of D.C. bias recording are shown graphically in Figs. 12 and 13. In Fig. 12, D.C. bias is applied to bring the magnetic material to saturation, which also serves to erase any previous traces of recording. Low level output of good wave form is obtained.

In Fig. 13 an increase of 6 db in the output level is achieved at some sacrifice of quality. Signal is applied about the point d of the hysteresis loop, and the iron is in a neutral magnetic condition during periods of zero input, resulting in the minimum of background noise during reproduction.



D. C. BIAS RECORDING.

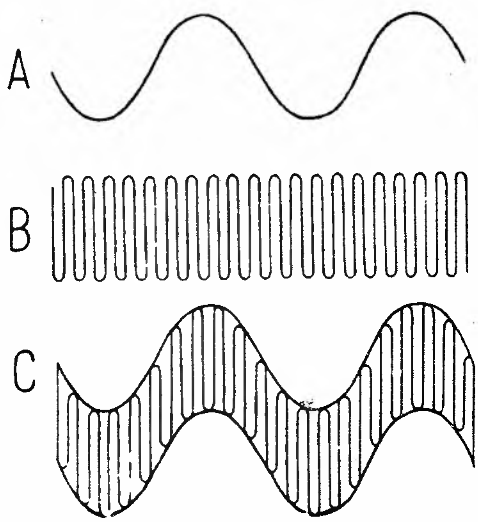
FIG. 12.



D. C. BIAS RECORDING. THE BIASING FIELD IS SO ADJUSTED THAT THE WIRE OR TAPE IS DEMAGNETISED DURING PERIODS OF ZERO SIGNAL INPUT.

FIG. 13.

High Frequency Bias. This method of bias results in improved fidelity and higher output than the methods previously described. The theory is beyond the scope of this book. The frequency of the bias should be several times that of the highest recorded audio frequency to avoid intermodulation. The combination of the audio frequency signal and H.F. bias is outlined in Fig. 14.



A = Audio signal.

B = H.F. bias.

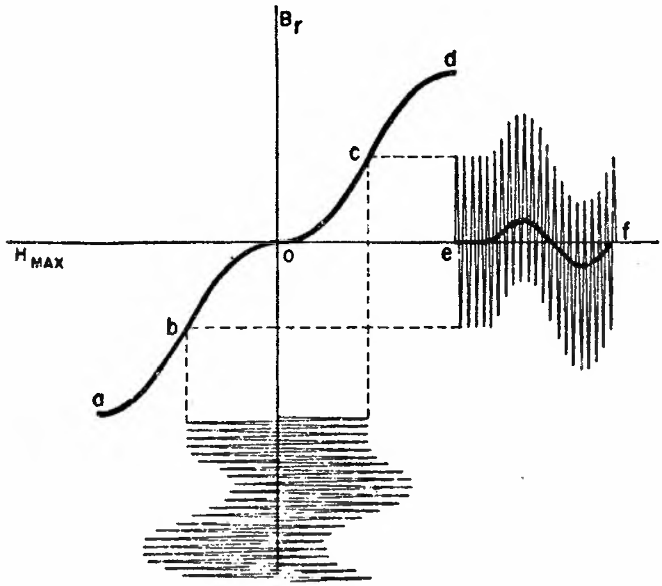
C = Combination wave.

COMBINATION OF AUDIO AND H.F. BIAS.

FIG. 14.

Provided the wave form of the high frequency is pure and the optimum ratio between signal amplitude and bias amplitude is maintained, a recorded signal of low distortion will result with high signal-to-noise ratio.

Fig. 15 shows the application of H.F. bias and the resulting wave form.



H. F. BIAS.

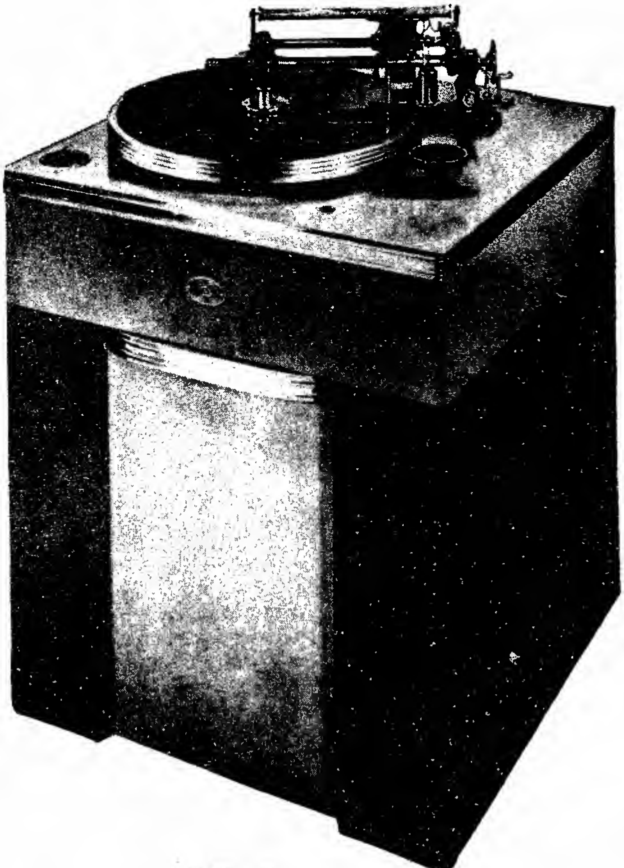
FIG. 15.

A very important factor in the recording process in the demagnetising force, which acts at the high frequencies and reduces greatly the amplitude of induction in the magnetic material after it leaves the recording gap. This effect is especially pronounced at the bias frequency. After the medium leaves the gap, demagnetisation takes place at bias frequency. A net induction then remains which is the difference between the positive and negative half-cycles of the bias wave, this difference being in turn a function of the audio signal. The remaining induction is shown by the curve of e-f of Fig. 15. Where the audio signal is zero the medium is almost completely demagnetised, with the result that the noise level is low. Furthermore, if the bias amplitude is properly adjusted and the audio signal is not too large, the audio signal is transferred to the medium according to the linear portions b and c of the characteristic curve, with the additional result that the distortion is low.

Sometimes the same oscillator that supplies the bias current is also used to supply the erasing current; this arrangement has an advantage over two oscillators in that there is no possibility of beats and other undesirable interactions between bias and erase frequencies.

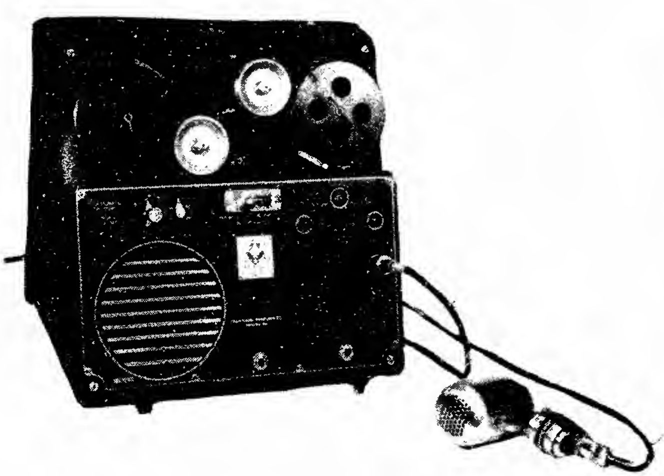
The current required for erasure is much higher than the bias current.

3.4 Wire Recorder. A combined recording and reproducing unit using steel wire is shown on Page 16. The unit contains a rectifier, audio amplifier, recording mechanism, drive motor and associated accessories. The general scheme of operation is as described for the tape recorder, the main difference being in the use of 4 mil medium carbon steel recording wire instead of tape. Two recording speeds are provided, giving 33 minutes of recording at high speed (400 r.p.m.) and 66 minutes at low speed (200 r.p.m.). During the play-back period, the audio output of the amplifier is connected to a built-in loud-speaker.

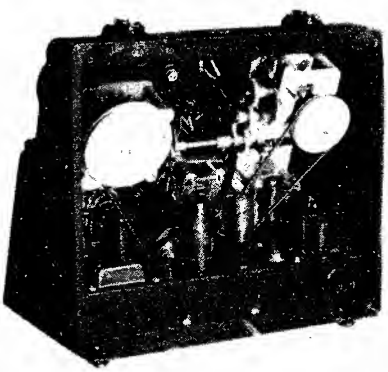


DISC RECORDER.

TAPE RECORDER (PORTABLE).



FRONT VIEW.



BACK VIEW.

WIRE RECORDER (PORTABLE, WITH PLAY BACK).

RECORDING UNITS.

3.5 Recent Developments in Tape Recorders. Because of the usefulness of tape recorders, efforts are being continually directed to effecting improvements in their performance. A recent development has been an instrument in which the tape consists of a plastic film, one side of which is coated with a ferro-magnetic material. This tape has less tendency to sag and stretch than does wire under varying atmospheric conditions, and this materially reduces flutter and wow effects.

The frequency response is flat within 2 db from 30 to 10,000 c/s and an improvement in signal-to-noise ratio has been achieved. The tape speed is about 30 inches per second and the coils contain about 3,300 ft. of tape giving approximately 20 minutes playing time. Other units giving up to one hours playing time with coated plastic tape have been developed. Tapes are readily joined with a special adhesive.

A number of tape units are in use or in the process of being installed in the A.B.C. studios throughout Australia and are giving good service.

All tape and wire recorders, however, have one particular disadvantage as compared with discs, they do not permit convenient programme "spotting." It is much easier to mark a certain spot on a disc and a number of spots may be rapidly selected.

3.6 Miscellaneous Notes. The H.F. response of a tape recorder is affected by the running speed of the tape and some indication of this effect is seen in the following table.

TYPICAL RESPONSE RANGES.

Tape Speed	Response Range c/s	Spool Size	Length of Tape	Playing Time
3-3/4 inches per second	100-3,500	5"	600 ft.	32 minutes
7-1/2 inches per second	50-7,000	7"	1,200 ft.	32 minutes
15 inches per second	30-12,000	11"	3,250 ft.	43 minutes
30 inches per second	30-15,000	11"	3,250 ft.	21-1/2 minutes

The above will naturally vary with the quality of tape, and the width of the replay gap, but are illustrative.

Paper Discs. An interesting development of magnetic recording is the use of magnetised paper discs. This system is primarily for office use, such as dictation, making extracts from discs or tape, etc. The recording medium is a coated paper disc and the recording head is driven from a drive screw as in disc recording, the sound track thus being a spiral magnetic track. In replaying the head is also driven by the drive screw since there is no groove for it to follow.

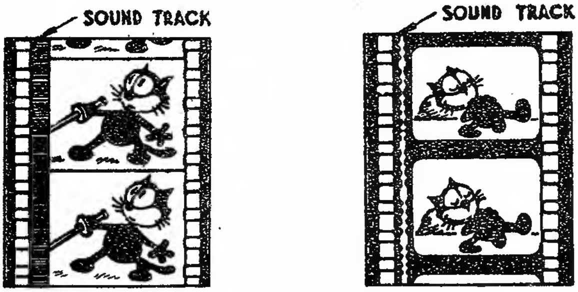
Erasure is by means of a permanent magnet held in contact with the disc. The discs are thin and may be mailed in an ordinary envelope, and can be used several hundred times.

The playing time is usually between 3 and 6 minutes which places some limit on its usefulness.

/ Tape

4. OPTICAL RECORDING.

4.1 The optical or photographic method of recording is used mainly for sound motion films, but some application of its principles has been tried with portable recorders for broadcasting use. A brief reference only will be made, since the system is not yet in common use for broadcasting.



(a) Variable Density Track. (b) Variable Width Track.

SOUND TRACKS.

FIG. 16.

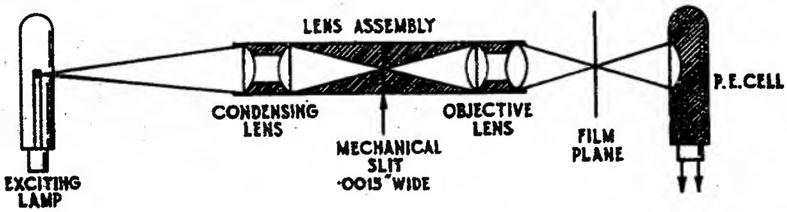
Two types of sound-on-film records are in common use today. One type is the variable density method - a series of striated bands as shown in Fig. 16a. The other type is the variable width method shown in Fig. 16b - a serrated band with tooth-like projections.

Optical recording is a constant amplitude method, that is, for a system having a uniform frequency characteristic, the electrical power input to the amplifying system is proportional to the amount or amplitude of light falling on the film in recording. Among the methods of producing such film sound tracks are -

(i) Reflecting galvanometer system.
(ii) Flashing-lamp method.
(iii) Light-valve system.

Method (iii) has seen very wide use in studio, location and newsreel recording, which use has shown it to be capable of making sound records of high quality. It can be adapted for either variable-density or variable-width recording.

4.2 In reproducing either the variable-density or variable-width sound films, the light from a straight incandescent filament is made to fall on a slit about 0.0015 inches wide. This image is, in turn, focused on the film as a slit about 0.001 inches wide. (See Fig. 17.)



BASIC SOUND-ON-FILM REPRODUCTION CIRCUIT.

FIG. 17.

The light which passes through the film falls on the sensitive portion of a photo-electric cell. The currents generated by this valve or cell will vary in intensity and frequency in accordance with the light variations of the film, thus reproducing as electric currents the original audio frequency modulation.

/ 5.

5. REPRODUCTION OF RECORDS.

5.1 The reproduction of tape and wire recorders has been referred to in previous paragraphs. The reproduction of disc type records requires the following equipment -

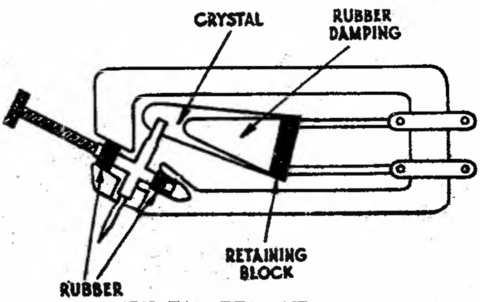
(i) Pick-up unit.
(ii) Gramophone machines.

The reproducer should be of the constant velocity type, and, provided resonances do not upset this condition, the electrical output will be similar to that of the recording head. It is necessary, therefore, to use electrical networks, that is, record equalisers, to compensate for the recording characteristic. The desired compensation required was shown in Fig. 3.

Of the many types of reproducers developed over a number of years, the following types have general use -

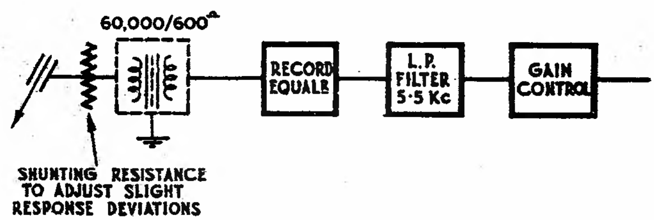
(i) Crystal.
(ii) Moving coil.
(iii) Moving armature.

5.2 Crystal Pick-up. The crystal type has been used extensively, due to its low cost, high voltage output and low distortion, and has been the best type available. Fig. 18 shows a typical type. The rear end of the crystal is securely fastened and the needle applies a bending motion to the free end. Rubber damping blocks are mounted on each side, and serve not only to position the crystal but to control the frequency response. This type is designed to operate into a high impedance load of 1/4 megohm. This is not suitable for broadcasting purposes and the arrangement used is shown in Fig. 19.



CRYSTAL PICK-UP.

FIG. 18.



CRYSTAL PICK-UP - TYPICAL CONNECTION.

FIG. 19.

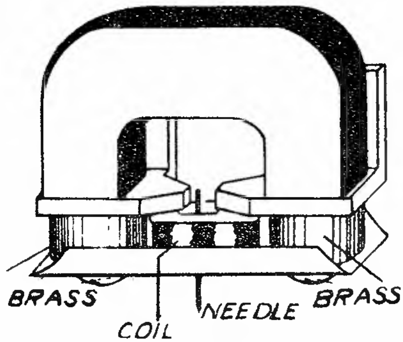
The impedance of 60,000 ohms effectively removes the low frequency in the response of the crystal pick-up and the record equaliser applies the correct equalisation, while the low pass filter cuts off the frequencies above 5,500 c/s. This is done to reduce the electrical noise produced at the surface of the record. Commercial discs have a surface of shellac with earth fillers, and, as the steel needle follows the grooves, a noise is produced which is chiefly of a high frequency nature, thus the filler reduces this noise without affecting, materially, the programme matter.

5.3 Moving Coil Pick-up. The principle of this type is similar to the moving coil loud-speaker. Although not in general use, experience has shown that moving coil pick-ups are not robust and the electrical output is low.

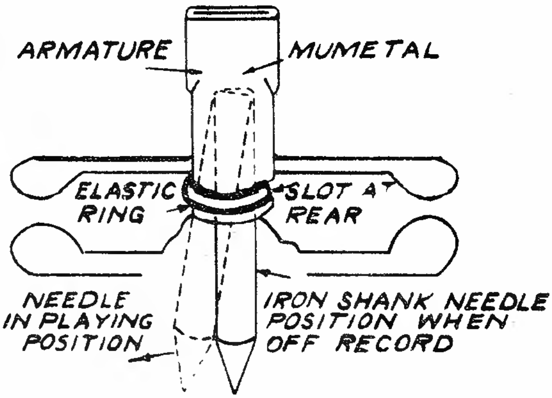
Another moving coil type, which is more robust, is somewhat like a simple alternator in construction. A permanent magnet supplies the field, and the coil (armature in shape but without the metal centre) is set in rubber at the ends. Although the needle does not rotate the armature as it does in an alternator, there is a partial movement, therefore, the electrical action is the same. The output of this type is low.

/ 5.4

5.4 Moving Armature Pick-up. Many forms of this type have been produced but, in general, although they have a satisfactory electrical output, the harmonic distortion is not easily controlled as the symmetry of the armature in relationship to the poles determines the degree of the distortion. Fig. 20 shows a Moving Armature Pick-up.



(a) Construction.



(b) Details of Needle Holder.

TYPICAL MOVING-ARMATURE TYPE MAGNETIC PICK-UP.

FIG. 20.

In this pick-up, the needle is pushed from side to side by the action of the record groove, this movement producing a corresponding variation of the armature position. Each movement of the armature causes a redistribution of flux in the magnetic field and changes the flux distribution through the coil, inducing a corresponding e.m.f. in it.

Later types of magnetic pick-ups use a light non-magnetic ribbon placed in a magnetic field in such a way that the motion of the needle is transmitted to the ribbon which, in cutting the magnetic field, has a voltage induced into it. The needle is also mounted in such a manner as to allow it to move in any direction, either vertical or lateral, thus it suits both lateral and vertical type recordings.

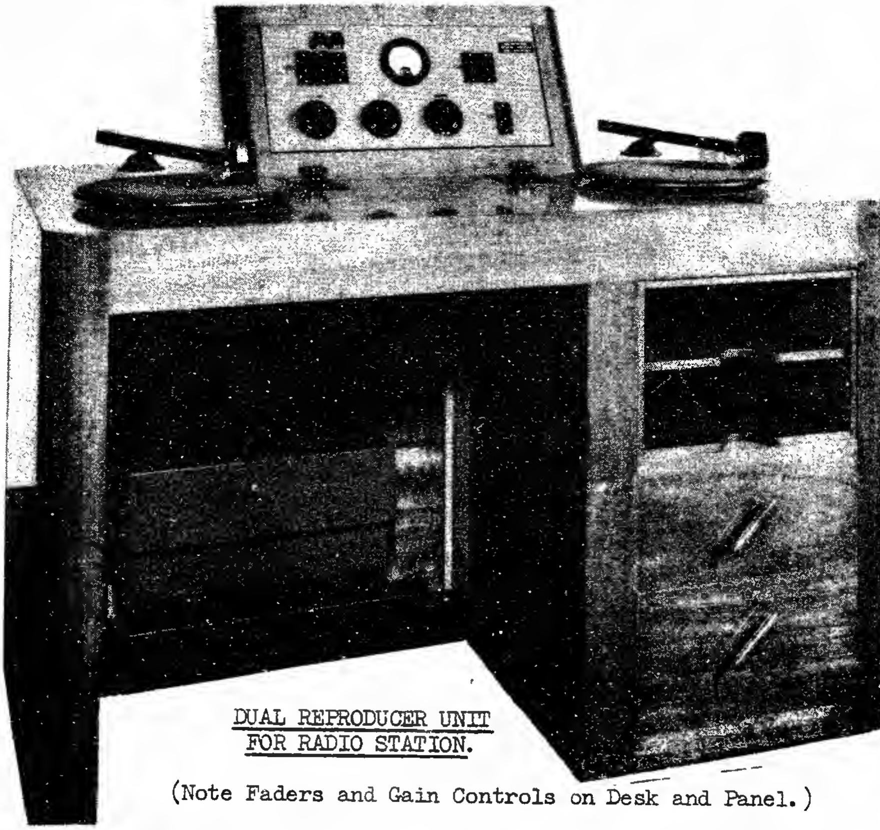
5.5 Requirements for Reproducer. The essential requirements of a satisfactory reproducer are as follows -

Must not be damaged when dropped on the needle point.
Satisfactory electrical output.
Low harmonic distortion.
Easy means of changing needle.
Low needle stiffness, otherwise needle will not follow the groove modulation, that is, will not track.
Weight at needle point should not exceed 1-1/2 ounces.
The arm and head should have sufficient mass, so that the mechanical resonance is below 50 c/s.
The construction of the arm and head must be mechanically good with low tracking error.
Electromagnetic fields must not cause excessive electrical noise.
The performance of the reproducer must be constant with time, temperature and humidity.
The frequency response should be preferably of the constant velocity characteristic within close limits. Failing this, the response must be of a type which can be readily equalised. This does not refer to the record equaliser.

5.6 The tone arms on all pick-up devices should be counter-balanced, in order to relieve the record of unnecessary weight, to reduce wear on the record and to increase

/ the

the life of the pick-up head. Pick-up heads may be wound to any desired impedance, 50 ohms, 200 ohms and 600 ohms being typical values for conveniently working into amplifiers. The usual output of a pick-up is -35 to -40 db.

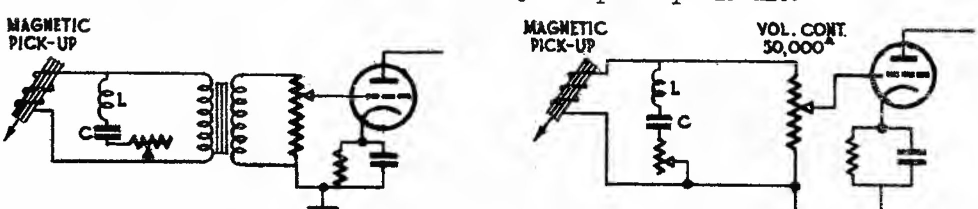


DUAL REPRODUCER UNIT
FOR RADIO STATION.

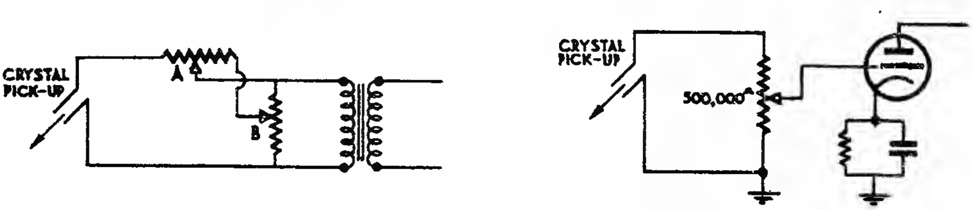
(Note Faders and Gain Controls on Desk and Panel.)

5.7 To minimise surface noise, a so-called scratch-filter is generally connected directly across the input transformer or pick-up device to attenuate the high-frequency hissing voltages present at that point. The values of L and C are usually chosen to resonate at about 5,000 or 6,000 c/s.

Typical magnetic pick-up circuit connections are shown in Fig. 21a and Fig. 21b shows the connection of a volume control to a crystal pick-up circuit.



(a) Magnetic Pick-Up and Scratch-Filter.



(b) Crystal Pick-Up and Volume Control.

FIG. 21. REPRODUCER CIRCUITS.

5.8 Gramophone Needles. Steel needles are used for the replay of records, except in the recording room where sapphire needles are generally used. The needles used for commercial discs are of the short thick type, as the longer and thinner the needle the greater the high frequency attenuation. For the replay of lacquer coated discs "shadowgraph" steel needles are used. Each of these needles has been examined for imperfect points by projecting the outline, many times magnified, on to a screen; an imperfect point would damage the grooves.

5.9 Gramophone Machines. The requirements for a satisfactory machine for studio purposes are stringent. The main requirements are as follows -

(i) Low acoustic noise.

(ii) Low mechanical vibration.

(iii) High torque, that is, high rotating force.

(iv) Speed must remain constant within close tolerances.

(v) Simple to operate.

(vi) Must operate at both 78 and 33-1/2 r.p.m.

A typical machine includes a synchronous alternating current motor with a speed of 1,500 r.p.m. which drives the 16" turn-table through a steel pulley fitted to the motor shaft. The turn-table has a soft rubber tyre fitted to the external edge, and the motor is rubber suspended. To obtain the change of speed, the steel pulley has two diameters and the motor is lifted or lowered by a simple mechanism. The rubber is used to attain the first two requirements listed above. The use of rubber, however, for the drive does not permit constant speed for high loading, but the speed is satisfactory for the loading of a gramophone reproducer.

Mechanical and acoustic noise is due entirely to the motor. The motor must be of special design and may be as low as 1/60 H.P. The turn-table of cast aluminium, generally about 8 lb., ensures almost negligible variations of speed.

5.10 Tracking Error. "Tracking error" is the angle between the tangent to the record groove at the point of needle contact and the needle itself projected into the plane of the record. In the ideal case, the needle axis and the groove tangent at the point of contact should both be in one plane perpendicular to the record surface, but the use of arms pivoted at one point prevents this, with the result that damage to the grooves results and distortion is introduced, unless the tracking error is maintained at low values. The bent arms and offset heads on various types of reproducers are aimed to reduce the tracking error.

The following troubles can be experienced when the error is large -

(i) The needle pressure is exerted mainly on one side wall of the groove, instead of being equally divided between the groove walls.

(ii) When a large error exists at one point on the record, it will be small at other points. This means that the shape of the needle will not fit all grooves equally well.

(iii) The needle will not produce sinusoidal modulation vibration from a sinusoidal groove, thus wave form distortion is produced.

Two methods are used to reduce tracking error -

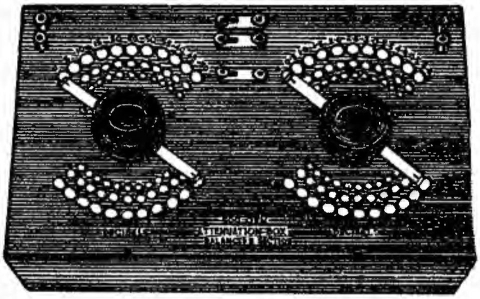
(i) Make the arm reasonably long.

(ii) Offset the head and adjust to give correct overhang of needle beyond centre spindle.

/ 6.

6. ATTENUATORS.

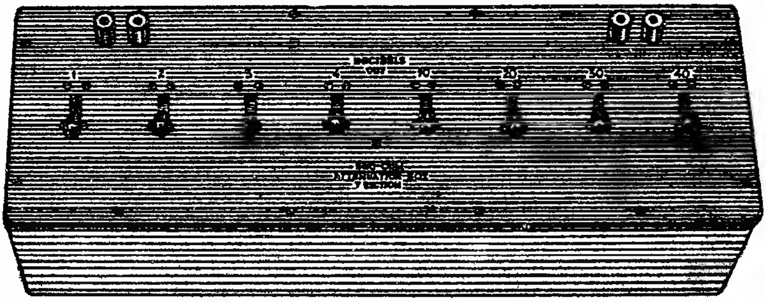
6.1 Attenuators may be defined as resistance networks used for the purpose of reducing voltage, current or power in known and controllable amounts. Attenuators can be of the balanced or unbalanced type, the former being in general use. Attenuators are designed for specific impedance conditions. For instance, 600 ohms is an impedance generally used for testing purposes, and it is necessary that an attenuator, to be used for testing in such a circuit, has a characteristic impedance (Z_0) of 600 ohms. The impedance of the attenuator will be 600 ohms if measured at, say, the input terminals, provided the output terminals are connected to a resistance of 600 ohms and vice versa. The impedance will be constant for all settings of the attenuator.



BALANCED H ATTENUATOR.

Attenuators are of many designs, but five forms, which are in general use, are -

"T" type. "H" type (balanced "T").
"π" type.
Bridged "T" type.
Bridged "H" type.
Matching networks.



KEY TYPE ATTENUATOR.

6.2 Fig. 22a shows the circuit of a "T" type attenuator with a characteristic impedance of 600 ohms. The variable arms are controlled from a single control. To increase the attenuation, the resistance of the series arms Rs is increased, while the resistance of the shunt Rp arm is decreased. When the attenuation is decreased, the reverse is the case, and, at zero attenuation, the series arms are short-circuited while the shunt arm is open-circuited. In Fig. 22a, where $Z_A = Z_B = Z$, then the values of arms are -

$$R_p = \frac{2Z\sqrt{N}}{N-1}$$

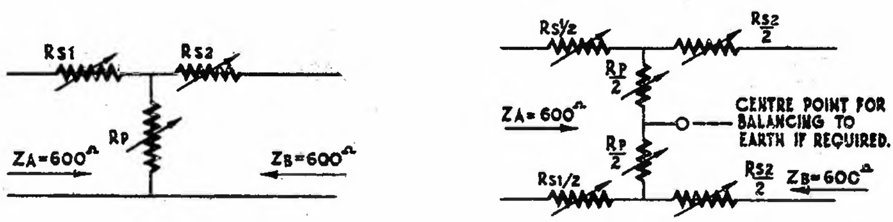
$$R_{s1} = R_{s2} = Z\left(\frac{N+1}{N-1}\right) - R_p$$

where N = ratio of $\frac{\text{Power absorbed by pad}}{\text{Power delivered to load.}}$

/ 6.3

6.3 Fig. 22b shows the circuit of a balanced "H" attenuator. It will be seen that this attenuator is really the balanced form of the "T" attenuator. The shunt arm is centre tapped, so that an earth can be connected to reduce longitudinal currents. The variable arms can be controlled by a single knob, or each decibel value can be key controlled. For instance, the single knob type could have 10 steps each of 5 db and the knob could have 10 steps each of 0.5 db, which is a total of 55 db. A key controlled type could have 8 keys arranged to have the following decibel values - 1, 2, 3, 4, 10, 20, 30 and 40, which is a total of 110 db in 1 db steps. Other types have a number of keys and then a knob control to give the fractional values, that is, 10 steps of 0.5 db. This will give 110 commonly used types for test purposes.

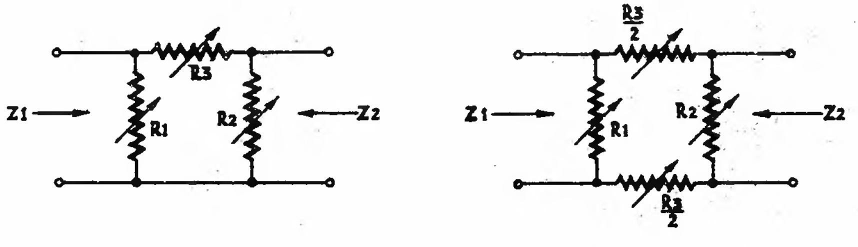
The formula as given for Fig. 22a applies, excepting that the values are halved.



(a) "T" Pad.   (b) Balanced "H."

FIG. 22. ATTENUATORS.

6.4 "π" Type Attenuator. The "π" or "pi" type attenuator is so called from the resemblance of its elements to the familiar symbol "π", as will be seen from Fig. 23 which shows this type of network.



(a) Unbalanced Type.   (b) Balanced Type. (Also Called "O.")

FIG. 23. π ATTENUATORS.

In Fig. 23, when $Z_1 = Z_2 = Z$, then -

$$R_3 = \frac{(N - 1)\,Z}{2\sqrt{N}}$$

$$\frac{1}{R_1} = \frac{1}{R_2} = \frac{1}{Z}\;\frac{(N + 1)}{N - 1} - \frac{1}{R_3}$$

6.5 Choice of T or π networks for attenuators may be based on the following considerations -

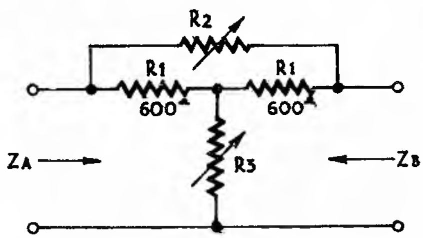
(i) For balanced circuits, the π network becomes a square of four resistances, while the T network becomes an H network requiring five resistances.

(ii) For variable attenuators for unbalanced circuits controlled by keys switching individual pads, the switching of T networks can be simpler than for π networks.

/ (iii)

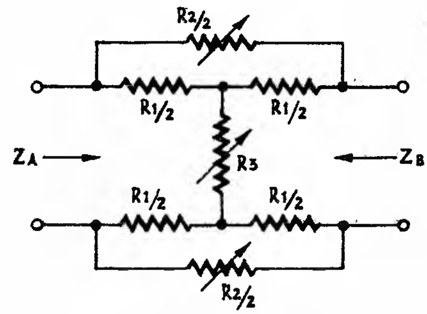
(iii) For variable attenuators with rotary controls, T network construction may be somewhat simpler than π, but errors at high or low attenuation due to poor brush contact are less serious with π networks.

6.6 Fig. 24 shows the unbalanced form of the "Bridged T" attenuator, and Fig. 25 shows the balanced T type.



UNBALANCED BRIDGED "T."

FIG. 24.



BALANCED BRIDGED "T."

FIG. 25.

Although not in general use the unbalanced bridged "T" attenuator has possibilities for use in fader circuits. It has only two variables and has zero insertion loss (this will be discussed under "ladder networks"). In addition, the characteristic impedance is constant irrespective of the attenuator setting. This network is designed to operate only between equal resistive terminal impedances $Z_A = Z_B = Z$. It is a useful form because only two variable elements are required.

In Fig. 24, $R_1 = Z$; $R_2 = Z\,(\sqrt{N} - 1)$; and $R_3 = \dfrac{Z}{(\sqrt{N} - 1)}$

6.7 An example in calculating a bridged attenuator will be worked out to show the use of the formula.

Let $Z_A = Z_B$ 600 ohms. $N = 100 : 1$ (that is, 20 db).

$R_1 = 600$ ohms.

$R_2 = 600\,(\sqrt{100} - 1)$.

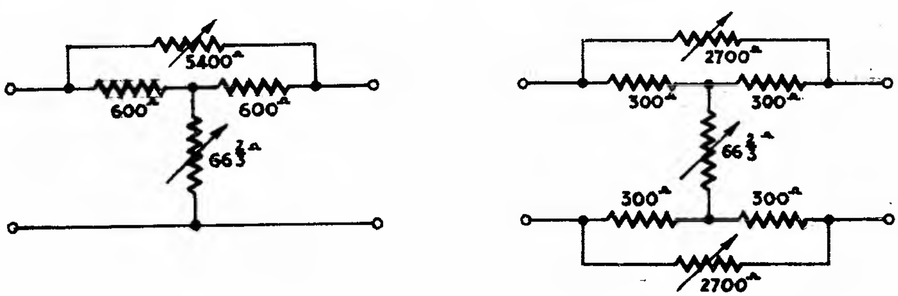
$= 600 \times 9$

$= 5{,}400$ ohms.

$R_3 = \dfrac{600}{(\sqrt{100} - 1)} = \dfrac{600}{9}$

$= 66\frac{2}{3}$ ohms.

Thus, a Bridged T for a 600 ohm circuit to give 20 db maximum attenuation would be as in Fig. 26a, and a Balanced T type as in Fig. 26b.

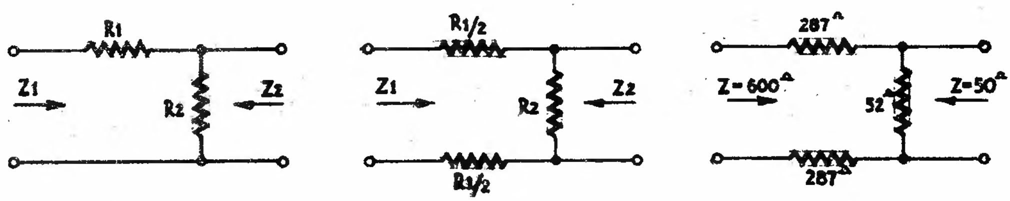


(a) Unbalanced Bridged T.

(b) Balanced T.

FIG. 26. EXAMPLE OF ATTENUATOR CALCULATION.

6.8 Matching Networks. Although transformers are used extensively for impedance matching, for testing purposes use is made of matching networks. For instance, if a pre-amplifier is to be tested, the 600 ohm oscillator circuit would be connected by means of a 600 to 50 ohms matching network. If a transformer were used instead, it is probable that certain errors would be introduced. Fig. 27 shows a typical 600 to 50 ohms network with an insertion loss of 16.5 db, which is the minimum loss for this type of network.



(a) Unbalanced. (b) Balanced. (c) Example.

MATCHING NETWORKS.

FIG. 27.

Networks of this nature have constant impedance conditions and are wound with non-inductive resistances, so that frequency errors will not be introduced.

For minimum loss pad matching Z_1 to Z_2 ($Z_1 > Z_2$), the formula is -

$$R_1 = \sqrt{Z_1(Z_1 - Z_2)}$$

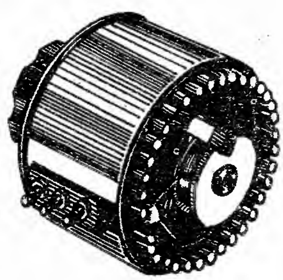
substituting and $R_1 = \sqrt{600(600 - 50)} = \sqrt{330,000} = \underline{575 \text{ ohms}}.$

$$R_2 = \frac{Z_1 Z_2}{\sqrt{Z_1(Z_1 - Z_2)}}$$

from which $R_2 = \dfrac{600 \times 50}{575} = \underline{52 \text{ ohms}}.$

7. FADERS AND GAIN CONTROLS.

7.1 Since the earliest days of broadcasting, there has always been a need for combining two or more sources of energy, such as microphones and amplifiers, into a common load impedance, and of controlling the various levels of energy relative to and independent of each other. Early methods of mixing utilised simple or parallel connections of microphones and associated equipment with consequent high insertion losses and distortion. The devices which control the relative levels of energy of each channel are known as faders, and are simply variable attenuators which may be smoothly changed over a range of attenuation from a small minimum to 80 to 100 db, although 40 to 60 db is a convenient design value. A Fader, as its name implies, provides the auditory sensation of a gradually diminishing sound, fading imperceptibly into the background noise and vanishing at maximum loss.

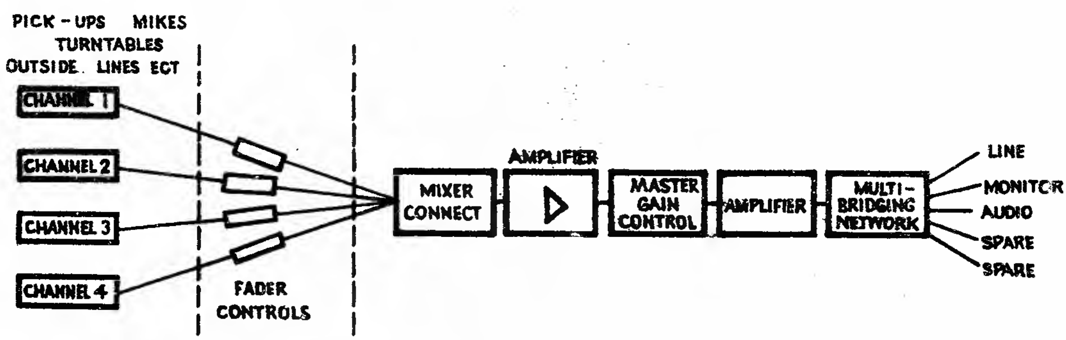


MASTER GAIN CONTROL.

(See Fig. 28 also).

The reverse condition is called "fading in," and the sound is heard gradually increasing without any unpleasant jarring of the auditory system, which would be the case in a sudden change from silence to a loud level of sound.

Devices which accept the energy of the combined groups of faders are known as Master Gain Controls, the actual mixing being done through the connections of the faders. The master gain control is another attenuator, and is used to regulate the over-all level of the total output of all channels. Fig. 28 shows a typical arrangement.



TYPICAL ARRANGEMENT FOR FADERS.

FIG. 28.

7.2 Mixing Methods. Mixer and fader control circuits may be of two general types -

(i) Electronic devices.
(ii) Resistive devices.

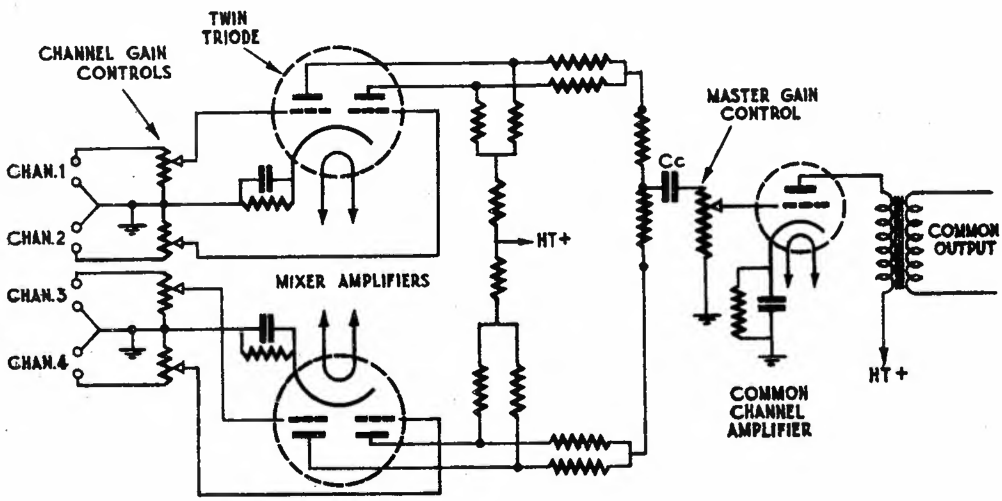
7.3 Electronic Mixing. Electronic mixers are variable-controlled amplifiers using one amplifier or more per channel. The output sides of the amplifiers may be fed through a suitable mixing network or commoned together through coupling condensers. Although seldom used, electronic mixers have a number of advantages -

(i) Freedom from direct interaction between input circuits, when mixing takes place in the output.
(ii) Because of the gain available, mixing may be done directly from low energy sources, such as microphones.
(iii) Ease of use with variable equalisation schemes for expansion or compression circuits.

Some disadvantages are -

(i) Relatively high initial cost.
(ii) High maintenance costs.
(iii) Inflexibility caused by power supply requirements.
(iv) Greater liability to faulty conditions through component failures.

Fig. 29 shows an electronic mixing circuit.



ELECTRONIC MIXER.

FIG. 29.

7.4 Resistive Mixers. This type of mixer has the following advantages -

(i) Relatively low initial cost.
(ii) Negligible maintenance cost.
(iii) Light weight and compactness.
(iv) Ease of construction.
(v) High flexibility for most purposes.

The disadvantages are -

(i) They may not be used for low level mixing without an accompanying amplifier, if the channel outputs are to go over a loop or line.
(ii) Relatively high losses, which must be compensated for by amplifiers.
(iii) Noise caused by dirty contacts.

/ Some

Some points to be considered in design are as follows -

(i) Number of channels required.
(ii) Impedances of inputs and outputs.
(iii) Limits of insertion loss, both maximum and minimum.
(iv) Whether low or high level mixing.
(v) If constant fader impedance required over attenuation range.
(vi) Freedom from interaction between fader attenuators.
(vii) Compactness, weight and initial cost.
(viii) Care in wiring, shielding and screening to avoid extraneous pick-up.

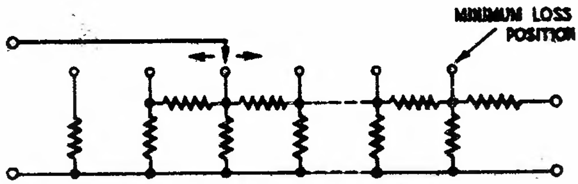
Control networks which include faders and gain controls are of two types -

(i) Unbalanced types.
(ii) Balanced types.

It is usual for faders to be of the unbalanced type, as they are isolated by the output transformers on one side and an input transformer on the other side.

On the other hand, a master gain control can be either of the balanced or unbalanced type.

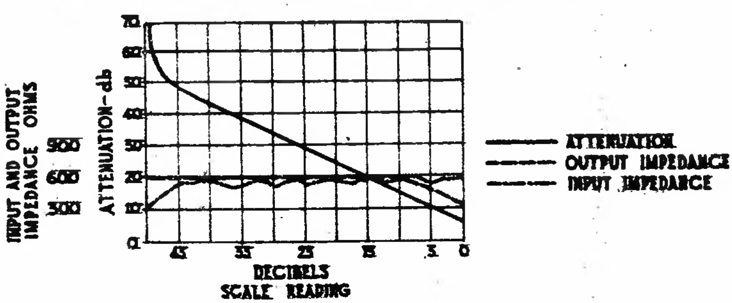
Faders and Gain Controls. Faders and Gain Controls are generally of the "ladder type," and a circuit is given in Fig. 30 of one form of this type. One variable only is needed; thus one arm with a collector ring is all that is needed for control purposes. As a result, it is possible to reduce the size to small dimensions, which enables several faders to be mounted in a relatively small space. Earlier types were 2-3/4 inches over-all diameter, but the midget type now available is only 2-1/4 inches in diameter.



UNBALANCED LADDER NETWORK.

FIG. 30.

The "ladder" type of network is not of the constant impedance type, and a typical impedance condition is given in Fig. 31. It will be seen that an appreciable variation takes place, depending on the attenuator setting. In addition, when the attenuator is at the zero position, a loss of 6 db still exists. This is the minimum insertion referred to earlier. This means that, whenever a ladder type network is wired into a circuit, a loss of 6 db is introduced.



IMPEDANCE AND ATTENUATION OF LADDER NETWORK - 600 OHMS.

FIG. 31.

The circuit given in Fig. 30 is only one form of this type of network, and, in other types, a shunt resistance may appear after every third or fourth series resistance.

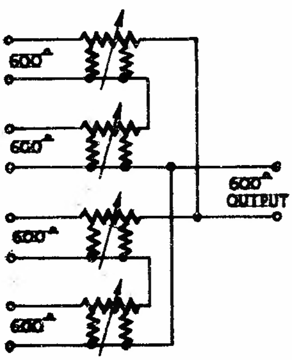
As shown earlier, it is possible to obtain a balanced form of the unbalanced attenuators. This also applies to ladder networks, and the balanced master gain control is now available. Physically, master gain controls have the same general make-up, but two units are mounted in tandem, that is, one behind the other, and the units are controlled with a single shaft and knob. The minimum insertion loss is 6 db.

Faders and gain controls have many steps to give a fine control of the attenuation, and, at the same time, to have as large a control as possible. The number of studs for midget types is either 28 or 31, and, in the latter type, the range is 45 db in steps of 1.66 db, and then larger steps are used to taper off to infinity or "off" position.

7.5 Combining Faders. In a broadcast studio, faders are connected in groups of 4 or 6, and the case of 4 faders will now be considered.

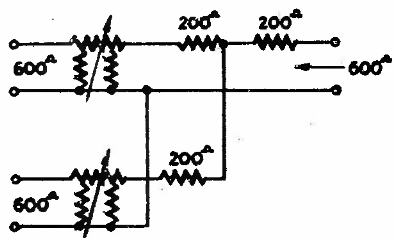
Fig. 32 shows the usual method; it will be seen that 4 unbalanced ladder networks are connected in series-parallel. This arrangement provides a correct impedance condition in each direction, and ignores the fact that the impedance of a ladder network is not constant. This series-parallel arrangement has an insertion loss of 12 db, 6 db due to the fader of any channel, and 6 db to the connections. When 2 or 6 faders are connected together, for instance, in parallel, the impedance conditions are not satisfactory, and it is necessary to use building out resistances.

Fig. 33 shows the arrangement for two 600 ohm faders connected in parallel and arranged to have correct impedance conditions. The minimum insertion loss is now 12 db.



SERIES-PARALLEL CONNECTED FADER.

FIG. 32.

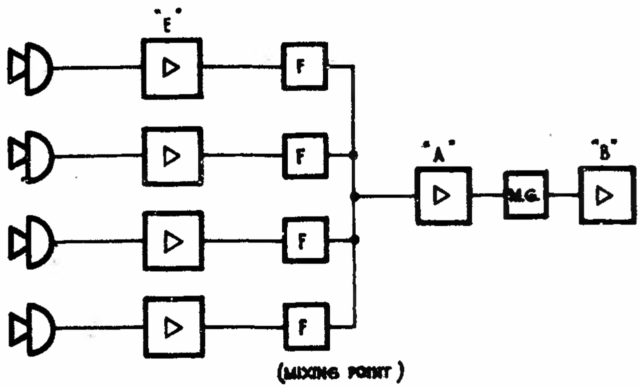


PARALLEL CONNECTED FADERS.
BUILT OUT TO 600 OHMS.

FIG. 33.

It will be seen that faders, when connected together, introduce a relatively high insertion loss, and it is necessary that the associated amplifiers have sufficient gain to compensate for this loss.

7.6 Mixing Levels. Mixing may be carried out prior to amplification, in which case it is termed "low-level" mixing. If mixing is accomplished after amplification, it is termed "high-level" mixing. The terms refer to the relative power of the audio-frequency signals at the mixing position. In Fig. 28, mixing takes place prior to amplification, and Fig. 34 shows a high-level mixing circuit, the amplifiers "E" being between the microphones and the mixing circuits.



4 CHANNEL FADER UNIT
WITH MASTER GAIN CONTROL.

FIG. 34.

Low level mixing has the following disadvantages -

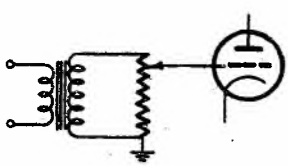
(i) As the microphone level is reduced by the amount of insertion loss of the faders, the signal to noise ratio will be reduced by the amount of this insertion loss, which can be as high as 12 db for 4 faders.

(ii) As the level at the faders is very low, the faders require to be well maintained to avoid contact noise.

It will now be seen why high level mixing is the preferred method of using faders.

Regarding the signal to noise ratio referred to, the ratio that exists at the input of the pre-amplifier is the signal to noise ratio for the system as a whole and cannot be improved at any other point. This is due entirely to the low output of the microphones used in studios. System, in this case, refers to the over-all circuit from the microphone to the transmitter output.

7.7 Potentiometers. In Fig. 35 a typical potentiometer to control the amplifier gain is shown. Radio receiver type of volume controls are not used; usually the physical make-up of the potentiometer used is the same as that used for faders. Balanced potentiometers are used for push-pull amplifiers.



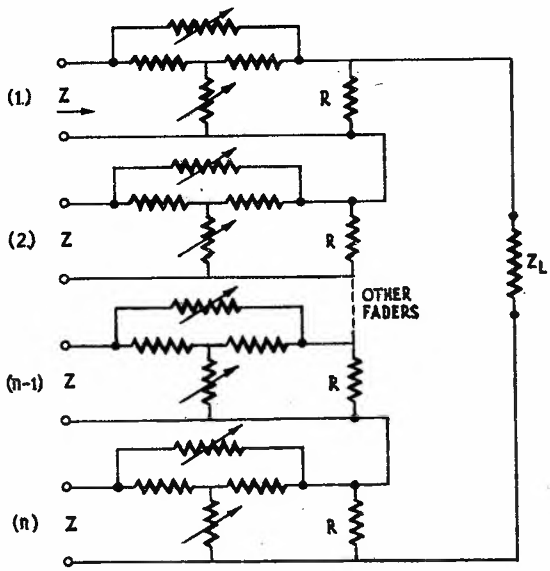
POTENTIOMETER TYPE GAIN CONTROL.

FIG. 35.

In some cases, only a small range of control is needed, for instance, in "D" or branching amplifiers. For these cases, a Yaxley or similar type switch is used. It should be noted that potentiometers are mounted directly on the amplifier as they have a high impedance. On the other hand, where a control is required away from the amplifiers, a low impedance type must be used.

/ Types

Types of Resistances Used. Attenuators, generally, are precision instruments, and the resistances are non-inductively wound to close tolerances as low as 0.25 per cent. On the other hand, faders and gain controls may have wire wound resistances, hand wound directly on to pins of the main base, and commercial type resistances of the 1/2 watt type are now being used extensively. This is possible as faders and gain controls are not of the precision type. It should be noted here that values of resistances below about 250 ohms are usually wire wound, not made of carbon. This type of resistance is also used for potentiometers.



TYPICAL SERIES MIXER CIRCUIT.

FIG. 36.

7.8 Design of Faders. Some points of interest on the design of combining faders are given here. Three cases will be considered.

(i) Series mixer circuit.
(ii) Parallel mixer circuit.
(iii) Series-parallel mixer circuit.

(i) Series Mixer. A typical series mixer circuit is given in Fig. 36, using bridged "T" faders.

Consider n mixer channels, each having an input impedance Z, each shunted with a resistor R and supplying a common load Z_L. It is assumed that constant resistance attenuators are used for faders. Although bridged "T" attenuators are shown, the use of other types will not affect the results, providing these other types are designed for the same impedance conditions.

Assuming Z and n are known then -

$$R = Z \frac{n}{n - 1}$$

$$Z_L = Z \frac{n^2}{2n - 1}$$

Insertion loss (each channel) = $10 \log_{10} (2n - 1)$ in decibel

or, if Z_L and n are known -

$$R = Z_L \frac{2n - 1}{n(n - 1)}$$

$$Z = Z_L \frac{2n - 1}{n^2}$$

Insertion loss as before.

/ Consider

Consider for an example a six-channel system working into 600 ohms, that is, $n = 6$ and $Z_L = 600$. Find Z, R and insertion loss.

$$R = Z_L \frac{2n - 1}{n(n - 1)} = 600 \times \frac{11}{30} = 220 \text{ ohms.}$$

$$Z = Z_L \cdot \frac{2n - 1}{n^2} = 600 \times \frac{11}{36} = 183 \text{ ohms.}$$

$$\text{Loss} = 10 \log_{10} (2n - 1) = 10 \log 11 = \underline{10.4 \text{ decibel}}.$$

Thus, the input circuits would require matching transformers unless their impedance was 180 ohms.

Let Z = 600 ohms, others as before.

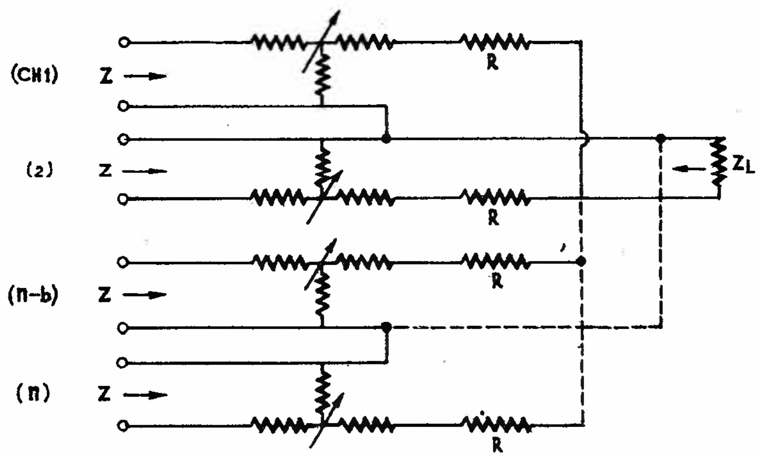
$$R = 600 \times \frac{6}{5} = 720 \text{ ohms.}$$

$$\text{Loss} = 10.4 \text{ decibel.}$$

$$Z_L = 600 \times \frac{36}{11} = 1{,}960 \text{ ohms.}$$

In this case, the output impedance is 1,960 ohms, and would need matching to the amplifier impedance (or whatever equipment following).

(ii) Parallel Mixer. Fig. 37 shows a parallel mixing circuit.



PARALLEL MIXING CIRCUIT.

FIG. 37.

This is a resistance compensated parallel type of mixer of n channels using constant resistance attenuators. Type "T" is shown, but no restriction as to type need be made here either. In this circuit -

$$R = Z \frac{n - 1}{n} \text{ and } Z_L = Z \frac{2n - 1}{n^2} \text{ when n and Z are known}$$

$$\text{or } R = Z_L \frac{n(n - 1)}{2n - 1} \text{ and } Z = Z_L \frac{n^2}{2n - 1} \text{ when n and } Z_L \text{ are known.}$$

The insertion loss is as in the series case.

/ Consider

Consider the same case as the previous example (n = 6) -

$$n = 6$$

$$\text{and } Z = 600 \text{ ohms (input impedance)}$$

$$\text{then } R = 600 \times \frac{5}{6} = 500 \text{ ohms}$$

$$\text{and } Z_L = 600 \times \frac{11}{36} = 183 \text{ ohms.}$$

$$\underline{\text{OR}}\ n = 6$$

$$\text{and } Z_L = 600 \text{ ohms (output impedance)}$$

$$\text{then } R = 600 \times \frac{30}{11} = 1{,}636 \text{ ohms}$$

$$\text{and } Z = 600 \times \frac{36}{11} = 1{,}960 \text{ ohms.}$$

Thus, impedance matching of input or output circuits would be required as in the series case.

(iii) Series-Parallel Mixer. A simple case only of the series-parallel case will be considered, say, for example, 16 faders arranged as a series-parallel mixer as shown in Fig. 38, 4 in series and 4 in parallel, that is, 4 paralleled groups of 4 in series.

In this example, let Z of each channel be 600 ohms and also the load impedance Z_L. The formulae for finding the compensating resistances R_s and R_p and the insertion loss for each channel are as follows -

$$\text{Series compensating resistance } R_s = Z\left(\frac{N_s}{N_s - 1}\right)$$

$$\text{Parallel compensating resistance } R_p = Z_L\left(\frac{N_p(N_p - 1)}{2N_p - 1}\right)$$

$$\text{Loss in decibel (each channel)} = 20 \log_{10} (2N_s - 1)$$

$$= 20 \log_{10} (2N_p - 1)$$

$$\text{where } N_s = \text{Number in series} = 4$$

$$N_p = \text{Number in parallel} = 4$$

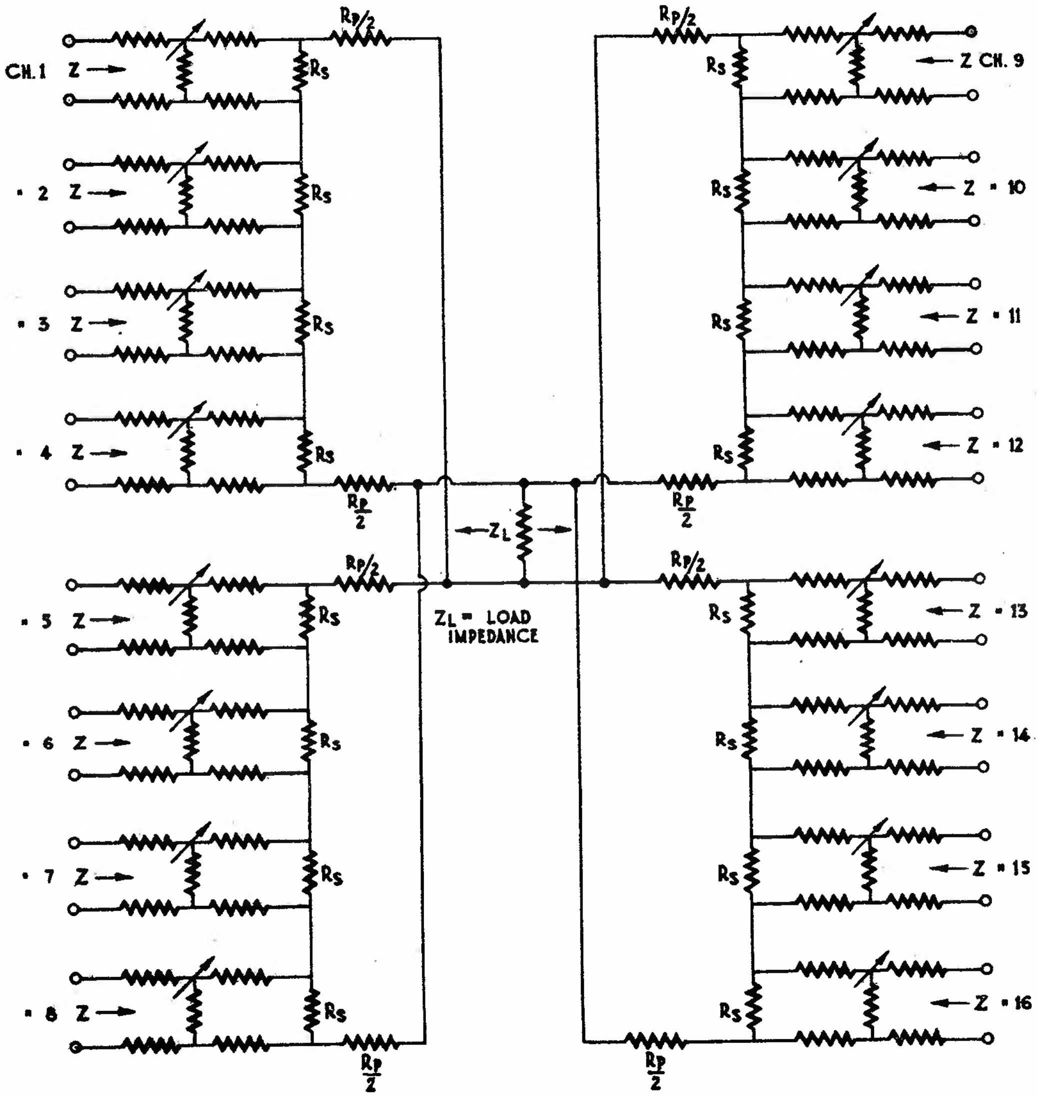
$$Z = \text{Channel input impedance} = 600 \text{ ohms}$$

$$\text{and } Z_L = \text{Load impedance} = 600 \text{ ohms}$$

$$\text{then } R_s = 600 \times \frac{4}{3} = \underline{\underline{800 \text{ ohms}}}$$

$$R_p = 600 \times \frac{12}{7} = \underline{\underline{1{,}028 \text{ ohms}}}$$

$$\text{Decibel} = 20 \log 7 = \underline{\underline{16.9 \text{ decibel.}}}$$



SERIES-PARALLEL ARRANGEMENT OF 16 FADERS.

FIG. 38.

8. TEST QUESTIONS.

1. What is an attenuator; state its desirable characteristics?

2. Sketch the essential parts of a magnetic type recording head. What are the more important points to be observed in making a recording?

3. What is meant by -

   (i) Constant velocity recording?
   (ii) Constant amplitude recording?

   Which of these methods is in more general use and why?

4. What is the main purpose for the use of each of the following in a Broadcast system -

   (i) Fader.
   (ii) Gain control, and
   (iii) Mixer unit.

   Give a block diagram showing typical use.

5. What does the term "tracking error" refer to as associated with disc recording and reproducing? What is its main effect?

6. What method is used to improve the signal-noise ratio at the high frequency end of recording and reproducing systems?

7. Name, describe and compare two commonly used methods of recording programme items.

8. What are the essential requirements of a reproducer? Describe briefly, with a sketch, one type of pick-up.

REFERENCES.

For additional information on recording and reproducing sound see the books -

"Telecommunication Journal of Australia," - Volume 6, No.5, Page 280.
"Telecommunication Journal of Australia," - Volume 7, No.1, Page 45.
"Telecommunication Journal of Australia," - Volume 7, No.3, Page 135.
"Radio Engineering." F. E. Terman.
"Elements of Sound Recording." J. G. Frayne and Halley.
"Recording and Reproduction of Sound." O. Read.
"Magnetic Recording." S. J. Begun.
"Sound Reproduction." G. A. Briggs.
"Science and Music." Sir James Jeans.
"Elements of Acoustical Engineering." H. V. Olson.

and the sound motion films -

"Fundamentals of Acoustics."
"Sound Waves and Their Sources."

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

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Postmaster-General's Department,
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COURSE OF TECHNICAL INSTRUCTION.

CONTENTS:

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1. INTRODUCTION.

1.1 This Paper covers the requirements associated with the origination, control and distribution of Radio Programmes insofar as the technical aspects are concerned. Each broadcasting organisation of course, has its own particular set up and type of equipment, but similar general principles apply in each case. The principal elements of a Broadcasting System are tabulated below, and Fig. 1 shows a typical broadcasting System in block diagrammatic form. The elements of a broadcasting system are -

(1) MICROPHONES -

(a) Studio.
(b) Remote Pick-ups.

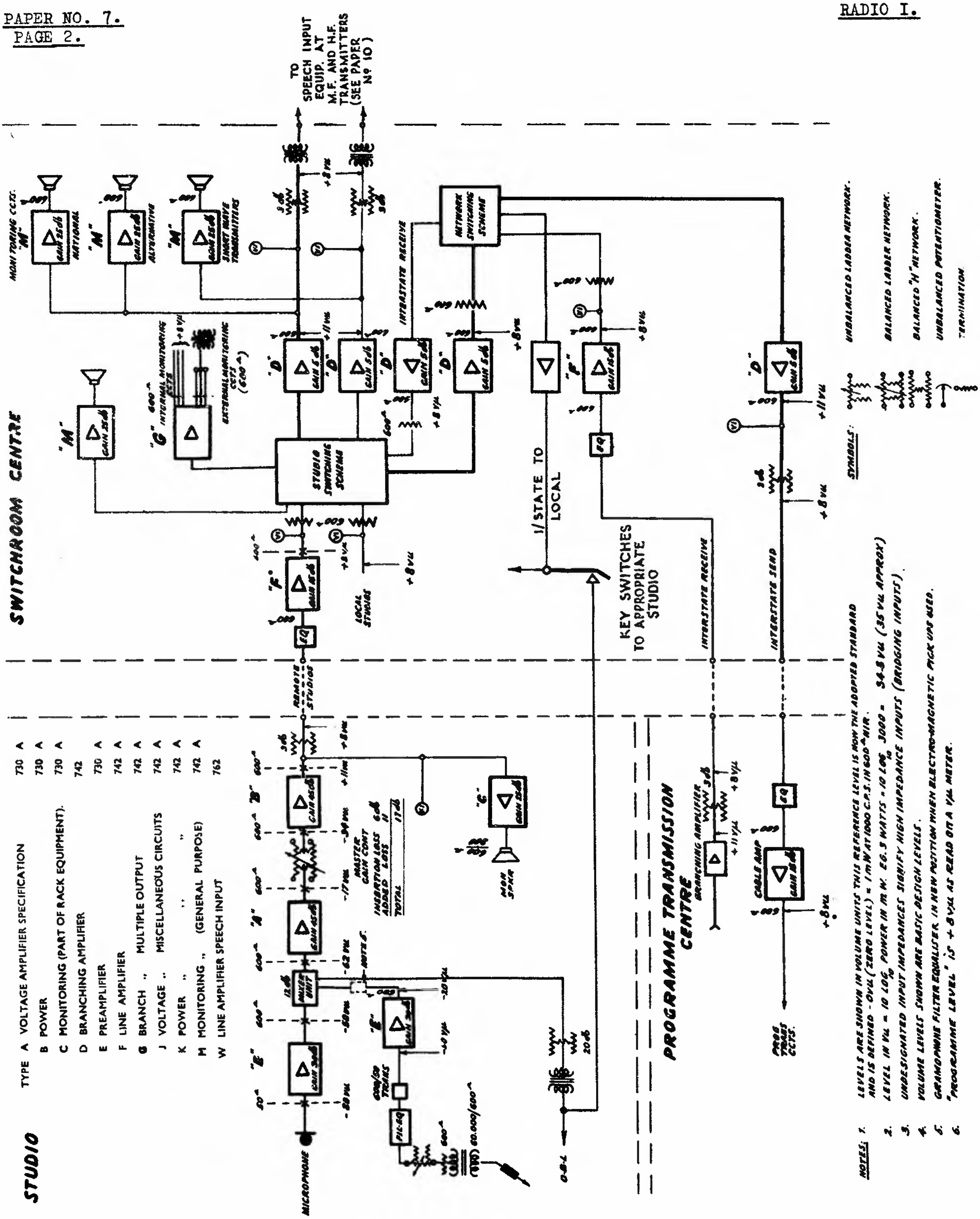
(2) APPARATUS FOR CONTROLLING AND CONVEYING MICROPHONE OUTPUT -

(a) Studio Control Booth -

(i) Preliminary amplifier (pre-amp.).
(ii) Microphone mixers.
(iii) Studio amplifier.
(iv) Volume controls or faders.
(v) Volume indicator.
(vi) Monitoring speaker.

(b) Remote Pick-ups -

(i) Preliminary amplifier.
(ii) Volume controls or faders.
(iii) Volume indicator.
(iv) Monitoring equipment.
(v) Intercommunication (radio, telephone or land line).



TYPICAL BROADCASTING SYSTEM.

(National Broadcasting Service Overall Transmission Circuit.)

FIG. 1.

(3) MASTER CONTROL ROOM APPARATUS -

(a) Volume Controls.
(b) Studio Amplifiers.
(c) Relays and Switching Apparatus.
(d) Network Channel Amplifiers.
(e) Volume Indicator.
(f) Monitoring Facilities.

(4) TELEPHONE FACILITIES TO LOCAL AND DISTANT TRANSMITTING STATIONS IN THE NETWORKS.

(5) RADIO TRANSMITTER -

(a) Line Amplifier or Limiting Amplifier.
(b) Volume Controls.
(c) Volume Indicators.
(d) Radio Transmitter.
(e) Monitoring Facilities -

(i) Monitoring rectifier and speaker.
(ii) Modulation-percentage indicator.
(iii) Carrier frequency monitor.
(iv) Modulation monitor (noise and distortion measurements).

(f) Aerial and Coupling Line.

1.2 By omitting duplication of items, the foregoing list may be rewritten as follows -

Microphones.
Loud-speakers.
Faders and Gain Controls.
Mixers.
Volume Indicators.
Amplifiers.
Switching Schemes.
Programme Transmission Lines to Stations and Centres.
Studios.

Reference to this list will show that microphones, loud-speakers, faders, gain controls and mixers have been described earlier, and the main principles of amplifiers have also been discussed. The other items are dealt with in the order in which they appear in the above list.

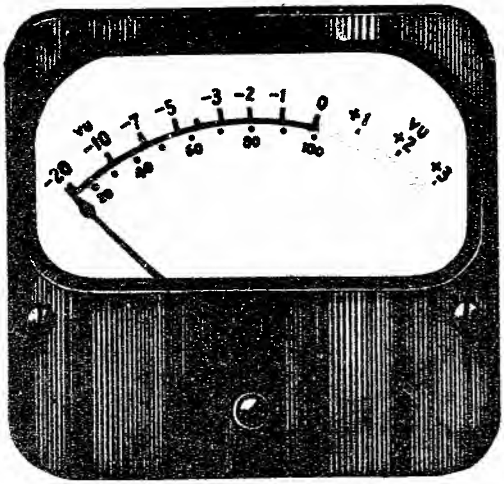
2. VOLUME INDICATOR.

2.1 The volume level of an audio signal at any particular point in a Broadcasting System is normally measured by means of a standardised instrument called a "Volume Indicator". The volume indicator is a direct current instrument fitted with a full wave instrument rectifier.

The Weston High Speed Level Indicator was the recognised standard instrument but was not satisfactory, as the rapid movement of the pointer was fatiguing to the eye. In addition, on account of the human element, accurate readings were not always obtained. This limitation led to the development of a Volume Indicator, particularly designed for the requirements of broadcasting and the telephone plant which provides the inter-connecting service among broadcast stations. The Volume Indicator gives satisfactory correlation of measurements under normal conditions of operation, and permits rapid and accurate readings.

/ The

The Volume Indicator consists of at least two parts -



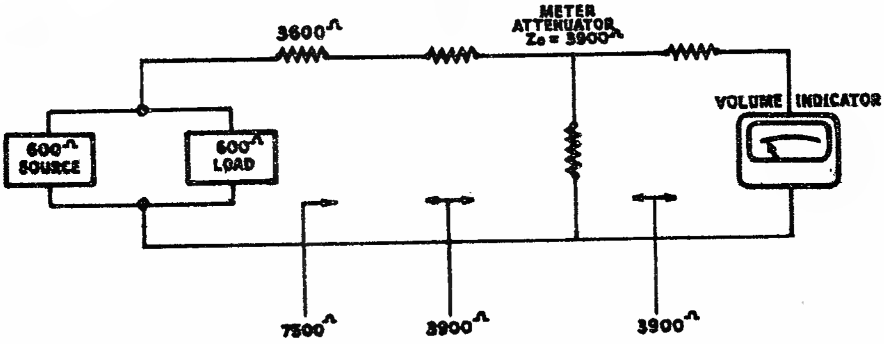
VOLUME INDICATOR.

FIG. 2.

(i) An indicating instrument (see Fig. 2), and
(ii) An attenuator (see Fig. 3).

As stated previously, it is used for the indication of volume, and has become generally known as a "VU" Meter. (Vu is pronounced "vee-you" and is customarily written with the lower case letters.) This word is used for the numerical expression of volume. The volume in vu is numerically equal to the relative strength of the waves in question in decibels above or below "reference volume".

The term vu should not be used to express results of measurements of complex waves made with devices having characteristics differing from those of the standard Volume Indicator.



VOLUME INDICATOR AND ASSOCIATED ATTENUATION NETWORK.

FIG. 3.

2.2 Terms Used.

Volume. This term applies to the strength of speech and music waves and is the reading given by the Volume Indicator. The Indicator has specific dynamic and other characteristics, and is calibrated and read in a prescribed manner.

Reference Volume. This is the base of the system of measurement of volume and is the scale reading for a steady 1,000 c/s wave of 1 milliwatt power into 600 ohms. (See paragraph 2.3 also.)

Reference Deflection. This is the deflection to the scale point at or near which the instrument is intended normally to be read.

/ Dynamic

Dynamic Characteristic. If a single frequency sinusoidal voltage between 35 and 10,000 c/s and of such amplitude as to give reference deflection under steady state conditions is suddenly applied, the instrument pointer will reach 99 per cent. of reference deflection in 0.3 second, $\pm$ 1.0 per cent., and will then overswing reference deflection by at least 1.0 per cent., and not more than 1.5 per cent. The time required for the instrument pointer to reach its position of rest on the removal of the voltage is approximately equal to the time of response.

Response Versus Frequency Characteristic. The sensitivity of the volume indicator will not depart from that at 1,000 c/s by more than 0.2 db between 35 and 10,000 c/s, or more than 0.5 db between 25 and 16,000 c/s.

2.3 Details. A correctly calibrated volume indicator, with its attenuator set at zero vu, will give reference deflection when connected to a source of single frequency sinusoidal voltage adjusted to develop 1 milliwatt in a resistance of 600 ohms, or with the attenuator set at n when the calibrating power is n db above 1 mW.

It should be noted here that the volume indicators available at present will not read 0 vu when 1 mW is developed in 600 ohms, and, in fact, the instrument will only read -4 vu. Therefore, to obtain a reading of 0 vu, a power 4 db above 1 milliwatt must be developed in the resistance. These remarks only apply provided the attenuator is out of circuit.

Method of Reading. The reading is determined by the greatest deflections occurring in a period of about a minute for programme waves, or a shorter period (for example, 5 to 10 seconds) for message telephone speech waves, excluding not more than one or two occasional deflections of unusual amplitude.

Impedance. The volume indicator is normally used as a bridging instrument, and, when so used, its impedance must be sufficiently high so as not to influence unduly the circuit with which it is used. The impedance must not be less than 7,500 ohms for use on a 600 ohm circuit.

Harmonic Distortion. When the volume indicator is connected to a simple resistive circuit through which a sinusoidal wave (between 25 and 8,000 c/s) is being transmitted, the root-mean-square sum of the harmonics produced will not exceed 0.2 per cent.

Ability to Withstand Overload. Because of the great variation in amplitude which this indicator may register, it has greater ability to stand overloads than average instruments. The volume indicator is able to withstand, without injury or effect on calibration, a momentary overload of ten times the voltage corresponding to reference deflection, and a continuous overload of five times that voltage.

Scale. Two scales are provided, one calibrated in vu and the other in Percentage Modulation. The latter scale is of particular use when the instrument is installed at radio transmitters.

Connections. The connections for the volume indicator are shown in Fig. 3. It will be seen that the attenuator has a characteristic impedance of 3,900 ohms.

2.4 Advantages of the Volume Indicator. The advantages of the Volume Indicator include -

(i) vu indications.
(ii) Controlled dynamic range.
(iii) Increased scale length.
(iv) Improved legibility (scale).
(v) Less eye strain and fatigue (pointer action).

In addition, the scale card, which is a special buff colour with a minimum number of markings, reduces eye-strain and fatigue.

/ When

When the volume indicator is adjusted to read reference level, the programme peaks are suppressed by approximately 8 db. This means that the peak power is approximately 8 db above the volume indicator reading.

2.5 Until recently, it was customary practice to measure the power level of the programme by means of a decibel meter, which is really an alternating current voltmeter calibrated to read 0 decibel when placed across a 600 ohm load carrying 0.006 watt. This corresponds to a voltage of about 1.73 volts, the normal understood reference level (when speaking of power in decibel) being 0.006 watt. Problems involving volume units are solved exactly as any decibel problems, inasmuch as a change of any number of vu's corresponds to a change of a similar number of decibel. <u>If 0 decibel is taken as 0.001 watt in 600 ohms, the terms vu and decibel may be substituted in any problem.</u>

3. <u>AMPLIFIERS.</u>

3.1 The principles of audio-frequency amplifiers were given in Paper No. 4, but the general forms of amplifiers as standardised for broadcasting use in Australia will be described. The alphabetical designations are merely arbitrary and serve as a quick reference guide. (See Fig. 1.)

3.2 <u>"E" Amplifiers.</u> "E" Amplifier lifts the level of the microphone to a satisfactory level to permit fading (that is, mixing). The number of "E" amplifiers in a control booth can be 4 or 6, followed by an equal number of faders. (The "E" amplifier is also termed "Pre-amplifier".) The gain averages 30 db, and the output power 10 mW.

<u>"A" Amplifier.</u> "A" Amplifier lifts the programme to a satisfactory level for the master gain control. This control is used to maintain a satisfactory level to the transmitters and to other States. Average gain 45 db, output power 10 mW.

<u>"B" Amplifier.</u> "B" Amplifier is a power amplifier which supplies the power to the line. The 3 db pad is used to smooth out variable impedance conditions which may exist in the cable circuits between remote studios and the switchroom. Average gain 45 db.

<u>"C" Amplifier.</u> "C" Amplifier bridges the output of the "B" amplifier and supplies power to the monitoring loud-speaker in the control booth. A similar unit is in the studio but is equipped with an "M" type amplifier. Average gain 25 db.

<u>"M" Amplifier.</u> "M" Amplifier is the monitoring amplifier and is normally fitted in the bottom of the loud-speaker cabinets. Average gain 25 db, output 3 watts.

<u>"F" Amplifier.</u> "F" Amplifier is used to lift the programme due to the equalised line between the remote studio and the switchroom. Average gain 15 db, output 120 mW.

<u>"G" Amplifier.</u> "G" Amplifiers serve to distribute the programme to the many points where monitoring facilities are required, that is, A.B.C. Offices, P.M.G. Offices, and so on. Each amplifier is of the bridging type and has six outputs. Gain 5 db, 6 channels output each 120 mW.

<u>"D" Amplifier.</u> "D" Amplifiers are of the bridging type, with a gain of 5 db. There is one "D" amplifier per channel for distributing the programme to the transmitters and network switching scheme. These amplifiers are also used between the studio switching scheme and the network switching scheme. Output 240 mW.

<u>Branching Amplifier.</u> "Branching Amplifier" is similar to the "D" amplifier, but is direct current operated. The Branching Amplifier is used to distribute the programme from programme channels to as many points as required. It has greatest use for "all station" broadcasts. Gain 5 db.

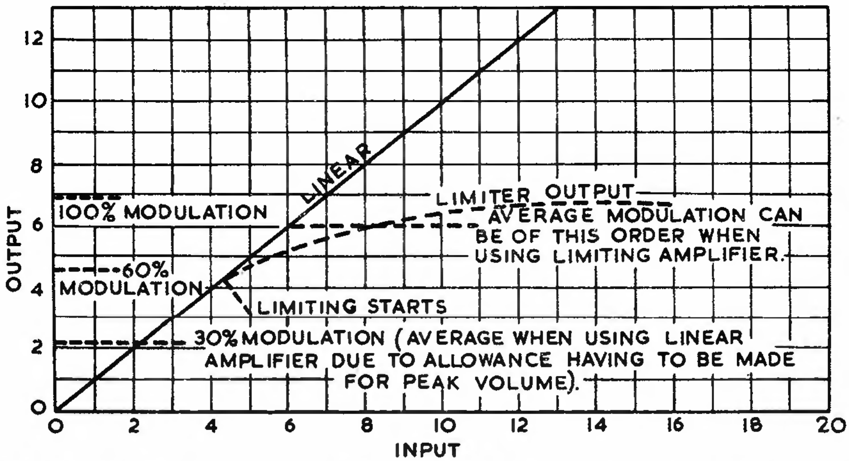
/ Cable

Cable Amplifier. "Cable Amplifiers" compensate for the losses introduced by the equalised cable pairs between the Network Switching Centre and Programme Room. This amplifier is also direct current operated. Average gain 15 db, output 120 MW.

"W" Amplifier. "W" Amplifier is the line amplifier to compensate for the loss introduced by the equalised permanent line to the transmitter. In addition, this amplifier has certain characteristics which suppress peaks above a certain pre-determined level to prevent overloading the transmitter. The "W" amplifier also permits the use of a high level into the transmitter, depending on the degree of limiting used. Gain 60 db, output 240 mW.

The "W" amplifier has an important application in the operation of high frequency transmitters, where it is desirable to keep the average modulation depth as high as possible. In medium-wave stations, the average modulation depth is 30 per cent., but, in the areas served by these stations, the signal is usually strong when compared with a short-wave signal. By increasing the average modulation content of the signal, a better signal-to-noise ratio is obtained at the receiver.

The amplifier is really an audio-frequency amplifier with automatic-gain-control. The degree of limiting or compression is adjustable to suit local requirements. Fig. 4 shows this action schematically.



SHOWING THE IMPROVEMENT POSSIBLE IN AVERAGE MODULATION DEPTH WHEN USING A LIMITING AMPLIFIER.

FIG. 4.

A typical adjustment of the amplifier is as follows -

Tone at programme level (+8 vu) is fed into the limiting amplifier, and its gain is adjusted until the amplifier is definitely limiting at 90 per cent. modulation. This condition is indicated when an increase of input to the amplifier produces no increase in modulation depth. When this adjustment has been made, the line level is increased by +6 db (by reducing the gain control 6 db) and the amplifier is said to possess 6 db of limiting. The provision of 3 db of limiting is equivalent to doubling the transmitter power.

4. AMPLIFIER NOTES.

4.1 Modern studio amplifier systems are designed to work with low-level microphones, such as the Dynamic, Crystal and Velocity types, and special precautions must be taken to keep extraneous disturbances from entering the circuits. To prevent the picking up of stray electrostatic and magnetic fields by the microphone cable, it must be thoroughly shielded. The connecting plugs and sockets must have absolute and firm contact at all times in order to prevent noises due to variation of contact resistance. It is desirable to keep the direct current input resistance of a high-gain amplifier as low as possible to reduce thermal-agitation voltages.

4.2 Carbon resistors, owing to their construction, should never be used at points carrying a low programme level.

4.3 Valves for the input circuits should be carefully chosen as regards quiet operation and microphonic characteristics, and should be shock-mounted if there is a likelihood of acoustic shock from a near-by loud-speaker. Good non-oxidising valve sockets should be used.

4.4 A specially shielded and screened transformer should be used in the input circuits of amplifiers operating at low levels, such as the "E" type.

4.5 It is preferable to distribute gain over several amplifiers, if a relatively high gain is required, since stable operation is difficult to obtain in high-gain amplifiers.

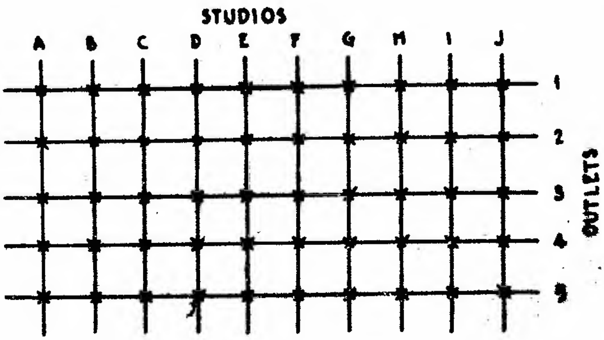
4.6 Wires carrying alternating currents, such as leads to power convertors, filaments, etc., should be kept away from low level wiring, which, in its turn, should be adequately screened and shielded.

5. STUDIO AND NETWORK SWITCHING.

5.1 Studio Switching. Broadcasting centres of necessity have more than one studio (because various types of programme items require different acoustics) and there are outside broadcasts also to be fed into the main system. It is necessary that any one or more of these programme originating points can be connected into the system as required. To provide this facility, Studio Switching was developed. Its position in the overall scheme can be seen by reference to Fig. 1.

The equipment used in this scheme consists of standard keys, jacks, lamps, etc., with telephone facilities and monitoring facilities to other studios as required. Circuit details are not given, since the scheme may be readily visualised.

5.2 Network Switching. Network Switching is situated at the Switchroom (as shown in Fig. 1) and is connected between the Studio Switching and the Programme Centre or Room. A number of links is provided between Network Switching and the Programme Room in each direction, so that incoming and outgoing programmes can be handled. . Network Switching can best be considered as the terminus for interstate programme channels, both send and receive.



SWITCHING SCHEME BASIC DESIGN.

FIG. 5.

5.3 Switching Principles. Basically, all switching is on the principle shown in Fig. 5. Assuming a switch or key at all the cross-over points (as indicated by "x" in Fig. 5), it will be seen that any studio or number of studios can be connected to any of the outlets, and the circuits can be so arranged that one group of studios can be on one outlet while another group is on another outlet.

/ In

In addition, a group of studios may be connected to more than one outlet, if necessary. The same basic method of connection is used for Network Switching.

In addition to the above, it is usual to have facilities so that the connections can be "Preset". It must be realised that most of the switching is done on a time basis, where even seconds are important, so that presetting enables an operator to "set up" in advance the connections for the next programme unit. The operation of a Master Key at the appropriate time completes the circuits.

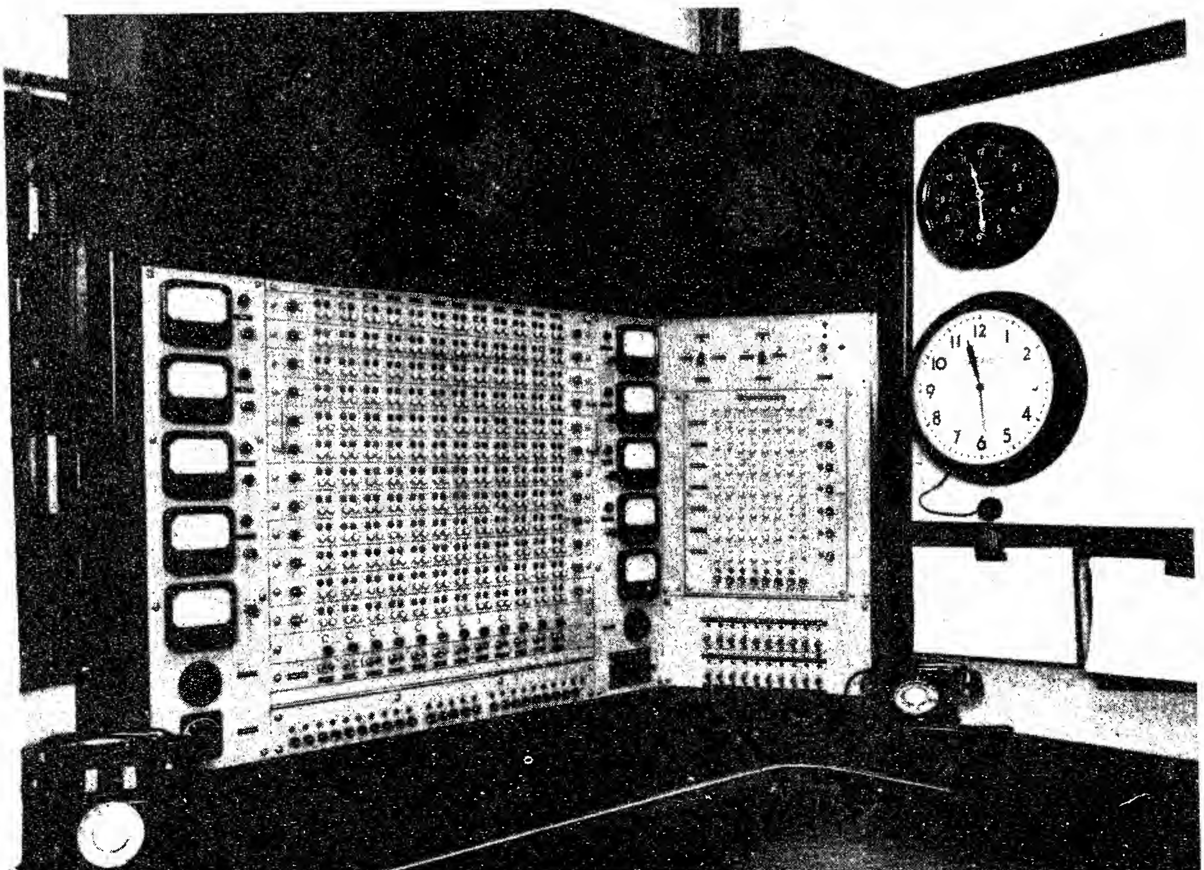
The Master Key appears on the Switchroom panel and also in the Control Booth and on the Announcer's Desk. This key enables the announcer to switch by remote control, and also provides two other points of operation to ensure that the switching is done at the correct place and time.

On the switching panels, the push-pull type of telephone key is used, as it requires little space and is safe in operation, that is, the accidental operation of other keys is less remote. However, lever type keys have been used in some schemes. Associated with each key are Indicating Lamps to indicate the circuit conditions, one colour to indicate the preset condition and another to show when the actual switching has been made.

Also on the panels are Monitoring Keys, so that a loud-speaker can be switched to any outlet. In addition, Volume Indicators are wired across each outlet as is shown in Fig. 1.

Two main methods of circuit selection are used, namely, relays and uniselectors, both of the telephone equipment type. It is usual, however, to use uniselectors to select the monitoring circuits.

A key-switching system and associated equipment is shown below.



N.B.S. STUDIOS ADELAIDE.
VIEW OF LOCAL AND INTERSTATE SWITCHING PANEL.

5.4 Programme Transmission Centre. The Programme Centre or Room is always in the same building as the Trunk Test Room and is the true terminal of all interstate and intrastate programme channels. As far as the National Broadcasting Service is concerned, these channels are extended to the Network Switching Centre for switching purposes.

Programme channels are now used by Commercial Stations also, and in much the same manner the channels are extended to the stations concerned. Insofar as technical performance is concerned, all Programme channels are under the control of the Programme Centre.

5.5 Transmitters or Stations. The transmitting stations are connected to the Studio Switching Centre by means of Permanent Programme Lines, and radiate the programme matter sent out on these lines by the Studio Switching Centre. The only switching that a transmitter can do is to change over from the permanent line to the emergency line when a fault condition occurs.

5.6 Typical Operations. A typical operation of the system shown in Fig. 1 is as follows -

Consider a programme originating from 2FC (Sydney) being broadcast by the station shown in Fig. 1, and which ceases at 8 p.m. when the National programme originates from 3AR (Melbourne). Before 8 p.m., the interstate channel from Sydney is connected from the Programme Room to Network Switching via a tie line, and is then extended through Studio Switching to the 3AR studio. From this point, the programme can be considered to have originated in that studio, for it is fed through the Control Booth back to Studio Switching and to the transmitters.

(It should be noted here that the interstate channels to Hobart and Adelaide are extended to the Network Switching Centre where the Sydney programme is connected to them.)

At 7.59 p.m., 2FC gongs the network off and, in the next minute, all States make local announcements but are monitoring the circuit from 3AR waiting for the theme music which introduces the next programme unit.

From what has been said above, Hobart and Adelaide are already connected to the Network Switching Centre, but Sydney, up to 7.59 p.m., is not. Prior to 7.59 p.m, a channel is extended from the Network Switching Centre to the Sydney Network Switching Centre, and from there to the Studio Switching Scheme as well as to Brisbane. It is seen now that all States are connected to the Melbourne Network Switching Centre, and can be connected to any Melbourne Studio through the Studio Switching Scheme.

The minute 7.59 p.m. to 8 p.m. is necessary, as the clocks in all States are not synchronised, and it also allows any one State to finish a local programme.

At 7.59 p.m., the Network Switching Centre is extended through to the 3AR studio, and all other national studios hear the 3AR announcements on monitoring headphones. When the theme is heard, other studios connect their transmitters to the network and the programme now commences.

From earlier information, it does not follow that the programme must originate at the 3AR studio. In fact, the programme can originate anywhere in Victoria and, in addition, any other studio can take control. However, for purposes of demonstration, it is assumed that the programme originated at the 3AR studio.

/ 6.

6. RADIO PROGRAMME TRANSMISSION.

6.1 The Broadcasting Network, as it is known today, has only been possible by the use of trunk lines, both intrastate and interstate, for the transmission of broadcast programmes. The National Network extends in Queensland as far north as Atherton and as far west as Geraldton in Western Australia.

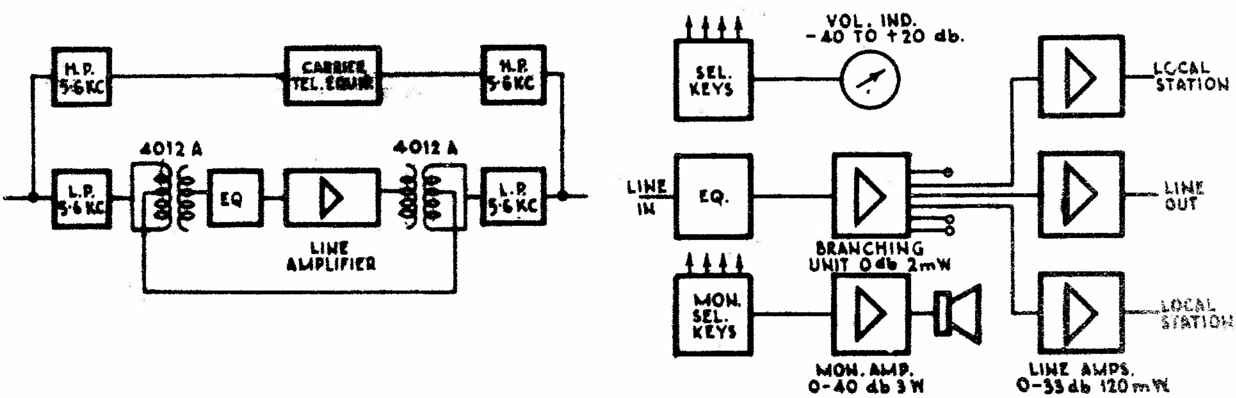
6.2 Programme Channels are of two types -

(i) Programme Physical Channels. Physical programme channels have a frequency response from 35 to 5,000 c/s, due to the type of 3 Channel Telephone Carrier Systems operating on the same pair of wires. In the near future. certain of the more important channels will be extended to 7,000 c/s.

(ii) Programme Carrier Systems. On the other hand, programme carrier systems have a frequency range from 35 to 7,500 c/s, but are being displaced by J12 Carrier Telephone Systems which have a frequency range that does not permit the use of both systems on the same trunk route.

6.3 On any trunk route, due to the attenuation caused by the copper wires, it is necessary to have an amplifier at about every 120 miles to restore the radio programme level to normal. These amplifiers are commonly known as line amplifiers or repeaters, and Fig. 6a shows a typical arrangement at a Repeater Station. It is seen that the line amplifier is only used for the amplification of the Radio Programme, and separate amplifying equipment is used for the Carrier Telephone Systems.

6.4 At Repeater Stations, provision is usually made for the programme, which normally passes through the repeater station, to be fed to local broadcast stations. A Branching Unit is used for this purpose and, as seen from Fig. 6b, this amplifier operates from a low level and has six outputs. In each output, as required, a line amplifier is patched in, and the programme is then at a satisfactory level to the next repeater station and also to the local broadcast stations. It is seen that an Equaliser is in the circuit before the branching unit. If a branching unit is not being used, the equaliser is fitted before the line amplifier. Every repeater section is fitted with an equaliser at the receiving end to restore the frequency response to a satisfactory condition.



(a) Amplifying the Programme.

(b) Branching the Programme.

REPEATER STATIONS.

FIG. 6.

6.5 At the Terminal Station, which on interstate routes is the Programme Room, the circuit arrangement is as shown in Fig. 7.

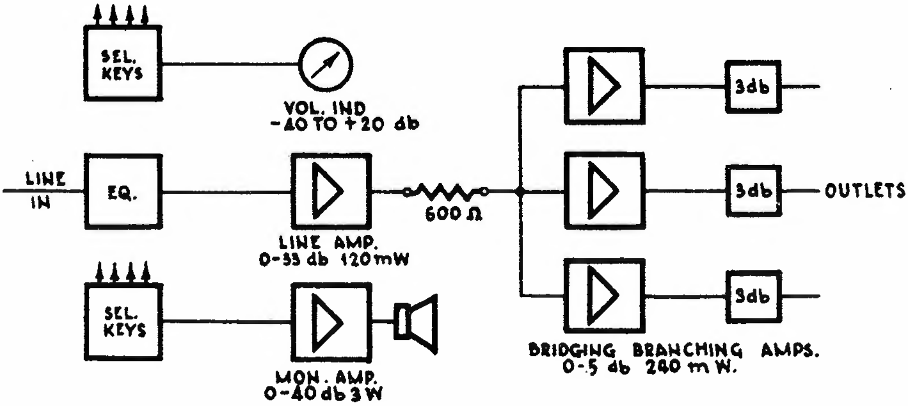


FIG. 7. TERMINAL INSTALLATION.

This is not the same arrangement as used at a Repeater Station. Although the line amplifier is the same, the splitting arrangement is not. In this case, Branching Amplifiers are used, one to each outlet, and are bridged across the programme channel as required. In addition, a Monitoring Amplifier - alternating current operated with a gain of 40 db - is wired through appropriate switching keys so that any point may be monitored as desired.

6.6 Line-up Procedure. Before a Programme Channel can be used for broadcast purposes, it is necessary that it be in the correct direction, as these channels are unidirectional but can be reversed as desired. When the direction is correct, then a line up is made. For this purpose, a General Radio Noise and Distortion Set is used, and the line up is made between the terminals of the channel under test.

The line-up procedure is as follows -

(i) A 1,000 c/s tone is sent to the Programme Channel, and the line amplifier is adjusted at each repeater so that a level of +8 vu is sent to line.
(ii) A frequency response test is made at selected frequencies between 35 and 5,000 c/s with the aid of an audio frequency oscillator and level indicator.
(iii) A 400 c/s tone is sent to line by the transmitting terminal at a level of +14 vu, and the distortion is then measured at the receiving end.
(iv) The 400 c/s tone is then removed and the line terminated at the sending end.

The receiving end then measures the noise and the resultant figure is corrected by reducing it 6 db. This is necessary as the test was made at a level 6 db above programme level.

The channel is now ready for use and is extended at each terminal to the switchroom.

The amplifiers, except the monitoring amplifier, operate from the 24 and 130 volt repeater or terminal batteries. These amplifiers use radio receiving valves, 6J7G and 6V6 G/GT types, and a power output of 400 mW can be obtained. However, from a line amplifier only 120 mW is required, while from the branching amplifier a power of 240 mW is required due to a 3 db pad in the output. This pad assists to improve the impedance conditions of the cable pairs to the switchroom and B class stations.

As seen from Fig 1, a cable amplifier is used to compensate for the losses introduced by the equalised cable pairs from the switchroom.

6.7 Programme Levels and Power Output. The basic programme level is +8 vu (6.3 mW), but a level of 16 vu is required to accommodate peaks in programmes.

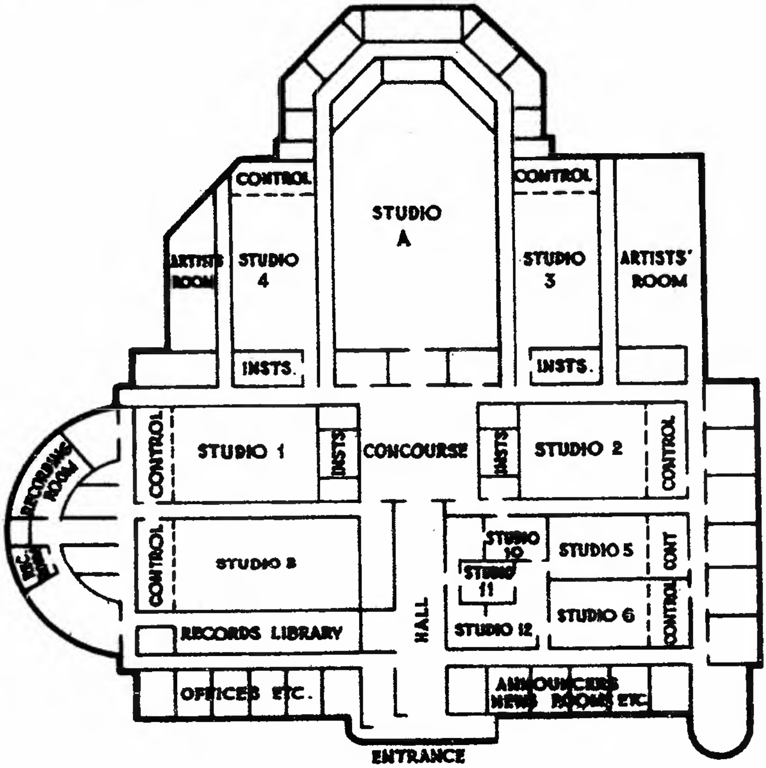
Since the vu meter is not designed to read peaks, this differenco in level is compensated by lining-up with an 8 db pad at transmitter. After completion of lining-up the pad is removed and the circuit is capable of taking normal level plus peaks to 16 vu. The power capability of the amplifier should be at least 16 vu (40 mW) but to allow a reasonable margin, amplifiers of 120 mW output are used. / Where

Where a 3 db pad is to be fitted after the amplifier, the power capability must be increased to 240 mW. Typical factors, for which the marginal allowance in power output is made, are -

(i) Ageing of valves.
(ii) Reduction of distortion.
(iii) Various impedance conditions.

7. RADIO STUDIOS.

7.1 General. As stated previously, one studio is seldom sufficient for a Broadcasting Station. The variety of programmes handled calls for a number of studios of various types, such a collection usually being termed a Studio Group. Fig. 8 shows a floor plan of a typical studio group, including studios of various sizes, control booths, record library and recording rooms.



TYPICAL FLOOR PLAN OF STUDIO BLOCK.

Studio A for Large Symphony Orchestras.
Studio B for Dance Bands and Small Orchestras.
Studios 1-4 for Dance Bands and Choirs.
Studios 5-6 for Smaller Bands, Piano Solos and Plays.
Studios 10-12 for Talks and News.

FIG. 8.

The studio group comprises not only a number of rooms specially allocated for the purpose of being the originating point of a programme, but must be considered as including permanent and temporary outside sources, such as Town Halls, Churches and Sports Grounds.

/ Within

Within the group, one or more programmes may be proceeding at the same time, together with rehearsals, auditions and recordings. Obviously, such a complex arrangement requires a central point of control. In some broadcasting systems, this central point is known as the Main Announcers' Studio or simply Main Studio; in other systems, the control is with the combined Network and Studio Switching Systems. The technical equipment of a studio must provide for the setting up of at least two broadcasting channels, which are directed from any studio to the main studio and to other outlets, as discussed under switching schemes. The more important functions of studios will now be referred to in a little greater detail.

7.2 Announcers' Studio. The Announcers' Studio has the following functions -

(i) Receives the interstate programmes and controls the regional switching.
(ii) Selects and mixes one of two microphones (announcer and speaker) and the output of two gramophone pick-ups.
(iii) Transmits programmes originating in the studio block to the network.
(iv) Effects the switching of outlets in association with the control operator.
(v) Switches outside lines, time signals, gongs, etc., into the programme.
(vi) Monitors programmes, loud-speakers or headphones, and gives necessary cues to speakers and performers in the studios.
(vii) Requires intercommunication facilities with control booths, switch-room, studio, P.B.X., etc.

TYPICAL ANNOUNCER'S DESK.

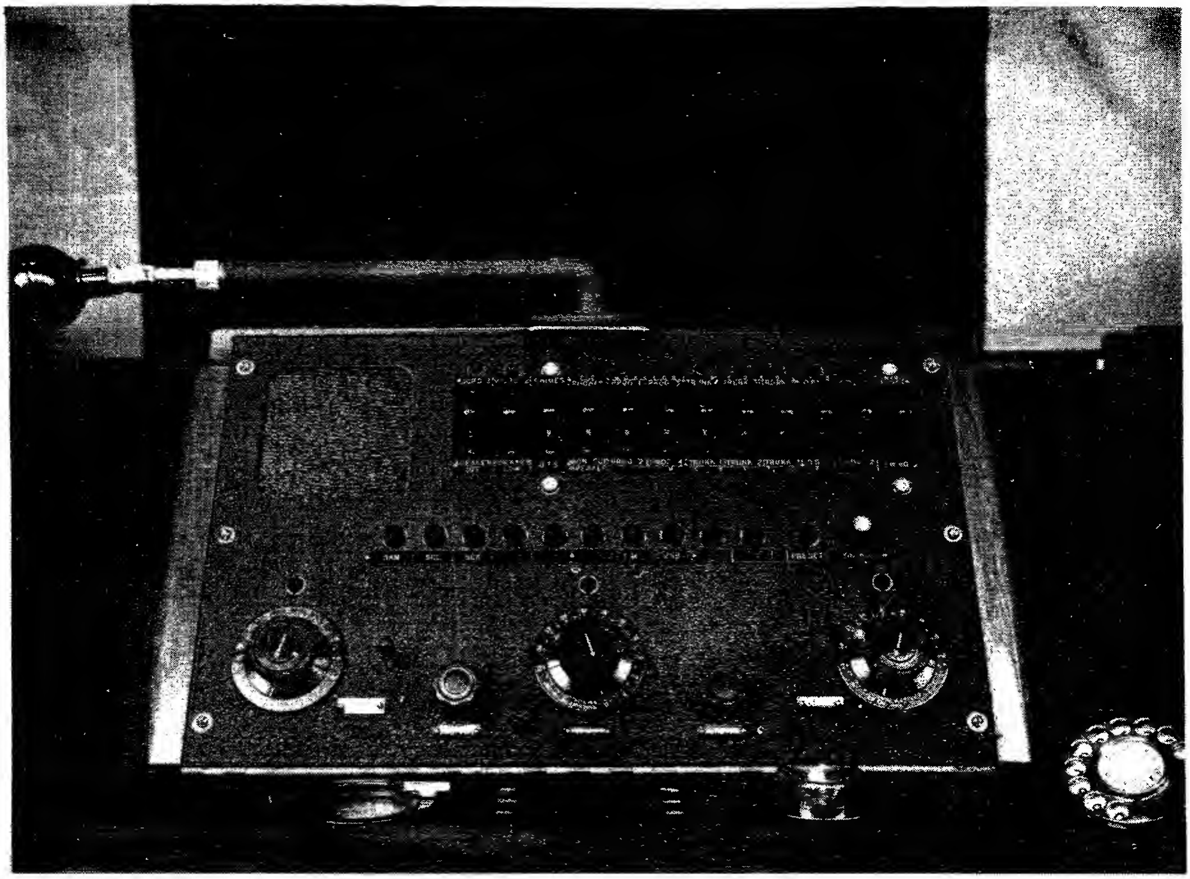
The majority of programmes originate in studios other than the main studio, but the control is still with the main studio. In addition, outside broadcasts, such as those from Sports Grounds and Churches, are controlled in this studio.

The term "Control", as used above, refers to the switching.

The control of programme levels is a function of the Control Booth associated with this studio.

A typical studio includes the Announcer's Desk, the Speaker's Desk, Monitoring Loud-speakers and Microphones. The Announcer's Desk includes two gramophone machines with motors remotely controlled by keys mounted on the desk. Keys are provided also to switch the microphones (including the microphone on the speaker's desk) and to select monitoring circuits. The Master Switching Key is also included on this desk. In addition, the usual indicating lamps are provided, one group of which indicates to the announcer the stations connected to that studio.

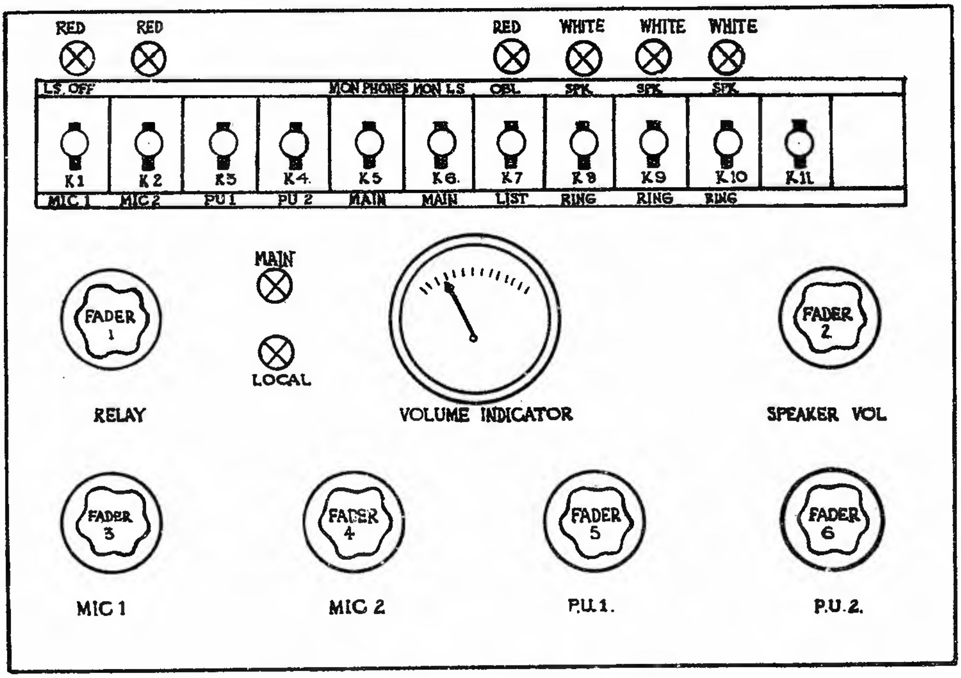
Certain other facilities are provided in both the control booth and studio. These facilities include telephone circuits between studio, control booth and switchroom, as well as a "speakback" microphone in the control booth, so that the operator or producer can speak to the studio over the studio monitoring loud-speaker.



N.B.S. STUDIOS, ADELAIDE.

ANNOUNCER'S DESK - CONTROL PANEL.

One important Guard Circuit is provided. This circuit provides that, during the time that the studio is broadcasting gramophone records or programme matter from some external point, the announcer can monitor the programme on the loud-speaker, but, the moment a studio microphone is to be used, the loud-speaker is automatically switched off. At times when the announcer is making local announcements while waiting to cross to the network, he will use monitoring headphones. Fig. 9 shows the panel of an Announcer's Desk, illustrating most of the above points.



NOTES.

Keys 1 and 2 connect microphones and associated faders into circuit. Appropriate red lamp lights.

Keys 3 and 4 connect gramophone pick-ups through faders 5 and 6

Key 5 permits monitoring of main or local programmes on headphones.

Key 6 permits monitoring of main or local programmes on speaker.

Key 7 (down) connects outside broadcast cue line to monitoring circuit.

Key 7 (up) connects outside broadcast line through.

Keys 8, 9, 10. Speaking and ringing facilities for local telephone lines, order wires, etc.

TYPICAL PANEL LAYOUT OF A REGIONAL TYPE ANNOUNCER'S DESK PANEL.

FIG. 9.

7.3 Additional Details. It is apparent that special precautions must be taken to prevent crosstalk and "singing" between circuits carrying the same programme but at different levels, for instance, the actual microphone circuit and the monitoring circuits. Low level circuits are wired in single lead-covered pairs, while other circuits are wired in lead-covered cables. Rubber covered shielded cable is used from the microphone to the connecting point, which may be on the skirting board or at a convenient point on the announcer's desk depending on the purpose of the microphone.

In a typical studio block, there is a vast quantity of telephone equipment, including relays, uniselectors and indicating lamps. It is necessary to provide power for this equipment, and it is usual to have a 24 volt battery operated on the floating principle associated with a Static Rectifier.

All amplifier equipment is alternating current operated from the commercial power supply, and it is usual to provide a standby engine with alternator fitted with a governor to ensure a stable frequency and voltage.

7.4 Outside Broadcast Lines. The outside broadcast lines terminate in the switchroom and can be extended to any studio as required. Permanent lines which have frequent use are fitted with equalisers to correct the frequency response, while other lines are equalised, when required, with a variable equaliser.

Facilities are provided at the terminating point of these lines in the switchroom, so that continuity and insulation can be measured whenever a line is required for broadcasting. Each line has a jack at the distant end, and this is so wired that the inner springs loop the line when it is not in use. This jack enables the continuity test to be made at any time. The insulation test between wires would be made with the outside broadcast operator at the distant end, but the insulation to earth can be made at any time. The loop resistance of each line is known, so that the loop test would detect shorts or partial shorts.

As it is usual to provide two lines to each point, one can be used as the programme line, while the other is the telephone circuit but can be used for broadcasting in case of failure of the programme line. (Further reference to these lines will be made later.)

7.5 Control Booth. On a typical Control Booth Table, the following equipment is to be seen -

(i) Electric clock,

(ii) Faders and master gain control.

(iii) Volume indicator.

(iv) Switching keys.

(v) Indicating lamps.

(vi) Jacks.

The purpose of the first three need not be discussed here. The switching keys provide for the monitoring of the various circuits, including the normal control booth output, outside broadcast lines and incoming lines from the switchroom. The keys also provide for the switching of these various circuits, including the time signal. The indicating lamps show the condition of the switched circuits. Two important keys are the key to extend the "preset" to the announcer's desk and the master switching key. In addition are facilities to control the "caution" and "on air" lights outside the studio door.

/ The

The following functions are associated with the Control Booth -

(i) Selecting and mixing a number of sources.

(ii) Control of level from associated studios by means of faders and volume indicator.

(iii) Effecting the switching of outlets from studio to studio as required.

(iv) Monitoring the programme originating in the associated studio and, if desired, from other studios.

(v) Switching outside lines to programmes, and feeding programme back to outside points when they are "off air" for cue purposes.

(vi) Intercommunication telephone system with switchroom, outside studios, studios and the announcer within the studio.

(vii) Giving of a warning light to the studio prior to commencement of a programme, and lighting the "on air" signals in the studio and outside the doors when broadcasting.

(viii) Talk-back facilities to studio.

(ix) Indicating when circuit conditions, depending on operations elsewhere, are suitable for the operation of the above facilities.

Some of the above equipment is shown in the pictures in this Paper.

STUDIO AND CONTROL BOOTH. (MELBOURNE.)

7.6 Switchroom. A brief reference was previously made to the switchroom when discussing switching principles, and the following is a summary of the principal facilities required at this centre -

(i) Presetting arrangements for switching outlets from one studio to another.

(ii) Effecting the switching of outlets, if required.

(iii) Testing and equalising outside lines prior to a broadcast and routing these to a studio.

(iv) Communication with studio control booths, broadcasting outlets, recording rooms, trunk equipment and test positions.

(v) Routine tests on all studio equipment, for example, frequency response, distortion and noise, and also on outside line connections.

(vi) Monitoring and observation on all outgoing programmes, as may be required.

(vii) The control and distribution of time signals and similar automatic signals.

Typical equipment includes -

(i) Line terminating rack.
(ii) Switching panels and relays.
(iii) Amplifier racks similar to the studio racks.
(iv) Branching amplifier racks.
(v) Test equipment rack.
(vi) Power distribution bay.
(vii) Monitoring loud-speakers.

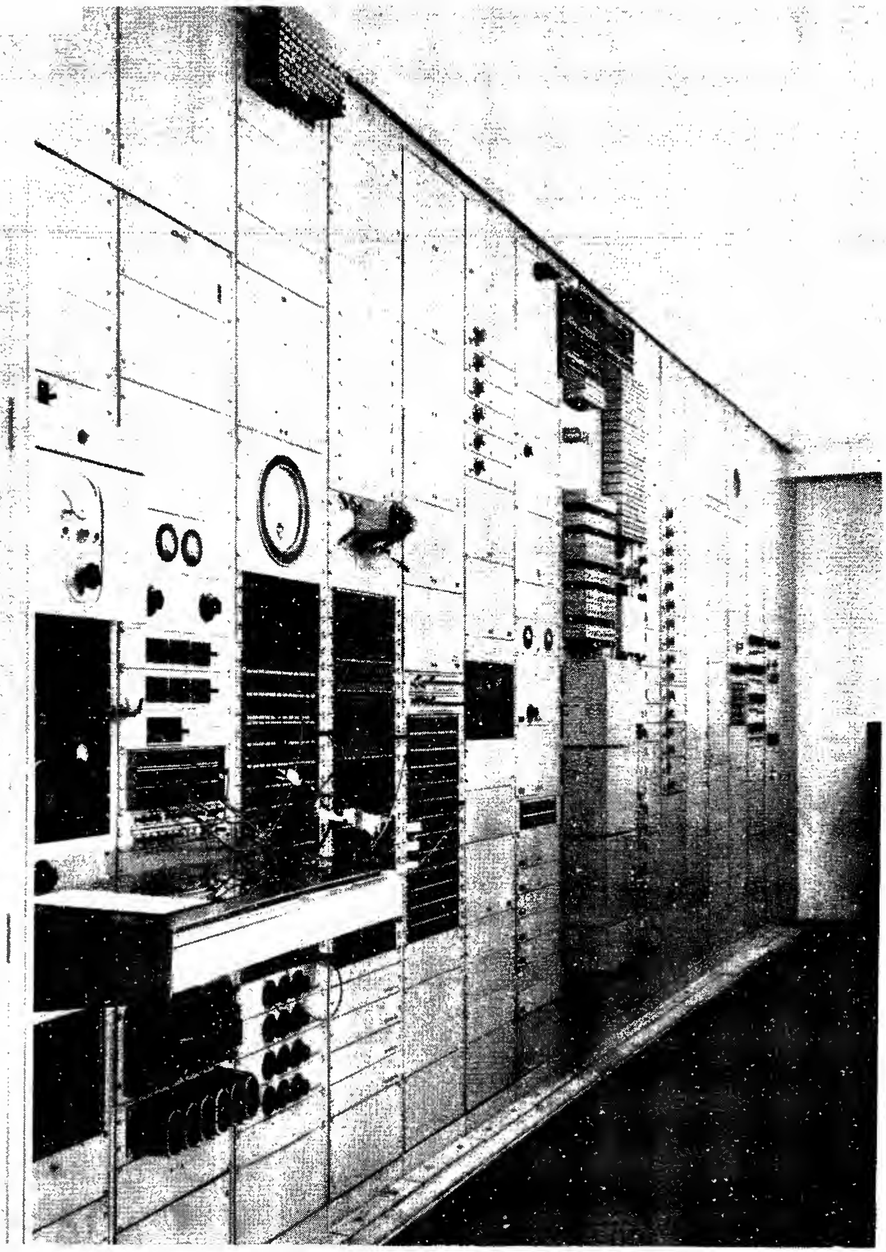
7.7 Some Aspects of Studio Design. Modern broadcasting studios are treated to give the exactly desired reverberation time. This is different from the reverberation time that would be taken as ideal for a concert hall or for direct listening to music. The proper reverberation time of a given studio is first calculated by formula and the studio is then treated with sound-absorbing materials to obtain the required reverberation time.

In designing a studio, it is necessary to keep in mind that any reflection paths must be broken up in order to avoid the production of standing waves. These standing waves are due to waves bouncing back and forth between, say, the floor and the ceiling. At some points their effects are additive and at some points subtractive, the net effect being that of a stationary wave. This effect may be neutralised by having the ceiling at an angle with respect to the floor, or vice versa, or by placing a highly absorbent plaque on the floor or ceiling.

Modern studio design aims at the use of little absorbing material, but the use of a great number of curved surfaces, so that any sound is reflected over a great many paths. This avoids the setting up of standing waves without the disadvantage of absorption of sound energy. The effect on reverberation time is to enable the use of a long reverberation time without producing echo effects. The effect also is to "liven" the studio by giving a good reproduction of the highest audio frequencies.

Because it is important to keep unwanted sound out of a studio and to keep originating sounds from getting out into other studios, many studios are "floated". The basic principle is a "room within a room", the inner room having no solid ties with the outer room, but being spring supported and separated. All connections, such as lighting conduits, power and air-conditioning systems, must be of a flexible non-sound transmitting nature.

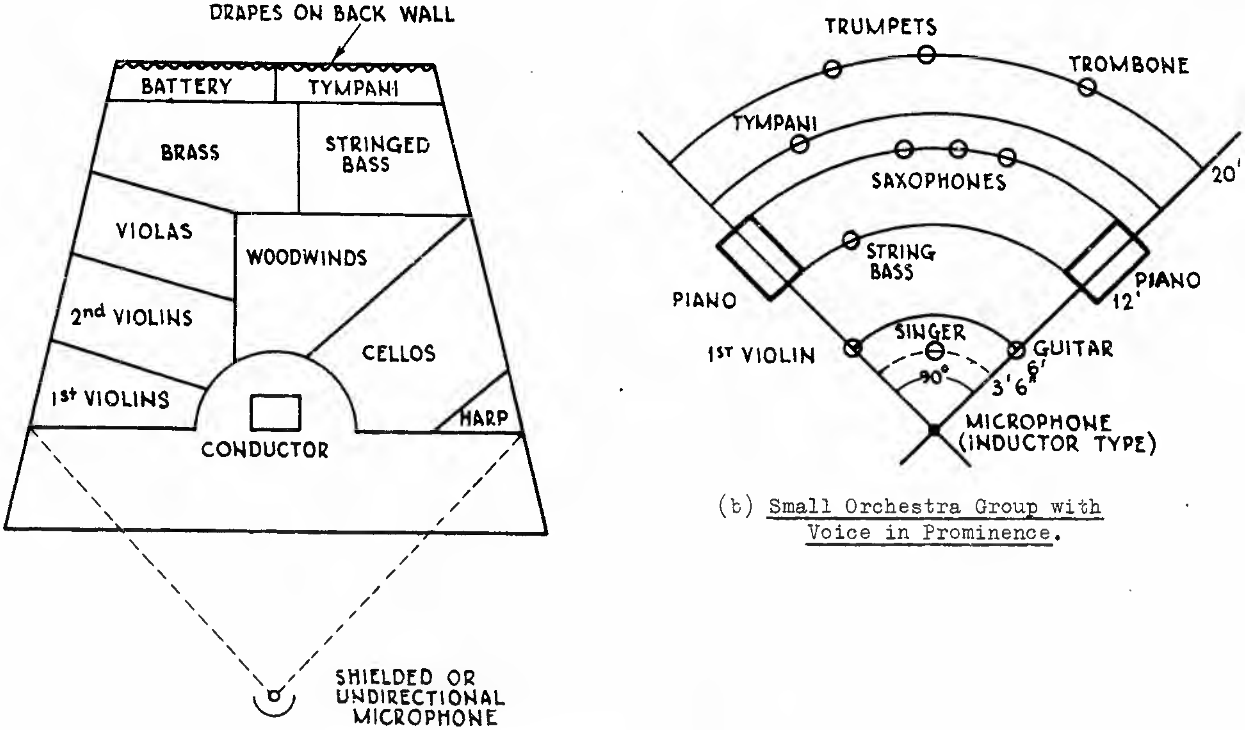
/ Typical



TYPICAL SWITCHROOM EQUIPMENT ASSOCIATED WITH STUDIO SWITCHING.

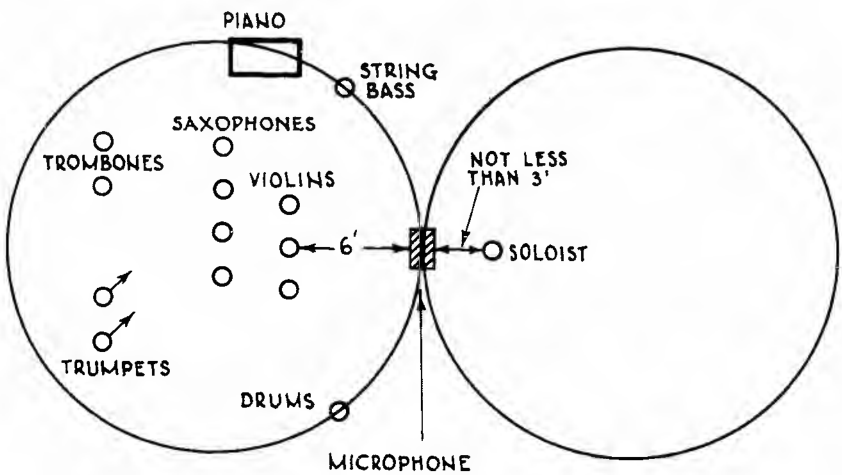
7.8 Miscellaneous Notes. Some typical studio arrangements are shown in Figs. 10a to 10c. Fig. 11 shows relationship of room volume and reverberation time.

Some pictures of studios are shown on page 24.



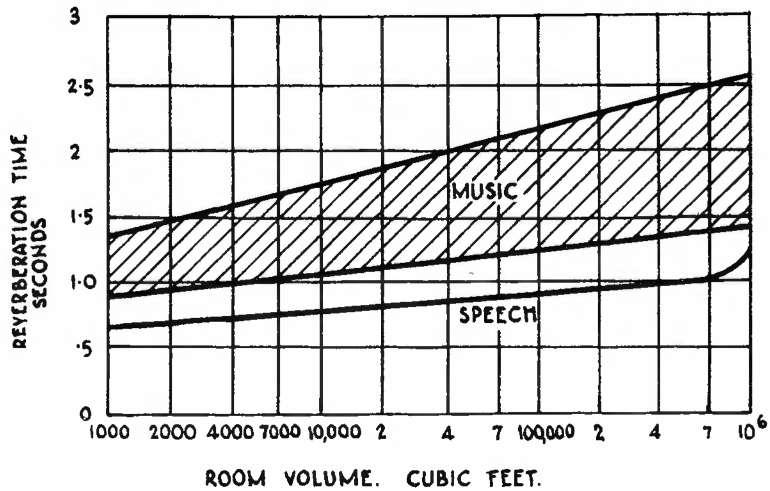
(a) Set Up for 110 Piece Symphony Orchestra.

(b) Small Orchestra Group with Voice in Prominence.



(c) Dance Orchestra with Soloist Using Velocity Microphone.

FIG. 10. TYPICAL STUDIO ARRANGEMENTS.



SHOWING HOW ROOM VOLUME AFFECTS THE REQUIRED REVERBERATION TIME FOR SPEECH AND THE RANGE OF REVERBERATION TIME FOR MUSIC.

FIG. 11.

Table 1 gives the peak power of various musical instruments playing triple forte (fff).

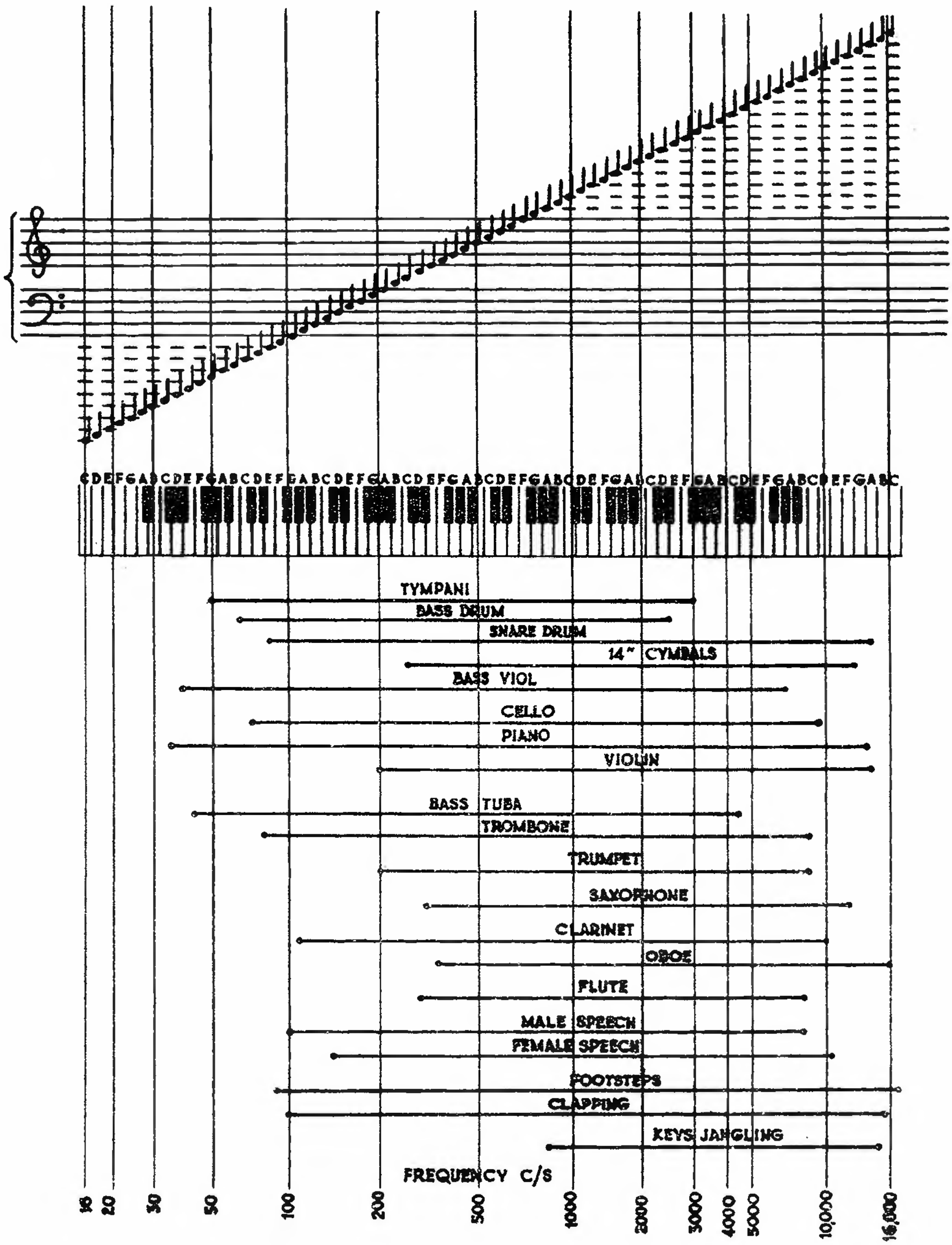
Instrument	Peak Power watts	Instrument	Peak Power watts
Heavy Orchestra	70	Bass Saxophone	0.3
Large Bass Drum	25	Bass Tuba	0.2
Pipe Organ	13	Bass Viol	0.16
Snare Drum	12	Piccolo	0.08
Cymbals	10	Flute	0.06
Trombone	6	Clarinet	0.05
Piano	0.4	French Horn	0.05
Trumpet	0.3	Triangle	0.05

TABLE 1.

Fig. 12 is of interest in showing the frequency range required for the reproduction of musical instruments, etc., without noticeable distortion.

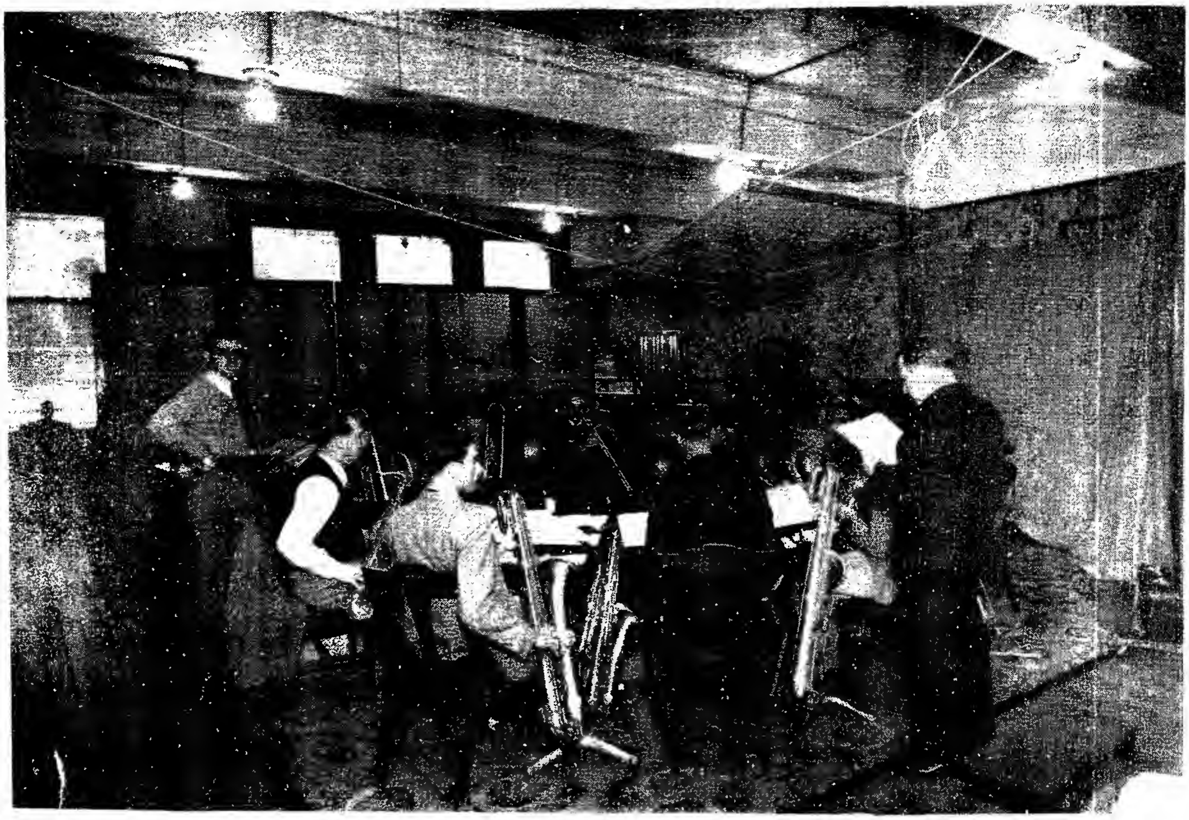
7.9 Line Equalisers. Long lines and circuits carrying programmes will deliver at their remote end an output lacking in high frequencies. The attenuation of these high frequencies is mainly due to the distributed capacity between the conductors, particularly if there is a section of cable involved. The reactance of this capacity decreases as the frequency increases and, at high frequencies, approaches a short circuit.

To correct this defect, adjustable filters called "equalisers" are inserted in the line as required (see Fig. 1). These equalisers are set to attenuate the low frequencies in exactly the same proportion as the line affects the high frequencies.



FREQUENCY RANGE REQUIRED FOR DISTORTIONLESS REPRODUCTION OF SOUNDS.

FIG. 12.

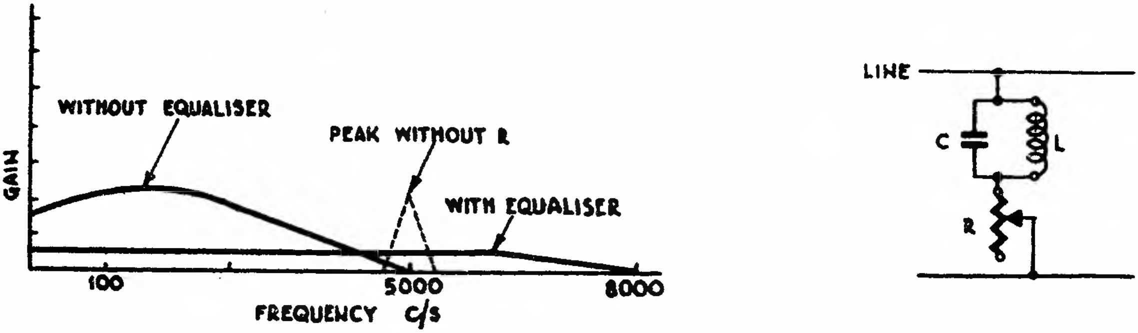


RADIO STUDIO OF YESTERDAY!

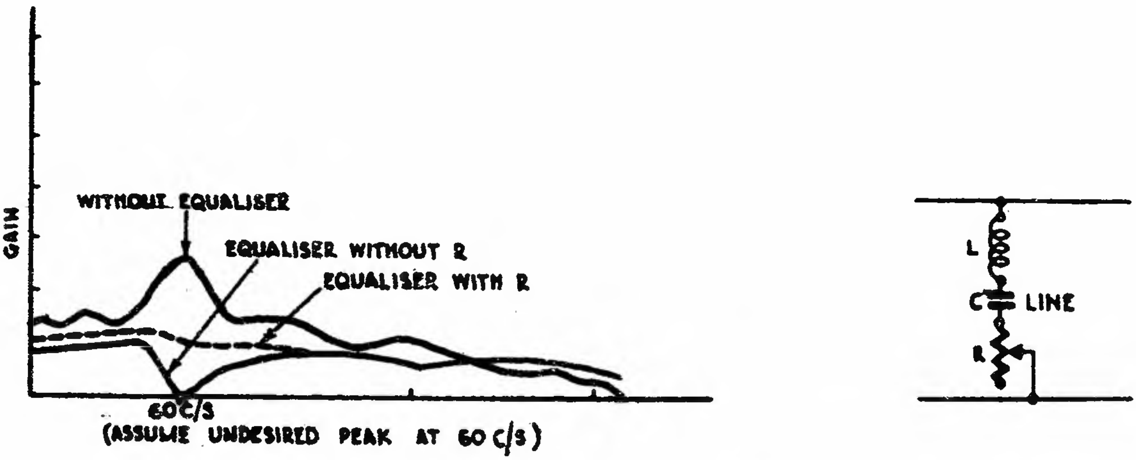
(Note the different means of endeavoring to improve the acoustical properties of the studio.)

RADIO STUDIO OF TODAY!

Two common types of equalisers and their effects are shown in Figs. 13a and 13b.



(a) Shunt Type.



(b) Series Type Equaliser for Particular Frequency.

LINE EQUALISERS.

FIG. 13.

A more detailed treatment of equalisers is found in Long Line Equipment. The equaliser is designed to resonate at the high frequency end (the exact frequency depending on the quality of service required), and resistance R serves to dampen the peak which occurs at the resonant frequency.

The series type equaliser (Fig. 13b) is of use in attenuating a particular frequency, which might be due to some characteristic of the line.

The component values are chosen to resonate at the frequency it is desired to attenuate, and the resistance R is used to dampen the effect at this frequency.

7.10 Typical Equalising Procedure. In order to equalise a line, an audio-frequency voltage of constant amplitude is impressed on the line. The source is usually a standard audio-frequency oscillator, and its output should contain negligible value of harmonics. An isolating attenuator pad is placed between the audio-frequency output and the line. It is very important that the impedance across the input of the volume indicator is constant over the entire frequency range to be transmitted, otherwise errors of measurements are introduced.

/ The

The equaliser is usually placed at the receiving end, and the output of the equaliser is fed to suitable level measuring equipment. A frequency run is then taken and the received level of each test frequency is noted, in order to furnish an approximate idea of the equalising required.

The equaliser is then adjusted until the volume indicator at the remote end gives the same reading, within the prescribed tolerances, for each frequency.

Preliminary adjustments of the equaliser resistance at three or four of the most important frequencies, say at 100, 1,000, 3,000 and 5,000 c/s, are made to determine the approximate value of the resistance required. Fine adjustments are then made until the response of the line is sufficiently flat over the required frequency range.

8. CONTROL BOOTH CONSOLETTE.

8.1 In some radio systems, the Control Booth equipment has been assembled into "consolette" form. This consolette provides a neat and useful arrangement in the case of Public Address equipment control, and in small transmitter installations where the programme facilities are not very extensive.

A block diagram of the facilities provided by a typical consolette is shown in Fig. 14.

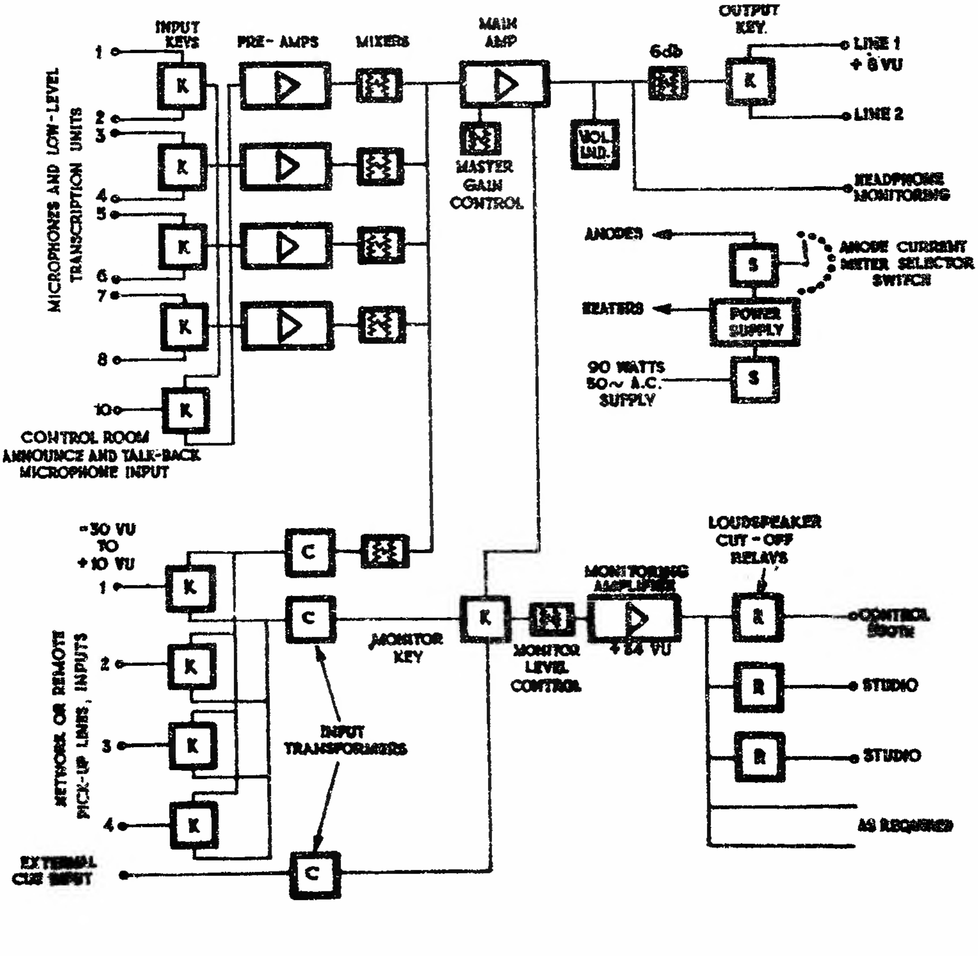
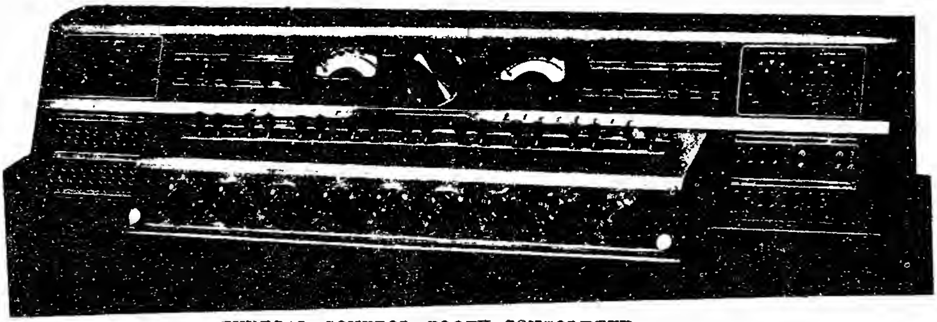
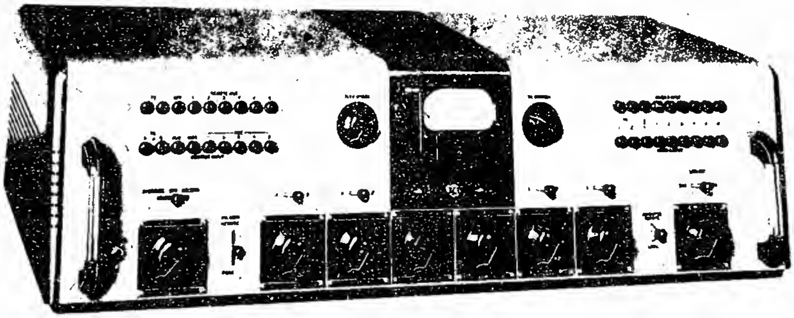
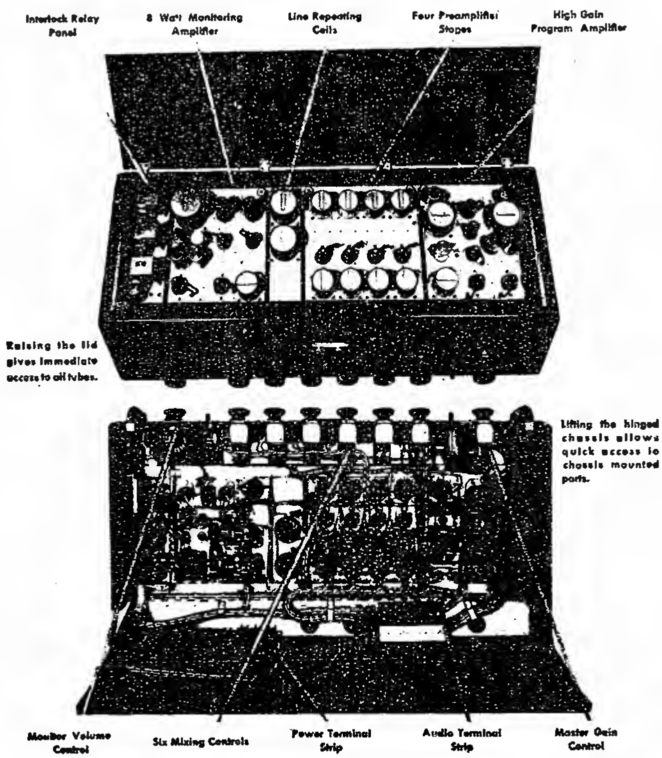


FIG. 14. FACILITIES PROVIDED BY TYPICAL CONTROL BOOTH CONSOLETTE.



TYPICAL CONTROL BOOTH CONSOLETTE.





ANOTHER TYPE OF CONTROL BOOTH CONSOLETTE.

9. TEST QUESTIONS.

1. Why is it necessary to equalise a programme line? Illustrate your answer with a typical circuit showing its effect by means of a graph.

2. (i) Define the term zero level as applied to a vu meter.

   (ii) What were the main reasons leading to the development of this type of level indicating meter?

3. State the general principles of acoustical design as applied to modern broadcasting studios, and what are the principal effects to be guarded against?

4. (i) What are the main functions performed by a control booth operator?

   (ii) Where is the control booth located in the general broadcasting scheme?

5. What is the function of a "pre-amplifier"? Why is particular attention usually paid to their design as far as noise is concerned?

6. Give a block schematic diagram of a broadcasting network showing the linking up of the following units -

   Microphone (talk studio).
   Gramo-unit.
   Outside pick-up.
   Fading/Mixing unit.
   A and B amplifiers.
   Master gain control.
   Monitoring speaker.
   Programme line to transmitter.
   Programme amplifiers and monitoring facilities at transmitters.

(Switching facilities, etc., are not required to be shown.)

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

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

COURSE OF TECHNICAL INSTRUCTION.

OSCILLATORS.

CONTENTS:-

1. INTRODUCTION.

1.1 For radio-communication purposes, an oscillator is defined as a source of alternating voltage of constant amplitude and of a frequency predeterminable by certain circuit parameters. (See Paper No. 1.) Although the frequency spectrum covered by radio extends from zero frequency to beyond 100,000 Mc/s, this Paper is primarily concerned with oscillators operating within the range up to approximately 30 Mc/s. Oscillators developed mainly for use at the higher frequencies are treated in later notes dealing with these frequency bands.

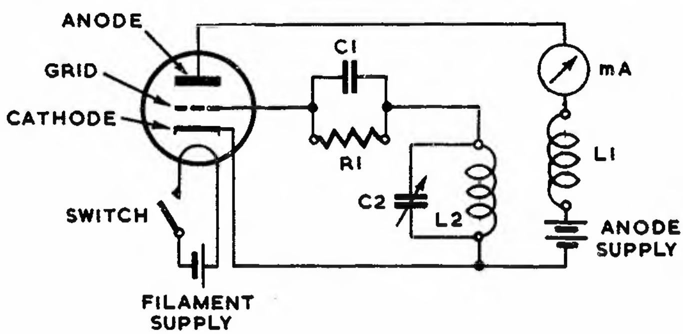
1.2 Due to its amplifying characteristic, a thermionic valve provides a simple and economic means of generating oscillations. Once a circuit commences oscillating, it continues indefinitely when there are no losses, but, due to the damping effect of the losses, oscillation gradually ceases, unless maintained by external means. That is where the amplifying action of the valve is useful, since, once the associated circuit is oscillating, sufficient energy is fed back (from the output circuit to the input circuit) to sustain the oscillations and make up for the energy lost in the circuit. This additional energy is not available without the amplification.

The feedback is likened to the gentle push given to a swing once it is in operation. This push serves to make up for the energy lost due to the résistance of the swivels, wind, etc., and keeps the swing moving at a steady rate in a constant arc. The valve also possesses little inertia, that is, it takes little energy to cause it to operate from a state of rest. In most cases, it is found that there is sufficient residual charge in the components comprising the circuit to cause oscillations to commence immediately power is applied.

Since the power supplies associated with the valve are D.C. and the output of the oscillator is A.C., the oscillator is regarded as a convertor changing D.C. to A.C.

1.3 Basic Oscillator Action. A valve oscillator consists essentially of an oscillatory circuit (generally a parallel-resonant circuit), a valve amplifier and a feedback circuit. In the oscillatory or input circuit, electrical oscillations occur according to the basic laws governing this type of operation. The feedback circuit takes a part of the energy from the output circuit and impresses it in the form of a voltage on the grid and cathode (the input circuit) of the valve. This energy is then amplified and again returned by the anode-cathode circuit to the output circuit, supplying the losses occurring in the latter and thus keeping it oscillating. The feedback circuit consists of either capacitive coupling through the grid-anode capacity of the valve, or inductive coupling through a coil in the grid circuit magnetically coupled to the output circuit. Thus, the valve itself does not oscillate, but merely amplifies the electrical impulses impressed on its grid from the oscillatory circuit, and delivers this amplified energy back to the input circuit to make up for the losses therein, in order to maintain oscillation.

1.4 Basic Oscillator. Fig. 1 is a basic oscillator circuit capable of producing waves of continuous character. The theoretical functioning of this circuit is identical for almost every form of valve transmitting-circuit oscillator and must be carefully analysed.



L1 = anode coil.
L2 = grid coil.
C2 = tuning capacitor.
C1 and R1 = grid leak-capacitor combination.

BASIC OSCILLATOR CIRCUIT.

FIG. 1.

When the switch in the filament circuit is closed, electrons are emitted from the cathode and are attracted to the anode. This electron flow constitutes a current which flows through anode coil L1 back to the cathode. The current through coil L1 results in a magnetic field around the coil.

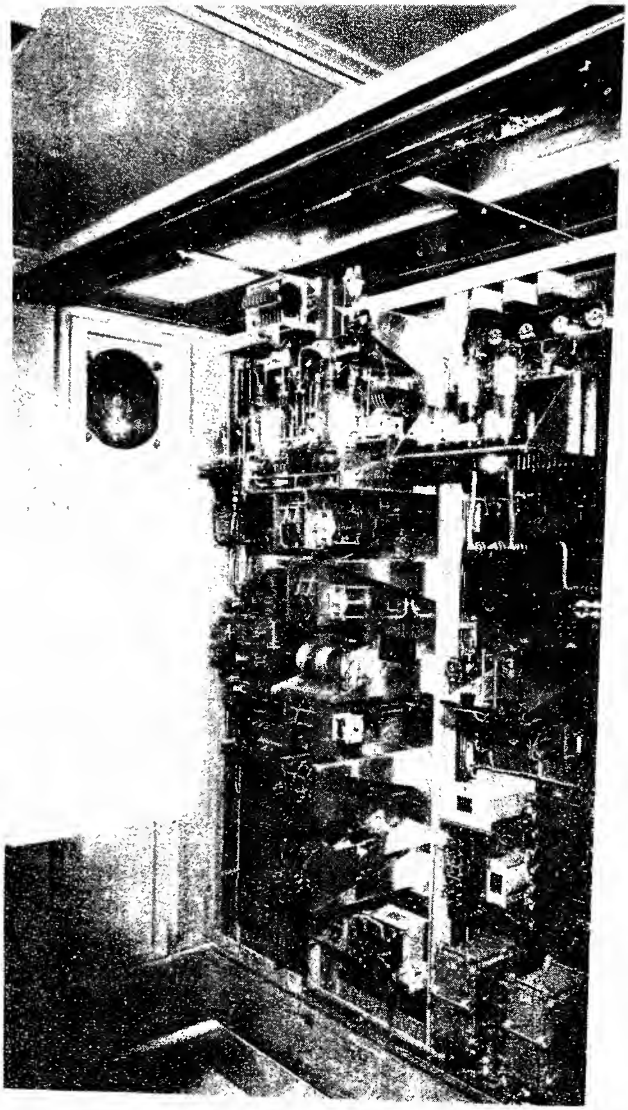
As the anode current is reaching its maximum value, an expanding magnetic field is formed around coil L1, and, as the coil L2 is near L1, the expanding magnetic field induces an e.m.f. in L2. When the relationship of the coils is such that the top of L2 is positive potential, the grid, which is connected to this end, is also at positive potential. This positive potential partly neutralises the space charge between the cathode and anode and allows more current to flow in the anode circuit. This increased flow results in a greater expansion and movement of the field in the same direction about L1 and, consequently, a greater e.m.f. of the same polarity is induced into L2, resulting in a greater charge upon the grid. Again, the anode current increases, and a heavier charge is placed upon the grid. This action continues up to a certain point, depending upon the characteristics of the valve and the resistance of the circuit.

As long as the rate of increase of current in LI is progressive, the e.m.f. induced in L2 increases and the rate of flow of electrons increases, resulting in further increases in the anode current. However, depending upon the characteristics of the valve and the resistance of the circuit, the rate of increase in the anode current decreases. This results in a lower e.m.f. being induced in L2 because the e.m.f. induced in that coil depends upon the rate of variation of the current through LI.

/ The

The potential on the grid, therefore, is reduced, and the anode current decreases. Some of the lines of force surrounding L1 collapse, thus cutting L2 in such a direction as to reverse the potential on the grid. As a consequence, the anode current is further reduced and decreases the potential on the grid. Further reduction in the anode current follows but, due to the characteristics of the valve and the resistance of the circuit, the rate of decrease of the anode current is reduced. Thus, the anode current rises and falls with a definite frequency, the period of which depends upon the values of the inductance and capacity L2 C2 in the circuit. By using the proper values, a circuit of this type can be arranged to produce frequencies ranging from the lowest audible range to the highest range in radio frequencies.

1.5 In all forms of oscillating systems, especially those in which the frequencies are high, precautions must be taken to keep the resistances of the constants low, otherwise the oscillations are apt to stop. At the higher frequencies, precautions must be taken against such additional losses as those due to high frequency resistance losses, dielectric absorption due to poor insulating materials, eddy currents and inefficient connections.



OSCILLATOR AND LOW POWER STAGES.

(VLG SHORT-WAVE STATION).

1.6 Type of Oscillators. The following list indicates some of the many types of oscillators which have been developed, but a complete list is outside the scope of these notes -

Feedback -

- Tuned anode.
- Tuned-anode tuned-grid.
- Hartley.
- Colpitts.
- Electron coupled.
- Push-pull.

Piezo-electric (Crystal),

Magnetostriction,

Negative Resistance -

- Dynatron.
- Negative transconductance.
- Push-pull.
- Negative grid resistance.

Heterodyne or Beat-Frequency,

Resistance Tuned,

Multivibrator (Relaxation Oscillator),

Magnetron, Barkhausen-Kurtz,

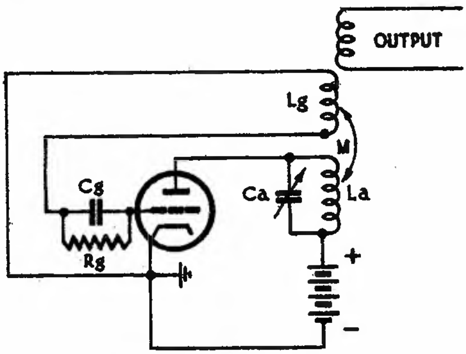
and many others for special applications.

/ 2.

2. FEEDBACK OSCILLATORS.

2.1 A feedback oscillator is considered as a tuned feedback amplifier, in which the amplitude and phase angle of the feedback are such as to cause oscillation. Suppose that a voltage Ei is applied to the input of an amplifier and that the resulting output voltage is Eo. When a portion of Eo is applied to the input and in phase with it, this feedback voltage acts in the same manner as Ei. When feedback is equal to Ei, then it replaces Ei, and the amplifier continues to deliver the original output Eo if Ei is now removed. In other words, the amplifier would oscillate at constant amplitude.

2.2 Tuned-Anode Oscillator. A typical feedback oscillator of the Tuned-Anode type is shown in Fig. 2, in which the feedback is obtained by mutual coupling between the grid and anode coils.



TUNED-ANODE OSCILLATOR.

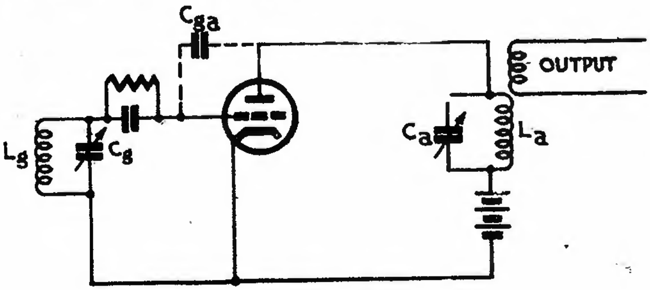
FIG. 2.

The valve is biased by the grid-leak/capacitor combination, which also functions to make the oscillator self-starting and is more likely to give a stable operating point under conditions corresponding to high efficiency operation. The use of a grid-leak also tends to make the oscillators self-adjusting at conditions corresponding to high efficiency and improves the frequency stability.

The grid capacitor must have a low reactance at the oscillator frequency so that the alternating voltage drop is small. The capacitor must not be too large, however, or the oscillations become intermittent. (See also "Miscellaneous Considerations" later in this Paper.)

The tuned-grid oscillator is similar to the above, except that the grid coil Lg is tuned instead of the anode coil La.

2.3 Tuned-Anode/Tuned-Grid Oscillator. A tuned-anode/tuned-grid oscillator is shown schematically in Fig. 3.



TUNED-ANODE/TUNED-GRID OSCILLATOR.

FIG. 3.

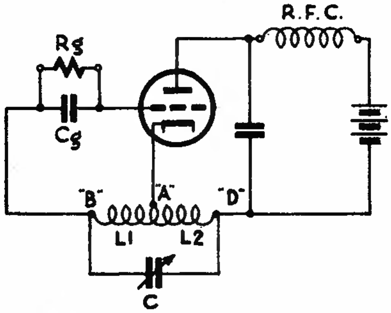
The grid tuned circuit LgCg and the anode tuned circuit LaCa are both adjusted to offer an inductive reactance at the frequency to be generated. The capacity Cga is usually supplied by the valve interelectrode capacity, although, at lower frequencies, added shunting capacity is often required to provide sufficient feedback.

However, too large a total value of this capacity is undesirable, because the circuit does not oscillate. In general, the output of the valve oscillator increases as Cga increases up to certain limits, depending upon the valve and circuits.

The energy fed back from the anode to the grid circuit through Cga causes an A.C. voltage to be built up across inductance Lg and thus across the grid and cathode. The valve then acts as an amplifier and develops a correspondingly high voltage in the anode-cathode circuit and across the oscillatory circuit, thus giving the latter additional energy and keeping it in sustained oscillation.

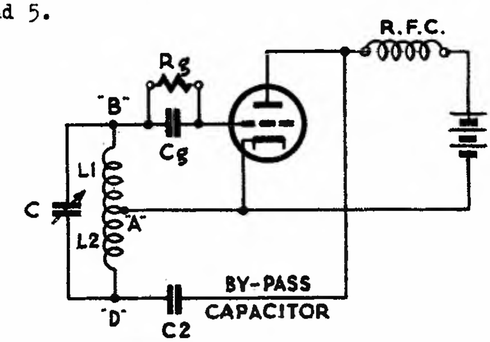
/ 2.4

2.4 Hartley Oscillators. The "Series-Feed" and the "Parallel or Shunt Feed" Hartley oscillator circuits are shown in Figs. 4 and 5.



SERIES-FEED HARTLEY OSCILLATOR.

FIG. 4.



SHUNT-FEED HARTLEY OSCILLATOR.

FIG. 5.

The difference between the Series-Feed and the Shunt-Feed Hartley Oscillators is that the anode-supply voltage is connected in series with the anode inductance in the Series-Feed type, and the tuned anode circuit is in parallel with the voltage supply fed through the radio-frequency choke (R.F.C.) in the Shunt-Feed type. The shunt method is used frequently, but has a disadvantage that parasitic oscillations may take place in the feed capacitor and choke. The Hartley oscillator is frequently used, as the criterion for oscillation is not critical. L1 and L2 are usually the two parts of a tapped coil, and the position of the tap is used to control the amplitude of oscillation. The circuit oscillates most readily when the ratio of L2 to L1 is in the range from approximately 0.6 to 1.0, the ratio increasing with the amplification factor. Because the tuning capacitor shunts both grid and anode coils, the Hartley oscillator gives a lower frequency for a given inductance than either the tuned-anode or the tuned-anode/tuned-grid oscillators.

The grid excitation takes place across the grid coil and the points A and B, owing to the voltage drop across the inductive reactance of the coil. The proper value of this reactance depends upon the type of valve used and the grid-excitation voltage required. The grid excitation in both the series and the parallel feed arrangements takes place through the medium of the induced e.m.f. from the anode portion AD of the coil to the grid portion AB, with the exception that, in the parallel feed (Fig. 5), the initial excitation takes place through the capacitor C2 in the form of a dielectric displacement of electrons. This displacement causes a movement of electrons in the anode coil which produces a magnetic field. Capacitor C2 also prevents the D.C. component of anode-current flowing in the oscillatory circuit, thereby allowing only the high-frequency currents to flow in the L and C circuits.

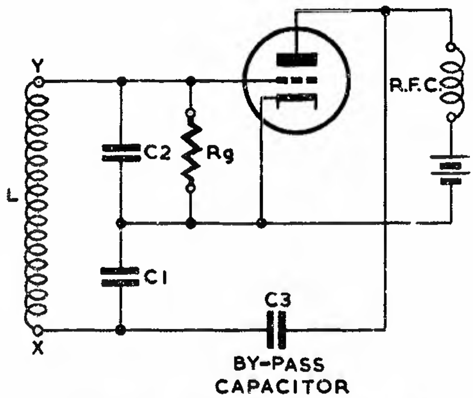
Capacitor C2 must have a low reactance to the operating frequencies to minimise the drop across it, so that it delivers the proper potential across the anode-excitation coil. The capacitor must, however, have a sufficiently strong dielectric to withstand the D.C. and oscillating potentials across it.

2.5 Colpitts Oscillator. Fig, 6 shows the Colpitts circuit which is similar to the Hartley oscillator, except that the coil and capacitor have changed places. Thus, the ratio of C2 to C1 determines the proper operating conditions instead of the ratio of L2 to L1.

The initial excitation is produced through the anode-blocking capacitor C3 in the form of an electron displacement, which, in turn, also produces a displacement in the capacitor C1. This results in a potential difference across C1 and C2, which excites the grid and produces sustained oscillations. The total reactive voltage across the two capacitors at the points XY is divided by the proper choice of C1 and C2. In practice, the ratio of C2 to C1 is about 3 : 1.

/ One

One point to be noticed in Figs. 4, 5 and 6 is the difference in the method of connecting the anode-supply voltage. In Fig. 4 the series-feed method is used, while in Figs. 5 and 6 the parallel-feed method is used. Each method has its advantages and special applications, depending on circuit and load conditions. In a series feed, the anode current flows through the output-circuit inductance. This arrangement permits the battery or other power-supply source to be practically at ground potential as far as A.C. is concerned, which is a desirable condition because it reduces losses and prevents inter-coupling with other apparatus. However, there is a disadvantage in that the output coil is at a high D.C. voltage above ground, necessitating additional insulation and also extreme care in the method of coupling to an external load.



COLPITTS OSCILLATOR (SHUNT FED).

FIG. 6.

In the parallel system, both the above disadvantages are eliminated, but the choke coil is required to withstand the full A.C. output voltage. This means that some A.C. flows through the choke coil, R.F.C., and is lost to the output circuit. This loss is very great unless the impedance of the coil is high, and the coil must be wound with many turns of wire to produce a large inductance, and, at the same time, have a low distributed capacitance shunting the winding. A properly designed choke coil offers from 5,000 to 10,000 ohms impedance at the operating frequency, so one or more coils connected in series prevent practically any current from flowing and thereby greatly reduce the possibility of circuit losses.

2.6 Electron-Coupled Oscillator. In the electron-coupled oscillator (Fig. 7), the inter-electrode anode-grid capacity is either neutralised or the anode is shielded by the use of a third screen-grid.

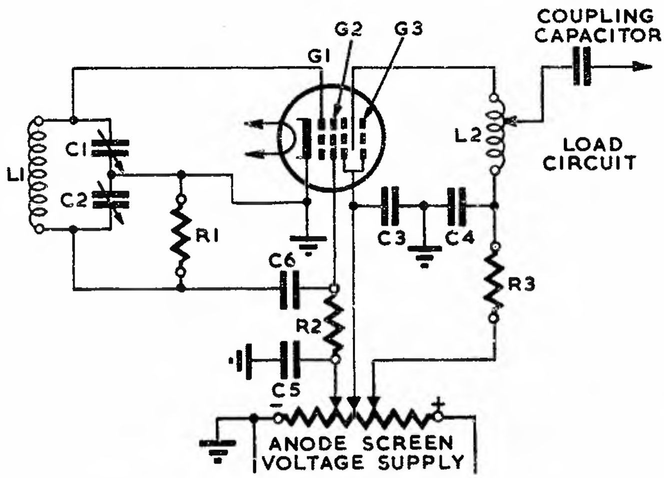


FIG. 7. ELECTRON-COUPLED OSCILLATOR.

In the latter case, the addition of a third grid near the anode acts as an electrostatic screen, making neutralising unnecessary. Hence when there is no capacitance between the anode and control grid G1 and no other electrostatic or magnetic relationship exists between the anode and grid circuits, any oscillations that are generated in the frequency-controlling circuit (L1, C1, C2) affect only the load circuit through the medium of the electron flow between the cathode and the anode. Hence, the variations in the anode circuit, because of the electron pulses transmitted to the anode from the cathode, generate in the anode circuit a pulsating D.C. variation of a frequency governed by the oscillator circuit. This, in turn, produces an alternating voltage in the anode circuit at the same frequency. A study of Fig. 7 shows that the only coupling between the load and oscillator circuits is that of the electron stream, and this results in an oscillator possessing very good frequency stability.

/ A

A further point of interest is that the cathode and grids G1 and G2 together form a Colpitts circuit with the screen grid G2 acting as the anode.

The capacitors C3, C4, C5 and C6 are radio-frequency by-passes to maintain a low reactance path for the A.C. variations in the load and screen circuits. R1 and C6 are the grid-leak combination to provide bias and stability, and resistors R2 and R3 decouple the power supply circuits. Radio-frequency chokes are a suitable alternative to the resistors.

The initial starting of the oscillations is due to a voltage drop across the split section of the tuning capacitor C2 at some instant of instability, such as when the cathode or anode supply circuit is first closed. This potential difference across C2 sets the oscillatory circuit (L1, C1, C2) into oscillation at a definite frequency determined by the constants -

$$f = \frac{1}{2\pi\sqrt{LC}}$$

During the initial oscillating pulse in the circuit L1, C1, C2, a voltage is built up across this combination. A part of this voltage is present across C1 which, since the grid-cathode terminals are connected directly across it, establishes an e.m.f. between G1 and the cathode. Hence, the alternating voltages built up across the oscillating circuit vary the grid voltages on G1 in accordance with the frequency characteristics of the circuit, and the grid, in turn, controls the rate at which the electrons impinge upon the anode.

2.7 Push-Pull Oscillator Circuit. A typical circuit arrangement of a push-pull oscillator is shown in Fig. 8a and an explanatory sketch is shown in Fig. 8b. The instantaneous operation of the valve V1 in Fig. 8b, is through the circuit shown by the solid lines.

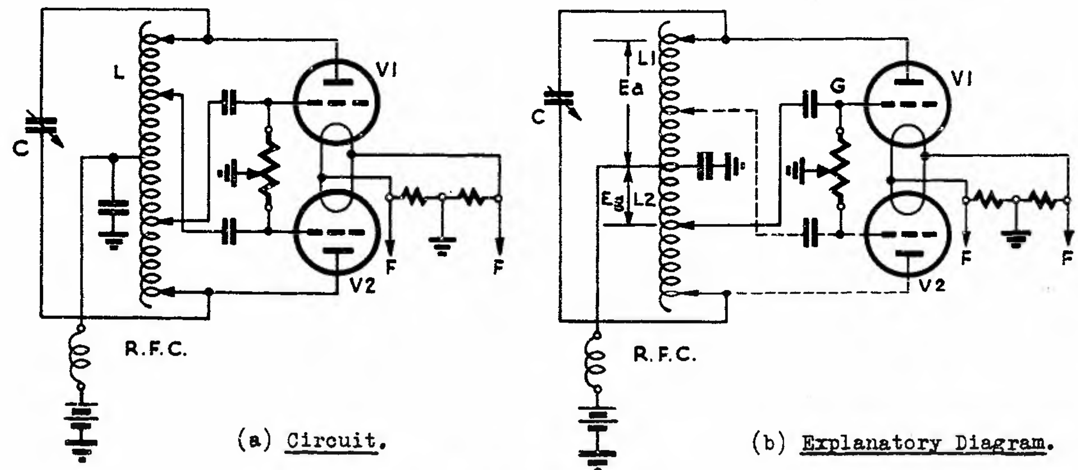


FIG. 8. BALANCED OR PUSH-PULL OSCILLATOR.

At the instant of any instability in the anode or cathode circuit, such as is caused by closing the power-supply circuit, a potential difference Ea is developed across the anode-excitation coil L1. During the period in which the instability takes place, a momentary moving of the magnetic field is developed across L1. In accordance with the laws of induction, therefore, an e.m.f. Eg is developed across L2. This e.m.f., in turn, excites the grid G of V1. Since C and L1 represent a shunted circuit connected across the valve, the circuit is set into self-oscillation at a frequency depending upon the L-C constants of the circuits. After the initial impact to the oscillatory circuit has occured, sustained oscillation takes place, resulting in an excitation of both grids 180° out of phase with each other. By the proper adjustment of the two grid connections to the inductance, perfect symmetry or balance is obtained.

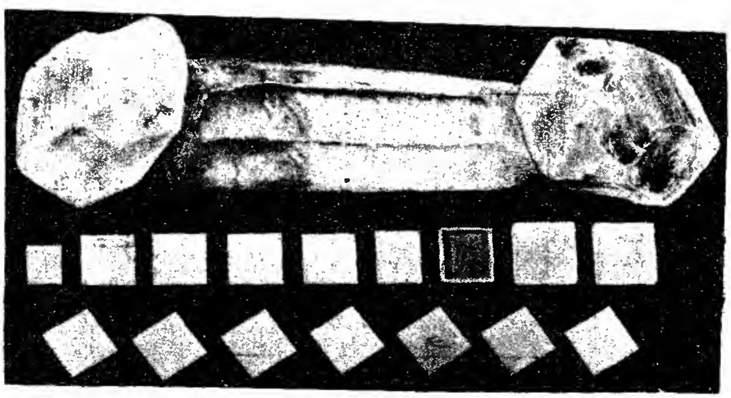
The advantage derived from an arrangement of this kind is greater power output and better frequency stability, owing to the reduction of the valve input and output capacitance, since, with this arrangement, both valves are effectively in series across the tuned-load circuit. This gradually reduces the circuit losses, due to valve capacitance and makes the form of oscillation highly desirable for use at high radio frequencies, where large power-oscillation outputs are required with good frequency stability. / 3.

3. CRYSTAL-CONTROLLED OSCILLATORS.

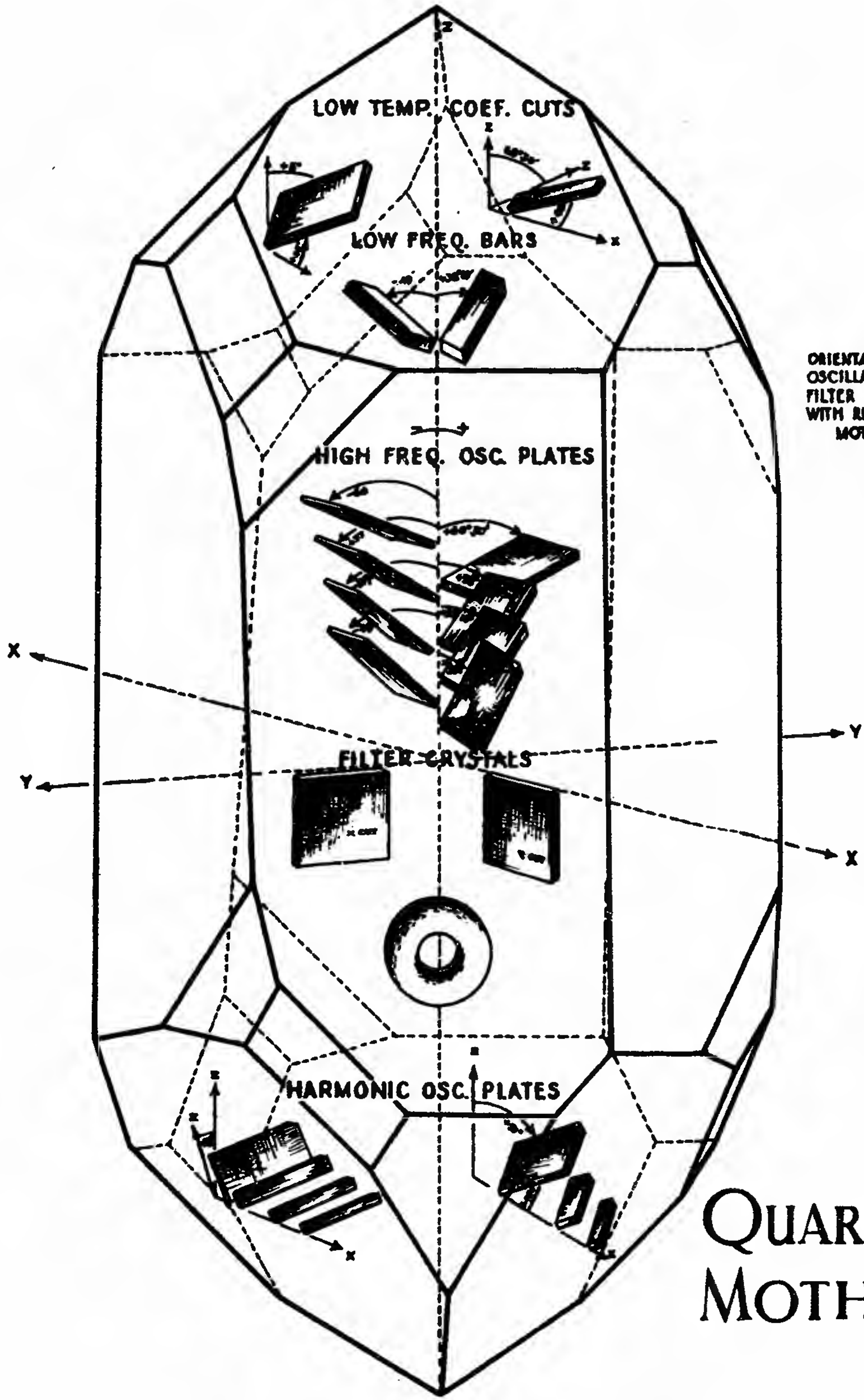
3.1 Although the oscillators just described have good stability, the long period stability is insufficiently good to render them entirely suitable for use in broadcast transmitters, where the tolerances required are approximately $\pm$ 10 c/s (in the range 500-1,500 kc/s). Fortunately, an oscillator was developed having its frequency controlled by a piezo-electric crystal.

3.2 Crystals. Several crystalline materials, such as quartz, tourmaline, Rochelle salt and cane sugar, have the property of producing an e.m.f. when the crystal is mechanically strained (compressed or expanded). Furthermore, this action is reversible, so that, when an electric charge is impressed upon the crystal, the latter changes its shape. This remarkable characteristic is called the "piezo-electric affect" and is present in all the so-called piezo-electric materials to a greater or less degree, dependent upon the material, upon the form in which it is cut and the manner in which the electric field is impressed. Rochelle salt is the most active in this property, but quartz has more mechanical advantages as it is practically unaffected by light shocks, aging and moisture. There is another advantage which is even more important. Quartz, due to its extreme hardness and low internal friction when vibrating, has low damping and consequently a high-Q characteristic. This property is the chief characteristic necessary in the oscillatory circuit of a valve oscillator where great frequency stability is required, and it is here that the quartz crystal plays a major role.

RAW QUARTZ.



FINISHED CRYSTALS.



ORIENTATION OF OSCILLATOR AND FILTER CRYSTALS WITH RESPECT TO MOTHER

QUARTZ MOTHER

An alternating voltage applied across a quartz crystal causes the crystal to vibrate and, when the frequency of the applied alternating voltage approximates a frequency at which mechanical resonance exists in the crystal, the amplitude of vibrations is very large. Any crystal has a number of such resonant frequencies, depending on the crystal dimensions, type of cut with reference to the axes of the natural crystal, the mode of mechanical stress and other factors.

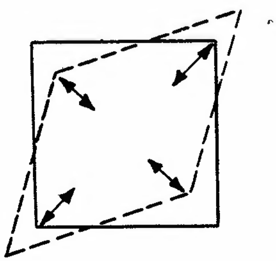
3.3 The desirable properties for a piezo-electric resonator are -

(i) Low temperature coefficient (that is, variation of frequency with temperature).
(ii) A high piezo-electric activity. ("Activity" indicates the ease with which the desired oscillations are started.)
(iii) Absence of parasitic oscillations and/or other frequencies near the required one.

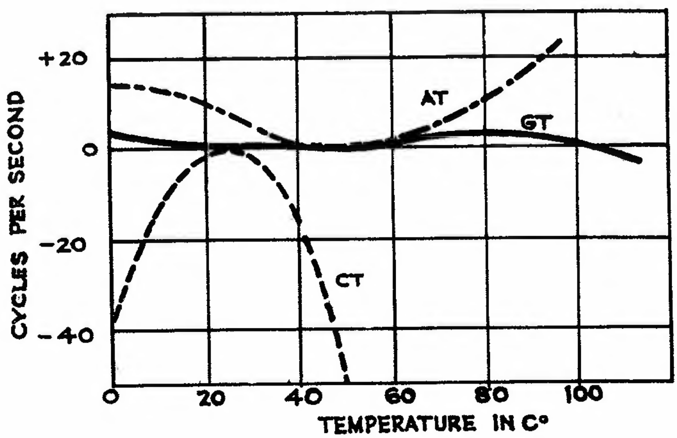
Before proceeding further, one point must be clearly understood. Although the piece of quartz used as a resonator is generally called a "crystal", it is not a whole crystal but only a section of a crystal cut to certain dimensions and specifications. This section may be a thin slab in rectangular form, or it may be cut like a disc. The smaller, thinner sections used for very high frequencies are of circular cut and similar in size to a coin. Still another form, which is used for precision work but not in commercial transmitters, is a flat ring. In these notes, the piezo-electric crystal is called a quartz plate, and this term indicates a thin slab of rectangular or circular form cut from a high-grade quartz crystal along a particular axis to make it more active piezo-electrically.

3.4 <u>Crystal Cutting and Temperature Coefficient.</u> There are many methods of cutting the plate from the natural crystal, as shown on page 9, and these have been given type letters such as AT, GT, CT, V, X and Y. The mechanical displacements of two of these types, when subjected to a force, are shown in Figs. 9a and 9b. These sketches show the method of vibration of the crystal when in operation. (The dimensions are greatly exaggerated, of course.) Fig. 9c shows the point regarding temperature coefficient, and the variation of resonant frequency with temperature of three typical cuts. It is noticed how superior the GT is in this regard.

(a) 'AT' Cut Displacement.



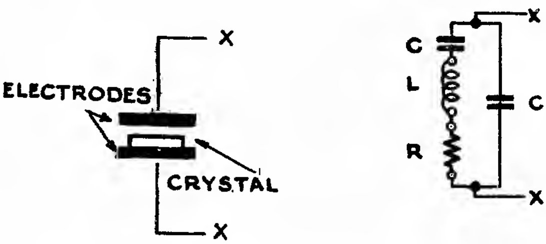
(b) 'CT' Cut Displacement.



(c) Variation of Frequency with Temperature.

FIG. 9. QUARTZ CRYSTAL PLATE DETAILS.

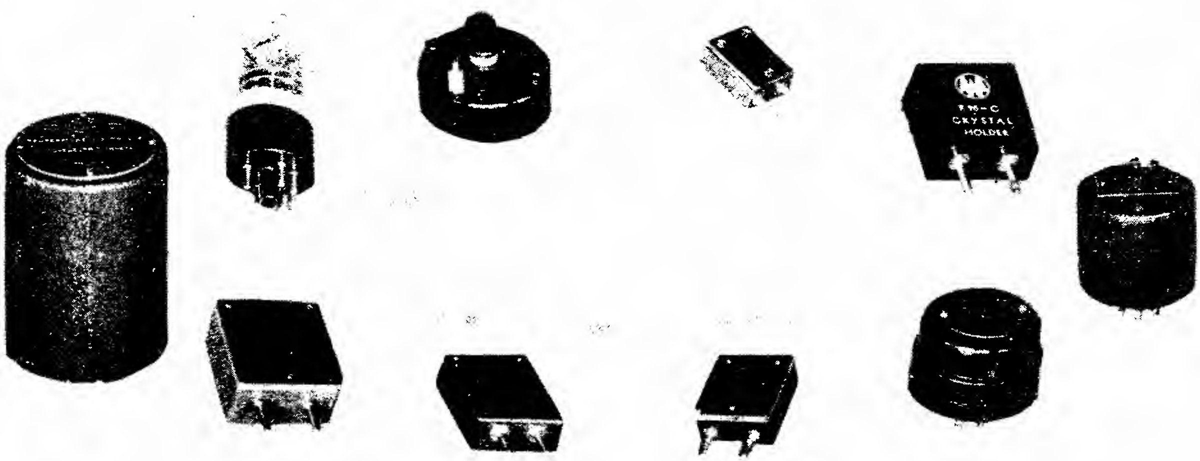
3.5 Crystal Resonator. When operating as a resonator, the piezo-electric crystal of Fig. 10a is considered equivalent to the electrical circuit shown in Fig. 10b. The series capacitor C represents the rigidity of the crystal, the inductor L its mass, and R the internal resistance due to internal friction. The parallel capacitor C1 is mostly due to the metal electrodes of the crystal holder acting like a capacitor with the crystal as the external connections and the grid-cathode capacitance of the valve when used.



(a) Crystal. (b) Equivalent Circuit.

FIG. 10. CRYSTAL RESONATOR.

A method of making electrical connections to the quartz plate is shown schematically in Fig. 10a. The quartz plate is usually placed in a horizontal position between two metal electrodes or plates. The lower electrode provides a base for the quartz plate and is an integral part of the whole mounting. The upper electrode may be spaced above the plate, touch it, or be clamped to it. The crystal, like a tuning fork, must be permitted a certain amount of mechanical freedom in order to oscillate. The early practice was to permit the upper electrode to rest lightly upon the surface of the plate, a flexible connection being made to the electrode. The difficulty with this method was that the electrode moved about and small particles of foreign matter settled under it, resulting in a change of spacing between the plate and the electrode. This condition caused considerable variation in frequency. To avoid such variations, the quartz-plate holder is now constructed so as to clamp the plate rigidly in one position. Each electrode is made with an uneven surface of rises and depressions. The rises make contact and clamp the plate, while the depressions provide air spaces between plate and electrode which allow the plate to oscillate or vibrate.



MOTHER QUARTZ AND ASSORTMENT OF CRYSTAL HOLDERS.

As a matter of interest, details of a relatively low-frequency quartz crystal (430 kc/s) are given below -

Dimensions -

Thickness	0.636 cm.
Width	3.33 cm.
Length	2.75 cm.

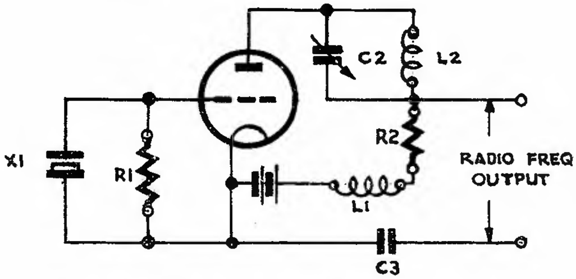
Resonant Frequency. 430 kc/s.

Equivalent Electrical Characteristics (Fig. 10b) -

$L = 3.3$ henrys.	$R = 4{,}500$ ohms (very approx.).
$C = 0.042\ \mu\mu F$.	$Q = 2{,}300$ (very approx.).
$C1 = 5.8\ \mu\mu F$.	

/ 3.6

3.6 Quartz-Plate Oscillator. A schematic diagram of a valve oscillator using a quartz plate resonator, is shown in Fig. 11.



OSCILLATOR WITH QUARTZ PLATE RESONATOR.

FIG. 11.

As previously explained, the valve with its associated circuits (such as cathode and anode supply, automatic bias by means of R1, and tuned output circuit) constitutes an amplifier. Feed-back to the grid circuit is obtained by the grid-anode capacitance of the valve. The resonant circuit that determines the operating frequency of the oscillator is the quartz plate X1 which is connected across the grid and cathode of the valve. The resistor R2 acts as a load to the amplifier, tending to stabilise it against external load changes and also to provide a means of coupling for the load. This resistance is only a few hundred ohms, so that variations in coupling from a high impedance load do not appreciably affect the oscillator.

When the anode voltage is suddenly applied to the valve, a surge is produced in the output circuit charging C2, which tends to discharge through L2 similar to the operation of an ordinary oscillatory circuit. This action in the output circuit causes a voltage to be built up across the anode and cathode of the valve, which, through the feedback coupling of the grid-anode capacitance impresses a voltage upon the grid-cathode circuit and across the quartz plate. The voltage in the latter circuit sets up an electrostatic field around the plate, causing it to change shape slightly. As soon as the impressed voltage dies down, the quartz plate returns to its normal shape and, in so doing, produces a voltage across the electrodes and across the grid and cathode of the valve. This voltage is amplified in the anode circuit and some of the energy is fed back again to the grid circuit and the quartz plate, which causes the latter to vibrate or oscillate, thereby applying a varying voltage to the grid. This action results in a continuous building-up process of the oscillations, until the valve oscillator is delivering its maximum power to its own and the external load for a given set of operating voltages and circuit constants. Actually, the output circuit is not tuned to the exact frequency but to a slightly higher one, so that it is somewhat inductive at the operating frequency. This condition provides greater stability and also aids the quartz plate to start vibrating more readily.

3.7 Tri-Tet Quartz Plate Oscillator. Fig. 12 shows a favourite oscillator circuit as used by amateurs and commonly known as the "Tri-Tet".

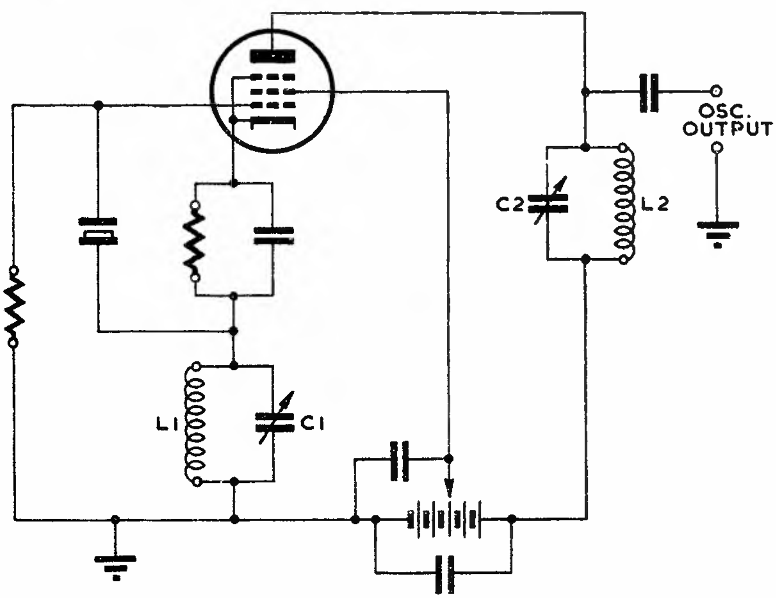
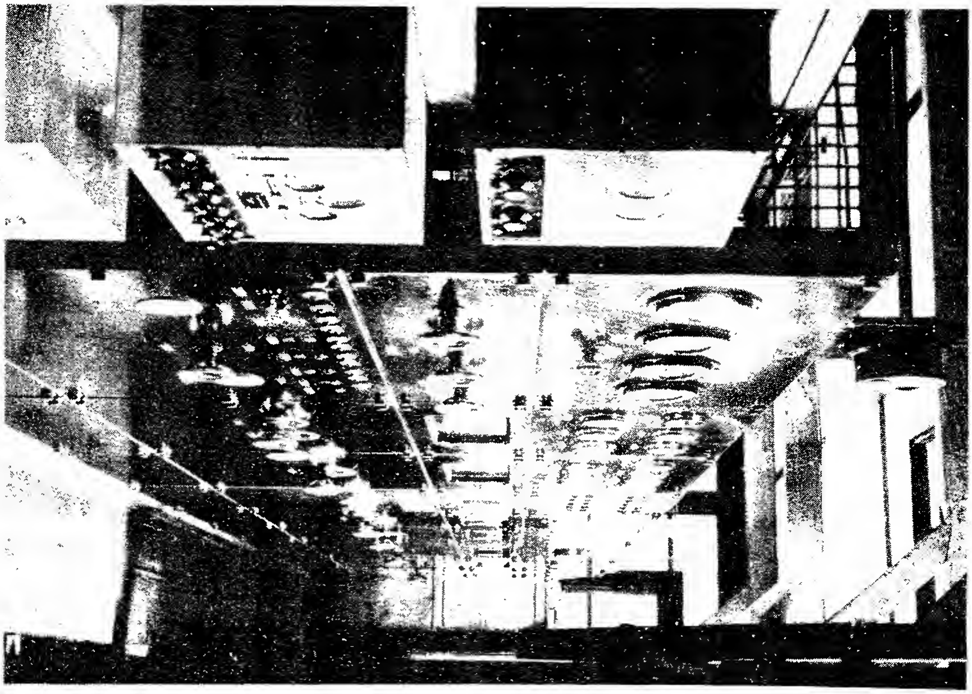


FIG. 12. TRI-TET OSCILLATOR.

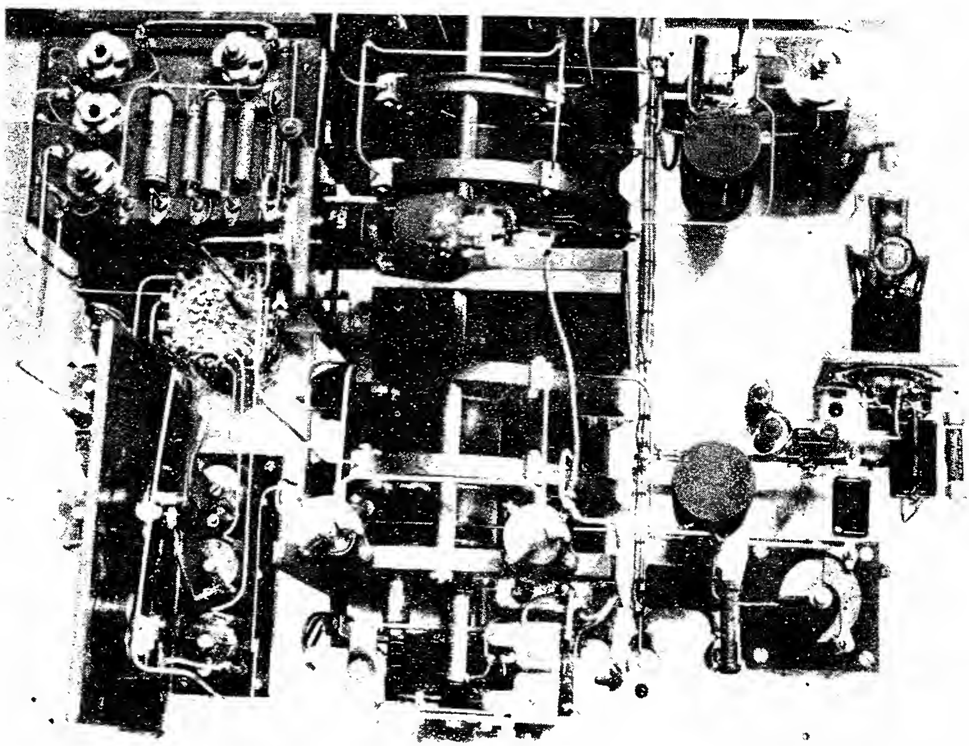
In this circuit the control-grid, screen-grid and cathode of a pentode or screen-grid valve function as a triode crystal oscillator with the screen serving as the anode, while the anode tank circuit is tuned to a harmonic. This arrangement is useful when a relatively high frequency is required, the crystal dimensions for which are impractically small.

In Fig. 12, the resonant circuit L1C1 is tuned to a frequency slightly higher than the crystal frequency and offers an inductive reactance to the crystal frequency. The resonant circuit L2C2 is tuned to a harmonic of the crystal frequency.

/ 3.8



6WF TRANSMITTER.
(CRYSTAL OSCILLATOR UNIT AT LEFT).

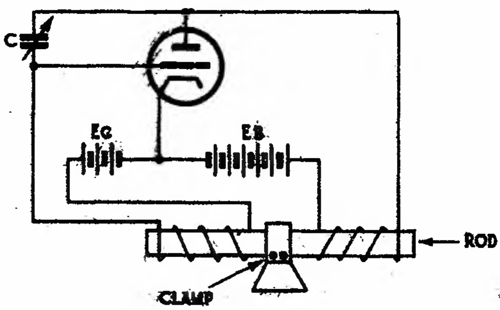


OSCILLATOR UNIT AT A H.F. TRANSMITTER.
(TWO "COLD" TYPE CRYSTALS AT LOWER RIGHT).

3.8 Crystal Oven. Where a very fine control of frequency is required, the crystal unit is mounted in a temperature-controlled oven. A thermostat provides a convenient means of switching a heating element on and off as the temperature varies between predetermined limits. By this means, the frequency of a crystal oscillator is controlled to the extent of a few parts in a million.

4. MAGNETOSTRICTION OSCILLATORS.

4.1 When a body is placed in a magnetic field, stresses are produced within the body tending to distort it. Inversely, when a body is distorted, there is a change in the magnetic permeability. Magnetostriction is the name given to this effect. Many metals and alloys exhibit magnetostriction, but it appears to be most pronounced in alloys of iron, nickel and chromium. When a rod of this material is magnetically polarised and placed in a coil carrying A.C. it vibrates longitudinally at the frequency of the A.C. When this frequency is the resonant frequency of the rod mechanically, the amplitude of the effect is large even for very small currents in the coil. This arrangement is equivalent to a parallel-tuned circuit coupled by means of the coil into the electric circuit.



MAGNETOSTRICTION OSCILLATOR.

FIG. 13.

The resonant frequency $= \frac{V}{L}$

where V = velocity of sound through the rod (approx. 4 kilometres per second),

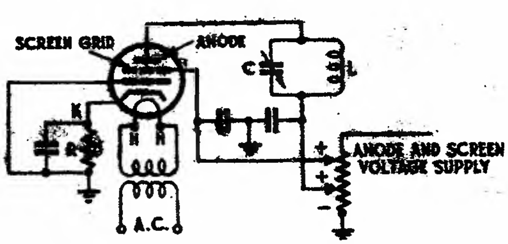
and L = the length.

Good stability (about 1 part in a million per degree Centigrade) is obtained, and thus compares favourably with the crystal resonator. A typical circuit is shown in Fig. 13. The capacitor C provides the feedback required, so that the circuit resembles the Tuned-Anode/Tuned-Grid circuit of Fig. 3 of this Paper.

5. NEGATIVE RESISTANCE OSCILLATORS (OR "TWO-TERMINAL" OSCILLATORS).

5.1 A two terminal circuit element is said to have a "negative A.C. resistance" when a positive increment of current through it results in a negative increment of terminal voltage.

5.2 The Dynatron Oscillator. The dynatron oscillator presents this characteristic of "negative resistance", which is well-known in the theory of regeneration and which is also used for other purposes requiring similar characteristics.



DYNATRON OSCILLATOR CIRCUIT.

FIG. 14.

The circuit arrangement of a screen grid valve used as a dynatron oscillator is shown in Fig. 14.

In a regenerative circuit, a portion of the signal voltage in the anode circuit is fed back to the grid circuit to overcome losses in the valve and its associated circuits. The losses in the circuit are referred to as the "positive resistance" of the circuit, and the energy fed back from the anode to the grid circuit to neutralise the effects of the positive resistance is called the "negative resistance". When sufficient energy is fed back, the losses due to positive resistance are so small that the circuit starts to oscillate.

/ The

The theoretical operation of the dynatron oscillator is described briefly as follows -

When the cathode is heated to incandescence and an anode potential is applied between the anode and the cathode, the electron flow from the cathode to the anode results. The screen grid is operated at a higher positive potential than the anode and electrons arriving at this grid cause an electron flow in the grid circuit of fairly large amplitude, which is, of course, true in conventional circuits when large signal potentials are applied to the grid. An important function, however, remains for the electrons that passed through the grid laterals toward the anode. These electrons on arriving at the anode cause a breakdown of the surface tension, owing to the collision impacts, which results in an emission of electrons from the anode surface. This phenomenon is known as secondary emission. Now, since the screen grid is at a higher potential than the anode, the electrons due to secondary emission are attracted to the grid. This presents a condition in which both the electrons from the cathode and from the anode (the latter due to secondary emission) are attracted to the grid. Now assume the anode voltage to be gradually increased. In ordinary valve circuits, an increase in anode voltage results in an increase in the anode current, but, in this circuit, the gradual increase in anode voltage reduces the anode-to-cathode current flow, owing to the neutralising effect at some critical point between the normal electron emission from the cathode circuit and the secondary electron emission from the anode. When this particular point is reached, the anode-to-cathode circuit resistance indicates negative resistance characteristic when an A.C. potential is applied to the circuit. This peculiarity is the basis of the action of a dynatron oscillator.

5.3 Negative Transconductance or Transition Oscillator. This type of oscillator (Fig. 15) makes use of the condition that increase of negative suppressor voltage of a pentode causes more electrons to go to the screen instead of to the anode, and thus increases the screen current and decreases the anode current. In other words, the anode-screen transconductance is negative.

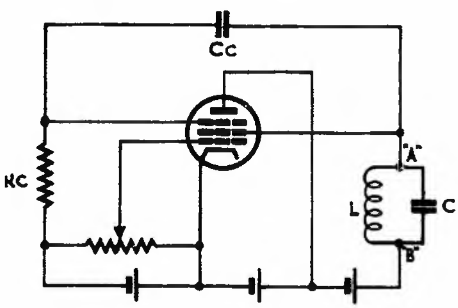


FIG. 15. TRANSITION OSCILLATOR.

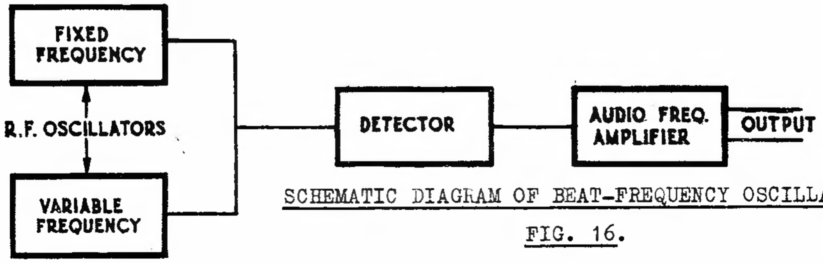
In Fig. 15 -

Cc = Feedback capacitor,
Rc = Coupling resistor,
LC = Tuned circuit determining frequency, and
"AB" = Points across which negative resistance appears.

5.4 The two circuits in Figs. 14 and 15 sufficiently show the principle of operation of Negative-Resistance Type Oscillators, and the information about Push-Pull and Negative Grid Resistance types are found in radio text books.

6. BEAT-FREQUENCY OSCILLATORS.

6.1 In the Beat-Frequency Oscillator, voltages obtained from two radio-frequency oscillators operating at slightly different frequencies are combined and applied to a detector or convertor valve, as shown schematically in Fig. 16. The difference-frequency current that is produced represents the output of the oscillator. The practical value of the beat-frequency oscillator (B.F.O.) arises from the fact that a small percentage variation in the frequency of one of the individual oscillators, such as can be obtained by a single turn of a dial controlling a variable-capacitor, varies the "beat" or difference frequency output over a considerable range (for example, 10 c/s. to 20 kc/s.) At the same time, the oscillator output can be made substantially constant as the frequency is varied.

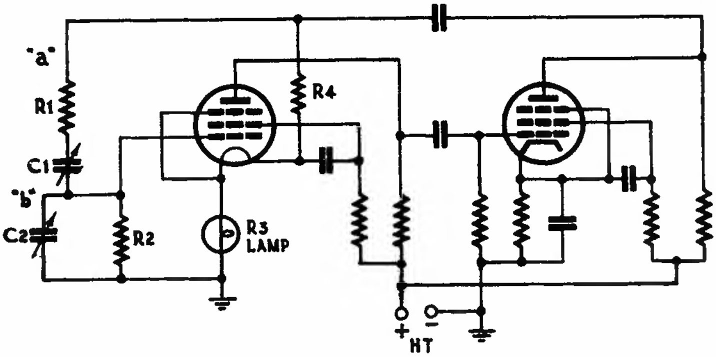


SCHEMATIC DIAGRAM OF BEAT-FREQUENCY OSCILLATOR.

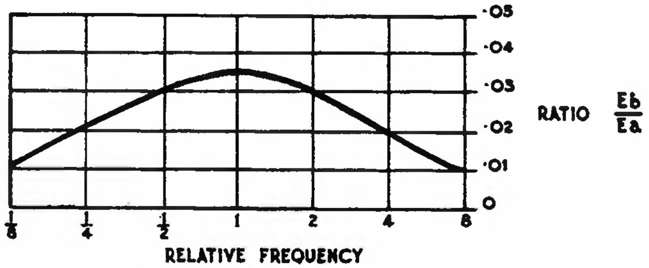
FIG. 16.

7. RESISTANCE-TUNED OSCILLATOR (OR RESISTANCE-CAPACITANCE TUNED OSCILLATOR).

7.1 In this type of oscillator, shown in Fig. 17a the frequency is determined by a resistance-capacitance network that provides regenerative coupling between the output and the input of a feedback amplifier. Here, the network consists of R1, C1, R2, C2, and, by proportioning so that R1C1 = R2C2 under all conditions, the ratio of the voltage developed across the input (Point b) of the amplifier to the voltage existing across the output (Point a) varies with frequency, as shown in Fig. 17b.



(a) Circuit.



(b) Variation of $\frac{Eb}{Ea}$ with Frequency.

FIG. 17. RESISTANCE-CAPACITANCE TUNED OSCILLATOR.

Oscillations tend to take place at the maximum of this curve, which is at a frequency of -

$$\frac{1}{2\pi\sqrt{R1.R2.C1.C2.}} \text{ c/s.}$$

The amplitude of the oscillations are satisfactorily controlled by using a ballast lamp to provide the feedback resistance R3. Increase of amplitude raises the current through the lamp and thus increases its resistance. This, in turn, increases feedback, which decreases the gain and thus the amplitude. The result is a substantially constant amplitude under widely varying conditions and absence of distortion due to overloading. The frequency generated is inversely proportional to the tuning capacitance instead of to the square root of the tuning capacitance as is usual, thus rendering it possible to cover a frequency range of 10 : 1 with an ordinary tuning capacitor as used in broadcast receivers. The wave-shape is extremely good, frequency stability is high and the output is virtually constant over very great frequency ranges.

The oscillations also readily synchronise in harmonic relation with voltages injected almost anywhere in the circuit, which is a useful property for applications such as frequency measurements of radio transmitters.

/ 8.

8. MULTIVIBRATOR OR RELAXATION OSCILLATOR.

8.1 The term "Relaxation Oscillator" is applied to those oscillators in which the frequency is controlled by the charge or discharge of a capacitor or inductor through a resistor. Such oscillators are characterised by highly distorted wave-shapes and are used to generate short pulse, saw-tooth, square or triangular waves according to the circuit arrangements and details. The frequency of relaxation oscillators is also readily controlled by voltages injected into the circuits.

8.2 The Multivibrator is one of a number of types of relaxation oscillators and consists of a two-stage resistance coupled amplifier, in which the output of the second stage supplies the input to the first stage, as shown in Fig. 18. Such an arrangement oscillates, because each valve introduces 180° phase shift. The frequency of oscillations is determined primarily by the time constants of the grid-leak grid-capacitor arrangements and approximates -

$$\frac{1}{R'_g C'_g + R''_g C''_g} \text{ c/s.}$$

The usefulness of the Multivibrator arises from the fact that the wave is so irregular that harmonics up to at least the several hundredth are present, and that the frequency is readily controlled by an injected voltage.

8.3 Fig. 19 shows how several multivibrators are used for frequency multiplication from a precision standard. The accuracy of harmonics is the equivalent accuracy of the 1 kc/s input.

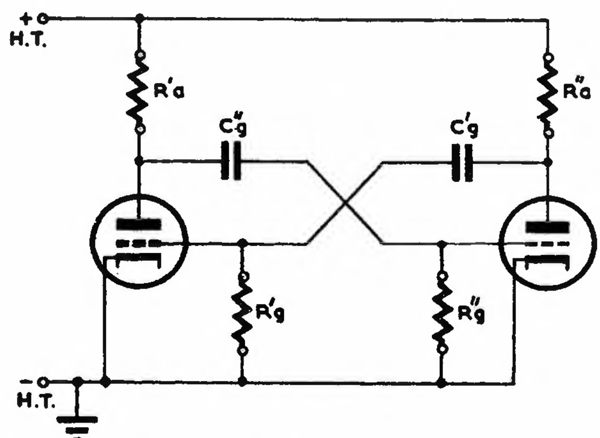


FIG. 18. MULTIVIBRATOR OR RELAXATION OSCILLATOR.

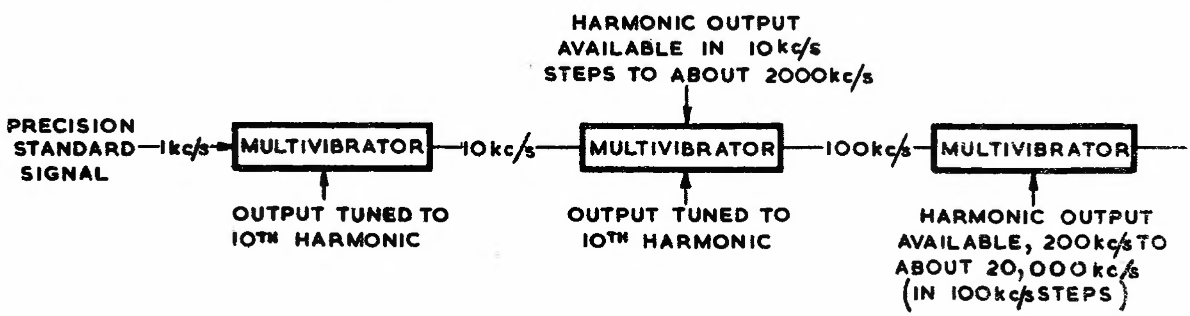


FIG. 19. SEVERAL MULTIVIBRATORS USED FOR FREQUENCY MULTIPLICATION.

9. MISCELLANEOUS OSCILLATORS.

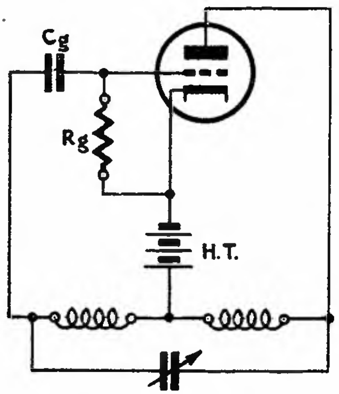
9.1 The Magnetron, Barkhausen, Klystron, etc., oscillators are types developed mainly for use at ultra-high frequencies, and the discussion of them therefore, is left for later notes dealing with these frequency bands.

10. MISCELLANEOUS CONSIDERATIONS RELATING TO OSCILLATORS.

10.1 Use of Self-Bias to Limit Amplitude of Oscillation. Reference was made earlier to the use of self-bias, and a more detailed explanation of its effects are given here.

Fixed bias is rarely used in practical oscillators, as it is necessary to limit the amplitude of oscillations to prevent excessive distortion and frequency instability.

The most common method of limiting the amplitude is by the use of a grid blocking capacitor and grid-leak resistor, as shown in Fig. 20.



Cg = Grid capacitor.
Rg = Grid Resistor (or Grid-Leak).

LIMITING THE OSCILLATOR AMPLITUDE.

FIG. 20.

The initial bias is zero but, as soon as oscillation commences, the grid is driven positive during a portion of the cycle, and so electrons flow from the cathode to the grid. During the remainder of the cycle, these electrons cannot return to the cathode but can only leak off the capacitor and grid through the resistor Rg. The trapped electrons make the potential of the grid negative with respect to the cathode, thus providing a bias. The greater the amplitude of oscillation, the more positive the grid swings and the greater is the average grid current. Thus, the bias builds up with oscillation amplitude, causing the transconductance to fall until equilibrium is established. In this manner, the amplitude is prevented from becoming too high without making the quiescent transconductance sufficiently low to prevent oscillation starting spontaneously when the power is switched on. Under equilibrium conditions, grid current flows during only a very small fraction of a cycle, and the bias closely approaches the amplitude of the alternating grid voltage.

Low power loss in the resistor, high-frequency stability and good wave form call for the use of a high grid-leak resistance, but it is found that, when the resistance is too high, oscillation is not continuous. After a number of cycles of oscillation, the bias becomes so high that the circuit stops oscillating. Because oscillation starts at a lower bias than that at which it stops, some time elapses while the capacitor discharges sufficiently to allow oscillation to recommence, and thus periods of oscillation alternate with periods of rest. This is termed "motorboating" from its characteristic sound. The period of motorboating depends upon the time required for the capacitor to discharge, and thus the "time-constant" of the resistor-capacitor combinations is of importance. (Time-Constant equals product of grid capacitance Cg and grid leak resistance Rg.)

/ 10.2

10.2 Frequency Stability. Undesired changes of frequency result from three major causes; changes in -

(i) The Mechanical Arrangement of the Elements of the Oscillating Circuit. These changes are produced by vibration, by mechanical, electrostatic or electro-magnetic forces, or by temperature changes.

Remedy - Careful design and by temperature control.

(ii) The Values of the Circuit Parameters. Likely causes are changes of temperature in inductors and capacitors, and from variation of load which alters effective resistance of tuned circuit.

Remedy - Temperature control, use of thermally compensated inductors and temperature-controlled compensating capacitors. Careful choice and location of component parts. Use of a "buffer" stage between oscillator and subsequent stages.

(iii) The Amplification Factor, Grid and Anode Resistances and Inter-electrode Capacitances of the Valve. These factors are affected by operating voltages, cathode emission and electrode spacing.

Remedy - Stabilised power by the use of voltage regulators. The dependence of frequency upon inter-electrode capacitances are minimised by the use of a high ratio of tuning capacitance to inductance (high "C" circuit), and by the use of circuits where the tuning capacitance shunts the inter-electrode capacitance.

10.3 Blocking. The phenomenon of blocking appears as a sudden stoppage of oscillations, accompanied by a reversal of grid current and an increase of anode current to a value much higher than is obtained with the full D.C. supply voltage and zero grid potential. A high-power valve is usually destroyed by blocking, since the energy dissipated at the anode is very great. Blocking is caused by operating conditions that permit secondary emission from the grid at a greater rate than the flow of electrons to the grid from the cathode. This causes a reversal of grid current, making the grid positive and increasing the flow of electrons to the anode (a negative grid reduces this flow). A factor which contributes to this condition is a high grid leak resistance.

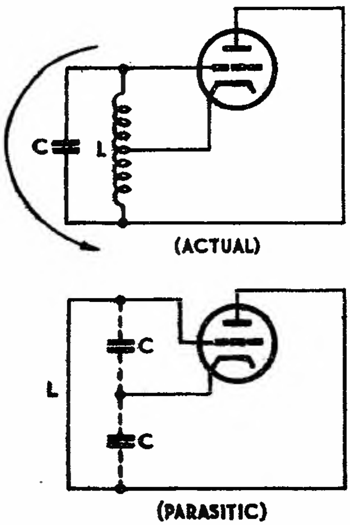
10.4 Effect of Harmonics on Frequency Stability. Harmonic voltages in the circuits of an oscillator adversely affect the frequency stability. This arises because the harmonics cross-modulate with one another and with the fundamental to produce fundamental frequency currents, which are not necessarily in phase with the fundamental frequency currents due to normal operation. The phase of the resultant fundamental frequency current affects the frequency of the oscillator.

10.5 Parasitic Oscillations. Parasitic oscillations are referred to in connection with radio-frequency amplifiers, but they are often present in oscillators and cause reduction of power output, spurious frequencies and distortion. Some examples of the manner in which parasitic oscillations occur are shown in Fig. 21. These oscillations are most likely to occur when large valves are used, because of the long leads, large inter-electrode capacitances and relatively large values of trans-conductance involved.

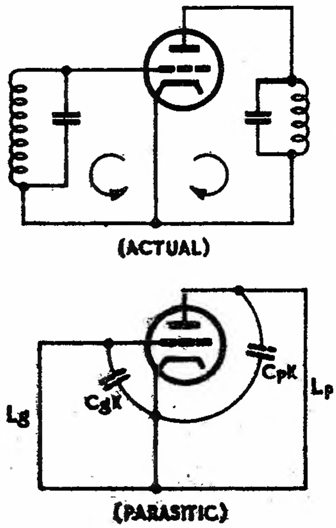
Many other sources of parasitics are found by using this method of simplified drawing, but these suffice to indicate the need for careful design of oscillator and amplifier circuits.

In Fig. 21, Lp and Lg are supplied by the one-turn loops from the valve electrodes through the tuning capacitances back to the cathode. The capacitance for the parasitic oscillation circuit is supplied by the inter-electrode capacitances Cpk and Cgk.

/ Fig. 15.



(a) Hartley Circuit Acting as a Colpitts with "L" Supplied by the Grid-Anode Lead.



(b) Tuned-Anode/Tuned-Grid Circuit showing the formation of a Parasitic Tuned-Anode/Tuned-Grid Circuit.

FIG. 21. PARASITIC OSCILLATIONS.

11. TEST QUESTIONS.

1. Sketch a typical oscillator circuit of the feedback type and briefly outline the principles of operation.

2. Why is a high ratio of capacitance to inductance desirable in an oscillator circuit?

3. When an oscillator having a high degree of stability is required, state the points in design to which you would pay particular attention?

4. What would be the probable effect on a valve in a power oscillator when the grid-leak resistor is open circuit? Give reasons for your answer?

5. What is meant by the terms -

   (i) Series feed oscillator?
   (ii) Parallel (or shunt) feed oscillator?
   (iii) Negative resistance?
   (iv) Tank circuit?

END OF PAPER.

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COURSE OF TECHNICAL INSTRUCTION.

RADIO TRANSMITTERS.

CONTENTS:

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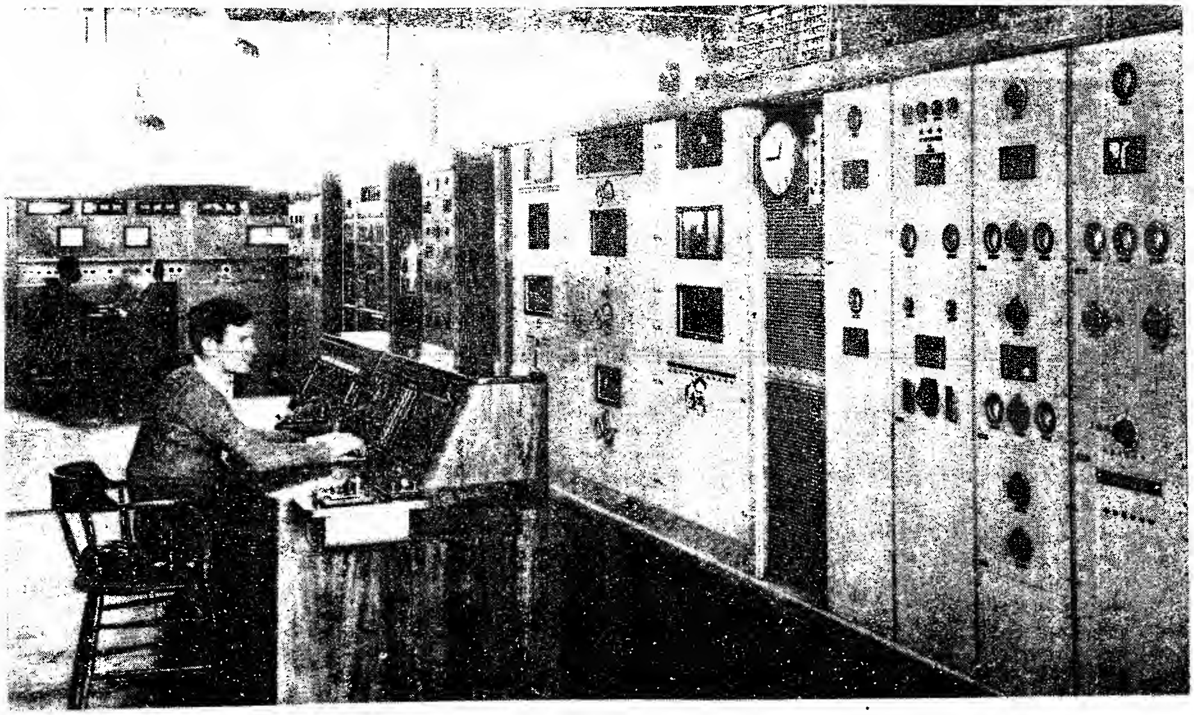
1. INTRODUCTION.

1.1 Previous Papers have dealt with Radio-Frequency Amplifiers and Oscillators. Radio Transmitters are combinations of these items designed to furnish predetermined amounts of power to radiating systems. For relatively low-power Transmitters (10 kW), the components required are of simple design, but, for those greater than 10 kW, the voltages and currents in the final amplifiers of the system necessitate careful design and workmanship of all components.

The basic circuits shown in the previous Papers, however, still apply. Apart from actual construction and wiring differences, only the power ratings and breakdown voltages of the components are different. For example, in a 10 kW transmitter, the peak power at full modulation is approximately 40 kW, which is a reasonable amount to deal with, but, in a 100 kW transmitter, the peak power is in the vicinity of 400 kW, which imposes a relatively large strain on the components of the circuit.

1.2 Radio Transmitters. Transmitters are conveniently discussed as follows -

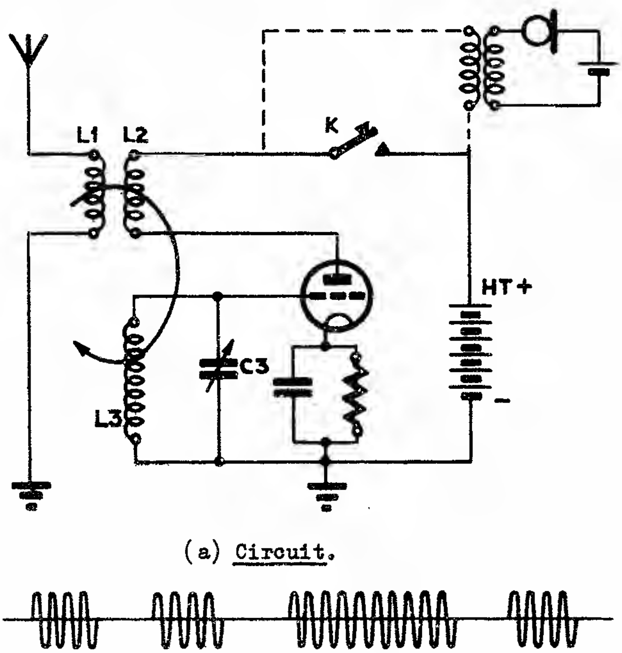
Fundamental or Elementary Type Transmitter.
Some Design Considerations.
Telegraph Transmitters.
Broadcast (or Telephony) Medium-Frequency Transmitter.
Broadcast (or Telephony) High-Frequency Transmitter.
Transmitter Adjustments.



VLG AND VLR RADIO TRANSMITTERS.

2. ELEMENTARY TRANSMITTER.

2.1 Any of the oscillators described previously become an elementary transmitter by suitably coupling an aerial system into the circuit. Fig. 1a shows a grid-tuned oscillator functioning as a Radio transmitter.



(a) Circuit.

(b) Code Signals.

FIG. 1. ELEMENTARY RADIO TRANSMITTER.

When the key is depressed, the circuit is completed for the anode voltage supply to the oscillator. Oscillations thus commence at a frequency determined by the values of L3 and C3. L2 and L3 comprise the feedback circuit for maintaining oscillations, and L2 and L1 form the output circuit, the impedance of the aerial circuit acting as the load on the valve. As long as the key is depressed, energy is transferred to the aerial and radiated therefrom. Messages are thus sent in Morse Code by pulses of radio-frequency waves in the form of long or short trains of oscillations governed by the operation of the key, as in Fig.1b.

A transmitter of this type, however, has a number of disadvantages, chief among which is frequency instability due to the variable load provided by the aerial system. However, this elementary transmitter serves to show the principles of transmission, and from this basic circuit the larger units are built up. For telephony, the key is replaced by the

/ secondary

secondary of a transformer having a microphone circuit in the primary. The results are not very satisfactory, but the illustration shows how Voice Modulation is obtained in lieu of Telegraphy.

2.2 The next step to improve the above transmitter is to use an amplifier between the oscillator and the aerial. This system is shown in Fig. 2, and has a decided advantage over the self-excited oscillator transmitter in that the frequency stability of the transmitter is greatly improved. This improvement is due primarily to the fact that, in this system, the oscillator is not coupled directly to the aerial and, therefore, is unaffected by any changes in the aerial-to-ground capacitance. The master-oscillator in this particular circuit is of the Hartley type, although any of the conventional oscillators previously explained can be used with equally good results. Maximum efficiency is not necessary in the master-oscillator which is designed to give maximum stability. In the power amplifier, however, circuits are designed to give maximum efficiency.

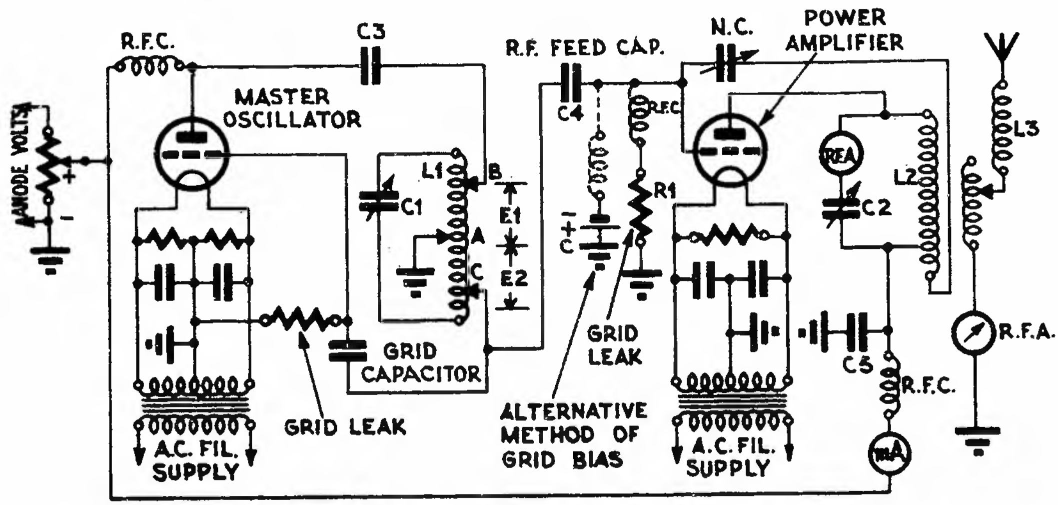


FIG. 2. MASTER-OSCILLATOR POWER-AMPLIFIER TRANSMITTER.

When the aerial-to-ground capacity varies, the reaction upon the power-amplifier circuit merely results in a decrease in the radiation efficiency, owing to the change in the resonance conditions between these two circuits. In practice, this decrease is usually unnoticeable. The frequency stability of the transmitter is, therefore, much improved by using the master-oscillator power-amplifier system in preference to the simple oscillator alone as a transmitter.

The master-oscillator power-amplifier system is commonly referred to as the M.O.P.A. transmitter. The circuit shown here is used on any frequency to which the oscillator (or "driver") is tuned.

The advantages of an Oscillator-Amplifier Type Transmitter over a Self-Excited Oscillator are summarised as follows -

(i) The efficiency of a properly-excited amplifier is greater than that of a self-excited oscillator, because the amplifier's grid losses are supplied from a separate source (that is, the oscillator).

(ii) The frequency stability is better, because the oscillator is isolated from the aerial system and, therefore, is little affected by changes in the aerial constants.

(iii) Changes in the amplifier anode-supply voltage have negligible effect on the frequency of the oscillator, so that anode-supply ripple does not result in frequency modulation, and, when the amplifier valve is keyed, a key filter with a large time constant can be used without causing a keying chirp.

/ The

The frequency of the oscillations in Fig. 2 is determined by the circuit constants C1, L1, known as the tank circuit because it acts as a reservoir of radio-frequency energy. The amount of anode and grid reactance is determined by the number of turns between the points AB and AC, respectively. These reactances govern the degree of radio-frequency grid-voltage excitation E2 across the grid-cathode of the oscillator valve.

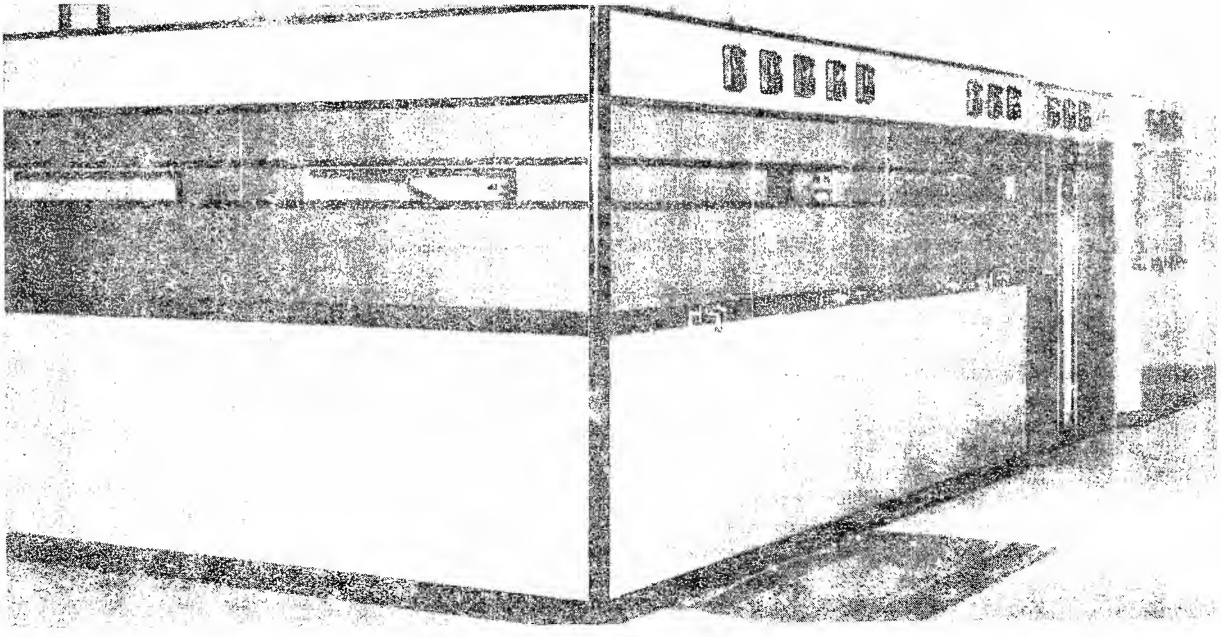
When the valve is oscillating at a definite frequency as determined by the constants L1, C1, the input (grid-cathode) to the power-amplifier valve is connected across a portion of the inductance L1 so that it is properly excited. In other words, the high-frequency e.m.f. built up across the grid section of the oscillator coil AC is applied across the input to the amplifier valve through the radio-frequency feed capacitor C4 and through the common ground connection to the cathodes.

The proper amount of grid excitation across the amplifier valve is determined by the reactance of the feed capacitor C4 and the point at which the amplifier grid connection is made on the oscillator coil. The power amplifier is adjusted to the proper grid and anode voltage in accordance with the class of amplification in which the valve is to be operated, namely, Class A, B or C (see Paper No. 4).

The output circuit, L2, C2, is now adjusted to resonance with the master-oscillator by an indication of the minimum current in the radio-frequency ammeter, R.F.A. in this circuit, or by a maximum dip in the reading of the anode milliammeter, mA, after it has first been neutralised to prevent self-oscillation by properly adjusting the neutralising capacitor, N.C. (Neutralising is explained in later paragraphs of this Paper.)

The aerial circuit is then tuned to resonance by adjusting the coupling and the tuning inductor L3 for a maximum indication on the aerial radio-frequency ammeter.

Capacitor C3 is the anode-blocking and radio-frequency by-pass capacitor combined. C5 is the power-amplifier radio-frequency by-pass capacitor, which serves as a low-reactance path for the radio frequencies in the anode circuit to flow through to the cathode. C5, in conjunction with the radio-frequency choke, isolates the high-frequency currents from the power-supply source, thereby increasing the efficiency and stability of the amplifier circuit.



MODERN RADIO TRANSMITTER. (R.C.A., U.S.A.). (50 kW SHORT WAVE TRANSMITTER, SHEPPARTON.)

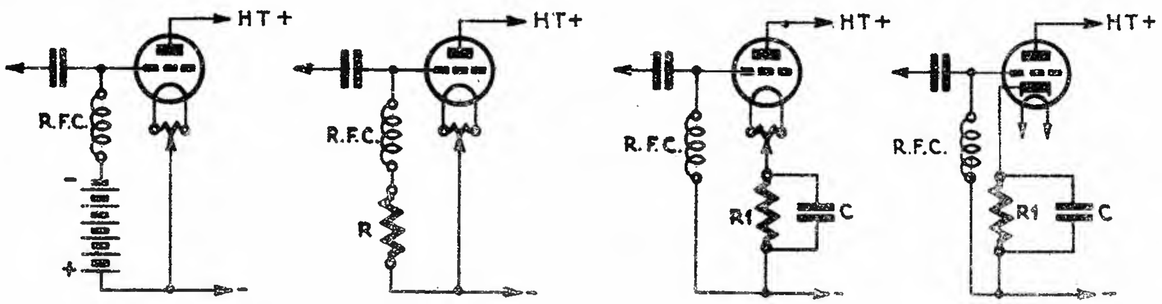
3. DESIGN CONSIDERATIONS.

3.1 Before giving further details of transmitters, several design and operational aspects common to most types are discussed. These are -

Amplifier Biasing Circuits.
Interstage Coupling Circuits.
Grid and Anode Circuits.
Neutralising Circuits.
Amplifier Adjustments.
Optimum Amplifier Output Conditions.
Aerial Coupling Circuits.
Protection (Equipment and Personnel).

General considerations only are discussed, as individual transmitters require individual treatment. Differing designs and applications require special adjustments and operational treatment, but these notes give a general idea of some of the more important points.

3.2 Amplifier Biasing Circuits. (Amplification of some remarks made in a short reference to biasing methods in a previous Paper.) It is necessary to make provision in the circuit for the application of proper negative grid bias voltage to the grid of the amplifier valve. (See Figs. 3a, b and c.)



(a) Battery Bias. (b) Grid Leak. (c) Cathode Biasing.

FIG. 3. BIASING CIRCUITS.

Transmitting valve tables issued by the valve manufacturers indicate the value of grid bias which is used under representative operating conditions for normal anode voltage. When the valve is operated at an anode voltage other than that indicated, the grid bias is increased or decreased accordingly. In Fig. 2, the amplifier is biased by the fall of potential across the resistor R1, which is in series with the grid radio-frequency choke. The purpose of this choke is to prevent the radio-frequency excitation current from leaking to earth through the grid leak. This method is called grid leak or automatic biasing. An alternative method of biasing is shown by the dotted lines, where a battery of the correct value is used to provide the negative grid bias. This method is known as battery or fixed biasing.

Battery Bias. Batteries have the advantage of giving practically constant voltage under all conditions of excitation or lack of it, although the grid current flow does have a charging effect which tends to raise the battery voltage. This effect increases as the internal resistance of the battery increases.

Besides the constant-voltage feature with varying grid currents, battery bias also protects the amplifier valve or valves in case of excitation failure. This advantage is common to all fixed bias systems (as contrasted with biasing systems which depend upon the flow of grid current). When excitation stops with anode voltage applied, the anode current drops to zero or a low value, depending upon whether the amplifier is biased beyond or near to cut-off.

/ Grid-leak

Grid-leak Bias. Grid-leak bias is economical, as batteries are not necessary, and has the desirable feature that the bias regulates itself in accordance with the amount of excitation available, thereby tending to give optimum amplifier operation under varying conditions of excitation. Without excitation, however, there is no grid bias, and, in the case of valves operating at fairly high voltages, especially those having low and medium values of amplification factor, a large anode current flows when the excitation fails while the anode voltage is connected to the valve. This may seriously damage the valve if not quickly corrected.

The advantages of battery and grid-leak bias are secured and their disadvantages eliminated by using a combination of both. Many circuits have sufficient battery bias to reduce the amplifier anode current to a safe value should excitation fail, and connected in series with the battery is a grid leak to obtain the additional bias needed under operating conditions. In general, the leak values recommended in the valve table are used without change when used in conjunction with a small amount of "safeguarding" battery bias. The bias power pack, when properly designed, offers the advantages of a battery grid-leak combination.

When grid-leak bias is used, the bias under operating conditions is calculated by multiplying the leak resistance by the grid current in amperes. When a battery is in series with the leak, the battery voltage is added to the voltage obtained by the calculation.

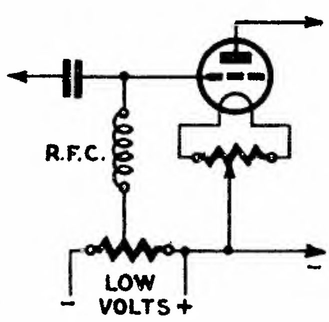
Cathode Biasing. Transmitting valves are often biased by the method universally used in receivers - by inserting a resistor of suitable value in series with the cathode and using the voltage drop resulting from the flow of anode current through the resistor as bias. Correctly applied, this method of biasing combines the self-adjusting features of grid-leak bias with some measure of protection to the valve in case of excitation failure. The voltage drop in the biasing resistor causes a reduction in anode voltage by the same amount, however, so that the anode supply is designed to have a terminal voltage equal to the desired operating anode voltage plus the grid bias voltage. For a valve intended to be operated with 1,000 volts on the anode and 200 volts negative bias, for example, the anode supply voltage should be 1,200 volts.

The value of the biasing resistor, R1, in Fig. 3c is calculated for normal operation. For example, when the valve is rated to draw 130 mA anode current and 20 mA grid current, and requires 150 volts grid bias, the resistance required is -

$$R = \frac{E}{I} = \frac{150}{0.15} = 1{,}000 \text{ ohms.}$$

The anode current which flows without excitation is found from the valve's characteristic curves. Assume some value of anode current, calculate the bias resulting and then check with the curves to see if that particular bias value will cause the assumed anode current to flow, keeping in mind the reduction in anode voltage due to bias. The power input is then determined to make sure it is below the safe anode dissipation rating of the valve.

Other Methods. When conditions permit, power amplifier bias is obtained from the anode supply used for low-power stages. A diagram showing the essentials is given in Fig. 4. The low-voltage and high-voltage supplies are not connected together at negative, as is usually the case, but the positive terminal of the low-voltage supply is connected to negative high-voltage. Since this places the cathode of the power amplifier and the cathodes of the low-voltage valves at opposite terminals of the low-voltage supply, a separate well-insulated transformer winding must be used for the amplifier filament. The tapped resistor, used to permit varying the bias, is the regular bleeder on the low-voltage supply.



LOW VOLTAGE ANODE SUPPLY FOR BIASING A POWER AMPLIFIER.

FIG. 4.

3.3 Interstage Coupling Circuits. Interstage couplings for audio-frequency circuits are dealt with in Paper No. 3 of this book. The requirements for radio-frequency circuits are somewhat different, due to the frequencies involved and also because, in most cases, the input and output circuits are tuned. The problem of coupling is further complicated by the characteristics of different types, and the various methods of coupling are used to meet effectively the varying character of the impedance. Some information on this subject appears earlier, and this is amplified here.

Coupling methods are divided into three different classes -

(i) Capacitive or Direct Coupling.
(ii) Inductive or Indirect Coupling.
(iii) Transmission Line or Link Coupling.

Capacitive or Direct Coupling. In this method, the grid circuit of an amplifier circuit is coupled to a preceding driver stage by means of a fixed or variable capacitor as shown in Fig. 5.

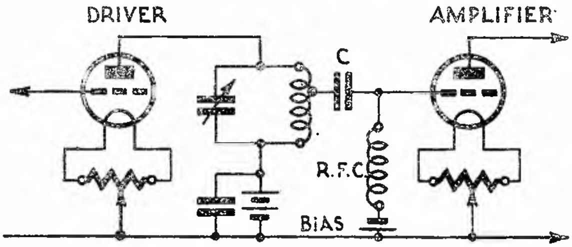


FIG. 5. CAPACITIVE COUPLING.

Capacitor C isolates the D.C. power supply from the grid of the amplifier valve and provides a low impedance path for the radio-frequency energy between the valve being driven and the driver valve. This method of coupling is simple and economical for low-power amplifiers or exciter stages, but has certain disadvantages for the coupling of a final amplifier to the preceding stage. The grid circuit leads in a neutralised amplifier must be as short as possible, but this is difficult to attain in the physical arrangement of a high-power amplifier with respect to a capacitively coupled driver stage. The radio-frequency choke in series with the bias supply lead must offer an extremely high impedance to the radio-frequency current, and this too is difficult to obtain when the transmitter is operated on several harmonically related bands. Another disadvantage of capacitive coupling is the difficulty of adjusting the load on the driver stage. Impedance adjustment can be accomplished by tapping the coupling lead a part of the way down on the coil of the tuned stage of the driver anode circuit. However, when this lead is tapped part way down on the coil, parasitic oscillations (spurious oscillations at a very high frequency) become very troublesome and are difficult to eliminate. If the driver stage has sufficient power output so that an impedance mismatch can be tolerated, the coupling capacitor is connected directly to the anode end of the coil and is made small enough in capacitance (for the particular frequency of operation) that not more than normal anode current flows.

The impedance of the grid circuit of a Class C amplifier is as low as a few hundred ohms in the case of a high mu valve, and ranges from that value up to a few thousand ohms for a low mu valve.

Capacitive coupling places the grid-to-cathode capacity of the driven valve directly across the driver tuned circuit, which reduces the L-C ratio and sometimes makes the radio-frequency amplifier difficult to neutralise, because the additonal driver stage circuit capacitances are connected into the grid circuit.

Capacitive coupling is used to advantage in reducing the total number of tuned circuits in a transmitter to conserve space and cost. It is also used between stages for driving pentode amplifiers, as these valves normally require relatively small amounts of grid excitation.

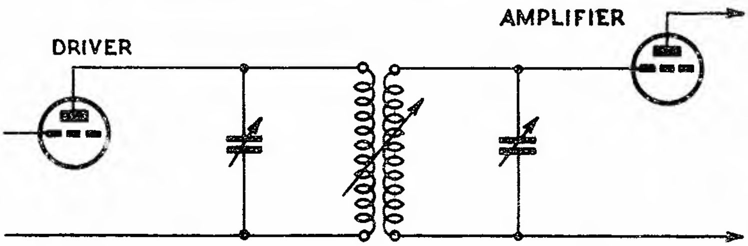


FIG. 6. INDUCTIVE COUPLING.

Inductive or Indirect Coupling. The driver stage is sometimes coupled to the amplifier by means of inductive coupling, which consists of two coils electromagnetically coupled to each other, as shown in Fig. 6.

The degree of coupling is controlled by varying the mutual inductance of the two coils which is accomplished by changing the spacing between the coils.

Inductive coupling is used extensively for coupling radio-frequency amplifiers in radio receivers and occasionally in radio-frequency transmitter circuits. The mechanical problems involved in adjusting the degree of coupling in a transmitter, however, makes this system of limited practical value in these circuits.

Transmission Line or Link Coupling. A special form of inductive coupling, which is applied to radio transmitter circuits, is known as link coupling. A low impedance radio-frequency transmission line, commonly known as a "link", couples the two tuned circuits together. Each end of the line is terminated in one or more turns of wire or loops wound around the coils to be coupled together. The loops are coupled to each tuned circuit at a point of zero radio-frequency potential. This nodal point is at the centre of the tuned circuit in the case of anode neutralised push-pull amplifiers, and at the positive "B" end of the tuned circuit in the case of screen-grid and grid neutralised amplifiers. The link coupling turns are as close to the nodal point as possible. A typical circuit is shown in Fig. 7.

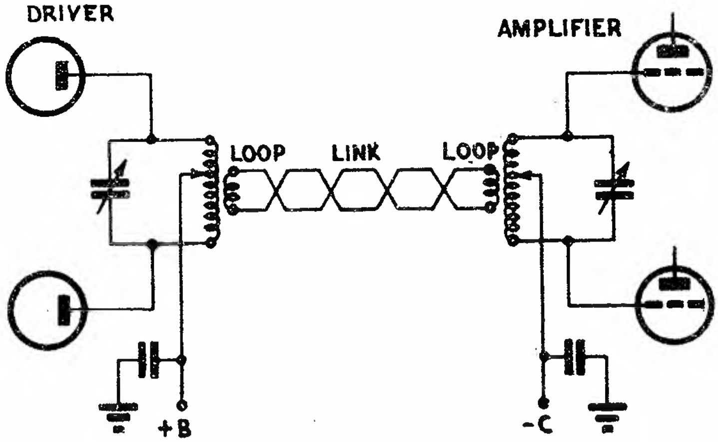


FIG. 7. LINK COUPLING CIRCUIT.

Some of the advantages of link coupling are -

(i) Eliminates coupling taps on tuned circuits.

(ii) Permits the use of "series" power supply connections in both tuned-grid and tuned-anode circuits, and thereby eliminates the need for radio-frequency chokes.

(iii) Allows separation between transmitter stages of distances up to several feet without appreciable radio-frequency losses.

(iv) Reduces capacitive coupling and thereby makes neutralisation more easily attainable in radio-frequency amplifiers.

(v) Provides semi-automatic impedance matching between anode and grid-tuned circuits, with the result that as much as 50 per cent. greater grid swing is obtained in comparison with capacitive coupling.

The impedance of a link coupling line varies from 75 to 200 ohms, depending upon the diameter and spacing of the conductors. For example, the impedance of a two-wire coupling line is determined from the following general formula -

$$Zo = 276 \log_{10} \frac{D}{r}$$

where Zo = Characteristic impedance in ohms,

D = Distance between centres of wires in centimetres, and,

r = Radius of wire in centimetres.

The above is one of a number of formulae used for calculating the impedance of various types of transmission lines, and is included for the purpose of illustration.

3.4 Grid and Anode Circuits. Reference was made in Paper No. 8 to the types of connections known as shunt and series-feed. These terms are also applied to distinguish the grid and anode connections of amplifier valves.

Shunt and Series-Feed. D.C. grid and anode connections are made either by shunt (parallel) feed or series-feed systems. Simplified forms of each are shown in Fig. 8.

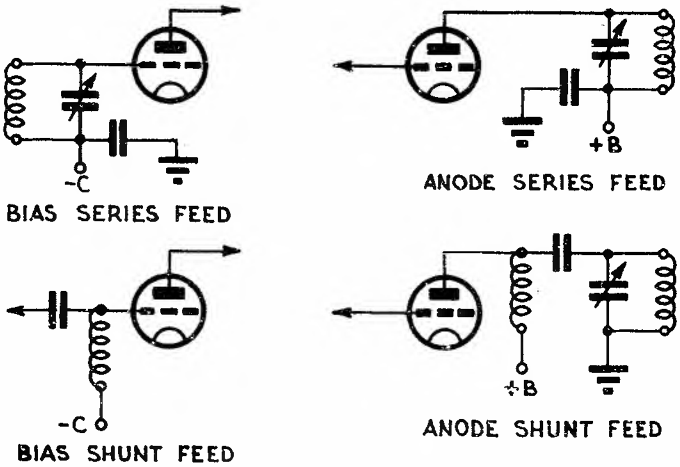


FIG. 8. D.C. GRID AND ANODE CIRCUITS.

Series-feed is defined as that in which the D.C. connection is made to the grid or anode circuit at a point of very low radio-frequency potential.

Shunt feed is always made up to a point of high radio-frequency voltage, and always requires a high impedance radio-frequency choke or resistance in the connection to the high radio-frequency point, in order to prevent loss of radio-frequency power.

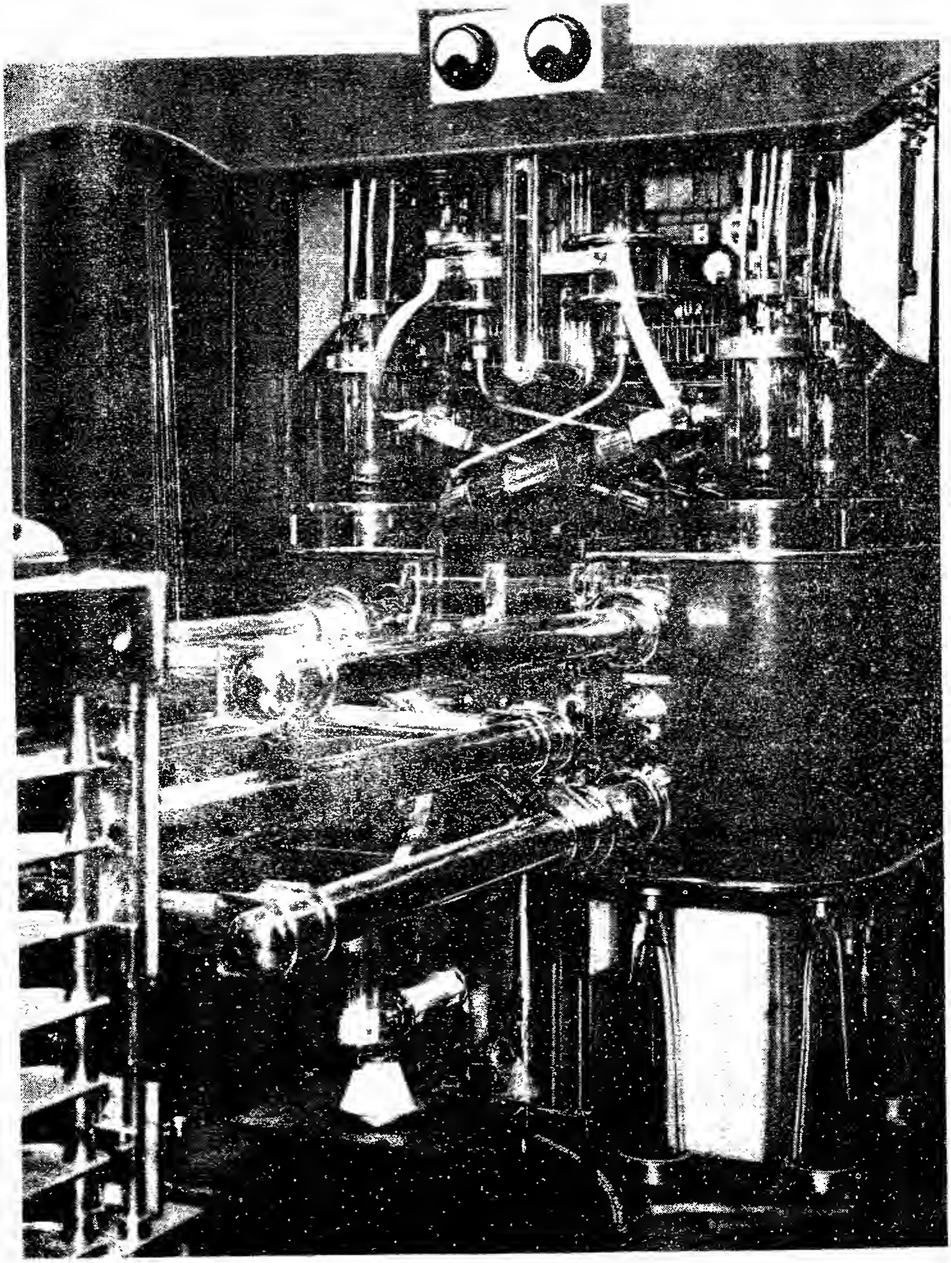
Series feed is used more than shunt feed, since radio-frequency choke coils are usually omitted and the loss of radio-frequency energy is minimised.

3.5 Neutralising Circuits. The principles of neutralising are referred to in Paper No. 4, and basic circuits are given. As this is a very important aspect of design and operation, some further details are now given.

As already explained, the process of neutralising consists of taking some of the radio-frequency voltage from the output or input circuit of the amplifier and introducing it into the other circuit in such a way that it effectively "bucks" the voltage operating through the grid-anode capacitance of the valve, thus rendering it impossible for the valve to supply its own excitation. For complete neutralisation, it is necessary, therefore, that the neutralising voltage is opposite in phase to the voltage through the grid-anode capacitance of the valve and equal to it in value.

The out-of-phase voltage is obtained quite readily by using a balanced tank circuit in either grid or anode, taking the neutralising voltage from the end of the tank opposite that to which the grid or anode is connected. The amplitude of the neutralising voltage is regulated by means of a small capacitor, the "neutralising capacitor" having the same order of capacitance as the grid-anode capacitance of the valve. Circuits in which the neutralising voltage is obtained from a balanced anode tank and fed to the grid of the valve through the neutralising capacitor are called "anode-neutralising" circuits. When the neutralising voltage is obtained from a balanced grid tank and fed to the anode of the valve, the circuit is known as a "grid neutralising" circuit.

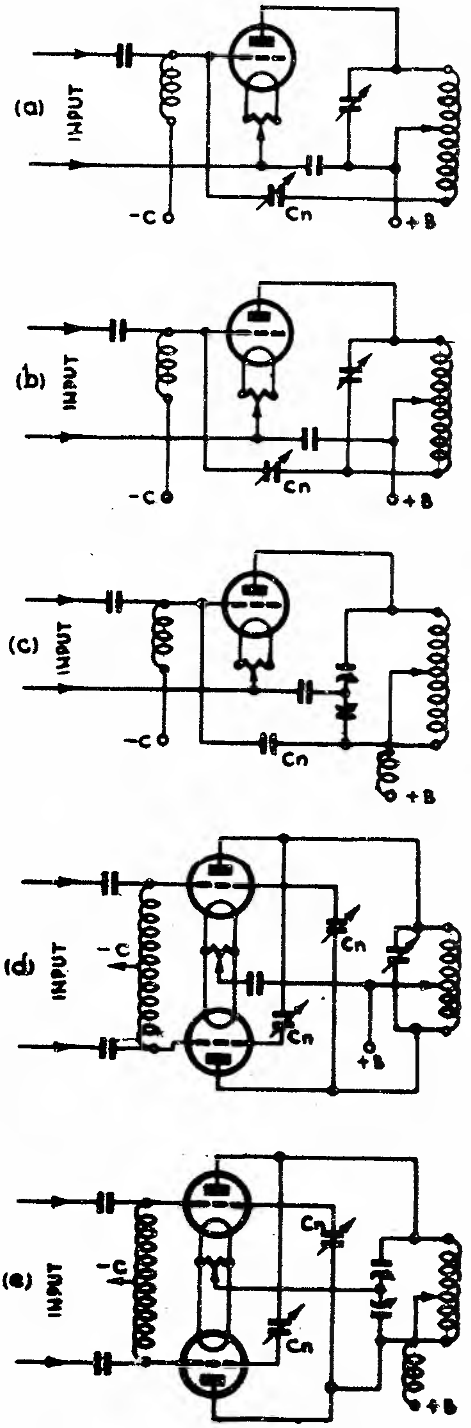
A neutralising circuit is actually a form of bridge circuit, the grid-anode capacitance of the valve and the neutralising capacitor forming two capacitive arms, while the halves of the balanced tank circuit form the other two arms.



(NOTE: Criss-cross neutralising capacitors. Four valves connected in parallel push-pull are used in this amplifier.).

OUTPUT COUPLING CIRCUIT OF A 100 kW RADIO TRANSMITTER. (SHEPPARTON.)

Anode-Neutralising Circuits. Several anode-neutralising circuits are given in Fig. 9. In the circuit shown the tank coil is centre-tapped, with the tank capacitor connected across only the upper half of the coil. The neutralising portion of the coil is connected back to the grid of the valve through the neutralising capacitor, Cn.



ANODE NEUTRALISING CIRCUITS.

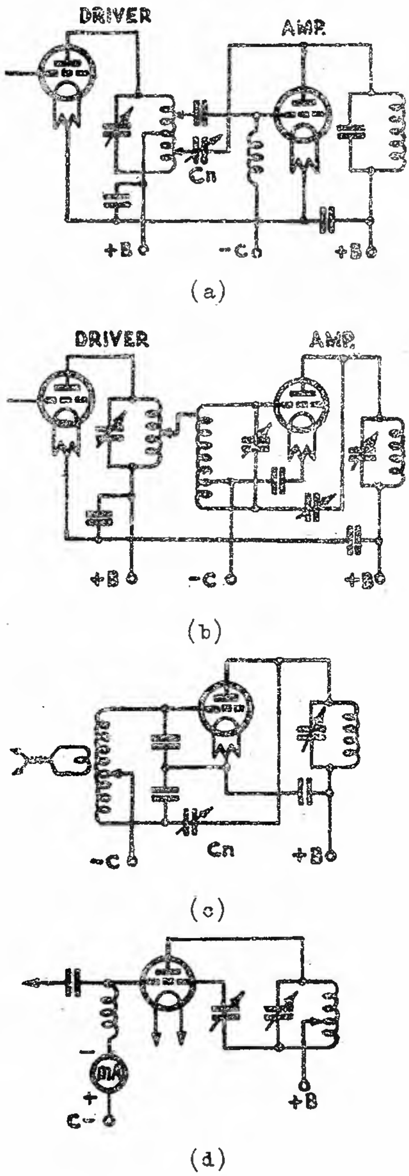
FIG. 9.

The circuit of Fig. 9b is similar, differing, however, in that the tank capacitor is connected across all of the tank coil. This method of connection is preferable as it tends to keep a battery voltage balance over a range of frequencies. The only reason for using the circuit of Fig. 9a is to get as high an impedance as possible in the part of the tank circuit included between the cathode return and anode for the sake of efficiency.

In both the circuits already described, the division of radio-frequency voltage between anode and neutralising portions of the circuit is obtained by balancing the tank coil. In Fig. 9c, the balance is capacitive, by the use of a split-stator tank capacitor with grounded rotor. The radio-frequency potential across the tank coil divides in the same way, a node (point of zero voltage) appearing at its centre. Hence, the anode voltage is introduced at the centre of the coil. The radio-frequency choke in the anode voltage lead is for the purpose of isolating the centre of the coil from earth for radio-frequency, since an earth through a by-pass capacitor, when not exactly at the point of zero potential, often causes circulating currents which reduce the anode efficiency of the amplifier.

The push-pull neutralising circuits shown in Figs. 9d and 9e are known as "cross-neutralised" circuits, the neutralising capacitors being cross-connected from the grid of one valve to the anode of the other. With the proper physical arrangement of parts, a more exact balance is obtained with push-pull than with a single valve, because both sides of the circuit are symmetrical. Hence, these circuits are often easier to neutralise than single-valve circuits. The split capacitor circuit of Fig. 9e is preferred for push-pull amplifiers.

Grid Neutralisation. Typical grid-neutralising circuits are shown in Fig. 10. These circuits resemble closely the anode-neutralising circuits, except that the neutralising voltage is obtained from a balanced input tank and fed to the anode of the valve. The circuit shown in Fig. 10a is used with capacity coupling between driver and amplifier. The grid coupling capacitor, being large in comparison with the valve and neutralising capacities in most circuits, has negligible effect on the operation of the neutralising circuit.



GRID NEUTRALISING CIRCUITS.

FIG. 10.

Grid neutralising systems are well adapted for use with transmission line or link-coupled amplifiers, since the separate grid tank offers a ready means for obtaining the neutralising voltage. It is somewhat harder to drive a valve with a balanced input tank, however, because only half the radio-frequency voltage developed in the tank is available for the grid-cathode circuit of the amplifier. This can be overcome to some extent by using the largest possible L-C ratio in the grid tank, in order to build up the radio-frequency voltage to the highest possible value. An advantage of the grid-neutralising systems is the fact that the single-ended anode tank circuit has high impedance, and hence gives greater anode efficiency than a balanced anode tank in which the anode-cathode circuit is connected across only half the turns or half the capacitance.

Neutralising Adjustments. The procedure in neutralising is the same regardless of the valve or circuit used. To neutralise satisfactorily, it is essential that some form of fairly sensitive radio-frequency indicator be available. A flashlight lamp with its terminals connected to a loop of wire, a neon bulb or a thermo-galvanometer connected to a wire loop are suitable.

The first step in neutralising is to disconnect the anode-voltage from the valve being neutralised. The cathode of the valve is heated, however, and the excitation from the preceding stage is fed to its grid circuit. Couple the radio-frequency indicator to the anode tank circuit (when a neon bulb is used, simply touch the metal base to the anode terminal), and tune the anode circuit to resonance, which is indicated by a maximum reading of the radio-frequency indicator. Then, find the setting of the neutralising capacitor which makes the radio frequency in the anode tank drop to zero. Turning the neutralising capacitor probably alters the tuning of the driver tank slightly, so the preceding stage must be retuned to resonance.

Now couple the radio-frequency indicator to the anode tank once more and again tune the anode circuit to resonance. Probably the resonance point occurs at a slightly different setting, and the second reading on the radio-frequency indicator is lower than the first one. Retune the preceding stage once more and go through the whole procedure again. Continue until the radio-frequency indicator gives no reading, when the anode tank circuit is tuned in the region of resonance. When this has been accomplished, the valve is neutralised.

/ The

The aim of neutralising adjustments is to find the setting of the neutralising capacitor which eliminates radio frequency in the anode circuit when the anode circuit is tuned to resonance. It sometimes happens that, while a setting of the neutralising capacitor is found which gives a definite point of minimum radio frequency in the anode circuit, the radio frequency is not completely eliminated. This is probably due to stray coupling between the amplifier and driver tank coils, or stray capacitances between various parts of the amplifier circuit tending to upset the voltage balance. A better layout with short, widely-spaced leads, or with coils so placed that coupling between them is minimised - usually when the axes of the coils are at right angles - should be tried. Shielding of the amplifier often eliminates troubles of this sort.

Neutralising Indicators. In the neutralising procedure outlined above, the use of a neon bulb or other radio-frequency indicator has been assumed. In circuits in which the neutralising bridge is entirely capacitive, as in those circuits using split-stator capacitors, touching the neon bulb to a high-potential point of the circuit often introduces sufficient stray capacity to unbalance the circuit slightly, thus upsetting the neutralising. This is particularly noticeable with high-power amplifiers, where the excitation voltage is considerable and a slight unbalance gives an indication. In such cases, a flashlight lamp and loop of wire tightly coupled to the tank coil gives a more accurate indication of the exact neutralising point. A thermo-galvanometer similarly connected to a wire loop has considerably greater sensitivity.

A D.C. milliammeter connected to read rectified grid current, Fig. 10d, makes a sensitive neutralising indicator. When the circuit is not completely neutralised, tuning the anode tank circuit through resonance changes the tuning of the grid circuit and affects its loading, causing a change in the D.C. grid current. With push-pull amplifiers, or single-ended amplifiers using a tap on the tank coil for neutralisation, the setting of the neutralising capacitor, which leaves the grid current unaffected as the anode tank is tuned through resonance, is the correct one. When the circuit is slightly out of neutralisation, the grid meter needle gives a noticeable flicker. With single-ended circuits having split-stator neutralisation, the behaviour of the grid meter depends upon the type of valve used. When the valve's output capacitance is not great enough to upset the balance, the action of the meter is the same as in other circuits. With high-capacitance valves, however, the meter usually shows a gradual rise and fall as the anode tank is tuned through resonance, reaching a maximum right at resonance when the circuit is properly neutralised. A sharp flicker at resonance shows that the circuit is not neutralised.

3.6 Amplifier Adjustments.

Efficiency and Output. The attainable anode efficiency - the ratio of radio-frequency output to D.C. input - is of great importance in determining the operating conditions for the amplifier in a radio transmitter. When the safe anode dissipation rating of the valve is the only consideration, it is desirable to obtain the highest possible anode efficiency, since the power output is limited solely by the efficiency. For example, a valve having an anode dissipation rating of 100 watts operating at an anode efficiency of 90 per cent. could handle a D.C. input of 1,000 watts, giving 900 watts radio-frequency output, while the same valve at 70 per cent. efficiency could handle a D.C. input of only 333 watts, giving an output of 233 watts radio frequency. The anode dissipation - the difference between input and output - is the same in both cases, 100 watts.

There are other considerations, however, which limit the useful anode efficiency. Assuming that the anode input is not to exceed the manufacturer's ratings for the valve, the difference between 70 per cent. and 90 per cent. efficiency is not so great.

/ For

For instance, taking the same 100 watt valve and assuming that the 70 per cent. efficiency condition corresponds with the ratings, an efficiency of 90 per cent. increases the output to only 300 watts (333 watts input). The additional 67 watts of output, an increase of about 27 per cent., requires excessively large driving power, because, as shown by Fig. 11, the efficiency increases very slowly beyond the optimum point, while the reverse is true of the driving power required.

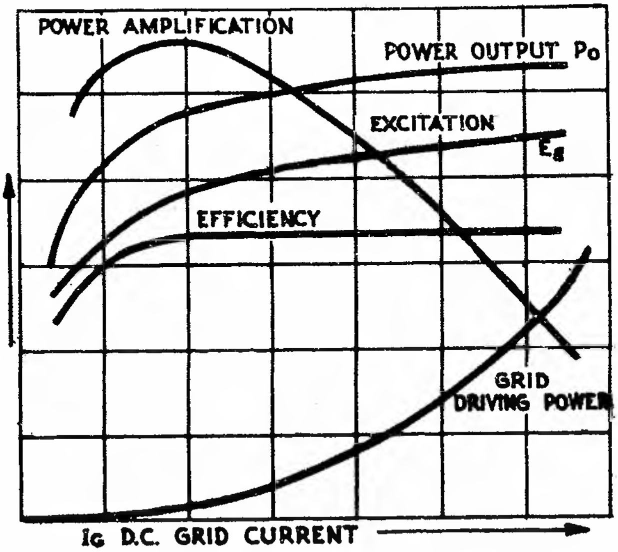


FIG.11. EFFICIENCY AND DRIVING POWER INCREASE.

A second factor which limits the usable efficiency is the fact that high values of efficiency are attained only through the use of high values of load resistance, which, in turn, require the use of very high anode voltage. Not all valves are suited to operation at anode voltages much above normal, while from an economic standpoint a high-voltage power supply represents greater cost than the installation of a second valve operating at lower voltage to give the same order of power output at lower anode efficiency.

Most valves are designed for operation as radio-frequency power amplifiers under average conditions, where the anode efficiency is in the vicinity of 70 per cent. This corresponds to the optimum point on the curves of Fig. 11. These valves deliver their rated power output at moderate anode voltages, considering the size of the valve, and with fairly low values of driving power. A few valves available, however, are operated at relatively high anode voltages and are provided with oversize cathodes to withstand high-voltage operation. Such valves are operated at moderate anode voltages with normal efficiency or used for obtaining large power outputs at high anode efficiency.

Impedance Matching. A load absorbs the most power from a generator source when its impedance "matches" the impedance of the supply source or generator. First take the case of a generator with internal resistance "r" feeding a pure resistance load "R" in a non-reactive circuit as shown in Fig. 12.

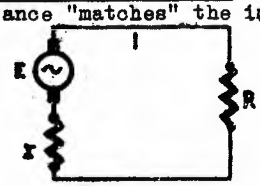


FIG. 12.

When the e.m.f. of the source is E, the power dissipated in the load R is given by -

$$P = I^2R$$

where the current $I = \dfrac{E}{r + R}$

Now, substituting this value for I -

$$P = \left(\frac{E}{r + R}\right)^2 R = \frac{E^2R}{r^2 + 2rR + R^2}$$

$$= \frac{E^2R}{(r^2 - 2rR + R^2) + 4rR} = \frac{E^2R}{(r - R)^2 + 4rR}$$

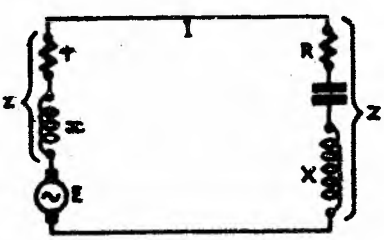


FIG. 13.

This expression has a maximum value when its denominator is a minimum, that is, when $r - R = 0$, or $R = r$.

Thus, the output power is a maximum when the load resistance equals the internal resistance of the supply generator. Where the circuit is reactive, it must be adjusted to a non-reactive or resonant condition for correct matching when $R = r$.

Consider Fig. 13 where -

the generator impedance $z = \sqrt{r^2 + x^2}$

and the load impedance $Z = \sqrt{R^2 + X^2}$

/ Suppose

Suppose z to be fixed, and determine the value of Z for the maximum output. Now, output power is given by -

$$P = I^2R$$

$$\text{but } I = \frac{E}{\text{Total } Z}$$

$$= \frac{E}{\sqrt{(r + R)^2 + (x + X)^2}}$$

$$\text{and } I^2 = \frac{E^2}{(r + R)^2 + (x + X)^2}$$

$$\text{giving } P = \frac{E^2R}{(r + R)^2 + (x + X)^2}$$

This expression shows that, as far as X is concerned, the output (P) is a maximum when X = -x. Thus, if x is (+) inductive, X is (-) capacitive.

When both magnitude and phase angle of the load impedance are adjustable, then the load or receiving impedance is equal in resistance component and equal and opposite in reactive component. When only the magnitude of the load is adjustable, then the absolute value of the load impedance equals the absolute value of the generator (that is, Z = z). In power work, the matching of impedance to secure the greatest possible power in the load is seldom necessary, as the overall efficiency under these maximum output conditions is only 50 per cent. Moreover, the voltage regulation is also 50 per cent., only half of the generator voltage being available across the load. Power generators are usually rated on a load which gives a comparatively small voltage difference between open and closed circuit conditions, the limits being the permissible heating and the efficiency required.

In radio circuits, on the other hand, where the cost of power is small in proportion to the total expense and the heating losses in conductors are easily minimised, extensive use is made of the fact that the matching of the load assures the maximum transfer of energy. Voltage regulation is not as necessary in radio circuits, where any voltage condition can be catered for, as in telephone power circuits, where the equipment operates within narrow voltage limits.

An important method of impedance matching is the use of parallel resonant circuits to alter the magnitude of a load resistance. Tuned parallel circuits are frequently used as loads for thermionic valves, and such valves correspond to generators with a high internal resistance. When the load is a "tank" circuit consisting of an inductance L henry with resistance R ohm in parallel with a capacitance C farad, its impedance at resonance is given by -

$$Z = \frac{L}{RC}$$

and it is non-reactive.

<u>Use of Anode-Tap to Adjust for Optimum Output Load.</u> When the ratio of L to C provides an unduly high impedance output, it is possible to adjust a tapping on the anode inductance, as shown in Fig. 14, until the load resistance correctly matches the valve's internal resistance.

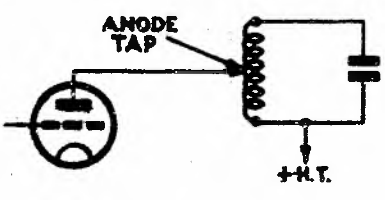


FIG. 14.

This principle of impedance matching is the most widely applied characteristic of parallel resonant circuits, and, without this, it is impossible to obtain appreciable power from transmitting valves at high frequencies.

Parallel and Series Resonant Circuits compared as Matching Impedances. The impedance of a parallel resonant circuit at resonance is given by -

$$Z_r = \frac{X_c^{\ 2}}{R} = \frac{X_L^{\ 2}}{R} = \frac{X_L X_c}{R} = \frac{L}{RC}$$

which is always a high impedance for the particular resonant frequency when R is small.

On the other hand, a series resonant circuit has an impedance $Z_r = R$ at resonance, and this may conveniently be made a very low value for the particular desired frequency.

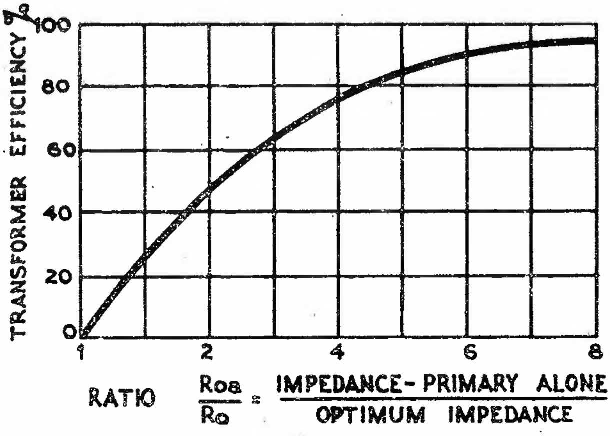
The use of resonant circuits allows correct matching for the desired frequency and discriminates against all other frequencies. Thus, these circuits not only match the load but also filter the output. Later, it is shown that coupled circuits are used to provide multiple resonance, which enables several frequencies or a small band of frequencies to be matched.

It can be shown that to secure sharpness of resonance with a low impedance source or feeder, series resonance is used in the load, while to secure sharpness with a high impedance source or feeder, a parallel combination is used. As most supply sources, such as valves, have a high impedance, parallel resonance is more frequently used than series resonance, and the methods of calculating it, as given in the above formulae, should be familiarised.

Tank Circuits - L-C Ratios. As previously stated, for a given set of operating conditions, there is a value of anode load resistance which gives highest efficiency. As far as the anode efficiency of the valve itself is concerned, it does not matter how this load resistance is obtained, that is, the valve works equally well into an actual resistor or into a tank circuit having any practicable constants, as long as the resistance or impedance represented by the tank is the desired value at the desired frequency. However, the relation between the loss of power in the tank circuit and that in the load is affected by the inherent (unloaded) impedance of the tank circuit. As power consumed in the tank circuit is a definite loss, the efficiency of the tank circuit in delivering power from the valve to the external circuit is increased by the unloaded tank circuit losses being kept at an absolute minimum.

The impedance of the tank circuit at resonance is equal to L/RC, where L is the inductance, C the capacitance and R the resistance. The higher the tank impedance (that is, the lower its losses), the greater the transfer efficiency; the relationship is shown by the curve of Fig. 15.

It is evident that the impedance of the tank unloaded should be at least ten times its loaded impedance for high transfer efficiency. The tank impedance is raised in two ways - by lowering the resistance through the construction of low-loss coils and by careful placement of parts, or by raising the L-C ratio. With practical circuits, it is much easier to raise the tank impedance by increasing the L-C ratio than by attempting to reduce the resistance.



VARIATION OF COUPLING TRANSFORMER EFFICIENCY WITH TUNED CIRCUIT IMPEDANCE.

FIG. 15.

Under normal conditions of operation, with efficiencies of the order of 70 per cent. and a fairly low value of optimum load impedance, a satisfactory L-C ratio results when the tank capacitance actually in use is approximately 200 µµF at 3.5 Mc/s, 100 µµF at 7 Mc/s, 50 µµF at 14 Mc/s, and proportionate values at other frequencies. For this type of operation, higher L-C ratios give a comparatively slight increase in transfer efficiency.

3.7 Optimum Amplifier Output Conditions. Unless handling modulated carrier, radio-frequency amplifiers generally operate as Class C Amplifiers. Strictly speaking, a Class C amplifier is any amplifier biased beyond cut-off.

There are four variable factors that must be balanced and compromised when optimum output and efficiency are to be obtained from a Class C amplifier. These factors interact on each other, and it is difficult to obtain the best set of operating conditions without a good working knowledge of the way that changes in these four factors affect the operation of the amplifier.

These four factors are -

Grid Bias, Grid Excitation,
Anode Voltage, and Aerial loading.

Grid Bias. In a Class C amplifier, the negative bias applied to the control grid is always greater than the negative voltage necessary to bring the anode current to zero. In other words, the bias is always greater than that necessary to produce anode current cut-off. Cut-off bias is either determined experimentally, or it is calculated approximately by dividing the actual measured D.C. anode voltage applied to the valve by the amplification factor of the valve.

Thus, when radio-frequency excitation is not applied to a Class C amplifier, the anode current is zero. The radio-frequency grid excitation voltage consists of A.C., so that, when it is superimposed on the constant negative grid bias, the instantaneous grid voltage supplied by the preceding driver stage swings alternately more and less negative about the axis determined by the negative bias voltage (see Fig. 16). When the instantaneous grid voltage is more negative than the D.C. bias, the anode current remains at zero and starts to flow only when the instantaneous grid voltage passes the positive side of the point marked "cut-off" in Fig. 16.

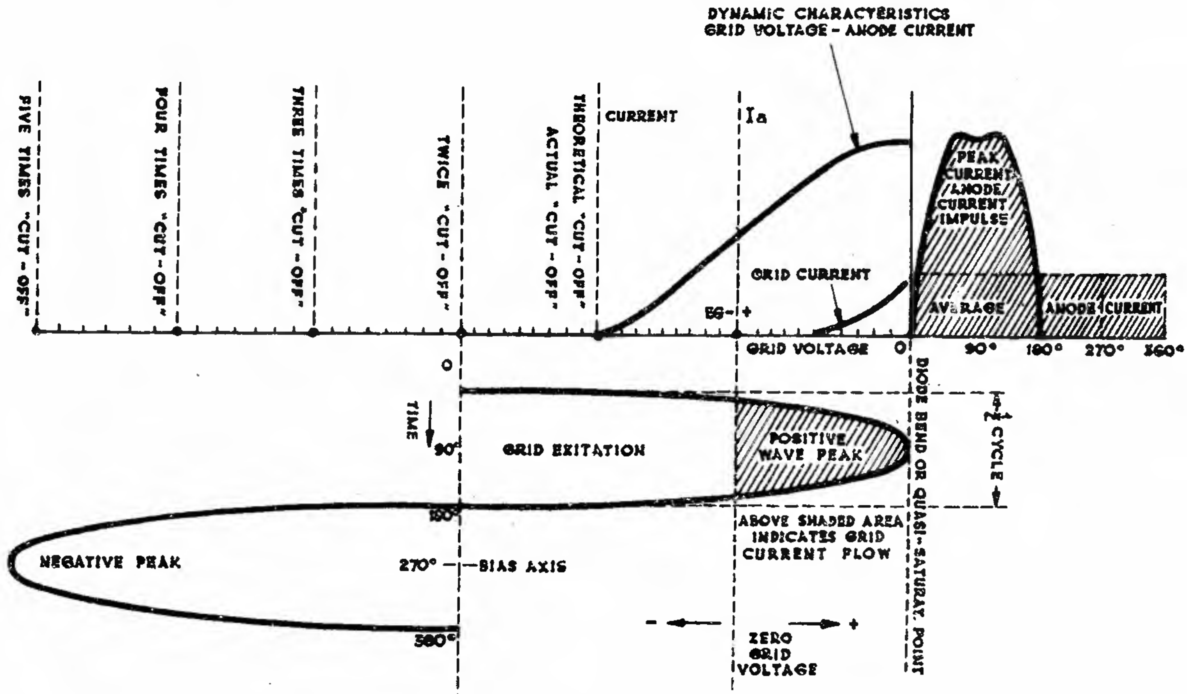


FIG. 16. "CLASS C" RADIO-FREQUENCY AMPLIFIER CURVES.

As the grid swings more positive (or less negative, which is the same), the anode current gradually increases until the grid is at its most positive point. Then, as the grid voltage swings back on the next succeeding negative-half excitation cycle, the anode current decreases to zero and remains there for some time, or until the grid again swings to the positive side of the cut-off. Thus, it is noted that anode current flows for a relatively short pulse, and that during most of the cycle the valve is not conducting. During this portion of the cycle, the valve "cools off", so that the instantaneous anode current and anode loss is high as long as the average anode loss (measured over a whole cycle) is kept below the rated anode loss of the valve. It is also noted that the higher the negative grid bias, the shorter the proportion of each cycle during which anode current flows. Thus, for a given instantaneous anode voltage and anode current, the anode efficiency rises as the negative bias is raised. However, the power lost in the grid circuit of a valve is divided between the grid itself and the source of grid bias (bias voltage times the D.C. grid current), so that raising the negative bias voltage raises the amount of grid excitation power dissipated in the grid bias supply, and, therefore, more radio-frequency power must be supplied from the preceding driver stage. Thus, in general, the power gain through a Class C amplifier decreases as the grid bias is increased.

It may also be said that the power gain decreases as the anode efficiency is raised, as long as raising the anode voltage is not the cause of the increase in efficiency. Raising the anode efficiency of a Class C amplifier is, to a certain point, desirable. The higher the anode efficiency, the more power output is obtained from a given valve, as the output from most modern valves is limited largely by the ability of the valve to dissipate heat. Thus, the smaller the valve loss compared with the power output (which ratio defines anode efficiency), the more power output is obtained from any given valve. However, as the required grid driving power increases as the anode efficiency is raised, it is uneconomical to raise the anode efficiency to the point where it takes more than 10 per cent. of the power output to drive the grid.

It is quite possible to get 90 per cent. anode efficiency from valves, but, in some cases, 33 per cent. of the power output (applied in the form of radio-frequency grid excitation to the grid of the final amplifier) is required to obtain this high anode efficiency. In cases of this kind, it is desirable to use a slightly bigger valve in the final amplifier, in order to use a smaller valve as the radio-frequency driver of the final amplifier. As a power gain of 10 usually represents a good compromise for the average Class C final amplifier, the bias voltage is chosen so that the power dissipated in the grid circuit is not more than about 10 per cent. of the radio-frequency power output of the amplifier.

When power gain is the principal objective (as it is in many buffer or radio-frequency driver stages), it is often better to bias the valve Class B instead of Class C. (Class B bias is that amount of bias that brings the anode current practically to cut-off.) Class B represents the best compromise where anode efficiencies of about 60 to 70 per cent. are satisfactory. It is possible with some of the newer high-frequency triodes to obtain a power gain of between 40 and 60 through a radio-frequency amplifier biased to cut-off. This means that, between a crystal oscillator and a 1 kW final amplifier, only one buffer-driver stage is necessary.

A fairly accurate rule for experimentally determining the proper amount of negative grid bias is to start with more bias than necessary, and then gradually reduce it, until the stage draws the normal maximum D.C. grid current recommended by the valve manufacturer. It is noted that, for any given amount of grid excitation power, the D.C. grid current rises as the bias is reduced.

Negative grid bias is best obtained from a combination bias source, except in the case of the extremely high μ valves, such as the 203Z, 838, 46, etc., which can use grid leak bias alone. Fixed bias about equal to cut-off can be provided from batteries, a separate bias pack, or from a cathode bias resistor. The balance, when more than cut-off bias is used, is provided from the voltage drop through a grid leak resistor as has been previously explained. In varying the bias, the resistance of the grid leak is varied. The actual value of the resistance is not critical. The aim is to use an amount of bias that brings the grid current to the optimum point.

The bias is roughly adjusted with the anode voltage off, although it will probably be slightly high as the D.C. grid current usually decreases when the anode voltage is applied (when the stage is perfectly neutralised). If the grid current increases when the anode voltage is applied, the stage is regenerative or oscillating. If the grid current decreases excessively when the anode voltage is applied, the stage is suffering from degeneration, and, in either case, the neutralisation is not perfect. After neutralising and then applying the anode voltage, the grid leak resistance is reduced slightly to restore the D.C. grid current to the proper value. It is usually found that changes in anode voltage or aerial coupling affect the grid current, so that, when the aerial coupling is adjusted to the limiting point (when either the anode heats or the valve draws maximum rated current), some final slight readjustment of grid bias is necessary.

Grid Excitation. An adjustment of the radio-frequency grid excitation is not usually necessary or desirable in a radio transmitter. This adjustment is usually left at the maximum obtainable from the oscillator or the preceding buffer stage, and any adjustment of the effect of the excitation is made by an adjustment of the bias voltage. However, when a large surplus of radio-frequency grid driving power is available, it is advisable to connect the excitation whilst adjusting the negative bias to a point between 2.5 and 4 times the cut-off value, so that normal grid current is flowing. By the rule of bias adjustment previously given in the paragraphs on grid bias, the amplifier operates satisfactorily even though the excess of available grid excitation necessitates raising the bias to 10 times cut-off value in order to reduce the D.C. grid current to the normal operating value. However, the use of a grid bias in excess of 4 times cut-off value is not justified by results, and materially increases the generation of undesirable radio-frequency harmonics in the output. When link coupling is used, the adjustment of radio-frequency excitation can be made by adjusting the degree of magnetic coupling between the coupling link and the tank circuit. When capacitive coupling is used between the driver and the driven stage, the coupling is adjusted either by reducing the grid bias on the preceding anode tank or by varying the size of the coupling capacitor between the driver anode and the grid of the driven stage.

When the driver stage does not draw normal anode current, the coupling to the grid of the driven stage is probably too loose. In this event, the coupling link is moved closer to the grid tank or the anode tank. When capacitive coupling (Fig. 5) is used, the size of the grid coupling capacitor is increased to about 0.0001μ F as a maximum. When the driver still does not draw normal anode current, it is necessary to reduce the tapping of the driver anode on its anode tank, leaving the grid of the driven stage connected to the top of the tank.

When the driver stage uses too much anode current, or its anode is excessively heated, it is necessary to reduce coupling. This is done by the opposite procedure, that is, by moving the coupling links to loosen the magnetic coupling with the grid or anode tank, using fewer turns on the coupling link or, with capacitive coupling, reducing the coupling capacitance or tapping the grid down on the driver tank coil. Tapping the grid down on the driver tank sometimes causes undesirable parasitic oscillation, especially with valves which have secondary and primary emission troubles with the control grid. In general, link coupling is desirable, due to the fact that a better impedance match is obtained and a greater transfer of energy is possible from stage to stage. Capacitive coupling gives good results, but it is difficult to get the grid impedance of the driven stage reflected back as the proper load impedance on the driver stage. Some transmitters use capacitive coupling almost entirely, but they have the laboratory facilities to get the excitation taps, anode voltages, etc., at exactly the right values to give optimum results. In the absence of such equipment, it is simpler to use the few extra parts required by link coupling, and so avoid the troubles that can occur when amplifiers are capacitively coupled by cut-and-try methods.

/ Anode

Anode Voltage. In a heavily driven Class C amplifier, the radio-frequency power output varies approximately as the square of the anode voltage, so that it is desirable to use fairly high values of anode voltage.

In most transmitting valves, the amount of anode voltage that can be used is limited by the internal insulation and gas content of the valve itself. For an example, 1,500 volts are applied to the anode of some types of valves in a continuous wave (unmodulated) transmitter operating on a frequency below 10 Mc/s, and 1,750 volts are applied to the anode of other types. However, when operating at this high value of anode voltage, the anode dissipation must not exceed, even momentarily, the valve rating. Some valves are not subject to insulation or gas limitations, but it has been determined that there is little advantage in using more than 3,000-4,500 volts. At higher anode voltages, there is little gain in efficiency or ease of driving, although the cost of the higher grade tank and filter capacitors required for this high voltage, increases.

Here are some basic relationships which must be kept in mind -

The higher the anode voltage, the easier a valve is to excite to a given output and anode efficiency.

The higher the anode voltage, the higher the anode efficiency for a given output and grid drive.

The higher the anode voltage, the looser the aerial coupling for a given power output.

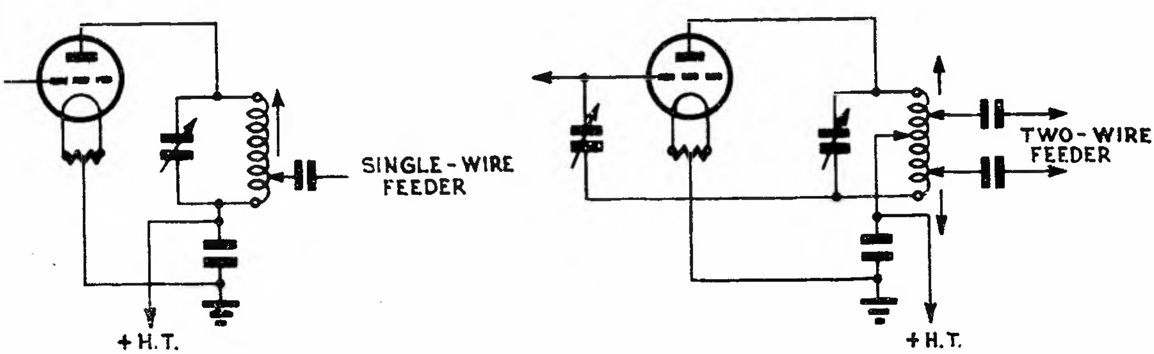
The higher the anode voltage and the looser the aerial coupling, the higher is the circulating current in the tank circuit. It follows that resistance losses in the tank are more troublesome at high anode voltage.

The higher the anode voltage, the less tank capacitance necessary with a given amount of circuit merit, or "Q", for a given valve and power output.

3.8 Aerial Coupling Circuits. To some extent, the choice of an aerial coupling system is dictated by the type of aerial (and feeder system) to be used.

As with other radio-frequency circuits, there are two possible types of coupling, inductive and capacitive, together with their variations. A good aerial coupling system transfers power with the least possible loss, and discriminates against harmonics to prevent off-frequency radiations.

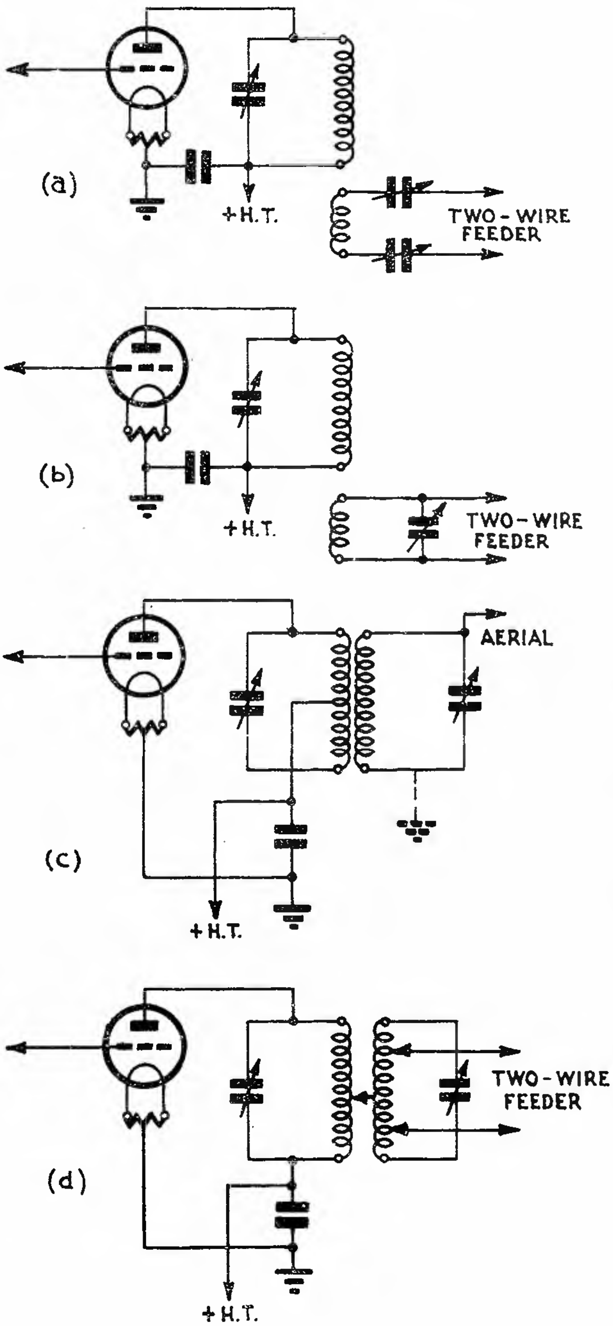
Capacitive Coupling. Capacitive coupling methods for single and two-wire feeders are shown in Fig. 17.



CAPACITIVE COUPLING TO THE AERIAL OR FEEDERS.

FIG. 17.

The adjustment is simple, as the tap is moved along the tank coil to change the value of coupling, increasing coupling being indicated by the direction of the arrow. The single wire feeder can be coupled on either side of the ground point with balanced tank circuits. While the two-wire line can be coupled to a single-ended tank circuit by connecting one wire to the ground end of the tank, the inductive coupling system shown in Fig. 18d is preferable in such a case.



INDUCTIVE COUPLING CIRCUITS FOR AERIALS.

FIG. 18.

Push-pull amplifiers can be used with direct capacitive coupling. The connections are the same as those already discussed.

Low-impedance feeder lines usually are tapped on the coil close to the ground end, while high-impedance feeder lines find the optimum coupling point nearer the anode end of the tank coil.

Inductive Coupling. Inductive coupling circuits are shown in Fig. 18. Inductive coupling is widely used with tuned transmission line feeders and also for working directly into the aerial itself without a feed system. These circuits are easy to adjust and readily adaptable to a wide range of impedance matching.

The circuit of Fig. 18a, which uses series capacitor tuning, is used for coupling to a low-impedance point on the aerial system. Fig. 18b, which uses parallel tuning, is used for coupling to a high-impedance point on the aerial system. Coupling increases by moving the aerial coil closer to the tank coil in the direction of the arrow. The coupling with series capacitor tuning is usually fairly close, and a relatively small coil is required. With parallel capacitor tuning, the coupling usually is quite loose and a large coil with small tuning capacitance should be used in the coupling system.

The arrangement of Fig. 18c, used to couple a transmitter to an aerial, is also parallel tuned (high-impedance feed point). The coupling tank may be grounded, as shown by the dotted lines, provided a short ground lead is obtained.

The circuit of Fig. 18d is used to couple a single-ended tank circuit to a balanced two-wire untuned feeder. The degree of coupling is changed by moving the coupling coil in relation to the tank coil, and also by changing the feeder taps on the coupling coil.

Adjustment of aerial coupling is dealt with in the next Paper.

3.9 Protection (Equipment and Personnel). Since radio transmitters are really convertors of energy at relatively low-frequencies to energy at high-frequencies, they require a primary source of power. This is generally taken from the main distribution systems and converted to the required voltages and currents by the power supply system associated with the transmitters (see Radio II). As the transmitter power is increased, the voltages around the system become more dangerous, and it is necessary to provide adequate protection to equipment and personnel against damage and/or injury. The main points to be guarded against are listed under two headings - PERSONNEL, and EQUIPMENT.

Details of circuits to provide the correct sequence of operations and functions are not necessary, as they vary with the transmitter design, but a general idea of the precautions to be adopted is given.

Personnel. All dangerous potentials must be removed or isolated when the transmitter is closed down.

Transmitters are usually totally enclosed and are entered only through gates which are fitted with control contacts. When a gate is opened, the circuit of the high-tension rectifiers is interrupted (sometimes the bias rectifiers as well) and all dangerous voltages removed from the transmitter. "Caution" and "Danger" notices are also provided where there is a liability of contact with relatively low voltages. Large value capacitors or capacitors with high voltages, are usually fitted with discharging contacts, which are operated to prevent a high voltage building up when the capacitor is out of circuit. This voltage gradually builds up due to dielectric absorption occurring in the insulating material forming the dielectric. It is necessary, however, that all personnel engaged on radio transmitters assist the above mechanical and electrical precautions by the exercise of common sense and care.

Equipment. To avoid damage to equipment, especially costly items such as high-power valves and rectifiers, protective circuits are required, firstly, to ensure the correct sequential application of filament, bias and anode voltages, and, secondly, to indicate faulty conditions and prevent damage therefrom.

It is now becoming customary to use circuit breakers, wherever possible, in lieu of fuses.

Some of the more important protective methods are given briefly hereunder -

(i) In the case of water-cooled or forced-air-cooled valves, protection against the application of filament voltages until the correct rate of water flow is attained.

(ii) Time delay on mercury-vapour rectifier valves to ensure that mercury has reached correct operating temperature before application of high voltage (about 15 to 20 minutes starting from cold), also about 1/2 minute delay (automatic) to recover from short breaks in transmission.

(iii) Time delay between application of final filament volts and bias.

(iv) Time delay between application of bias voltages and high tension voltages.

(v) Fault or guard circuits -

- (a) Water or air failure.
- (b) Arc back on rectifier valves.
- (c) Overload conditions.
- (d) No load conditions.
- (e) Filament volts failure.
- (f) Bias.
- (g) Anode.
- (h) Temperature alarm on water or air-cooled systems.

The above are the general conditions met with, but by no means exhaust the protective or guard circuits required. The type of transmitter used causes variations and additions to the above, as in the case of a transmitter using Class B modulation. Here, it is necessary to guard against the removal of the load from the modulator with the modulation on, a monitoring diode being used to control the high tension. Failure of radio frequency causes a relay in the diode circuit to remove the high tension from the modulator, thus preventing damage to modulator valves. Reference to these protective circuits is made again in a later Paper.

/ 4.

4. RADIO TELEGRAPH TRANSMITTERS.

4.1 Radio Telegraph Transmitters differ from Radio Telephone Transmitters primarily in that the keying system is provided in place of the modulator, and that Class C amplifiers are used throughout, even where linear amplifiers are used in the corresponding telephone transmitter. In telegraph work it is not generally necessary to have the transmitter particularly free from noise and hum modulation.

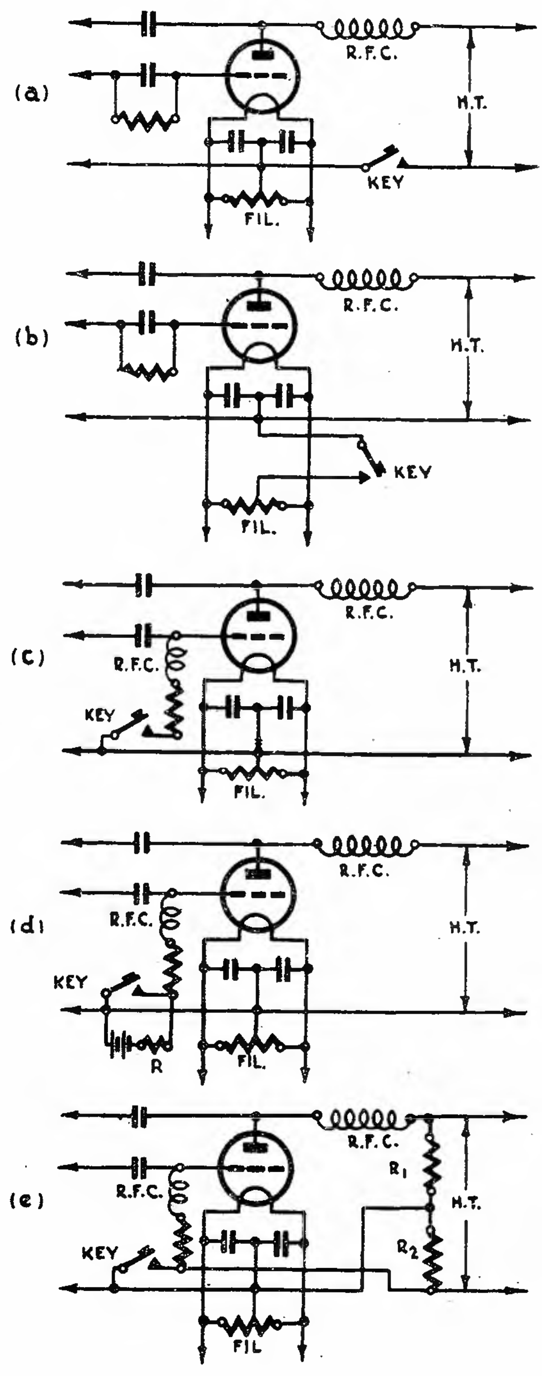


FIG. 19. TYPICAL KEYING METHODS.

For telegraphic transmission, the continuous oscillations (C.W. - as explained in Paper No. 10 of this book) must be divided into the dot and dash trains of the morse code. A hand key operating at low or moderate power levels is generally used, so that the power that need be controlled is small. The keying operation is normally performed with the aid of a relay controlled by the key, but, in low-power transmitters, the key is sometimes inserted directly in the transmitter circuits.

4.2 Methods of Keying. Keying is accomplished by an arrangement which reduces the output of the transmitter to zero when the key is "open" and permits full output when the key is "closed".

One way of doing this is to fit the key in series with one of the anode-supply leads to the valve, as shown in Fig. 19a. When the key is open the anode power is cut off and, thus, there is no output.

The keying method shown in Fig. 19b is known as "centre-tap" keying, because the key is connected between the cathode centre-tap (which may be the midpoint of a resistor or a centre tap on the cathode transformer) and the point where the negative side of the power supply and the grid return circuit are connected together. This system differs from Fig. 19a because, in addition to breaking the anode supply from the valve, the key also breaks the D.C. return circuit from the grid and thus prevents the flow of grid current.

In Fig. 19c, the key opens only the D.C. grid return circuit, leaving the anode supply connected to the valve. This circuit operates because a break in the D.C. path between grid and cathode of oscillators or radio-frequency power amplifiers

/ causes

causes electrons to accumulate on the grid, giving it a negative charge which prevents the flow of anode current. Since the negative voltage which the grid must assume, in order to prevent anode-current flow, depends upon the valve's amplification factor, this method of keying is more successful with valves having high μ's, because a smaller "blocking" voltage is required than with low μ valves. Good insulation in the key is also a requisite, since poor insulation permits some of the charge to leak off and thus reduce the negative grid voltage to a value which allows anode current to flow with the key open. This causes some energy to be radiated during spaces in the keying; such radiation is termed a "back-wave".

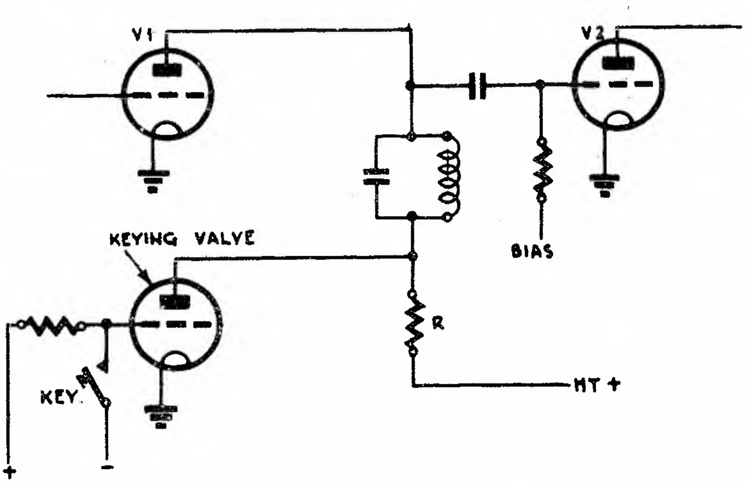
Other grid-blocking keying systems, in which the negative voltage applied to the grid during keying spaces comes not from natural accumulation of electrons but from a bias source having a definite voltage, are shown in Figs.19d and 19e.

In Fig. 19d, the grid-blocking voltage is supplied by a battery. When the key is open, the full blocking voltage is applied to the grid through the resistor R; when the key is closed, the blocking voltage is short-circuited as far as the grid of the valve is concerned. Resistor R is in the circuit to prevent actual short-circuiting of the blocking-voltage source. This resistor must be of such value as to limit the current flow to a few milliamperes when the key is closed - roughly 5,000 ohms for each 50 volts of bias. The extra bias or blocking voltage required in this keying method depends upon the type of valve, the anode voltage and the excitation. In normal oscillator or amplifier stages, a first approximation is a blocking voltage equal to the anode voltage divided by one-third the amplification factor (μ) of the valve. The actual value of bias required is usually somewhat greater or less, however, and is best determined by experiment.

The system shown in Fig. 19e is similar to that of Fig. 19d, but, in this method, blocking bias is obtained from the anode supply through a voltage divider. The centre-tap of the cathode is connected to the junction of R1 and R2, and the grid return of the negative side of the power supply, so that, when the key is open, the voltage drop across R2 is applied as bias to the grid of the valve. With the key closed, R2 is short-circuited. R1 is the regular power-supply bleeder. The resistance of R2 is about half that of R1 in practically all cases.

In all of these diagrams, the centre-tapped resistor across the cathode supply is usually omitted when the cathode transformer winding is centre-tapped. In this case, the connection shown to the mid-point of the resistor across the cathode supply in Fig. 19, is made to the centre-tap of the transformer winding. When a secondary battery is used for the transmitting valve filament, the centre-tapped resistor is omitted, and the connection which goes to its mid-point should be connected to the negative terminal of the cathode. The by-pass capacitors across the cathode are not necessary.

The relay in keying systems is sometimes replaced by a valve. An example of one of the many ways of accomplishing is shown in Fig. 20.



With the contacts closed, the keying valve is biased beyond cut-off, and normal operation takes place. When the key is open, however, the grid of the keying valve is slightly positive, causing this valve to draw a large anode current through the series resistance R and hence, to reduce the potential that is applied to the anode of V1. This reduces the output of V1 to a value insufficient to drive the grid of the Class C amplifier valve V2 above cut-off.

FIG. 20. KEYING VALVE.

/ Valves

Valves on the high-powered side of the keyed stage must be operated, so that, with the exciting voltage removed, the anode current is not sufficient to make the anode dissipation of the valve excessive. This requires either that the class C amplifier stages have fixed bias, self-bias or a combination of the two. Grid-leak bias is seldom permissible.

Unless the oscillator is keyed, the low-powered portions of the transmitter operate continuously. It is important that this part of the telegraph transmitter is carefully shielded to prevent radiation during the spacing periods. When such a "back-wave" has appreciable intensity, it causes severe distortion of the keyed characters.

4.3 Choice of Keying Systems. Although any of the circuits shown in Fig. 19 may be used with any type of oscillator or amplifier, the systems shown at b, c and e are generally used. Centre-tap keying is positive - that is, completely prevents output during keying spaces - but is more likely to cause key clicks (see paragraph 4.4) than either of the grid-blocking systems. Keying a high voltage lead as at Fig. 19a causes key-clicks. The method of Fig. 19d is good, but requires an extra source of voltage for the blocking bias.

With oscillator-type transmitters, either self or crystal controlled, the circuit chosen must be one which gives clean keying with least tendency toward key clicks. An actual trial of a few of the circuits soon gives the desired information. In multi-stage transmitters, the operator has the additional choice of applying the keying system to one or more stages. It is preferable to let the oscillator in an oscillator-amplifier set run continuously, keying one or more of the following amplifier stages. As an alternative, an intermediate stage is keyed, in which case the output stages, although operative, do not receive excitation during the keying spaces and, hence, give no power output. Also, the oscillator itself may be keyed, an arrangement which is practically a necessity when break-in operation is to be used at the station.

In general, it is advisable to key in a low-power stage preceding the higher-power amplifiers. When this is done, key clicks are reduced and there is less possibility that a back-wave will be emitted. It should be noted, however, that the amplifier valves following the keyed stage must be provided with sufficient negative bias from a fixed voltage source to cut off anode current when there is no excitation, otherwise the amplifier valves are likely to be damaged.

A separate cathode transformer or winding for the keyed valve is necessary with centre-tap keying (Fig. 19b) and the blocked-grid arrangement (Fig. 19e) when a single stage is to be keyed.

4.4 Key Clicks. A properly designed radio-telegraph transmitter produces radiation in the form of clean-cut dots and dashes having rounded edges as shown in Fig. 21b.

(a) Code Characters.

(b) Envelope of Wave with no Key Clicks.

(c) Envelope of Wave with Key Clicks.

RADIO TELEGRAPH SIGNALS.

FIG. 21.

The rapid starting and stopping of power output, however, produces transient oscillations of very short duration, which do not have a well-defined period but which spread over a large portion of the frequency spectrum and often cause interference in nearby receivers tuned to a frequency widely different from that of the transmitter. Interference of this kind manifests itself in the form of clicks or thumps in the output of the affected receiver. The interference is usually noticeable only in the immediate vicinity of the transmitter, seldom travelling more than a few hundred yards except on frequencies within a few kc/s of the transmitter frequency.

To prevent key clicks, it is necessary to prevent the radiation of transient oscillations caused by keying.

Transients can be prevented by slowing up the rate at which power is applied to the transmitter. Provided the "slowing-up" is not too pronounced, the keying is unaffected while the transient is eliminated. The slowing-up is done at an audio-frequency or low-radio-frequency rate by introducing an inductor in series with the key, as shown in Fig. 22a, b and c. This inductor varies from a large radio-frequency choke (about 10 mH) to an iron-cored choke of a few henrys inductance. Experiment is usually necessary to determine the value of inductance which eliminates the clicks in a particular transmitter. The inductance must be large to prevent clicks, but not so large that the crispness of keying is spoiled.

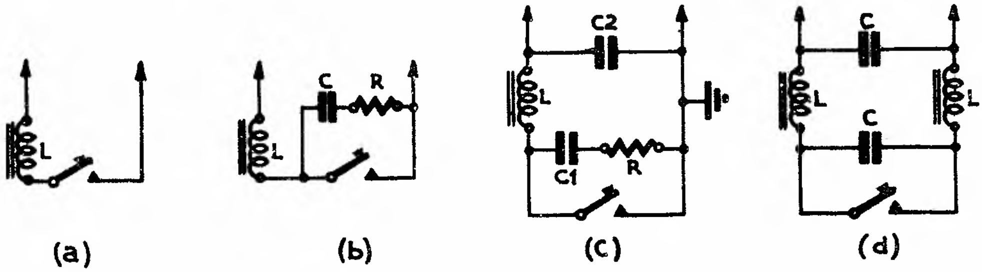
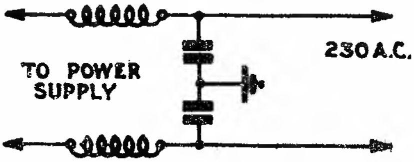


FIG. 22. KEY CLICK FILTERS.

The simple circuit of Fig. 22a, however, is not usually enough in itself to prevent clicks. Introduction of inductance in series with the key is likely to cause sparking at the key contacts, a prolific source of the second type of keying transient. It then becomes necessary to absorb the energy released by the inductance when the key is opened, the capacitor C and resistor R (Fig. 22b) being provided for this purpose. R is usually variable for ease in adjustment. With an inductance of a few henrys at L, C is usually between 0.25 and 1 μF, and R is a variable resistor having a maximum resistance of 50 to 100 ohms.

Oscillations arising in the key circuit - the type that travels over the power wiring - are further reduced by connecting a second capacitor, as shown at C2 in Fig. 22c, in which L, C1 and R have values corresponding to those in Fig. 22b. C2 is 0.1 μF or less. The side of the circuit which connects to an earth or low-potential part of the transmitter (usually the negative high-voltage terminal) is indicated by the earth symbol. Another circuit which often proves effective is the balanced arrangement shown at Fig. 22d, in which the inductors are large radio-frequency chokes and the capacitors about 0.1 μF each.

These circuits are usually most effective when installed right at the key, with very short connecting wires. Their purpose is to damp out the transients generated at the key contacts, so that the oscillations do not get into the keying line. When long leads are used to connect the key to the click filter, the oscillations which still exist in this part of the circuit are radiated from the connecting wires.

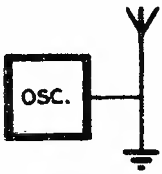


R.F. FILTERS FOR POWER LINES.

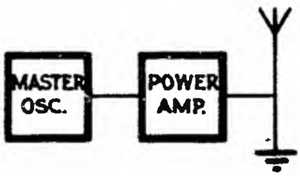
FIG. 23.

Further Click-Prevention Methods. To prevent keying transients from being carried over house wiring and power lines from the transmitter to nearby receivers, a filter is installed in the 230-volt line which feeds the power transformers. Such a filter is shown in Fig. 23. This filter consists of a pair of radio-frequency choke coils, one in each leg of the line, and a pair of capacitors in series across the line with their mid-connection earthed.

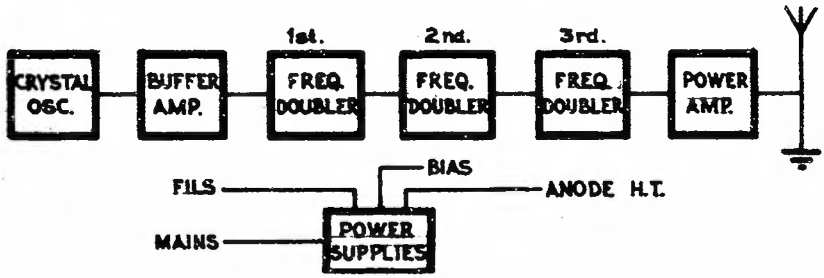
4.5 Schematic Diagrams of Typical Radio-Telegraph Transmitters. Fig. 24 shows, in block schematic form, the manner in which a Radio-Telegraph Transmitter is built up from a simple keyed oscillator. The main difference between the medium and high-frequency transmitters is the addition of frequency multiplying stages. These stages enable the oscillator crystal to be of a relatively low frequency, which renders manufacture easier. Crystals are now readily available up to about 7 Mc/s, but the use of the lower values simplifies oscillator design.



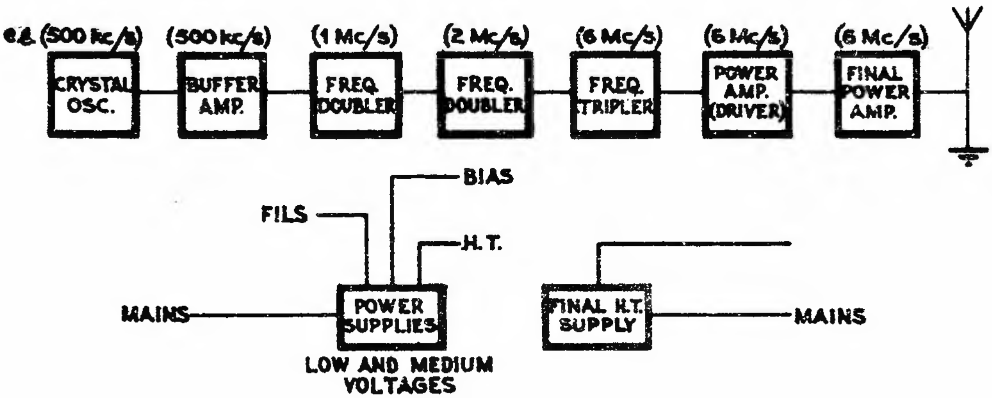
(a) Simple Valve Transmitter (Medium and High Frequencies).



(b) Improved Transmitter for Medium Frequencies.



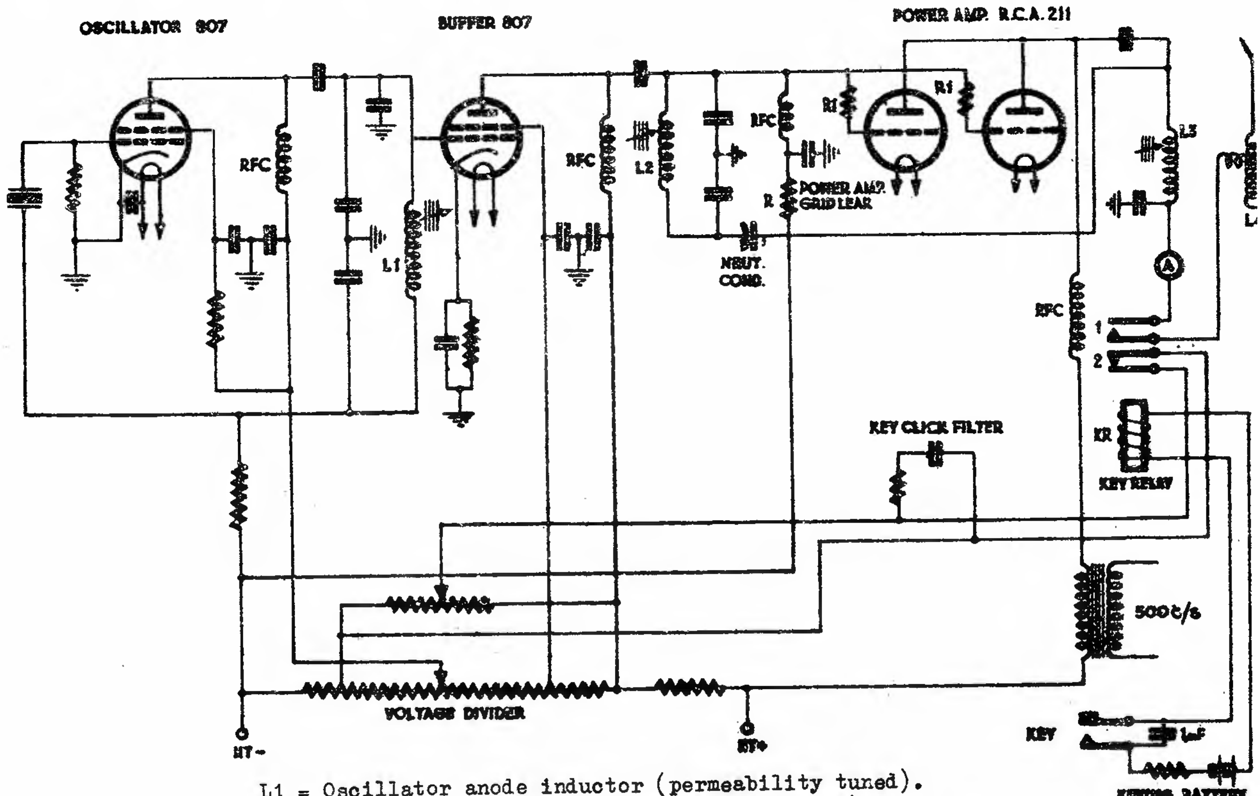
(c) Low-Power High-Frequency Telegraph Transmitter.



(d) High-Power High-Frequency Telegraph Transmitter.

FIG. 24. DEVELOPMENT OF A RADIO TELEGRAPH TRANSMITTER.

Fig. 25 gives the simplified circuit of a transmitter of about 300 watts power suitable for the medium frequency band. It is noticed that the secondary of an audio transformer is in series with the anode supply to the power amplifier, the primary being connected to an audio source of 500 c/s. This arrangement places an audible tone on the keyed morse characters, and renders reception easier (see next Paper).



L1 = Oscillator anode inductor (permeability tuned).
L2 = Buffer anode inductor (permeability tuned).
L3 = Power amplifier anode inductor (permeability tuned).
L4 = Aerial loading coil.
R1 = Parasitic suppressors.

TELEGRAPH TRANSMITTER (300 WATT).

FIG. 25.

The oscillator includes an 807 valve followed by an 807 valve as buffer driving a pair of 211 valves in parallel for the final output. Keying is accomplished via the keying relay KR. This relay performs two functions when the key is depressed - contacts "1" connect power amplifier output to the aerial, and contacts "2" remove a short-circuit from a resistor in the voltage divider circuit. This circuit is simplified in Fig. 26, and it is noted that R1 and R2 comprise a bleeder circuit across the high tension supply. With the key "open", contacts K2 are closed, increasing the current through the bleeder circuit and reducing the effective voltage. With the key "down", contacts K2 are opened, thus inserting the 15,000 ohm resistance in series with the 20,000 ohm resistance and reducing the bleeder current.

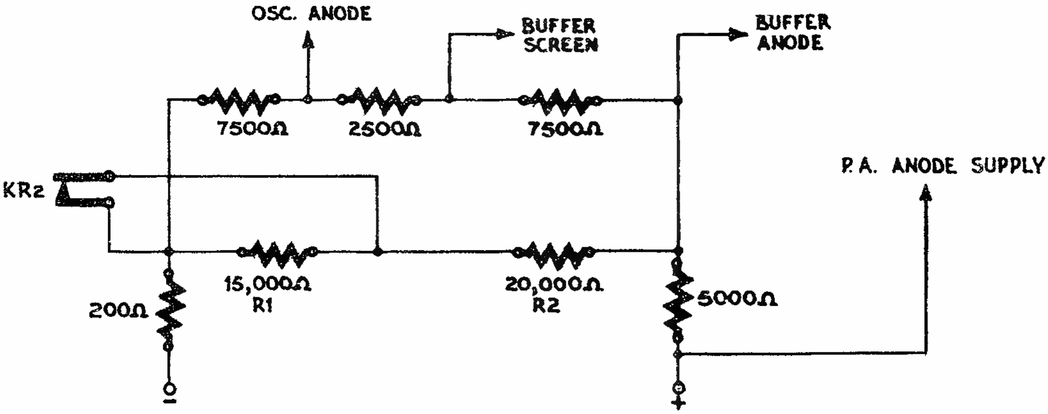


FIG. 26. HIGH TENSION SUPPLY.

4.6 High-Frequency Radio-Telegraph Transmitters. Fig. 27 shows a simplified sketch of a high-frequency radio-telegraph transmitter, consisting of the following valves -

Oscillator	Type 76
Buffer Doubler	Type 6L6
2nd Doubler	Type 6L6
3rd Doubler	Type 6L6
Power Amplifier (2 Valves in parallel).	Type 6L6

Keying is accomplished by short-circuiting the bias resistor R13. This removes from the grid circuits of all the valves a bias which normally renders them inoperative. This arrangement places only a low voltage across the key. The power input to the aerial system is about 200-300 watts.

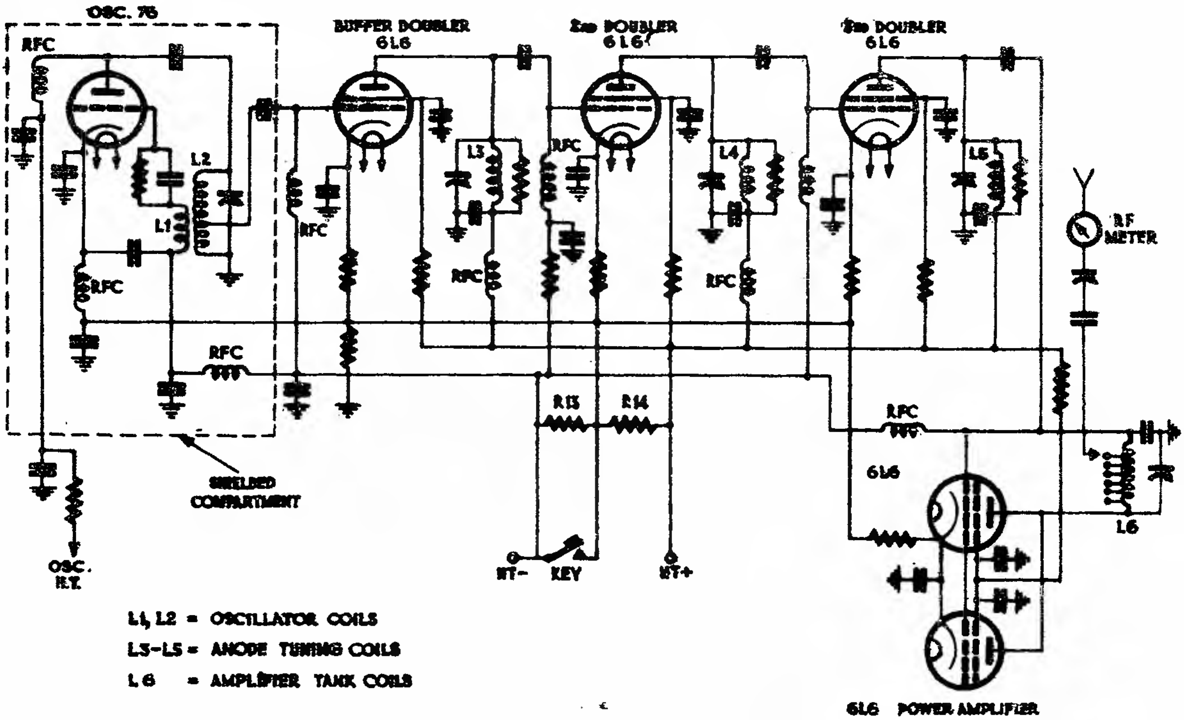
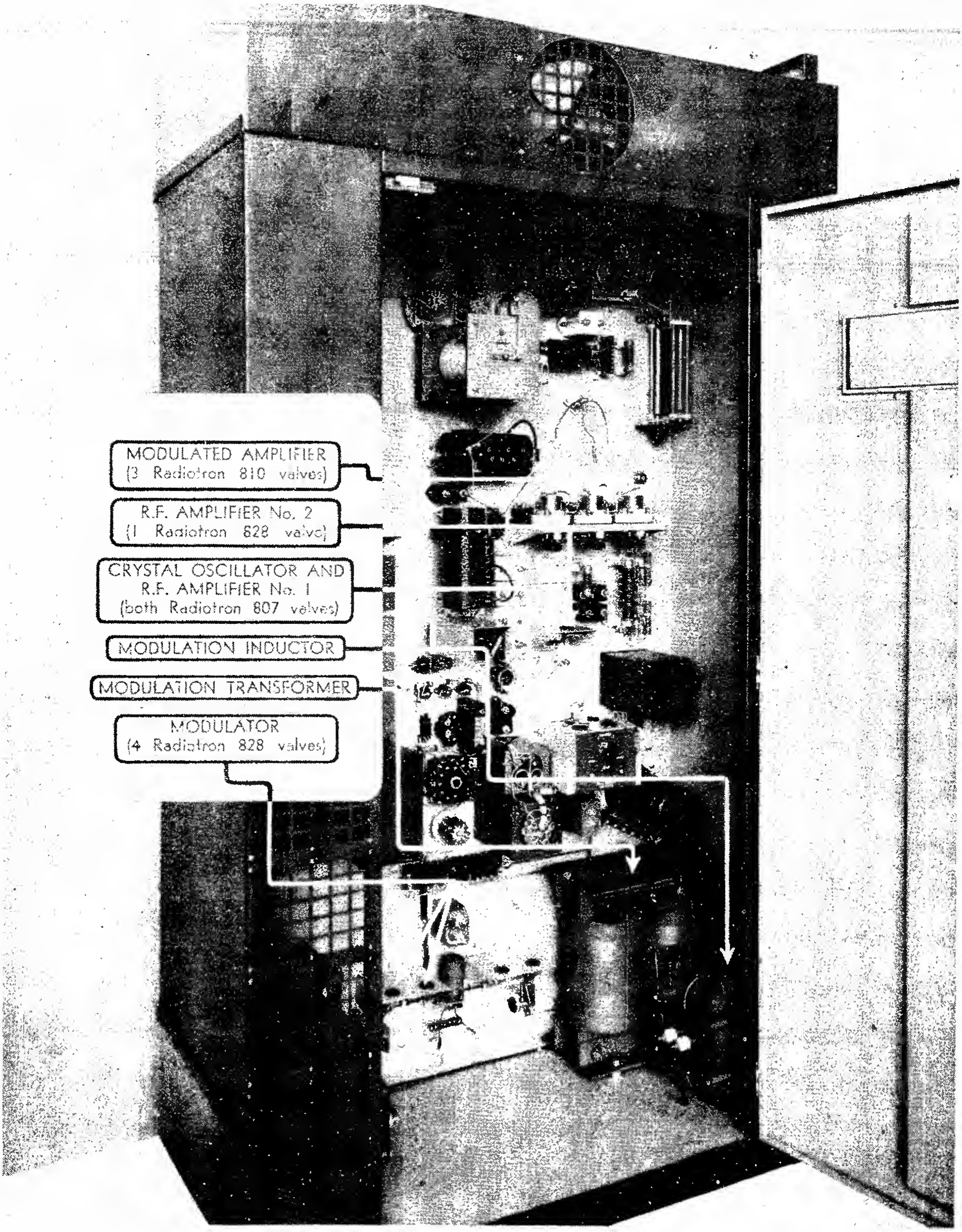


FIG. 27. HIGH-FREQUENCY RADIO-TELEGRAPH TRANSMITTER.



500 WATT BROADCAST TRANSMITTER (A.W.A.).

5. BROADCAST TRANSMITTERS FOR MEDIUM-FREQUENCY BANDS.

5.1 Broadcast transmitters represent the highest development of radio-telephone transmitters with respect to stability of carrier frequency, band-width, low distortion and noise, etc. Such transmitters normally include a crystal oscillator followed by several buffer/amplifier stages, a modulated amplifier and an audio-frequency modulating system, plus accessories such as protection equipment and monitoring facilities.

5.2 The crystal oscillator of a broadcast transmitter operates at the carrier frequency, and normally uses a zero-temperature-coefficient crystal operating at low-power level and under conditions such that the greatest frequency stability is realised. (Tolerances for medium frequency stations in Australia are ± 20 c/s, but most stations are kept to within a few c/s.)

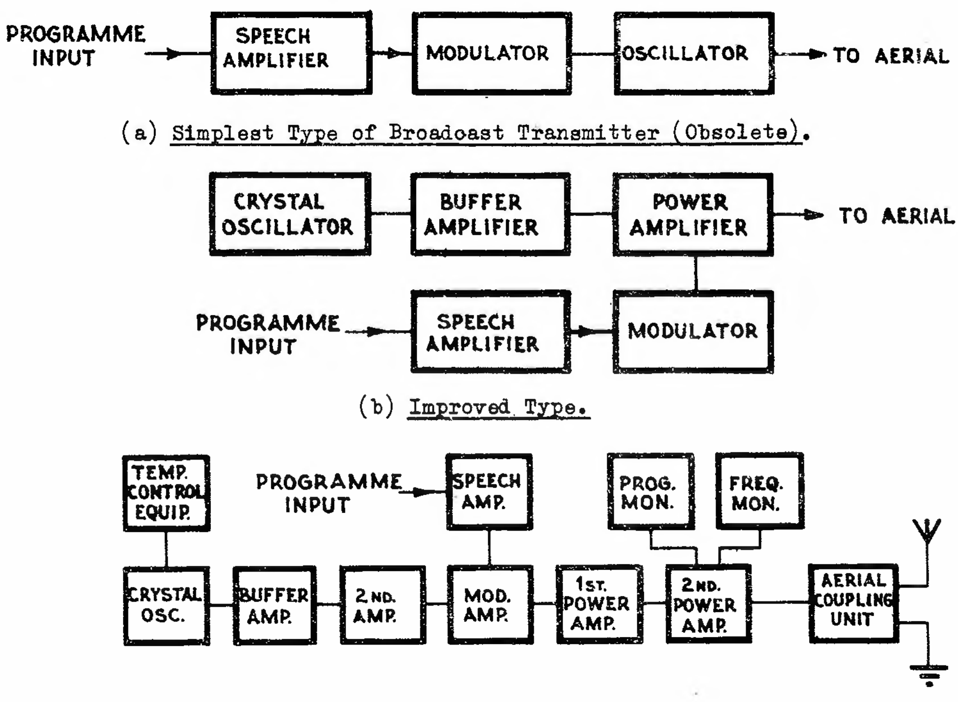
5.3 The buffer amplifiers are for the purpose of isolating the crystal from the modulated amplifier, and use screen-grid, pentode or neutralised triode valves.

5.4 Modulation is accomplished either in the last stage of radio-frequency amplification, termed high-level modulation, or at lower power level, then termed low-level modulation. In low-level systems, the radio-frequency amplifiers following the modulated stage are usually high-efficiency type linear amplifiers. High-level modulation usually takes the form of anode modulation with a Class B audio amplifier.

A variety of modulation methods (see Paper No. 10) is in use for low-level modulation, including anode, conventional control-grid and suppressor-grid modulation, in this order of popularity. Negative feedback is used to reduce distortion, and is in common use.

5.5 The main difference between telegraph and telephone transmitters is that telephone transmitters require some means of impressing the programme on the carrier; this is termed "Modulation".

Fig. 28 shows the development of Broadcast Transmitters in a similar manner to Fig. 24 for Telegraph Transmitters.

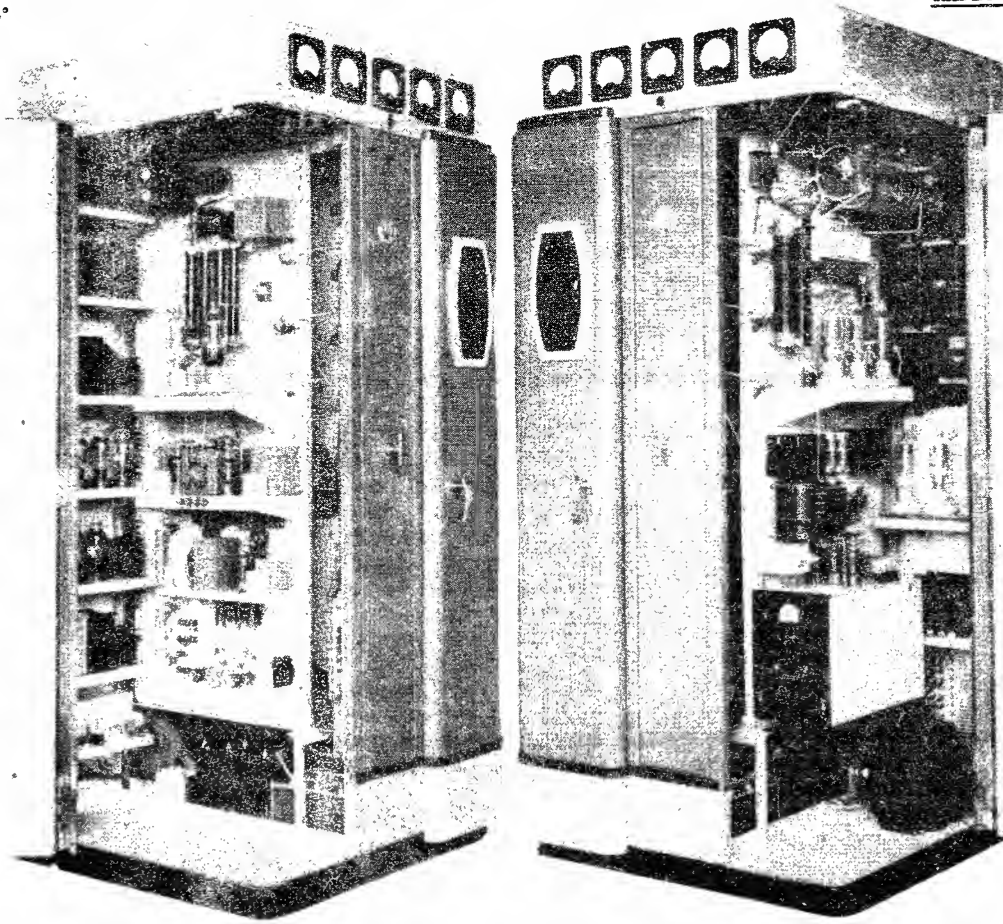


(a) Simplest Type of Broadcast Transmitter (Obsolete).

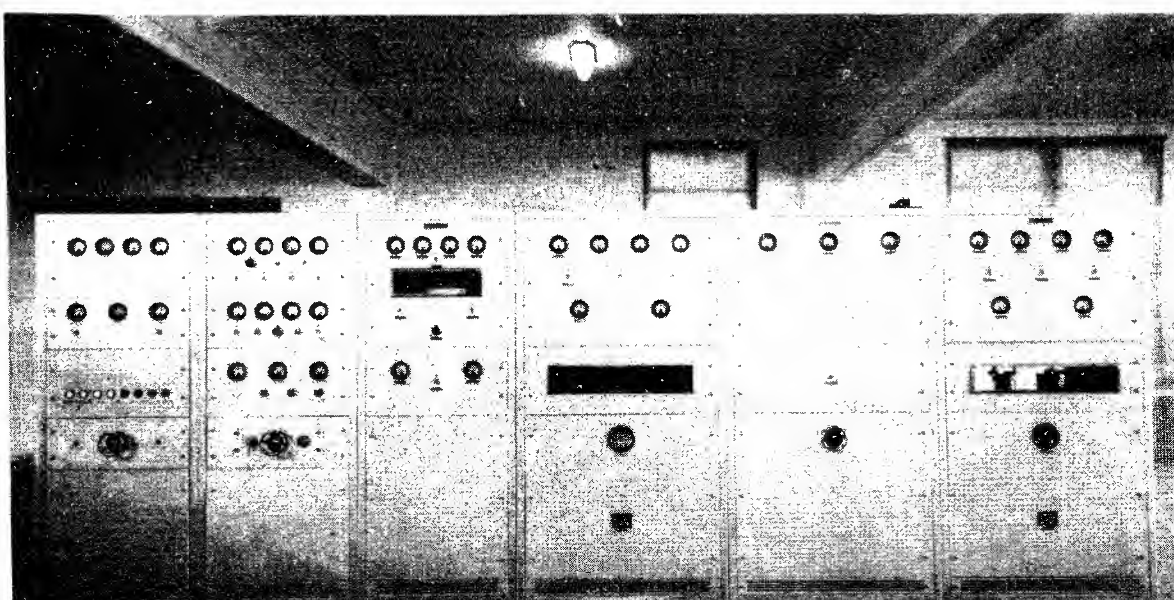
(b) Improved Type.

(c) Modern Medium-Frequency Transmitter.

FIG. 28. DEVELOPMENT OF BROADCAST TRANSMITTER.



500 WATT BROADCASTING TRANSMITTER (A.W.A.).



MEDIUM FREQUENCY BROADCAST 10 kW TRANSMITTER.
(2FC, 3LO, 3AR AND OTHERS).

5.6 As mentioned previously, negative feedback is commonly applied to broadcast transmitters to improve their performance. The most important practical benefits gained from the use of negative feedback in transmitters are the reduction in distortion and in noise modulation. The reduction in distortion makes it possible to sacrifice linearity of modulation and amplification for efficiency, and, in particular, makes it practicable to use high-efficiency systems which, without feedback, have high distortion. The reduction in noise modulation makes it possible to use A.C. for heating the filaments in the radio-frequency valves and to use relatively light filtering in the power supply systems without the hum modulation exceeding a reasonable value. Fig. 29 shows the application of negative feedback to a radio-telephone transmitter.

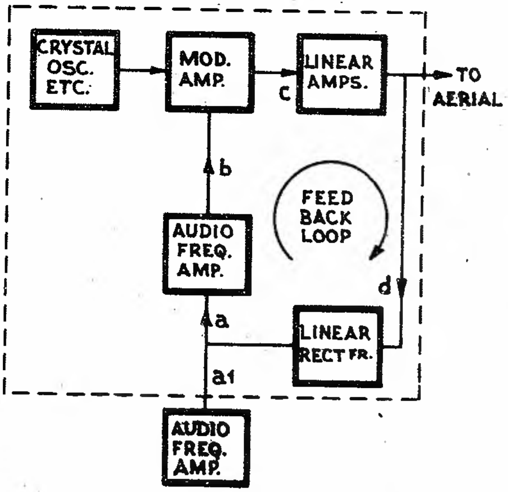


FIG. 29. FEEDBACK CIRCUIT.

In operation, a portion of the output wave is rectified and fed back into the audio-frequency system of the transmitter in phase opposition to the audio-frequency voltages being amplified, as shown in Fig. 29. The consequences are briefly summarised as follows -

The audio gain is reduced, so that more amplification is required in the modulator. Amplitude, phase and frequency distortion generated anywhere within the loop a, b, c, d in Fig. 29 are reduced as far as their effects upon the transmitter output are concerned, and all forms of noise modulation, including hum, induced anywhere in the part of the transmitter enclosed by the dotted rectangle are likewise discriminated against.

The reduction in distortion, noise, etc., resulting from the use of negative feedback, is exactly proportional to the reduction in gain caused by feedback, and is accordingly commonly measured in this way. In a typical broadcast transmitter, the amount of feedback is usually approximately 15 to 30 db.

5.7 By way of interest, in Fig. 30 is given the block diagram of a 10 kW broadcast transmitter such as 3LO and 2FC, etc., and Figs. 31, 32 and 33 show briefly some of the separate stages. The transmitter is of the low-power modulated type and incorporates feedback. It must be noted that the modulated amplifier has four valves in push-pull/parallel arrangement, and the power amplifier is similarly arranged. The valves in the final amplifier are of the water-cooled type.

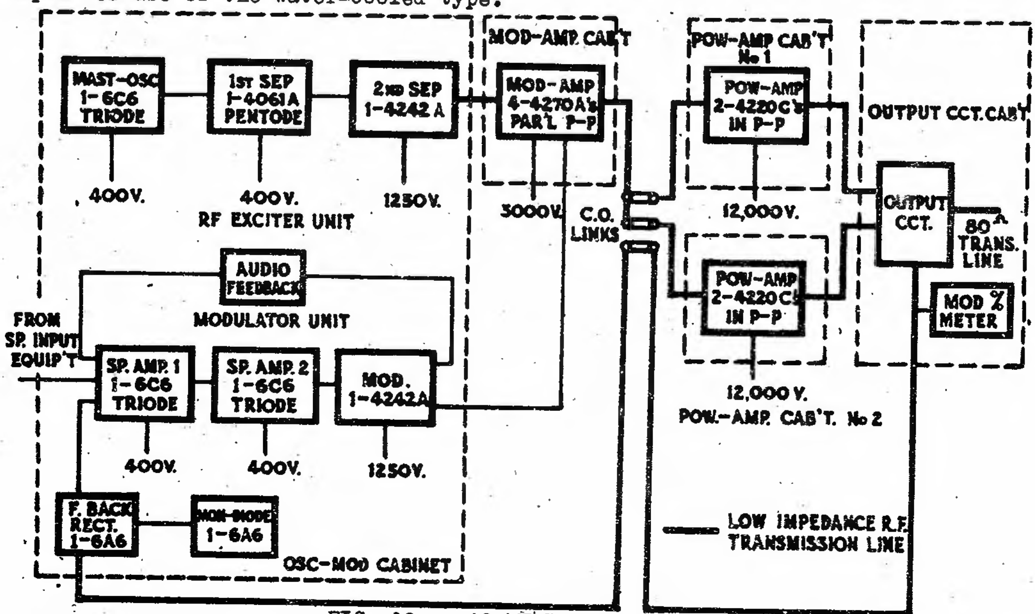
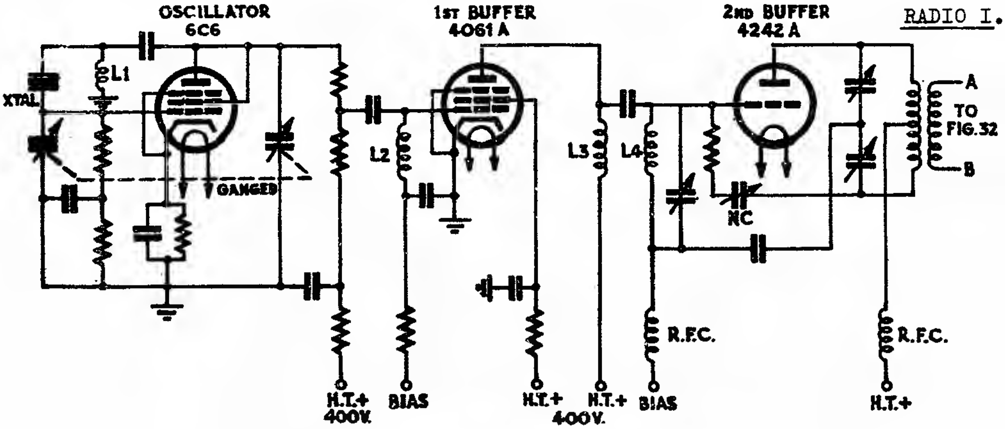


FIG. 30. 10 kW TRANSMITTER.



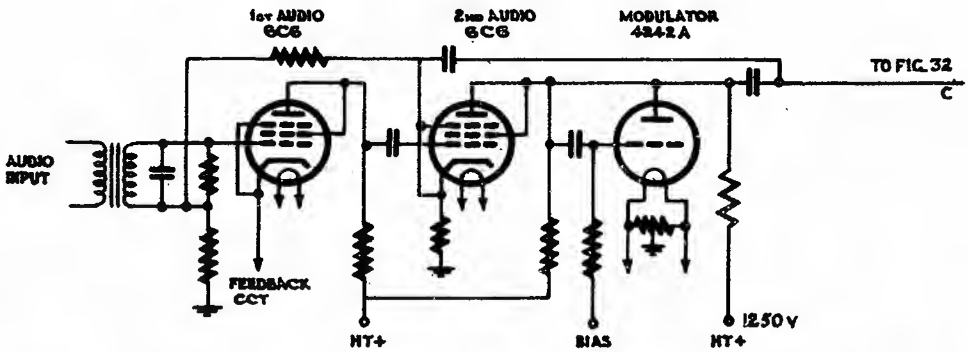
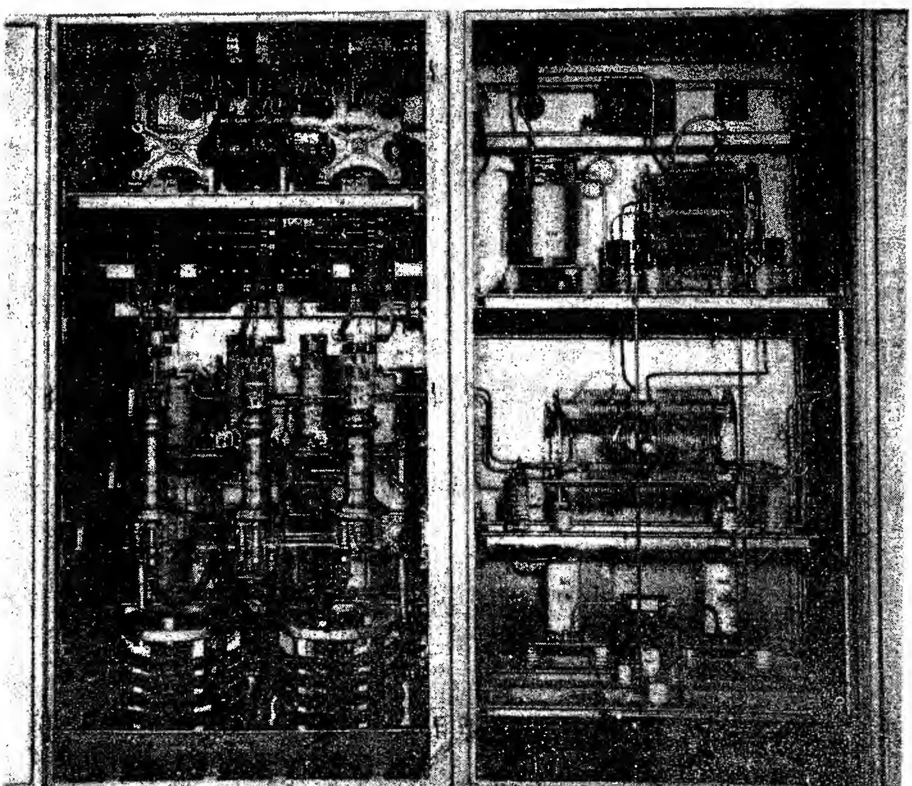


FIG. 31. OSCILLATOR-BUFFER, SPEECH AMPLIFIER AND MODULATOR STAGE.



5 kW POWER AMPLIFIER AND OUTPUT TUNING UNIT OF A 10 kW MEDIUM FREQUENCY BROADCAST TRANSMITTER. (S.T.C.).

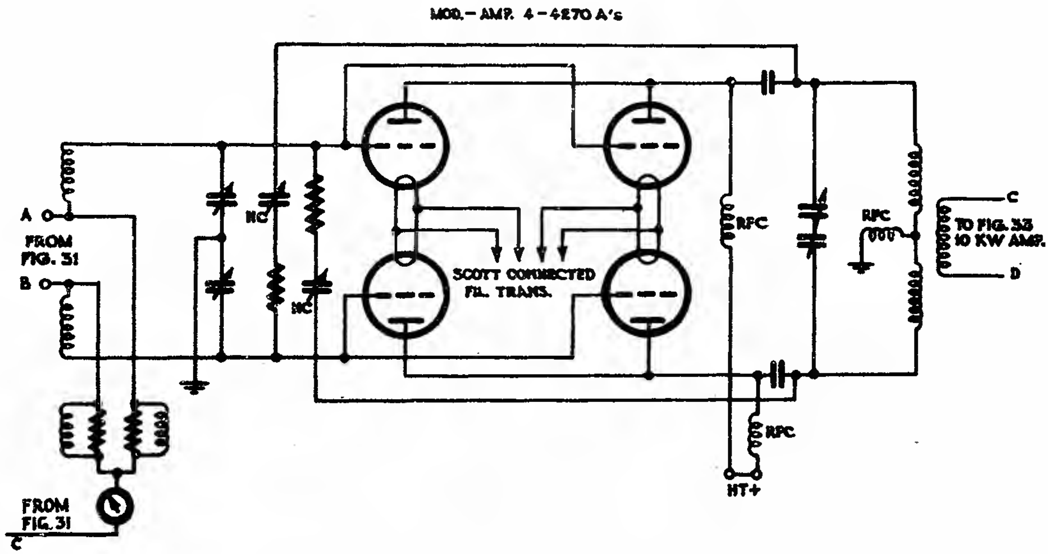


FIG. 32. MODULATED AMPLIFIER STAGE.

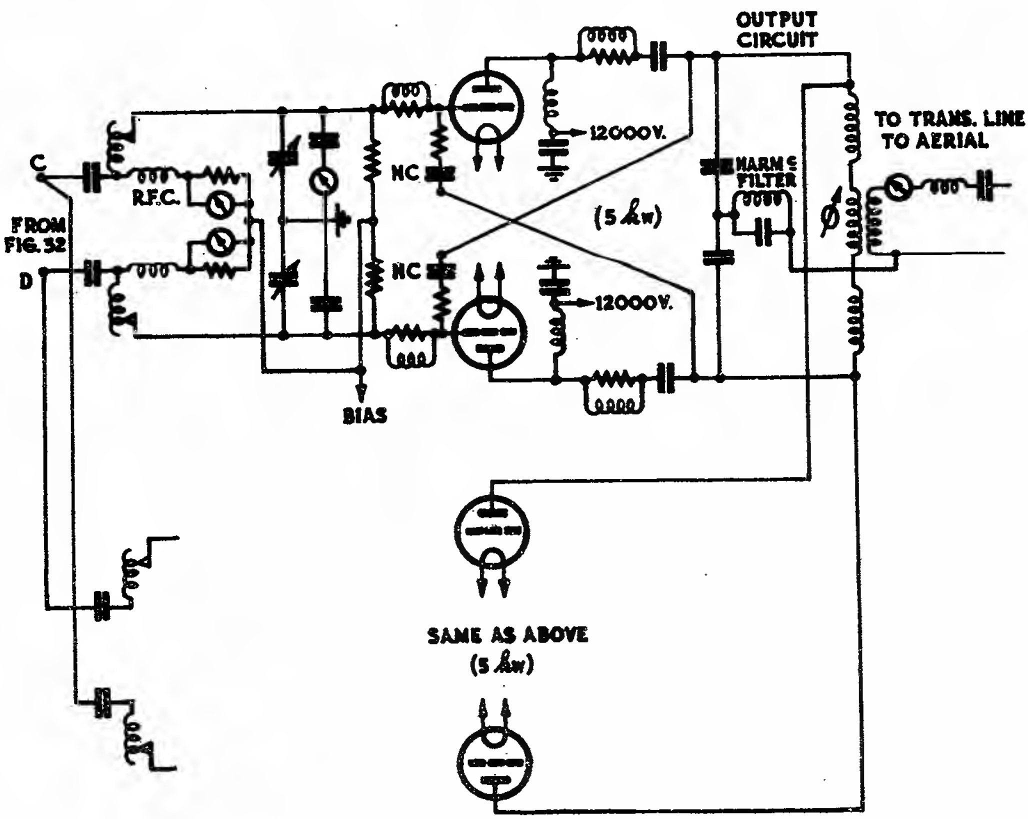


FIG. 33. 10 kW AMPLIFIER AND OUTPUT CIRCUIT.

6. BROADCAST TRANSMITTERS FOR HIGH-FREQUENCY BANDS.

6.1 The main differences between these transmitters and those in Section 5 are the following -

(i) Provision for frequency changing (range usually about 6-22 Mc/s).

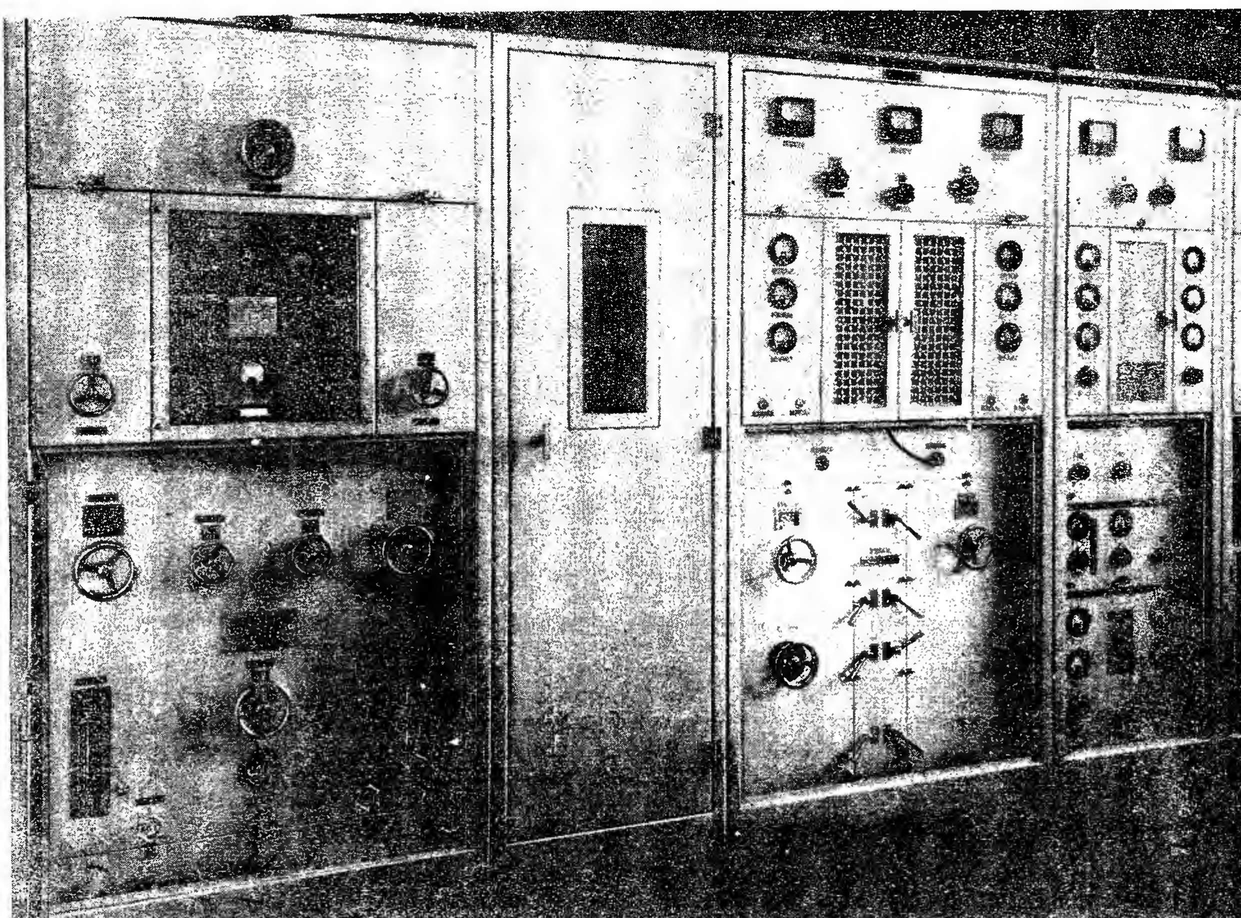
(ii) Provision for aerial switching.

(iii) Provision for greater average depth of modulation by the use of limiting amplifiers. (Sometimes used with M.F. transmitters also.)

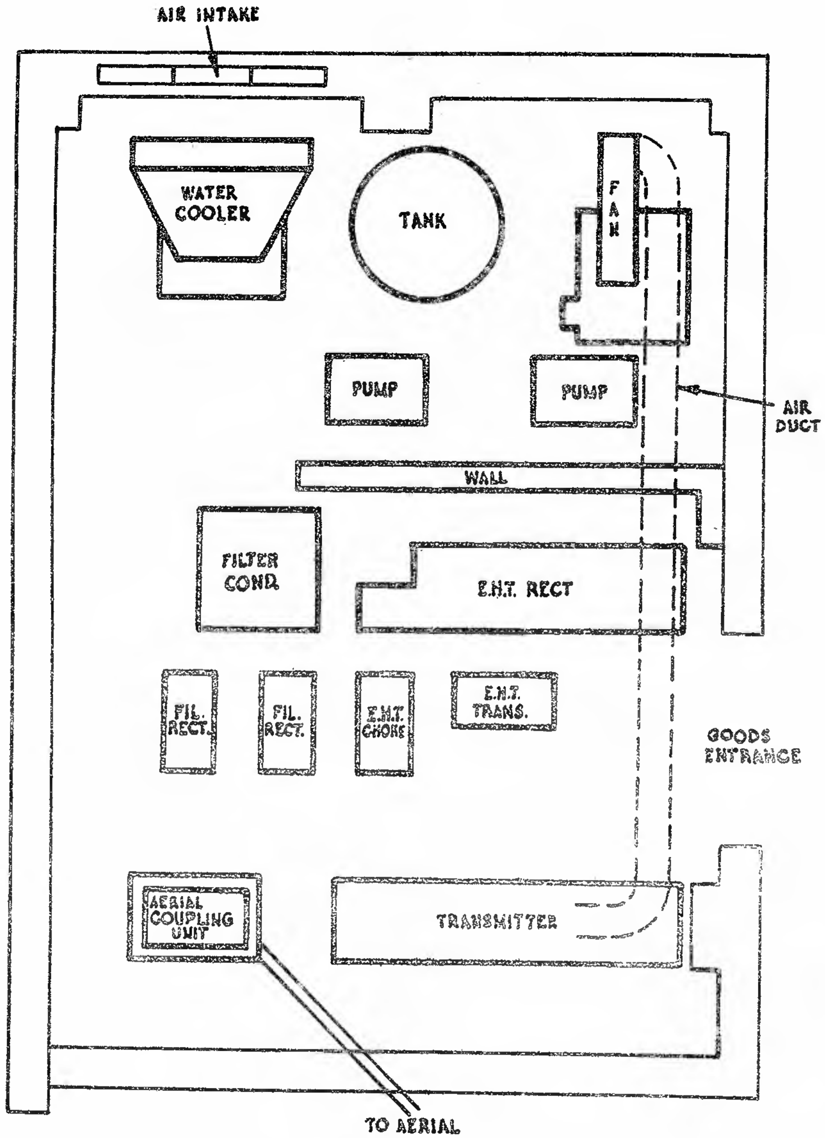
The frequency changes required vary from the day to night change for fixed broadcast coverage to the relatively large number which are required for international transmissions.

The aerial changes likewise vary according to the type of service being provided, the number of countries being served and the times of the services.

The transmitters follow the same lines as those previously described, plus the additions of the requisite number of multiplier stages to bring the crystal frequency to the carrier frequency. The floor plan of a 10 kW Short-Wave Transmitter is shown on Page 37.



VLH TRANSMITTER 10 kW (S.T.C.).



10 kW SHORT-WAVE TRANSMITTER BLOCK FLOOR PLAN.

6.2 For the purpose of illustration, the schematic diagrams of a 50 kW high-frequency transmitter using high-level modulation are given in Figs. 34, 35, 36, 37 and 38. Schematic diagrams of some of the stages are included.

The aerial switching is shown very simply. Actually, there are three transmitters sharing access to these aerials, and suitable protective circuits are necessary to prevent such conditions as two transmitters on one aerial, etc.

The schematic diagrams show that the circuits of high or low-power transmitters are basically similar.

Detailed descriptions of the circuits and functions are beyond the scope of these notes, the figures and diagrams being given to add interest to the text, and to give an overall idea of the different transmitters.

/ Fig. 34.

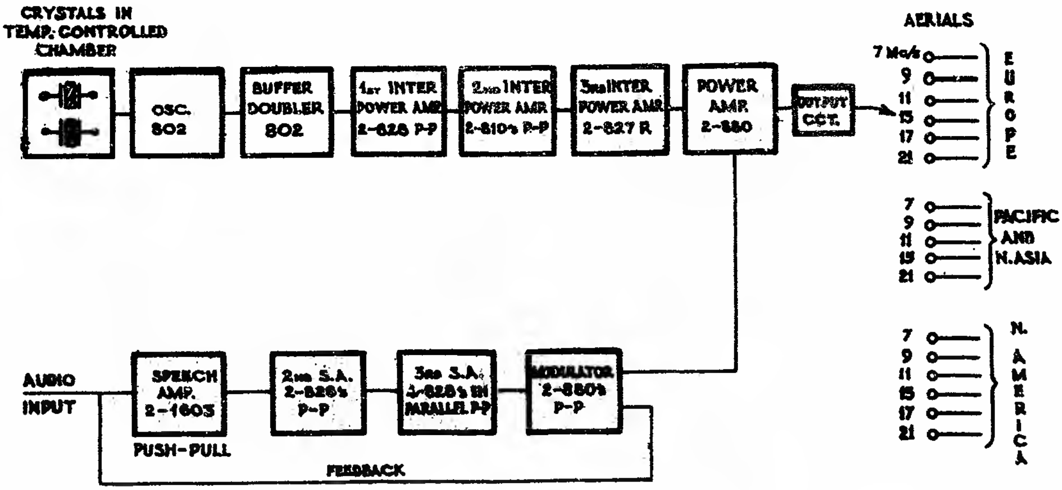


FIG. 34. 50 kW HIGH-FREQUENCY BROADCAST TRANSMITTER.

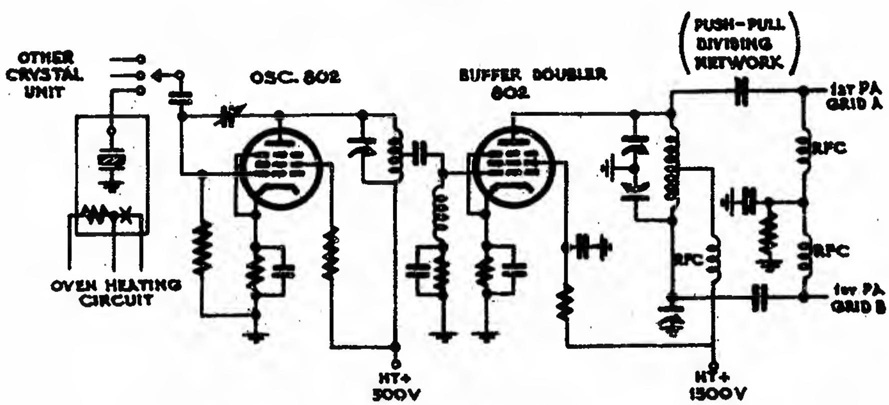


FIG. 35. OSCILLATOR-BUFFER STAGES.

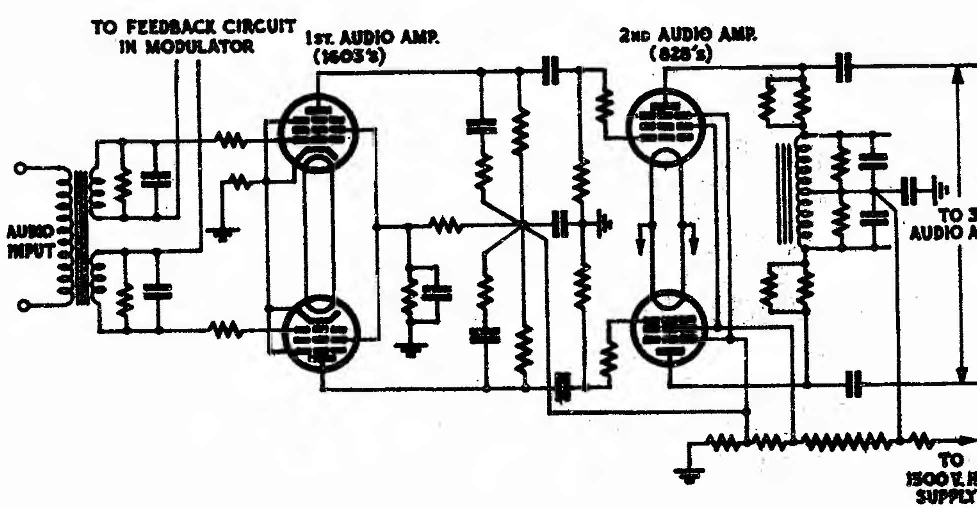
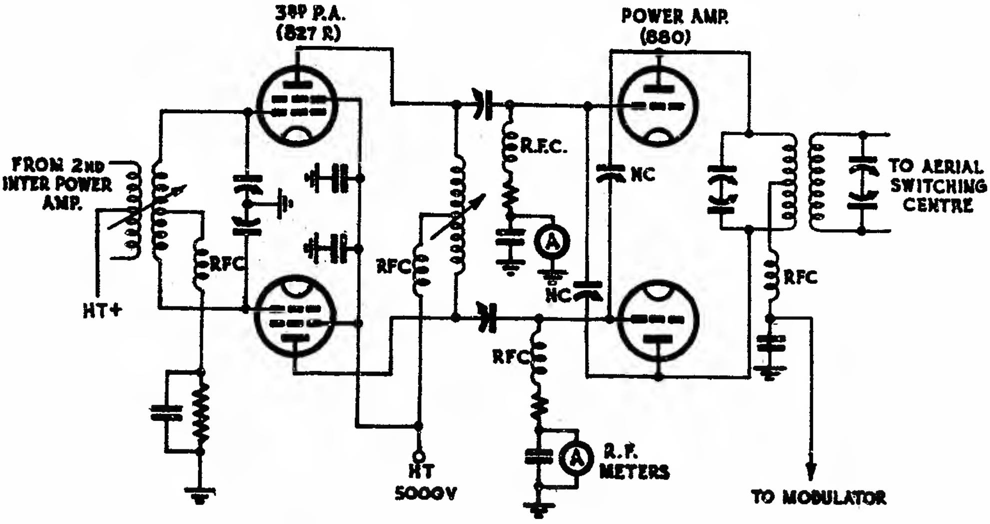
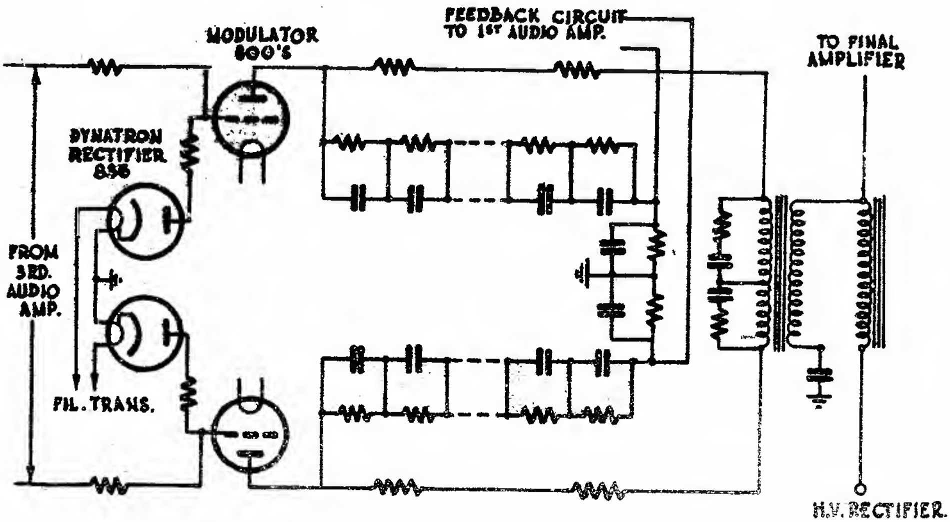


FIG. 36. FIRST AND SECOND AUDIO AMPLIFIERS.



THIRD POWER AMPLIFIER AND FINAL AMPLIFIER.

FIG. 37.



MODULATOR.

FIG. 38.

7. RADIO TRANSMITTER ADJUSTMENT.

7.1 Some general notes on tuning and adjustment of Radio Transmitters are now given. It must be understood, however, that these notes are fairly broad and give only a general idea of the principles involved.

The performance of a transmitter depends to a large extent on correct tuning.

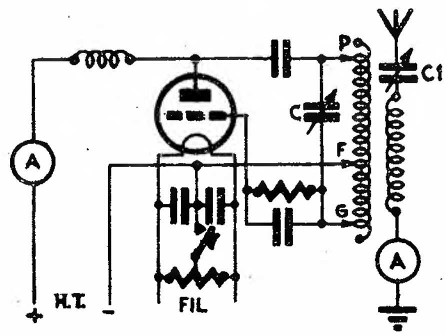
To tune a transmitter effectively, it is necessary that an operator be provided with the means of listening to the signals. A calibrated frequency meter and a monitor are required in order that the signal is heard while the transmitter is tuned. In addition to the monitor, an extremely desirable aid to tuning is a "tuning indicator". One such piece of equipment comprises a radio-frequency meter connected to a single turn of heavy wire for coupling to the tank coil of the oscillator or amplifier. This loop is usually about 2 or 3 inches diameter and absorbs a small amount of energy from the tank coil of the transmitter, which causes a reading on the galvanometer. With this instrument, it is readily ascertained whether radio-frequency currents are circulating in the tank circuit. (Paper No. 11 covers Measuring Instruments.)

It is also necessary to provide meters to enable the operating conditions of the transmitter to be checked. The more important of these meters are D.C. ammeters and voltmeters to read the anode currents and voltages. These meters enable the power input to the valves to be calculated and checked, and thus minimise the risk of damage to valves and equipment right at the start. Another very desirable meter is a thermo-couple ammeter which is connected in the aerial or feeder circuit. When this meter is connected at a point of maximum current in the aerial (a current antinode), its reading gives a good indication of changes of power in the aerial with changes in the transmitter adjustment.

A D.C. ammeter suitably connected in series in the grid bias circuit of an amplifier provides a convenient method of measuring the rectified grid current. The value of rectified grid current provides a means of ascertaining the radio-frequency voltage and power supplied to the grid circuit of the amplifier.

7.2 Tuning the Oscillator. Hartley Circuit. The self-excited oscillator of the Hartley type shown in Fig. 39 is used as a master oscillator in some transmitters operating in the medium wavelength. A description of the adjustment for best output of this type of equipment is given as a typical case.

Assuming that the correct values of cathode and anode voltage are available, the first step in adjusting a Hartley oscillator is to disconnect the aerial circuit from the transmitter and to connect the voltage to the valve cathode.



HARTLEY OSCILLATOR.

FIG. 39.

The successful functioning of such a transmitter depends largely on the correct positioning of the inductance coil clip connections P, F and G (Fig. 39). The cathode clip F is usually connected near to or at the centre of the inductance, and the anode clip P and the grid clip G are varied until the correct feedback is secured to maintain steady oscillation while the tuning capacitor C is varied from minimum to maximum values. More turns are usually included from F to P than from F to G. The ratio of turns between the grid and cathode clip are found to be between 1 : 4 or 1 : 5.

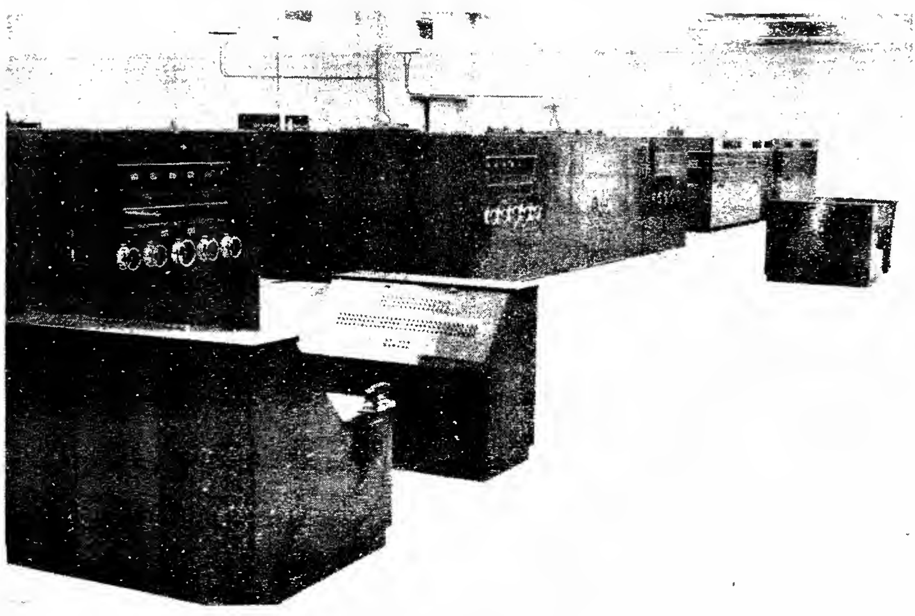
When the inductance clips are placed in the approximately correct positions on the coil, the high tension voltage is connected. The tuning indicator is loosely coupled to the inductance and the key closed. When the circuit is oscillating, a reading is given on the tuning indicator and the anode current is of low value. When the circuit is not oscillating, the tuning indicator indicates zero and the anode current is abnormally high, possibly two or three times the rated value. This condition is harmful to the valve, and the anode high tension supply must be disconnected.

Note that, when oscillators obtain the grid bias from the self-rectification of a portion of the high-frequency current (grid leak bias), very high anode currents are obtained when oscillation ceases, whereas with a fixed bias on the grid this is avoided.

It is now necessary to readjust the inductance taps to obtain oscillation. When the grid tap is too close to the cathode tap, the valve does not oscillate on account of the lack of excitation. When the cathode tap is too close to the anode tap, the excitation is excessive and the efficiency is low.

When the above adjustments have been satisfactorily made, the frequency meter is loosely coupled to the tank circuit and the latter adjusted by means of the capacitor C, Fig. 39, to the correct operating frequency.

When it is known definitely that the operating frequency is correct, the aerial coil is connected and very loosely coupled to the tank coil. The aerial is now tuned by varying the capacitor C1 and increasing the coupling until a reading on the aerial ammeter is obtained. The aerial coupling is now increased until the valve is drawing the rated anode current value. This setting, however, is too critical for stable oscillation, so that aerial coupling is reduced until the point is reached at which the anode current is about 85 per cent. of the previous value. The frequency is again checked with the frequency meter to ensure that the frequency has not shifted with the change in coupling of the transmitter. After compensating for any frequency variations, the aerial is retuned for maximum output at the correct value of anode current for stable operation.

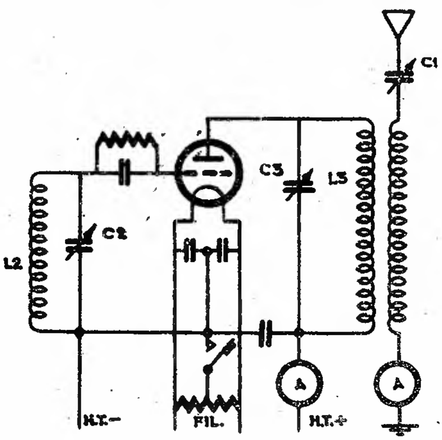


RADIO AUSTRALIA.

(VLC in far corner, VLB in centre, and part of VLA at left of picture. Aerial switching panels and control desks are in left foreground and at right.)

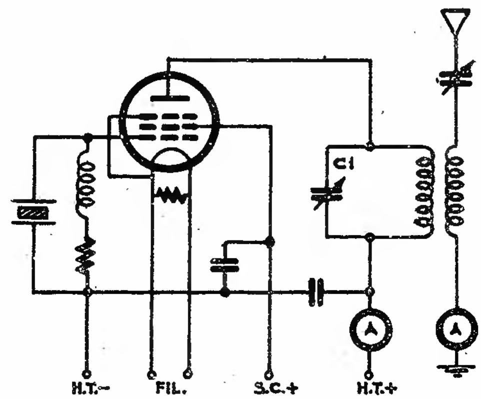
7.3 Tuned Anode Tuned Grid Circuit. In the tuned anode tuned grid circuit of Fig. 40, the tuned circuit L3C3 is tuned to a frequency slightly higher than L2C2, and, under these conditions, sufficient energy is fed back from the anode circuit to the grid circuit through the grid anode capacitance of the valve to maintain continuous oscillation. Failure of oscillation in this circuit is generally associated with an incorrect value of grid leak and capacitor.

7.4 Crystal Oscillator. The crystal oscillator, Fig. 41, is easy to adjust as there is little the operator can do to change the frequency or to affect adversely the frequency stability. Tuning, therefore, becomes chiefly a matter of obtaining the optimum amount of power from the oscillator.



TUNED ANODE TUNED GRID CIRCUIT.

FIG. 40.

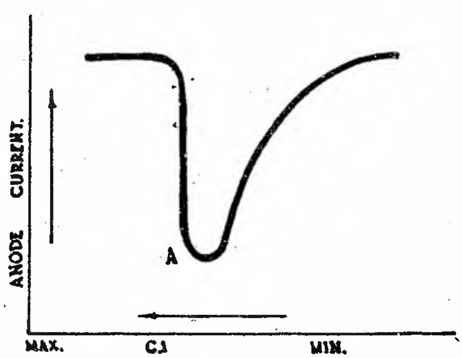


CRYSTAL OSCILLATOR.

FIG. 41.

Using an anode milliammeter as an indicator of oscillation, the anode current is found to be steady when the circuit is in the non-oscillating state, but dips when the anode capacitor is tuned through resonance at the crystal frequency. The behaviour of anode current, when the tank capacitor is varied, is shown in Fig. 42. When the capacitance is increased from minimum, there is a rather gradual decrease in anode current when oscillations commence. This decrease continued until the point A is reached, when there is a sharp rise in anode current, followed by cessation of oscillation. The tuning indicator shows maximum reading at point A.

When power is taken from an oscillator, the dip in the anode current shown at resonance is less pronounced. The greater the power output, the less is the dip in anode current. When the load is too great, oscillations cease. The greater the loading, the smaller the voltage fed back to the grid circuit for excitation purposes. This means that the radio-frequency voltage across the grid filament circuit is reduced and oscillation ceases.



BEHAVIOUR OF ANODE CURRENT WITH VARYING TANK CAPACITY.

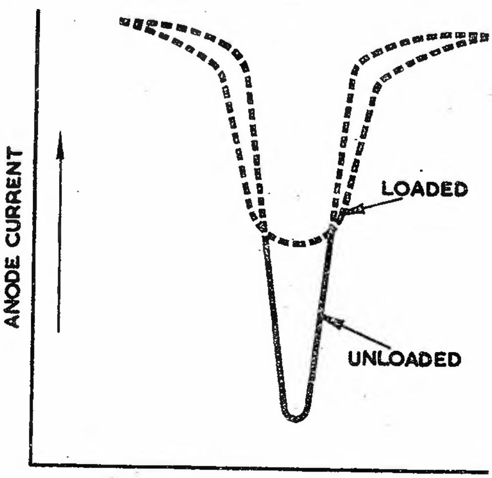
FIG. 42.

7.5 Adjusting the Amplifiers. The M.O.P.A. Transmitter. When the transmitter being adjusted is of the master-oscillator power-amplifier type, the next step, after the adjustment of the oscillator section, is the tuning of the amplifier. In describing this process, it is assumed that the neutralising has been carried to a satisfactory conclusion. A valve which is not properly neutralised behaves erratically when the anode voltage is applied.

The first step is to adjust the driver stage tuning for maximum amplifier grid current. Then the coupling between the stages is adjusted to give a further increase, if possible. The driver circuit is retuned to resonance every time the coupling is changed, no matter what coupling system is used, since a change in coupling is likely to throw the driver tank circuit slightly off tune. A few minutes spent in changing coupling and readjusting tuning shows quickly the optimum coupling for maximum grid current or excitation.

Once the proper grid-coupling adjustment is found, the amplifier anode tank capacitor is set approximately at resonance. With the excitation connected, the anode voltage is then applied and the anode tank circuit tuned to resonance, which is indicated by a very pronounced dip in anode current. This effect is explained in paragraph 7.7. This adjustment must be made quickly, since the valve cathode is damaged by continued application of anode voltage with the tank circuit tuned off resonance. (In preliminary adjustments, it is desirable to use low anode voltage until the amplifier is properly tuned.)

The off resonance anode current is usually much higher than the rated anode current for the valve - sometimes several times as great - but at resonance drops to 10 or 20 per cent. of the rated value. The higher the excitation power, the greater is the dip in anode current at resonance. When the dip in anode current is not pronounced, the excitation may be low or the valve may not be properly neutralised.



BEHAVIOUR OF ANODE CURRENT
WITH TANK TUNING AND LOADING.

FIG. 43.

When the tuning process has been carried this far, the amplifier is connected to the output load circuit. This load circuit is the aerial itself, through the coupling apparatus, or the grid circuit of a following amplifier. When the load is connected, the amplifier anode current rises. The anode tank circuit must be retuned for minimum anode current - this "minimum", however, is no longer the low value obtained at no load, but a new value nearer the rated anode current of the valve - since connecting the load probably will detune the tank circuit to some extent. The coupling to the load circuit must be adjusted, so that the new minimum anode current value is approximately the rated anode current of the valve. The typical behaviour of anode current with tank tuning and loading is shown in Fig. 43.

/ 7.6

7.6 Adjusting the Aerial Coupling. The last step in setting the transmitter in operation is to adjust the aerial coupling.

The fundamental rule in adjusting the aerial coupling is to start with very loose coupling and then couple tighter in small steps until one of three things occurs -

(i) The anode starts to show colour,

(ii) The anode current reaches the maximum allowable for the particular valve used, or

(iii) Aerial current stops increasing with an increase in aerial coupling.

When high anode voltage is used, it is probable that anode colour is the first limit reached. When low anode voltage is used, it is quite possible that the anode current reaches the maximum allowable value before the anode begins to show undue colour.

When the aerial current stops increasing with increases in aerial coupling before either the anode shows undue colour or the anode current reaches maximum, it indicates that there is not sufficient capacity in the anode tank circuit. In other words, less L and more C must be used to produce resonance.

The question of L to C ratio must be checked, even when anode colour or maximum anode current is the first limit reached. The L to C ratio is checked by the following test. Tune the anode tuning capacitor through resonance. When the maximum aerial current occurs at exactly the same point as minimum anode current, there is enough C in the circuit. When maximum aerial current and minimum anode current do not correspond, then more C and less L is necessary in the anode tank circuit.

In other words, it is seen that the tank tuning is sharp when the aerial is uncoupled, and the tuning becomes broader as the aerial coupling is increased. When there is not enough Q (circuit merit) in the tank circuit, it is possible to overcouple the aerial, which tremendously reduces anode efficiency and power output.

There can be too much C in the anode tank circuit, and this condition is verified by measuring the minimum anode current with the aerial uncoupled. When the minimum unloaded anode current does not fall below 10 per cent. of the anode current when properly loaded, there is too much loss in the anode tank circuit. When the tank coil is properly designed and built to have low losses and there are no high resistance joints in the tank circuit connections, the chances are that the C is too high, and more L and less C must be used to produce resonance at the operating frequency.

The following rules show the relationship between aerial coupling and the other circuit parameters.

The tighter the aerial coupling (everything else remaining constant), the higher the anode current, the greater the power output, the greater the anode loss and the lower the anode efficiency.

The tighter the aerial coupling, the more harmonics are radiated, and the lower are the losses in the anode tank, due to the lower circulating current.

/ 7.7

7.7 Theory of Adjustments. Tuning by the Anode Milliammeter Dip. In considering the adjustment of an amplifier stage (Figs. 44 and 45) of a radio transmitter, it is well to have clear ideas as to what is happening to the radio-frequency currents and how the readings of D.C. meters indicate any particular condition.

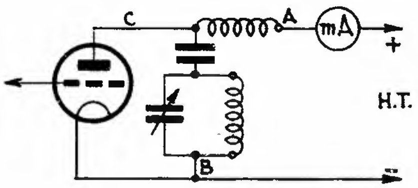


FIG. 44. PARALLEL FEED.

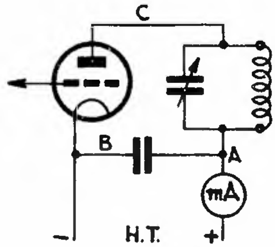


FIG. 45. SERIES FEED.

Two sources of power are supplied to the amplifier stage -

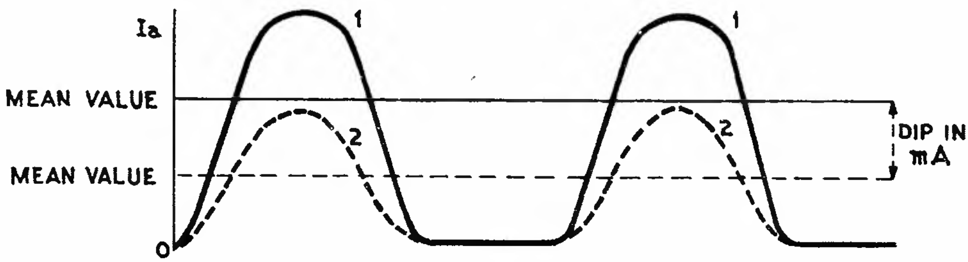
(i) The radio-frequency excitation from the driver or previous stage.
(ii) The D.C. anode input.

If purity of radio-frequency output wave form were the only requirement, all consideration of efficiency, as represented by the ratio of radio-frequency output to D.C. anode input, could be neglected and a high mean anode current provided (as indicated by the anode milliammeter), about which value the sinusoidal radio-frequency output would fluctuate. Such conditions sometimes apply in low power devices such as receivers, where high efficiency is not important, but, in transmitters using a considerable amount of power, efficiency becomes important. It is, therefore, customary to sacrifice purity of output for efficiency, and to purify the output by the use of output circuits carefully tuned and designed to provide the optimum impedance only to the desired frequency component of the distorted radio-frequency output. To this end, Class C amplifiers are usually associated with carefully designed output circuits.

In such an amplifier, the anode current is practically zero without radio-frequency excitation on the grid circuit, but, as soon as excitation is applied, a reading is obtained on the anode milliammeter, depending on the excitation applied and on the impedance of the output circuit to the radio-frequency currents produced therein.

Consider an amplifier energised from a driver but with its anode high tension disconnected. The radio-frequency voltage applied to the grid-cathode circuit of the valve has the same effect as in the "diode", and rectification is produced giving a reading on the grid milliammeter which shows the amount of excitation as the coupling or tuning of the input circuits is adjusted.

Now, knowing that the grid is being adequately excited, the anode high tension supply is connected. A radio-frequency current flows in the anode circuit, its value depending on the impedance of the circuit. But what does the anode milliammeter register? Not radio-frequency current, for it is a D.C. meter. It registers the mean value of current about which the radio-frequency currents fluctuate. (See Fig. 46.).



Curve 1 with low impedance tank circuit.
Curve 2 with high impedance tank circuit at resonance.

FIG. 46. WAVE FORM AT C IN CIRCUITS FIGS. 44 AND 45.

When considering parallel resonance of a tank circuit, it was explained that such circuits have a high impedance at resonance.

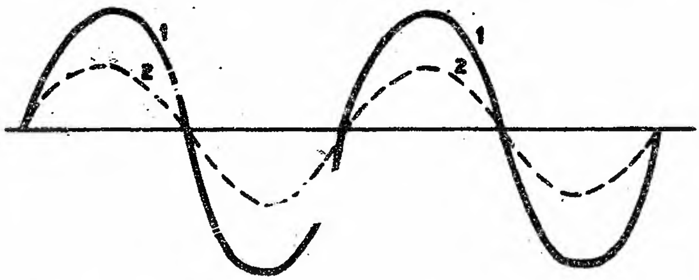
(In a parallel resonant circuit, $Z = \frac{L}{RC}$, and, although a high circulating current is developed in the tank circuit itself, the anode radio-frequency current is a minimum at resonance, being just sufficient to provide for the losses in the load R.)

Anode circuits of amplifiers are designed to give a high impedance at the desired frequency by having suitably designed tank circuits (with high Q) which are adjusted to resonance. When off resonance, the impedance of the anode circuit is greatly reduced, being an inductive impedance when the capacitor value is below the resonant value and capacitive when it is above the value.

As the capacitor of the anode circuit is adjusted for resonance, the radio-frequency current is varied in value and phase until at resonance it is a minimum. It has been seen, however, that with heavily biased amplifiers the mean anode current is proportional to the impedance of the circuit, that is, to the amplitude of the radio-frequency oscillations; therefore, the anode milliammeter gives a valuable indication and shows a dip (Fig. 43) as the tuning capacitor is passed through resonance.

This may be understood more clearly after studying the curves of a Class C amplifying valve, provided the fact is appreciated that every alternating or radio-frequency current which does not fluctuate about a zero value is equivalent in all respects to a D.C., and equal to the mean value of the A.C., plus an A.C. component which fluctuates about a zero axis. These two components can be separated, so that the alternating or radio-frequency component passes through one branch of a circuit having a suitable impedance and the D.C. component passes through another branch, including a meter. It is instructive to trace these separate branches in a typical amplifier anode circuit, and sketch the wave form of the electron flow in the several portions of the circuit.

Such an example is shown in Figs. 46, 47 and 48, which shows also the effect of reducing the amplitude of the radio-frequency currents.



Curve 1 with low-impedance tank circuit.
Curve 2 with high-impedance tank circuit at resonance.

FIG. 47. WAVE FORM AT B IN CIRCUITS FIGS. 44 AND 45.

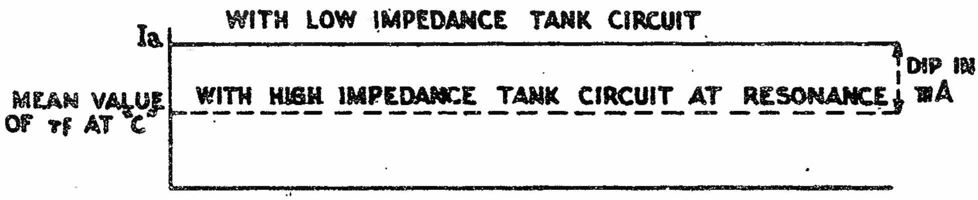


FIG. 48. WAVE FORM AT A IN CIRCUITS FIGS. 44 AND 45.

Now, on coupling the load or aerial circuit to the output tank circuit of the transmitter, the impedance is reduced and the anode current is again increased, the difference in reading being an indication of the power absorbed by the aerial circuit.

Power Input and Output. Considering the relation of power output and input, the product of the average values of the anode current and anode voltage is the measure of the average anode input power. The useful power output is determined graphically by analysing the anode current wave form for the fundamental frequency component (R.M.S. value) and taking the square of this value by the equivalent tuned circuit impedance.

A simple method of determining the power output from a radio-frequency power amplifier is by making a comparison of the power input conditions which exist when the amplifier is in a loaded condition and an unloaded condition. An amplifier is said to be loaded when it is supplying power to an external source, such as an aerial or the grid circuit of a following amplifier.

For example, in an amplifier with the aerial disconnected and the tank circuit tuned to resonance, the anode circuit milliammeter reads 30 mA. When the anode voltage is 1,000 volts, the power input in watts is -

$$IE = 0.03 \times 1{,}000 = 30 \text{ watts.}$$

When the amplifier is loaded by coupling up the aerial, the anode current rises to 150 mA and the power input is now $0.15 \times 1{,}000 = 150$ watts, which is an increase of $150 - 30 = 120$ watts.

It is seen that the increased power is being absorbed in the aerial circuit, but this does not mean the whole of this power - the 120 watts - is being radiated. The radiated power is somewhat less than this value, depending upon the aerial losses (such as leakage, brush discharge, I^2R losses, absorption and losses in the ground connection) and chiefly upon the type of aerial used. However, an indication of the general efficiency of the aerial system is obtained by this method of calculation.

The foregoing helps in understanding the operation of the transmitter shown in Fig. 49. The operation of the power plant is described in Paper No. 9 of Radio II.

The transmitter shown in Fig. 49 is of the master-oscillator power-amplifier type, also used in ships and mobile service. The master oscillator is of the Hartley type, while the power amplifier uses two valves in parallel to give the desired output. The grid bias for the amplifier is developed across the 1,000 ohm resistor, connected to the lower portion of the oscillator tank inductor. Resistors of 100 ohms are connected in each amplifier grid lead. These resistors are not provided for the purpose of biasing, but to suppress parasitic oscillations at a very high frequency which are very prevalent in circuits operating with valves in parallel. It must also be noted that the amplifier tank circuit capacitance is the aerial to ground capacitance in this case.

The tank circuit is resonated by varying the positions of the plate and filament-ground taps. The grid system of neutralising is used in the amplifier.

/ Fig. 49.

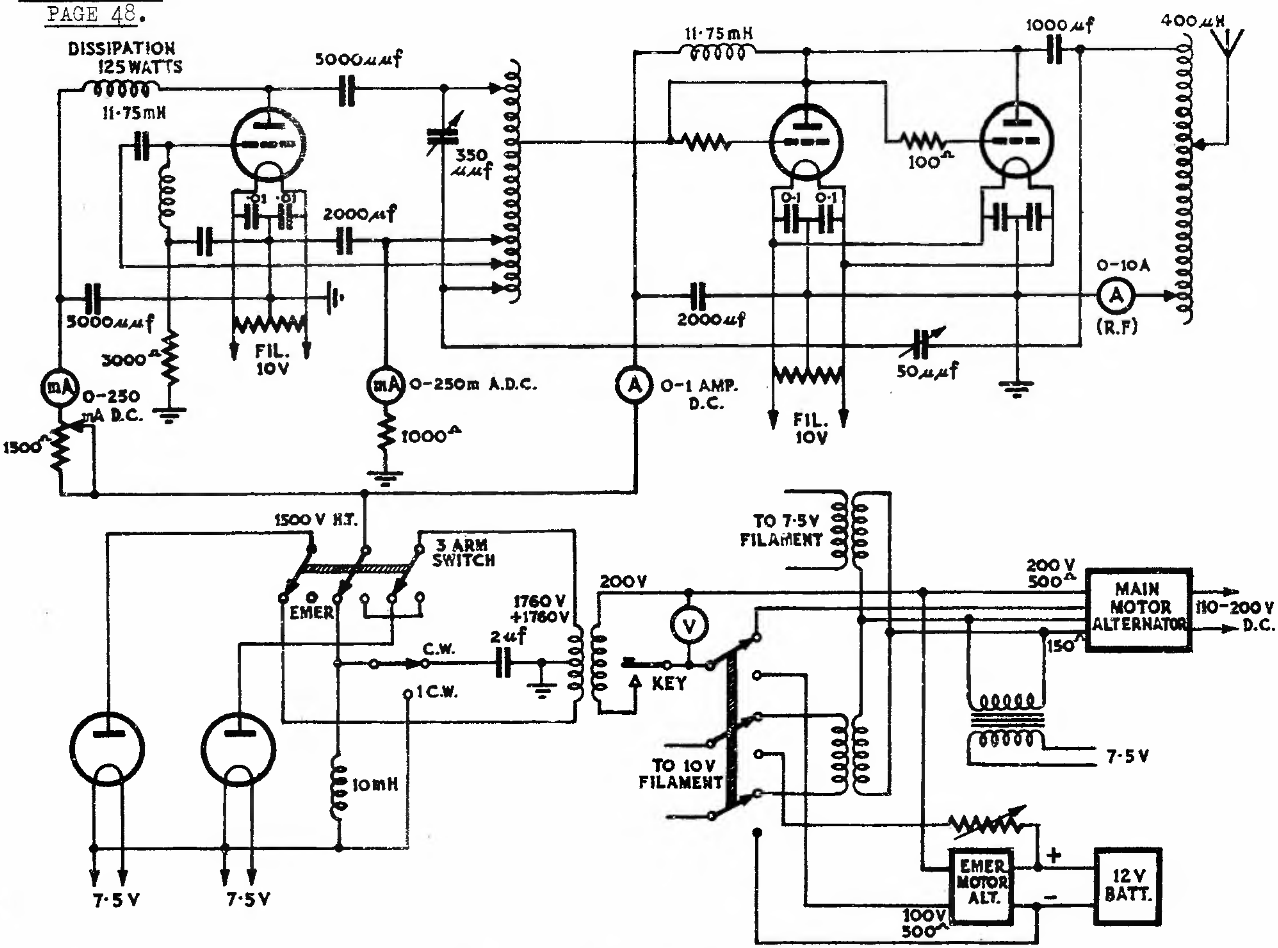


FIG. 49. TYPICAL MARINE TRANSMITTER (500 W).

8. TEST QUESTIONS.

1. Illustrate two methods of coupling radio-frequency amplifiers in a transmitter and mention their advantages or disadvantages relative to one another.

2. Define the terms shunt-feed and series-feed as applied generally to radio-frequency amplifiers.

3. Name and briefly explain three essential differences between broadcast transmitters as used in the medium-frequency and high-frequency bands.

4. Draw a sketch of an elementary one-valve transmitter and outline the principles of operation.

5. Explain by means of sketches three methods of obtaining grid bias voltages for radio transmitting valves. Which of these methods increases the liability of damage to valves? Under what condition would the damage be likely to occur?

6. Draw a block schematic diagram of a low-power medium-frequency broadcast transmitter designed to provide a high quality service.

7. Briefly describe two methods of coupling the final amplifier to the aerial, and refer to any particular advantages or disadvantages associated with their use.

8. Feedback is often used to improve the performance of a broadcast transmitter. Explain briefly a method of applying this and the advantages resulting therefrom.

9. With Fig. 30 as a guide, assemble Figs. 31-33 into a complete transmitter.

10. Assemble Figs. 34-38 in a similar fashion to Question 9.

END OF PAPER.

COMMONWEALTH OF AUSTRALIA.

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Postmaster-General's Department,
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COURSE OF TECHNICAL INSTRUCTION.

RADIO COMMUNICATION.

CONTENTS.

1. INTRODUCTION.

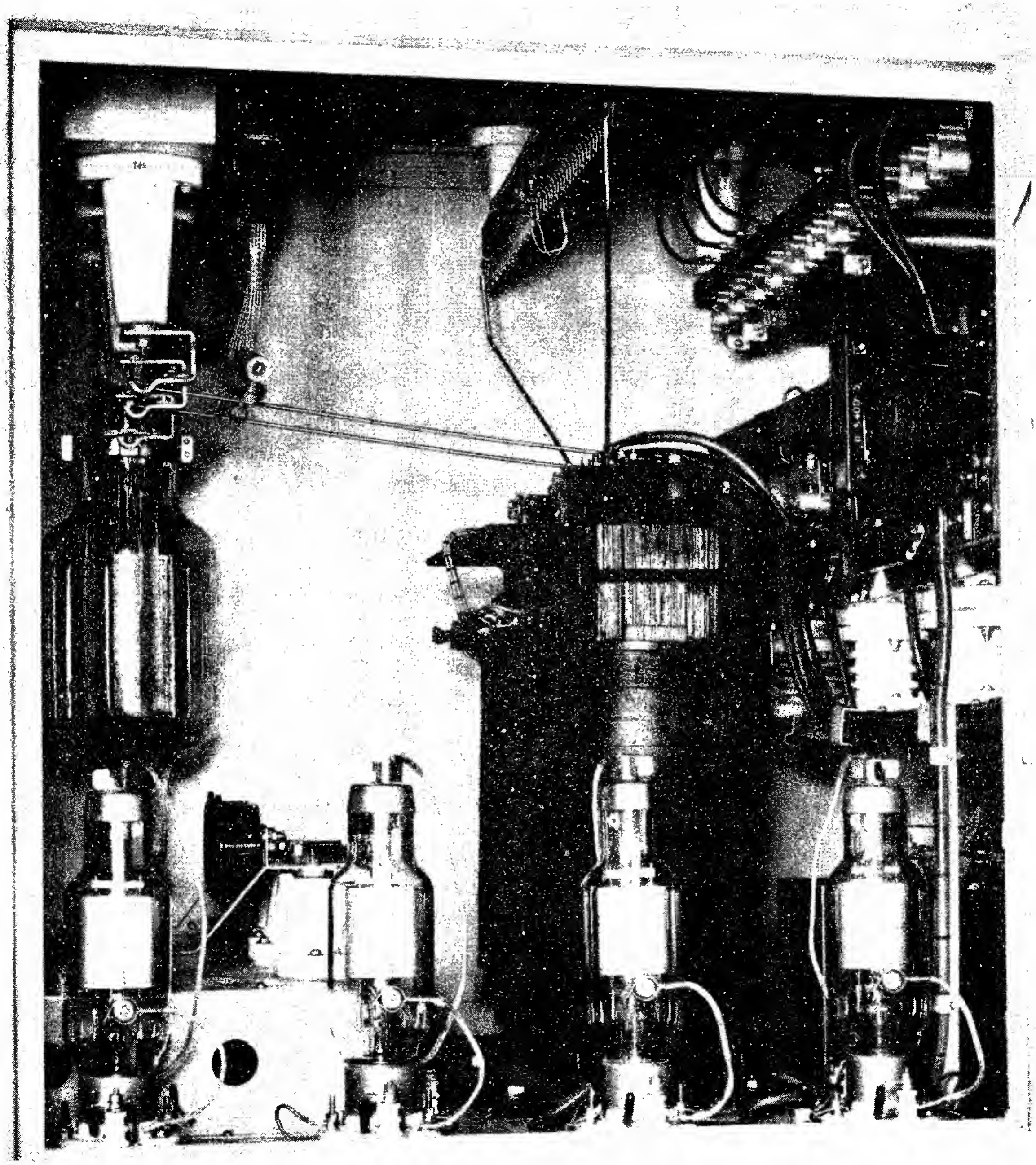
1.1 Radio-frequency currents are generated by a transmitter and used to stress effectively the propagation medium by means of an open oscillatory circuit or aerial designed to produce radio-frequency electric and magnetic strains in a maximum amount of space. These strains in space set up radio waves having a predetermined radio frequency, but move out into space at a constant velocity (3×10^8 metres per second). These waves are used to convey intelligence from the transmitting station to the receiving station.

Intelligence is usually one of two forms, either telegraphic or telephonic, but both are concerned with using these radio or high-frequency waves to convey audio or low-frequency effects which are discerned by the ear. Therefore, the problem is to use the radiation properties of the radio frequencies to convey the audio-frequency intelligence. The basic principles involved in the process are outlined briefly before describing the special equipment used at the transmitter and receiver to enable the audio-frequency intelligence to be conveyed by the radio-frequency waves and to be made available in an audio form at the receiver.

2. RADIO TRANSMISSION SYSTEMS.

2.1 The primary radio frequency oscillations sent out from a radio transmitting aerial are known as the Carrier Wave. This carrier wave provides the means for carrying the audio frequency, which is the electrical equivalent of the sound to be transmitted. In the ordinary type of radio receiver, the sound reproducing device is unable to respond to these radio frequencies, nor could the human ear respond to such radio-frequency vibrations, as they are above the limits of audibility. In order to produce an audible sound in a receiver, it is necessary to break up the radio-frequency waves into groups, or to "modulate" these at an audible frequency.

Before dealing with modulation systems and modulation, the several types of emission are defined. These types are as defined by the International Radio Convention, 1938, (Cairo).



MODULATOR UNIT AT 2 kW H.F. TRANSMITTER (S.T.C.).

2.2 Type A1 Continuous Waves (C.W.). The radio-frequency carrier at the radio station is usually controlled by a key arranged to suppress emission in its normal position, the result of keying being the transmission of the radio-frequency carrier in long or short pulses according to the requirements of the code. (See Fig. 1a.)

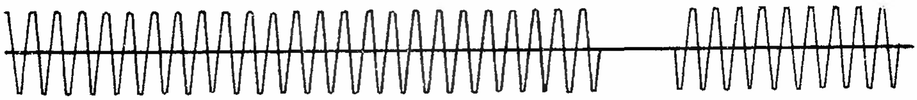


FIG. 1a. CONTINUOUS WAVE. (KEY CONTROLLED.)

The method has the merit of simplicity, but requires a special type of receiver to convert the received radio-frequency signals to an audible note, as the only sound heard on an ordinary receiver is a slight hissing sound. A receiver having reaction, or a local oscillator which generates a signal to "beat" with the received carrier and thus give an audible note, is required to hear these signals.

2.3 Type A2 Interrupted Continuous Waves (I.C.W.). This name originated from the method of breaking or interrupting the radio wave at an audio-frequency rate. A dash consists of a long train of radio-frequency carrier broken into short groups of radio-frequency oscillations, the number of groups depending on the rate of interruption. For example, suppose the carrier frequency is 500,000 c/s and a dash of 1 second duration is sent, then the dash consists of 500,000 complete oscillations (C.W.). When the carrier is interrupted 1,000 times per second, then the dash consists of 1,000 groups of radio-frequency waves each containing $\frac{500,000}{1,000} = 500$ radio-frequency oscillations (I.C.W.). (See Fig. 1b.)

FIG. 1b. INTERRUPTED CONTINUOUS WAVE.

Modulated Continuous Waves (M.C.W.). This wave is similar to the I.C.W., but the chopper is replaced by an audio-frequency modulating note having a sinusoidal wave form. This method enables pure reproduction of any note or group of notes or even speech. (See Fig. 1c.)

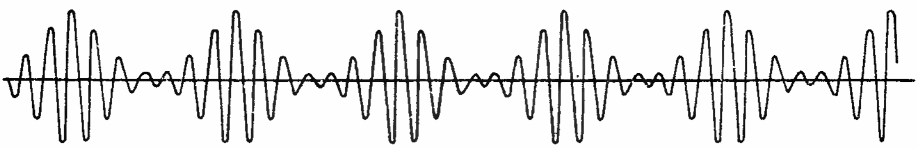


FIG. 1c. MODULATED CONTINUOUS WAVE. (KEY CONTROLLED.)

2.4 Type A3 Radio Telephony and Broadcasting. This is a special case of M.C.W., the difference being that the carrier wave is radiated continuously during transmission and that the equipment is designed for high quality service. (See Fig. 1d.) Reference to Fig. 1d shows that the carrier wave amplitude is not constant, and any system of modulation which thus affects the carrier is termed Amplitude Modulation.

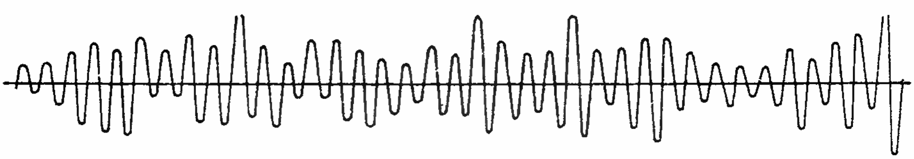
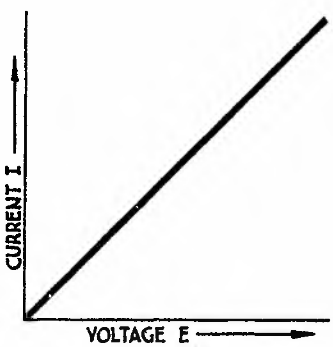


FIG. 1d. M.C.W. TELEPHONY.

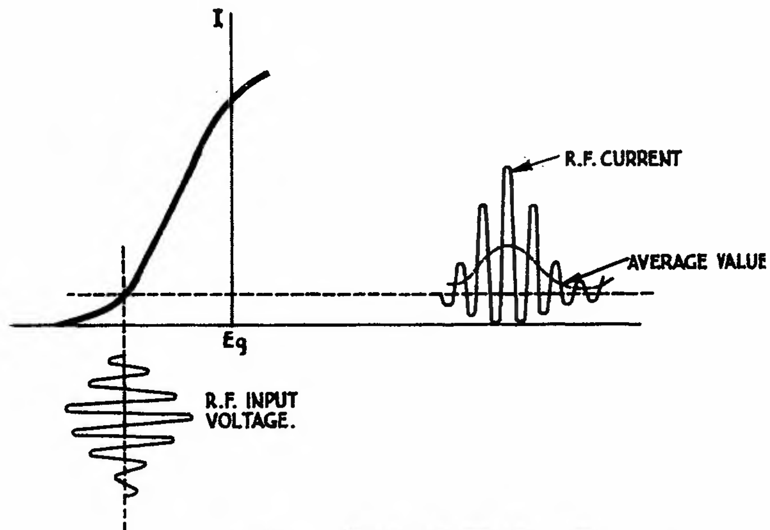
2.5 Much work has been done in recent years on a system whereby the amplitude of the carrier is kept constant and the modulating frequency causes a variation of the transmitter frequency at a rate corresponding to modulating frequency. This system is known as Frequency Modulation.

2.6 When I.C.W. or M.C.W. signals are received on an aerial, radio-frequency currents in similar groups or of similar varying amplitude to that in the transmitter aerial flow in the receiving circuits, which are tuned to provide a minimum circulating impedance. These currents could be used to operate any thermal device if they were of sufficient intensity, but, under practical conditions, they are so small in value that amplification is usually desirable, and even then there is little advantage in using them to operate any mechanical device. The requirements of radio telephony are the production of audio-frequency currents in a telephone receiver or loud-speaker, so that the appropriate sounds are heard by the ear. This is brought about by making use of the fact that certain substances and devices have a resistance, which varies with the amplitude and direction of the applied current. The characteristic curves of a Linear and a Non-Linear device are shown in Figs. 2 and 3 respectively.



CHARACTERISTIC CURVE OF "LINEAR" RESISTANCE.

FIG. 2.

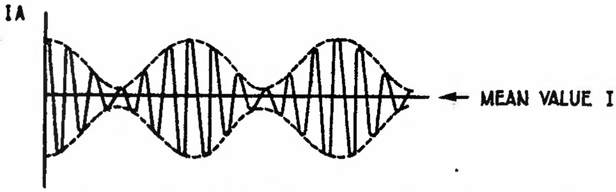


CHARACTERISTIC CURVE OF "NON-LINEAR" DETECTOR.

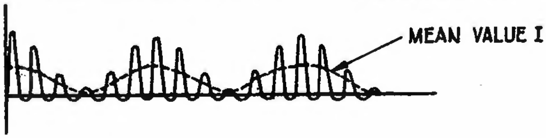
FIG. 3.

"Non-linear" devices are known as Detectors, as they "detect" the presence of radio-frequency current by producing a varying D.C. equal to the mean value of radio-frequency current passing through these and varying in accordance with the amplitude. Thus, a detector is used to produce a varying D.C. or an audio current from a circuit containing radio frequency current of varying amplitude. A detector does this by distorting the output, because of the non-linearity of its voltage/current curve, and producing an output which, instead of swinging symmetrically about a steady value, swings symmetrically about a varying mean value (Fig. 4) which follows the variations in amplitude of the radio-frequency wave. This varying mean current follows exactly the intelligence impressed on the transmitter. The current is passed through a low-frequency circuit of suitable impedance in the output of the detector of a receiver, in which is incorporated a telephone receiver which follows this fluctuating "mean value" current and producessound waves therefrom.

/ Fig. 4.



(a) Symmetrical Radio-Frequency Received Current.

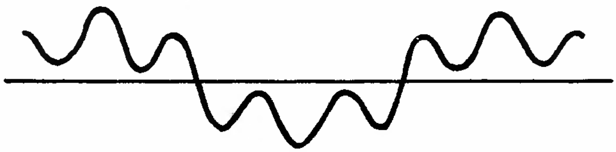


(b) Distorted Radio-Frequency Output from a Detector Stage of a Receiver (Showing Fluctuating Mean).

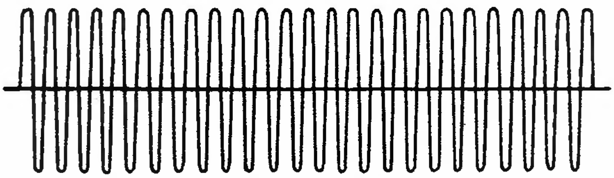
FIG. 4. CURVES OF DETECTION.

2.7 Basic Radiation is pictured, as shown in Fig. 5, as a "carrier" of the audio-frequency intelligence.

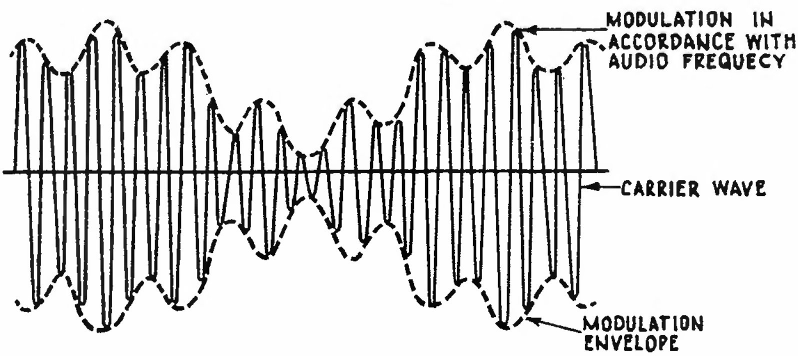
In the transmitter, the "carrier wave" is MODULATED to produce the variations of amplitude which convey the "signal frequency," while, in the receiver, a type of distortion or partial rectification (known as DETECTION) takes place to produce a current at the "signal frequency" which actuates the telephone receivers or loud-speakers.



(a) Modulating Current (Audio Frequency).

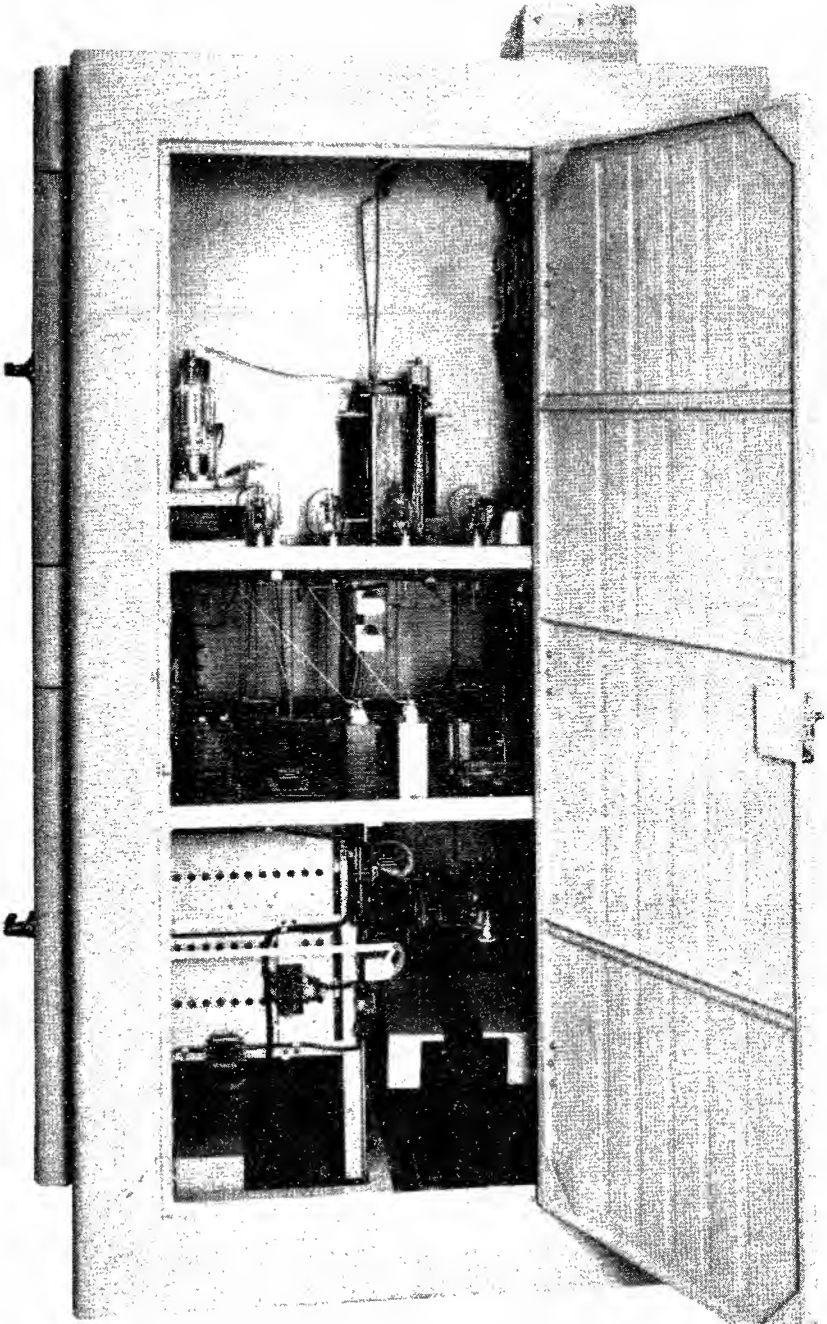


(b) Carrier Wave (Radio Frequency.)



(c) Modulating Carrier Wave (Radio Frequency).

FIG. 5. REQUIREMENTS FOR THE TRANSMISSION OF INTELLIGENCE.



MODULATOR-OSCILLATOR IN THE 500 WATT M.F.
TRANSMITTER (S.T.C.).

3. MODULATION.

3.1 Modulation, therefore, is defined as the process of imparting to the radio-frequency carrier wave the intelligence it is desired to transmit. There are three types of modulation now in use -

(i) Amplitude Modulation, in which the carrier frequency remains constant, but its amplitude is varied in accordance with the level of the applied audio frequency and at a rate equal to this latter frequency.

(ii) Phase Modulation, in which the carrier amplitude remains constant but the phase angle varies at a rate equal to the applied audio frequency, the degree of variation depending on the magnitude of the applied audio frequency.

(iii) Frequency Modulation, in which the carrier amplitude remains constant, but the frequency is periodically varied about the unmodulated value at a rate equal to the applied audio frequency, the deviation from normal frequency depending on the magnitude of the applied audio frequency.

Types (ii) and (iii) are not dealt with at this stage.

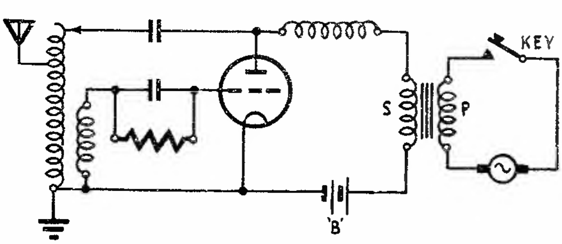
3.2 Amplitude Modulation. The power developed in the tuned circuit of an amplifier, and hence the amplitude of the carrier wave, depend on several different factors including the following -

(i) The value of the H.T. voltage applied to the amplifier anode.
(ii) The steady potential of the grid.
(iii) The relation between the impedance of the tuned circuit and the A.C. resistance of the valve.

An alteration of any one of these factors produces a change in the carrier amplitude. It is seen that, when the alteration is an audio-frequency variation whose amplitude corresponds to that of music or speech vibrations (Fig. 5a), the carrier amplitude is modulated in amplitude in a similar manner as shown in Fig. 5c.

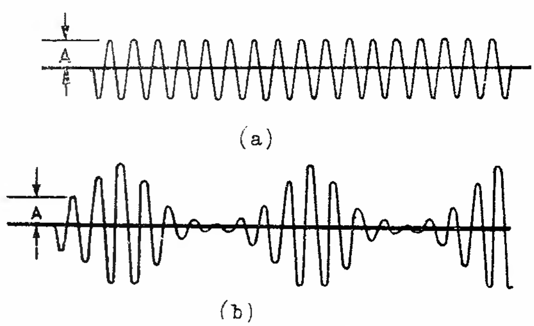
3.3 Theory of Amplitude Modulation. The effect of amplitude modulation is explained as follows. Let the anode circuit of the transmitter (Fig. 6) contain a source of continuous current and an induced A.C. from the transformer PS. Normally, the anode receives a supply of "B" battery current, and oscillations of amplitude A (Fig. 7a) are set up in the aerial. When the key is closed, an A.C. is impressed on the anode current, and the voltage of the anode varies according to the frequency of this impressed alternating voltage. When the peak value of the alternating voltage is in the same direction as the D.C. anode voltage, the total anode voltage is increased. When the A.C. cycle reverses, the anode voltage is reduced. The anode voltage, therefore, can be made to vary between zero and twice the D.C. voltage.

When such a transmitter is correctly adjusted, the aerial current varies with the applied anode voltage. Therefore, in this case, the aerial current varies between zero and twice the D.C. value, as shown in Fig. 7b. The oscillations represented are completely modulated and are said to be modulated 100 per cent.



SIMPLE CIRCUIT FOR PRODUCING A.M. WAVES.

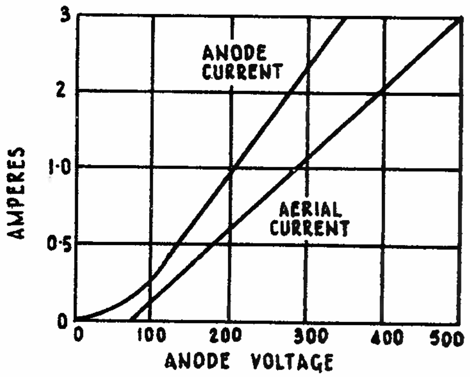
FIG. 6.



AMPLITUDE MODULATED WAVES.

FIG. 7.

Fig. 8 shows the relation of anode voltage to aerial and anode feed current.



RELATION OF ANODE VOLTAGE TO AERIAL AND ANODE FEED CURRENT.

FIG. 8.

Power is required to cause an aerial current to vary at audio frequencies, that is, to modulate the carrier, It can be shown that, when the aerial current is caused to vary according to a sine wave, the power in the aerial increases according to the following formula -

$$P = \left[(I1)^2 + 0.5\ (I2)^2\right] R.$$

where P = Power in watts for any degree of modulation,

(I1) = Aerial current without modulation,

(I2) = Maximum variation of aerial current, and

R = Effective aerial resistance in ohms.

If the carrier is not modulated, the power in the aerial is -

$$P = (I1)^2\ R.$$

With complete modulation, that is, when I1 equals I2, the aerial power from the previous formula is -

$$P = 1.5\ (I1)^2\ R.$$

The 50 per cent. increase in aerial power, which corresponds to approximately 22 per cent. increase in the aerial current for 100 per cent. modulation, is the additional power being radiated because of the forced variation of the aerial current due to modulation. In the above case, the added or sideband power is supplied by external sources, or by a change in anode circuit efficiency resulting from suitable grid excitation of the radio-frequency amplifier.

The aerial current from a Class C radio-frequency amplifier is essentially proportional to the anode current and voltage supplied to the anode of the amplifier valve, so that, when the anode current is caused to vary, the aerial current varies in the same manner. This indirect way of causing a variation of aerial current requires considerable power to modulate the carrier 100 per cent., because of the losses in the radio-frequency amplifier output valve. The modulation of the anode supply to a Class C radio-frequency amplifier is commonly called anode modulation, and the power for modulation must be supplied from an external source. Therefore, the greater the efficiency of the output valve and coupling system, the greater is the output power and corresponding sideband power. To determine the modulation efficiency, it is necessary to make a few calculations on the anode circuit values to determine the power requirements.

As has been previously stated, the anode current, when modulated, is in reality an A.C. impressed on the steady D.C. supplied to the amplifier valve. This condition is represented in Fig. 9, in which Ia is the value of anode current without modulation and Imax is the maximum audio variation of anode current when modulated. The power represented by these two currents is found by using the formula -

$$P1 = I_a^2 R1 \text{ watts.}$$

/ Fig. 9

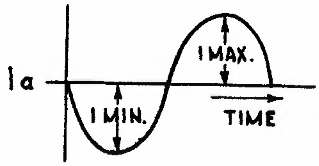


DIAGRAM OF INSTANTANEOUS VALUES OF MODULATED ANODE CURRENT.

FIG. 9.

The anode voltage is known as well as the D.C. value of the anode current, so that -

$$P1 = IaEa \text{ watts}$$

in which Ea = Anode supply voltage,

and P1 = Power input to amplifier valve.

$$\therefore IaEa = I_a^2 R1$$

$$\text{and } R1 = \frac{Ea}{Ia}.$$

Since Ia and Ea are constant, the power required to modulate Ia is $\frac{Imax^2 R1}{2}$ watts, which modulates the anode current $\frac{Imax}{Ia}$ per cent. When the modulation is 100 per cent., Imax = Ia and the power required to modulate in the anode circuit is 50 per cent., of the D.C. power to the anode.

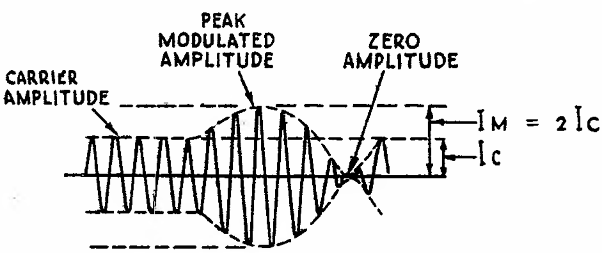
3.4 Percentage Modulation. The degree of amplitude modulation is described in terms of the amplitude variation of the transmitted wave, and is usually given as a decimal modulation factor or as a percentage. The modulation factor expressed in percentage is 100 times the maximum departure (positive or negative) of the envelope of a modulated wave from its unmodulated value divided by its unmodulated value.

Fig. 10 shows curves of amplitude modulation, with the amplitude relations for determining percentage modulation shown.

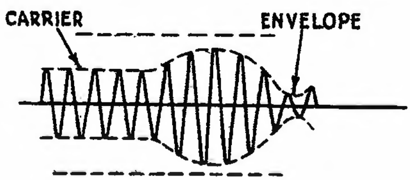
In the form of an equation, the expression for percentage modulation is -

$$\text{Percentage M} = \frac{Im - Ic}{Ic} \times 100$$

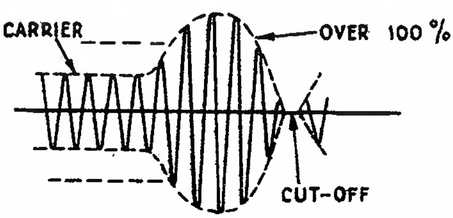
where Im = the maximum amplitude of the positive peak or the maximum amplitude of the negative peak, and Ic = the unmodulated carrier amplitude. In the case of over-modulation shown in (c), the positive percentage is greater than 100. However, the negative percentage is never greater than 100, because the amplitude cannot become less than zero. Such a condition results in a distortion of the wave envelope, the envelope being the outline of the radio-frequency cycle peak.



(a) 100 Per Cent. Modulated.



(b) Less than 100 Per Cent. Modulated.



(c) Over-Modulated.

GRAPHICAL REPRESENTATION OF AN AMPLITUDE MODULATED WAVE.

FIG. 10.

/ 3.5

3.5 Sideband Frequencies. While it is usual to modulate a carrier wave by varying its amplitude at the desired audible frequency, it is possible to produce an exactly equivalent modulated output (where one note only is to be conveyed) by mixing with the carrier wave another radio-frequency wave slightly different in frequency, and this mixture is effected at any point. It can be proved mathematically or graphically that a mixture of two pure waves gives a resultant wave whose amplitude varies at a frequency equal to the difference between the two equal amplitude radio-frequency waves. For instance, a wave with a frequency of 1,000 kc/s, when mixed with a wave having a frequency of 1,001 kc/s or 999 kc/s, is modulated or varied in amplitude with a frequency of 1 kc/s. This effect is noted when two stations interfere by transmitting carrier waves with adjacent radio frequencies, and is well-known to all broadcast listeners. This effect also is produced by feeding the final amplifier from two master oscillators differing slightly in frequency. Also, it is produced in a C.W. receiver, where the extra radio frequency is not mixed with the carrier wave until it has reached the receiver.

At this stage, it is important to appreciate that a modulated wave is merely a mixture of radio-frequency waves however it is represented, graphically or mathematically. In fact, all modulated or fluctuating radio-frequency waves are shown to be mixtures of pure sine waves of constant amplitude, the number and frequency distribution of the components increasing with the irregularity or abruptness of the changes in the amplitude of the combined mixture. This applies no matter how the complex non-sinusoidal wave form is produced.

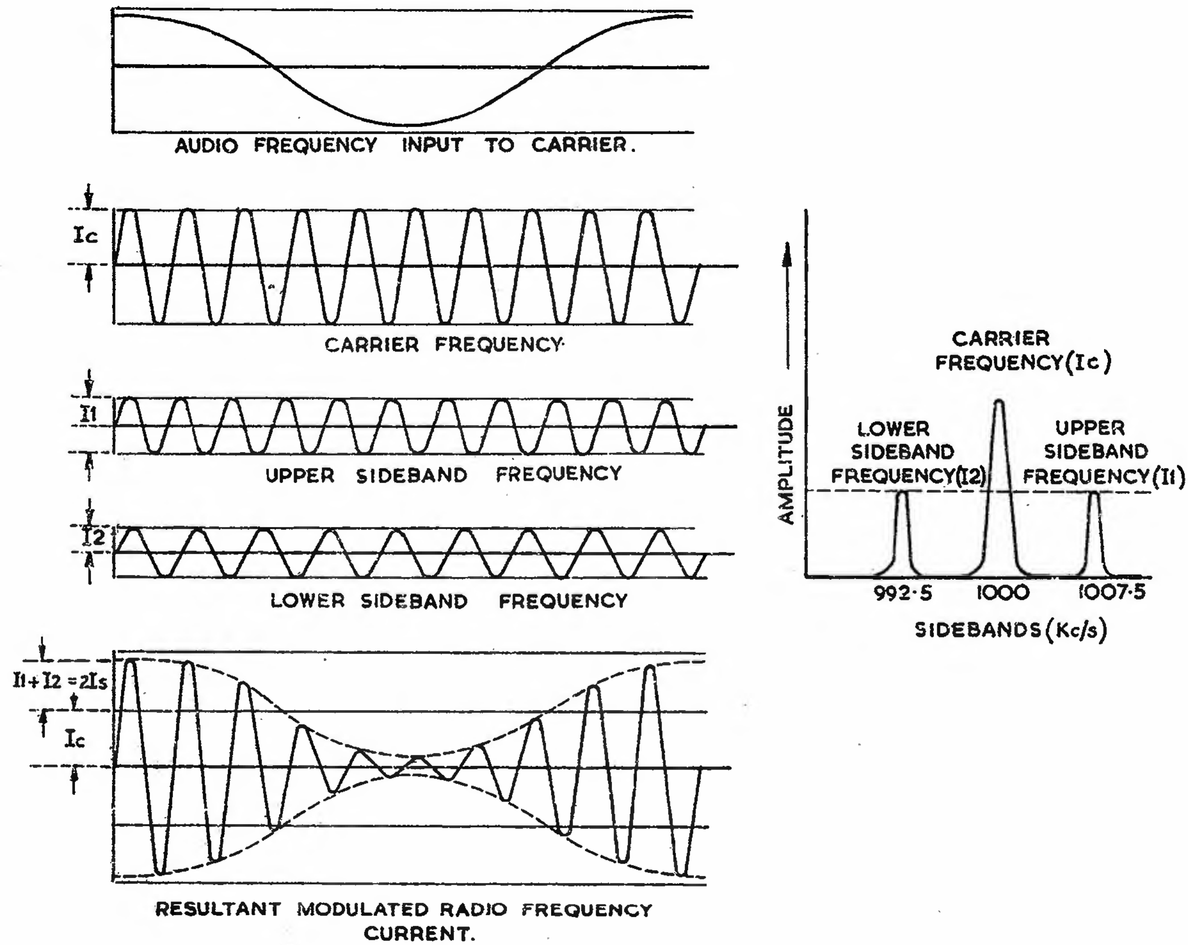
It is only necessary to mix one frequency with the carrier to produce modulation at any given note, and such "single sideband" systems exist and are becoming more important. These systems have been developed to the stage where the extra or "sideband" frequency alone is transmitted, the carrier frequency being provided at the receiver. However, in the simple modulation system as used in M.C.W. telegraph transmitters and in broadcast transmitters, it can be shown that the complex varying amplitude wave modulated by a single note is equivalent in all respects to a mixture of three waves. When "c" c/s is the radio-frequency carrier frequency and "m" c/s is the modulating note, the complex modulated wave is equivalent in all respects to three component pure sine waves with frequency $(c - m)$, c and $(c + m)$ c/s.

Thus, radio frequencies having numerical values equal to the carrier frequency plus and minus the modulation frequency are produced by the process of amplitude modulation. These frequencies are symmetrically distributed on both sides of the carrier frequency in what are called the upper and lower sidebands. At 100 per cent. modulation, the combined maximum amplitude of the two sidebands is equal to the carrier amplitude or the maximum amplitude of one sideband is one-half the carrier amplitude. With 100 per cent. modulation by a single tone, the two sidebands contain one-third of the total average power.

The sideband view-point in some cases, presents a more desirable picture of amplitude modulation, as it shows that, during the process of modulation, the modulated signal appears as three components (or more) on three (or more) different frequencies. This view-point is shown in Fig. 11.

Assume that a carrier frequency of 1,000 kc/s is modulated by a steady audio frequency of 7.5 kc/s. The process of modulation results in the generation of two additional frequencies, one of 1,007.5 kc/s and another of 992.5 kc/s. These frequencies are known as the upper and lower sidebands respectively. As this interaction of frequencies occurs in the transmitting circuit, these reactions produce a resultant amplitude change in the output wave form, and its form is as shown in Fig. 11. In other words, the resultant wave-shape represents the instantaneous sum of the three radio frequencies.

/ Fig. 11



RELATIONSHIP OF SIDEBANDS TO CARRIER FREQUENCY.

FIG. 11.

It is this wave which represents the true picture of the modulated output emanating from the transmitter during the process of modulation. The ratio of the sum of the two sidebands divided by the carrier frequency is generally referred to as the modulation factor, expressed -

$$m = \frac{I1 + I2}{Ic} = \frac{2Is}{Ic}.$$

The interaction of the sideband frequencies with the carrier frequency, produces (on detection) the resultant or beat frequency. Hence, in the case of 7.5 kc/s modulation, there are sideband frequencies of 7.5 kc/s above and 7.5 kc/s below that of the carrier frequency or a band-width of 15 kc/s, but the actual resultant frequency due to the interactions is 7.5 kc/s. The intelligibility, therefore, is entirely due to the sideband frequencies and their reactions with the carrier and not to the carrier frequency alone.

/ The

The power contained in the sidebands depends upon the variations in the amplitude of the wave. Consequently, when the modulation cycle is complete or 100 per cent., the maximum amount of power is contained in the sideband. As the intelligence bearing portion of the modulated wave is not the carrier but the sidebands, it follows that the greater the sideband amplitude the greater is the useful portion of the wave. For a wave completely modulated (that is 100 per cent.), two-thirds of the total wave energy is concentrated in the carrier with the remaining one-third in the sidebands when the modulating signal is a pure tone. The sideband power varies between this one-third maximum and zero directly as the square of the modulation degree.

When the sideband power is calculated, it is found that the following relations hold.

Percentage Modulation.	Percentage of Total Power in Carrier.	Percentage of Total Power in Sidebands.
0	100	0.0
25	97	3.0
50	89	11.0
75	78	22.0
100	66.6	33.3

(Note. These power relations hold strictly only with a sinusoidal (pure tone) modulating signal. For speech modulation, the sideband power values are approximately halved.)

Hence, in a wave modulated 25 per cent. by a pure tone signal, only 3 per cent. of the total power is useful power. For 50 per cent. modulation, the useful power is increased to 11 per cent. of the total with single tone modulation. A 100 per cent. modulated wave delivers 11 times as much energy as one modulated 25 per cent., as shown in the above table.

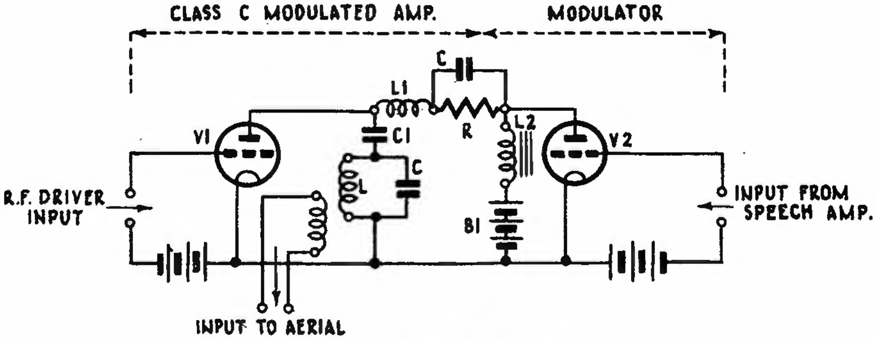
3.6 Amplitude Modulation Methods. Modulation takes place through any of the valve electrodes. Typical types of modulation are -

(i) Anode modulation,
(ii) Grid modulation,
(iii) Suppressor-grid modulation,
(iv) Screen-grid modulation,
(v) Cathode modulation, and
(vi) Push-pull grid modulation.

Anode Modulation is perhaps the most commonly used, as it provides excellent linearity up to high degrees of modulation, has low distortion and reasonable efficiency. A typical anode-modulated amplifier is the "Heising", and it is essentially an ordinary Class C amplifier in which the modulating voltage is superimposed upon the D.C. anode-supply voltage. This makes the total effective anode voltage consist of the sum of the D.C. anode-supply voltage and the modulating voltage, and so corresponds to the desired modulation envelope. Typical circuit arrangements are shown in Figs. 12 and 13, the difference being the use of transformer coupling in Fig. 13, instead of the choke in Fig. 12. Consider Fig. 12 in which V1 is the modulated amplifier (mod-amp) and V2 the modulator. In this arrangement, both V1 and V2 receive their anode supply from a common source through a large choke coil or anode reactor L2. The successful operation of this system depends almost entirely upon the design of the reactor choke coil.

/ In

In Fig. 12, the anode power supply for the Class C amplifier and modulator valves is from a common source B1, the anodes being fed through the choke L2 which operates as a transformer having a 1 : 1 ratio. The modulation frequency voltage developed across this choke is added to and subtracted from the supply voltage impressed on the anode of the amplifier valve, giving the resultant modulation as previously indicated. To maintain fidelity at low audio frequencies, the choke L2 must be of high inductance.

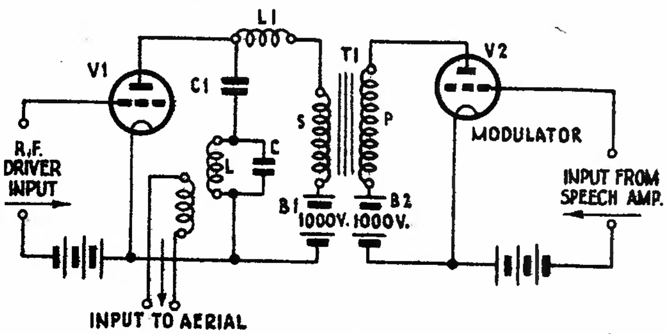


BASIC HEISING ANODE MODULATOR.

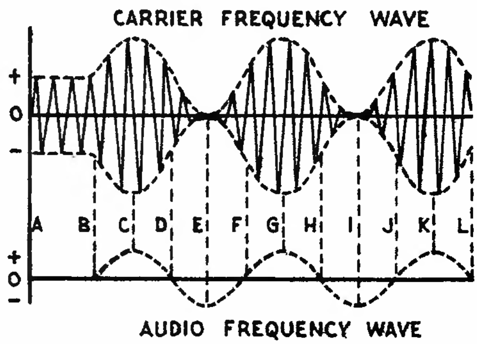
FIG. 12.

Resistor R in series with the anode of V1 is provided to operate this valve at a relatively lower anode potential than that of the modulator valve, so that the percentage of modulation of the carrier frequency is very high. The capacitor shunted across this resistor is necessary to allow the low-frequency variations to be applied to the anode of the radio-frequency amplifier valve. Without this capacitor, the low-frequency variations are taken up by the resistor in the form of IR-drop losses and serious distortion results.

Fig. 13 is known as the "modified" Heising and has two advantages over Fig. 12, firstly, the elimination of R and C from the high voltage side of L1 (these must withstand very high voltages), and, secondly, the transformer is designed to match the modulators to their load. In this circuit, the valve V1 amplifies the radio-frequency signal applied to its grid and cathode from the driver stage. C1 is a blocking capacitor to prevent a D.C. bridge across the anode-supply battery B1. LC is the amplifier tank circuit.



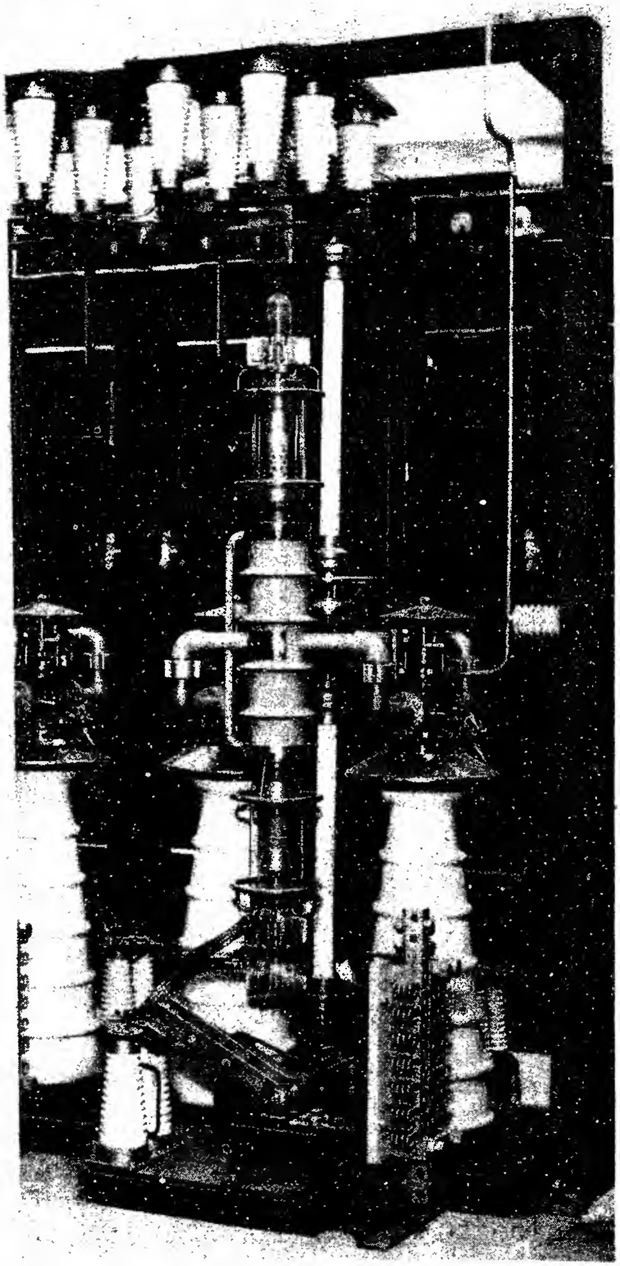
(a) Modified Heising Modulator.



(b) Modulating Effect of Audio-Frequency Wave on Carrier Frequency.

ANODE MODULATED CLASS C AMPLIFIER.

FIG. 13.



MODULATOR VALVE (4030/C) 100 kW H.F. RADIO TRANSMITTER.

The valve V2 is the modulator, which is an audio-frequency amplifier. Speech or music from the preceding speech amplifier is impressed upon the input circuit of this valve. The modulator output is delivered to transformer T1, the secondary of which is connected in series with the anode supply to V1. L1 is a radio-frequency choke provided to prevent radio-frequency currents from entering the modulator and power supply. The modulator is designed to produce across the secondary of T1 a voltage just equal to that of the anode voltage on V1 supplied by battery B1, when the audio-frequency input voltage to V2 just reaches its maximum.

Suppose that the anode-voltage on the valve V1 (Fig. 13a) is 1,000 volts and that, without modulation, the peak radio-frequency current in the aerial circuit is 1 ampere. Then, let an audio-voltage be applied to the input of V2 of such a wave form as shown in the lower wave of Fig. 13b. When the audio wave starts at point B on this curve, there is no change of anode current in V2, hence a voltage is not built up across the secondary (S) of T1. However, as the audio voltage increases to the point C on the curve, the secondary voltage of T1 increases, finally reaching a maximum of 1,000 volts. Assume that for this half of the audio wave the secondary voltage is in series, adding to the battery potential of B1, that is, the upper side of the secondary is positive, thus adding to the voltage of B1. The total voltage on the anode of V1 is now 2,000 volts. With double the anode voltage, the anode current doubles and the peak radio-frequency currents in the aerial circuit increase. Conversely, on the other half of the audio wave (from D to F on the curve of Fig. 13b), a secondary voltage is built up in T1 and is equal to 1,000 volts at the point E, but this voltage is negative and opposes that of B1. The anode voltage on V1 now totals zero, hence the anode-current and radio-frequency output-current peaks are zero.

At the point F on the audio wave, the voltage to V2 is zero, there is no voltage induced in T1 and the valve V1 operates normally with 1,000 volts on the anode. Thus, in passing through the entire audio cycle, the anode voltage on V1 effectively varies from normal, up to twice normal, back to normal, down to zero and finally back to normal, completing the cycle. Accordingly, the carrier-frequency peak currents are caused to vary correspondingly, that is, the amplitude of the carrier wave is made to vary in accordance with the audio wave. This is called amplitude modulation.

In these paragraphs, only 100 per cent. modulation has been considered, in which the peak values of carrier current vary from zero to twice normal. When the audio-input voltage is less than that required to produce this degree of change, then the percentage of modulation is correspondingly less. This condition is necessary, because the changes in the radio-frequency currents must not only represent the different frequencies in speech or music but must also follow all the variations in amplitude of the speech or music. The modulating amplifier together with the modulator must, therefore, have a linear-frequency response and a linear-amplitude characteristic, and it is these features which require careful design of the modulating stage.

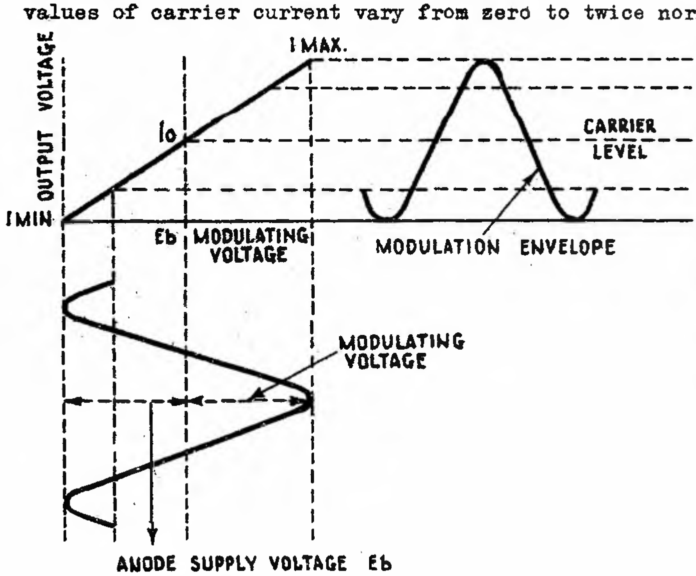


FIG. 14. TYPICAL LINEARITY CURVE.

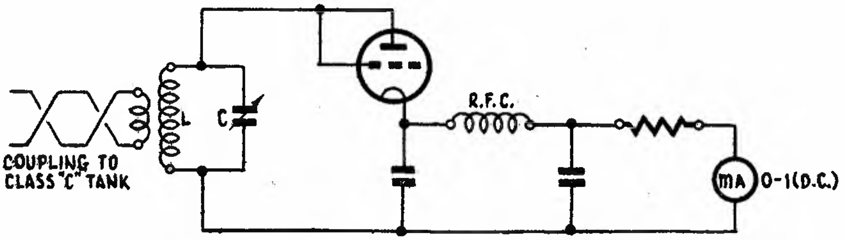
A linear-amplitude characteristic means that a unit change of anode voltage causes a unit change of radio-frequency output current. When a curve is plotted for anode voltage versus output current for the modulating amplifier, with the radio input to this stage held constant, the curve must be practically a straight line. This type of curve is called a "dynamic" characteristic of the valve. In the development of a radio transmitter many such curves are plotted, in order to determine the necessary voltage and output impedance to produce this desired straight characteristic.

A typical linearity curve is shown in Fig. 14.

/ Adjustment

Adjustment for Proper Modulation. When anode modulation is to be undistorted, the radio-frequency amplifier's anode circuit must behave as a pure resistance, that is, the ratio of anode voltage to radio-frequency output current must remain constant for anode voltages from zero to twice the operating value. This condition is only met when the grid is driven to saturation, that is, when sufficient excitation is applied so that any addition fails to result in increased output power. Given grid saturation, the output impedance must be adjusted by varying the coupling to the load (aerial), until the linear relation between anode voltage and current is obtained at the same time that the power input and effective modulating resistance called for by the design are established. Furthermore, this relation is never obtained with the Class C radio-frequency amplifier unless grid bias is secured automatically by means of a grid resistor. Fixed C-battery bias does not give the necessary change at low instantaneous anode voltage, where the peak grid voltage almost equals the minimum anode voltage.

Assuming all these conditions to be fulfilled, the output is tested for linearity by means of a very simple device. This instrument is essentially a linear rectifier or detector of the type shown in Fig. 15.



DIODE TYPE CARRIER SHIFT INDICATOR.

FIG. 15.

When modulation is symmetrical, which it must be for distortionless transmission, the positive peaks of the modulation envelope are the same amplitude as the negative peaks for corresponding modulation cycles (refer back to Fig. 5c). In other words, the envelope amplitude varies equally on both sides of the unmodulated carrier amplitude value. It follows that the average radio-frequency amplitude during modulation is equal to the unmodulated carrier amplitude. This fact provides a very convenient and certain method for checking linearity by means of a carrier shift indicator.

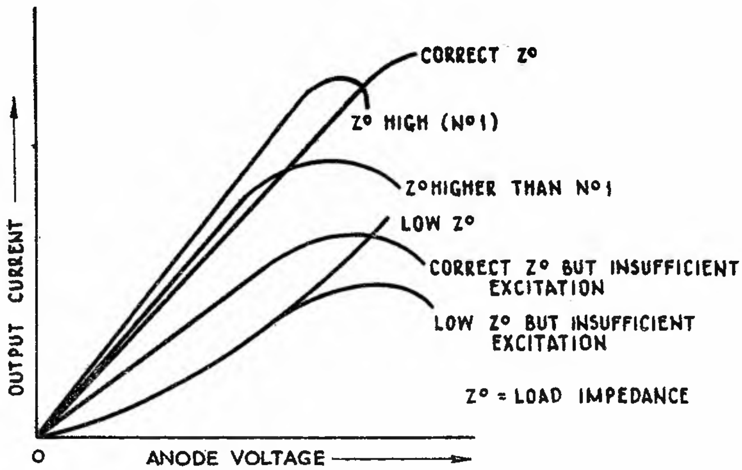
The linear rectifier is coupled to the radio-frequency amplifier under test, and a convenient deflection is obtained on the meter with the carrier unmodulated. When modulation is applied, no change in the rectifier output appears unless the modulation characteristic of the modulating amplifier is non-linear and the modulation capability is exceeded. This change in deflection, when it occurs, is known as carrier-shift. By carrier-shift is meant the shift in the average envelope amplitude during the modulation cycle. Carrier-shift is not to be confused with frequency-shift, which has nothing to do with the processes under discussion. A positive shift indicates that the time-average of the positive half-cycle is greater than that of the negative half-cycle and conversely.

In the interpretation of results, the following points, in conjunction with Fig. 16, are helpful when testing Class C amplifiers.

(i) A negative shift usually indicates insufficient grid excitation or too high an output impedance, or both. When it is known from previous tests that grid saturation exists for the anode voltage in use, then negative shift definitely indicates a load impedance in excess of the proper value for linearity.

(ii) A positive shift usually denotes too low an output impedance or over-modulation, which causes the negative peaks to cross the zero axis for an appreciable portion of the modulation cycle.

/ Fig. 16.



CURVES FOR CLASS C AMPLIFIERS.

FIG. 16.

Adjusting Output Impedance to the Proper Value. In practice, most radio-frequency amplifiers work into a tuned circuit adjusted for resonance at the working frequency. When resonance exists, the load presented to the valve is a pure resistance having a numerical value equal to -

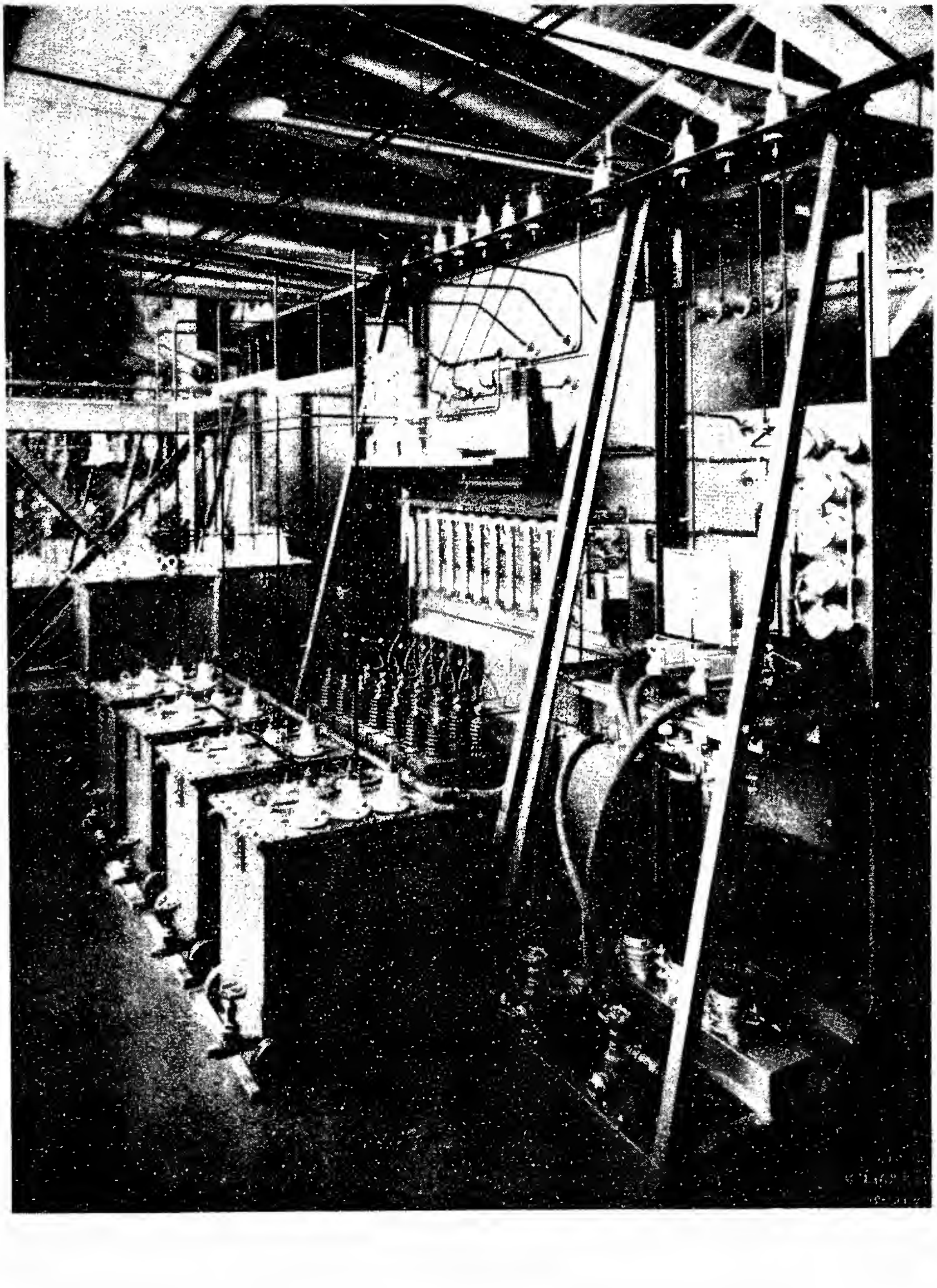
$$Zr = \frac{Xc^2}{R} = \frac{XL^2}{R} = \frac{XLXc}{R} = \frac{L}{RC}.$$

Increasing the reactance, that is, the number of turns, increases the output impedance, and vice versa. Increasing the coupling to the load reduces the effective output impedance, and the converse of this is also true. A fault in some transmitters designed for audio operation is that high-C output circuits are used, in which case, to obtain resonance, the inductances are sufficiently small to make it impossible to adjust the output impedance properly.

Note. The anode load impedance may be too large, especially in high-power transmitters where inadequate tank capacitance is provided. As shown in a previous Paper, there is an optimum L-C ratio, considering excitation requirements, efficiency and harmonic output. This generally suitable optimum ratio requires tank capacitances of approximately 200 μμF at 3.5 Mc/s, 100 μμF at 7 Mc/s and 50 μμF at 14 Mc/s, the tank capacitance being inversely proportional to frequency.

When the foregoing procedure has been rigorously followed, the modulating amplifier should be able to do its job properly. The actual modulation degree can be checked with a cathode-ray oscilloscope, or, approximately, with the usual current-squared galvanometer, provided a linear rectifier is also used with the latter to ensure that no carrier-shift is present. A current-squared instrument, or any other effective current or voltage indicating device, shows both the increase due to modulation and carrier-shift at the same time, and there is no way of separating them. These methods of indicating modulation are not satisfactory on anything but steady tone, because of the inherent inertia of thermo-instruments.

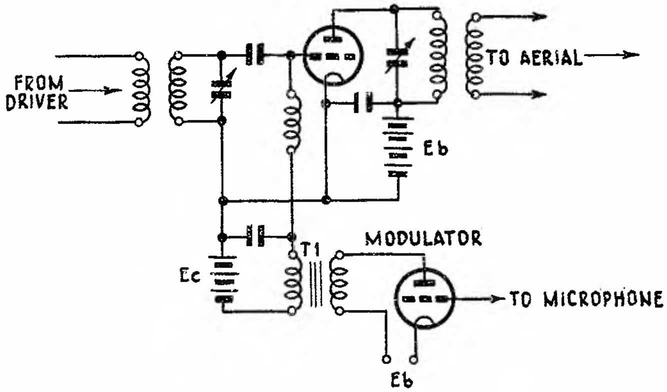
/ Grid



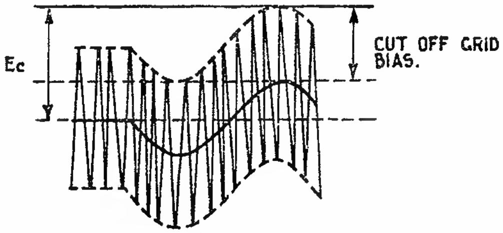
BROADCAST TRANSMITTER (A.W.A.)

Rear entrance showing artificial aerial (left), modulation transformer and choke.

Grid Modulation. In this method of modulation, the output of a Class C amplifier is controlled by varying the grid bias of the valve. A suitable circuit arrangement is shown in Fig. 17a and consists of an ordinary Class C amplifier, in which the effective bias voltage consists of a D.C. component upon which is superimposed the alternating modulating voltage through transformer T1. With proper circuit proportions, the radio-frequency output voltage of such an arrangement is made to vary almost linearly with changes in the effective bias.



(a) Circuit.



(b) Voltage applied to Grid.

(c) Anode Current.

GRID MODULATED CLASS C AMPLIFIER AND CURVES SHOWING DETAILS OF OPERATION.

FIG. 17.

For proper operation of a grid-modulated amplifier, the conditions at the crest of the modulation cycle correspond to ordinary Class C operation. The only other special consideration is that it is desirable to operate with less grid excitation or drive than is customary with Class C amplifier operation.

In designing a grid-modulated amplifier, a valve is chosen so that the normal Class C amplifier rating is at least four times the desired carrier power. This valve is able to handle the crest of the modulation cycle when the power output is four times the carrier power. The valve is then adjusted for conditions at the crest value of the modulation cycle in the same manner as any other Class C amplifier.

When modulation occurs in this stage, that is, when audio-frequency voltages are induced in the secondary of transformer T1 from the modulation, the audio voltage acts with the fixed grid bias to make it larger or smaller, depending upon which half of the cycle is effective at that particular moment.

Thus, the total grid bias is made to vary in accordance with the audio input, and, following the normal valve action of grid voltage upon anode current, the anode current is made to increase when the total negative bias is low and to decrease when the negative bias is high. These changes of anode current produce corresponding changes in the radio-frequency output, which is the effect explained under amplitude modulation. Therefore, the audio frequency causes a change of the bias, a change of the anode current, and resulting variations of radio-frequency output corresponding to the variations in the audio-frequency input voltages - hence the term grid-bias modulation. One of the advantages of this system is that, although the modulated stage may be handling considerable power, very little audio-frequency power is required to modulate it.

The disadvantage of grid modulation is the very low average anode efficiency. This is because the crest alternating voltage between the cathode and anode must have a value less than half the anode supply voltage during unmodulated intervals, when this alternating voltage is still to be less than the anode supply potential at the crest of the modulation cycle. As a consequence, the anode efficiency during modulation conditions is only half as great as for an ordinary Class C amplifier anode modulated.

/ Suppressor

Suppressor-Grid Modulation. The output of a pentode Class C amplifier is controlled by applying to the suppressor-grid a modulating voltage superimposed upon a suitable bias, as shown in Fig. 18.

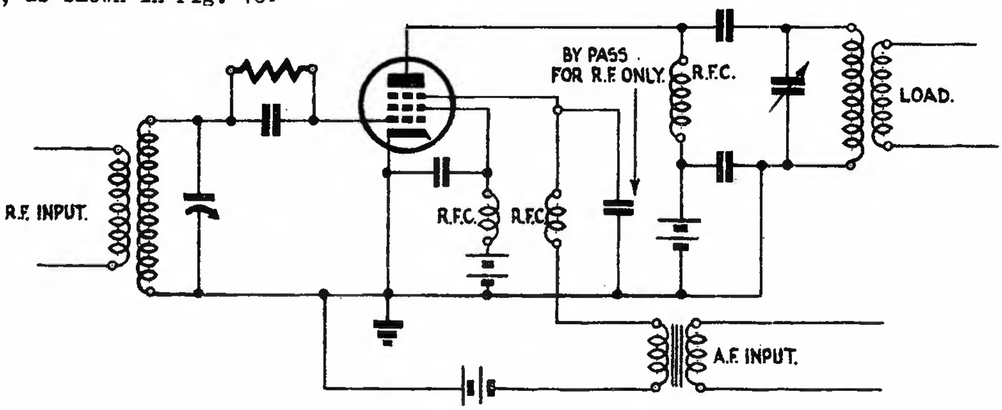


FIG. 18. SUPPRESSOR-GRID MODULATION.

The operation of this arrangement makes use of the fact that, as the suppressor-grid potential is made increasingly negative, a virtual cathode is formed between suppressor and screen. The anode current and hence the output then become less, the more negative the suppressor voltage.

The anode efficiency is approximately half the efficiency obtained in normal Class C operation; the overall efficiency is somewhat less because of the screen-grid losses.

The linearity of modulation is not particularly high. Also, the screen current rises during the negative portion of the modulation cycle, so that care must be taken to avoid exceeding the allowable screen dissipation when suppressor-grid modulation is used.

Screen-Grid Modulation. Modulation is occasionally obtained by applying the modulating voltage to the screen grid of the valve superimposed upon the screen voltage. This is analogous to anode modulation of a triode, and requires considerably less modulating power in proportion to output than with ordinary anode modulation. At the same time, the modulation characteristic is not particularly linear, and the anode efficiency has the same low value obtained with control-grid modulation.

Cathode Modulation. In cathode modulation, the modulating voltage is applied between the cathode and earth of a Class C amplifier, as shown in Fig. 19. This arrangement is essentially a combination of control-grid and anode modulation, with the former predominating. The anode efficiency, modulating power and carrier power obtainable from a given valve are accordingly intermediate between the corresponding cases of pure anode and pure grid modulation, with a tendency towards the latter.

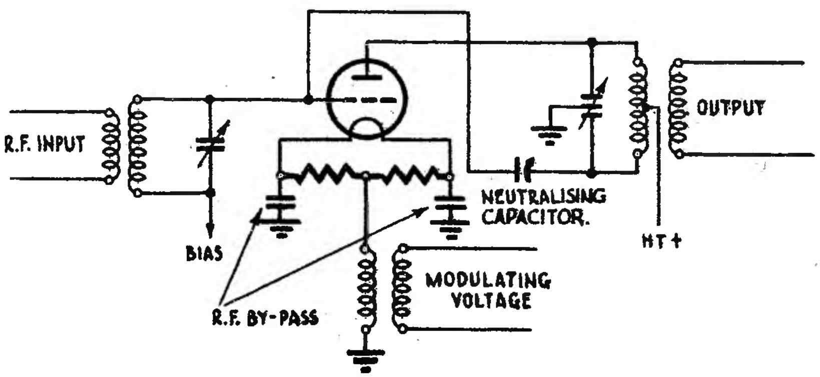
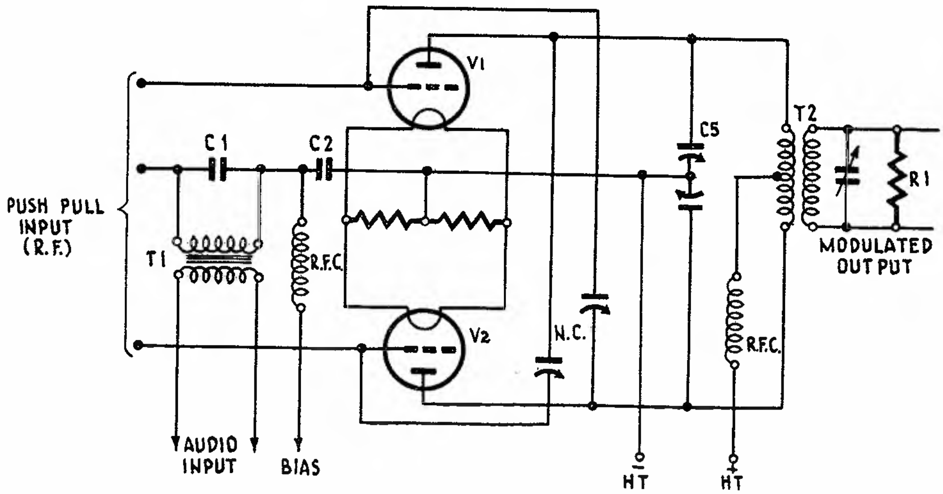


FIG. 19. CATHODE MODULATION.

Push-Pull Grid Modulation. This method of grid modulation is superior to that described previously, and a typical circuit is given in Fig. 20.



PUSH-PULL GRID-MODULATED AMPLIFIER.

FIG. 20.

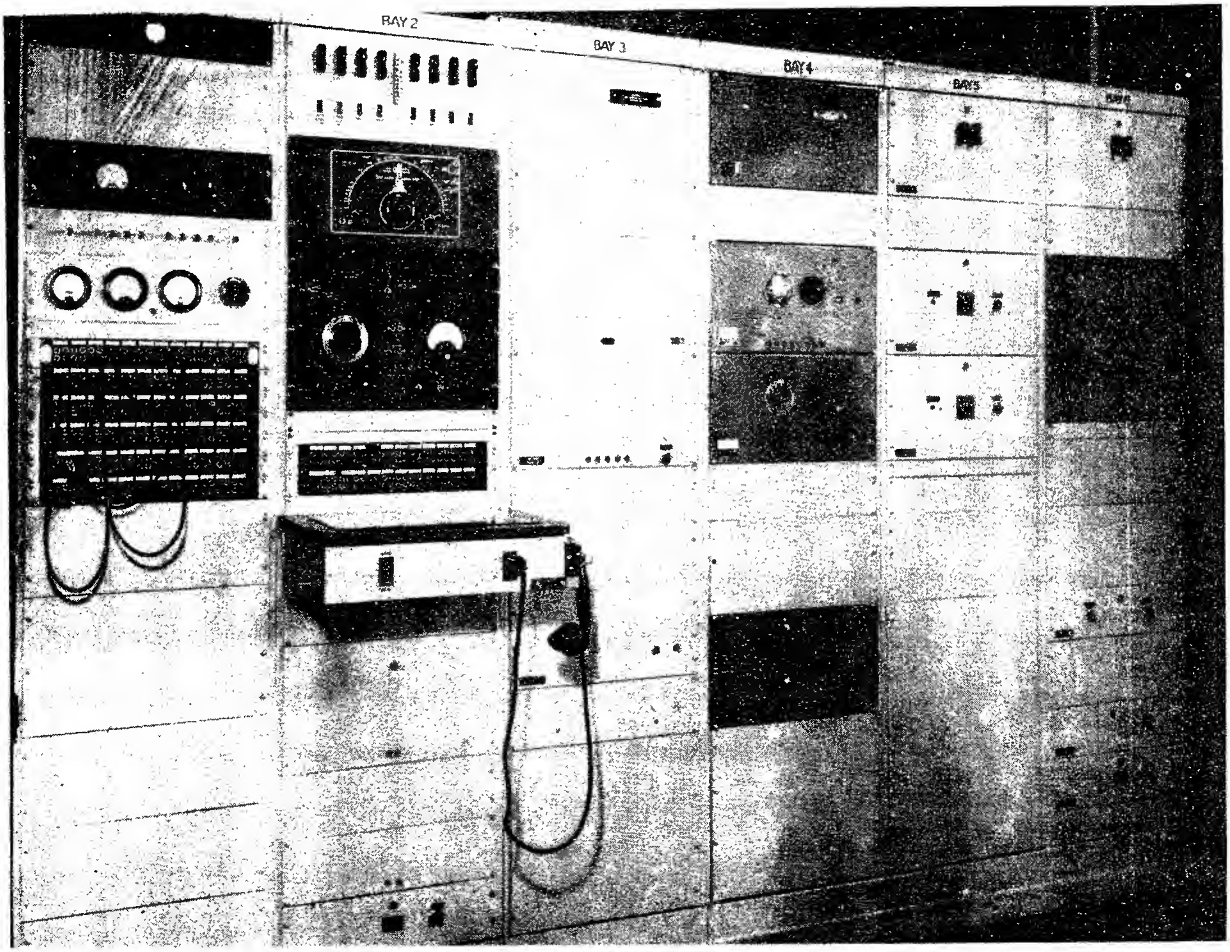
The radio-frequency input voltage is impressed upon the grids of V1 and V2 out of phase, that is, when one grid is negative the other is positive, and the amplified radio-frequency voltage appears across the anodes similarly, but in reverse order. The output of the valves is delivered to the succeeding stage through the transformer T2, which must be tuned to the carrier frequency by the centre-tapped capacitor C5. The secondary of T2 is tuned, and delivers the radio-frequency power to the resistance R1 and the succeeding stages.

The grid bias is adjusted to anode current cut-off or beyond, depending on the required operating conditions.

When audio-frequency voltages are applied via transformer T1, they are impressed upon the grids in phase, both grids being either positive or negative at the same time. Since this in-phase condition is the same as that previously existing as far as the grid bias is concerned, the audio voltages act with the grid bias, adding to it or subtracting from it, according to which half of the cycle is effective at the particular instant. Thus, the total grid bias varies in accordance with the audio input, and, following the normal action of grid voltage upon anode current, the anode current increases during the positive half-cycles. These changes of anode current produce corresponding changes in the radio-frequency output, thus the amplitude of the radio-frequency carrier wave varies in accordance with the applied audio-frequency voltages. One advantage of this system is that, although the modulated stage is handling considerable power, very little audio power is required to modulate it.

Conclusion. Numerous other circuits for producing Amplitude Modulation have been devised, but the foregoing notes show the principles involved.

/ Speech



SPEECH INPUT EQUIPMENT RACKS AT H.F. RADIO STATION, LYNDHURST.

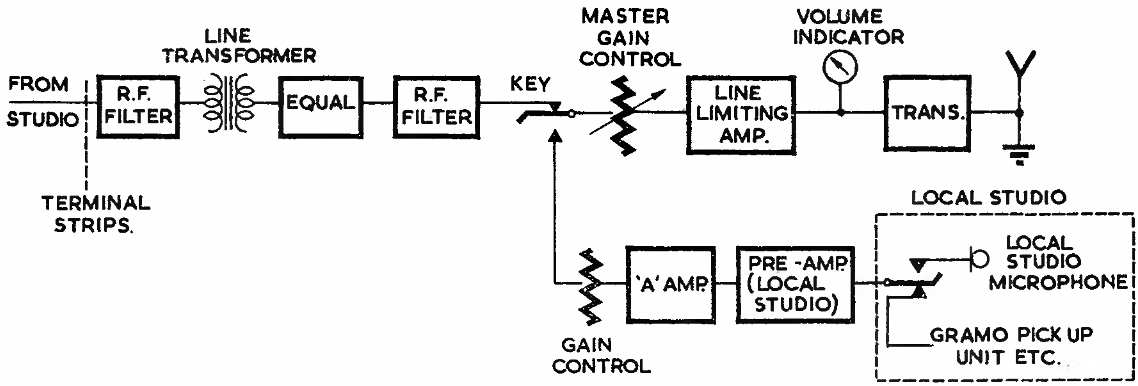
4. SPEECH INPUT EQUIPMENT.

4.1 It is convenient at this juncture to refer to some items of equipment installed in the transmitter building, and known as "Speech Input Equipment". As the name implies, it is a terminating point for the incoming programmes from the studios, Speech input equipment also provides a convenient centre for switching programmes, for conducting modulation and overall response tests, and for connecting an emergency studio to the transmitter in the event of a breakdown of the circuits from the studios. (Transmitters are, in most cases, located some distance from their associated studios.)

4.2 The equipment generally comprises audio-frequency amplifiers, limiting amplifiers, audio-frequency oscillators, local studio amplifiers, gain controls, telephone and monitoring lines, etc., mounted on suitable racks with power supplies and jack fields. Details of amplifiers, oscillators, etc., are given in other Papers dealing with Measurements and Studio and Programme Equipment.

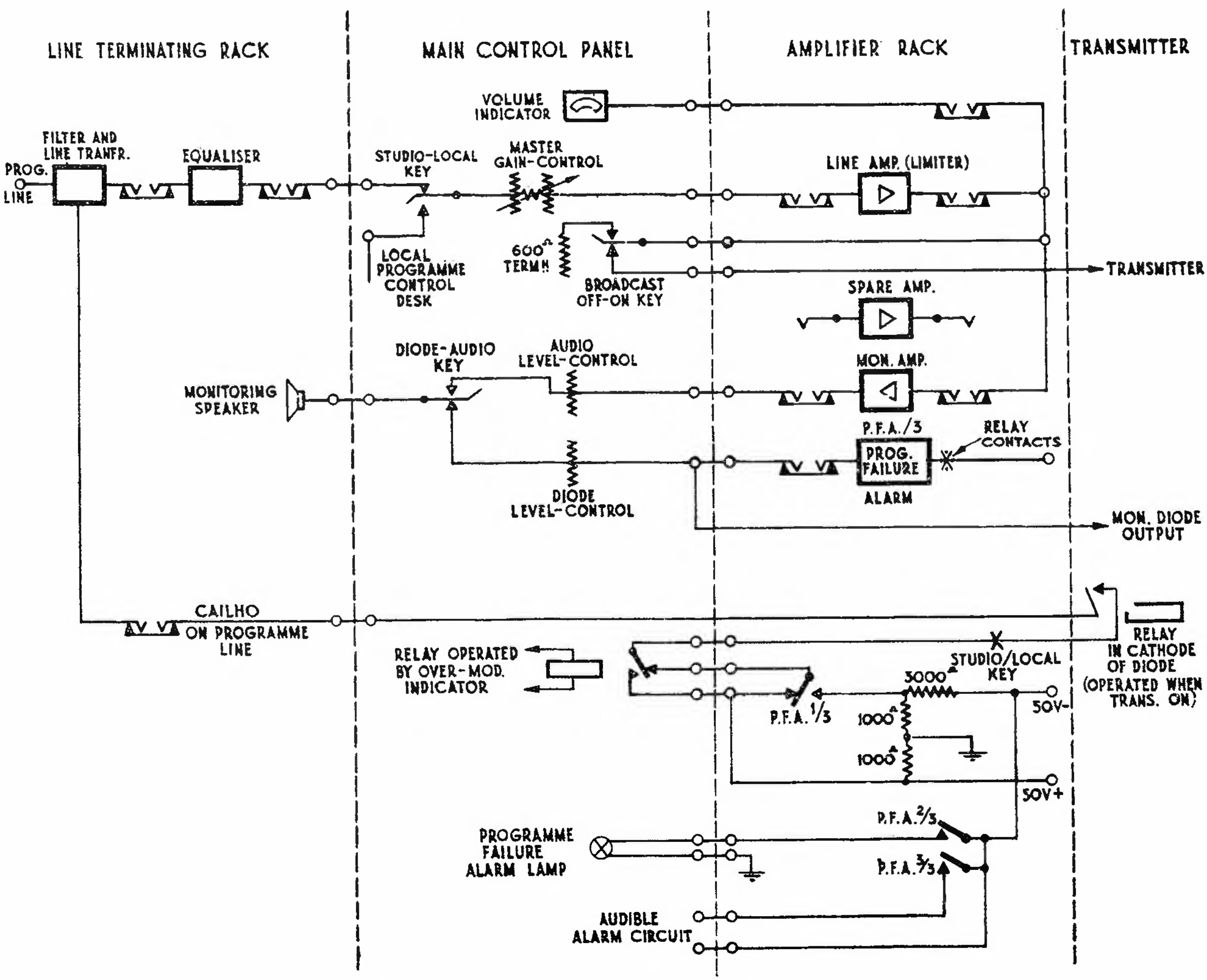
Block schematic circuits of typical speech input equipment are shown in Figs. 21 and 22.

/ Typical



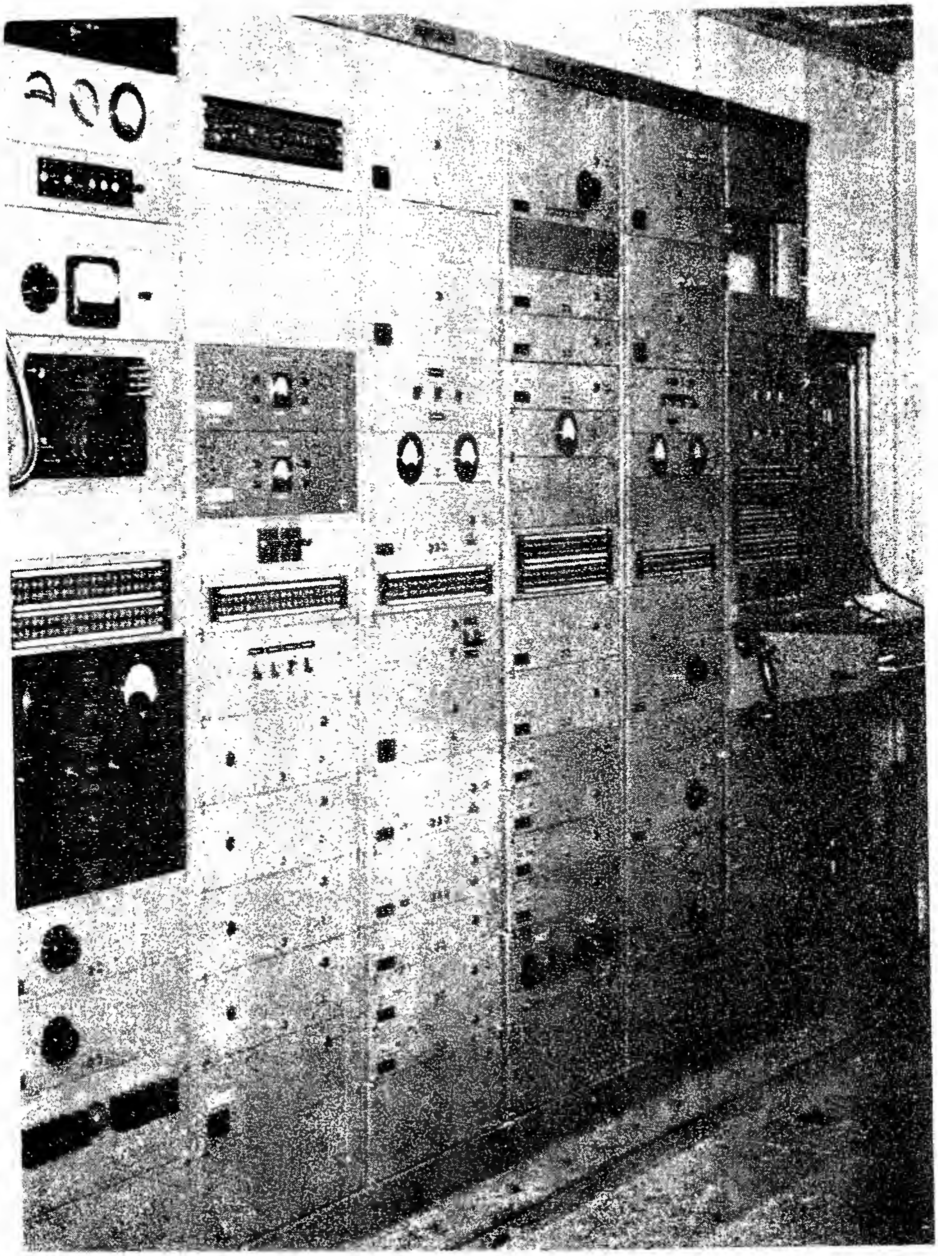
TYPICAL LINE CIRCUIT FROM LINE INPUT TO TRANSMITTER.

FIG. 21.



ANOTHER TYPE OF SPEECH INPUT CIRCUIT INCLUDING PROGRAMME ALARM CIRCUITS BACK TO STUDIO.

FIG. 22.



SPEECH INPUT EQUIPMENT RACKS AT H.F. RADIO STATION, SHEPPARTON.

5. DETECTION OR DEMODULATION.

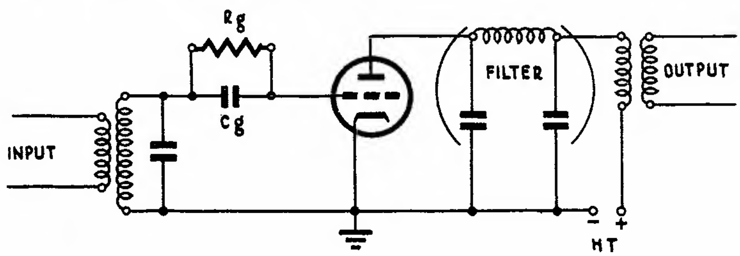
5.1 Detection or demodulation is defined as the process of extracting from the modulated radio-frequency carrier wave the intelligence imparted to it by the process of modulation. In broadcasting, this intelligence is of the form of audio-frequency voltages. In the case of amplitude modulation, detection accordingly means deriving from the modulated wave a voltage that varies in accordance with the modulation envelope. In all practical cases, this is accomplished by rectification of the modulated wave.

Three methods are in general use -

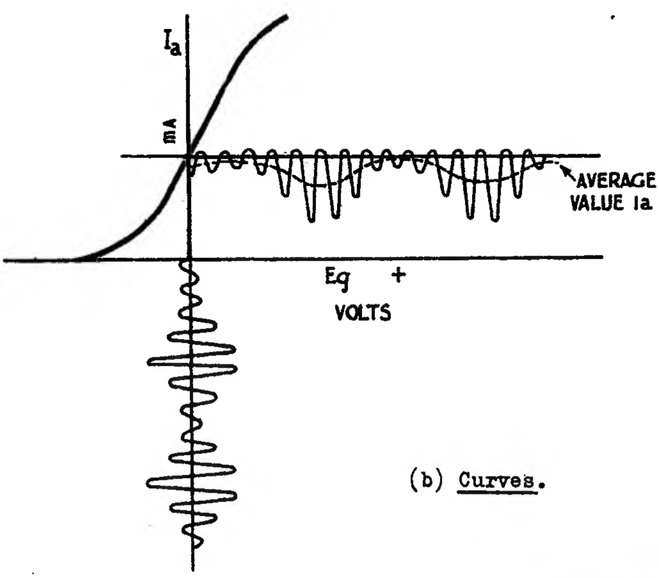
(i) Grid-Current or Grid-Leak Detection,
(ii) Anode-Bend Detection, and
(iii) Diode Detection.

5.2 Grid-Leak Detection. In this method, rectification takes place in the grid circuit by taking advantage of the grid-voltage/grid-current characteristic of the valve. The grid-cathode circuit functions, in effect, as a diode rectifier, and the rectified voltage on the grid is amplified in the anode circuit by ordinary amplifier action.

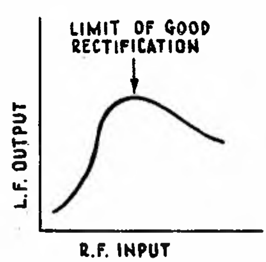
Fig. 23a shows a typical circuit, Fig. 23b the Eg/Ia characteristic of a typical valve and Fig. 23c the limiting action.



(a) Circuit.



(b) Curves.



(c) Limiting Action.

FIG. 23. GRID-LEAK DETECTION.

The grid is normally given a slight positive bias which places it in its most sensitive condition.

/ When

When a modulated wave is applied to the input circuit, the positive peaks of the signal cause the grid to attract electrons. These electrons are confined to the grid, due to the presence of the grid capacitor, and each positive half-cycle contributes its quota of electrons. Thus, a negative charge is gradually built up on the grid which reduces the current in the anode circuit. The grid eventually blocks under these conditions, so a grid-leak, Rg, is included to allow a percentage of the electrons to leak away and prevent the grid blocking.

The amplitude of each radio-frequency peak varies, due to the presence of modulation, so the electron accumulation varies according to the modulation envelope. Thus, the negative charge on the grid varies according to the modulation. Since the grid exerts a control on the anode current, this current varies in accordance with the modulation envelope, and, due to the normal amplifying properties of the valve, the output is an amplified rectified version of the input.

As well as the rectified audio voltage, the grid carries radio-frequency voltages. These radio-frequency voltages are greater than the audio voltages, and, therefore, cause the valve to be overloaded long before the full audio output (that is, full amplification) has been obtained.

Thus, the input which this type of detector handles is limited and is, in fact, about 40 per cent. of the valve's normal capacity.

With a relatively large input, the working point of the valve moves down towards the lower bend, and anode-bend rectification begins to occur. This acts in the opposite phase to grid detection, and the anode current is due to the additive effects which introduce serious distortion.

Grid rectification causes a decrease of anode current, and anode rectification causes an increase of anode current. The resultant limiting action is shown in Fig. 23c.

Time Constant. Regarding the grid-leak-capacitor combination, it is important to choose values so that their time constant (CR) is long compared with one cycle of the carrier frequency, but short compared with one cycle of audio frequency. The capacitor should not discharge appreciably between the positive peaks of the radio-frequency wave, but should discharge sufficiently between the audio-frequency peaks to preserve the modulation frequency. Cg varies between 0.0001 and 0.00025 µF, and Rg varies from 1 to 5 megohms.

Advantages of Grid-Leak Detection.

Good sensitivity to weak signals.
Amplifies the rectified signal.

Disadvantages.

Draws current from input, and therefore, lowers the selectivity.
Cannot handle large signals without distortion.
Cannot supply automatic volume control voltages.

5.3 Anode-Bend Detection. This title is derived from the fact that the valve is operated on the lower end of its anode characteristic, that is, near the lower bend. The upper bend could also be used, but the lower bend is preferable because -

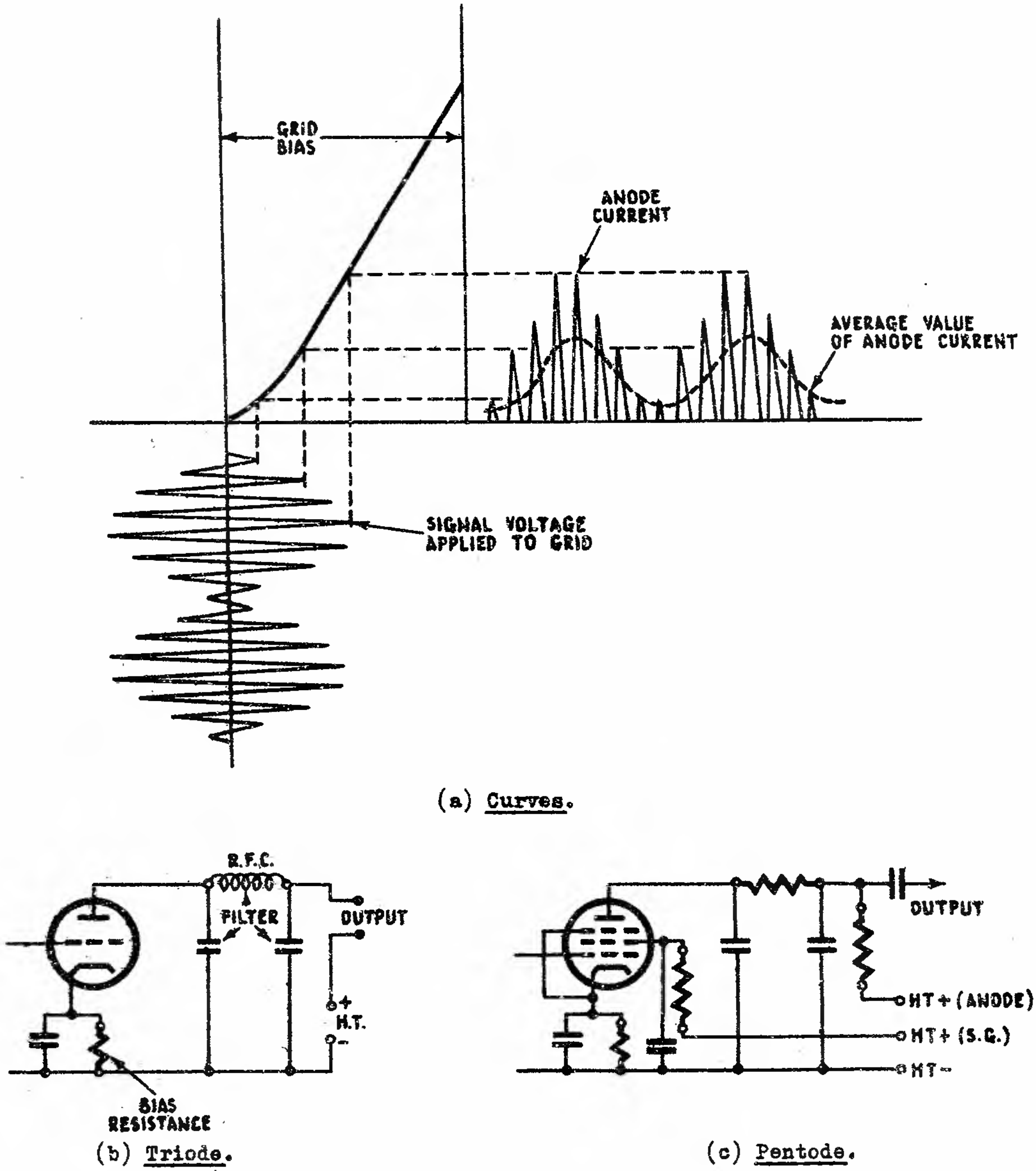
(i) No grid current flows, therefore, there is no distortion or decrease of selectivity of the previous circuit by damping.
(ii) The current drain of the anode circuit is less.

In operation, an ordinary amplifier valve is biased approximately to cut-off.

A radio-frequency signal voltage applied to the grid of such a valve gives pulses of anode current on the positive half-cycles and no current on the negative half-cycles, as shown in Fig. 24a. The resultant average anode current is then dependent upon the average amplitude of the applied signal, thus giving demodulation or detection. This current develops an output voltage by being passed through an ordinary audio-coupling system, as shown in Fig. 24b, showing the circuit for a triode, and in Fig. 24c for a pentode.

/ The

The audio-frequency coupling network is similar in design to that used in audio-frequency amplifier coupling, except that it is desirable to insert a radio-frequency filter between the anode in the output, as shown in Figs. 24b and 24c.



ANODE-BEND DETECTION.

FIG. 24.

The bias voltage is obtained from a fixed source, or it is self-bias. In the latter case, the bias resistor is such that the valve operates at cut-off when the rated carrier voltage is applied. Resistance coupling is normally used with pentodes, while transformer coupling is customary with triodes. Resistance coupling is not used with triodes, since it introduces an A.C./D.C. impedance ratio problem not present with transformer coupling or with a pentode.

/ Advantages

Advantages of Anode-Bend Detection.

Can handle large inputs as compared with grid-leak detection.
Has infinite input impedance as compared with a diode.
Gives more output voltage in proportion to applied signal than a diode.

Disadvantages.

About one-third as much output voltage as valve acting as Class A amplifier.
Compared to diode -

(i) Has greater distortion.

(ii) More critical as to operating conditions for satisfactory operation.

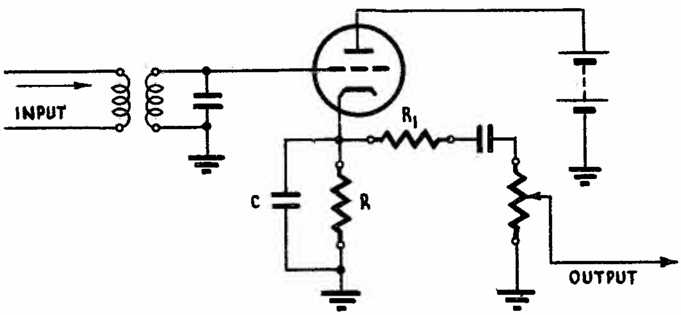
(iii) Can develop only a limited output voltage without excessive distortion.

(iv) Cannot furnish automatic volume control voltages readily.

Linear Detection. With relatively large inputs, the anode bend detector operates linearly, that is, the output varies approximately in a linear manner with the input.

Square Law. In the case of weak signals, however, the valve operates over a curved portion of its characteristic, and the output current tends to vary as the "square" of the input oscillatory voltage. This condition is known as "square-law detection", and is sometimes made use of for particular requirements, such as single-sideband transmissions.

5.4 Infinite Impedance Detector. A triode valve, self-biased nearly to anode-current cut-off by a large cathode resistance by-passed for radio frequency, comprises what is termed an "Infinite Impedance Detector." The circuit is shown in Fig. 25.



INFINITE IMPEDANCE DETECTOR.

FIG. 25.

When a signal voltage is applied, the positive peaks of grid voltage cause pulses of anode current to flow, which builds up a voltage across the impedance R that is only slightly less than the envelope amplitude. These pulses act in the RC circuit of Fig. 25 in much the same manner as in a diode, with the exception that the energy is obtained from the anode supply circuit. The grid does not draw current, and therefore, does not load the preceding circuit, in effect, presenting an infinite input impedance to it.

The peak modulation-frequency voltage obtainable across the output without overloading is quite large, approaching half the anode supply voltage. The infinite input impedance detector is subject to negative peak clipping when the degree of modulation exceeds the A.C./D.C. impedance ratio of the load. For proper operation, this detector is adjusted so that it operates reasonably close to cut-off when no signal is applied. It is also necessary that a reasonably large signal voltage be applied when distortion is to be minimised. The operation, however, is linear up to an output level limited only by the anode supply voltage.

The bias increases with carrier signal voltage, and follows the modulation up to the limit when the degree of modulation is -

/ M

$$M = \frac{Io + Is}{Is\sqrt{(\omega CR)^2 + 1}}$$

Where Io = Anode current for zero signal,
Is = Increase in anode current with signal,
M = Modulation index,
$\omega = 2\pi f$,
f = Modulation frequency, and
R and C are as in Fig. 25.

By proper choice of R and C, peak clipping is avoided.

A disadvantage of this circuit is its inability to supply voltage for conventional A.G.C. circuits.

5.5 Diode Detection. This circuit is, in effect, a half-way rectifier operating under special conditions, but, whereas an ordinary half-wave rectifier supplies D.C. in accordance with its A.C. input, the diode reproduces the audio-frequency amplitude variations of its radio-frequency input. (Fig. 26a shows the basic circuit.)

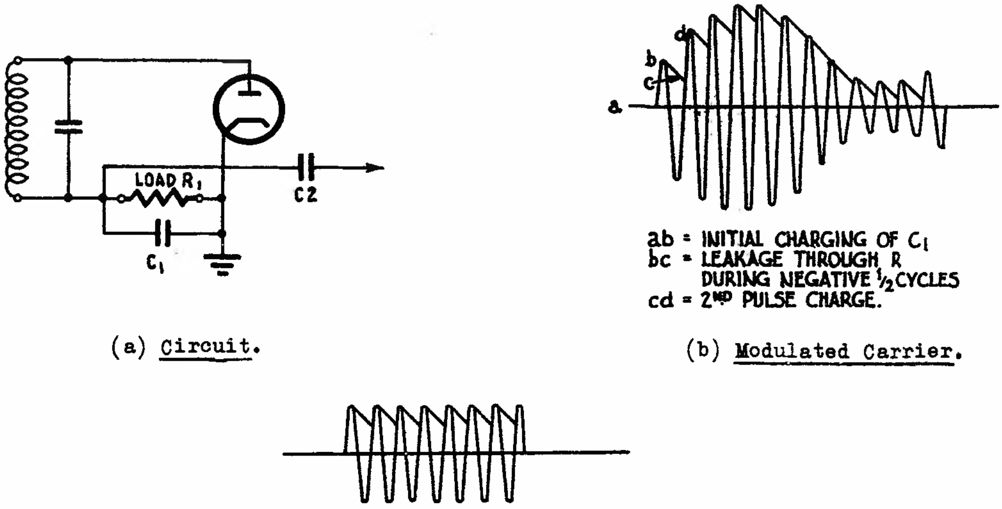
As a result of this latter requirement, the first filter capacitance (C1 in circuit) must be chosen with two liminations on its capacitance.

(i) A lower limit to ensure that the output does not contain an excessive percentage of the cyclic variations of the radio-frequency input.

(ii) An upper limit set by the necessity of avoiding attenuation of the higher modulation frequencies.

The circuit operates as follows -

Due to the unidirectional conduction property of a diode, current only flows when the anode is at a positive potential, that is, during the positive half-cycles of the applied signal. This current charges the diode capacitor C1 to the peak value of the radio-frequency voltage. After reaching its peak, the radio-frequency voltage begins to fall away and, when the capacitor is not present, the diode output falls with it. However, the capacitor is discharging relatively slowly through the diode load resistor, R1, and builds up a potential across it equal to the peak radio-frequency voltage. Each positive peak thus charges the capacitor to its peak value, and, since the peak values vary in accordance with the modulation envelope, the voltage across the resistor R1 varies likewise, reproducing the original modulation voltages. When modulation is absent, the constant peaks of radio-frequency pulses (positive half-cycles) simply result in a steady voltage being developed across the load. Fig. 26b shows the above points.



(a) Circuit. (b) Modulated Carrier.

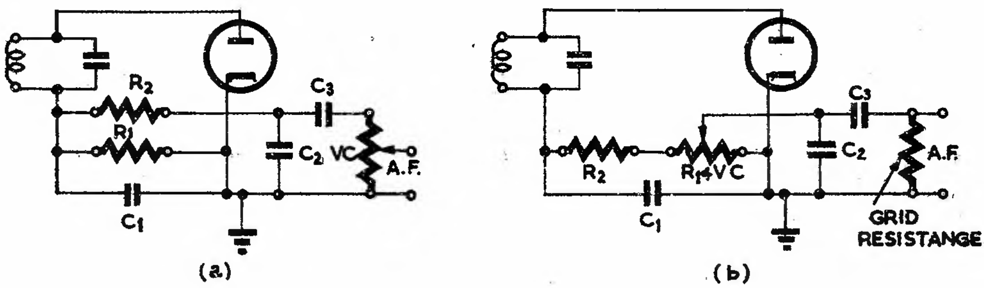
(c) Unmodulated Carrier.

FIG. 26. DIODE DETECTION.

Although the resultant voltage across the capacitor appears ragged, in practice there are a great number of radio-frequency cycles for each audio cycle and the curve is smoother. For example, a 1 Mc/s signal with 1 kc/s modulation would have 1,000 cycles of radio frequency for each cycle of modulation frequency.

The ragged form of the rectified current in Fig. 26b constitutes a cyclic ripple, and its amplitude is dependent on the time constant of the capacitor and leak resistor, being lowest when large values of capacitance and resistance are used. This condition also results in higher D.C. output voltage.

Load Circuit. Fig. 27a shows a diode detection circuit which includes, as well as the load R1C1, a filter circuit R2C2, a coupling capacitor and a volume control which is alternatively a grid leak of fixed value. From a D.C. point of view, the resistor volume control is isolated from the diode load R1 by means of the capacitor C3, but, because this capacitor is intended for audio-coupling the volume control is effectively in parallel with the diode at modulation frequencies. As referred to previously, there is a D.C. voltage, varying at modulation frequency, developed across the diode load. This variation of the diode load voltage is regarded an an audio-frequency voltage, and it is this audio-frequency voltage which is fed to the audio amplifier input across which volume control is shunted through C3.



DIODE DETECTION CIRCUITS.

FIG. 27.

It has already been pointed out that volume control is effectively in parallel with the diode load at audio frequency. This means that the audio control impedance of the load circuit is less than its D.C. resistance, because the value of volume control and the diode load resistor in parallel is always less than that of the diode load resistor alone. The effect is that the capacitor C1 is discharging into one value of load for D.C. (that is, the carrier) and into a lower value for A.C. (modulation). This results in wave form distortion, and, consequently, it is essential that the value of volume control be kept as high as possible so that its shunting effect on the diode resistor is kept at a minimum.

Fig. 27b was developed to minimise these effects, and the differences are -

Filter resistor R2 and load resistor R1 in series with volume control.

/ Results

Results.

(i) Slight loss of output voltage, but not serious.
(ii) Distortion reduced because audio-input circuit is never shunted across the full diode lead resistor, that is, better balancing of automatic control and direct control loading is obtained.
(iii) Audio load is only shunted across that portion of the diode load which is actually in use.
(iv) Value of filter capacitor C2 may be reduced, since that portion of the volume control not in use, plus the filter resistor R2, constitutes the total filter resistance.

Advantages of Diode Detector Circuits.

Less distortion because dynamic characteristic is made more linear than that of other detectors.
Ability to supply automatic volume control voltage.

Disadvantages.

Draws current from input circuit which reduces selectivity of input tuned circuit.
Requires about 10 volts peak for satisfactory operation.
Gives no amplification (except perhaps in compound valves.)

Detection Efficiency is usually expressed as the ratio -

$$\frac{\text{Output voltage developed across load resistor R1}}{\text{Voltage represented by the envelope of modulated wave.}}$$

5.6 Summary of Three Types of Detectors.

GRID-LEAK DETECTOR.

Advantages.

Very sensitive for small signal voltages.
Amplification, approximating that when valve used as audio-frequency amplifier, is obtained.
Distortion is not high unless carrier voltage exceeds about 1 volt.

Disadvantages.

Damping of input circuit due to grid-current and by a resistance component reflected from the anode circuit by the anode-grid capacitance.
Maximum carrier voltage limited, due to the bottom bend of EgIa characteristic.

ANODE-BEND DETECTOR.

Advantages.

There is no damping of the input circuit due to grid current.
Anode-grid capacitance damping is generally low, owing to the valve operating on the low gain bottom bend of the EgIa characteristic.
Amplification is obtained from input to output.

Disadvantages.

Sensitivity is low for small carrier voltages.
Distortion of audio-frequency modulation is high, except for large carrier input voltages with low modulation percentages.
Maximum carrier voltage is limited by grid current.

DIODE DETECTOR.

Advantages.

Distortion of audio-frequency components of modulation decreases as the carrier voltage is increased.
The carrier modulation percentage is high without introducing distortion, provided certain coupling conditions are fulfilled.
Negligible damping of input circuit, except that due to conduction current.
Maximum permissible carrier voltage is almost unlimited.

Disadvantages.

The value represents a loss of amplification. Suitable choice of R1C1 (Fig. 27) helps to lessen this loss.
Conduction current produces damping of input circuit. The use of a high value of R1 reduces this effect.

6. TEST QUESTIONS.

1. Explain briefly the theory of amplitude modulation. When an R.F. carrier of 1,280 kc/s is modulated by a pure tone of 3,600 c/s, what sideband frequencies result?

2. When the unmodulated power of a transmitter is 5,000 watts, what is the peak power radiated at 100 per cent. modulation by a pure sinusoidal tone, and what is the power contained in -

   (i) the carrier, and
   (ii) the upper sideband,

   under this condition?

3. Briefly outline the theory of anode modulation and give a typical circuit.

4. Sketch a circuit for obtaining amplitude modulation and outline the operation of the circuit selected.

5. Describe the anode-bend method of detection, and compare it briefly with one other method with which you are familiar.

6. Illustrate and explain briefly the principles of a diode detector.

7. What is grid-leak detection? Illustrate your answer.

8. Give a brief definition of the following terms -

   Modulation.
   Detection.
   Sidebands.
   Amplitude Modulation.
   Linear Detection.
   Square-Law Detection.
   Types of Radio Emission.
   Carrier Shift.
   Percentage Modulation.

END OF PAPER.

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Melbourne, C.2.

COURSE OF TECHNICAL INSTRUCTION.

RADIO MEASUREMENTS.

CONTENTS.

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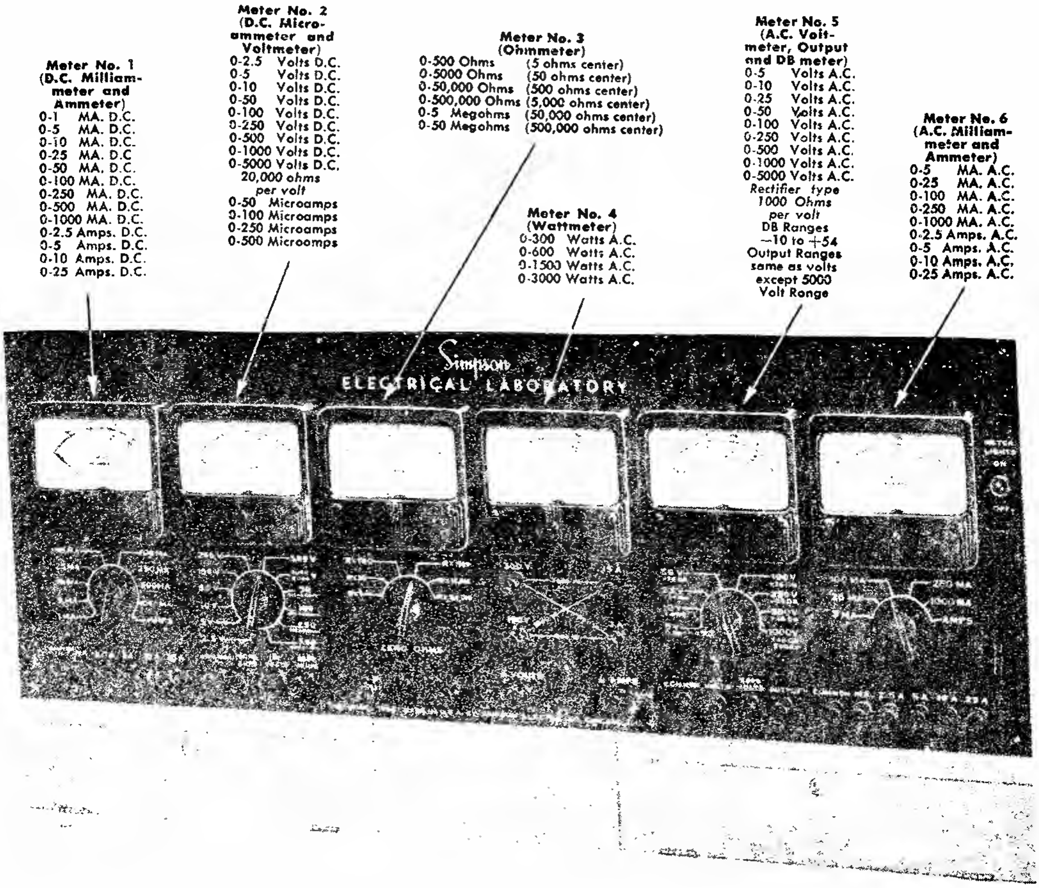
1. INTRODUCTION.

1.1 A brief outline of measurements and tests used for radio equipment generally is given in this Paper. In a branch of science covering such a wide field as Radio, there are many tests and measurements associated with design and laboratory development. These, however, are not discussed, since, in a large number of cases, the instruments used are not readily available for general use. Many of the principles of measuring instruments are covered in other books of this Course, but, in order to keep a continuity of treatment, these principles are briefly revised in this Paper.

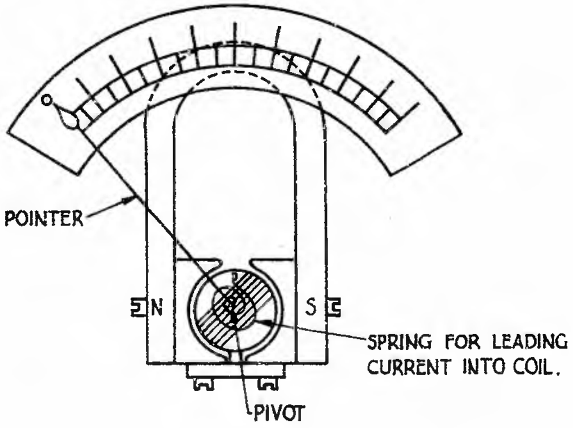


THE FUNCTIONS OF OVER 60 SEPARATE INSTRUMENTS ARE COMBINED IN THIS MODERN MEASURING UNIT!

This complete test unit is adaptable to the testing of electrical appliances, such as small motors, circuits and radio sets. This unit consists of six individual instruments, indirectly illuminated, each with a complete set of ranges. In addition to the wide variety of A.C. and D.C. voltage and current ranges, a multi-range ohmmeter and a single phase wattmeter have been incorporated. Also, to meet the need for extreme sensitivity required in testing circuits where only a small amount of current is available, an instrument is provided with a sensitivity of 50 microamperes, providing 20,000 ohms per volt on all D.C. voltage ranges. The unit incorporates a rectifier type instrument for measuring A.C. voltage with a resistance of 1,000 ohms per volt on all ranges.

2. CURRENT, VOLTAGE AND POWER MEASUREMENTS.

2.1 Measurement of Direct Voltage and Current. Direct voltages and currents of the magnitudes usually encountered in radio-communication work are measured with instruments of the D'Arsonval (moving coil) type. Such instruments are rugged, stable and consume relatively little power, and are also commercially available in a wide variety of ranges, types and degrees of accuracy. Fig. 1 shows this type of meter.

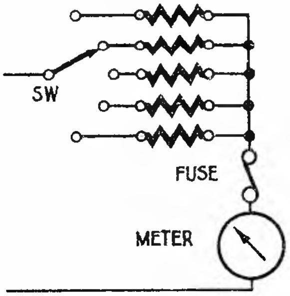


MOVING COIL METER.

FIG. 1.

Voltmeters, as described in other books, are essentially current instruments provided with a series resistance. The power consumed by a voltmeter depends upon the current sensitivity of the instrument, and is commonly expressed in "ohms per volt". Thus, when a meter is rated at "10,000 ohms per volt", it absorbs $\frac{E^2}{R}$ watts from the circuit, that is, $\frac{1 \times 1}{10,000}$ = 0.1 mW when measuring one volt or, $\frac{200 \times 200}{200 \times 10,000} = \frac{1}{50}$W = 20 mW for a 200 volt reading.

Multi-Range Voltmeters. Multipliers are used to extend the range of a voltmeter. A typical circuit for a voltmeter multiplier is shown in Fig. 2. The resistance values must be chosen so that, at each position of the switch, the total resistance in the circuit, including fuse and meter, extends the range of the meter as required. For high precision instruments, the multiplier resistances are made of wire having zero temperature co-efficient, such as manganin.



EXTENDING THE RANGE OF A VOLTMETER.

FIG. 2.

The value of a resistance for use as a multiplier is determined from the formula -

$$R = Rm \left(\frac{E}{Em} - 1\right) \text{ ohms}$$

where R = Multiplier resistance in series with Vm,

Rm = Resistance of meter,

E = Full scale deflection with multiplier, and

Em = Full scale deflection without multiplier.

When a milliammeter movement is used, Em is ordinarily small compared with E. The above expression then reduces to -

$$R = E\left(\frac{Rm}{Em}\right) \text{ ohms,}$$

which is satisfactory for general use. The factor $\frac{Rm}{Em}$ is a constant of the meter in ohms per volt, as referred to previously.

/ Multi-

Multi-Range Current Meters. To provide a current measuring instrument with a number of ranges, care must be taken to arrange the circuit so that contact and fuse resistances ar not included in the shunts.

Fig. 3a shows a typical circuit arrangement using individual shunts to provide a current measuring meter with a number of ranges.

The resistance value of the shunts are calculated from the formula -

$$Rs = \frac{Rm\ Im}{I - Im} \text{ ohms}$$

where Rm = Resistance of meter,

Rs = Resistance of shunt,

Im = Full scale current of meter without shunt, and

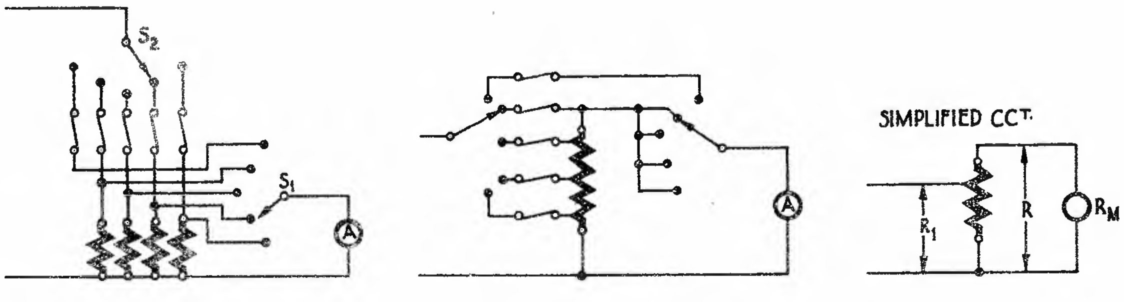
I = Full scale current of meter with shunt.

The Multiplication Factor of shunts is -

$$K = \frac{Rs + Rm}{Rs} \text{ times}$$

$$\text{where } K = \text{ratio of } \frac{\text{full scale current with shunt}}{\text{full scale current without shunt.}}$$

Universal Shunt. Another method of shunting ammeters is by the use of a "universal shunt", and is shown in Fig. 3b. It is seen from the simplified circuit in Fig. 3b that the relative multiplying ratio is proportional to R/R1, and is independent of both the meter resistance and also the total shunt resistance R. Hence, when the resistance R is tapped at points which make R/R1 = 1, 2, 5 and 10, then the relative multiplying factors at these points are 1, 2, 5 and 10 respectively and are the same for any meter, whatever its resistance.



(a) By Individual Shunts.　　(b) By Universal Shunt.

FIG. 3. EXTENDING THE RANGE OF AN AMMETER.

2.2 A.C. Measurements. The problem of measuring A.C. in communication work is complicated by the wide frequency range which must be covered and by the resulting calibration difficulties.

Typical of the meters used for A.C. measurements are -

Moving-Iron Meter.
Dynamometer.
Rectifier Meter.
Thermo-couple Meter.
Thermionic Valve Meter.
Electrostatic Voltmeter.

/ Moving

Moving-Iron Meter. The iron-vane types of voltmeter and ammeter are widely used at power frequencies, and are used to a limited extent on the lower audio frequencies. These types are relatively inexpensive and are sufficiently accurate for most applications. The magnetic pull is proportional to the square of the current, so that the meter is insensitive to small currents, and the scale is not uniform. The meters read R.M.S. values. In one type, the pointer is fixed to an iron vane, which is attracted by a coil to an extent which depends on the current through the coil (see Fig. 4). Another type depends on the mutual repulsion between two vanes, one fixed and one carrying the pointer. The two vanes are magnetised by the same coil, and, with A.C., the repulsion is constant and independent of the polarity of the A.C.

Dynamometer. The dynamometer type depends on the reaction produced between a movable and a stationary coil (see Fig. 5). Dynamometer types are more accurate than the moving-iron type, but they are considerably more expensive and require more operating energy. Dynamometer instruments are designed for frequencies up to about 1,000 c/s, but have not a large application. The "Megger" and "Wattmeter" are adaptations of this design.

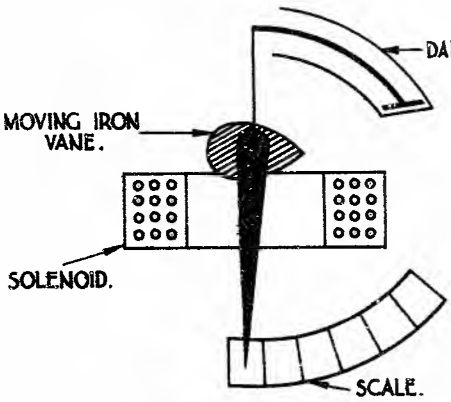


FIG. 4. MOVING-IRON METER.

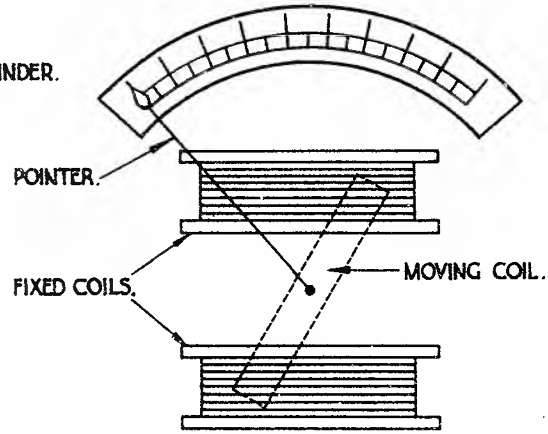


FIG. 5. DYNAMOMETER.

Rectifier Meter. In the rectifier meter, the current to be measured is passed through a full-wave copper oxide rectifier unit, and the resulting D.C. is shown by a moving coil D.C. meter (see Fig. 6). Rectifier meters can be built to give full scale deflection with A.C. of less than 1 mA, and so make possible the construction of A.C. voltmeters having sensitivities of 1,000 ohms per volt and more. The ruggedness and overload capacity compare favourably with moving-coil D.C. meters. Variation of rectifier characteristics with temperature limit the best accuracy to about 5 per cent. The electrostatic capacity of the rectifier element, which partially by-passes the higher audio-frequencies around the rectifier, causes inaccuracy. One disadvantage of this meter is the comparatively high-voltage drop, which is approximately 1/2 to 1 volt for full scale deflection.

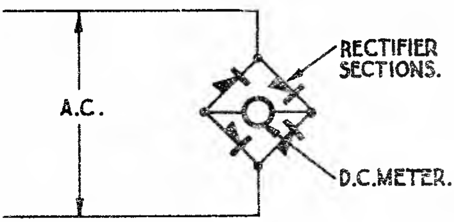


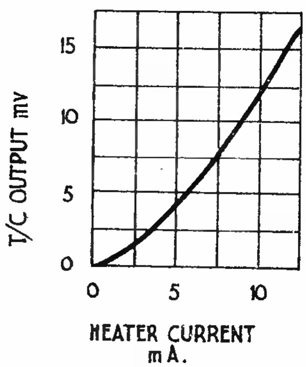
FIG. 6. RECTIFIER METER.

Rectifier meters give an indication which is proportional to the average amplitude of the A.C. wave. These meters are normally calibrated on a sine wave, and the scale indicates the equivalent effective value of this wave. Inaccuracy results when a non-sinusoidal wave form is measured.

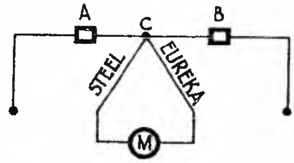
Thermo-couple Meter. This type operates on the principle that a current flows from the junction (called a thermo-couple) of two different metals when the junction is

/ heated.

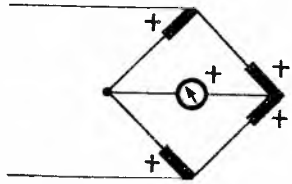
heated. The output of the thermo-couple is recorded by a sensitive D.C. micro-ammeter of the moving-coil type (see Fig. 7). The thermo-couple is the standard method for measuring currents at audio and radio frequencies. This method is accurate and stable, and the calibration is independent of frequency up to extremely high frequencies.



(a) Calibration Curve.



(b) Typical Circuit.



(c) Bridge Circuit.

FIG. 7. THERMO-COUPLE METER.

The meter shows R.M.S. values, and, when D.C. is used for calibration, it is necessary to take the average of results obtained with the same D.C. flowing through the heater in opposite directions.

The main disadvantage is the low overload capacity, since currents exceeding full load rating by more than 50 per cent. may burn out the heater.

Thermionic Valve Meter. These meters are covered in Radio II, in connection with receiver tests and also in Applied Electricity III.

Electrostatic Voltmeter. The electrostatic voltmeter depends for its operation on the electrostatic attraction between two bodies subject to a potential difference. The force of attraction is proportional to the area of the exposed surface and to the square of the potential difference, and is inversely proportional to the distance between, that is -

$$\text{force} \propto \frac{A\ V^2}{d}$$

where A = area of plates,
V = potential between plates, and
d = distance between plates.

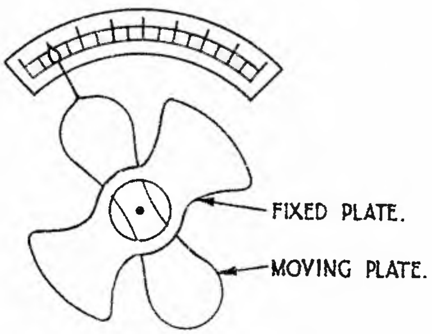


FIG. 8. ELECTROSTATIC VOLTMETER.

A typical design consists of a bank of moving plates, so provided that the attractive force draws them into mesh with a bank of fixed plates. The plates are usually so shaped that, at low values of potential, the surface areas are sufficiently great to compensate for the lesser attractive force existant.

Since the attraction varies with the square of the potential, the meters read R.M.S. values and are suitable for A.C. or D.C.

/ Electrostatic

Electrostatic voltmeters draw very little power from the circuit being measured and are, therefore, suitable for measurements in high resistance circuits, and are designed for use at radio frequencies. The scale is non-linear and contracted at the lower end. This affects the accuracy, which is of the order of approximately 2 per cent. to 5 per cent. This type of instrument is not generally suitable for ranges below 100 volts, but is used up to hundreds of kilovolts. One type of electrostatic voltmeter is shown in Fig. 8.

2.3 Summary of Features of Radio Measuring Meters.

Moving-Coil Meters.

Indicate average value of current.
Reliable, sensitive, flexible and of high precision.
Suitable for D.C. or rectified A.C. only.
Scale, linear.

Moving-Iron Meters.

Indicate R.M.S. values.
Suitable for A.C. or D.C.
Limited frequency range (usually "commercial" range).
Accuracy moderate.

Dynamometers.

Indicate R.M.S. values.
Suitable for D.C. or A.C.
Non-linear scale.
Sensitivity not as good as moving coil.
Frequency range limited to about 100 c/s.

Rectifier Meters.

Suitable for A.C. generally to upper audio frequencies.
Modern design has extended range to several Mc/s.
Moderate accuracy and stability.
Susceptible to wave form errors and frequency errors.

Thermo-couple Meters.

Sensitivity equivalent to that of the associated moving coil meter.
Indicate R.M.S. values.
Suitable for A.C. and D.C.
Suitable up to high radio frequencies.
Scale non-linear; somewhat sluggish in action.
Do not possess much overload factor of safety.

Thermionic Valve Meters.

High input impedance, therefore negligible power consumption from source.
Moderate accuracy.
Very extensive voltage range.
Frequency range to over 100 Mc/s.
Reasonably linear scale.

Electrostatic Voltmeters.

Suitable for medium and high A.C. and D.C. voltages.
Power consumption negligible.
When calibrated on D.C., electrostatic voltmeters indicate R.M.S. value of A.C.
Can be designed for use at radio frequencies.
Somewhat fragile.

/ 2.4

2.4 Measurement of Power.

Direct Current. Power is measured by determining the voltage (E) across the load and the current (I) through the load -

$$\text{Power} = EI$$

Alternatively, a Wattmeter is used. The Wattmeter consists of an instrument based on the principles of Fig. 5. The interaction of the current and voltage coil fields produces the torque required to move the pointer. The scale is calibrated directly in watts.

Alternating Current (Low Frequencies). At commercial power frequencies and audio frequencies up to about 1,000 c/s, a wattmeter, as referred to above, provides for reasonably accurate measurement of power. When the circuit has no reactive components, that is, zero power factor, then the product EI gives the power.

Audio and Radio Frequencies. The usual method of measuring power at radio and audio frequencies is to measure the effective resistance of the circuit and the R.M.S. current flowing. The power is then calculated from the relation $P = I^2R$. This method gives satisfactory results at all frequencies, since circuit resistance can be measured with good accuracy.

3. CIRCUIT CONSTANTS OR PARAMETERS.

3.1 D.C. Resistance. Resistance is measured, as stated in other books of this Course, on a Wheatstone Bridge, by the Voltmeter-Ammeter method or with an Ohmmeter. Other departmental books deal with these more thoroughly, and a brief reference only is given here.

Fig. 9a is the basic Bridge Circuit for measuring resistance values and, assuming R1 = R2, then, when balance condition appears on the galvanometer (zero current), the unknown resistance R4 is equal to the reading of variable resistance R3, which was adjusted until balance was obtained, that is -

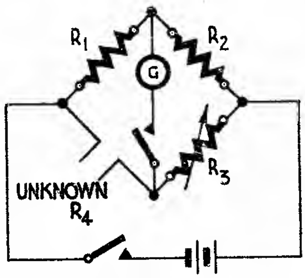
$$R1R3 = R2R4 \qquad \text{or,} \qquad R4 = \frac{R1R3}{R2}$$

Another method of measuring resistance is shown in Fig. 9b. The method (i) is used when the current drawn by the voltmeter is not negligible by comparison with the current via Rx, and method (ii) is used when the voltage drop in the ammeter is not negligible when compared with the applied voltage.

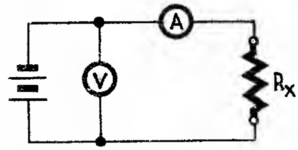
Ohmmeters are also suitable for making approximate resistance measurements, such as in servicing radio receivers and similar equipment. Two typical simplified circuits are shown in Fig. 9c. The arrangement (i) is based on the assumption that the battery generates a constant voltage during its life, but that internal resistance increases. To operate, the terminals XX are short-circuited and R is adjusted for full scale deflection on the meter. The resistance is then placed across XX and the new reading noted on the meter. Ohmmeters are usually calibrated in ohms to facilitate measurements. The arrangement (ii) operates on the theory that, with use, the battery voltage drops but the internal resistance is unchanged. These two assumptions contribute to the inaccuracy factor in the measurements, but the method has a large number of useful applications.

3.2 A.C. Bridges. The most satisfactory method of measuring resistance, capacitance, inductance and mutual inductance of a circuit at audio frequencies is by means of an A.C. bridge. A.C. bridges are derived from the Wheatstone balance bridge shown in Fig. 9a. The A.C. bridge is similar, except that the arms become complex impedances and the source of current is A.C. For general use, a pair of head-phones serves as a detector.

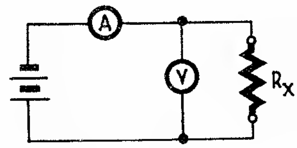
/ Fig. 9.



(a) Wheatstone Bridge.

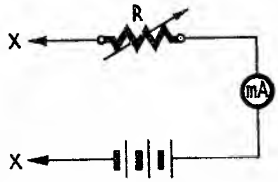


(i)

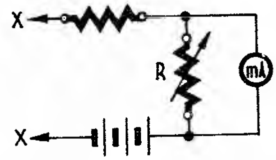


(ii)

(b) Voltmeter/Ammeter Method.



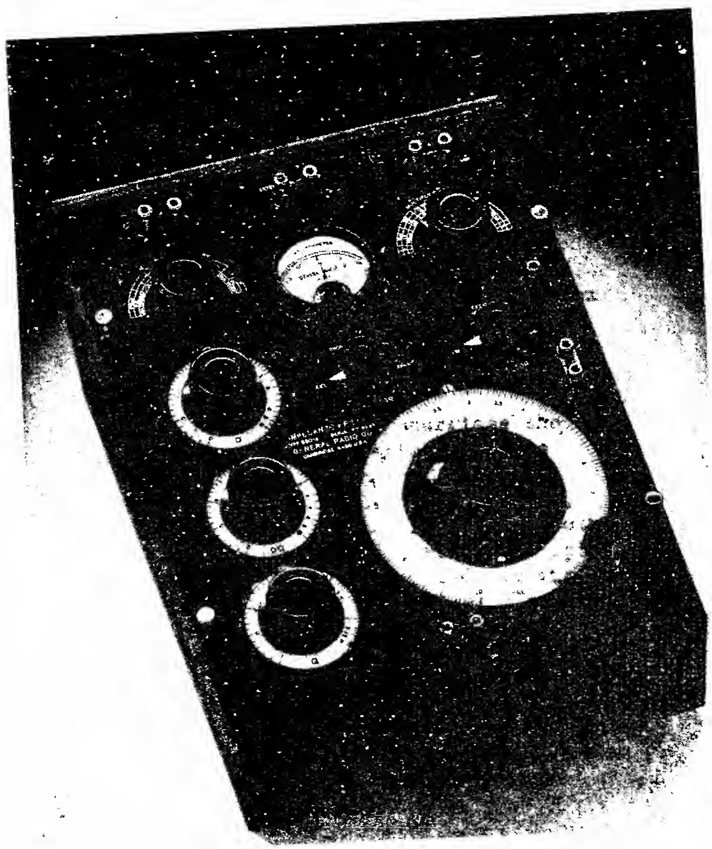
(i)



(ii)

(c) Ohmmeter Method.

FIG. 9. MEASURING RESISTANCE.



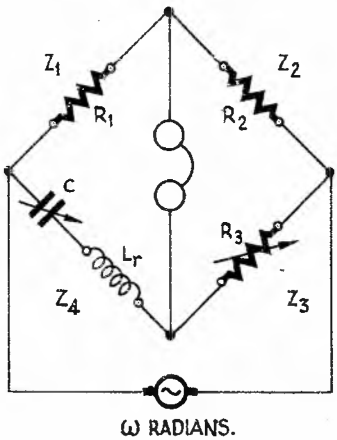
TYPICAL IMPEDANCE BRIDGE.

There are a large number of A.C. bridges in use, each with some particular application, but the following types have been selected as representative -

Resonance Bridge.
Owen's Bridge.
Capacity Bridge.

Resonance Bridge. The Resonance Bridge is used mostly to measure the effective resistance of radio apparatus at audio frequencies. This arrangement is shown in Fig. 10, where the symbols are -

R1 and R2 = Ratio arms (or Z1 and Z2),
R3 = Variable resistor (or Z3),
Lr = Unknown inductor with effective resistance r, and
C = Calibrated variable capacitor.



RESONANCE BRIDGE.

FIG. 10.

From this -

The fourth arm Z4 (which includes C + Lr)

$$= \sqrt{r^2 + (\omega L - \frac{1}{\omega C})^2}$$

When the bridge is balanced -

$$Z1Z3 = Z2Z4$$

$$\text{or } R1R3 = R2\left[\sqrt{r^2 + (\omega L - \frac{1}{\omega C})^2}\right]$$

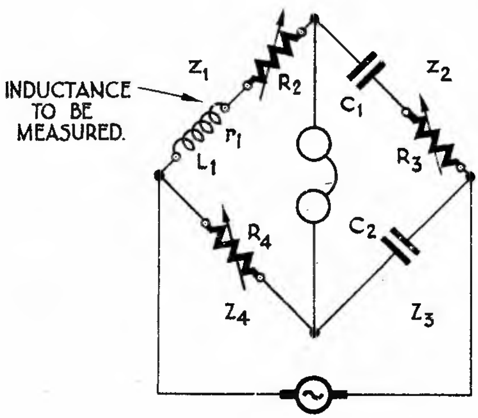
Now, at resonance, $\omega L = \frac{1}{\omega C}$ and, eliminating these then -

$$R1R3 = R2r$$

$$\therefore r \text{ (effective resistance of L)} = \frac{R1R3}{R2}$$

In practice, R1 usually equals R2, thus -

$$r = R3$$



OWEN BRIDGE.

FIG. 11.

Owen Bridge. The Owen Bridge measures inductance in terms of capacitance, but, in this bridge, the arms are so arranged that the balance equations for both resistance and inductance are independent of frequency, provided that the capacitor has low losses (see Fig. 11).

In Fig. 11 -

C1 and C2 = Two standard capacitors (approximately equal values),
R2 = Variable resistor (decade),
R3 = Variable resistor (decade),
R4 = Variable resistor (decade), and
L1r1 = Inductor of effective resistance r1.

/ Then

$$\text{Then } Z1 = r1 + R2 + j\omega L$$

$$Z2 = R3 + \frac{1}{j\omega(C1)}$$

$$Z3 = \frac{1}{j\omega(C2)}$$

$$Z4 = R4$$

When the bridge is balanced by successive adjustments of R2 and R3 -

$$Z1Z3 = Z2Z4$$

$$\text{or } (r1 + R2 + j\omega L)\frac{1}{j\omega(C2)} = \left(R3 + \frac{1}{j\omega(C1)}\right) R4$$

$$\text{or } \frac{r1 + R2}{j\omega(C2)} + \frac{L}{C2} = R4R3 + \frac{R4}{j\omega(C1)}$$

Equating non-reactive terms and -

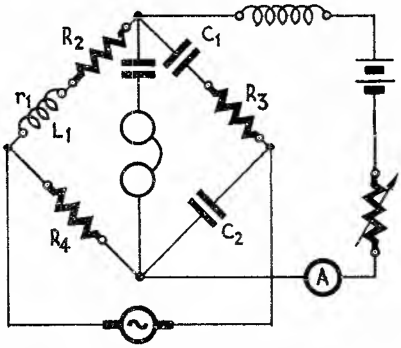
$$\frac{L}{C2} = R4R3$$

$$\text{or } L = R3R4C2$$

Equating reactive terms and -

$$\frac{r1 + R2}{j\omega(C2)} = \frac{R4}{j\omega(C1)}$$

$$\text{or } r1 = \frac{R4C2}{C1} - R2$$



OWEN BRIDGE.
(For choke with D.C. flowing.)

FIG. 12.

The Owen Bridge is also used for measuring inductance and effective resistance of a coil carrying D.C. (see Fig. 12). The equations are the same as before, as is the method of balancing.

Capacitance Bridge. Fig. 13 shows a simple Capacitance Bridge for determining the capacitance value of capacitors other than electrolytic capacitors. The arms consist of -

C = Z1 = Unknown capacitor,
Q = Z2 = Fixed known resistor,
S = Z3 = Variable known resistor, and
Cs = Z4 = Standard capacitor.

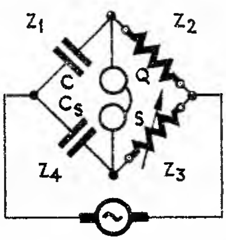
When this bridge is balanced -

$$Z1Z3 = Z2Z4$$

$$\text{or } \frac{1}{\omega C}S = Q\frac{1}{\omega Cs}$$

and the unknown capacitance C =

$$\frac{S}{Q}(Cs)$$

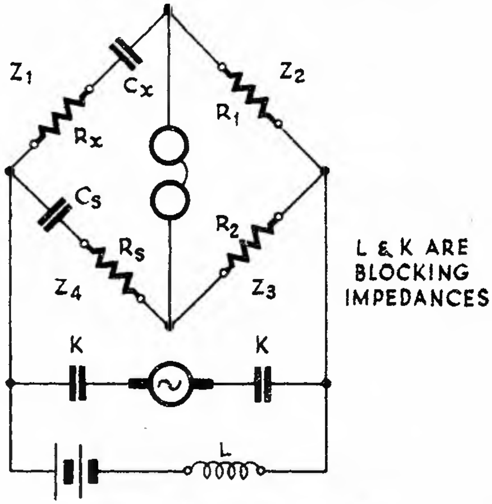


CAPACITANCE BRIDGE.

FIG. 13.

/ Electrolytic

Electrolytic Capacitors. Fig. 14 shows a method of measuring capacitance and effective resistance of an electrolytic capacitor. In this figure -



MEASURING CAPACITANCE OF ELECTROLYTIC CAPACITOR.

FIG. 14.

Z1 = Cx and Rx = Capacitance and effective resistance to be measured,
Z2 = R1 = Ratio arm,
Z3 = R2 = Ratio arm,
Z4 = Rs and Cs = Standard variable resistor and capacitor.

At balance, Z1Z3 = Z2Z4, or

$$\left(Rx + \frac{1}{j\omega Cx}\right) \quad R2 = R1 \left(Rs + \frac{1}{j\omega Cs}\right)$$

$$R2Rx + \frac{R2}{j\omega Cx} = R1Rs + \frac{R1}{j\omega Cs}$$

Equating non-reactive terms and -

$$R2Rx = R1Rs$$

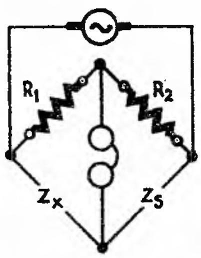
$$\text{or } Rx = \frac{R1Rs}{R2}$$

Equating reactive terms and -

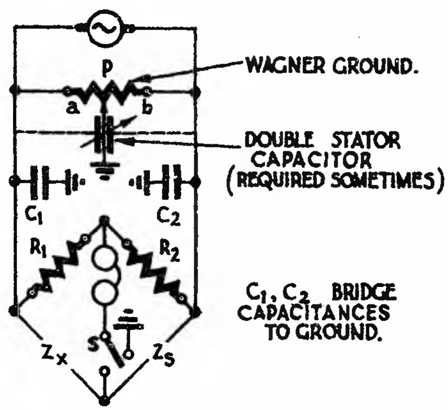
$$\frac{R2}{j\omega Cx} = \frac{R1}{j\omega Cs}$$

$$\text{or } Cx = \frac{R2Cs}{R1}$$

Wagner Earth. When the bridge shown in Fig. 15a is used to measure high impedances (such as 50,000 ohms and over), the results are usually in error, and the balance point is affected by placing the hands upon any part of the telephone receivers or their leads. This is caused by the fact that the neutral arm of the bridge is not at ground potential, and, as a result, spurious currents flow from the neutral arm to ground. Difficulties from this action are eliminated by using a Wagner earth to bring the neutral arm of the bridge to ground potential, as shown in Fig. 15b. The Wagner earth consists of a potentiometer P of perhaps 500 to 1,000 ohms resistance. The principal ground capacitances that cause trouble are lumped together as C1 and C2 in Fig. 15b. The low resistance sections a and b of the potentiometer are in shunt with these capacitances and so practically short-circuit them, causing the grounding point to be controlled by the slider on P rather than the capacitances to ground. By adjusting the slider so that $\frac{a}{b} = \frac{R1}{R2}$, the neutral arm is brought to ground potential, the bridge is balanced without body effects and the results are correct.



(a)



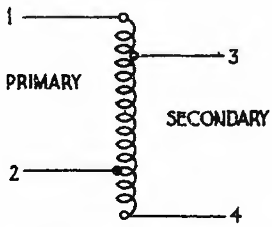
(b)

FIG. 15. USING THE WAGNER EARTH FOR CORRECTION.

The bridge is first balanced as well as possible without regard to the Wagner earth adjustment. The switch S in Fig. 15b is then operated so that the telephone receivers are connected between the neutral arm and ground and the slider on the potentiometer adjusted until no sound is heard, which is an indication that the neutral arm is at ground potential. The switch S is now returned to its original position, placing the receivers across the neutral arm, and the balance is completed.

3.3 Mutual Inductance and Coefficient of Coupling. The usual procedure for measuring the mutual inductance between two coils consists of connecting the two coils in series and measuring the total inductance of the combination, after which the terminals of one coil are reversed and the process repeated. The mutual inductance is then one-fourth of the difference of the two measured inductances. (For basic principle of this, see Paper No. 1, Radio I.

In the case of an auto-transformer, as shown in Fig. 16, the foregoing procedure is not possible. The following is a suitable method -



MUTUAL INDUCTANCE OF AUTO-TRANSFORMER.

FIG. 16.

(i) Measure impedance between 1 and 2 with 3 and 4 open-circuit = Lp.

(ii) Measure impedance between 1 and 2 with 3 and 4 short-circuit = Lc.

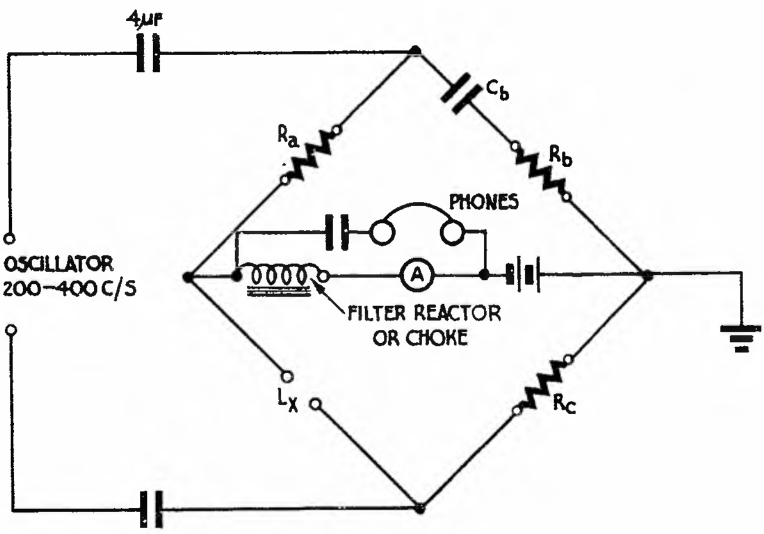
(iii) Measure impedance between 3 and 4 with 1 and 2 open-circuit = Ls.

Then $M = \sqrt{(Lp - Lc)Ls}$

also -

$$\text{Coefficient of coupling} = K = \frac{M}{\sqrt{LpLs}}$$

3.4 Incremental Inductance. Incremental inductance is the inductance which is offered to the flow of an A.C. superimposed upon a D.C., and is of particular importance in audio-frequency transformers and filter reactors. Fig. 17 shows the Hay Bridge adapted for this measurement. With this bridge, inductance is measured in terms of resistance and capacitance, and balance is obtained by varying Ra and Rb. The D.C. is introduced, measured and controlled in the neutral arm of the bridge, and is not affected by the process of obtaining a balance.



MEASUREMENT OF INCREMENTAL INDUCTANCE.

FIG. 17.

When the bridge is balanced, the following relations hold -

$$Lx = \frac{Ra\ Rc\ Cb}{1 + (Rb\ \omega Cb)^2} = \frac{Ra\ Rc\ Cb}{1 + \frac{1}{Q^2}}$$

$$\left(\text{where } Q = \frac{\left(\frac{1}{\omega Cb}\right)}{Rb} = \frac{\omega Lx}{Rx}\right)$$

When the losses in the inductance are low -

$$Lx = Ra\ Rc\ Cb$$

$$\text{and } Rx = \frac{Ra\ Rb\ Rc\ (\omega Cb)^2}{1 + (Rb\ \omega Cb)^2}$$

3.5 Reactance. It is more usual to calculate reactance from known constants, but it is approximately evaluated as follows. (See Fig. 18.)

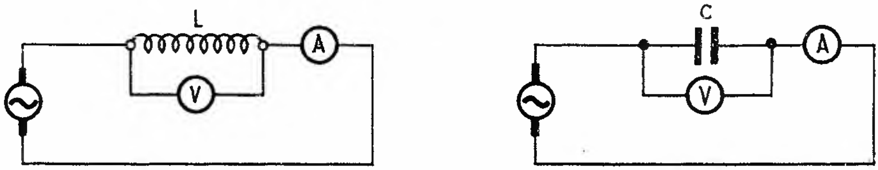


FIG. 18. APPROXIMATE VALUE OF REACTANCE.

When the voltage across the inductor or capacitor is measured by a high-resistance voltmeter, and the current in the circuit by an ammeter, then, by Ohm's Law -

$$X_L \text{ and } X_C = \frac{E}{I}$$

when E is in volts and I is in amperes.

It is apparent from this that the resistance of both the component and ammeter must be considered, together with any shunting effect of the voltmeter V.

3.6 "Q" of a Coil or Circuit. "Q" is regarded as the figure of merit of a coil or circuit, and instruments are designed to measure this characteristic.

In the Q-meter shown schematically in Fig. 19, a small voltage "e" is introduced in series with a tuned circuit as a voltage drop across a small resistance R. The circuit is then tuned to resonance, and the voltage E developed across the tuning capacitor is measured by a valve voltmeter. The circuit Q then equals E/e. The valve voltmeter is usually calibrated in terms of Q by holding the injected voltage e at a fixed value.

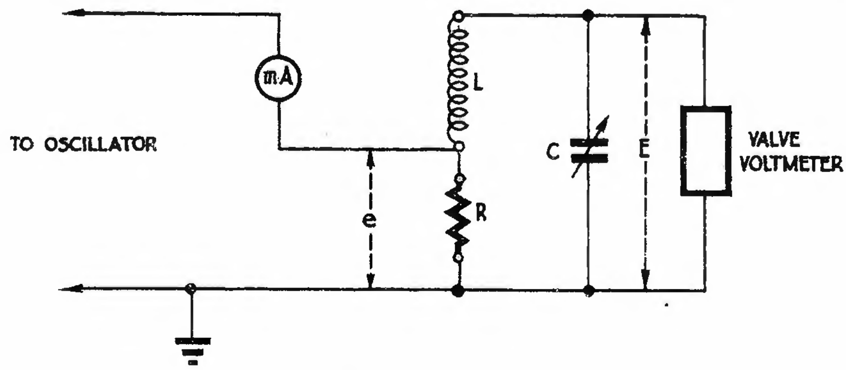
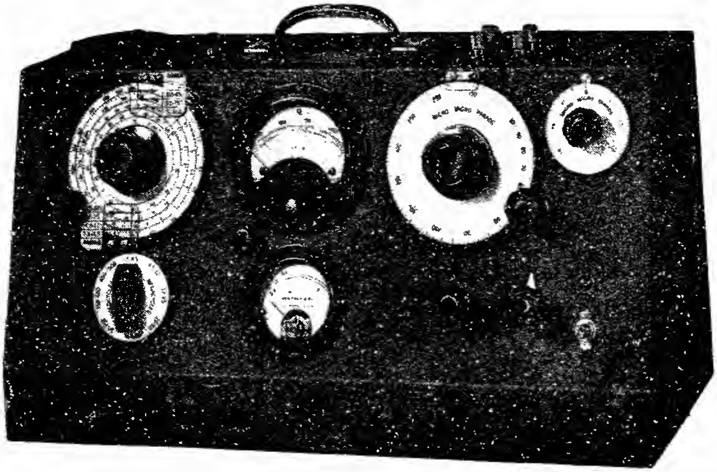
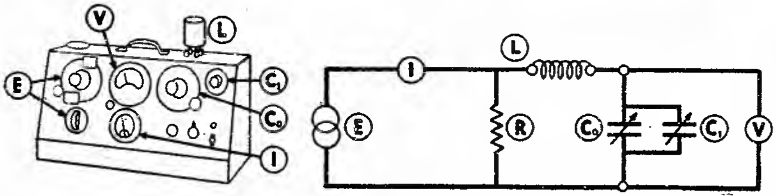


FIG. 19. BASIC CIRCUIT OF Q-METER.

A typical Q-meter usually is provided with a calibrated tuning capacitor, so that the apparent coil inductance is determined from the capacitor setting and frequency. Careful attention to design details minimises errors. Q-meters are frequently used to measure reactance, resistance or conductance of choke coils, dielectrics, etc., by the substitution method. The procedure involved consists in making two measurements. First, a convenient coil is used to form a resonant circuit, and the circuit Q is measured. The unknown impedance is then placed in parallel or series with the coil, as desired, the circuit returned to resonance, and the Q determined again. The reactance of the unknown is given by the change in tuning capacitance required, and the resistance or conductance is calculated from the effect on the Q and a knowledge of frequency and tuning capacitance.

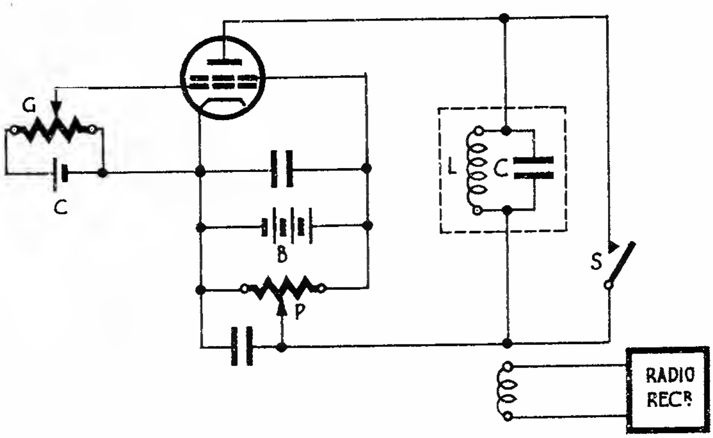
/ TYPICAL





TYPICAL Q-METER.

3.7 Parallel Resonant Impedance. A convenient method of experimentally determining the resonance impedance of a parallel circuit is shown in Fig. 20.



MEASUREMENT OF PARALLEL RESONANT IMPEDANCE.

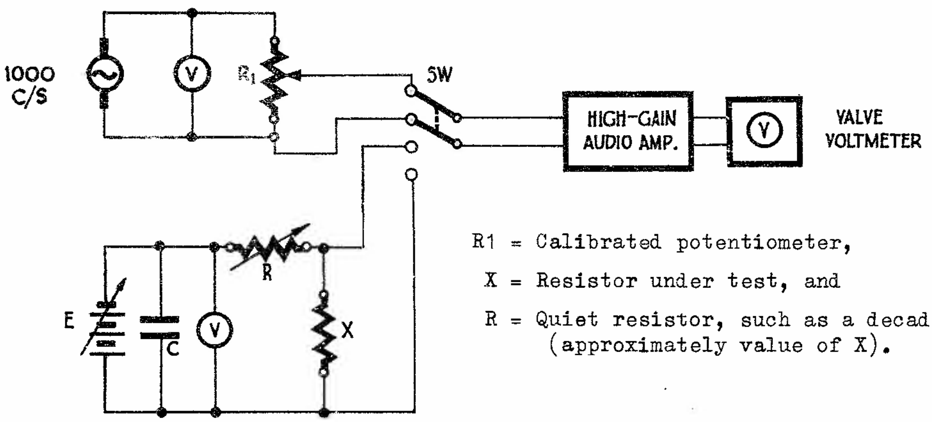
FIG. 20.

LC is the circuit to be measured, and the method is based on the fact that the circuit just commences to oscillate when the impedance of the LC anode circuit is numerically equal to the "negative resistance" of the valve characteristic. The potentiometers G and P control the grid bias and anode voltages respectively. A receiver or other indicating device is loosely coupled to LC to detect the point where oscillation starts. G and P are adjusted until the circuit is on the verge of oscillation, then LC is short-circuited by closing the key S, P is varied a few volts above and below the setting at which oscillation occurred and the values of anode current are noted. The values of C and B are unchanged during this latter adjustment. The slope of the Ea-Ia curve through the value of Ea where oscillation occurred is the negative resistance and is numerically equal to the impedance Zo. If L and C are known, R can be computed from -

$$Zo = \frac{L}{RC} \text{ or } R = \frac{L}{Zo\ C}$$

$$\text{that is, } Zo = \frac{Ea1 - Ea2}{Ia1 - Ia2} = \frac{L}{RC}$$

3.8 Measurement of Noise in Resistors. In those items of radio equipment operating at very low levels, it is of paramount importance to use components which do not contribute appreciably to the noise level of the circuit. Resistors, unless carefully designed, are capable of introducing appreciable noise voltages into circuits. To enable resistor noise-generating capabilities to be measured, the circuit of Fig. 21 has been found satisfactory in use.



R1 = Calibrated potentiometer,
X = Resistor under test, and
R = Quiet resistor, such as a decade (approximately value of X).

FIG. 21. MEASURING NOISE IN RESISTORS.

In operation, E is adjusted to the normal operating voltage of the circuit in which the resistor is to be used.

(i) The switch SW is operated to the down position, thus connecting the resistor X to the amplifier circuit. The noise generated across the resistor is thus amplified, and the amplifier gain adjusted to give a convenient deflection on the valve voltmeter.

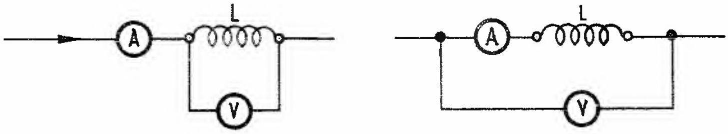
(ii) The switch SW is then moved to the up position and the potentiometer adjusted until the valve voltmeter reading is the same as in step (i).

(iii) The setting of the calibrated potentiometer (which is calibrated in microvolts) shows the equivalent R.M.S. voltage variation existing across the particular unknown resistor being tested. It can then be stated that the noise of the resistor is equivalent to so many microvolts R.M.S. for the particular voltage drop.

4. MISCELLANEOUS MEASUREMENTS.

4.1 Some Other Methods of Measuring Approximate Inductance of a Coil.

(i) Inductance of a Coil (see Fig. 22).



(a) (b)

FIG. 22. MEASURING INDUCTANCE.

When ω = frequency in radians/second, reactance = ωL ohms.
Neglecting the voltmeter resistance, the voltage across the coil L = $I\omega L$.

$$\therefore L = \frac{E}{\omega I} \text{ henrys.}$$

when E is in volts and I is in amperes.

(ii) Three-Voltmeter Method (see Fig. 23).

The resistance of the inductor is indicated by r.

R1 = Standard resistance. The resistance of V1 is very high compared with R1. R1 approximates impedance of L.

V, V1 and V2 = Voltmeter readings.

Voltmeter readings are plotted in the form of a graph as in Fig. 23b.

(a) Plot along horizontal the value V1 in phase with current I.
(b) With centre A and radius V2, describe an arc AC.
(c) With centre B and radius V, describe an arc BC.
(d) Join intersection of arcs to AB by a perpendicular to D.
(e) Call the angle DBC, θ.

$$\text{from this, } V = I\sqrt{r^2 + \omega^2 L^2}$$

$$\text{and } V1 = I(R1).$$

$$\therefore Z = \sqrt{r^2 + \omega^2 L^2} = \frac{V(R1)}{V1}.$$

θ is the phase angle of the choke and BC is the impedance.

$$\therefore BC = Z = \sqrt{(BD)^2 + (CD)^2} = \sqrt{r^2 + (\omega L)^2}$$

from which -

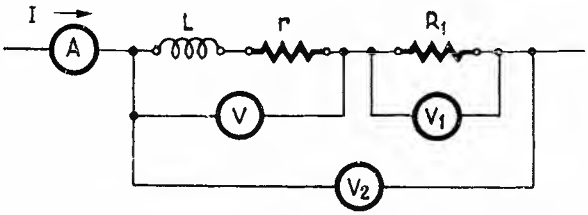
$$BD = \text{resistance } r = Z \cos \theta$$

$$CD = \text{inductive reactance } \omega L = Z \sin \theta$$

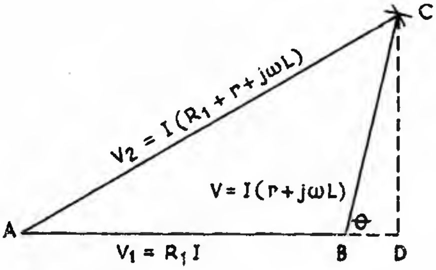
$$\text{this is, } L = \frac{Z \sin \theta}{\omega}$$

Fig

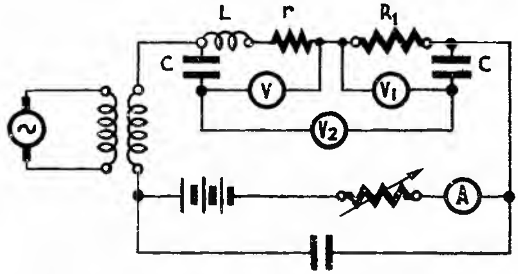
Fig. 23c shows the application of this test to a choke carrying D.C.



(a) Circuit.



(b) Plotting the Readings.



(c) 3 Voltmeter Method for Choke with D.C.

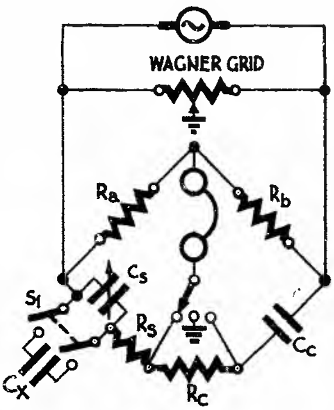
FIG. 23. MEASURING INDUCTANCE.

4.2 Substitution Methods of Measuring. Measurements of several fundamental constants or characteristics are often conveniently carried out by adjusting known constants until similar effects are obtained with either the known or unknown values in the circuit.

Three examples have been chosen to illustrate this method, and they serve to indicate the application of the principle -

Capacitance (low values).
Capacitance, inductance and impedance at radio frequency.
Gain or loss.

Capacitance Measurement. The substitution method is the most satisfactory method of measuring capacitances of 0.001 µF and less. This method can be applied to any bridge circuit capable of measuring capacitance, with results depending on the accuracy of the calibration curve of the standard capacitor. Fig. 24 shows the circuit arrangements -



CAPACITANCE MEASUREMENT BY SUBSTITUTION.

FIG. 24.

where Ra and Rb = Ratio arms,

Cs = Standard capacitor,

Cc = Any available capacitance a little larger than Cx, and,

Cx = Unknown capacitor.

Balance by varying Cs (with Cx in parallel) and equalising the power factors by Rc.

Remove Cx and rebalance by varying Cs.

Unknown capacitance is then the difference between the two readings of Cs, and is independent of any bridge errors, as bridge conditions do not vary.

Radio-Frequency Measurements. The simplicity and directions of the substitution method make this the most widely used of all measuring methods at radio frequencies.

/ Capacity.

Capacity. A calibrated variable capacitor and the capacitance to be measured are connected in parallel and used to tune a circuit to resonance at some convenient frequency. The unknown capacitance is then disconnected and the circuit retuned for resonance.

The unknown capacitance is equal to the change in capacitance of the calibrated capacitor. The exact frequency is unimportant, but some indication of resonance is required, such as a meter or heterodyne frequency meter. (The Q-meter is also useful for this test.)

Apparent Inductance of a Coil. The apparent inductance of a coil is measured by connecting it across a calibrated variable capacitor and determining the capacitance required to make the resulting tuned circuit resonant at some particular frequency.

$$\text{From the data available, } L = \frac{1}{\omega^2 C}$$

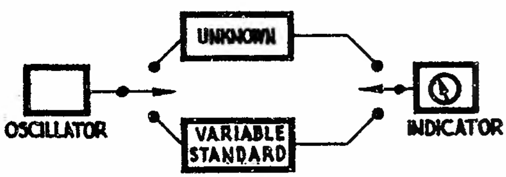
Another method is to tune a circuit to resonance, then add the coil in parallel and retune for resonance.

$$L \text{ then equals } \frac{1}{\omega^2 (C1 - C2)}$$

where C1 and C2 are the readings of the capacitor.

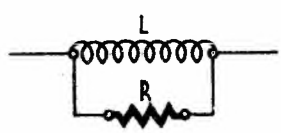
Gain or Loss. Fig. 25 shows a method of measuring gain or loss by the substitution method. The main point is to ensure that all impedances are matched. The circuit is adjusted until the output reading is the same whichever unit is in the circuit.

4.3 Radio-Frequency Choke Coils. The impedance of a radio-frequency choke coil can be represented by a reactance shunted by a resistance, as shown in Fig. 26.



GAIN OR LOSS BY SUBSTITUTION METHOD.

FIG. 25.



R.F. CHOKE COIL EQUIVALENT.

FIG. 26.

The choke coil reactance is obtained by the following procedure - A resonant circuit containing a variable capacitor is tuned to the frequency at which the choke coil characteristics are desired. The choke coil is then connected in parallel with the tuned circuit, and the capacitor adjusted to restore the original resonant frequency.

The equivalent reactance of the choke coil is then -

$$\text{Reactance } X1 = \frac{1}{\omega \Delta C}$$

Where ΔC is change in calibrated capacitor required to compensate for the addition of the choke coil.

/ The

The equivalent resistance R1 is determined by measuring the parallel resonant impedance at the same resonant frequency of the tuned circuit before and after the addition of the choke coil. When the equivalent parallel resistances before and after the addition of the choke coil are R' and R" then -

$$R1 = \frac{R''R'}{R'' - R'}$$

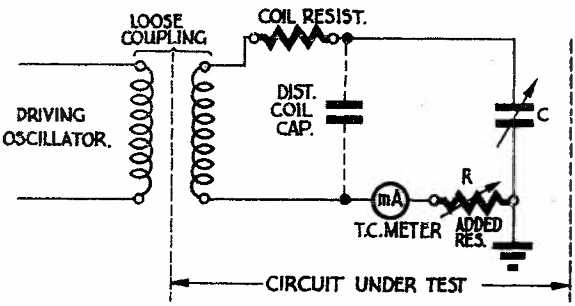
$$\text{From these, } X = X1 \frac{R1^2}{X1^2 + R1^2}$$

$$R = R1 \frac{X1^2}{X1^2 + R1^2}$$

$$Z = \sqrt{R^2 + X^2}$$

4.4 <u>Resistance Variation Method</u>. The resistance variation method of determining the resistance of tuned circuits makes use of the fact that, at resonance, the current in a circuit is equal to the applied voltage divided by the circuit resistance. When the applied voltage is kept constant, it is then possible to deduce the actual circuit resistance by the current change that results when a known resistance is added to the circuit.

Circuit arrangements suitable for carrying out the necessary measuring operations are shown in Fig. 27.



MEASUREMENT OF RADIO-FREQUENCY RESISTANCE BY RESISTANCE VARIATION METHOD.

FIG. 27.

The circuit under test is loosely coupled to a driving oscillator and has in series with it a thermo-couple ammeter mA and an adjustable resistor R. The circuit is first tuned to resonance with the driver, and the current in the milliammeter (Io) is observed when R is zero. A known amount of resistance is then added by R, the circuit returned to resonance (if necessary) without changing the coupling to the driver, and the resulting current noted (I1). The apparent series resistance of the circuit is then given by the following formula -

$$\text{Apparent series resistance of tuned circuit} = R\left(\frac{I1}{Io - I1}\right)$$

This principle has many applications, for example, measurement of input impedance of a radio receiver.

5. VALVE CHARACTERISTICS.

5.1 The term "characteristic" is used to designate the graphical relation between two or more variables, such as voltage and current. As applied to any electrode circuit in a valve, the characteristic designates the relation between the voltage on an electrode and the current flowing in the circuit of that electrode. Another characteristic of basic importance is the transfer characteristic, which is the relation between the voltage on an electrode and the current in the circuit of another electrode.

The three characteristics which are generally considered are -

$$\text{Amplification Factor or } \mu = \frac{\Delta Ea}{\Delta Eg} \quad \text{(Ia constant).}$$

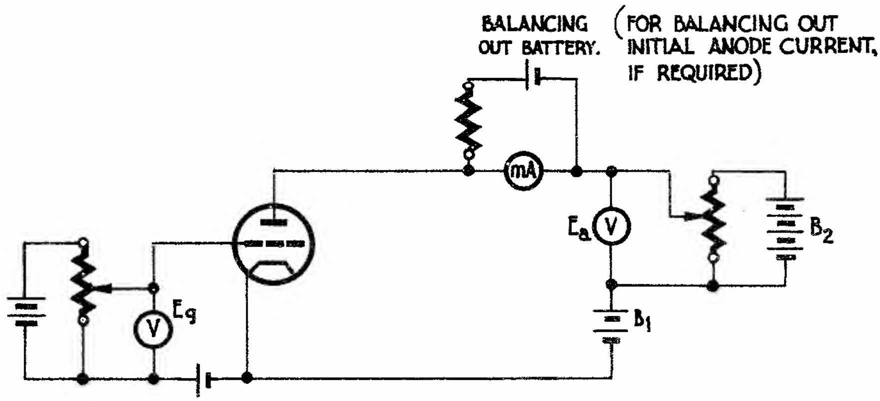
$$\text{Anode Resistance or Ra} = \frac{\Delta Ea}{\Delta Ia} \quad \text{(Eg constant).}$$

$$\text{Mutual Conductance or Gm} = \frac{\Delta Ia}{\Delta Eg} \quad \text{(Ea constant).}$$

5.2 As was mentioned in an earlier Paper, it is possible to determine the above graphically from the characteristic curves of the valves. A high standard of accuracy is not obtained by this method, however, and it is usually more desirable to measure these quantities. Two methods are available for this, the choice depending on the desired accuracy. These methods are -

(i) Incremental method.
(ii) Bridge methods.

5.3 Incremental Method. Fig. 28 shows a circuit for determining valve coefficients by incremental adjustments of voltages and currents.



MEASURING VALVE COEFFICIENTS BY INCREMENTAL ADJUSTMENTS OF ELECTRODE VOLTAGES AND CURRENTS.

FIG. 28.

Although a triode is shown, pentodes or tetrodes are substituted by adding the necessary meters to the other electrodes as required. Reasonable accuracy is obtained, but the practical difficulty is that, when the increments are small they cannot be read accurately without complicating the circuit, and when they are large, the accuracy is poor because of the requirements of the definitions given above. The exact definitions call for very small increments.

/ To

To determine the amplification factor by the incremental method, a convenient increment is added to the grid bias, say 2 volts, and the anode voltage is changed until the original current is restored. Suppose, in this case, the anode voltage change is 20 volts, then, by definition, the Amplification Factor would be $\frac{20}{2} = 10$.

The Anode Resistance is measured similarly by adding an increment to the anode voltage and reading the resulting increment of anode current. The Mutual Conductance is obtained by adding an increment to the grid bias and noting the change of anode current while the anode voltage is kept constant.

5.4 Bridge Methods.

Amplification Factor. Fig. 29 is a circuit by which the Amplification Factor is determined. The value of R1 is about 10 ohms, and R2 is adjusted for no sound in the head receivers. The value of μ is given by -

$$\mu = \frac{R2}{R1}$$

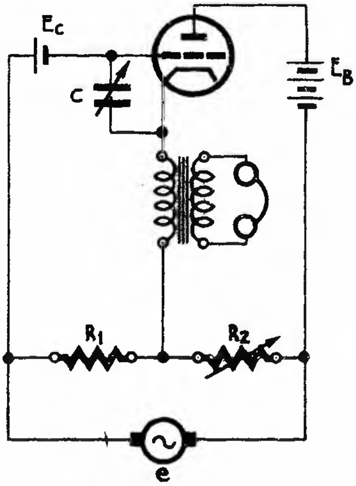


FIG.29. MEASUREMENT OF μ.

The capacitor C is sometimes necessary to balance out the internal valve capacitances and secure a good null point. The head receivers are preferably connected to the secondary of a small step-up transformer offering low primary resistance to the flow of the D.C. component of anode current. Shunting the receivers across a low resistance choke coil is an alternative method. The impressed alternating voltage "e" is no larger than is necessary to secure a good balance.

Anode Resistance. The Anode Resistance is measured by using the anode circuit of the valve as the fourth arm of an ordinary bridge, as shown in Fig. 30. When the bridge is balanced -

$$Ra = \frac{R2R3}{R1}$$

R3 is about 10,000 ohms, or some value comparable to Ra, and R2 is fixed at 10 or 100 ohms. A balance is secured by varying R1, which is comparable to R2 in magnitude. The variable capacitor C balances out inter-electrode capacitances as before. When Ra is large, the value of C required may be inconveniently large, in which case C is shunted across R3. A choke coil or transformer should be used with the head receivers, as in the case of μ.

Mutual Conductance. Mutual Conductance is measured by the arrangement of Fig. 31.

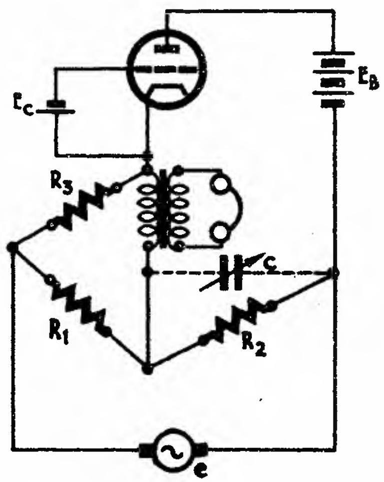


FIG.30. MEASUREMENT OF Ra.

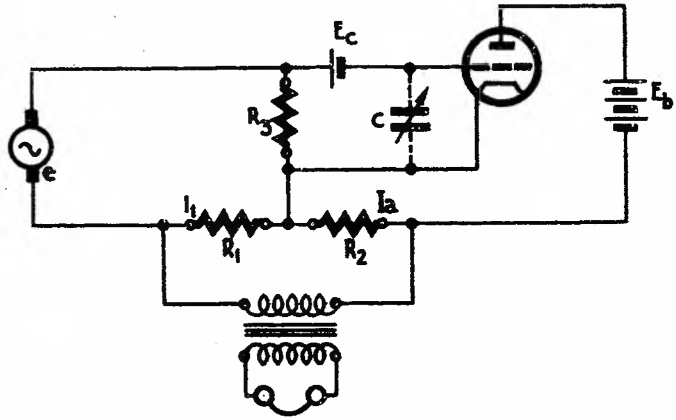


FIG. 31. MEASUREMENT OF Gm.

For a balance -

$$I1R1 = IaR2 = \frac{\mu Eg}{Ra + R2} R2 = \frac{\mu I1R3}{Ra + R2} R2.$$

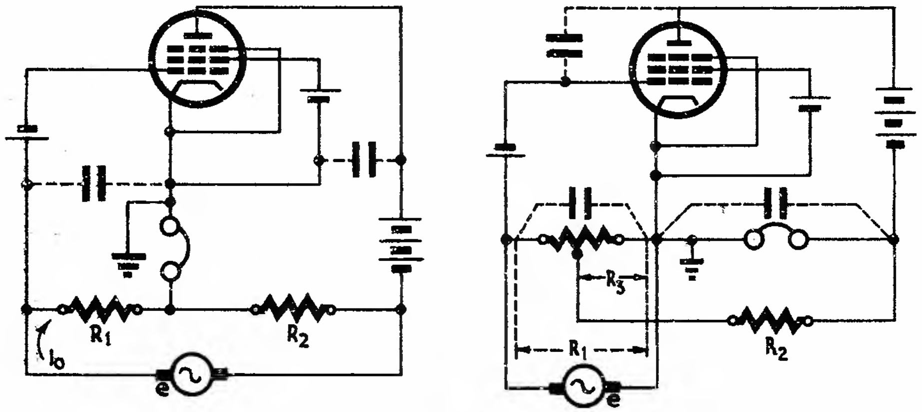
When R2 is neglected in comparison with Ra, the above expression becomes -

$$R1 = \frac{\mu}{Ra} R3R2 = Gm\ R3R2$$

$$\text{or } Gm = \frac{R1}{R2R3}$$

The variable capacitor C is used, as before, to balance out inter-electrode capacitances, where necessary. The resistance R1 is varied to secure a balance, while R3 is a fixed value of 1,000 ohms and R2 fixed at 100 ohms. The mutual conductance is usually expressed in micromhos, so that, with the above values of resistance, Gm in these units is ten times the setting of R1.

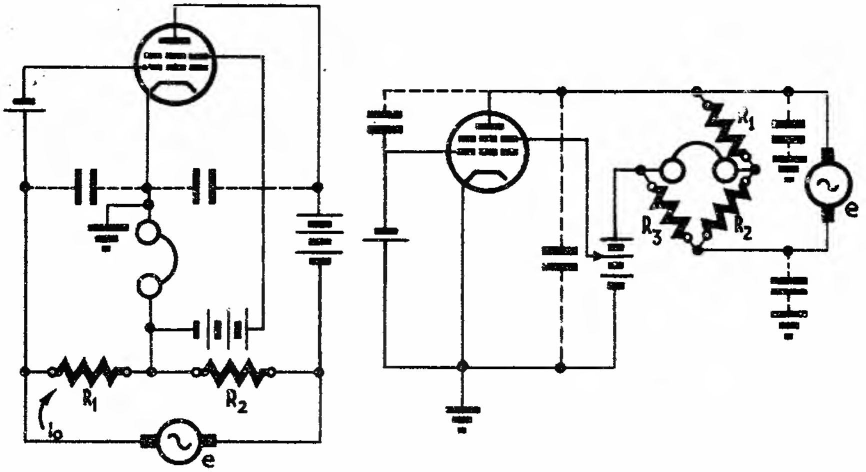
Pentode and Screen-Grid Valves. Similar measuring principles are applied to pentode and screen-grid valves, with minor modifications. Typical examples for a pentode are shown in Fig. 32.



(a) Measurement of μ. (b) Measurement of Gm.

FIG. 32. MEASUREMENT OF μ AND Gm (PENTODE VALVES).

Typical examples for a screen-grid valve are shown in Fig. 33.



(a) Measurement of μ. (b) Measurement of Ra.

FIG. 33. MEASUREMENT OF μ AND Ra (SCREEN-GRID VALVES).

The dynamic resistance of any electrode circuit is measured by making this circuit the unknown arm of a Wheatstone Bridge. Likewise, the mutual conductance between any two electrodes is determined by applying the drop in the resistor R1 to one electrode, and balancing out the resulting current that flows in the circuit of the other electrode by the resistances R2 and R3. The amplification factor is determined by applying the voltage drop developed across R1 to one electrode, balancing the resulting effect produced in the other electrode circuit by a potential developed across R2, and locating the head receivers in the part of the circuit where the current is to be constant. The methods for balancing the reactive current in the receivers are the same as in the case of triodes.

The above bridge methods of measuring valve characteristics give reasonably accurate results for most purposes. More elaborate set-ups are used to obtain greater accuracy, or for laboratory use, but the description of these is beyond the requirements of these notes.

5.5 Conversion-Transconductance. The conversion-transconductance of frequency converter valves is determined from measurements of the magnitude of a single beat-frequency component (f1 - f2) or (f1 + f2) of the output current, and of the magnitude of the input voltage of frequency f1.

The direct voltages applied to the electrodes are held constant, the magnitude of the alternating voltage of frequency f2 is constant, and impedance to the beat frequency of the load in the anode circuit is low compared with the anode resistance of the valve. For greatest accuracy, the smallest practical input signal voltage of frequency f1 is used in making these measurements. A circuit diagram suitable for making these measurements is shown in Fig. 34.

$$\text{Conversion Conductance} = \frac{\text{Magnitude of } (f1 + f2) \text{ or } (f1 - f2)}{\text{Magnitude of control-electrode voltage } f1}$$

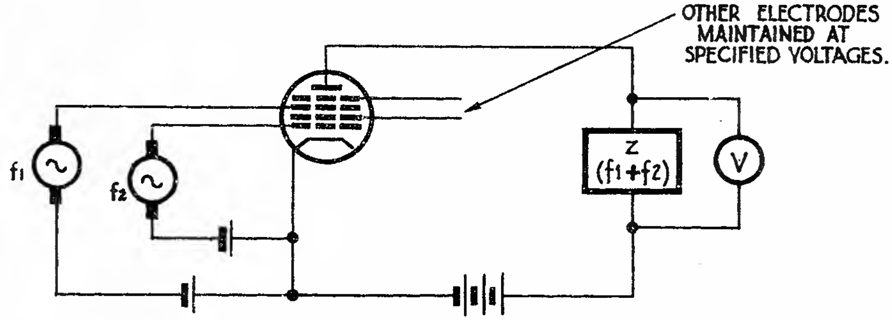
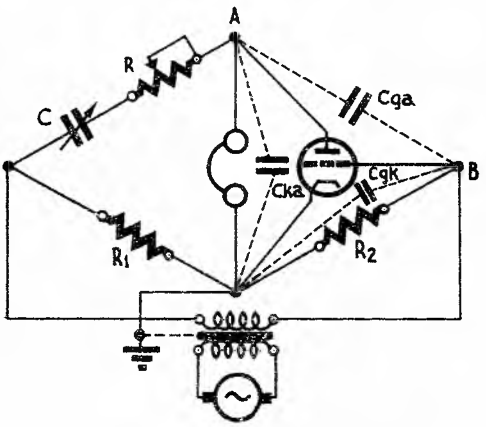


FIG. 34. CONVERSION-TRANSCONDUCTANCE MEASUREMENTS.



BRIDGE METHOD FOR MEASURING DIRECT INTER-ELECTRODE CAPACITANCE.

FIG. 35.

5.6 Inter-electrode Capacitance. Inter-electrode capacitance is measured with the cathode cold and with no direct voltages present. The valve is in its complete form, but not in its socket or holder.

A bridge method for the measurement of direct inter-electrode capacitance in a triode is shown in Fig. 35. In this figure, the capacitance Cga is shown under measurement and is connected across an arm of the bridge, the other capacitances being in shunt across the receivers and R2. The other capacitances are measured in turn by suitable interchange of connections, the capacitance under measurement being placed in the upper right arm of the bridge between points A and B.

The resistor R balances the capacitance Cgk (grid-cathode) which is in parallel with R2, and also corrects any accidental phase shifts present elsewhere in the bridge. When the bridge is balanced, the capacitance is -

$$Cx = Cga = \frac{R1\ C}{R2}$$

/ 6.

6. CHARACTERISTICS OF AUDIO-FREQUENCY WAVES AND CIRCUITS.

6.1 Measurement of some of the characteristics of audio waves is covered in the Paper on radio receiver tests in Paper No. 8 of Radio II, together with a description of some of the instruments. The chief characteristics of audio-frequency waves and circuits may be regarded as being -

Frequency Variation.

Harmonic Distortion.

Phase Shift.

Signal-Noise Ratio.

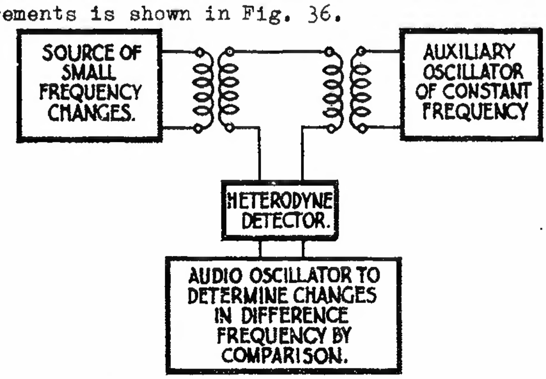
Frequency Response.

Power Output of Valves and Amplifiers.

Gain or Loss of Circuits, Components and Networks.

Overload Characteristics.

6.2. Frequency Variation. There is an important group of measurements which involves the accurate determination of small frequency differences. Typical of these are the measurement of radio frequencies by heterodyning to an audio frequency and measuring small capacitances by the change produced in the frequency of an oscillator, where frequency changes of only a few cycles need to be measured. A typical circuit for making such measurements is shown in Fig. 36.



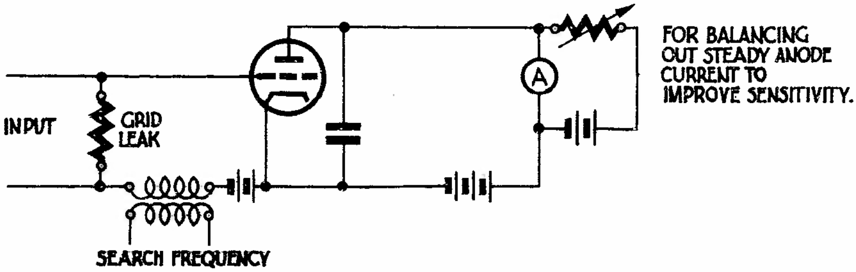
HOW A SMALL CHANGE IN FREQUENCY MAY BE ACCURATELY MEASURED.

FIG. 36.

An auxiliary oscillator of good stability over short time intervals is adjusted until its frequency differs by a convenient amount from the frequency of the oscillator under test, and some means (commonly comparison with an audio-frequency oscillator) is used to measure the difference frequency produced when the two radio-frequency oscillations are applied to a detector. When the frequency of the oscillator under test now changes by a small amount, even only two or three cycles, the audio-frequency beat rate changes by the same number of cycles, and the difference is accurately determined because it represents a comparatively large percentage change of the audio frequency.

6.3 Harmonic Distortion. A number of methods of measuring or determining harmonic distortion has been developed, and several typical methods are described.

Valve Voltmeter Method. This method is shown in Fig. 37, and involves superimposing a search voltage upon the wave being analysed and then applying the resultant wave to a full-wave square-law valve voltmeter. In such an arrangement, the volt-meter gives a steady deflection, depending only upon the effective value of the applied wave. When the search voltage frequency is within a fraction of a cycle of some frequency component of the unknown wave, then beats are superimposed upon this steady deflection. It is thus possible to measure the amplitude and frequency of each component of the unknown wave by varying the search frequency, and noting the frequencies at which beats occur and the amplitude of the beats.



FULL-WAVE SQUARE-LAW VALVE VOLTMETER HARMONIC ANALYSER.

FIG. 37.

This type of harmonic analyser is most satisfactorily calibrated by the substitution method, that is, by substituting for the unknown wave an adjustable known voltage, which gives the same amplitude of beats as developed by the unknown wave and has the same frequency. For example, when the search frequency gives beats at 300, 600, 900, 1,200, etc., c/s, it is assumed that the fundamental frequency is 300 c/s, and the other frequencies are the harmonics (2nd, 3rd, etc.). The amplitudes of these beats are noted and then the instrument is calibrated as described.

Tuned-Amplifier or Harmonic Analyser. A complex wave is analysed by applying it to a tuned amplifier and adjusting the resonant circuits to separate the component to be measured so that it can be evaluated. This method of analysis is capable of measuring very small frequency components, even when other components of large amplitude are present. However, the equipment required for this method makes it more suitable for use in laboratory measurements.

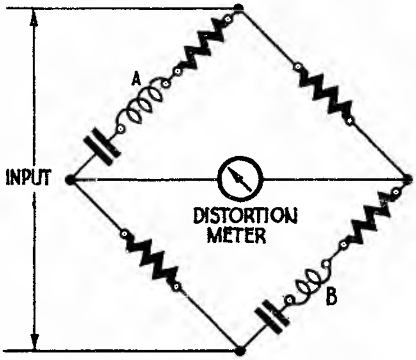
Heterodyne Method of Analysis. This method is referred to in Radio II. In this method, the component to be measured has its frequency increased to a predetermined value by heterodyne action, and is then amplified and measured at this fixed frequency. The amplitude of the harmonic component is obtained by means of a suitable calibration, while its frequency is determined from the calibration of the tuning dial on the instrument.

Resonance Bridge Method. The total R.M.S. value of harmonics contained in a current wave can be obtained by suppressing the fundamental frequency and measuring the remaining portion of the wave with the use of a thermo-couple or square-law valve voltmeter. Suppression of the fundamental can be accomplished by the use of a high-pass filter, so designed that the harmonics lie in the pass band, while the fundamental is severely attenuated. An alternative arrangement is to use a bridge, or bridged T network, which is balanced for the fundamental but is unbalanced for harmonics. Examples of such networks are shown in Fig. 38, and in Fig. 39 is shown a practical distortion meter based on this principle.

Here, the switch is at first operated to A, and the output indication is observed after capacitor C and resistor R (Fig. 38b) of the filter network have been adjusted for fundamental suppression, as indicated by minimum output. Switch S is now operated to B, and the attenuator is adjusted to give the same output indication as before. The attenuator reading is then the R.M.S. distortion in db below the fundamental.

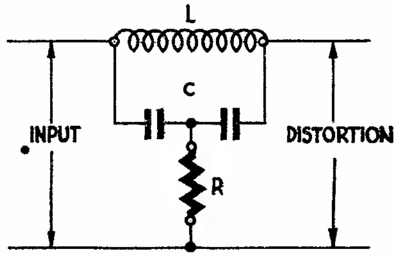
A disadvantage of this method is that it does not discriminate between the individual harmonics, and also includes any noise and hum that is present.

/ Fig. 38.



A and B are circuits resonant to fundamental frequency

(a) Bridge Circuit.



LC resonant at fundamental.
R adjusted for fundamental suppression.

(b) Bridged T Circuit.

FIG. 38. RESONANCE METHOD.

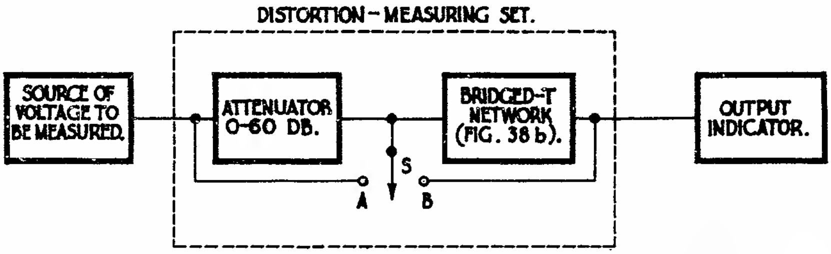


FIG. 39. DISTORTION MEASURING SET BASED ON FIG. 38b.

SET-UP FOR ADUIO-FREQUENCY MEASUREMENTS.

6.4 Phase Shift. The Cathode-Ray Oscillograph is the most widely used instrument for obtaining the phase differences between two voltages.

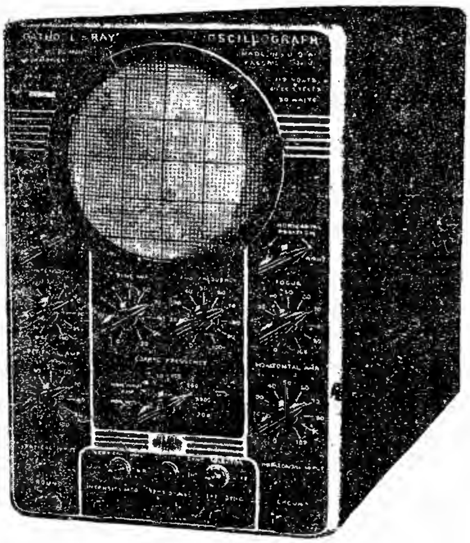
The usual procedure consists in applying one wave to the horizontal deflecting plates and the other wave to the vertical deflectors. This gives an elliptical pattern on the cathode-ray oscillograph, the exact character of which depends upon the phase and relative amplitudes of the voltages concerned. Patterns of typical cases are shown in Fig. 40, and the phase difference between two waves is given by the formula -

$$\text{Sin } \theta = \pm \frac{B}{A}$$

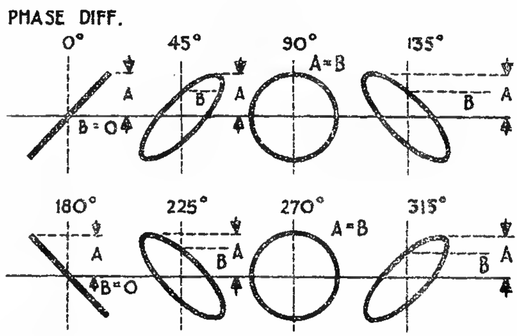
where B and A are as in Fig. 40.

The quadrant must be worked out from the orientation of the major axis of the ellipse and the direction in which the spot travels. Uncertainty as to the direction in which the spot travels is always eliminated by shifting the phase of one of the deflecting voltages in a known direction and noting the effect on the pattern.

There are many other ways of determining phase shift, but the above indicates the principles involved.



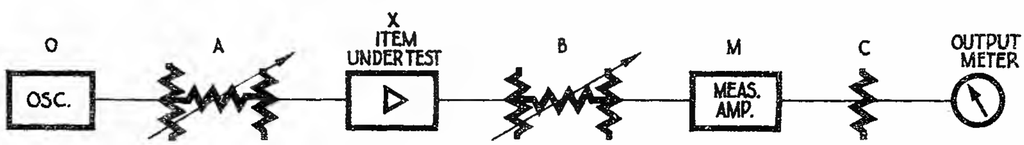
CATHODE-RAY OSCILLOGRAPH.



TYPICAL PHASE SHIFT PATTERNS.

FIG. 40.

6.5 Signal-Noise Ratio. Noise measurements of audio-frequency equipment, such as audio amplifiers, are made to determine the margin by which the noise is below the output signal, hence the term signal-to-noise ratio. This test is described in connection with the testing of radio receivers, but the principle applies equally well to other equipment. Fig. 41 is a typical set-up for measuring noise.



MEASUREMENT OF NOISE.

FIG. 41.

Fig. 41 is a typical case of noise measurement, the units used being an oscillator O, attenuators A and B, X the item under test, an amplifier M (required when noise is very low), a terminating impedance C and an output measuring meter.

Noise measurements are usually made with reference to some predetermined signal level and to a reference frequency, say, in this case, 1,000 c/s.

The attenuator A is to prevent overloading of the test unit.

Method. A 1,000 c/s tone is supplied to X, and the input level adjusted for the rated output of X. Without further adjustment of the input, attenuator B is adjusted for a convenient deflection of the output meter, as it is advisable that the attenuation in B is greater than the anticipated signal-to-noise ratio.

The tone is then removed from the input of the amplifier under test and the input circuit terminated in its correct impedance. Attenuator B is then adjusted (reduced) until the output meter reads the same as before. The difference between the two settings of "B" is the signal-to-noise ratio in db.

When amplifier M is included, the procedure is similar, B being set as before and the gain of M being adjusted for a convenient deflection of the output meter.

Noise measuring meters are available for making these measurements, as described in Paper No. 8, Radio II, in connection with Radio Receivers.

6.6 Frequency Response. Frequency response or electric fidelity is a measure of the relative amplification (or attenuation) of frequencies within a given band by an amplifier or other item of equipment designed for operation in audio-frequency circuits. Frequency response is a simple measurement, and Fig. 42 shows a typical case.

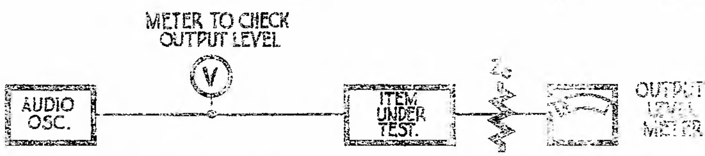


FIG. 42. FREQUENCY RESPONSE (OR ELECTRIC FIDELITY) TEST.

Procedure. After checking that the input and output circuits of the item under test are terminated in their correct impedances, the reference tone is sent from the audio oscillator (usually 1,000 c/s) and the output adjusted for a convenient output reading.

The audio oscillator is then varied over the desired range of frequencies, with constant output from the oscillator. The readings on the output level meter for each frequency are noted. A curve is then drawn, relating the relative levels of the different frequencies to the level of the reference frequency.

Amplifiers, attenuators, tone-control circuits, etc., are all tested in this manner by making the minor modifications required.

6.7 Power Output of Valves and Amplifiers. The power output of valves, amplifiers, etc., is measured by the well-established technique of using power output meters having suitable power and frequency characteristics for the particular case. There are a number of general precautions which should, however, be observed -

(i) In Class A1 amplification, the grid is not driven positive with respect to the cathode, hence the peak input grid voltage is approximately equal to the grid bias.

(ii) When the grid is driven positive, the essential characteristics of the driving circuit should be specified.

(iii) The effects of the regulation of the power-supply voltages should be specified.

(iv) The effects of feedback due to common circuit elements should be considered.

The

The power output capability of an amplifier is obtained by measurements of ~~gain at~~ a fixed frequency (usually 400 c/s) and at output levels increasing in steps from a very low value. The output power is taken as the value at which the gain is 1 db less than the gain at a very low output.

Fig. 43, which is included for interest, shows a circuit arrangement for measuring the undistorted power output of a pentode valve. The power output of the harmonic components is also determined. To ensure that any distortion present in the output is not due to the audio oscillator, a low-pass filter is inserted as shown. When an iron-cored choke is used for shunt feed in the anode circuit, care must be taken in selection of the choke to avoid the generation of harmonics in it, due to the non-linear and hysteretic behaviour of the iron.

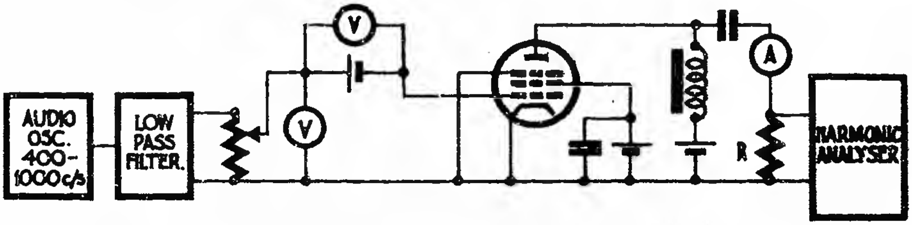
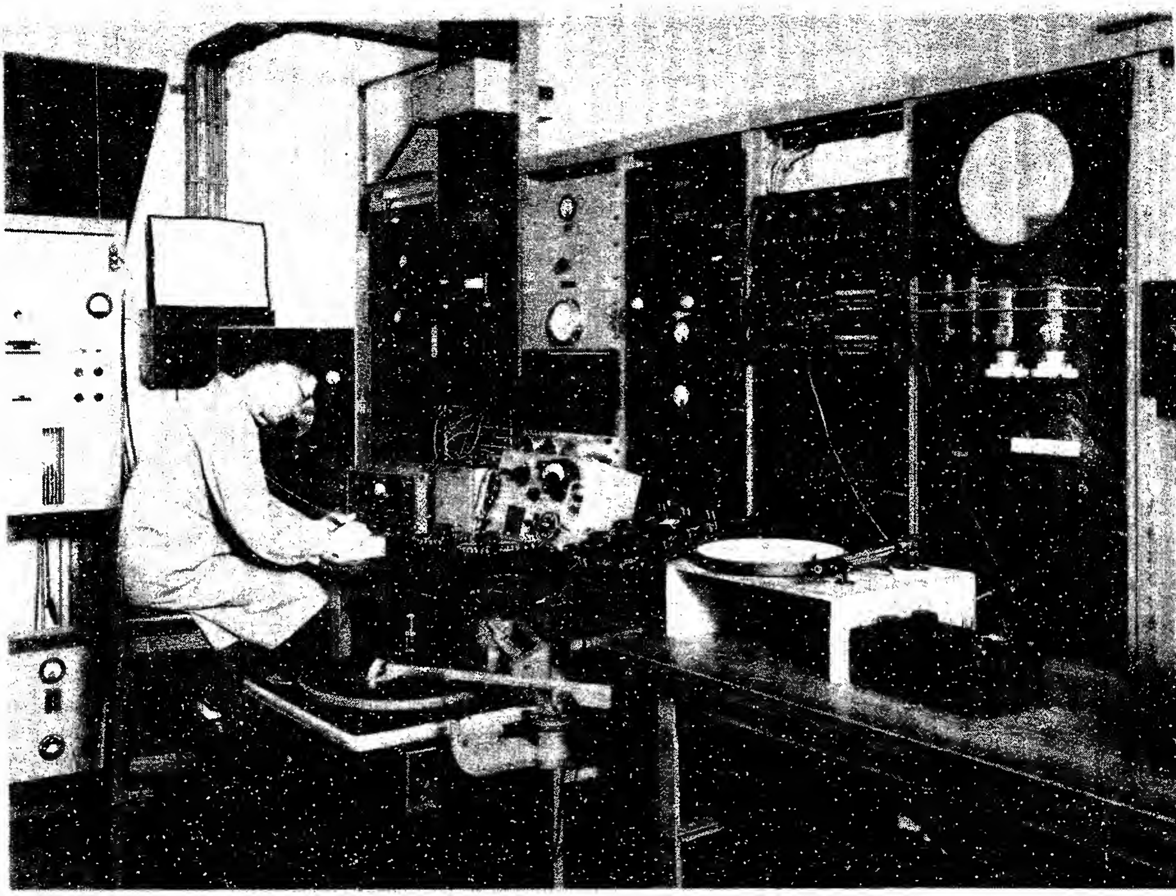


FIG. 43. MEASURING THE UNDISTORTED POWER OUTPUT OF A PENTODE.



AUDIO FREQUENCY LABORATORY.

6.8 Gain or Loss of Circuits, Components and Networks. The measurement of gain usually has two aspects, one is the determination of the gain at a fixed reference frequency, normally 1 kc/s, while the other is the determination of gain versus frequency over the design range of the unit under test. A suitable set-up is shown in Fig. 44, in which the gain of an amplifier is compared with the loss in a variable attenuator network.

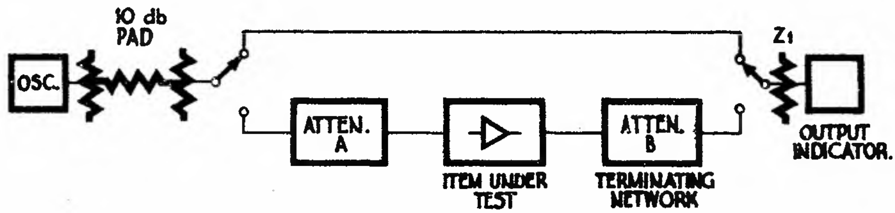


FIG. 44. MEASURING AMPLIFIER GAIN.

Points to be observed are -

(i) Impedances of all items must match throughout, or be suitably matched.

(ii) The output meter must not be overloaded.

(iii) The power fed into the amplifier must not be such as to overload it. This is prevented by keeping the attenuation of A as high as possible. After each measurement, the loss in A is increased, say 6 db, and that in B reduced 6 db without altering the reading of the output meter. Any alteration of the reading indicates overloading or oscillation.

(iv) The resistance of the voltmeter must be high compared with the circuit impedances across which it is connected.

(v) The sensitivity of the voltmeter must be such as to allow readings to be made at a power level which is within the output capabilities of the amplifier.

(vi) The load Z1 must match the 10 db and also the output impedance of B.

When these conditions are observed and the attenuator is adjusted, so that the voltmeter gives the same reading in both positions of the switch, the amplifier gain equals the sum of the losses in the attenuator and the terminating network. The voltmeter need not be calibrated, as it is only required to indicate equality of voltages at two points in the circuit. It is sometimes necessary to connect an amplifier in front of the voltmeter, in order to meet condition (v). Provided the amplifier has a high input impedance and an input transformer for bridging a balanced circuit, its frequency characteristic does not affect the results obtained. The attenuators generally available are designed to work between 600 ohm terminations. Where the amplifier under test is also designed to work between 600 ohm terminations, the test circuit is simplified, and the terminating network is then omitted. In the case of an amplifier designed to be bridged across a 600 ohm circuit (D amplifier), the attenuator must be provided with a 600 ohm terminating resistance. In the case of a microphone pre-amplifier designed to work from, say, a 50 ohm source, the attenuator must be terminated with a 600/50 ohm matching pad and the pad loss included with the attenuator loss. Where the load impedance of the amplifier differs from 600 ohms, the terminating network takes the form either of a matching pad or of a series or shunt resistor.

In taking a gain frequency characteristic, the gain is observed at the lowest and highest frequencies in the specified range, and at sufficient frequencies within the range to ensure that all significant variations of gain are observed.

/ Performance

Performance of a Single Stage. The performance of a single stage of a multi-stage amplifier is determined by the method shown in Fig. 45. In making measurements upon individual stages of a multi-stage amplifier, it is always absolutely necessary that the stage in question operates into its normal load and that all subsequent stages of amplification be in operation. This is because the input impedance of the grid into which the individual stage operates has considerable effect on the amplification characteristics, and also because the subsequent stages are the cause of regeneration that alter the amplification of the stage being investigated. Referring to Fig. 45, AB is the stage to be measured. The method is first to apply the test voltage across the input to the stage, that is, across AA, and measure the amplification from this point to the output. The test voltage is then applied across the output of the stage being studied, that is, across BB, and the amplification between this point and the load is obtained.

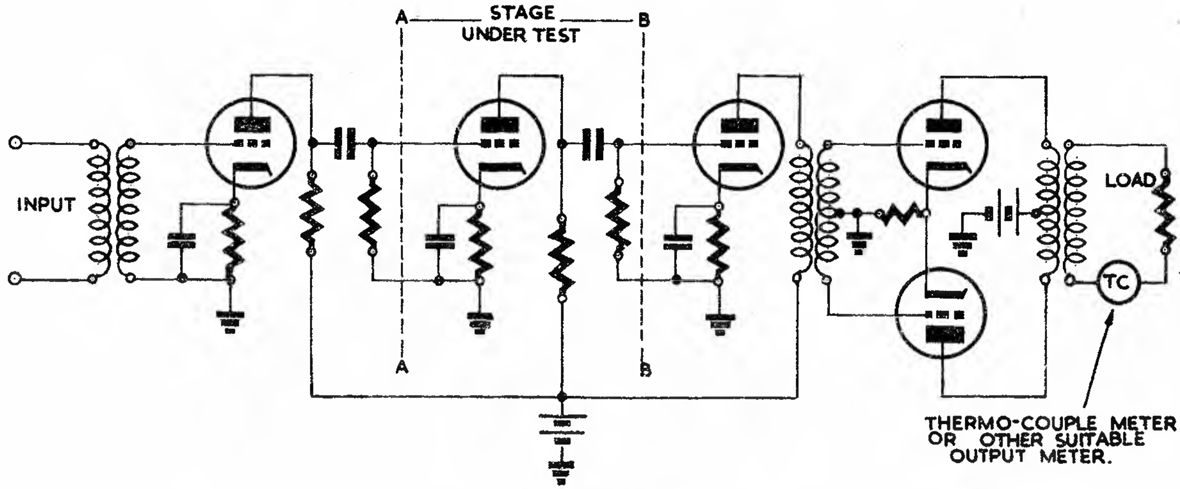


FIG. 45. METHOD OF TESTING INDIVIDUAL AMPLIFIER STAGE.

The ratio of these two amplifications is the gain of the stage between points AA and BB. The result corresponds to actual operation, since the stage operates into its normal load impedance, is subject to the same regenerative action with respect to subsequent stages of higher power level as in actual operation, and no shunt impedances that change the amplification characteristic are introduced into the circuit.

Attenuators and Networks. The principal test on a network relates to its attenuation, which is a function of frequency. A suitable set-up is shown in Fig. 46, where the network loss is compared with the loss in an adjustable resistance network of known loss. Using the set-up shown, the reference network is adjusted, so that the same meter readings are obtained in both positions of the switch, and the losses in the reference network and the network under test are then equal. This test is repeated at as many frequencies as required.

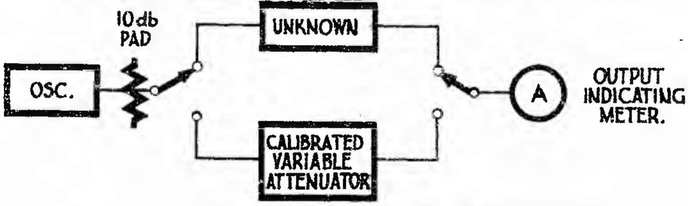


FIG. 46. MEASUREMENT OF NETWORK LOSS.

Points to be noted are -

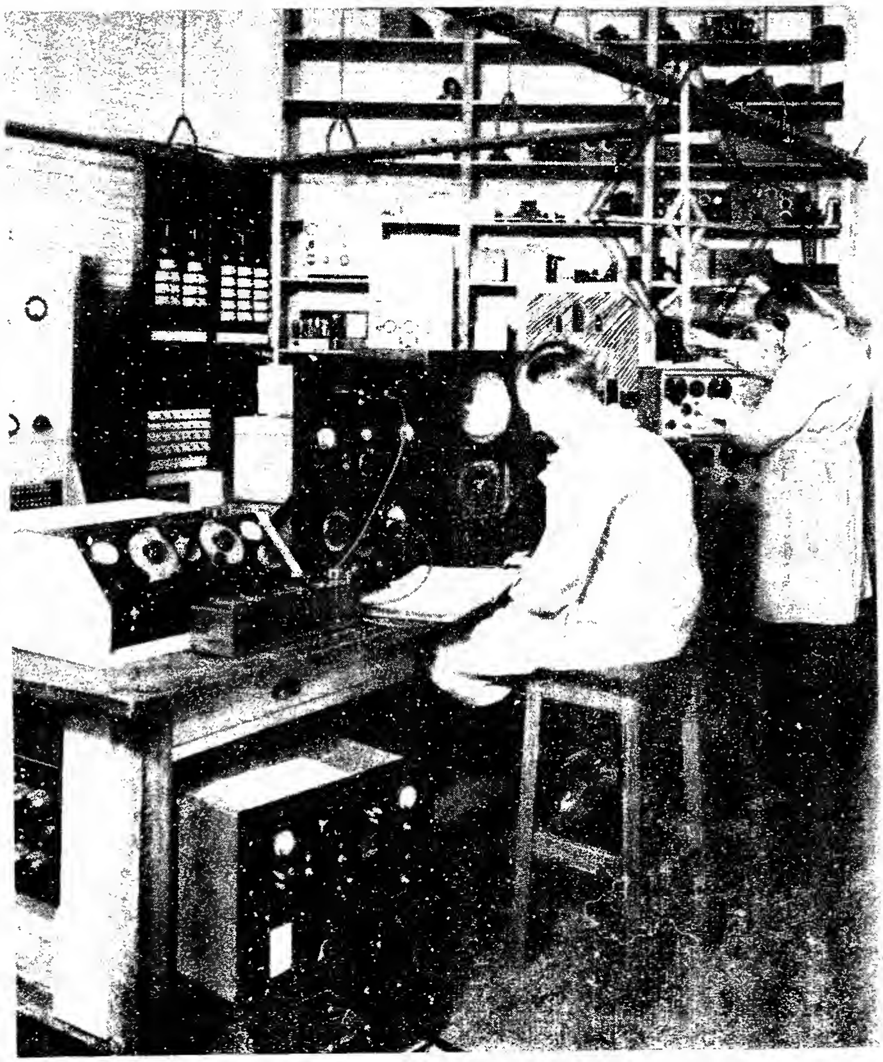
(i) The db pad is to make oscillator output impedance immaterial.
(ii) The impedances must be matched throughout, making sure that the network (unknown) is working between its correct impedances.
(iii) The variable attenuator must also work between its correct impedances.
(iv) The input power to the whole set-up must not be such as to lead to overloading of any component.

/ In

In respect of the last point, the usual testing levels do not cause overheating of any component, but some networks incorporating iron-cored inductors are apt to give rise to non-linear distortion when too high a level is applied. It is desirable to test such networks at a level not higher than that at which they are required to operate.

6.9 Overload Characteristics. The approximate point at which an amplifier overloads in actual operation is determined by relatively simple measurements of the amplifier behaviour. To test the overload point of an amplifier, a D.C. microammeter is connected in the grid circuit of the output stage valve, and a sensitive D.C. milliammeter is placed in the anode circuit. With 1 kc/s input to the amplifier, the gain is increased until either grid current flows (for a Class A amplifier) or there is a change in Ia, whichever happens first. Either occurrence indicates overload. Reduce the gain until this effect disappears, and then measure the power output. This power is then the overload point. To detect this point in intermediate stages, the same measurement is made to each stage in turn, or probably a quieter method is to bring the amplifier to the overload point of the output stage and check each stage with a cathode-ray oscillograph.

Another method suitable for various types of amplifiers is (see Fig. 44) to decrease attenuator "A" and increase "B", until a decrease of 1 db in "A" requires less than 1 db to keep the output constant. This is the overload point.



RADIO-FREQUENCY LABORATORY. (RESEARCH LABORATORIES).

7. MEASUREMENTS OF RADIO-FREQUENCY WAVES AND CIRCUITS.

7.1 Transmitters. The more important characteristics of radio transmitters, in regard to the measurements to be made, are -

Power Rating.

Modulation Percentage or Depth.

Carrier Noise.

Distortion.

Spurious Radiations.

Frequency Stability.

Wavelength.

Programme Monitoring Facilities.

Frequency Response.

7.2 Power Rating. The following are methods of measuring the radio-frequency power developed by a transmitter -

Current-Resistance Method. In this method, the current through a known resistance is measured, a thermo-ammeter and non-inductive resistor being the instruments generally used.

Photometric Method. In this method, a lamp filament heated to incandescence provides the resistive load. The D.C. or A.C. power required to heat this or a similar lamp to the same brightness is a measure of the radio-frequency power dissipated in the load. (See Fig. 47.)

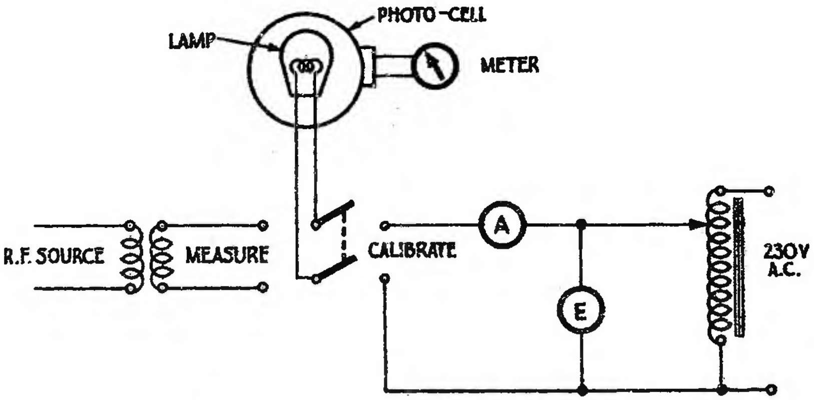
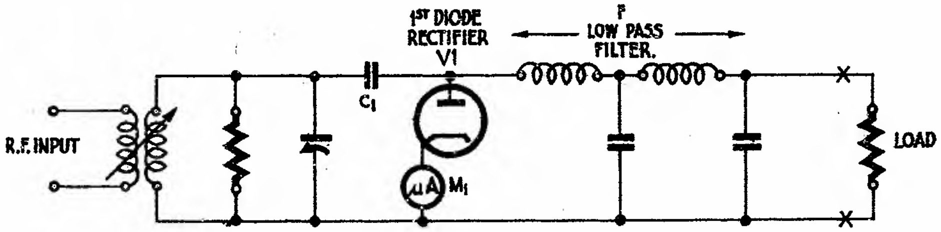


FIG. 47. PHOTOMETRIC MEASUREMENT OF RADIO-FREQUENCY POWER.

Anode Dissipation Method. In this method, the total power delivered to the filament, grid and anode circuits is measured. The power dissipated by the cooling fluid is observed, and the difference between this and the total power delivered to the valves of the output stage gives the radio-frequency power delivered by the transmitter into the output circuit and load. The loss in the output circuit is measured and subtracted thus giving the power delivered to the load.

7.3 Modulation Percentage or Depth. In all instruments for modulation measurement, other than cathode-ray instruments, the first step is the reduction of the radio-frequency voltage under investigation to a D.C. voltage. Audio-frequency variations in the radio-frequency voltage are then represented by corresponding audio-frequency variations in the D.C. voltage. This function is carried out by a diode rectifier, as in Fig. 48.

/ Fig. 48



INPUT CIRCUIT AND DETECTOR.

FIG. 48.

Certain conditions must be observed -

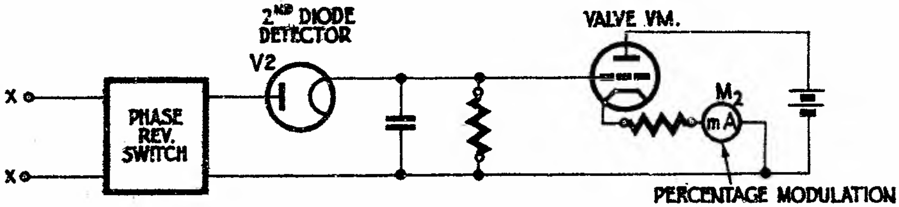
(i) The response of the input circuit must be uniform over the sideband range of the wave under investigation (± 20 or 30 kc/s with respect to the carrier frequency when harmonic distortion measurements are being made). A suitable value of resistance in parallel with the tuning capacitor allows this condition to be fulfilled over a wide range of carrier frequencies, even when plug-in coils are used to extend the range.

(ii) The load of the diode circuit must be equivalent to a resistance of constant value over audio-frequency range. The load resistance must, therefore, be fairly low, so that its value is not appreciably modified by the shunting effect of other components of the measuring circuits which follow. The usual value is from 5,000 to 10,000 ohms.

(iii) The output must be free from radio-frequency components. This requires a two or three stage filter, which must be correctly designed to match the terminating resistance in order that condition (ii) is fulfilled. (See Fig. 48.)

(iv) The input voltage must be high enough to ensure that the diode output is substantially proportional to the radio-frequency input. The usual value is from 50 to 200 volts.

For indicating modulation percentage, a valve voltmeter is used to indicate the peak A.C. voltage across the diode load. This has a series capacitor C1 in its input circuit to block the D.C. component (Fig. 48), and a phase reversing switch so that upward and downward peaks of modulation are indicated. (See Fig. 49.)



PERCENTAGE MODULATION INDICATOR.

FIG. 49.

The radio-frequency wave to be investigated is rectified by a diode detector V1 (see Fig. 48) operated as a linear detector. This gives pulsating D.C. output voltage that is almost an exact reproduction of the modulation envelope. The average value of this voltage gives the carrier amplitude and is read on a D.C. microammeter M1. Any change in this meter reading with variation in the degree of modulation indicates carrier shift.

/ The

The peak and trough values of modulation are obtained by separating the modulation-frequency component of the rectifier output and rectifying this audio wave by a second diode detector V2. This detector is arranged to indicate either positive or negative peaks, according to the connections. In one typical make of instrument, the voltmeter diode connections are changed. Another arrangement is to use a phase splitting stage and take the output as required from the cathode or the anode side. (See Fig. 50.)

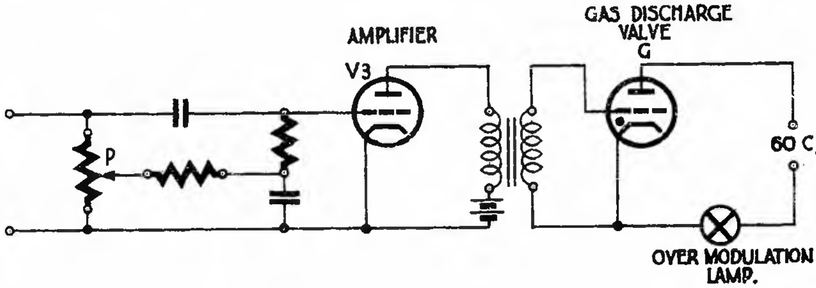


FIG. 50. PEAK INDICATOR.

The voltmeter diode develops an output voltage (across a load of several megohms) equal to the peak A.C. input.

The triode and associated meter provide an indication of this voltage. In order that the arrangement is direct reading, it is necessary that the D.C. output from the carrier diode be adjusted to a value equal to the peak value of the A.C. voltage which gives 100 per cent. reading. A meter in series with the cathode of the carrier diode carries an index mark for this purpose (meter M1 of Fig. 48). An alarm circuit is often provided for indicating modulation peaks in excess of a pre-set value.

The equipment required is shown in Fig. 50, and consists of an amplifier V3 so biased that, when the degree of modulation reaches the maximum allowable level, anode current begins to flow. This current fires a gas discharge valve G, which then causes current to pass through the lamp. The 60 c/s voltage shown would give a flashing signal more likely to attract attention. Fig. 51 shows a simplified schematic circuit of this equipment.

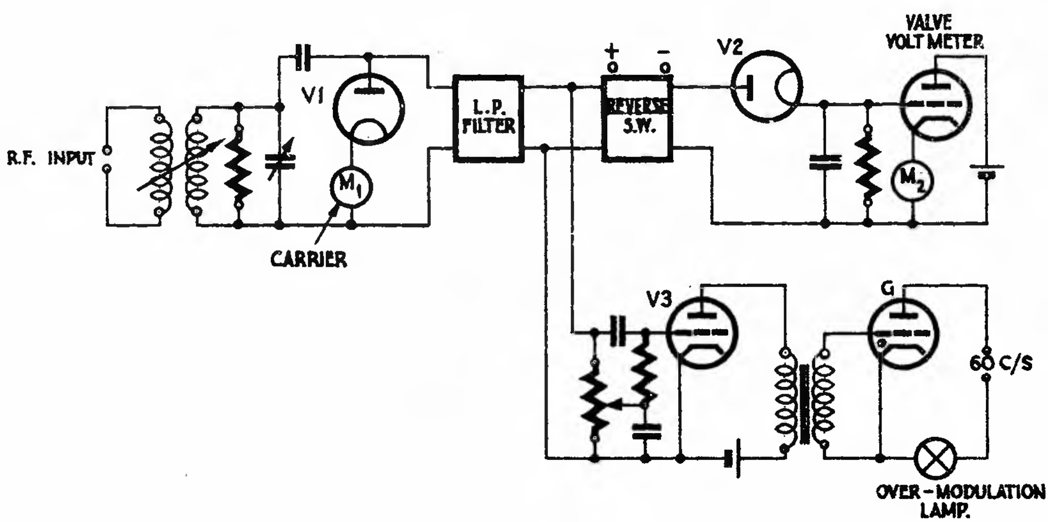
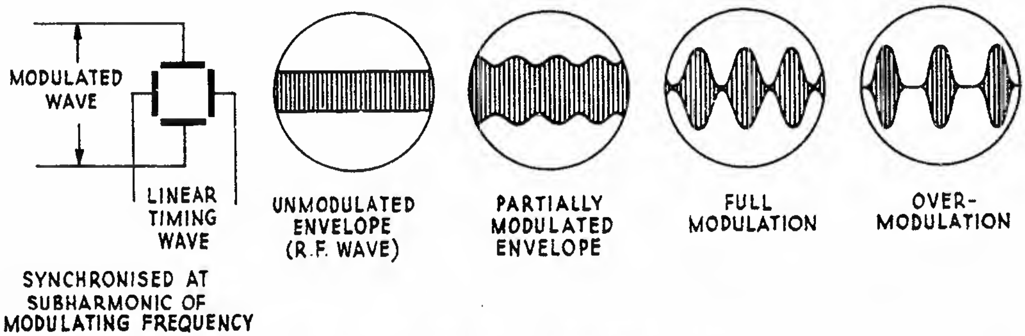


FIG. 51. BASIC SCHEMATIC CIRCUIT OF MODULATION METER.

Cathode-Ray Oscilloscope Method. The Cathode-Ray Oscilloscope offers a simple means of checking modulation depth. The results are, in general, less accurate than other methods, but they are conveniently obtained and the equipment is less expensive. There are several ways of determining the modulation depth by the cathode-ray oscilloscope.

The peak modulation is readily observed by applying the wave to be investigated to the vertical deflectors of the tube. The length of line that results with no modulation corresponds to twice the carrier amplitude, while the length of line with modulation corresponds to twice the peak amplitude of the modulation envelope.

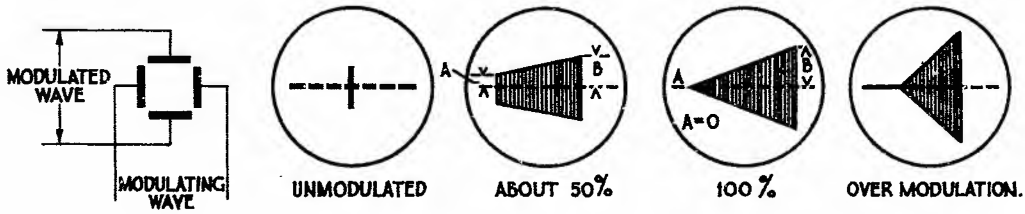
A very satisfactory method, however, is to apply the modulated carrier to the vertical deflector plates, and adjust the "timing" frequency of the sweep oscillator to a sub-harmonic of the modulation frequency. In other words, when 1,000 c/s is the modulating frequency, then a sweep frequency of 500 c/s causes 2 audio cycles to appear on the screen; a sweep of 250 c/s gives 4 audio cycles, etc. When the cathode-ray oscilloscope is synchronised correctly, a steady pattern appears on the screen. The radio-frequency voltage is seen as a solid illuminated pattern on the screen surrounded by the audio envelope. The picture is actually a radio wave in action, and, by comparing the height of the peaks with the unmodulated envelopes, an accurate estimation of modulation percentage is obtained. Since overmodulation is the condition usually to be guarded against rather than exact evaluation of percentage modulation, this method enables a continuous picture of the transmitter performance to be obtained. Fig. 52 shows typical patterns obtained by this method.



MODULATION PATTERNS ON A CATHODE-RAY OSCILLOSCOPE.

FIG. 52.

Another method is to apply the modulated wave to the vertical deflector plates and a small portion of the audio output of the modulator to the horizontal plates, instead of the "timing" frequency. This results in the so-called "trapezoid patterns" with straight sides, from which the degree of modulation is readily calculated. Fig. 53 shows typical patterns obtained by this method, which has the disadvantage that distortion of the envelope is not observable.



$$\text{MODULATION PERCENTAGE} = \frac{B - A}{B + A} \times 100.$$

FIG. 53.

7.4 Carrier Noise. Commutator ripple, A.C. hum and other carrier noise is compared on a power basis to a single tone at 100 per cent. effective modulation, and is expressed in terms of percentage modulation or in db below 100 per cent. modulation.

The basic principle of most methods of measuring transmitter noise is the application of a tone-modulated carrier wave to a rectifier, adjusting the resultant audio-frequency level to a predetermined value, removing the modulating tone and measuring the remaining noise content of the carrier. A calibrated attenuator and suitably calibrated meter enable these measurements to be read directly. A typical circuit is shown in Fig. 54.

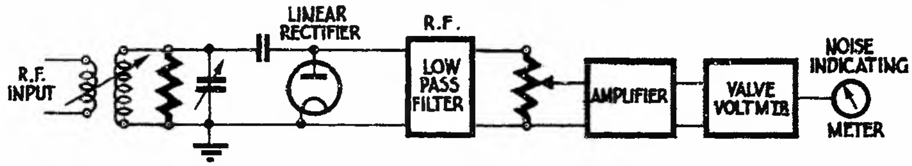


FIG. 54. TYPICAL CARRIER NOISE MEASURING SET-UP.

7.5 Distortion. Measurement of distortion in transmitters is made by modulating the carrier with a sinusoidal tone to the modulation depth desired, rectifying the signal by a linear rectifier and measuring the resultant audio component. Total distortion is the general measurement made, but a wave analyser is used when it is desired to determine individual harmonic values. Fig. 55 shows a method of making this measurement.

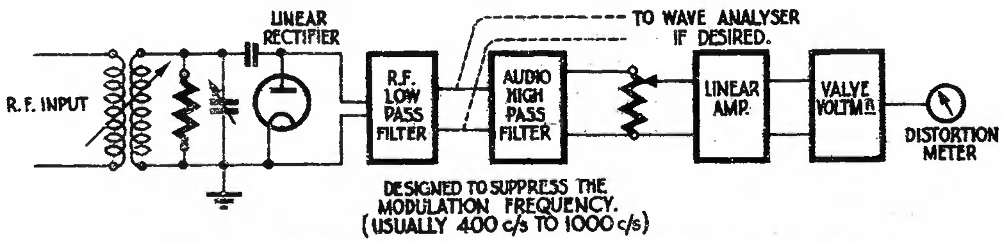


FIG. 55. TYPICAL TRANSMITTER-DISTORTION MEASURING SET-UP.

7.6 Spurious Radiations. Any radiation from a transmitter, the frequency of which is outside the communication band of the transmitter (that is, the normal band-width), is considered spurious. The following are representative of these radiations -

(i) Key Clicks.
(ii) Parasitic Oscillations.
(iii) Harmonics and Subharmonics.
(iv) Higher-Order Modulation Products.

(i) Key Clicks. Key Clicks are discussed and typical filter circuits are shown for suppressing them, in Paper No. 9 of this book.

(ii) Parasitic Oscillations. Parasitic Oscillations are oscillations whose frequency is not identical with the operating frequency. They are due to faulty design or improper operation.

(iii) Harmonics and Subharmonics. These terms refer to any component whose frequency is an integral multiple or submultiple of the carrier frequency.

Measurements of (ii) and (iii) are best made with a suitable field intensity measuring set, which covers the desired frequency spectrum. Alternatively, a suitable radio receiver having a calibrated signal-level meter serves the purpose.

Owing to the irregularity of field intensity patterns, such measurements are made at several points about the station, all at the same radial distance and approximate equal angular separation.

/ When

When the harmonic frequency is high, a measure of the intensity is not always indicative of the interference value, as it sometimes behaves as a "skywave".

A method of determining the harmonic power consists in comparing (by means of a sufficiently selective measuring set) the unknown harmonic power, which is present with the fundamental, to a known power of the same frequency, which is supplied and measured in the absence of the fundamental.

(iv) Higher-Order Modulation Products. Non-linearity in transmitters produces not only harmonics of the frequencies being amplified, but also other frequency components which may fall either inside or outside the communication band and thus cause interfering effects. These products are due to the presence in the equipment characteristic of terms of higher order than those required for normal transmission, hence the designation.

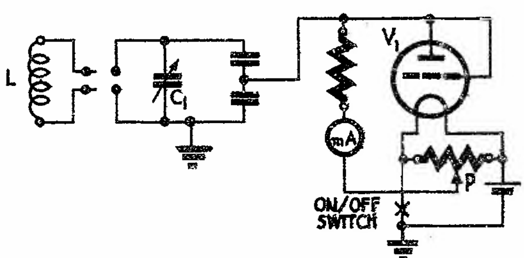
In a typical test, two equal tones, whose total amplitude is assumed to be equivalent to that of the speech, are applied simultaneously to the transmitter. The higher-order products which result in this case are steady oscillations, whose intensities are measured with the aid of a sufficiently selective radio receiver used as a frequency analyser. Their amplitudes are then expressed in terms of db below each of the fundamentals.

7.7 Frequency Stability. This term, as applied to a radio transmitter, is defined as its ability to maintain at a constant value a given carrier frequency. Frequency stability is usually expressed as a percentage deviation within which the operating frequency is maintained. Frequency meters, controlled or calibrated by standard frequency emissions, are a necessary adjunct to transmitters in some countries, but, insofar as Australia is concerned, there is maintained a frequency measuring service which regularly monitors and measures the frequency of each transmission channel originating a service in Australia. Thus, the responsibility of maintaining transmitters on their correct frequencies falls on the controllers of this service, and the installing of suitably designed frequency generating units is the responsibility of the licensee of the station.

7.8 Wavelength. As mentioned in the previous paragraph, frequency stability is of paramount importance, in the broadcast bands especially, but it is often desired to know the wavelength of an emission without requiring the precise frequency. Typical of this is the preliminary lining up of radio transmitters, checking oscillator frequency ranges, a quick search for parasitic radiations, etc., in locations where a small portable instrument is desirable.

Such an instrument is a Wavemeter, and many types of these are available with varying degrees of accuracy, etc.

In general, wavemeters consist of a tuned circuit with some form of resonance indicator and a pick-up coil.



TYPICAL WAVEMETER CIRCUIT.

FIG. 56.

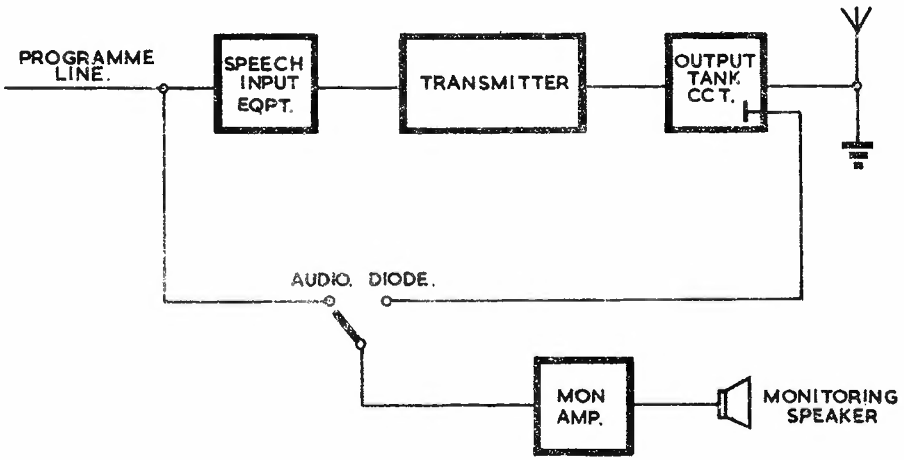
The basic circuit is shown in Fig. 56, in which L is a plug-in coil (to allow for a wide-range instrument), C1 is the tuning capacitor, and V1 is a low drain valve acting as a valve voltmeter with the meter M acting as a resonance indicator. The potentiometer P enables the meter M to be adjusted to zero as required. In operation, the circuit is tuned for maximum reading on the meter, and the wavelength read either directly off the scale or from a suitable table.

/ 7.9

7.9 Programme Monitoring Facilities. These facilities are a very essential part of broadcast station equipment. In broadcasting technique, "programme monitor audio" refers to a monitoring check on the audio signal input to the transmitter, whereas "programme monitor radio" refers to a check on the demodulated signal secured by rectification of the carrier envelope as produced at the broadcast transmitter output. By switching from one to the other, the station personnel determine by listening tests and measurements, the relative amount of distortion produced in the broadcast station equipment.

These programme monitoring facilities are usually provided in a suitably treated room, such as the control room where the speech input equipment is situated. The equipment consists of high quality audio-amplifiers and loud-speakers, with associated switching equipment.

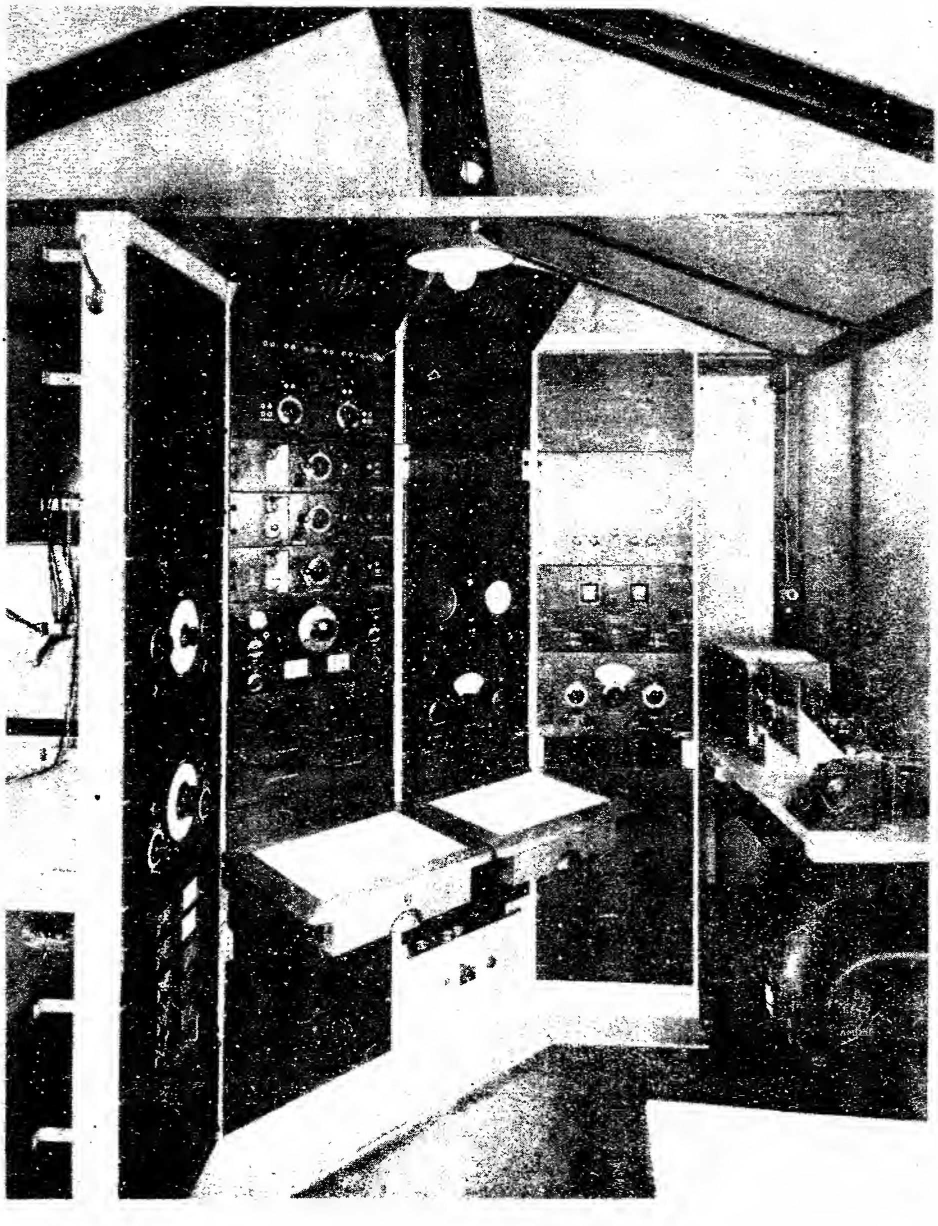
Additional to this for "radio" monitoring is a well-designed monitoring rectifier capable of demodulating the carrier signal as picked up in the output tank circuit of the transmitter or at the aerial. It is essential that this diode operates linearly to avoid the introduction of distortion. Fig. 57 is a schematic diagram of these facilities.



PROGRAMME MONITORING FACILITIES.

FIG. 57.

7.10 Frequency Response. The frequency response of a transmitter is usually measured from the studio to the output tank circuit, and is made by sending selected frequencies in the audio-frequency band at a predetermined programme level over the normal programme line. The output of the diode is measured on the normal programme monitoring equipment, and the levels of the various frequencies are compared to that of 1,000 c/s, which is the usual reference frequency. Thus, the whole programme circuit is included in this test.



FREQUENCY MEASURING EQUIPMENT. PRIMARY CENTRE. (MONT PARK).

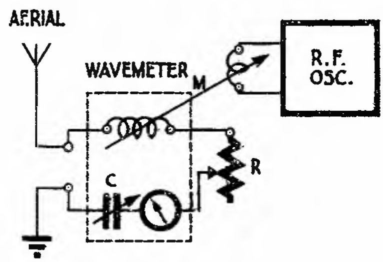
8. MEASUREMENTS ON AERIALS.

8.1 Measurements are usually made on aerials to determine the following characteristics -

Aerial Resistance.
Radiation Resistance.
Effective Height.
Resonant Frequency.
Radiation Output Power.
Radiation Efficiency.
Effective Aerial Reactance.

8.2 Aerial Resistance. There are several methods of determining the resistance of an aerial, including the resistance-variation method, reactance-variation method, substitution method and the use of a radio-frequency bridge. Two methods of making this measurement are described.

Resistance-Variation Method. Fig. 58 shows a method of making this measurement, the apparatus being -



RESISTANCE-VARIATION MEASUREMENT OF AERIAL RESISTANCE.

FIG. 58.

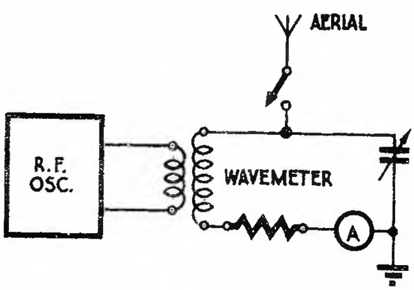
(i) A radio-frequency oscillator covering the frequency range of the transmitter.

(ii) A non-inductive variable resistance network R, such as a decade box.

(iii) A wavemeter of good accuracy having a resonance indicator in the form of a meter.

With R at zero resistance, set oscillator at desired frequency. Tune wavemeter by capacitor C for maximum reading on meter (keep this about 1/2 scale by adjusting coupling M). Add resistance in R until meter reading is half of the previous reading. The aerial resistance is then equal to the amount of added resistance. The usual precautions regarding stray couplings, capacitances, etc., must be observed in making this measurement.

Substitution Method. Fig. 59 shows the equipment used for this method of measuring aerial resistance.



SUBSTITUTION METHOD OF MEASURING AERIAL RESISTANCE.

FIG. 59.

The wavemeter is first grounded, with aerial disconnected, and reading for maximum indication noted. The aerial is then connected, and the meter retuned for maximum indication.

Finally, calibrated reactances and resistances are added to the wavemeter circuit in parallel until the same difference in readings is observed, indicating the value of resistance and reactance of the aerial by substitution. This method is suitable for high, as well as low, aerial resistance values.

/ For

For this test the following precautions are necessary -

(i) The oscillator must possess sufficient power and regulation to be unaffected by the wavemeter.
(ii) The standards of impedance and resistance used for substitution must be essentially free from stray capacitance.
(iii) The readings and adjustments must be made with care.

8.3 Radiation Resistance. It is not practicable to measure radiation resistance directly, but, for aerials up to $\frac{\lambda}{4}$, the following formula gives reasonably good accuracy -

$$Rr = 1579 \left(\frac{h}{\lambda}\right)^2$$

where λ = Wavelength in metres,
h = Effective height in metres, and
Rr = Radiation resistance.

8.4 Effective height. This term is not ordinarily used with other than low and medium-frequency aerials.

Effective height is measured using the aerial as a radiator. The vertical field ϵ in microvolts per metre is measured at distance d kilometres. The effective height h is then obtained from the formula -

$$h = \frac{\epsilon\, d}{1.26 f I}$$

where ϵ = Microvolts per metre,
d = Distance in km.
f = Frequency in kc/s, and
I = Current at the point where power is fed to the apparatus.

8.5 Resonant Frequency. Connect the aerial directly to ground through a radio-frequency current instrument of adequate sensitivity, and couple a variable frequency oscillator lightly to the system by proximity only. Search for the lowest frequency at which resonance is indicated by a maximum current.

8.6 Radiation Output Power. The power output or radiation output of the aerial system is calculated from the radiation resistance and the aerial current by the usual formula -

$$Pr = I^2 Rr$$

where Rr is as discussed in paragraph 8.3. A more direct measurement is impracticable.

8.7 Radiation Efficiency. The radiation efficiency of an aerial is often defined and calculated by the formula -

$$\text{Radiation efficiency} = \frac{Pr}{Pi}$$

where Pr is as in paragraph 8.6.
and Pi = total power in aerial.

Aerial efficiencies are often compared, however, by the field strength at one mile or km for 1 kW, as mentioned in Paper No. 11 of Radio II.

8.8 Effective Aerial Reactance. When not measured directly with a radio-frequency impedance bridge, aerial reactance is measured by resonating the aerial at the desired frequency, using a calibrated inductor or capacitor. At resonance, the aerial reactance is equal and opposite in sign to that of the tuning device.

9. MISCELLANEOUS ITEMS.

9.1 Transmission Line Impedance. There are diverse methods of measuring the characteristic impedance of a transmission line. A simple but effective method is shown in Fig. 60.

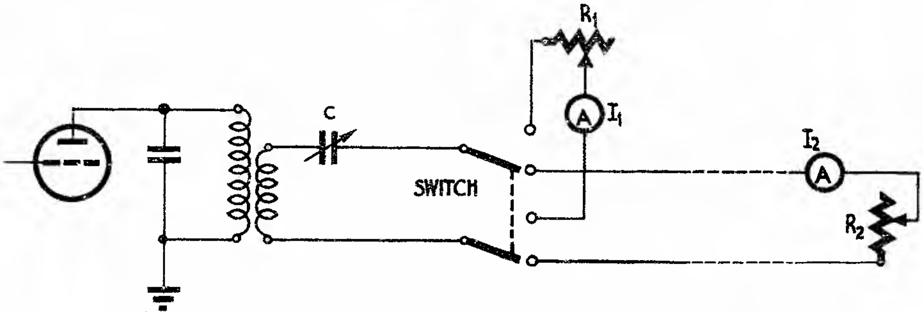


FIG. 60. MEASUREMENT OF SURGE IMPEDANCE OF TRANSMISSION LINE.

With this set-up and the switch thrown to the line position, a trial value of R2 is inserted. C is adjusted for maximum I2. Then, with the switch thrown in the opposite position and R1 set to equal R2, the capacitor C is adjusted for maximum I1. By trial, a combination is found where there is a maximum value of I1 and I2 for the same setting of C with R1 equal to R2. This value of R is the characteristic or surge impedance of the line.

9.2 Harmonics in Aerials. For determining harmonic components of small amplitude in aerials the following substitution method is useful (see Fig. 61).

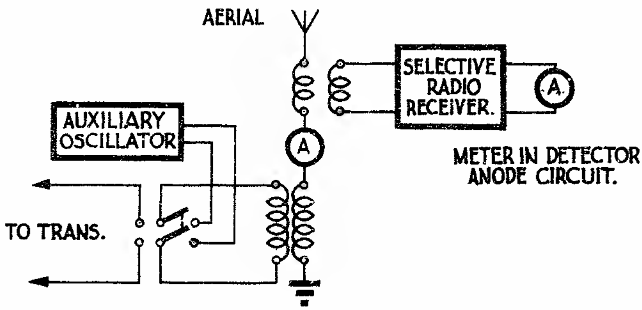


FIG. 61. SUBSTITUTION METHOD FOR AERIAL HARMONIC COMPONENTS.

A highly selective properly shielded radio receiver is coupled to the aerial system, tuned to the desired harmonic frequency, and the gain adjusted until a reasonable deflection is observed in the detector-anode circuit meter. The radio transmitter is now tuned off, and current from an auxiliary oscillator is passed through the coil. The amplitude of the auxiliary oscillator current, which gives the same detector-anode current, is then equal to the amplitude of the unknown current. The auxiliary oscillator current is readily measured, as it is not combined with currents of other frequencies.

9.3 Aerial and Line Impedances. Fig. 62 shows a typical set-up for measuring the impedance of aerials and transmission lines by the substitution method. The circuit consists of an oscillator with which is associated a calibrated capacitor C, a resistance box R for high-frequency service and a wavemeter. A sensitive meter is placed in the anode circuit and provided with a balancing arrangement for balancing out normal anode current, so that greater accuracy is obtained in reading small changes.

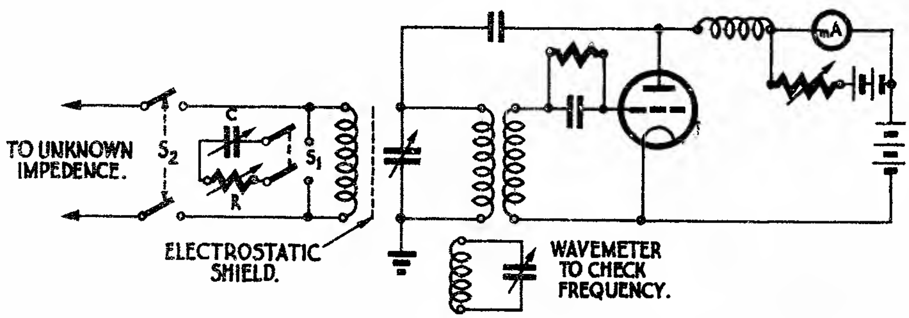


FIG. 62. MEASUREMENT OF AERIAL OR TRANSMISSION LINE IMPEDANCE BY SUBSTITUTION METHOD.

Procedure. When impedance is capacitive -

(i) Close S2, open S1.
(ii) Adjust oscillator to proper frequency with wavemeter.
(iii) Balance out D.C. anode current.
(iv) Open S2, close S1.
(v) Adjust C until original frequency is restored.
(vi) Adjust R until original anode current is obtained.

Unknown impedance is equivalent to C and R in series.

When impedance is inductive -

(i) Close S1 and S2 and set R to zero.
(ii) Adjust for proper frequency as before.
(iii) Open S2.
(iv) Adjust C for original frequency.
(v) Adjust R for original anode current.

$$\text{Unknown Reactance} = \frac{1}{\omega\Delta C},$$

$$\text{and Series Resistance of Unknown} = R\left(\frac{C}{\Delta C}\right)^2,$$

where ΔC is change in capacitance, and C is the capacitance after opening S2.

9.4 Measurement of Modulation Percentage. There is a method of determining modulation percentage by observing the rise in line current or circulating current in the modulated amplifier stage, provided that the load Z is non-reactive for determining the real power. It is first necessary to prepare a graph or table relating line or circulating current readings to percentage modulation, using a sinusoidal modulating voltage for this purpose. Once this graph or table has been prepared, it is only a simple matter to determine the modulation depth by rating the current and referring to the graph.

This method is based on the following relationship between the average modulated current and the unmodulated carrier current values -

$$\frac{Im}{Io} = \sqrt{1 + \frac{M^2}{2}}$$

where Im = Average modulated current,
Io = Average unmodulated current, and
M = Modulation factor = $\frac{K}{100}$ where K = percentage modulation.

For the initial calculation, let Io equal 1 ampere and find the ratio $\frac{Im}{Io}$ for a sufficient number of values of K to prepare Table "A".

Modulation Percentage K	Modulation Factor M	Ratio $\frac{Im}{Io}$
10 per cent.	0.1	1.0025
20 " "	0.2	1.0100
30 " "	0.3	1.0222
40 " "	0.4	1.0392
50 " "	0.5	1.0607
60 " "	0.6	1.0863
70 " "	0.7	1.1158
80 " "	0.8	1.1489
90 " "	0.9	1.1853
100 " "	1.0	1.2247

For example -

when M = 0.1, $\frac{Im}{Io} = \sqrt{1 + \frac{0.1^2}{2}} = \sqrt{1.005} = 1.0025$

when M = 0.3, $\frac{Im}{Io} = \sqrt{1 + \frac{0.3^2}{2}} = \sqrt{1.045} = 1.022$

and when M = 1.0, $\frac{Im}{Io} = \sqrt{1 + \frac{1}{2}} = \sqrt{1.5} = 1.225$ etc.

TABLE "A".

The value of Io (unmodulated carrier current) for the particular transmitter in use will be known, and it is multiplied by the ratios found above to give the line or circulating currents for the particular transmitter. These values are then plotted against modulation percentage, and the graph is ready for use as in Fig. 63.

Any significant departures from normal readings are considered as indicating the existence of distortion or other faulty condition, and a further detailed investigation must be carried out, using the cathode-ray oscilloscope or other check.

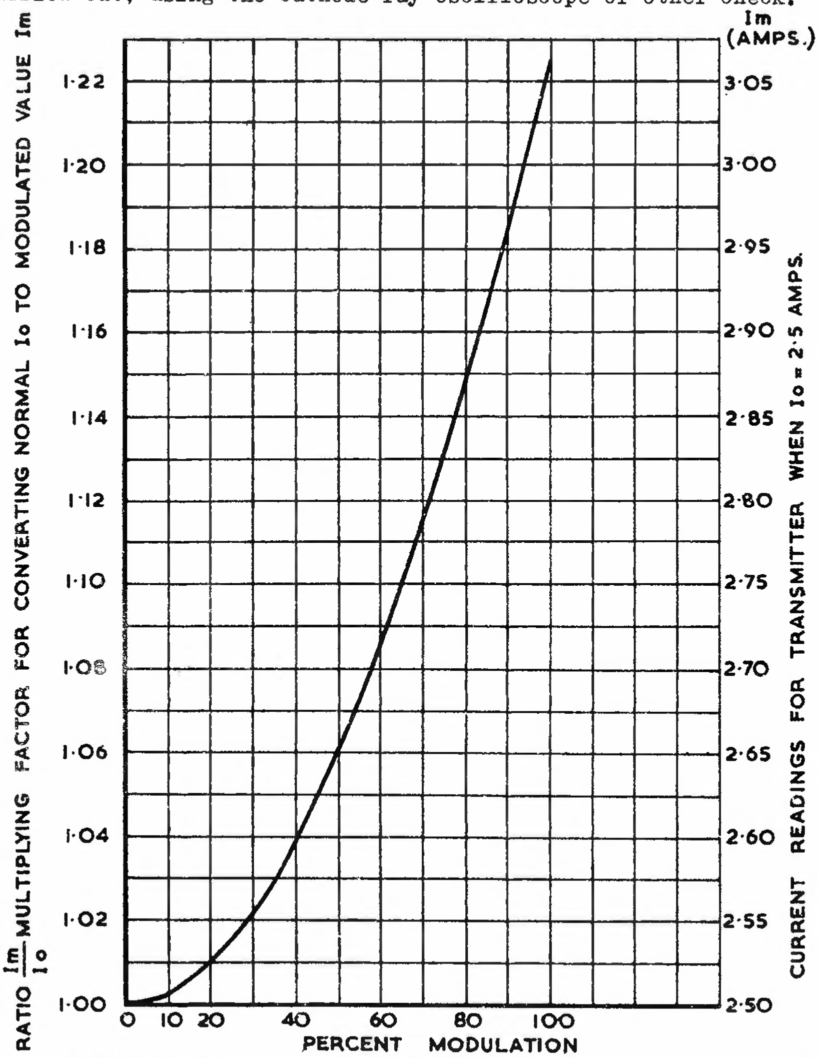


FIG. 63. ILLUSTRATING THE RISE IN MODULATED CURRENT WITH MODULATION PERCENTAGE.

The process of deriving the given formula is as follows.

The equation of a sinusoidally-modulated wave is shown to be -

$$i = Io \sin 2\pi f1t + M\frac{Io}{2} \cos 2\pi(f1 - f2)t - M\frac{Io}{2} \cos 2\pi(f1 + f2)t$$

(Carrier) (Lower Sideband) (Upper Sideband)

where i = Instantaneous value of current,
Io = Unmodulated value of current,
f1 = Carrier frequency,
f2 = Modulation frequency, and
M = Modulation percentage factor = $\frac{K}{100}$ where K = percentage modulation.

/ It

It is noticed from the above that -

(i) The carrier amplitude, Io, is independent of modulating frequency or percentage.

(ii) The sideband amplitudes are equal to half the carrier amplitude when $M = 1$ (100 per cent. modulation), and they are proportionately less as the modulation percentage decreases. Thus, the maximum total sideband power is one-half of the carrier power.

For example, when load impedance is Z then -

$$\text{Upper Sideband Power} = \left(\frac{MIo}{2}\right)^2 Z = \frac{M^2Io^2}{4} Z$$

$$\text{Lower Sideband Power} = \left(\frac{MIo}{2}\right)^2 Z = \frac{M^2Io^2}{4} Z$$

$$\text{Total Sideband Power} = \left(\frac{M^2Io^2}{4}\right) Z + \frac{M^2Io^2}{4} Z$$

$$= \left(\frac{M^2Io^2}{2}\right) Z \text{ or } M^2 \frac{Io^2}{2} Z \quad \ldots\ldots\ldots\ldots (1)$$

The sideband power is also proportional to the square of the modulation factor M.

By definition, Io is the unmodulated carrier current, thus (1) is rewritten -

$$\text{Total Sideband Power} = \frac{1}{2} M^2 \, Pc \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (2)$$

$$\text{where } Pc = \text{Carrier Power} = Io^2 Z$$

Letting Pa = Average modulated power (Im^2Z),

Pc = Average carrier power (Io^2Z),

Ps = Average sideband power, and

Im = Average modulated current,

$$\text{then } Pa = Pc + Ps$$

$$= Pc + \frac{1}{2} M^2 \, Pc \text{ (from (2))}$$

$$= Pc \left(1 + \frac{1}{2} M^2\right)$$

$$\text{from which } Im^2 Z = Io^2 Z \left(1 + \frac{1}{2} M^2\right).$$

$$\text{Dividing by Z, } Im^2 = Io^2 \left(1 + \frac{1}{2} M^2\right)$$

$$\text{and } \frac{Im^2}{Io^2} = 1 + \frac{1}{2} M^2$$

$$\therefore \frac{Im}{Io} = \sqrt{1 + \frac{M^2}{2}}$$

/ This

This is the ratio $\frac{\text{average modulated current}}{\text{average unmodulated current}}$ which is the formula required.

In Fig. 63, an example is shown for a case where Io (normal unmodulated current) is 2.5 amperes, that is -

For 50 per cent. modulation, the meter reads 2.655 amperes,

For 80 per cent. modulation, the meter reads 2.875 amperes, and

For 100 per cent. modulation, the meter reads 3.062 amperes.

10. TEST QUESTIONS.

1. Name three types of meters used in making A.C. measurements at relatively low frequencies, and describe one of the types selected.

2. It is desired to adapt a certain D.C. meter to measure 45 amperes D.C. and also, by suitable switching, to read 60 volts D.C. If the resistance of the meter is 200 ohms and full scale deflection is obtained when 5 mA is passed through, what values of shunt and multiplier are required to obtain the desired results?

3. Describe a meter suitable for making current and voltage measurements of good accuracy at frequencies between 1 kc/s and 30 Mc/s.

4. Explain, and illustrate by a sketch, the principles of the Wheatstone Bridge for measuring resistance.

5. Describe a bridge suitable for measuring inductance at audio frequencies, and briefly detail the operation.

6. Under what conditions is the power measured in an A.C. circuit by means of an ammeter and voltmeter reasonably accurate?

7. Detail a method of measuring the gain of an amplifier and state precautions to be observed.

8. Illustrate a simple method of measuring the approximate amplification factor of a valve and briefly describe the procedure. Which method is used when greater accuracy is desired? (Detailed answer not required here.)

9. What principal characteristics of audio-frequency waves are usually measured to determine the performance of amplifiers, etc. Give a block schematic diagram of the set-up for measuring <u>one</u> of the characteristics referred to.

10. Describe one method of measuring the power output of a low-power transmitter.

11. What are the more important characteristics of a high-quality broadcast transmitter.

12. Describe one method of measuring the resistance of an aerial system, and state the general precautions which are adopted when making such measurements.

END OF PAPER.