CHAPTER 6

Valve Amplifiers

D.C. Amplifier. Suppose some such device as a photocell, thermocouple or glass-electrode for pH measurements produces a small direct current of the order of 1 μA., which it is required to amplify so that it can be registered on a milliammeter, or made to actuate a circuit requiring greater power than the device itself yields. A basic circuit arrangement whereby this can be arranged is indicated in fig. 48, where a triode valve is used as a D.C. amplifier.

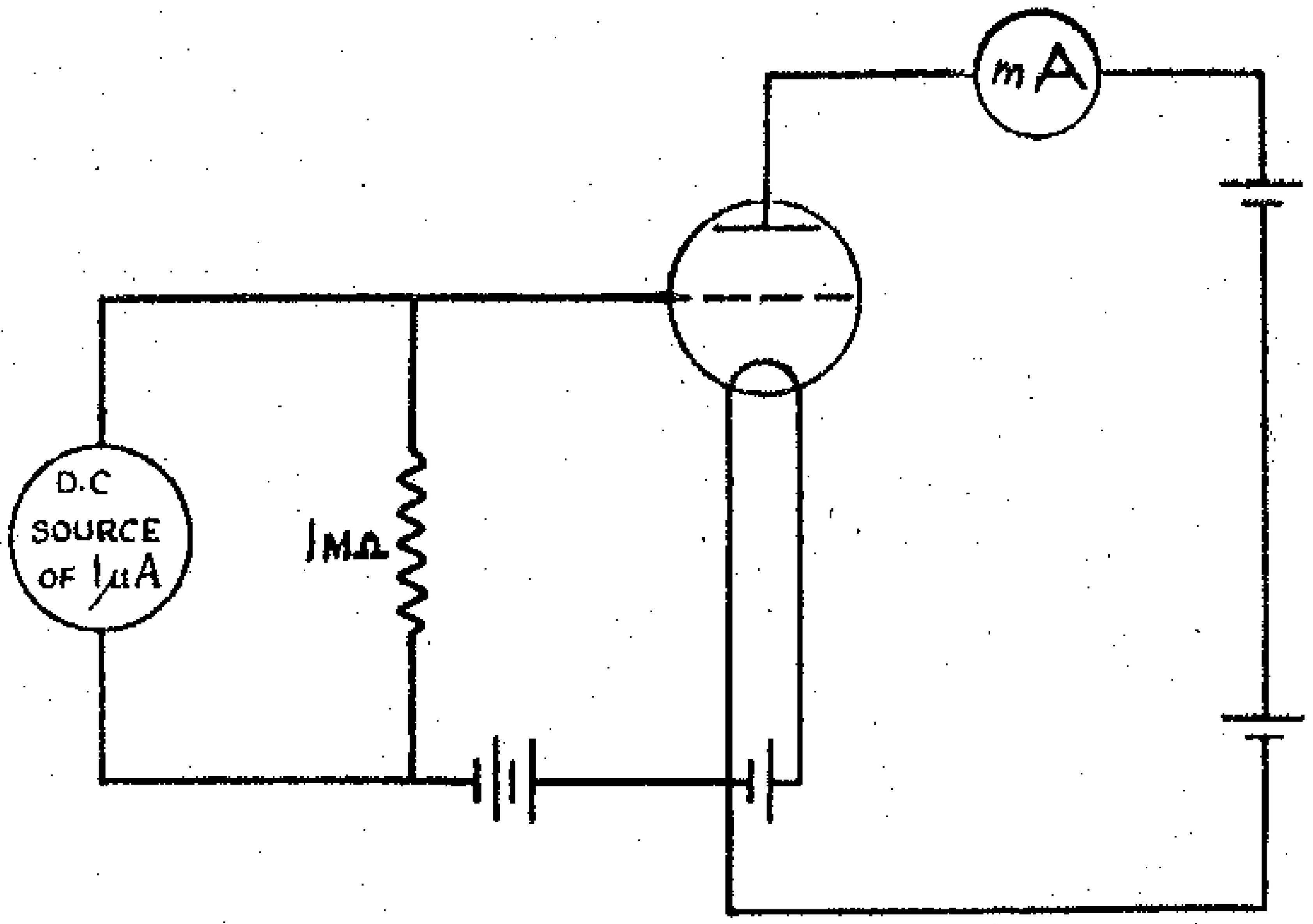


FIG. 48. Basic D.C. Amplifier Circuit.

The small current available is caused to pass through a resistance of, say, 1 MΩ. Thus a current of 1 μA. passed through 1 MΩ. will produce a P.D. of 1 V. This P.D. is placed across the grid-cathode of a triode valve in series with a negative grid bias. The negative grid bias is large enough to prevent the grid going positive, and so drawing grid current on application of the D.C. input. The input resistance of the triode is therefore high, of the order of 5 to 8 MΩ. for an ordinary triode valve. Positive ion current and electrical leakage prevents the input resistance of a normal valve from being higher, though these disadvantages can be overcome using a special type of electrometer triode.

The great virtue of the triode valve as a D.C. amplifier lies in its high input resistance, since it enables an E.M.F. input of high internal resistance and low current to be applied without damping, i.e. the high resistance is not reduced by a lower resistance due to the valve being put in parallel with it. Even so an input E.M.F. source of resistance greater than 2 MΩ. cannot be effectively employed using a standard triode in a normal circuit.

If the triode of the circuit has a mutual conductance of, say, 2 mA./V. at the anode potential used, then the anode current is changed, by the 1 V. P.D. applied to the grid, by 2 mA. Hence a device giving a current of only 1 μA. has given rise to an anode current change of 2 mA. This is a D.C. current amplification of 2 mA./1 μA., or 2000 times. Likewise, if the H.T. battery potential is raised so as to accommodate in the valve anode circuit a series resistance of 50 kΩ., and yet achieve the same anode potential as before, despite the D.C. volts drop in this anode load, then the change of 2 mA. across 50 kΩ. produces a voltage change across the load of $2/1000 \times 50{,}000 = 100$ V. A grid potential change of 1 V. has then given rise to an anode potential change of 100 V.; a D.C. voltage amplification effect of 100 times is achieved.

Alternating Voltage Amplifier. Class A. A common requirement, especially in radio receivers, is that a small alternating potential should be amplified, without distortion of wave-form, to give rise to a larger alternating potential. This is achieved using a triode Class A amplifier circuit.

The circuit of fig. 49*a* is used. An anode load resistance R is essential in order that a voltage variation of the anode is produced by the variation of anode current due to the grid potential changes. The valve mutual characteristic indicated in fig. 49*b* hence has to be the so-called dynamic characteristic, which is obtained exactly as the static characteristic, but with the necessary anode load resistance in the circuit. This load resistance will reduce the mutual conductance g_m of the valve to a lower value g_m', and, moreover, the dynamic mutual characteristic will be more linear than the static curve. If the grid potential is increased positively, the corresponding increase of anode current causes an increased volts drop across the anode load resistance. Since the actual anode potential is the H.T. potential minus the voltage drop across the anode load, so the anode potential falls. As the grid volts are increased so the anode volts are decreased, and, vice versa, a decrease of grid volts causes an increase of anode volts.

If the grid voltage is alternating, then the anode voltage, with an anode load resistance in circuit, alternates in anti-phase. Obviously the change of 1 V. on the grid will produce an anode voltage change which affects the anode current in the reverse direction to the effect of the grid volts change. In other words the static mutual conductance is reduced to the dynamic mutual conductance.

In operating the amplifier without distortion, Class A conditions are observed:

(1) The straight portion only of the dynamic mutual characteristic is utilised, i.e. the input alternating potential to the grid must produce grid voltage changes which are such that the anode current variations are in direct proportion.

(2) The grid must not be driven positive, otherwise grid current will flow, and power be drawn from the input source, causing damping.

Thus if the input voltage has a R.M.S. value of 3 V., then the positive peak input volts is $3\sqrt{2}=4{\cdot}2$ V. The negative steady grid bias used must therefore be at least $-4{\cdot}2$ V., so that the alternating plus direct inputs in series do not give rise to a positive grid.

In accordance with requirements (1) and (2), the sinusoidal wave-form representing the input to the circuit of fig. 49*a* is accommodated under the negative straight portion of the dynamic characteristic of the chart, fig. 49*b*.

If the R.M.S. input voltage $=E_G$, then the R.M.S. anode current variation produced in series with the steady anode current is $g_m' \, . \, E_G$, where g_m' is the dynamic mutual conductance of the valve. Consequently, the R.M.S. value of the alternating voltage across the anode load R, is $g_m' \, . \, E_G R$.

The voltage amplification factor (V.A.F.) or stage-gain (m) of the circuit is defined as

$$\frac{\text{Output R.M.S. voltage across anode load}}{\text{Input R.M.S. voltage to grid}}.$$

$$\therefore \quad m=g_m'\,R.$$

This is not the most convenient stage-gain formula to use in practice, since g_m' is not known unless it is obtained by plotting

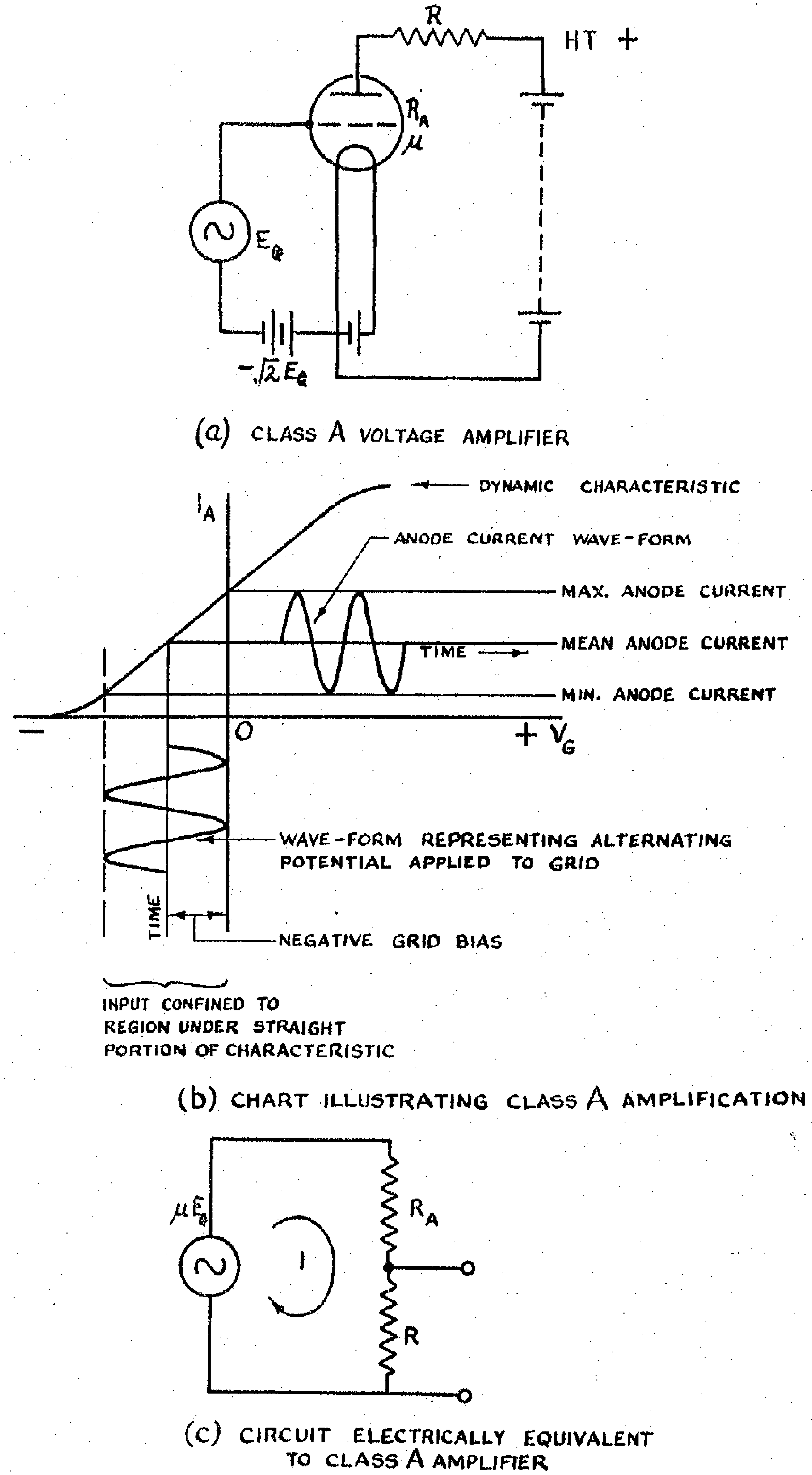


FIG. 49. *a*, Class A Voltage Amplifier. *b*, Chart illustrating Class A Amplification. *c*, Circuit Electrically Equivalent to Class A Amplifier.

the mutual characteristic for the valve with the anode load to be used.

Stage-Gain Formula. A circuit which is electrically equivalent to that of fig. 49*a* is shown in fig. 49*c*.

The valve is dispensed with in this circuit on the assumption that the change of potential E_G on the grid is equivalent to an anode voltage change of μE_G, by definition of the amplification factor, μ. Across μE_G there is effectively the anode load R, and the A.C. resistance of the valve R_A, in series. Since linear operation of the circuit is pre-arranged, so Ohm's law is applicable, and the R.M.S. value of the series current in the circuit is

$$I = \mu E_G/(R+R_A).$$

Correspondingly, there appears across the anode load R, a voltage variation of R.M.S. value given by IR, equal to $\mu E_G \,.\, R/(R+R_A)$.

The gain m is therefore

$$\frac{\mu E_G \,.\, R}{R+R_A} \div E_G = \frac{\mu R}{R+R_A}. \qquad (166)$$

Selection of Anode Load Resistance Value. In fig. 50 a curve is plotted of gain vs. anode load resistance R for a particular valve.

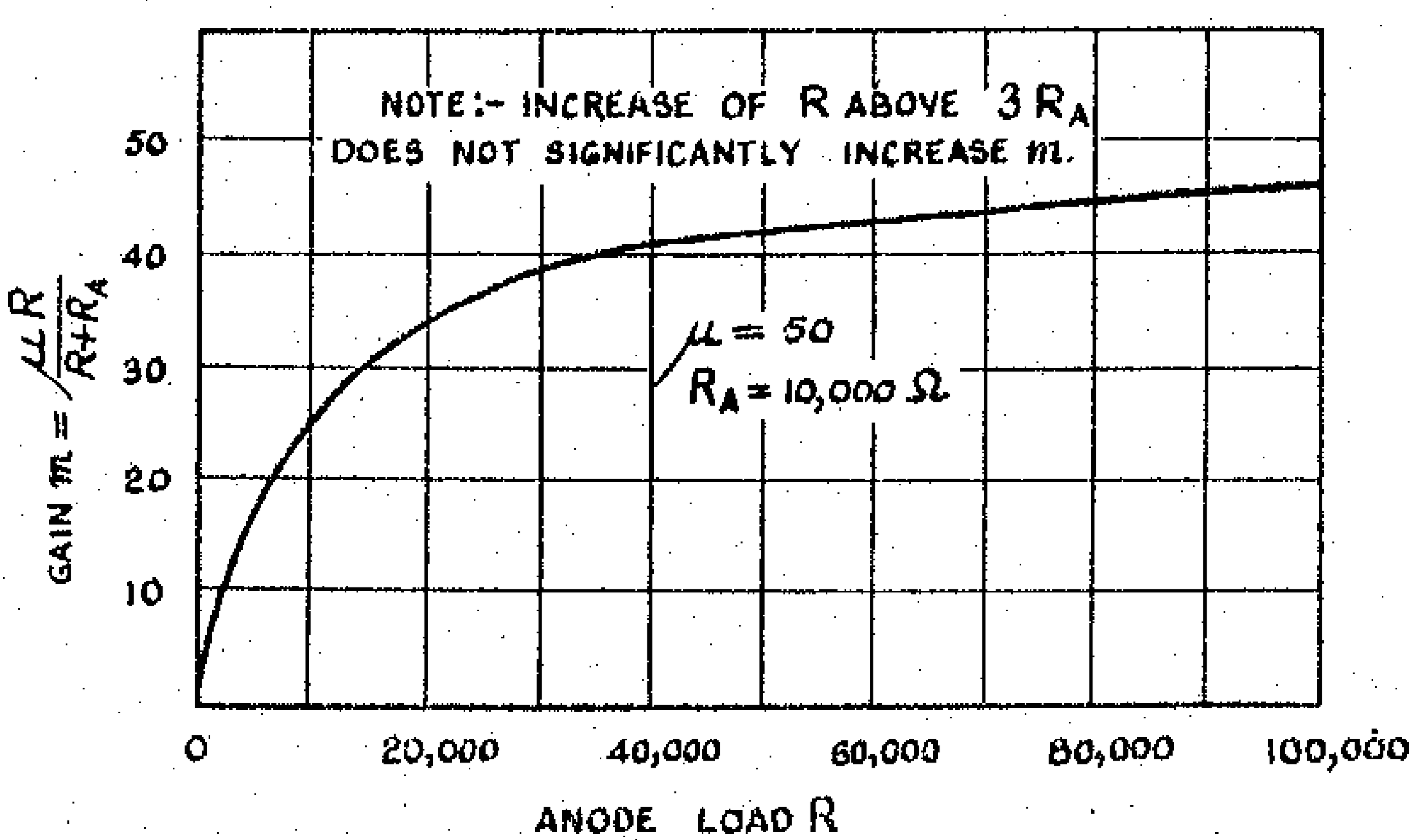


FIG. 50. Graph of Gain vs. Anode Load R, for Class A Amplifier.

As R is increased, so the gain increases to a maximum value of μ, the valve amplification factor. It would therefore seem desirable to use as high a value for R as possible. Against this consideration, however, it must be realised that the presence of a great value for R will introduce an excessively large D.C. volts drop between the H.T. supply and the valve anode. This voltage drop is inevitable using a resistance load because of the necessary direct anode current through the valve. In practice, therefore, a compromise is adopted whereby to obtain the maximum output alternating voltage without using an excessive H.T. supply, the anode load resistance is chosen to be three to four times the valve A.C. resistance. From the curve of fig. 50 it can be readily seen that this does not produce any serious loss of gain. In such circumstances the valve steady anode potential will be about one-third of the full H.T. supply volts.

Multi-stage D.C. Amplifiers. The circuit of fig. 48 can be extended to obtain another stage of amplification by using a second valve to amplify the output of the first. Thus a direct current or voltage amplification of, say, 100 times can be obtained from each stage, giving a total amplification of 100^2, or 10,000 times. Indeed, more stages can be added so that a very small D.C. voltage input to the first valve can be magnified by as much as a million times: an input of 1 μV. can give rise to an output of 1 V. Theoretically, there is no limit to the total amplification obtainable in this manner, but a practical limit is set by "shot effect", thermal agitation "noise", and fluctuations of H.T. supply voltages (see pp. 164 and 166).

A circuit showing D.C. amplifying stages connected in cascade is shown in fig. 51*a*.

A small input D.C. voltage applied to valve V_1 grid gives rise to a larger voltage across the anode load R_1, between the points A and B. The point A is connected to the cathode of valve V_2 via the H.T. battery of voltage E_B; the point B is connected to the grid of V_2 via a grid bias battery of voltage V_{G2}. The total D.C. potential appearing across the grid to cathode of V_2 is therefore: $E_B - I_{A1}R_1 - V_{G2}$, where I_{A1} is the anode current of valve V_1.

If the valve V_2 is to operate without grid current flow, then this voltage must be zero, or somewhat negative, so the grid bias battery supplying V_{G2} must have a potential of at least $-(E_B - I_{A1}R_1)$.

The chief disadvantage of such a method of connection is, therefore, the large grid bias battery required for the second stage. Again, this amplifier exhibits a tendency to "drift", which is more pronounced the greater the amplification attempted. This drift

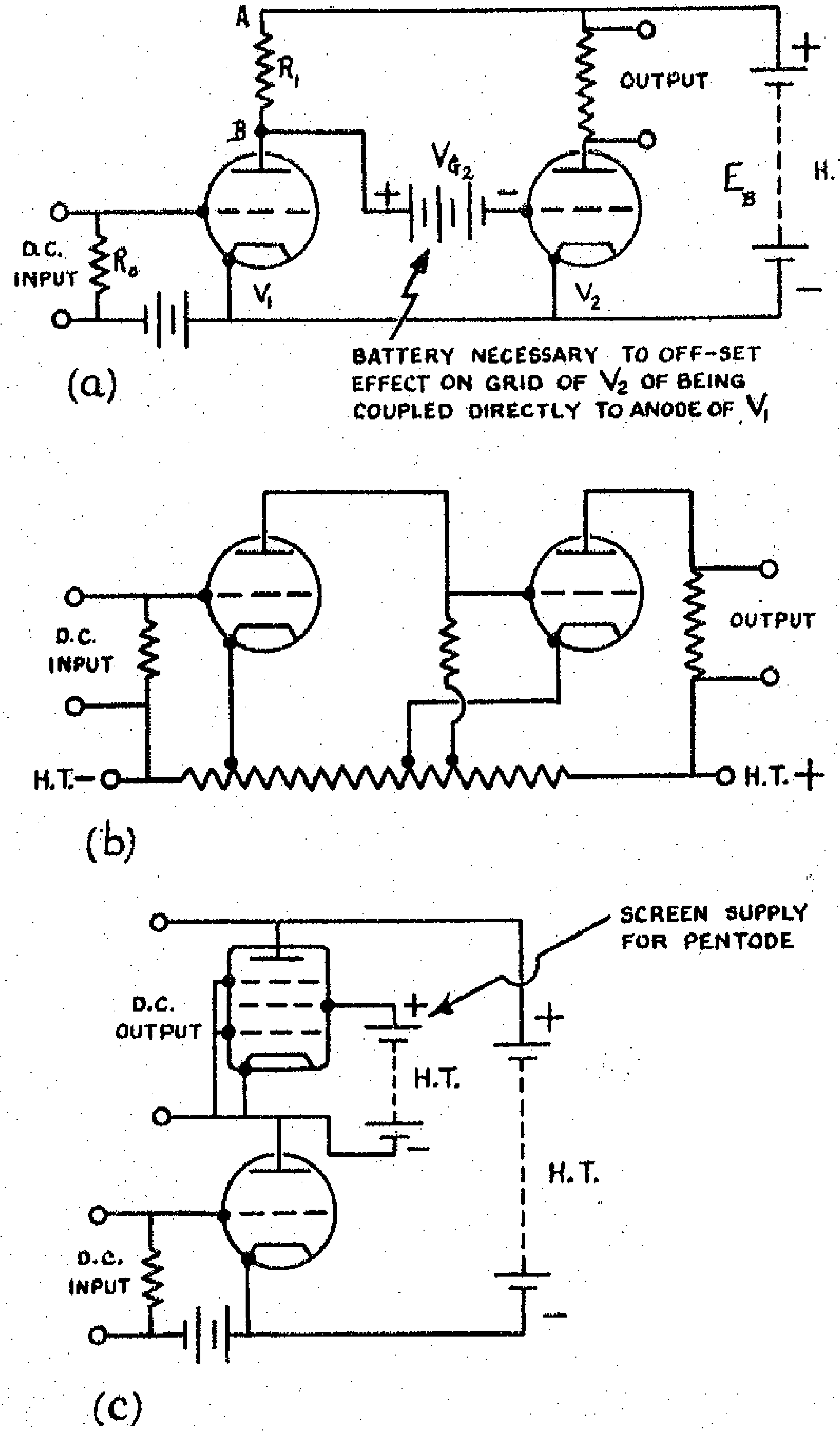


FIG. 51. *a*, Two-stage D.C. Amplifier. *b*, D.C. Amplifier of Loftin and White. *c*, D.C. Amplifier due to Horton.

is particularly prevalent since any D.C. supply potential change, anywhere in the circuit, causes the potentials in all succeeding stages to vary.

As the input across R_0 to valve V_1 increases positively, so the anode current I_{A1} increases, and the potential $(E_B - I_{A1}R_1)$ decreases, causing a drop of the anode current I_{A2} of valve V_2. If more stages are added, each succeeding valve works alternately; if the first, third and other odd stages undergo an anode current increase, then the second, fourth and other even stages suffer an anode current decrease.

Loftin and White* have developed a D.C. amplifier circuit operated by single H.T. supply of E.M.F. equal to the sum of all the grid and anode voltages. Such amplifiers can also be used for A.C. operation (fig. 51*b*).

Horton† introduced an ingenious D.C. amplifier circuit in which the high A.C. resistance of a pentode valve (see p. 139) is used as the anode load instead of the usual resistance. This provides an effective anode load resistance of one to two megohms, yet without introducing an excessive D.C. volts drop between the main H.T. supply and the anode of the amplifier valve proper (fig. 51*c*).

Multi-stage Low-frequency A.C. Amplifiers. The most widely used amplifier for the magnification of alternating voltages or currents at frequencies less than 20 kc./s. is the resistance-capacity coupled amplifier. Alternative circuits which give a greater overall amplification, but with more distortion, are the choke-capacity coupled arrangement, and the transformer coupled amplifiers of the series-fed and parallel-fed types.

The Resistance-Capacity Coupled Amplifier (R.C.C.). The discussion on p. 120 shows that the alternating voltage input to the grid of valve V_1 appears in amplified form across the anode load resistance R_1. The circuit of fig. 52 enables this voltage to be amplified again by making it the input to valve V_2. The method of connection, and a calculation of the amplification obtainable, can be appreciated by considering the equivalent electrical circuit, fig. 52*b*.

Let E_G be the R.M.S. alternating voltage input to V_1 grid, and m be the gain furnished by the first stage. The alternating voltage component across the load resistance R_1 will then be mE_G.

This alternating voltage mE_G is, in effect, supplying a circuit consisting of condenser C_G and resistance R_G in series. The fraction of this voltage appearing across R_G is applied to valve V_2 grid. The condenser C_G is essential to avoid the positive anode voltage of valve V_1 being superimposed on the grid of valve V_2.

A decision as to the component values C_G and R_G that should be adopted demands consideration of the following factors:

(*a*) As large a fraction as possible of the alternating voltage across R_1 needs to appear across R_G if the utmost possible voltage amplification is to be achieved. Since R_G and C_G in series form an

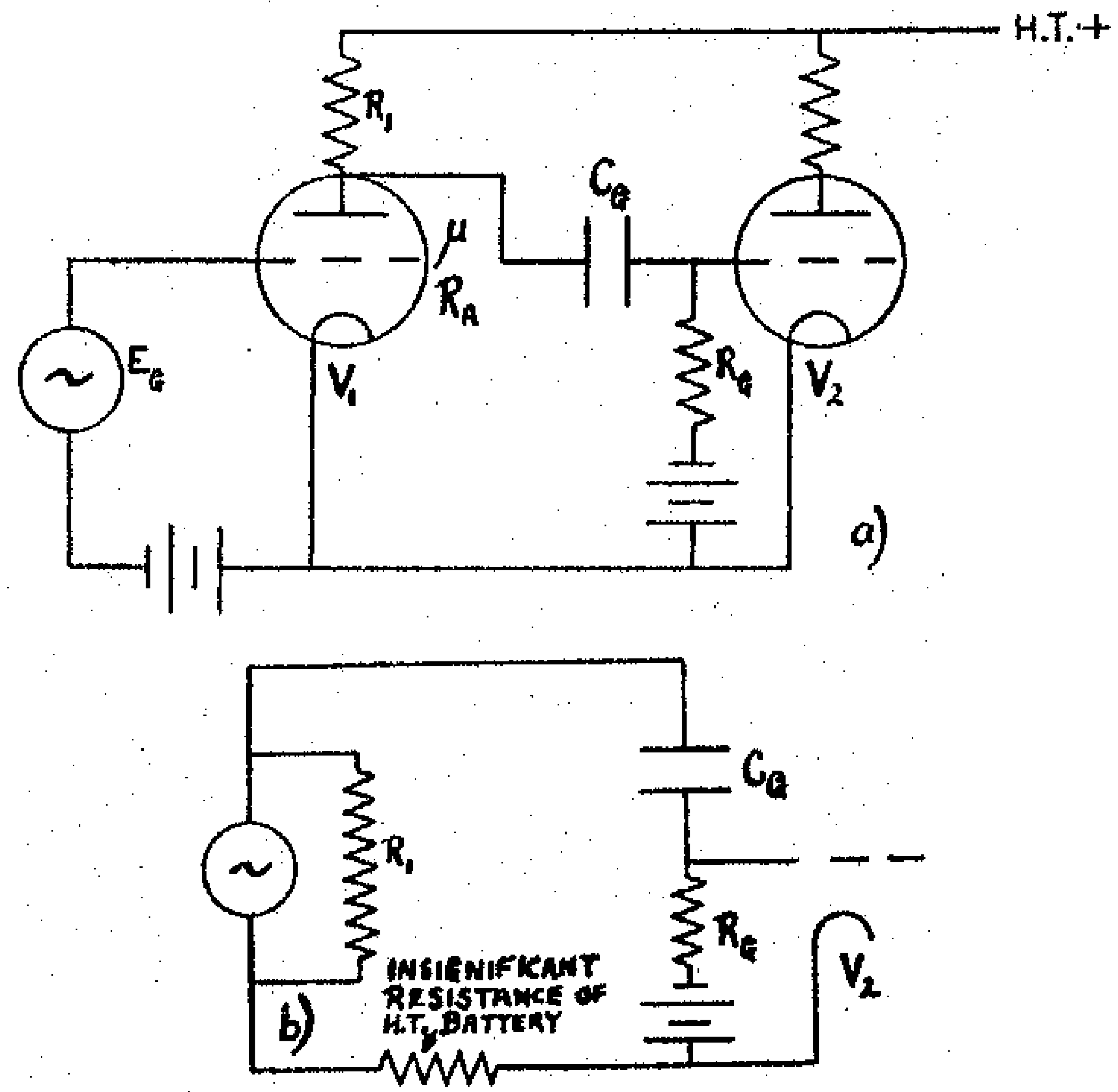


FIG. 52. *a*, Resistance-Capacity Coupled Amplifier. *b*, Electrically Equivalent Circuit.

A.C. potential divider across the supply of voltage mE_G from R_1, so the reactance of the condenser C_G needs to be small compared with the value of the resistance R_G. From the formula on p. 51, the voltage across R_G is

$$\frac{m \, . \, E_G R_G}{\sqrt{[R_G{}^2 + (1/\omega^2 C_G{}^2)]}},$$

where $\omega = 2\pi f$, f being the frequency of the alternating voltage input.

If this expression is to be as nearly as possible equal to mE_G, then $1/\omega C_G$ must be small compared with R_G, necessitating a large value of the coupling condenser C_G.

(*b*) The effective anode load resistance of valve V_1 is not R_1 alone, but less than R_1 because of the presence of the parallel circuit formed by R_G and C_G in series. If C_G is made sufficiently large in accordance with (*a*), then, to a first approximation, the anode load of V_1 may be considered as R_1 and R_G in parallel, equivalent to a resistance of $R_1R_G/(R_1+R_G)$. So m, the gain of the first stage will not be $\dfrac{\mu R_1}{R_1+R_A}$ but $\dfrac{\mu\,[R_1R_G/(R_1+R_G)]}{[R_1R_G/(R_1+R_G)]+R_A}$. In order that this actual value of gain should be as nearly as possible equal to the full gain, R_G must be large compared with R_1. A value of R_G of five to ten times R_1 is usually chosen.

(*c*) It would seem that the ideal alternating voltage amplifier is obtainable by using values of C_G and R_G as large as possible. However, there are practical limitations to the maximum values of these components which can be used. If R_G is made excessively large, say 10 MΩ., then any positive ion current,* say 1 μA., collected by the negative grid of valve V_2 will pass through the resistance R_G, producing a D.C. potential difference across it. In the case of the figures quoted, this P.D. will be 10 MΩ. × 1 μA. = 10 V., making the grid positive to this extent. Such a potential will usually overcome the effect of the negative potential brought about by the grid bias battery, so the resultant grid potential will be positive. The grid will then collect electrons; the effect will be cumulative, and excessive grid current will flow. This grid current will form a resistance in parallel with the high value of R_G selected, reducing its effective value considerably. In certain cases, the positive grid current may be so great as to overheat the grid inordinately. So values of R_G used with normal valve types rarely exceed 2 MΩ.

The value of C_G can be made large, provided that a large capacity condenser of good insulation resistance is available. If, however, as in radio communication receiver amplifiers, the audio-frequency voltage input to the amplifier varies over a frequency-range of 50 c./s. to 10,000 c./s., and also varies greatly in magnitude, then a transient powerful signal may temporarily drive the grid of valve V_2 positive, causing grid current to charge the condenser C_G. To prevent the condenser charge blocking the grid potential, the time constant C_GR_G must be sufficiently short to enable the condenser to leak away rapidly: a value of 0·005 is recommended,

* Such positive ion current is due to the ionisation of the residual gas in the valve by the electrons accelerated away from the cathode. If this ion current, as collected by a negative grid, exceeds 1 μA., then the vacuum in the valve is not good enough.

so that with $R_G = 0{\cdot}5$ MΩ., $C_G = 0{\cdot}01$ μF. An R.C. coupled amplifier operating with such component values will exhibit a reduced amplification at the lower frequencies because of the increased value of $X_{C_G} = 1/2\pi f C_G$ as f is reduced.

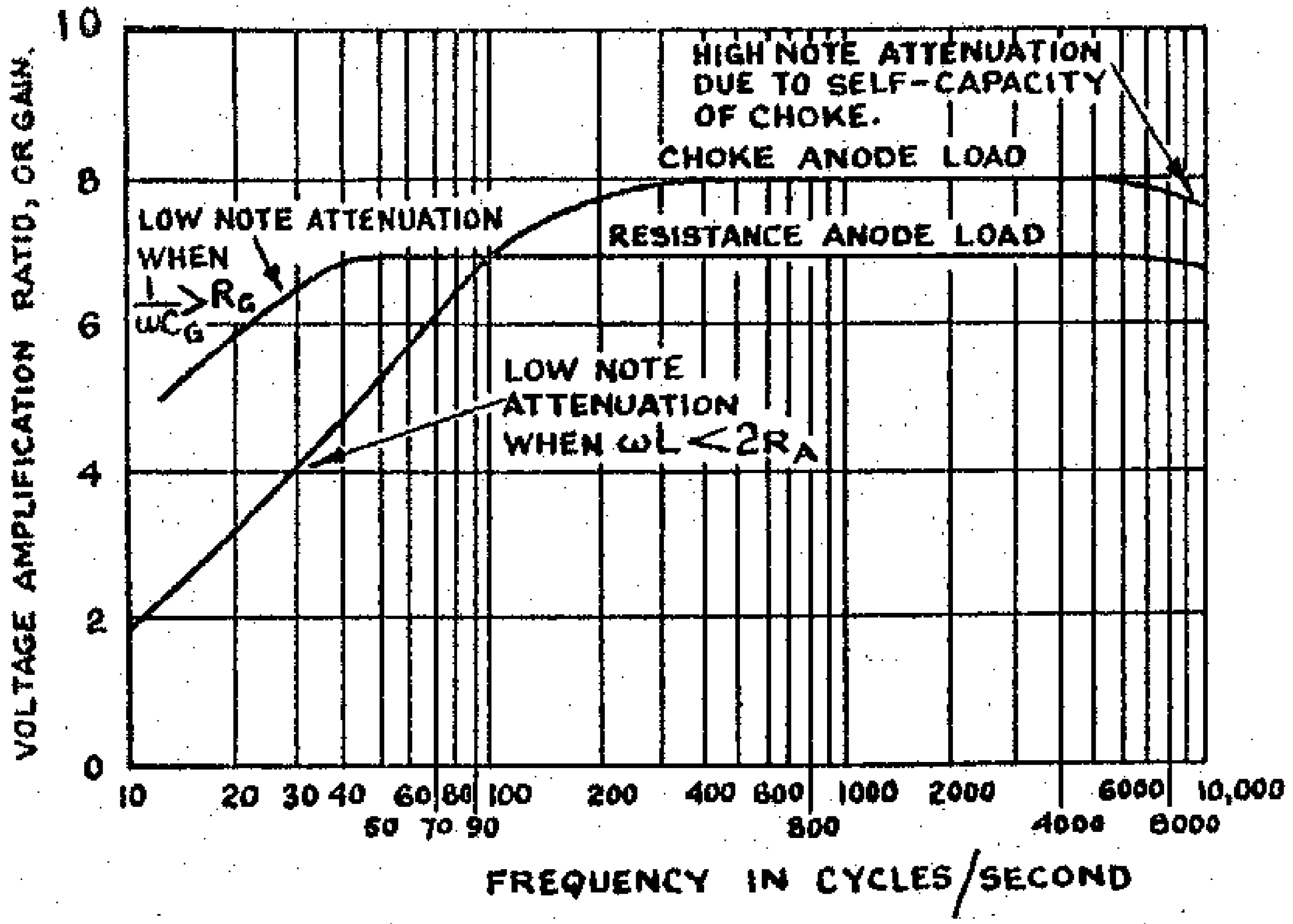


FIG. 53. Voltage-amplification vs. Signal-frequency Characteristic for R.C.C. Amplifier, and C.C.C. Amplifier.

The magnification of the first stage of a resistance-capacity coupled amplifier may be conveniently expressed by the formula:

$$\frac{E_{G2}}{E_{G1}} = \frac{\mu[R_1R_G/(R_1+R_G)]}{[R_1R_G/(R_1+R_G)]+R_A} \cdot \frac{R_G}{\sqrt{[R_G{}^2+(1/\omega^2C_G{}^2)]}}, \qquad (167)$$

where E_{G2} is the input alternating voltage to the grid of valve V_2, and E_{G1} is the input to valve V_1. In this formula, the effect of condenser C_G is ignored in calculating the effective anode load of valve V_1.

Choke-capacity Coupling. A choke can be used as the anode-load of valve V_1, in place of the resistance; otherwise the circuit is the same as the R.C.C. amplifier. The advantage is that the choke can have a high reactance to A.C. but a negligible resistance to D.C., and can therefore act as a high value load to produce a high stage-gain, yet not introduce any significant D.C. voltage-drop between the H.T. supply and the valve anode. If the low D.C. resistance of the choke is ignored, then the gain for the first stage will be $\mu\omega L/\sqrt{(\omega^2L^2+R_A{}^2)}$ (see fig. 54*a*).

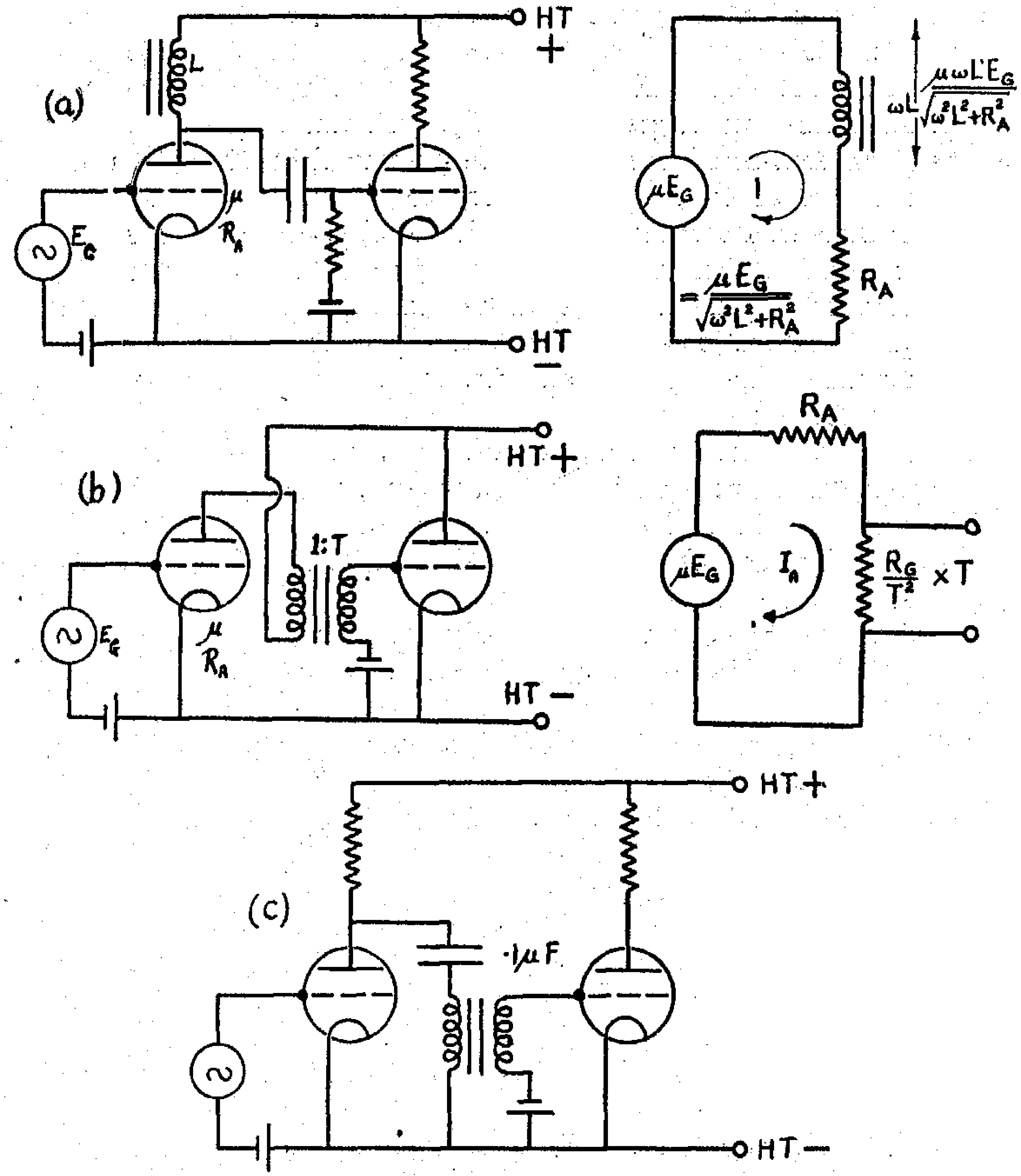


FIG. 54. *a*, Choke-capacity Amplifier, and Equivalent Circuit. *b*, Series-fed Transformer Coupled Amplifier, and Equivalent Circuit. *c*, Parallel-fed Transformer Coupled Amplifier.

The disadvantages attending the use of a choke as an anode load are:

(*a*) From the formula for the stage-gain it is apparent that the gain will be less for the lower values of the frequency $f = \omega/2\pi$.

The attenuation of low frequency signals, compared with higher values, will be reduced by the use of the maximum practicable value for the inductance L, so that ωL, even if ω is small, is still at least three times greater than R_A. In audio-frequency amplifiers, an iron-cored choke of value greater than 20 H. is used. The use of an iron core, necessary to ensure sufficiently high inductance in a component of moderate physical size, leads to a second difficulty in that the D.C. component of the valve anode current flowing through the choke winding will tend to cause magnetic saturation of the core, causing a reduction of effective inductance. The choice of an efficient component is made with both these points in mind.

(*b*) Since a large inductance necessarily has an associated self-capacity, a reduction of gain is also experienced at the higher frequencies, since then the choke impedance is considerably reduced by the effect of a parallel capacitance of value $1/\omega C$, where C is the choke self-capacity.

Transformer Coupling. The advantage offered by the use of a transformer as the coupling component between two valves in an amplifier is that an additional gain is furnished by the use of a step-up transformer with a ratio of as much as 1 : 8 times. Series-fed and parallel-fed connections are used (fig. 54): the latter circuit method has the advantage that the D.C. component of the valve anode current does not pass through the transformer primary, so its effective inductance is not thereby reduced.

At medium frequencies (200 to 3000 c./s.), the effects of the capacitances and inductances in the circuit shown in fig. 54*b* can be neglected in order to calculate, as correctly as is usually necessary, the gain of such an amplifier. A much simplified equivalent circuit can then be considered, as in fig. 54*b*. In this circuit, the primary inductance reactance and the effects of circuit capacitances are negligible compared with the effect of the reflected resistance R_G/T^2 due to the secondary load R_G, which is the input resistance of the second valve.

$$I_A=\frac{\mu E_G}{R_A+(R_G/T^2)}.$$

$\therefore$ the R.M.S. voltage across the transformer primary is

$$I_A\cdot\frac{R_G}{T^2}=\frac{\mu E_G(R_G/T^2)}{R_A+(R_G/T^2)}.$$

The output voltage across the transformer secondary is T times this amount, and thus will be $\dfrac{\mu E_G(R_G/T)}{R_A+(R_G/T^2)}$.

$$\therefore \text{ gain}=\frac{\text{voltage at grid of valve } V_2}{\text{voltage at grid of valve } V_1}=\frac{\mu R_G/T}{R_A+(R_G/T^2)}. \qquad (168)$$

This gain will be a maximum for a value of the transformer ratio T given by differentiating (168) with respect to T, and equating to zero.

$$\frac{-\left(R_A+\dfrac{R_G}{T^2}\right)\dfrac{\mu R_G}{T^2}-\dfrac{\mu R_G}{T}\left(-\dfrac{2R_G}{T^3}\right)}{\left(R_A+\dfrac{R_G}{T}\right)^2}=0.$$

$$\therefore \quad \frac{\mu R_G}{T^2}\left(R_A+\frac{R_G}{T^2}\right)=\frac{2\mu R_G{}^2}{T^4}.$$

$$\therefore \quad R_A+\frac{R_G}{T^2}=\frac{2R_G}{T^2}.$$

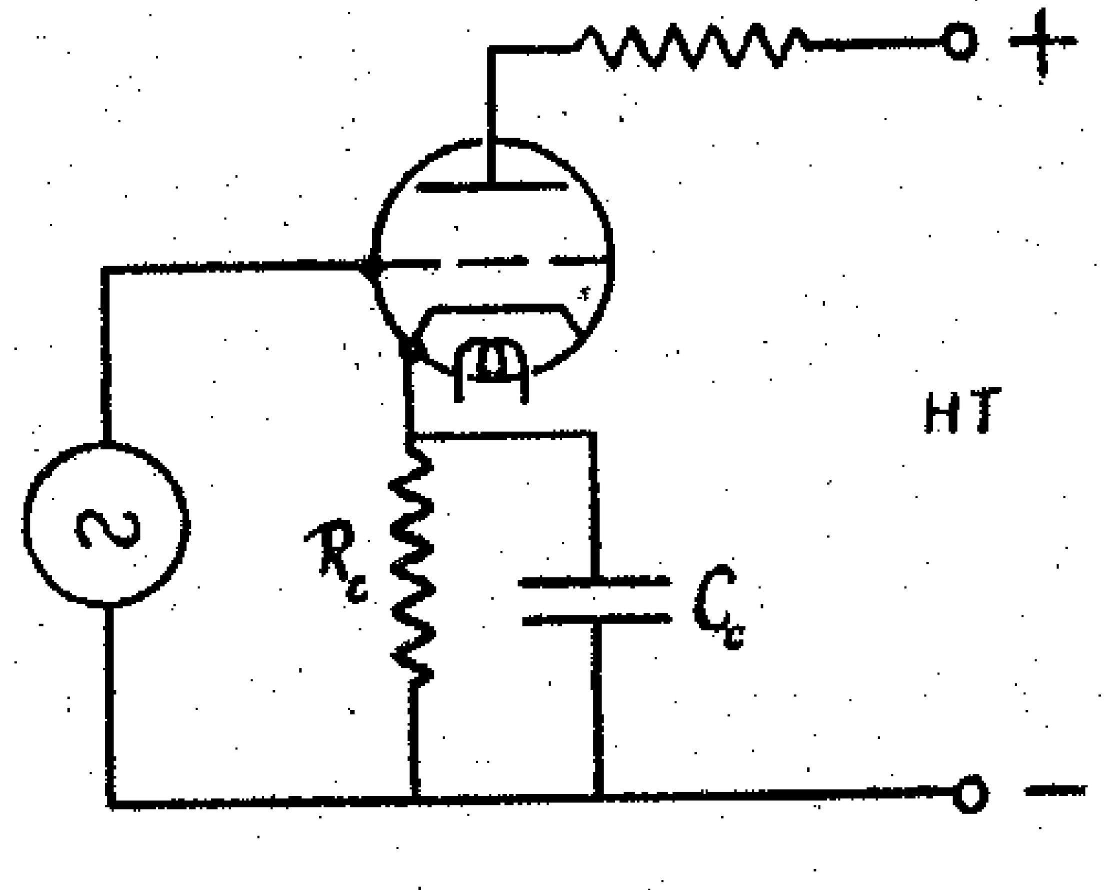
$$\therefore \quad \frac{R_G}{T^2}=R_A.$$

$$\therefore \quad T=\sqrt{\frac{R_G}{R_A}}. \qquad (169)$$

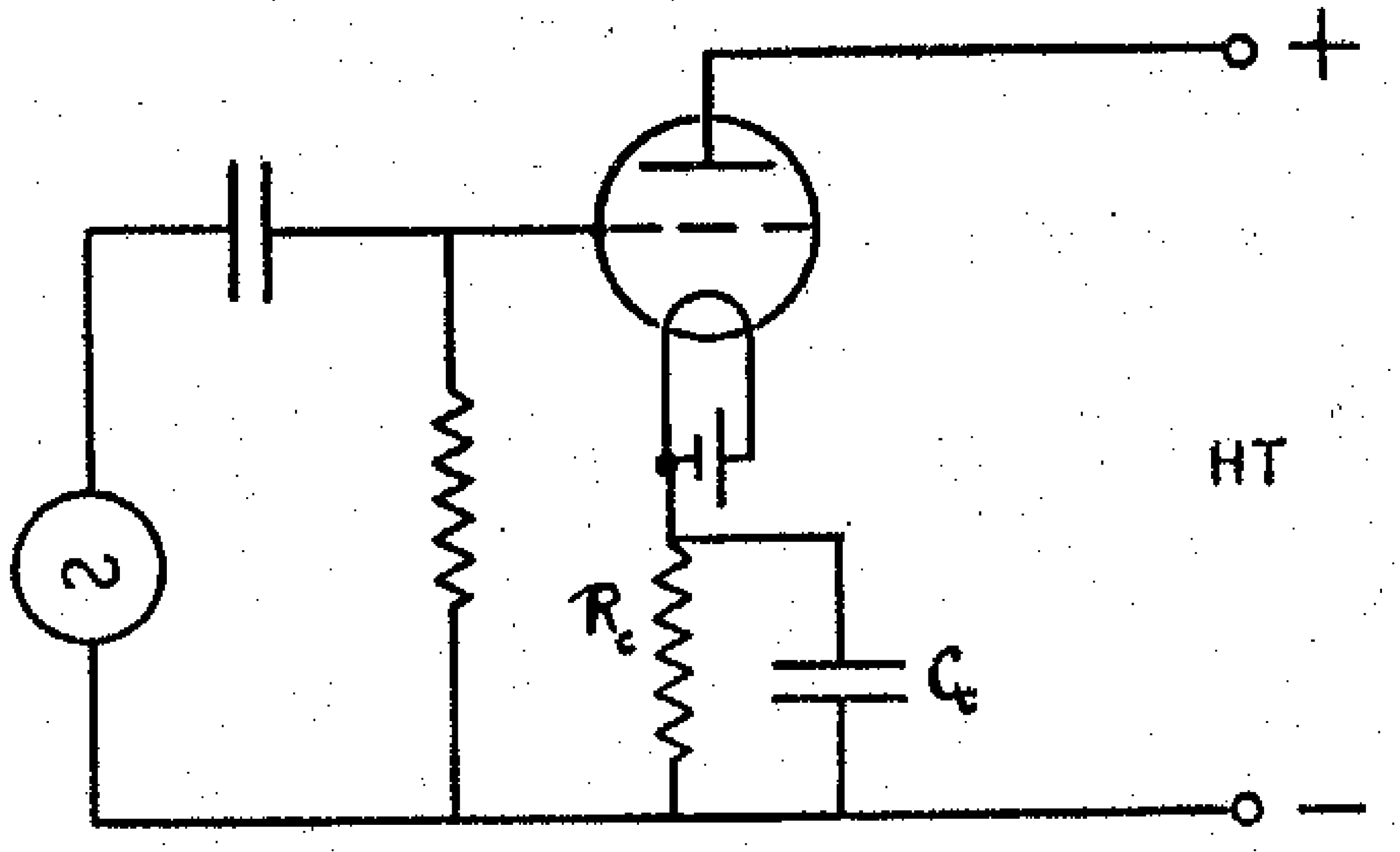
Since R_A, for an average triode valve, is about 50 kΩ., whilst R_G, the input resistance to a valve (grid negatively biased) is, about 5 MΩ. so the maximum values of transformer ratio T encountered in practical audio-frequency amplifiers are given by $T=\sqrt{(5\times10^6)/(5\times10^4)}=10$.

Automatic Bias. In amplifier practice the use of auxiliary grid-bias batteries is deprecated; a preferable method is to use an automatic bias arrangement. A resistance is arranged in the anode current circuit of the valve so that its cathode, or filament, depending on whether the valve is indirectly or directly heated, is maintained positive to the extent of the necessary bias relative to H.T.—, whereas the steady grid potential is zero (i.e. at H.T.—), (figs. 55). This is equivalent to arranging the grid potential negative with respect to the cathode.

The positive cathode bias is then equal to $I_A R_C$, where I_A is the steady component of the anode current, and R_C is the bias resistance used. Since I_A is known from valve data at the bias required, so R_C can be readily calculated, using Ohm's law, to obtain $I_A R_C$=bias required.



(a) CATHODE BIAS



(b) AUTOMATIC GRID BIAS, DIRECTLY-HEATED VALVE

FIG. 55. *a*, Cathode Bias Applied to an Indirectly heated Valve. *b*, Automatic Grid Bias used with a Directly heated Valve.

The anode current contains an A.C. component superimposed on the D.C. when it is amplifying an alternating signal input. The bias developed across the resistance R_C will, therefore, fluctuate at the signal frequency. To avoid this usually undesirable effect a by-pass condenser C_C is put in parallel with R_C of such capacity that its reactance $1/2\pi f C_C$ is less than 10% of R_C at the lowest signal frequency encountered.

For example, a valve has an anode current of 32 mA. when the bias is −10 V. What values of bias-resistance and condenser are necessary if the lowest signal frequency at the valve input is 100 c./s.?

$$\text{The bias voltage}=10=I_A R_C=\frac{32}{1000}R_C.$$

$$\therefore \quad R_C=\frac{10{,}000}{32}=313\ \Omega.$$

$$\frac{1}{\omega C_C}=\frac{1}{2\pi\,.\,100\,.\,C_C}\ \text{should equal } 31{\cdot}3\ \Omega.\ \text{max.}$$

$$\therefore \quad \text{minimum } C_C=\frac{10^6}{200\pi\times 31{\cdot}3}\ \mu\text{F.}=\frac{320}{2\pi}=51\mu\text{F.}$$

Grid Bias developed by Alternating Input. By the insertion of a condenser-resistance combination *CR* in the grid input circuit to a valve as shown in fig. 56, the negative bias developed on the grid relative to the cathode can be made to depend on the amplitude of the input alternating voltage. The condenser *C* is chosen to have such a capacity that it acts as a low reactance at the input frequency concerned. Such a circuit, therefore, is used only when the input is at high frequency (usually above 100 kc./s). A large fraction of the alternating input voltage will appear across the grid-cathode of the valve. Grid current will flow only when the valve grid is at a positive potential because of the input. Such grid current pulses will charge the condenser *C*, the plate *B* of this condenser acquiring a negative charge due to the accumulation of the electrons which reach the grid. Since plate *A* of the condenser is connected, via the resistance of the input generator circuit, to the valve cathode, so the D.C. potential developed across *C* will put a negative bias on the valve grid. If there is no leak resistance *R* across the condenser, then the bias developed will very soon be as great as the positive peak input voltage from the generator. The circuit will then cease to amplify the input and there will be no corresponding anode current alternations. On the other hand, if an appropriate value of resistance *R* is connected across the condenser, or connects the plate *B* to the cathode, then the steady P.D. across the condenser, and so the negative grid bias, will reach an equilibrium value which is some fraction of the positive peak input from the generator; this fraction depending on the magnitudes of *C* and *R*, and increasing towards unity as *R* is increased.

Such a method of producing a negative grid bias on a valve is widely used in oscillator circuits (Chapter 7). Moreover, if the input alternating potential has an amplitude which varies in accordance with a lower frequency modulation voltage (Chapter 8), then a detector circuit can be developed.

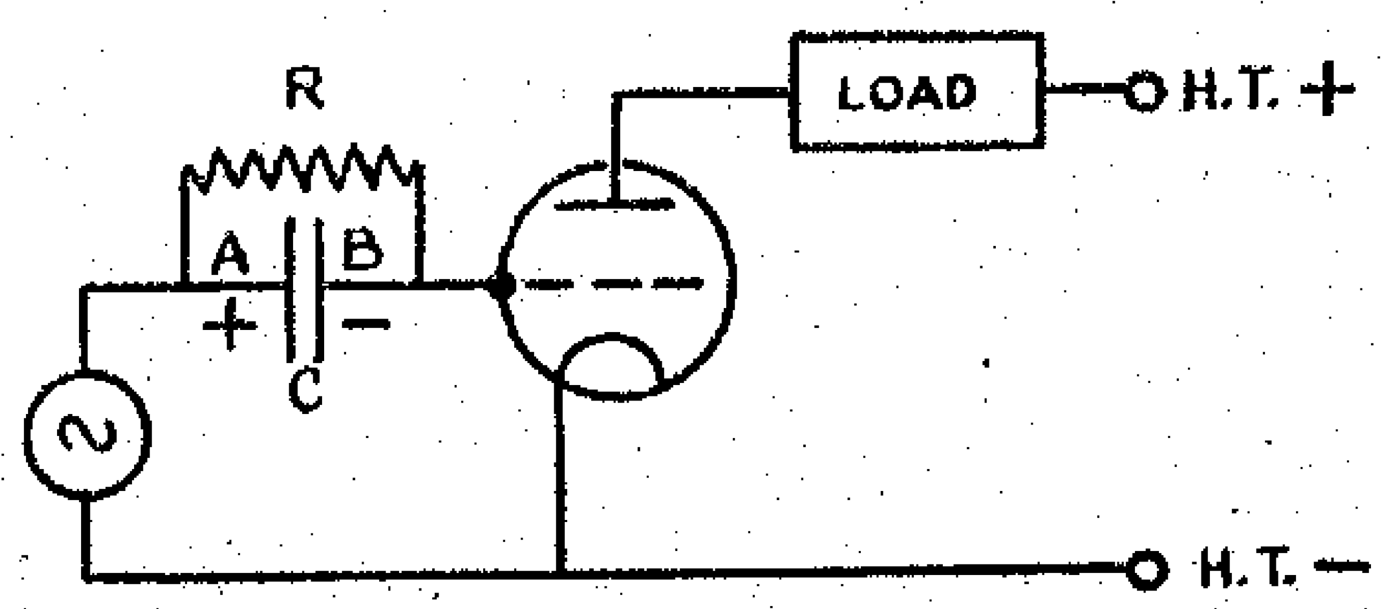


FIG. 56. Self-bias depending on Alternating Input.

High-frequency* Amplifiers. When amplifying signals at frequencies in excess of 50 kc./s. the effect of capacity in the circuit becomes important. In particular the capacitances between the electrodes of the valves used have to receive serious consideration. The effect of inter-electrode capacity in a valve is considered by Miller.

The Miller Effect.† The input admittance of a triode valve is controlled mainly by the effect of the inter-electrode capacitative link between the input circuit to the grid, and the output circuit at the anode.

In fig. 57 let the anode-grid capacity be C_{GA}, the anode-cathode capacity be C_{AC}, and the grid cathode capacity be C_{GC}. In the analysis which follows, the effect of C_{AC}, which is the smallest of these capacities, is neglected. In any case, it can be considered as simply adding slightly to the other two capacities. Using the notation indicated in fig. 57, the following equations are developed, on applying Kirchhoff's laws:

$$I_G = I_G' + I_G'', \qquad (170)$$

$$I_G' = E_G[(1/R_G) + j\omega C_{GC}], \qquad (171)$$

$$I_G'' = (E_G + mE_G\underline{/\theta})\, j\omega C_{GA}, \qquad (172)$$

where m is the stage-gain, and θ is the phase-angle between $-E_G$, the grid input voltage, and mE_G the voltage across the anode load impedance Z. The magnitude of θ will depend upon the relative values of the resistance, inductance and capacitance comprising Z.

* The terms high frequency (H.F.), and radio frequency (R.F.), are usually considered to be synonymous.

I_G is the total effective grid current from the supply at pulsatance ω, whilst I_G' and I_G'' are the components of this current via C_{GC} and C_{GA} respectively. R_G is the input resistance to the valve.

Equation (172) is conveniently rewritten in the form:

$$I_G'' = j\omega C_{GA}[E_G + mE_G(\cos\theta + j\sin\theta)]. \qquad (173)$$

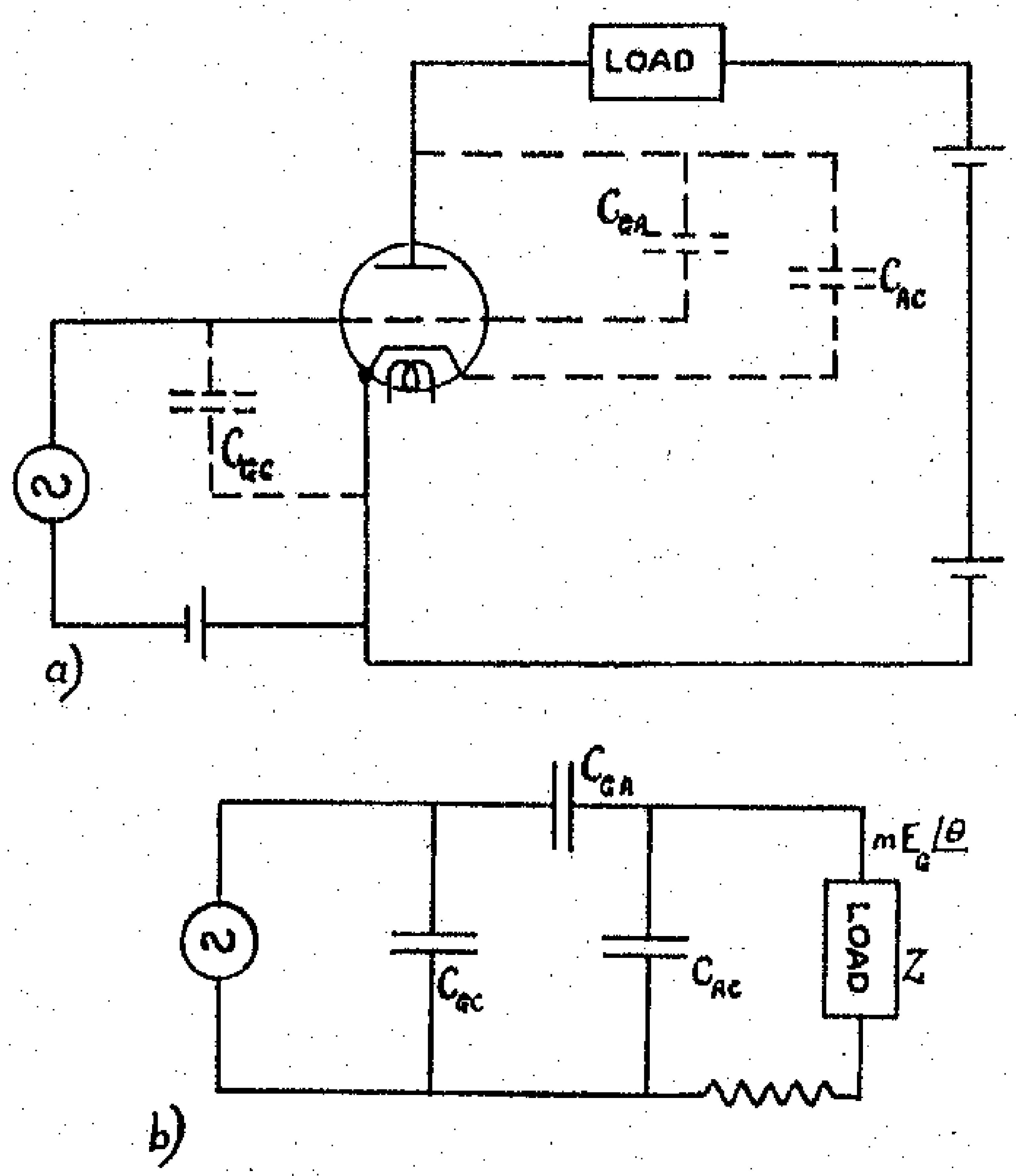


FIG. 57.—The Input Admittance of a Triode Valve (Miller Effect).

Substituting for I_G' and I_G'' from (171) and (173) in (170) gives

$$I_G = E_G[(1/R_G) + j\omega C_{GC}] + j\omega C_{GA}[E_G + mE_G(\cos\theta + j\sin\theta)].$$

$$\therefore \quad \text{Input admittance } Y_G = \frac{I_G}{E_G}$$

$$= \frac{1}{R_G} - m\omega C_{GA}\sin\theta + j\omega(C_{GC} + C_{GA} + mC_{GA}\cos\theta). \qquad (174)$$

Equation (174) clearly indicates that both the in-phase and 90° out-of-phase components of the input admittance are affected by m, and hence by the anode load. The input admittance will have a negative in-phase, or resistive component if $m\omega C_{GA}\sin\theta$ is positive, and exceeds $1/R_G$ in value. This implies that energy is being fed back via the anode-grid capacity to the input circuit in such a manner as to overcome the effects of positive resistance energy-loss at the input. With suitable input and output LC circuits, the input LC circuit can have its resistance loss reduced to zero, or made negative, so that continuous oscillations may be sustained (cf. the T.A.T.G. oscillator p. 182).

It is seen from (174) that the effective input capacity of a triode valve is considerably greater than its nominal value, which depends on electrode geometry. For example, suppose the anode load is resistive, so that $\theta=0°$, and let C_{GC} be 5 $\mu\mu$F., C_{GA} be 8 $\mu\mu$F. and the stage-gain be twenty times. Substitution of these values in equation (174) shows that the effective input capacity C_i is given by:

$$C_i=C_{GC}+(m+1)C_{GA}=5+(21\times 8)=173\ \mu\mu\text{F.}$$

At a frequency of 10,000 c./s., at the upper limit of the audio-frequency range, a nominal grid-cathode capacity of 5 $\mu\mu$F. will be a reactance of $\dfrac{1}{\omega C_{GC}}=\dfrac{1}{2\pi\times 10^4\times 5\times 10^{-12}}=3{\cdot}18\ \text{M}\Omega.$, which is negligible in its effect. But the true effective input capacitative reactance is $\dfrac{1}{2\pi\times 10^4\times 173\times 10^{-12}}=92{,}000\ \Omega.$, which is certainly not negligible. At radio-frequencies of the order of 1 Mc./s. such an input capacitative reactance will be only 920 Ω., so that an R.F. voltage input of only 0·92 V. will produce a reactive current component of 1 mA.

An amplifier intended for signals at radio-frequency cannot then be efficient if resistance-capacity coupled triode valves are used. The use of a tuned rejector-circuit as an anode load in place of a resistance will partly overcome the deleterious effects of the inter-electrode capacities discussed above, since the capacities are then simply in parallel with the necessary tuned circuit capacity for resonance, the gain per stage being $\dfrac{\mu(L/CR)}{(L/CR)+R_A}$, where L/CR is the dynamic resistance of the rejector-circuit at resonance (see p. 57). However, the use of input- and output-tuned circuits associated with a triode valve, tend to make the circuit prone to oscillate (see p. 172). Again, the effective input capacity of $C_{GC}+(m+1)C_{GA}$ will limit the minimum capacity which can be associated with the tuned circuit used. The remedy is to use a valve of much lower anode-grid capacity of the screen-grid or pentode type. Such valves give an added advantage for use as R.F. voltage amplifiers, in that their A.C. resistance and amplification factor are much higher than those of triodes.

The Screen-grid Tetrode. A second grid is introduced between the grid and anode of a triode, reducing the capacity between these electrodes, giving a tetrode, or four-electrode valve. This screen grid is usually maintained at a positive D.C. potential of value about two-thirds the anode D.C. potential. As regards R.F. voltages, however, this screen is earthed by a by-pass condenser, usually 0·1 μF., connected between the screen and earth (see fig. 58).

In the construction of this valve particular attention is paid to the problem of ensuring an anode-grid capacity which is as low as possible. To this end the anode area is reduced in size; the grid and anode leads are taken to separate ends of the valve, and skirts at screen potential are arranged at the top and bottom of the electrode assembly to reduce the electrostatic field which exists between the control grid ends and the anode. In this fashion, anode-grid capacities as low as 0·01 $\mu\mu$F. are achieved compared with values of the order of 5 $\mu\mu$F. obtained in the usual receiver triodes.

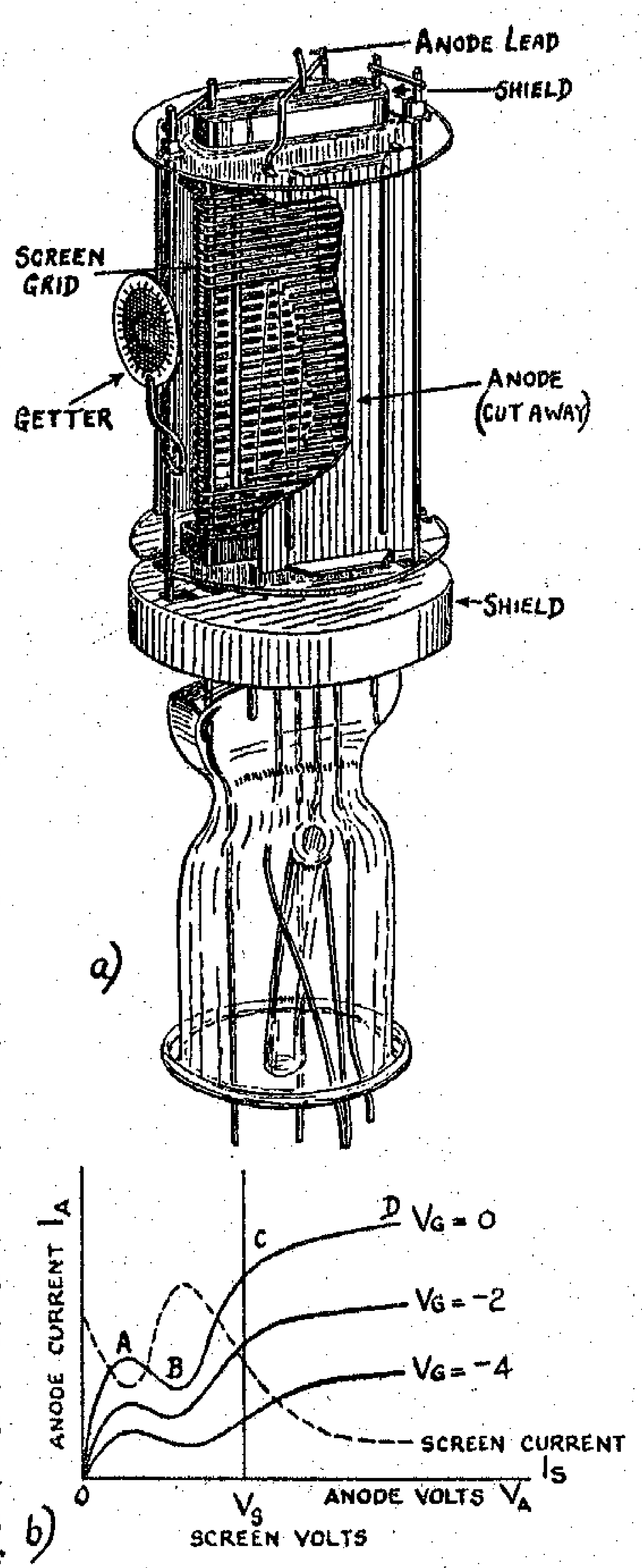


FIG. 58. *a*, The Screen-grid Valve.
b, Anode Characteristic of Screen-grid Valve.

The anode-current vs. anode-voltage ($I_A : V_A$) characteristic of this tetrode is peculiarly kinked, as is shown in fig. 58*b*. This effect is due to secondary emission from the anode.

The useful portion of this characteristic in R.F. amplifier practice is between points C and D; it is essential to confine the anode load voltage variations obtained in the amplifier to within these limits, since departure from this restricted linear region will result in distortion. The line CD is inclined by only a very small angle to the anode volts axis, and this inclination represents a high value of A.C. resistance. This high resistance is directly due to the introduction of the positive screen. A change of anode voltage ∂V_A will not be very effective in altering the anode current, since the electric field at the valve cathode, due to the positive anode, is much reduced by the shielding effect of the screen grid. For a change ∂V_A the corresponding anode current change ∂I_A will be very small; consequently the ratio $\partial V_A / \partial I_A = R_A$ is large. Likewise, the amplification factor $\partial V_A / \partial V_G$ will be considerably greater than in a triode valve, since the anode voltage will have to be changed by a large amount to be as effective as a grid voltage change in altering the anode current. In screen-grid valves, the A.C. resistance is between 0·5 and 2 MΩ., whilst the amplification factor is between 500 and 2000. The mutual conductance will be much the same as in the corresponding triode, since the electric field of the control grid at the cathode is little affected by the presence of the screen.

If the anode voltage of this type of tetrode is reduced to below the screen voltage, then the anode current decreases rapidly, as is indicated by the region of the curve CB. The anode will emit secondary electrons. In the normal operation of the valve, with the anode voltage greater than the screen volts, these secondaries will simply return to the anode, as they do in a triode valve. If, however, the screen is more positive than the anode, then many of the secondary electrons will go to the screen-grid instead; the anode current will then be reduced because electrons are leaving it in numbers almost as great as those arriving, whereas the screen-current will increase. If the anode voltage is much reduced—to only 20 to 30 V.—secondary electrons are no longer produced since the primaries will have insufficient energy. A subsidiary rise of anode current to a minor peak value at A is therefore experienced, indicated by the region BA of the characteristic curve. Then as the anode voltage is decreased to zero, so the anode current falls to zero.

Over the region AB of the curve, it is remarkable that the anode current decreases with increase of anode voltage, so $R_A = \partial V_A / \partial I_A$ is negative. The dynatron oscillator (p. 183) utilises this effect, since the negative resistance of a screen-grid valve, operated at the correct potentials, is balanced against the positive dynamic resistance of a tuned circuit in the anode. This negative resistance is due to the fact that, within the range of anode voltages experienced between points A and B of the curve, an increase of anode voltage brings about a greater increase of secondary emission than it does of primary electrons arriving at the anode.

The Pentode and Beam Tetrode Valves. In modern electronic practice the screen-grid valve is rarely used: the pentode or beam tetrode, which developed from the screen-grid type, are preferable. In both these valves, the kink in the characteristic due to secondary emission is eliminated: in the pentode by the use of a third grid, the suppressor, situated between the screen and anode; in the beam tetrode by the use of a more critical anode spacing in conjunction with the use of "beam" plates. Both radio-frequency and audio-frequency patterns of these valves are made: in the former, care is taken in the electrode construction and connections to ensure low anode-grid capacity; in the low-frequency model this low capacity is not an essential requirement, the chief attention being paid to ensuring high efficiency working concomitant with low distortion.

In the pentode valve, the suppressor grid, which is usually at cathode potential, has fewer turns per cm. than the control or screen grids, and is placed just inside the anode. As a result, the suppressor does not significantly impede the fast primary electrons accelerated through the screen to the anode, whereas the comparatively slow secondaries, travelling in all directions away from the anode plane, are readily decelerated to zero velocity before this third grid, and so return to the anode instead of penetrating to the screen grid.

When the anode potential is lower than the screen potential, this suppression of secondaries is still effective, but then the slower primary electrons become somewhat impeded by the suppressor grid, giving rise to the rounded "knee" typical of the pentode characteristic (fig. 59).

The beam tetrode valve dispenses with a suppressor grid, but instead is furnished with a pair of plates disposed about the screen-grid supports, and maintained at cathode potential, so that primary electrons which reach the anode are confined to a beam (see fig. 59). Electrons which leave the sides of the cathode, and which travel through the distorted electrostatic fields around the grid support wires, do not reach the anode, since they impinge on the beam plates. Consequently, a more rigid relationship is established between electrode shape and disposition, and anode current. In some types of beam tetrode, particularly the 6L6 (KT 66) valves, the screen and control grids have the same number of turns per cm., and are aligned so that the screen wires are exactly behind the control grid wires. This leads to further confinement of the effective electrons to beams, and also reduces the screen current.

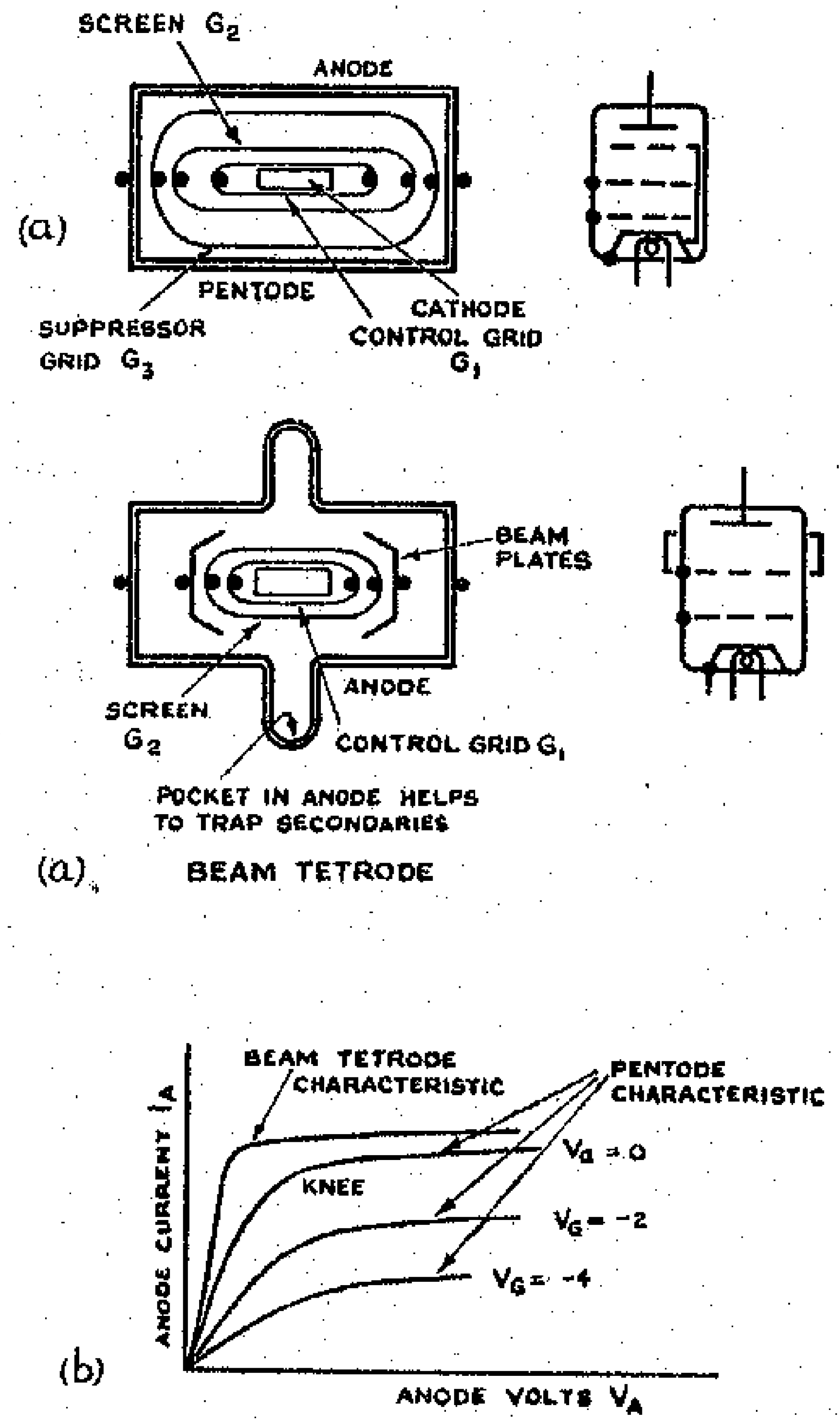


FIG. 59. *a*, Pentode and Beam Tetrode Valves. *b*, Pentode and Beam Tetrode Characteristics.

To prevent secondary electrons leaving the anode from reaching the screen of this valve, the anode is placed at such a distance that when its potential is less than the screen potential, then the primary electrons build up a negative space-charge between screen and anode. A potential minimum is thereby introduced which does not prevent the majority of the fast-moving primaries from reaching the anode but, on the other hand, slows down the slower secondaries to such an extent that they return to the anode instead of reaching the more positive screen. This potential minimum is further enhanced by the presence of the beam plates. The result is an anode characteristic similar to that of the pentode valve, but with a rectangular-shaped rather than rounded knee, occurring at a lower anode potential. Such a valve can therefore be used with a bigger variation of anode load voltage without distortion occurring than is the case in the equivalent pentode. As a power-amplifier valve it is therefore capable of greater working efficiency.

The R.F. pentodes and beam tetrodes are furnished with an anode of small area, a top-cap grid connection, and skirts at the upper and lower ends of the electrode assembly, as in the case of the screen-grid valve, to ensure low anode-grid capacity. Likewise the grids have many turns of wire per cm., and the anode spacing is great to obtain a valve of high A.C. resistance and amplification factor. These requirements are assisted by the introduction of a suppressor grid which, shielding the anode, enables even higher μ and R_A to be obtained than in the screen-grid valve.

In the case of the low-frequency beam power valve, used in particular as the output valve of an amplifier supplying considerable load current, the designer concentrates on ensuring a high mutual conductance, sufficient anode dissipation and with the knee of the characteristic at as low an anode potential as possible.

Radio-frequency Tuned Voltage Amplifiers. To amplify a signal voltage alternating at high frequency (100 kc./s. to 50 Mc./s.), R.F. pentode valves are usually employed, with parallel tuned circuits as anode loads. The commonest types of coupling are known as (*a*) tuned anode, (*b*) tuned grid, (*c*) R.F. transformer, and (*d*) tuned R.F. transformer coupling (see fig. 60).

(*a*) *Tuned-anode Coupling.* In this case a tuned rejector circuit (see p. 57) is used as the anode load for the first valve. Since such a circuit behaves as a high resistance of value L/CR at resonance, so the gain exhibited by the first stage is given by

$$m = \frac{\mu(L/CR)}{(L/CR)+R_A}. \qquad (175)$$

Since $L/CR = Q\omega L$, from equation (78), so a high value of m is obtained in this circuit by using a coil of high "Q". This necessarily implies a sharply selective tuned circuit, which may involve the cutting of the higher audio-frequencies in the sidebands (see p. 197) if such an R.F. amplifier is used in a domestic receiver. It is noteworthy that the A.C. resistance, R_A, of an H.F. pentode is of the order of 1 MΩ. The dynamic resistance of a tuned rejector circuit cannot usually be more than 200 kΩ. Substitution of these figures in equation (175) evinces that the gain m obtainable is much smaller than the maximum possible value, μ. Thus if μ is 1000 for a particular H.F. pentode, then

$$m=\frac{1000\times 100{,}000}{100{,}000+1{,}000{,}000}=\frac{1000}{11},$$

only $\frac{1}{11}$th of the maximum possible amount.

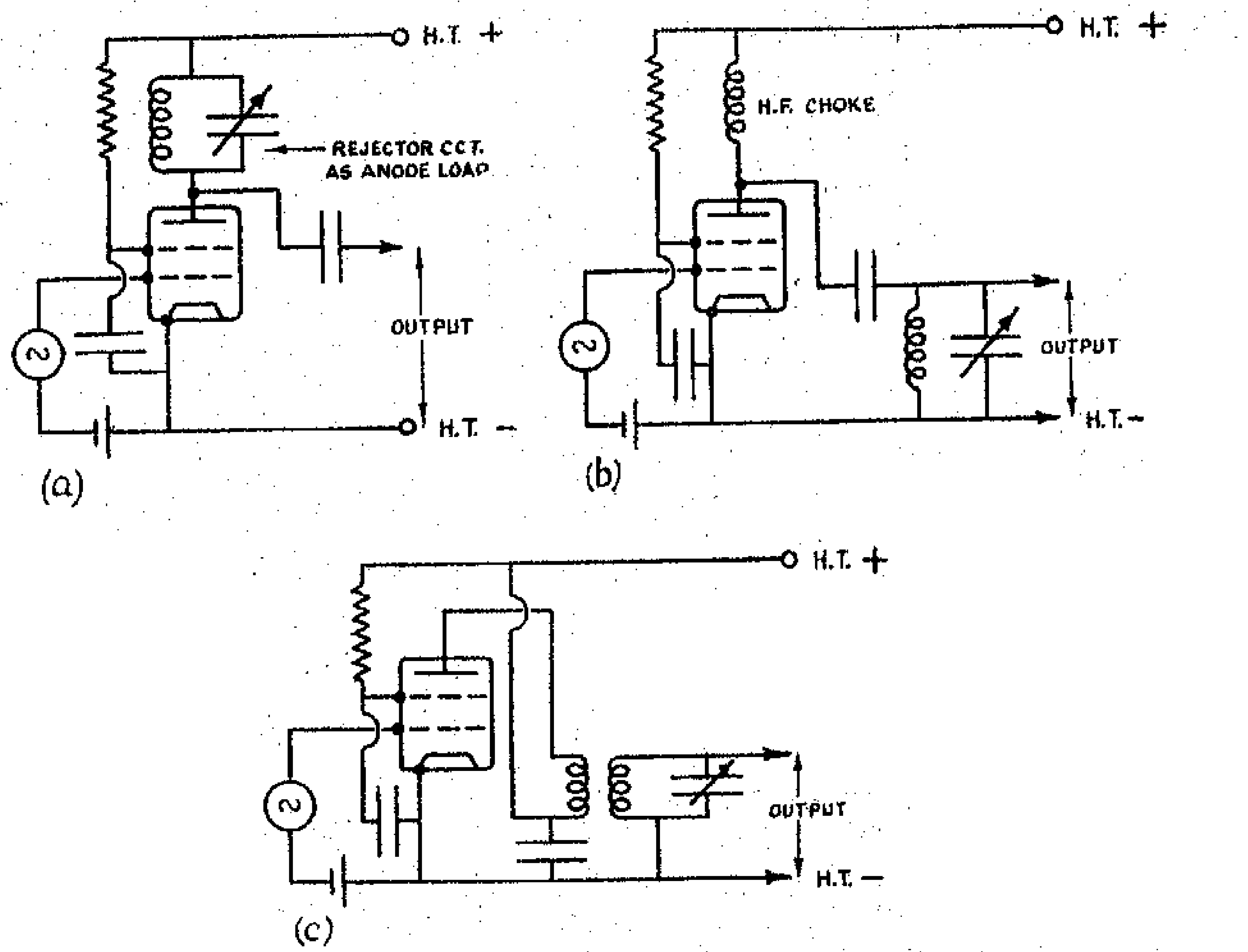


FIG. 60. R.F. Voltage Amplifiers. *a*, Tuned-anode Coupling. *b*, Tuned-grid Coupling. *c*, R.F. Transformer Coupling.

The gain m can be conveniently expressed, in such circumstances, as

$$m=\frac{\mu(L/CR)}{R_A}=g_m\frac{L}{CR}=g_m\omega LQ, \qquad (176)$$

where L/CR is neglected compared with R_A in the denominator. Thus the effective Q of the amplification curve, deciding the selectivity of the circuit, is approximately the same as the Q of the rejector circuit alone.

(*b*) *Tuned-grid Coupling.* This coupling method is illustrated in fig. 60*b*. An air-cored inductance, or choke, which has a high impedance to A.C. at the radio frequency used, is employed as the anode load. The voltage variations produced across it are coupled by a condenser to a tuned circuit which forms the input to the second stage of the amplifier. This tuned circuit improves the selectivity of the amplifier, but reduces the effective anode load presented to the valve, because it is effectively in parallel with the choke. This type of coupling is not frequently used, methods (*c*) and (*d*) being preferable.

(*c*) *R.F. Transformer Coupling.* This method of connection has the practical advantage that the grid input circuit of the second valve is isolated from the H.T. supply. The primary or secondary or both can be tuned circuits. In the third case the transformer can have band-pass characteristics (see pp. 68 and 197) to enable it to select a required frequency band width. Considering the case where the secondary only is tuned, then from formula, p. 66,

$$R=R_1+\frac{\omega^2M^2R_2}{Z_2^{\,2}} \quad \text{and} \quad X=X_1-\frac{\omega^2M^2X_2}{Z_2^{\,2}},$$

where ω is the pulsatance of the signal concerned, M is the mutual inductance of the R.F. transformer, R_1 and X_1 are the primary resistance and reactance respectively and R_2, X_2 and Z_2 are the secondary resistance, reactance and impedance respectively. R is the effective primary resistance, increased by the tuned circuit coupled to it; X is the effective primary reactance, decreased by the coupled tuned circuit.

Since the secondary is tuned, $Z_2=R_2$ and $X_2=0$, and the reflected resistance in the primary due to the secondary becomes $\frac{\omega^2M^2R_2}{X_2^{\,2}+R_2^{\,2}}=\frac{\omega^2M^2}{R_2}$. Since R_2 is small, and ω is a large value at radio frequency, so this reflected resistance is great compared with R_1 and X_1, and these latter can be neglected.

$\therefore\quad Z_1=\frac{\omega^2M^2}{R_2}$ approximately gives the effective primary impedance.

$$\therefore\quad \text{Primary current } I_1=\frac{\mu E_G}{R_A+(\omega^2M^2/R_2)},$$

where E_G = alternating voltage applied to grid of the first valve, and μ = amplification factor of first valve.

The voltage induced in the secondary $=\omega MI_1$.

Secondary current $=\frac{\omega M I_1}{R_2}=I_2$.

Voltage across secondary $=E_G'=\omega L_2 I_2$, where L_2 is the inductance in the tuned secondary circuit.

$$=\omega L_2 \cdot \frac{\omega M}{R_2} \cdot \frac{\mu E_G}{[R_A+(\omega^2 M^2/R_2)]}.$$

$$\therefore \quad \text{Gain} \quad m=\frac{E_G'}{E_G}=\frac{\omega^2 \mu M L_2}{R_2 R_A+\omega^2 M^2}.$$

This gain will be a maximum for a value of coupling given by putting $dm/dM=0$.

$$\therefore \quad \frac{d}{dM}\left[\frac{\omega^2 \mu M L_2}{R_2 R_A+\omega^2 M^2}\right]=0.$$

$$\therefore \quad (R_2 R_A+\omega^2 M^2)\,\mu\omega^2 L_2=\mu\omega^2 M L_2 \,.\, 2\omega^2 M.$$

$$\therefore \quad R_2 R_A+\omega^2 M^2-2\omega^2 M^2=0.$$

$$\therefore \quad \omega^2 M^2=R_2 R_A.$$

Then $m=\frac{\mu\omega^2 M L_2}{2\omega^2 M^2}=\frac{\mu L_2}{2M}$.

But $M=K\sqrt{(L_1 L_2)}$, where K is the coefficient of coupling between the two circuits, where L_1 is the primary inductance.

$$\therefore \quad m=\frac{\mu L_2}{2K\sqrt{(L_1 L_2)}}=\frac{\mu}{2K}\sqrt{\frac{L_2}{L_1}}. \qquad (177)$$

Again, since $\omega^2 M^2=R_2 R_A$

$$\therefore \quad \omega^2 K^2 L_1 L_2=R_2 R_A$$

and putting $\omega^2=\frac{1}{L_2 C_2}$, therefore $\frac{K^2 L_1 L_2}{L_2 C_2}=R_2 R_A$, where C_2 is the tuned secondary capacitance.

$$\therefore \quad K^2=\frac{R_2 R_A C_2}{L_1} \qquad (178)$$

gives the required optimum coupling value, which on inserting practical values for R_2, R_A, C_2 and L_1, where the frequency is 1 Mc./s., makes K about 0·8.

(*d*) *Tuned R.F. Transformer Coupling.* If both primary and secondary circuits are tuned, it can be shown by a complete analysis of the circuit that optimum coupling occurs when

$$\omega^2 M^2=\left(R_1+\frac{L_1}{C_1 R_A}\right)R_2 \qquad (179)$$

and the gain $\quad m=\frac{\mu(L_1/C_1 R_A)}{2\sqrt{\{[R_1+(L_1/C_1 R_A)]R_2\}}}. \qquad (180)$

Optimum coupling at 1 Mc./s. occurs for values of K of the order of 0·01.

Variation of Gain with Frequency. In all these methods of coupling together the stages of an R.F. amplifier, it is important to note that the gain and condition for optimum coupling depend on the signal frequency, $\omega/2\pi$. Hence in normal broadcast receiver practice, where it is required to accept and amplify a widely varying range of radio-frequencies depending on the station being received, it is apparent that the receiver performance must needs vary considerably from one end of the wave-band to the other. This variation is largely overcome by employing the superheterodyne principle whereby all receiver aerial signals are changed to a constant frequency value, called the intermediate frequency, I.F., and I.F. amplifier stages employ coupling transformers with both primary and secondary tuned, and designed to work at the optimum for the constant frequency value chosen.

The Variable-mu Valve. Equation (176) indicates that the gain of an H.F. pentode, employing a tuned circuit directly or indirectly coupled as an anode load, depends directly on the mutual conductance of the valve. In the variable-mu valve, the control grid of a screen-grid tetrode or H.F. pentode is more open at the centre than at the ends (see fig. 61), or is wound so that the spaces between adjacent grid turns increase exponentially from one end of the grid to the other. As a result the mutual characteristics ($I_A : V_G$) of such a valve exhibit pronounced "tail", since the cathode surface behind the more open spaces of the grid continues to furnish electrons to the valve anode current when the grid bias is sufficiently negative to prohibit by space-charge the flow of electrons from the rest of the cathode. Obviously such a valve has a large cut-off bias, and a value of mutual conductance which increases continuously as the negative grid bias is reduced from cut-off value to zero.

If the grid bias on a variable-mu valve is altered by means of a manually operated potentiometer, or is automatically controlled depending on the strength of the alternating input to the valve grid, so either manual or automatic control of the gain of an R.F. amplifier is achieved. This leads to the common forms of manual and automatic volume control methods used in broadcast receivers.

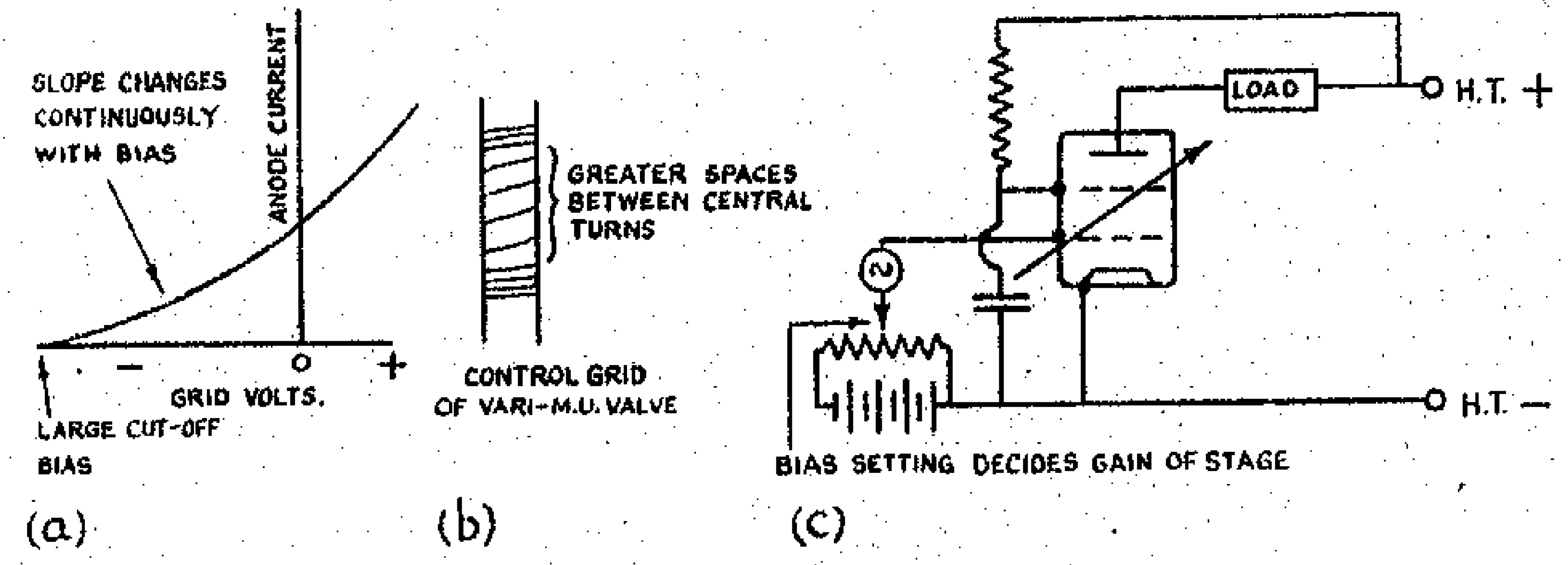


FIG. 61. The Action of the Variable-mu Valve.

Power Amplification. In using a valve as a power amplifier, the aim is to supply maximum power to the anode load with as little distortion as possible, instead of maximum voltage variation across the load. This demands large current variations through the load, so that if the use of excessive voltage variations across the valve is to be avoided, the valve must necessarily have a low A.C. resistance. Hence the power valves used as output valves in receivers, transmitters and other electronic apparatus where electrical power is converted into mechanical power or electromagnetic radiation, have much lower A.C. resistances than the voltage amplifying valves in electronic equipment. The latter usually serve to supply the power valve with a sufficiently large grid swing voltage, demanding little or no power consumption.

Let a valve amplifier be operated so that

E_G = input voltage to grid, R.M.S. value.
μ = amplification factor of valve.
R_A = A.C. resistance of valve.
R_L = anode load resistance.

Assuming the grid is not driven positive, so that grid current is avoided, then the effective anode voltage variation due to an input of E_G volts is μE_G. The corresponding current I is $\mu E_G/(R_A+R_L)$. The power developed across the anode load is therefore $I^2R_L=\dfrac{(\mu E_G)^2 \cdot R_L}{(R_A+R_L)^2}=P$.

This will be a maximum for a value of R_L given by putting $dP/dR_L=0$.

$$(\mu E_G)^2\left[\frac{(R_A+R_L)^2-R_L \cdot 2(R_A+R_L)}{(R_A+R_L)^4}\right]=0.$$

$$\therefore \quad R_A+R_L-2R_L=0.$$

$$\therefore \quad R_A=R_L. \qquad . \quad . \quad . \quad . \quad (181)$$

As in the case of any other form of electric generator, maximum power is supplied to the load when the load resistance equals the internal resistance of the source of supply. However, the requirement of low distortion must also be fulfilled, requiring extra consideration of the optimum load resistance value to be used.

Class A Power Amplification. The anode load of a power amplifier must necessarily be a resistive device; power cannot be dissipated in the wattless reactive components, condensers and inductances, in which the current differs in phase by 90° from the voltage. The usual loads are therefore ordinary resistances, or resistive elements such as loudspeakers, headphones, electric meters, telephone meters or aerials radiating energy or such resistances coupled to the power valve by a matching transformer or tuned circuits of the rejector circuit type with a make-up current in phase with the applied A.C. voltage: simulating a resistance in their effect. If a resistance is inserted directly in the anode circuit of the valve, then class A amplification conditions must be observed using an ordinary amplifier if distortion is to be avoided. This brings about the limitation that the grid swing voltage is restricted, and also that an increase of anode current brought about by a reduction of negative grid voltage is accompanied by an increase of anode load voltage, with a consequent reduction of anode volts, and hence power output.

Suppose a power valve has constants μ, g_m and R_A, and it operates with a mean, steady anode voltage V_A. Let $-V_G$ be the cut-off grid bias with this anode voltage. An alternating grid voltage is applied which, at its positive peak value, just brings the grid potential to zero, whereas at the negative peak value the dynamic characteristic ($I_A - V_G$) of the valve just begins to bend, so that class A conditions are observed (cf. p. 119). Let $2I_A$ be the total peak-to-peak anode current excursion brought about by the maximum permissible grid voltage change. Then the anode potential will be caused to vary from $V_A + I_A R$ to $V_A - I_A R$, if R is the resistive anode load. A change of anode potential of $I_A R$ is compensated by a change of grid voltage of $I_A R/\mu$, so that when the anode potential is a maximum at $(V_A + I_A R)$, the necessary cut-off bias will be $[-V_G - (I_A R/\mu)]$. Assuming the valve $I_A - V_G$ characteristic is linear down to the cut-off bias value, then the maximum permissible peak grid voltage input is $\frac{1}{2}[V_G + (I_A R/\mu)]$.

But $I_A = \dfrac{\mu E_G}{R+R_A}$, and substituting for E_G

$$I_A = \frac{\mu}{R+R_A} \cdot \frac{1}{2} \cdot \left(V_G + \frac{I_A R}{\mu}\right).$$

$$\therefore \quad I_A\left\{1 - \frac{R}{2(R+R_A)}\right\} = \frac{\mu V_G}{2(R+R_A)}.$$

$$\therefore \quad I_A(2R_A + R) = \mu V_G.$$

$$\therefore \quad I_A = \frac{\mu V_G}{R+2R_A}.$$

The power output P will be $I^2_{R.M.S.}R$, where $I_{R.M.S.} = I_A/\sqrt{2}$.

$$\therefore \quad P = \frac{\mu^2 V_G^2 R}{2(R+2R_A)^2}.$$

To find the value of the anode load R at which this power output is a maximum, put $dP/dR = 0$.

$$\therefore \quad \frac{\mu^2 V_G^2}{2} \cdot \frac{d}{dR}\left\{\frac{R}{(R+2R_A)^2}\right\} = 0.$$

$$\therefore \quad (R+2R_A)^2 - R \,.\, 2(R+2R_A) = 0.$$

$$\therefore \quad R^2 + 4RR_A + 4R_A^2 - 2R^2 - 4RR_A = 0.$$

$$\therefore \quad 4R_A^2 = R^2.$$

$$\therefore \quad R = \pm 2R_A. \qquad (182)$$

So the maximum power output is obtained when the anode load is twice the A.C. resistance of the valve.

The power output at this optimum load value will be

$$P = \frac{\mu^2 V_G^2 R}{2(R+2R_A)^2} = \frac{V_A^2 \,.\, 2R_A}{32R_A^2} = \frac{V_A^2}{16R_A}.$$

The Load Line. The selection of an anode load of value $R = 2R_A$ leads theoretically to the attainment of maximum undistorted power output from a power valve. Though serving as a useful guide in practice to the choice of load values in the case of triodes, yet a more accurate assessment of the load is required if distortion is to be kept to a minimum, since allowance must be made for the departure from linearity of the valve characteristics when operating near anode current cut-off. Moreover, using pentode power valves, the selection of a load $R = 2R_A$ gives far from satisfactory results. The approach adopted is to draw a load line, which is a line representing the voltage-current characteristic of the anode load, superimposed on the anode voltage-anode current characteristics for the valve in question. If the anode load is a resistance then, by Ohm's law, the load line is straight. For an impedance as anode load which is sensibly reactive in its effect, the load line will be an ellipse.

In general, two methods of approach are used. Either the load resistance value is known, and it is required to estimate the percentage harmonic distortion if the valve operating conditions are also specified, or alternatively, a certain power valve is to be used, and the graphical method outlined below is adopted as a means of estimating the best operating potentials for the valve, and the best load resistance value to use in order to keep the distortion satisfactorily low.

Suppose a triode power valve with a total H.T. supply of 400 V. is used. A limitation to the total power which can be dissipated is set by the maximum temperature at which the anode can be safely operated. Let this correspond to 16 W. maximum anode dissipation. The static anode voltage : anode current curves are known, as in fig. 62. The curve representing 16 W. dissipation is drawn on these characteristics by plotting $V_A I_A$ = constant, 16. When the anode current is zero (obtained in practice by setting the grid bias at cut-off), the voltage drop in the anode load is zero so the anode will be at the full H.T. voltage, 400. Hence the point A where $V_A = 400$, $I_A = 0$ is one point on the required load line. If class A conditions are to be observed, the peak positive grid input voltage just makes the grid potential zero or, at the most, slightly positive. Let B be the point where the zero grid voltage characteristic intersects the maximum dissipation curve. Then AB is the load line for maximum allowable power output. The slope of this load line obtained from $\dfrac{\text{AC in volts}}{\text{MC in amp.}}$ gives the corresponding resistance value to be used.

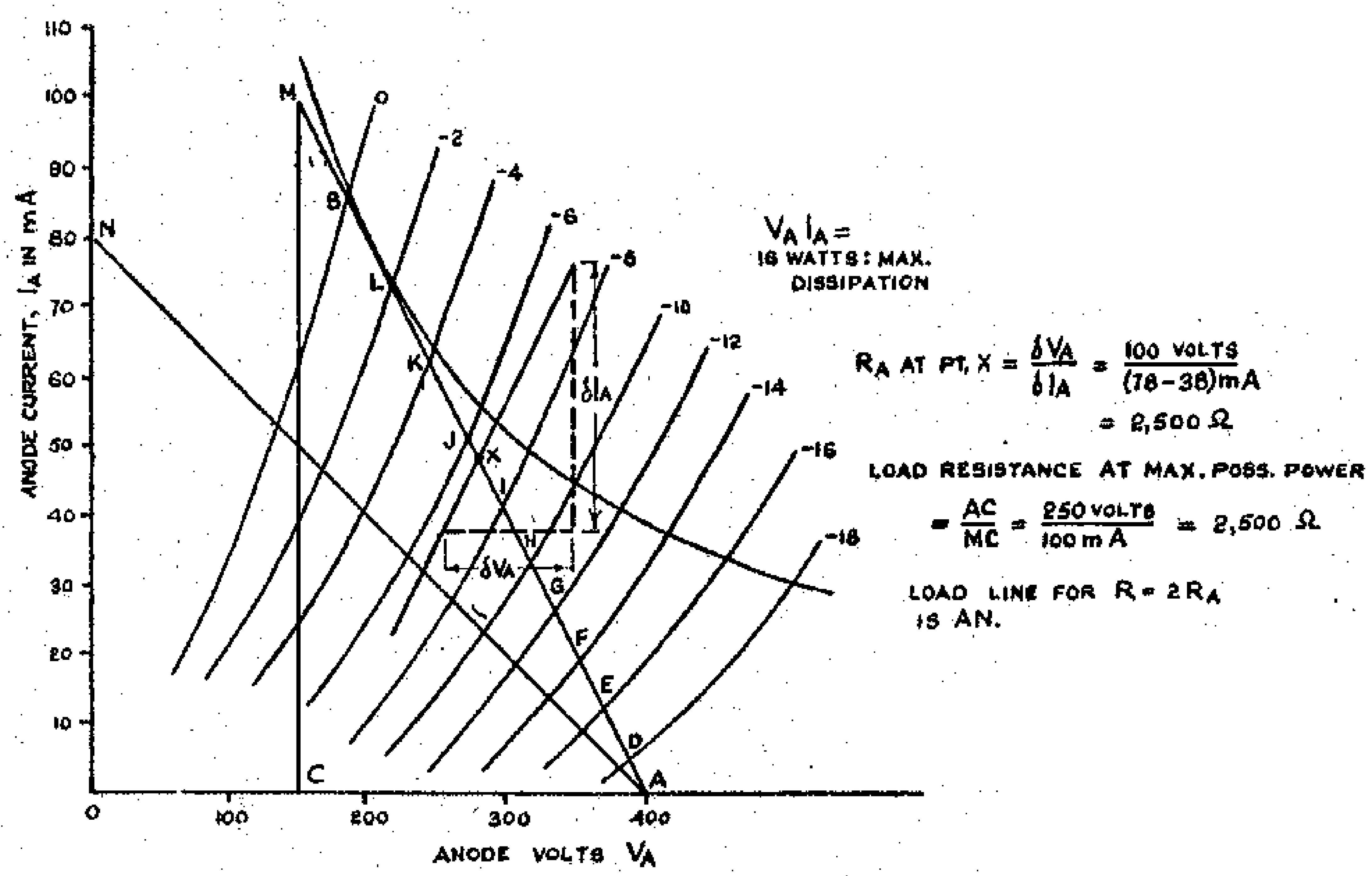


FIG. 62. The Load Line.

What is to be the criterion as regards distortion? Manifestly a small change of grid voltage anywhere in the region from $V_G=0$ to $V_G=$cut-off should bring about the same change in current through the anode load irrespective as to whether this grid voltage change takes place near cut-off bias, near the operating bias or near zero. Interpreted graphically, this means that the intercepts cut off along the load line by neighbouring $V_A : I_A$ characteristics should be equal for equal change of grid voltage in moving from one characteristic to the next, i.e. the intercepts $DE, EF, FG, GH, HI, IJ, JK, KL, LB$ (fig. 62) should all be equal. If the load line as drawn does not produce such equal intercepts then it must be varied. Graphically this implies swinging the load line about the fixed point A to arrange it below the line AB so that these intercepts are as nearly equal as possible. At the same time the slope of the line does not want to be too small (i.e. the resistance too high) otherwise the power dissipated in it will depart too seriously from the maximum possible. A useful, simple rule is to realise that the total percentage harmonic distortion will be less than 10% if the maximum departure from equality of the intercepts referred to does not exceed 10 : 11.

Having determined the best load resistance value, the operating bias can be determined as being at the point X, where X is the mid-point of the load line such that $DX=XB$. If a fuller investigation is then required, the dynamic mutual characteristic I_A-V_G, with the appropriate anode load, should be drawn to establish that this operating grid bias is midway between $V_G=0$ and the value of V_G at which this I_A-V_G characteristic commences to bend.

An examination of the most suitable anode loads for power pentodes and beam tetrodes in accordance with these conceptions indicates that, for maximum undistorted power output, the anode load needs to be $\frac{1}{8}$th to $\frac{1}{16}$th of the valve A.C. resistance.

Anode Efficiency. Defined as the ratio

$$\frac{\text{alternating power output}}{\text{D.C. power input}}.$$

For example, consider the case of a class A amplifier operated for maximum power output, disregarding distortion. It has been shown (p. 147) that then $R_L=R_A$. Obviously the anode efficiency will be 50%.

This can be proved in an alternative manner. Suppose a class A amplifier operates with steady anode voltage V_A and anode current I_A. Then the D.C. power input is $V_A \cdot I_A$. The maximum positive to negative peak excursion that the anode voltage can execute by virtue of an alternating input to the grid will be from V_A to zero, and to a positive maximum of $2V_A$. Correspondingly the maximum possible anode current excursion will be $2I_A$. But the R.M.S. value of the alternating anode voltage will then be $V_A/\sqrt{2}$, and for the current $I_A/\sqrt{2}$. Consequently the alternating power output is $\dfrac{V_A \cdot I_A}{\sqrt{2}\cdot\sqrt{2}}=\dfrac{1}{2}V_AI_A$, and the anode efficiency is 50%.

Class B Amplification. A maximum possible efficiency of 50% is a limitation to the power handling capabilities of a valve, especially in the case of radio transmitters where, with apparatus of restricted size, it is required to supply the aerial with as much energy as possible. For this reason class B and class C amplification practices have been developed. These involve using the valve with a steady, operating grid bias at, or beyond, cut-off.

Such practice is inadmissible in cases where a single valve is used with a resistance as anode load, since the negative half-cycle of the input does not then produce any corresponding anode current change: rectification is involved. However, if the anode load is an oscillatory rejector circuit, then the introduction of such distortion is not of importance. Fortunately such anode loads are those most useful in R.F. power amplifier practice.

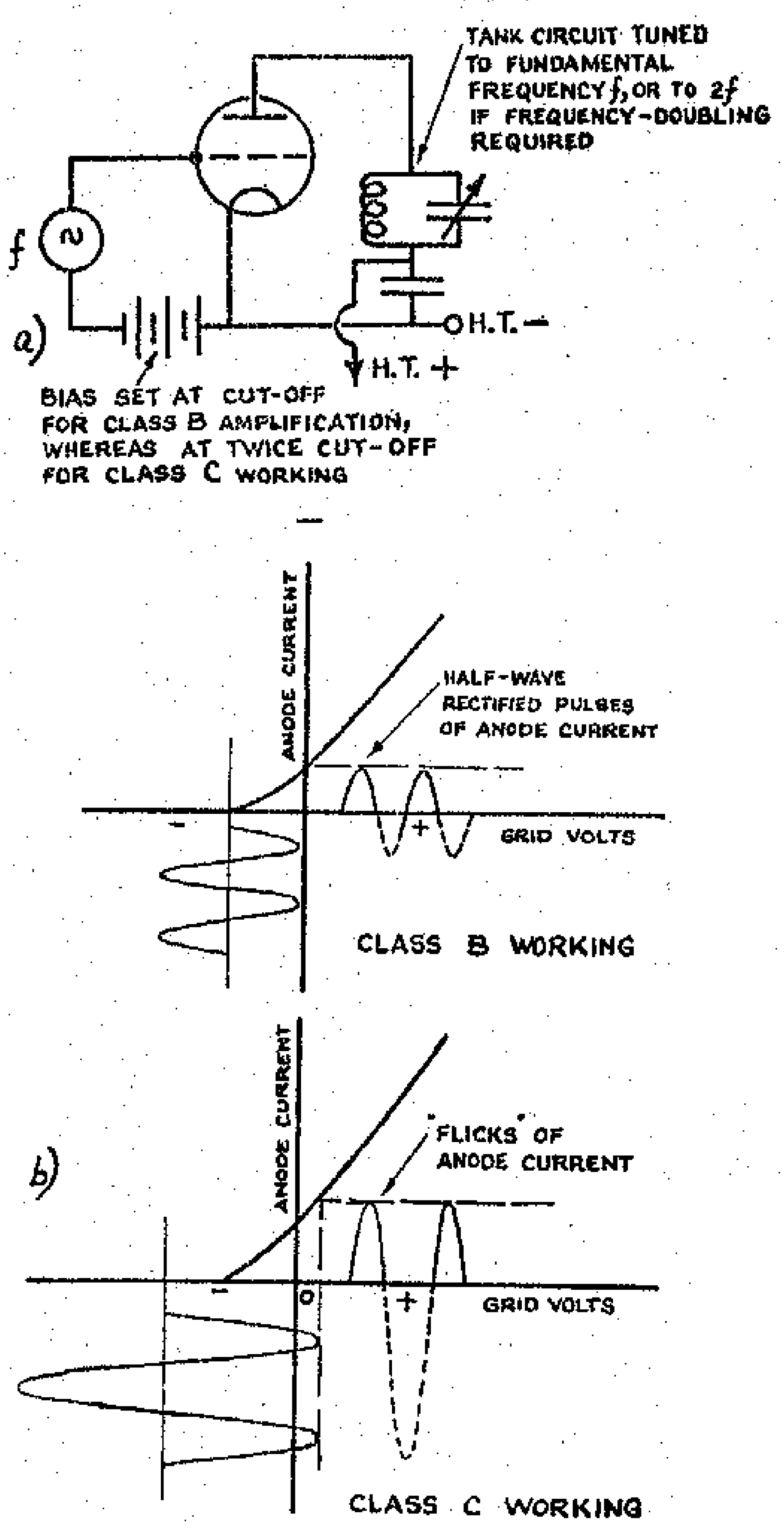


FIG. 63. Class B and Class C Amplification.

Though the anode current variations of such amplifiers have wave-forms which are rectified and distorted replicas of the input wave-form to the grid, yet such distortion necessarily corresponds to the introduction of second and higher order harmonics (see p. 69), whereas the rejector circuit anode load will only respond to the fundamental. The deliberately introduced distortion is, therefore, of no consequence as regards the ultimate output wave-form. Again, if the distortion introduced is predominantly second harmonic distortion, as in the case of a class B operated triode, then a class B push-pull amplifier, which eliminates even harmonic distortion, is valid (see p. 154), though the anode load is a resistance.

From the analysis given on p. 71, it is seen that the wave-form in the case of half-wave rectification possesses a first harmonic component of which the peak value is approximately half the maximum voltage attained. Hence, if the grid alternating input is approximately doubled, in the class B case (fixed bias at cut-off) compared with the class A case, then the fundamental component of the rectified output is capable of supplying the same power to a rejector circuit as would be supplied by a class A amplifier.

The anode efficiency of the class B amplifier is, however, greater because there is a smaller demand on the H.T. supply than when class A working operates. This is because anode current only flows when the alternating grid input is positive, and not when it is negative.

From the analysis given on p. 71, the average value of anode current during one-half cycle is seen to be $2I/\pi$, where I is the peak current. But during every alternate half-cycle the current is zero. Therefore the average current drain on the H.T. supply in class B working is I/π. In class A working, the mean D.C. anode current is $I/2$. Hence class B working demands an average current supply which is only $2/\pi$ of that in class A practice. Consequently a class B amplifier is $\pi/2$ times as efficient, giving a maximum possible efficiency of $\pi/2 \times 50\% = 78\%$.

Class C Amplification. The operating fixed negative bias can be made greater than cut-off, still further restricting the fraction of the input cycle time during which anode current flows. Thus in class C working, the grid bias is as much as twice cut-off value, giving anode current pulses for less than one-third of the operating time, with a consequent anode efficiency of as much as 85%. The considerable second and third harmonic distortion introduced restricts the use of such amplifier practice to solely those cases in which a rejector circuit is used as anode load. Push-pull working is now inadmissible, owing to the odd harmonics involved.

R.F. Power Amplification. To amplify the output of an oscillator, as in the usual radio transmitter master-oscillator, power amplifier system, a class C amplifier is commonly used. Apparatus with a total power output in excess of 250 W. commonly makes use of triode valves, but pentodes are frequently encountered in smaller gear.

To ensure the maximum anode efficiency, the alternating voltage applied to the grid (grid drive) is made great enough to produce a positive grid at the positive peak grid input volts. Thus the input circuit to the amplifier needs to be able to supply a moderate amount of power. When the grid is most positive, the valve anode current will be large, and the alternating voltage across the tuned circuit anode load will be at its maximum. Since the anode voltage alternates in anti-phase to the grid drive volts (see p. 119), the maximum positive grid voltage will occur simultaneously with the minimum possible anode potential, so that overheating of the anode due to the large current flicks brought about by driving the grid positive is not excessive.

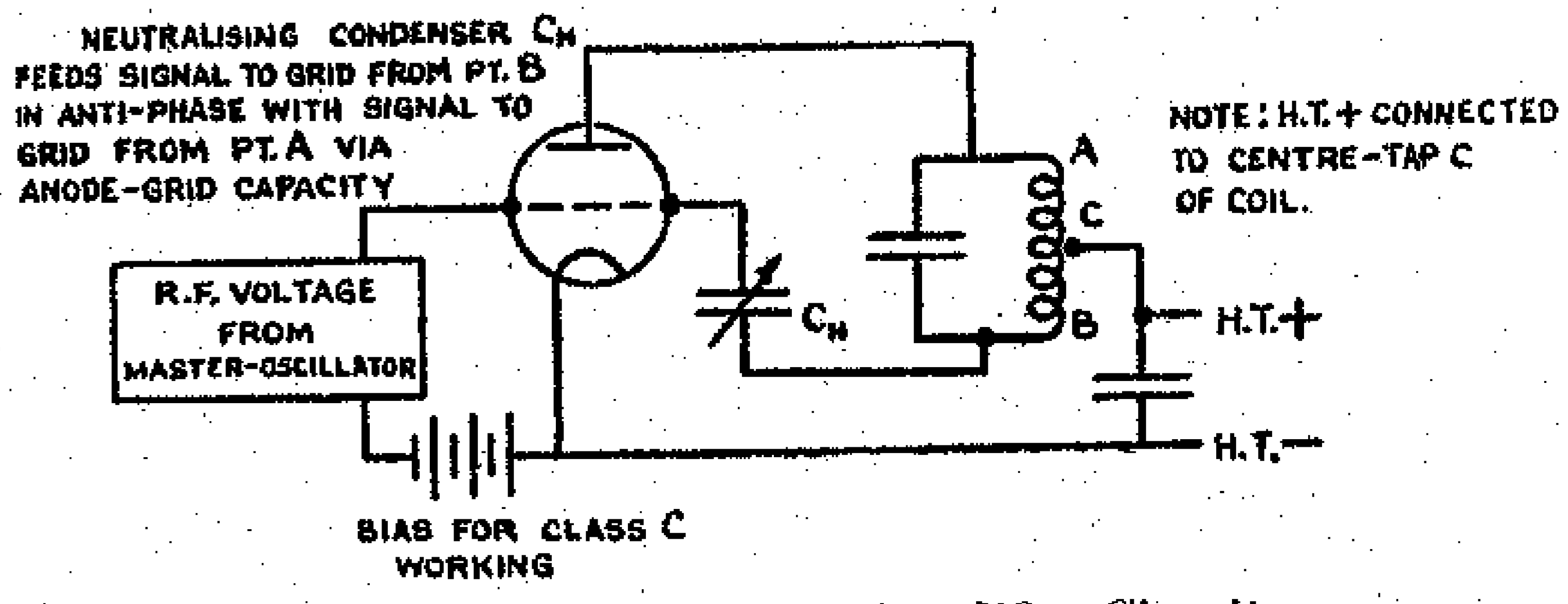


FIG. 64. R.F. Power Amplifier Circuit.

Such R.F. power amplifier circuits need neutralising, as is indicated in fig. 64, to prevent the power amplifier from going into oscillation due to feedback via the anode-grid capacity from the anode tuned circuit to the grid input tuned circuit.

Push-pull Amplification. A circuit arrangement which eliminates distortion at the second and other even harmonics which may be introduced by the valve characteristic, in which a pair of valves is used, generally as a power amplifier, where the grid inputs to the two valves are equal in magnitude, but differ in phase by 180°.

Suppose an A.C. input of sinusoidal wave-form is applied at the primary of the input transformer (fig. 65). Since the centre-tap of the secondary of this transformer is connected to fixed bias battery, voltage V_G, so the grid of valve V_1 will have potentials decided by $-V_G$ plus the alternating voltage across half-secondary A_1C_1, whereas V_2 grid will vary in accordance with $-V_G$ plus the half-secondary voltage across B_1C_1. But the potential at point A_1 will be positive with respect to C_1 when the potential of B_1 is negative to the same extent. Therefore the grids of V_1 and V_2 will have potentials varying in anti-phase.

A decrease of the negative potential on V_1 grid will cause the anode current of V_1 to rise, giving a rise of potential across the half-primary of the output transformer C_2A_2, which is the anode-load presented to V_1. The potential of point A_2 must therefore decrease, since the potential of the centre point C_2 is constant at H.T. potential. This action will necessarily be accompanied by a

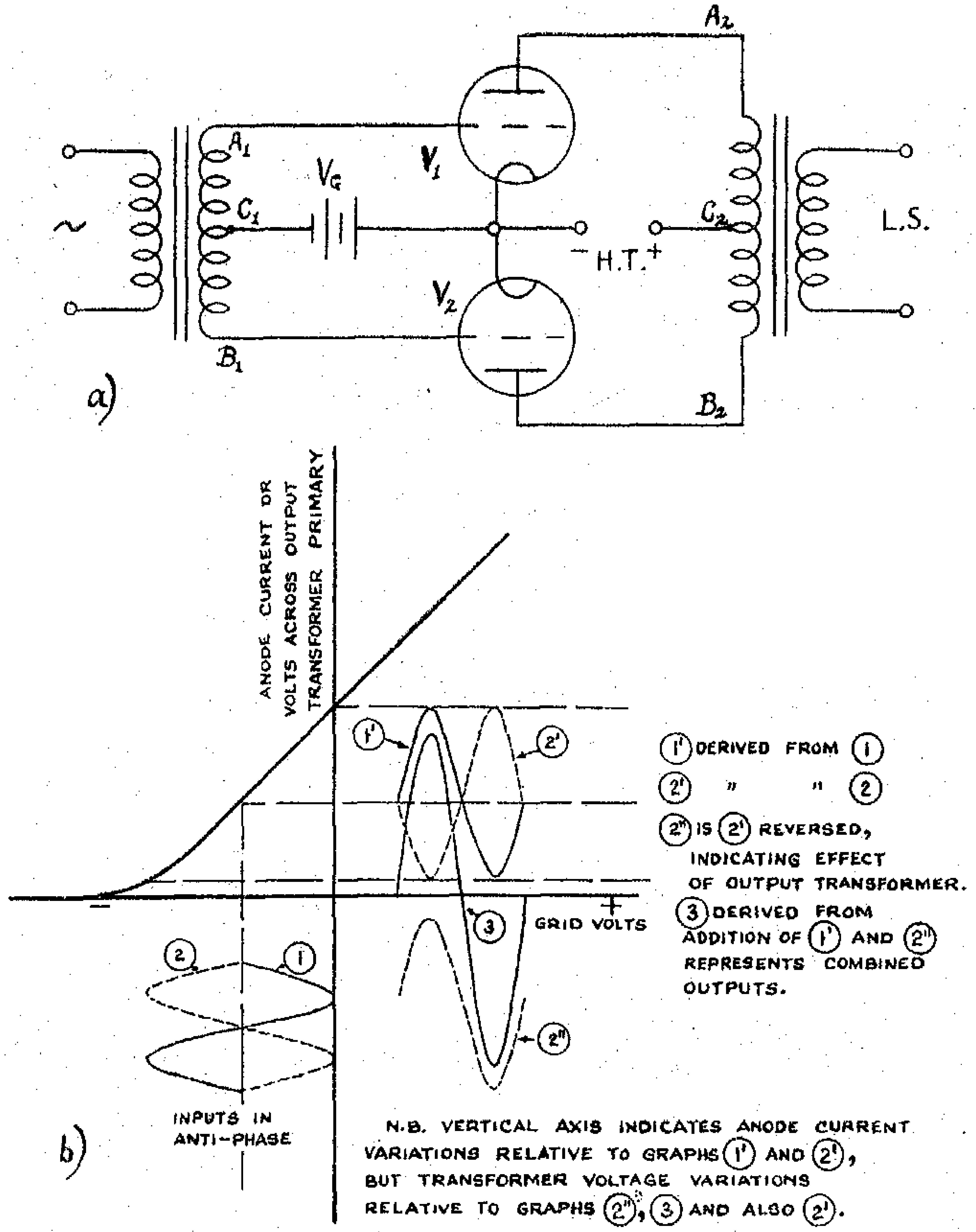


FIG. 65. *a*, Push-pull Circuit. *b*, Graphical Illustration of Push-pull Action.

corresponding increase of the negative potential of V_2 grid, producing a fall of the anode current of V_2. The voltage across the half-primary C_2B_2 of the output transformer must consequently decrease, implying an increase of the potential at point B_2, since the potential of centre-tap C_2 is fixed. The total potential variation across the whole of the output transformer primary A_2B_2 is then double the potential variation across either half, A_2C_2 or B_2C_2, since A_2 potential rises with respect to C_2 to the same extent to which B_2 falls. Hence the two valve outputs are effectively combined across the primary, giving a double output across the secondary, which is connected to a resistive load, such as a loudspeaker.

The D.C. components of the valve anode currents through the output transformer primary are in opposite directions, so the nett D.C. magnetisation of the transformer core is practically zero. The effect of D.C. magnetic saturation of the transformer core causing distortion is therefore eliminated.

A second, and more important, advantage of the push-pull technique is that second and other even harmonic distortion, due to lack of linearity of the valve characteristic over the operating region, is eliminated. Since triode valves produce distortion due to characteristic curvature which is almost exclusively at the second harmonic, so the grid bias may be set at the bend, or even at the valve cut-off region, giving class B working conditions with improved efficiency, and little distortion (see p. 151).

That even harmonic distortion is reduced to zero may be realised on considering a simple mathematical analysis of push-pull action. Suppose the relation between the valve anode-current I_A and the grid potential E_G is given by the general relationship

$$I_A = A + BE_G + CE_G^2 + DE_G^3, \text{ etc.} \qquad (183)$$

The alternating grid potential on one valve will be of the form $V_0 \sin \omega t$, whilst that on the other valve will be represented by the anti-phase voltage, $-V_0 \sin \omega t$, V_0 being the peak potential across half the input transformer secondary, $\omega/2\pi$ being the frequency of the input voltage.

Hence for valve V_1

$$I_A = A + B(V_0 \sin \omega t) + C(V_0 \sin \omega t)^2 + D(V_0 \sin \omega t)^3 + \text{etc.} \qquad (184)$$

and for valve V_2, *which must have exactly the same shape characteristic,*

$$-I_A = A + B(-V_0 \sin \omega t) + C(-V_0 \sin \omega t)^2 + D(-V_0 \sin \omega t)^3 + \text{etc.} \qquad (185)$$

Note I_A is written with negative sign here, since it is varying in opposition to the current from valve V_1.

The combined output across the output transformer primary is $2I_AZ$, where Z is the primary impedance, and $2I_A$ is given by subtracting equation (185) from equation (184).

$$\therefore \quad 2I_A = 2BV_0 \sin \omega t + 2C(V_0 \sin \omega t)^3 + 2E(V_0 \sin \omega t)^5 + \text{etc.}$$

Note that the squared terms, and other terms raised to an even power, vanish in this expression which determines the effective output, since $(-V_0 \sin \omega t)^2 = V_0^2 \sin^2 \omega t$. Moreover,

$$V_0^2 \sin^2 \omega t = V_0^2\left(\frac{1+\cos 2\omega t}{2}\right),$$

corresponding to a current component at the second harmonic of the input signal frequency. It is seen that such second, and other even harmonic distortion introduced by the curvature of the valve $I_A - V_G$ characteristics, are therefore eliminated.

It is noteworthy, however, that the percentage of third and other odd harmonic distortions remains the same as if a single valve were used. Since such third harmonic distortion is not noticeably introduced by triode valves, this push-pull method lends itself admirably to the design of receiver and other amplifier push-pull output stages where freedom from distortion is essential. Pentode valves, however, are prone to third harmonic distortion, and should therefore be used with caution in a low-frequency power amplifier push-pull arrangement.

Application of Feed-back to an Amplifier. If a fraction of the output voltage from a valve amplifier is fed back to be placed in series with the input circuit, then desirable or undesirable effects can be produced, depending on the phase of the feed-back voltage with respect to the input signal, and the purpose of the amplifier. If the feed-back is so arranged as to be in anti-phase with the initial input to the amplifier, then *negative* feed-back is achieved. A feed-back that is in the same phase as the input gives *positive* feed-back. Negative feed-back is also called "reverse" or "degenerative"; positive feed-back is also known as "regenerative", or "reaction".

If E=input voltage to the amplifier which is combined with a series feed-back voltage βE_0 derived from the output circuit (see fig. 66*a*), then the actual voltage to the amplifier input is

$$E_G = E \pm \beta E_0. \qquad (186)$$

A plus or minus sign is attached to β depending on whether the feed-back is positive or negative.

The nominal gain, m, of the amplifier, is given by

$$m=\frac{\text{output voltage}}{\text{input voltage}}=\frac{E_0}{-E_G}\left(=\frac{\mu R}{R+R_A}\text{ if resistive anode load used}\right). \quad . \quad . \quad . \quad (187)$$

A negative sign precedes E_G, since the anode voltage of an amplifier varies in anti-phase with the input grid voltage (see p. 120).

$$\therefore \quad E_G=\frac{-E_0}{m}. \quad . \quad . \quad . \quad . \quad . \quad (188)$$

Substituting for E_G in (186) from (188)

$$\frac{-E_0}{m}=E\pm\beta E_0.$$

$$\therefore \quad E_0(1\pm m\beta)=-mE.$$

$$\therefore \quad E_0=\frac{-mE}{1\pm m\beta}. \quad . \quad . \quad . \quad . \quad . \quad (189)$$

The nett gain, including the effects of feed-back, m_f, is given by

$$m_f=\frac{\text{output voltage}}{\text{signal input voltage}}=\frac{E_0}{-E}.$$

Therefore from (189) $\quad m_f=\dfrac{m}{1\pm m\beta}$. . . . . . (190)

If $m\beta$ is made considerably greater than unity, then (190) becomes

$$m_f=\pm\frac{1}{\beta}\text{ approx.} \quad . \quad . \quad . \quad . \quad (191)$$

indicating that the overall gain of the amplifier depends only on the factor β, which is independent of the characteristic of the valve, the effects of valve "noise" and variations in supply voltage, but depends only on the feed-back network. The gain of the amplifier is, however, much reduced by such feed-back.

If $m_f=+1/\beta$, in the case of positive feed-back, then the feed-back voltage is in phase with the input voltage. Such an amplifier is generally unstable, since any increase of the output voltage

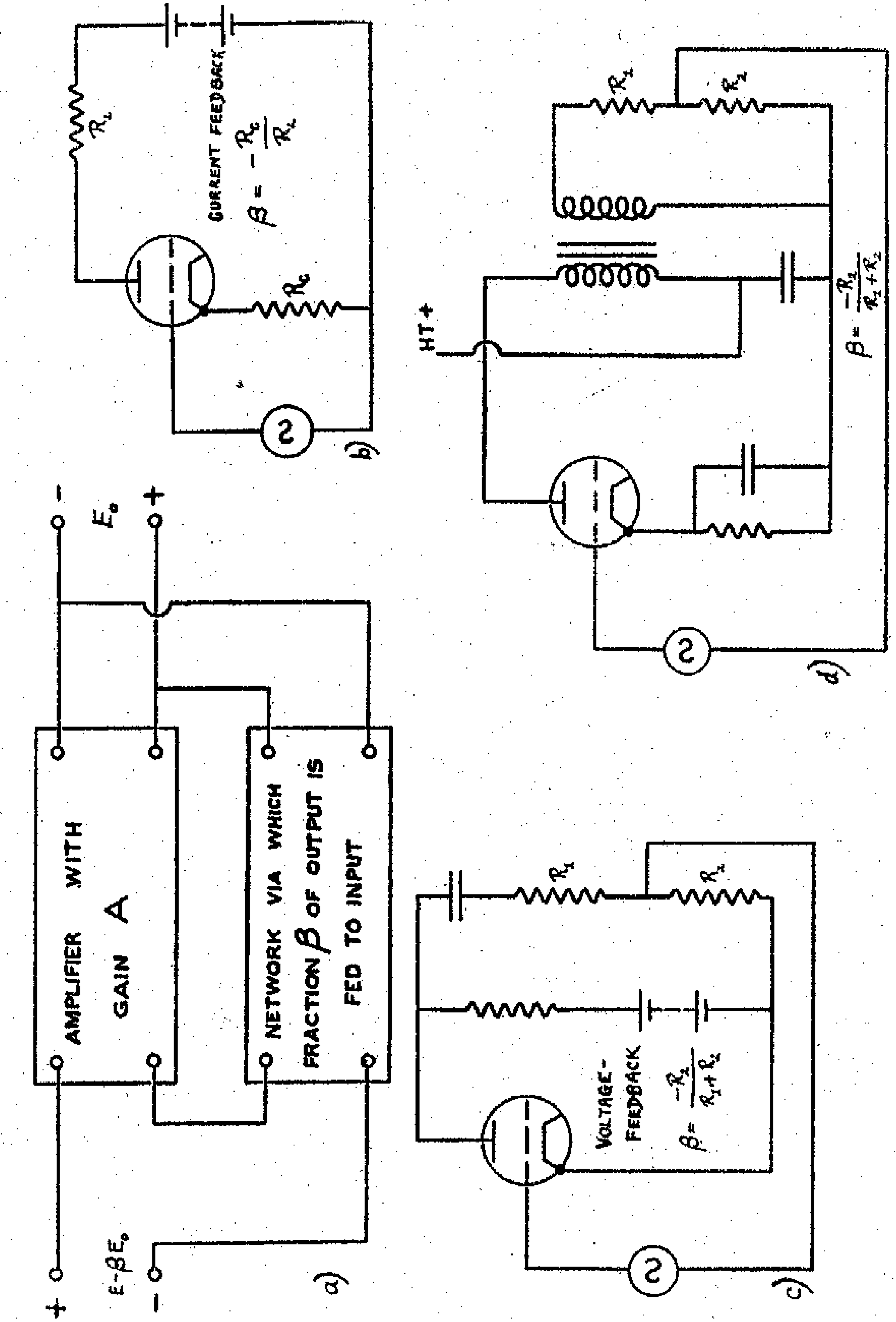


Fig. 66. Negative Feed-back Amplifier Circuits.

causes an increase of the total input signal, causing the output voltage to increase still further, the effect being cumulative.

If $m_f = -1/\beta$, in the case of negative feed-back, then the amplifier is much more stable than is the case of the amplifier without feed-back. This is particularly so the larger the value of $m\beta$, since then the actual input voltage to the amplifier is a small difference between comparatively large signal and feed-back voltages. If the gain m, due to the valve, changes, then the difference between signal and feed-back voltages changes, increasing if m decreases, and vice versa. The actual input voltage therefore changes in such a manner as to compensate for change of the gain m.

From equation (189) $\quad \dfrac{E_0}{E} = \dfrac{-m}{1-m\beta}$.

The change in gain (E_0/E) caused by a change in amplifier nominal gain m is given by evaluating $\dfrac{d(E_0/E)}{dm}$.

$$\frac{d(E_0/E)}{dm} = -\frac{d[m/(1-m\beta)]}{dm}$$

$$= -\left[\frac{(1-m\beta)+m\beta}{(1-m\beta)^2}\right] = \frac{-1}{(1-m\beta)^2}. \qquad (192)$$

Hence the change in gain with feed-back for a given change in gain without feed-back is decreased as β is increased negatively, i.e. as the negative feed-back is increased. Vice versa, if β is increased positively, in the case of positive feed-back, then the amplifier stability becomes poorer. This equation (192) indicates also how negative feed-back reduces amplitude distortion, since such distortion results generally from variations of m during the operating time.

The circuits involved in negative feed-back are of (*a*) the current feed-back type in which the voltage inserted at the input is proportional to the current in the load, (*b*) the voltage feed-back class where the voltage inserted at the input is proportional to the voltage across the load and (*c*) current-voltage feed-back in which a combination of (*a*) and (*b*) is used. These circuits are illustrated in fig. 66.

Since the gain of these amplifiers is decided by $1/\beta$, in the case where β is a considerable fraction $(m\beta > 10)$, and the network whereby β is achieved may be other than a resistive arrangement, such as a combination of inductance and resistance or capacitance and resistance, so an amplifier of desired frequency-response can be designed, whereas if a resistive feed-back circuit is used, the gain is largely independent of frequency, except for the influence of stray inductance and capacity.

Noise reduction (see p. 164), is brought about in a negative feed-back amplifier. However, this reduction is only for that arising within the valve concerned. Any noise present at the input to the amplifier will be present to the same extent in the output. The reduction of noise as regards that introduced by the amplifier valve itself may be expressed as

$$\frac{\text{Signal to noise ratio with feed-back}}{\text{Signal to noise ratio without feed-back}} = \frac{m_f}{m(1-m\beta)}.$$

The Cathode Follower. If a negative feed-back amplifier is used in which the feed-back is obtained by the use of a resistance in the cathode circuit, as in fig. 67*a*, but the anode load is omitted, the anode being connected directly to the H.T.+ supply, the equation (190) $m_f = m/(1 \pm m\beta)$ is suitably modified by putting $m = \mu R_C/(R_C + R_A)$, where μ and R_A are the constants of the valve used, R_C is the cathode resistance (cf. equation 166) and $\beta = 1$.

$$\therefore \quad m_f = \frac{\mu R_C/(R_A+R_C)}{1+\mu R_C/(R_A+R_C)} = \frac{\mu R_C}{R_A+(1+\mu)R_C}. \qquad (193)$$

The gain of such an amplifier is thus necessarily less than unity. It is therefore used as a current amplifier and not as a voltage amplifier, of which the purpose is to feed the grid with a high impedance input but in which the cathode load can be a low impedance output, yet without serious loss of voltage.

From equation (193),

$$I_A = \frac{\mu E_G}{R_A + R_C(\mu+1)}, \qquad (194)$$

where E_G is the alternating grid input, and I_A the alternating anode current, R.M.S. values.

Dividing numerator and denominator of this expression by $(\mu+1)$ gives

$$I_A = \frac{[\mu/(\mu+1)]E_G}{R_A/(\mu+1)+R_C}. \qquad (195)$$

By comparison with the circuit of fig. 49*c*, it can be seen that this cathode-coupled stage corresponds to a normal amplifier in which the amplification factor of the valve μ is reduced to $\mu/(\mu+1)$, and the A.C. resistance R_A is lowered to $R_A/(\mu+1)$, the equivalent circuit becoming that shown in fig. 67*b*.

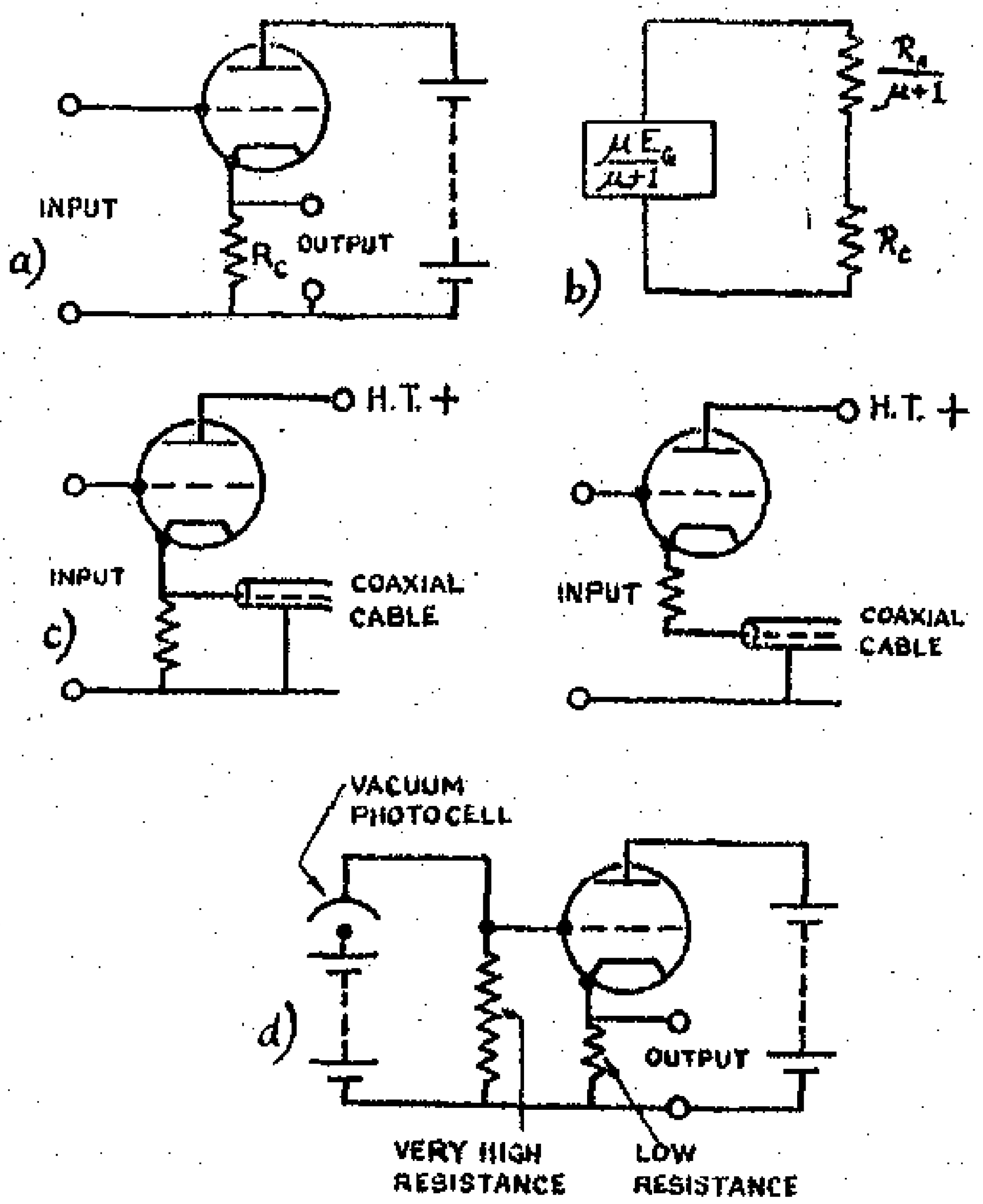


FIG. 67. *a*, The Cathode Follower. *b*, Equivalent Circuit for Cathode Follower. *c*, Cathode Follower used to couple to Transmission Line, *d*, Cathode Follower as D.C. amplifier.

The output resistance R_0 is effectively the valve resistance and R_C in parallel, and equals

$$\frac{R_C[R_A/(\mu+1)]}{R_C+R_A/(\mu+1)}=\frac{R_CR_A}{R_A+R_C(\mu+1)},$$

which in the case of a pentode,* where $\mu \gg 1$, becomes

$$R_0=\frac{R_C}{1+R_Cg_m}, \qquad (196)$$

on employing the relationship $g_m=\mu/R_A$. For example, in the case of a pentode, suppose $\mu=1000$, $R_A=250$ kΩ., $R_C=1000$ Ω. and $g_m=4$ mA./V.$=0{\cdot}004$ ohms, then

$$R_0=\frac{R_C}{1+1000\times0{\cdot}004}=\frac{1000}{5}=200\,\Omega.$$

$$\text{and the gain}=\frac{\mu R_C}{R_A+R_C(\mu+1)}=\frac{1000\times1000}{250{,}000+1000(1001)}=0{\cdot}8.$$

This low impedance output, one side of which is at earth potential, makes such circuits eminently suitable for coupling the output of an oscillator to a transmission line or aerial circuit in a transmitter, where a careful match is required between the output circuit and the load (see fig. 67*c*). Moreover, because of the low output impedance, the output voltage has good regulation where the input to the cathode-follower stage may have poor regulation. Another advantage of this circuit is that there is no polarity inversion: the output voltage variation is in phase with the input alternating voltage.

The cathode follower circuit also exhibits reduced Miller effect (cf. p. 134), the effective input capacity being reduced in accordance with the relationship

$$\text{Equivalent input capacity} =\text{actual input capacity } (C_{GC})\times\left(1-\frac{E_0}{E_G}\right).$$

A cathode follower may be used with advantage in photoelectric cell amplifiers, since this circuit technique enables the grid resistance across the input to be very high, yet with a low output cathode resistance. Thus using ordinary valve types, the presence of positive ion current and grid emission prohibits the effective use of an input resistance greater than 2 MΩ. (see p. 119). Suppose the photocell current is $0{\cdot}05\,\mu$A., then a D.C. voltage input to the amplifier of $0{\cdot}05\times2=0{\cdot}10$ V. is achieved on switching on the cell illumination. If the D.C. amplifier is normal (fig. 48) then a voltage gain of 100 times, and so an output of 10 V. is easily obtained. Compare the use of a cathode-follower amplifier. Even using normal triode and pentode valve types, the input resistance can now be as much as 100 MΩ., giving an effective input voltage of $0{\cdot}05\times100=5$ V. If a gain of 0·8 is realised, then the output voltage across, say, a 5 kΩ. cathode load will be 4 V. It would

* In using a pentode in a cathode-follower circuit, the screen must be kept at constant potential with respect to the cathode.

seem as though the cathode technique was thus only 40% as effective as the provision of the ordinary D.C. amplifier. In practice, however, great benefit is obtained from the use of the cathode-follower because it is virtually insensible to the supply voltage fluctuations and circuit variations which make normal D.C. amplifier practice so tedious and unreliable.

"Noise" in Amplifiers. It would seem that a consideration of the methods of coupling both A.C. and D.C. amplifiers in cascade would lead to the belief that there was no limit to the total amplification possible: it was simply a matter of ensuring constant voltage supplies and using a sufficient number of valve stages coupled together to obtain amplifications of several million. In practice this is not so: a limit is set by the "noise" which arises in the amplifier, due to minute random fluctuations of the current. These are particularly obnoxious in the first stage, since it is there that the true signal is small, and the "noise" becomes of comparable magnitude, giving a signal to "noise" ratio which is sufficiently small to make the eventual output unintelligible.

This "noise" arises in two places:

(*a*) Thermal agitation noise, or Johnson noise, brought about by fluctuations of the current in ordinary electrical conductors due to the thermal agitation of the conducting particles. In this connection the input resistance to the first amplifier is usually the only one which need be of concern, because it is only there that the signal current is sufficiently small for the "noise" current to be of comparable magnitude.

An electric current is due to the motion of electrons through conducting material, and the current due to an applied signal voltage is combined with the small currents due to the thermal agitation of the electrons. During any long period of time the nett effect of these thermal currents is zero. At any given instant, however, this is not the case—there may be a slight preponderance of random electron motion in one direction over that in the other. Such transient currents are of very complex wave-form, and produce voltages across the resistance of which the harmonic components extend over a very wide frequency band.

An expression for thermal "noise" is

$$E_N{}^2 = 4kT\int_{f_1}^{f_2} R\,df. \qquad (197)$$

Due to Nyquist,* this equation gives E_N, the R.M.S. value of the E.M.F. produced in series with the resistance of the conductor R, where k is Boltzmann's constant, T is the absolute temperature, f is the frequency and R is the actual or equivalent shunt resistance of the circuit, being L/CR in the case of a rejector circuit.

Integrating this equation over the frequency range (f_1-f_2) gives

$$E_N = \sqrt{[4kTR(f_1-f_2)]}. \qquad (198)$$

Substituting for k, then at normal room temperatures

$$E_N = 1{\cdot}25\times10^{-10}\sqrt{[R(f_1-f_2)]}. \qquad (199)$$

An immediate partial remedy to the elimination of this noise voltage is to restrict the band width (f_1-f_2) which the amplifier passes. If $R=100$ kΩ., and $(f_1-f_2)=10$ kc./s., then substitution in (199) gives

$$E_N = 1{\cdot}25\times10^{-10}\sqrt{(10^5\times10^4)} = 4\ \mu V.\ \text{approx.}$$

(*b*) Shot or "schrot" noise. First investigated by Schottky,† is due to a comparable fluctuation of the current in valves due to the random emission of individual electrons. Thus the anode current of a valve is constant considered over any finite time interval, but at a particular instant the number of electrons arriving may be slightly different from the number at another instant, though the average number is constant. In other words, "shot noise" is due to the finite magnitude of the electron charge motions which constitute an electric current, where the number of electrons flowing per second will be constant, but where the number per microsecond will fluctuate slightly about the mean value.

Added to this inevitable cause of current variation, there is a second source of fluctuation which is decided by the physical nature of the emitting cathode surface. This is called "flicker effect", and is particularly prevalent if the cathode surface is not smooth, but pitted with minute holes which occasionally release bursts of electrons. Added to this there are noise effects due to ionisation of the residual gas in the valve and due to secondary emission from the valve electrodes.

Shot noise, considered as separate from flicker effect, etc., can be expressed by the equation

$$I_N^2 = 2I_A e(f_1 - f_2) \qquad . \quad . \quad . \quad (200)$$

in the case of a saturated diode, where I_A is the anode current, and I_N is the R.M.S. value of the variation of I_A.

In the case of non-saturated triode and multi-electrode valves, the effect is reduced by the presence of space-charge, which acts as a cushion, or reservoir of electrons. Thus valves in which the space-charge effect is high are preferable for reduction of shot noise. In these cases, the effect of total noise is stated as the equivalent resistance which would give the same noise voltage due to thermal effects, and the magnitude of this resistance is proportional to I_A/g_m^2. Hence a valve of large mutual conductance g_m for a given anode current I_A, is necessary for low noise levels.

The triode valves are most free from this noise defect. The pentode has an equivalent shot noise resistance about five times that for the corresponding beam tetrode. This is because the beam tetrode has low screen current, high space-charge effect between beam plates and low secondary emission. Frequency-changer valves (see p. 210) are bad because their conversion conductances are low, and they also produce noise by virtue of their oscillator section. Thus it is preferable to use an R.F. beam tetrode amplifier before the frequency-changer in a super-heterodyne receiver if a particularly high signal/noise ratio is required.

A minimum signal/noise ratio of 5 to 6, or some 15 db.* is desirable in amplifier and receiver practice, where the noise concerned is the total noise brought about by the above effects.

Supply Voltage Regulation. If electronic circuits, such as amplifiers, are used in any kind of measuring or indicating device where a constant output reading is required from day to day for the same input conditions, then it becomes essential to pay considerable attention to the question of the supply of constant valve filament and anode voltages. This is particularly the case where the H.T. and L.T. supplies depend ultimately on the A.C. or D.C. mains, as in power-pack practice.

* Decibels=db.$=10 \log_{10}(P_1/P_2)$, where P_1 and P_2 are two powers,
$=20 \log_{10}(V_1/V_2)$, where V_1 and V_2 are two voltages.
$\therefore$ 15 db.$=20 \log_{10}(V_1/V_2)$. $\therefore$ $V_1/V_2=\text{antilog}(0{\cdot}75)=5{\cdot}63$.

Apart from manual control of a potentiometer, or variac across the mains, bringing an indicating voltmeter to a constant reading, there are four ways in which stabilisation of a voltage (or current) supply can be arranged. They are (*a*) by means of a constant-voltage transformer; (*b*) using a barretter; (*c*) by gas-filled discharge tube, or stabilovolt; (*d*) by use of thermionic vacuum tubes. In all these devices an unavoidable difficulty arises in that the voltage (or current) change must occur before it can be compensated; it is a desirable feature of a stabiliser, therefore, that such compensation takes place with as little time-lag as possible. In this connection, and in other ways, the use of thermionic valves as regulating means are the most successful.

(*a*) Constant voltage transformers employ magnetic saturation to achieve a constant A.C. voltage output. Thus transformers are available with ratings up to 25 kW. which give a secondary voltage constant to within $\pm 1\%$ for variations of primary voltage up to $\pm 15\%$. In such transformers the usual arrangement is for two transformers to be used with their primary windings in series additively, and their secondary windings in series with the load, but so wound that the A.C. voltage outputs are in anti-phase with each other. One of these transformers has an iron core which is partially magnetically saturated over the range of primary voltage required, and gives the larger secondary voltage. When the A.C. voltage applied increases in magnitude the fraction of the total applied to the saturated transformer decreases. Thus the secondary voltages become more nearly equal, it being a matter for the designer to arrange that such change of output compensates exactly for the primary supply change. Frequently the saturated transformer has a condenser connected across its secondary to improve the regulation, and to correct the wave-form. The chief difficulty with such arrangements is that the output wave-form is no longer sinusoidal, and in some circuits the change of wave-shape with change of supply voltage may be as disturbing as a change of voltage would be. Again, if the frequency of the A.C. mains supply varies, then the voltage variations from such transformers are worse than from an ordinary transformer. Transformers are available, however, which are guaranteed to be unconscious of both primary voltage and frequency changes within certain limits.

A constant voltage transformer feeding the primary of an ordinary transformer power pack can regulate the ultimate H.T. supply from such a pack. Again, these transformers can be used for maintaining constant A.C. heater voltages, or lamp filament supplies.

(*b*) The barretter is somewhat like an ordinary electric lamp of which the filament is iron wire operating in a hydrogen atmosphere. The filament and hydrogen pressure are so adjusted that the current through the lamp is largely independent of the supply voltage. The commonest use of these barretters is in *universal* A.C.-D.C. mains receivers where a barretter is placed in series with all the valve heaters across the A.C. mains supply, fig. 68, barretters being available which have a current capacity equal to the current rating of the 13 V. heaters commonly employed in A.C.-D.C. valves.

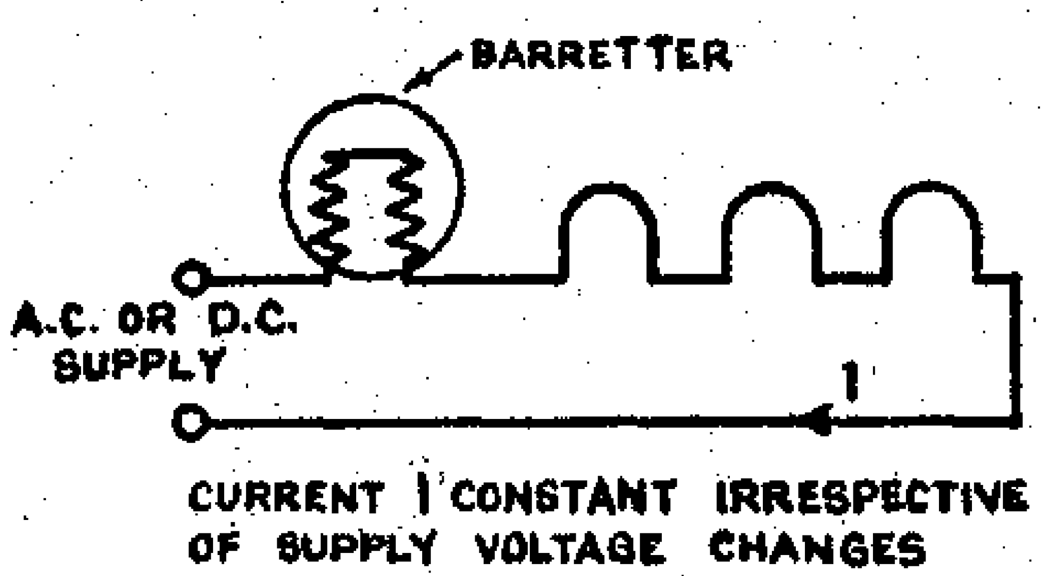


FIG. 68. The Barretter.

The iron filament presents a resistance to the supply E.M.F. which varies in such a manner that the series current is constant. Thus if the supply E.M.F. increases the current will tend to rise. However, such current increase causes greater heating of the filament, and if the filament resistance temperature coefficient is correct, then the increased filament resistance can compensate for the increased supply voltage. This coefficient is arranged to be correctly compensative over a wide range of voltages by the use of iron as the resistance element, in conjunction with the rate of conduction of heat away from the filament depending on the pressure of the surrounding hydrogen atmosphere.

The time lag for compensation to occur is a snag in using this method of stabilising. Again, the glass bulb of the barretter takes more than half an hour to reach an equilibrium temperature. Apart from its great use in providing valve heater currents in universal receivers, the author has found that barretters are of little use in trying to arrange constant current supplies for any purpose where laboratory measurements are to be made.

(*c*) Gas-discharge tube regulators, such as the neon lamp and the specially prepared stabilovolt, are of considerable value in achieving constant H.T. supplies. This is due to the peculiar characteristic of the two-electrode tube filled with an inert gas at low pressure whereby, once a glow discharge begins, the voltage across the tube remains nearly constant irrespective of the current through the gas.

The current through the tube increases with voltage until the striking potential is reached. Thereafter the voltage drop across the tube falls to a lower value, and further current increase is accompanied by practically no voltage change. The cathode glows, and as the current is raised, this glow extends to cover the whole cathode area.

The power supply voltage needs to be more than the striking potential. The circuit used is shown in fig. 69*a*, constant voltage being obtained across the load shown. The series resistance R is necessary to prevent exceeding the tube current rating. $R=(E_S-E_0)/I_m$, where E_S is the total E.M.F., E_0 is the load voltage, and I_m the maximum tube current. Any increase of current through the load, due to fall of load resistance, is accompanied by a tendency to greater voltage drop across R. This tends to reduce the voltage across the tube and load, but less current is then taken by the tube, compensating for the increase of load current, so that the tube and load voltage remain constant, because the voltage drop across R remains the same. Again, if the supply voltage increases, the additional current through the regulator tube will be sufficient to make the voltage drop across R compensate for the change, so the load voltage remains the same.

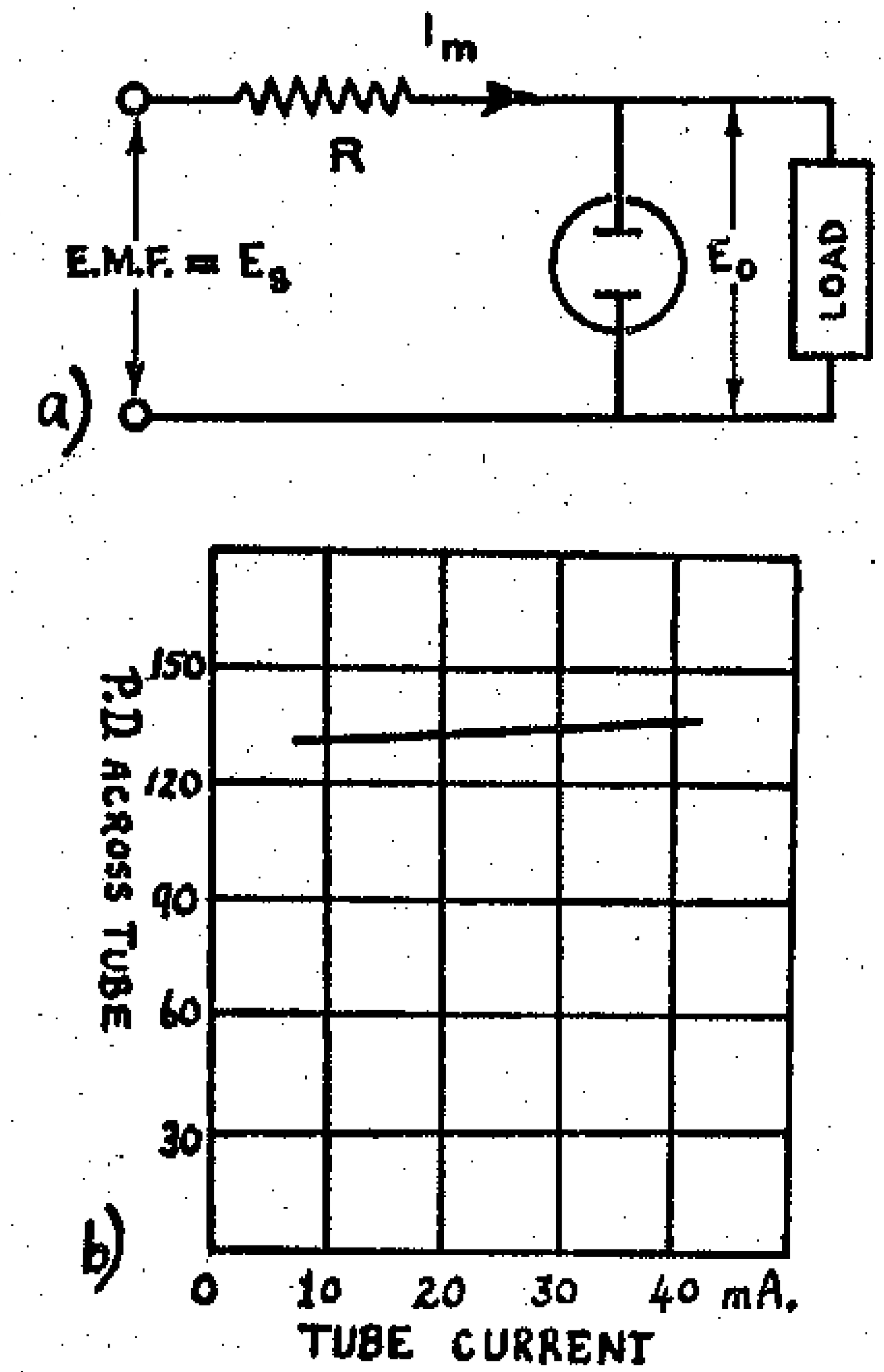


FIG. 69. *a*, Gas-filled Voltage Regulator. *b*, Voltage-current Characteristic.

These tubes are especially useful in the form of the stabilovolt in which the cathode is in the form of a number of metal cylindrical electrodes, one inside the other and insulated from one another. With appropriate external resistances, such an arrangement can be used to provide a total constant voltage of some 300 V. or more which can be subdivided into three or four parts as a potential divider. The alternative is to use a number of neon tubes, or preferably stabilisers like the Cossor S130, in series with one another.

Electronic Regulators. The best types of voltage and current regulators are those employing valves. Whereas a gas-discharge tube cannot readjust itself in less than 1/10th msec., an electron tube is practically instantaneous in its action. The chief disadvantage associated with the use of valves is in obtaining large current values. Thus a regulated current of 0·5 amp. would either require the use of a large type of power valve, or a number of normal valve types in parallel.

A saturated diode, with its constant anode current above saturation anode potential can be used. However, valves with oxide-coated cathodes do not operate for long under saturation current conditions. The bright emitter tungsten filament could be used, but then the provision of a carefully regulated filament current is a problem.

The basic circuit employed using a triode valve regulator is shown in fig. 70*a*. If the H.T. supply voltage rises, the voltage across the load tends to rise, and so does the cathode potential of the triode valve in series with the load. With a given bias, obtained as shown, an increase of H.T. is accompanied by a tendency to a positive increase of cathode potential, and as a result an increase of negative grid bias relative to cathode. So the valve current is reduced, and the voltage drop across it rises, which can be made to compensate for the increased supply volts, the load voltage remaining constant if correct working conditions are arranged. The opposite action takes place if a fall of supply voltage occurs. Similarly, any undesirable changes of the load current are regulated. A resistance is inserted in series with the valve grid to prevent excessive grid current flow when the H.T. supply is switched off, and the grid goes positive because of the continued supply from the bias battery. Alternatively, a gas-discharge tube can be used to provide the grid bias, as in fig. 70*b*.

A mutual-conductance, or g_m-regulator can be arranged as in fig. 70*c*. Here the H.T. supply E.M.F. to be regulated is placed across a potential divider, R_1 and R_2 in series. Let the total H.T.$=E$ volts, and let E_L=voltage across the load supplied. Suppose E changes by an amount ΔE, where ΔE is positive or negative. Then the change of grid voltage is $R_1\Delta E/(R_1+R_2)$. This bias change will produce a change of anode current through R_3 of $g_m R_1 \Delta E/(R_1+R_2)$, where g_m=valve dynamic mutual conductance. If this is equal to the change ΔE of the supply E.M.F.,

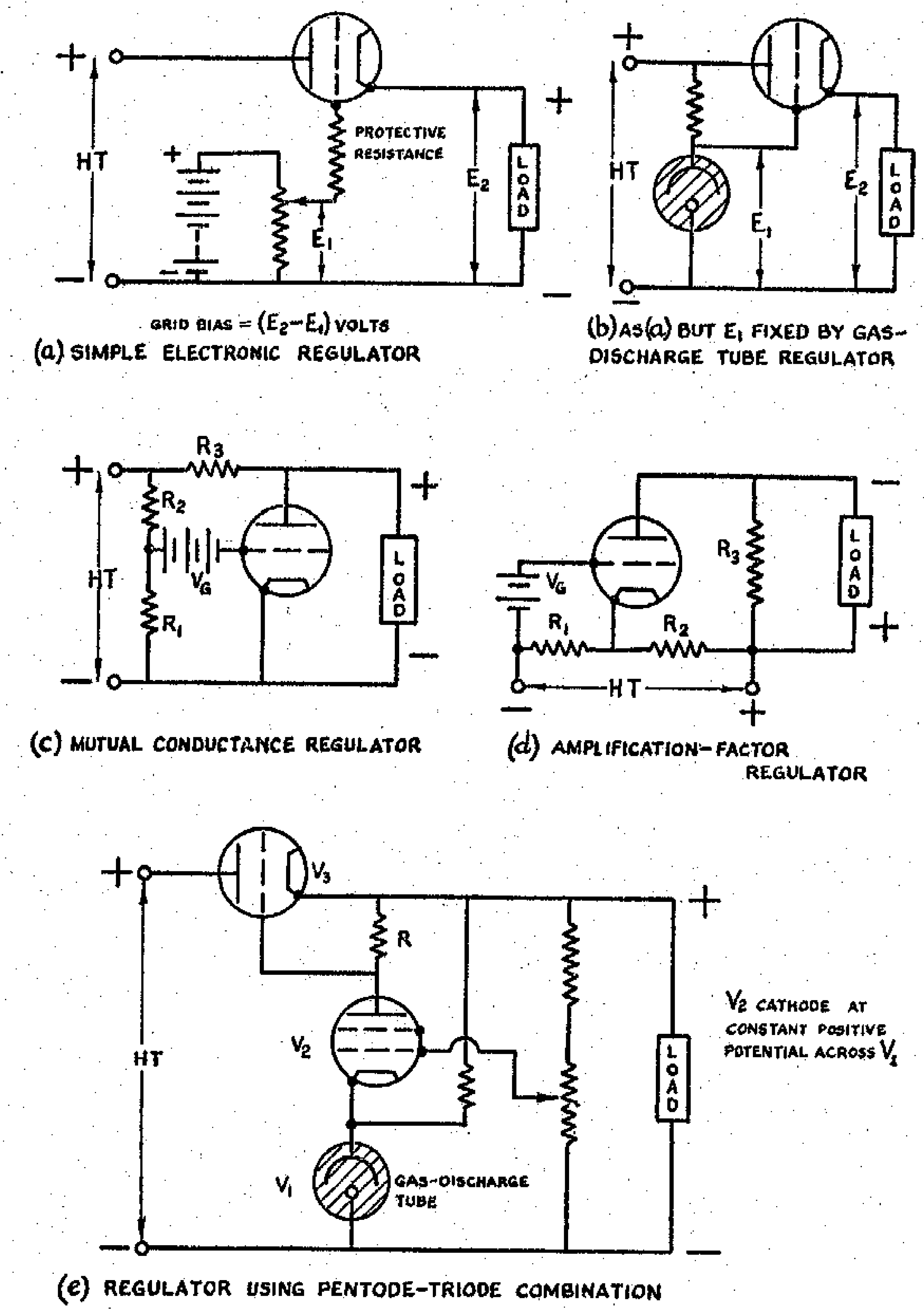


FIG. 70. Electronic Regulator Circuits.

then the anode voltage, and so load voltage, will remain constant, because then an increase of E is accompanied by an equal increase of the P.D. across R_3, and correspondingly for a decrease of E there is an equal decrease across R_3, whilst the anode voltage is equal to the difference between E and the P.D. across R_3. Hence for effective stabilisation of the load voltage

$$\Delta E = g_m \frac{R_1 R_3}{R_1 + R_2} \cdot \Delta E.$$

$$\therefore \quad g_m = \frac{R_1 + R_2}{R_1 R_3}.$$

The actual values of R_1, R_2 and R_3 chosen will depend on the valve characteristics considered in conjunction with the load E.M.F. and current required. The grid bias relative to the cathode will have a mean value of $[R_1/(R_1+R_2)E - V_G]$, which decides the value of the bias battery voltage necessary to achieve a suitable negative bias for class A operation of the valve.

An amplification-factor, or μ-regulator, is shown in fig. 70*d*. Here a change of E by an amount ΔE causes a grid voltage change of $R_1 \Delta E/(R_1+R_2)$ and an anode voltage change of $R_2 \Delta E/(R_1+R_2)$. Since the change of grid volts is here necessarily of opposite polarity to the change of anode volts, so the anode current through R_3, and hence the P.D. across R_3 and the load will remain constant, if the effect on the anode current of the grid voltage change is equal and opposite to the effect on the anode current of the anode voltage change. But in accordance with the definition of amplification-factor, this will be the case if the change of anode voltage divided by the change of grid voltage equals μ. Therefore the circuit is an effective regulator when

$$\frac{R_2 \cdot \Delta E}{R_1 + R_2} \div \frac{R_1 \Delta E}{R_1 + R_2} = \mu.$$

$$\therefore \quad \frac{R_2}{R_1} = \mu.$$

The ratio of R_1 to R_2 is therefore decided. The values of R_1, R_2 and R_3 will depend on the valve chosen, and the demands of the load. To operate the valve at its usual grid bias for class A working, it is to be noted that the grid potential relative to the cathode is $[V_G - R_1 E/(R_1+R_2)]$.

Regulators using pentode valves with their constant-current characteristic and high amplification factor are capable of very great control ratio. A typical circuit is shown in fig. 70*e*. The bias on V_3 depends on the P.D. across R, which rises negatively if the current through R rises, and decreases if the current through R falls. But the rise and fall of the current through R depends on the grid bias on V_2, which will change in direction in the same way as any change of load voltage. So a tendency for the load voltage to increase, sensitively controls the increase of negative bias on V_1, which makes the A.C. resistance of V_3 greater. The P.D. across V_3 therefore rises, which can be made to compensate exactly for any tendency for the load voltage to increase. Vice versa, a tendency for a decrease of load voltage to occur is offset.