CHAPTER XI.—AMPLIFICATION

GENERAL PRINCIPLES

1. In preceding chapters reference has been made to the employment of the thermionic valve as an amplifier, and it now becomes desirable to consider this function in greater detail. In this connection the term valve must be understood to exclude the two-electrode valve or diode, which possesses no amplifying property. The latter is dependent upon the introduction into the space between cathode and anode of one or more grid-like structures which exercise a control upon the electron current without necessarily acting as electronic collectors, and therefore without the necessity for power expenditure in the input circuit. The triode, possessing only a single grid as a control electrode, is the prototype of all amplifying valves, and its use will be first discussed in order that the advantages obtained under certain conditions by the use of tetrodes, pentodes and other special types will be appreciated in due course.

Classification of amplifiers

2. (i) Amplification may be defined as the process by which either current, voltage or power may be increased without serious change of wave-form, although this definition does not hold in certain special cases, e.g. when it is desired to emphasize one component of a complex wave-form compared with other components. An amplifier is an assembly of valves and circuits by which amplification is achieved. Current amplification is of little practical importance, and it is usual to divide amplifiers into two main classes, (*a*) voltage amplifiers and (*b*) power amplifiers. A further classification is also made according to the portion of the frequency spectrum in which the amplifier is designed to operate, namely audio-frequency (A.F.) and radio-frequency (R.F.) amplifiers. The extent to which amplification is employed has been considerably increased in the last few years. Its use was at one time entirely confined to receivers, radio-frequency amplification being incorporated in order to increase the input voltage to the detector valve, and audio-frequency amplification of the rectified signal in order to increase the volume of sound produced. Of late years, however, radio-frequency amplification has been an important feature of many transmitters owing to the necessity for control of the radiated frequency, while audio-frequency amplification is also often required in the sub-modulator stages of R/T transmitters.

(ii) Amplifiers are also sometimes classified under the following headings :—

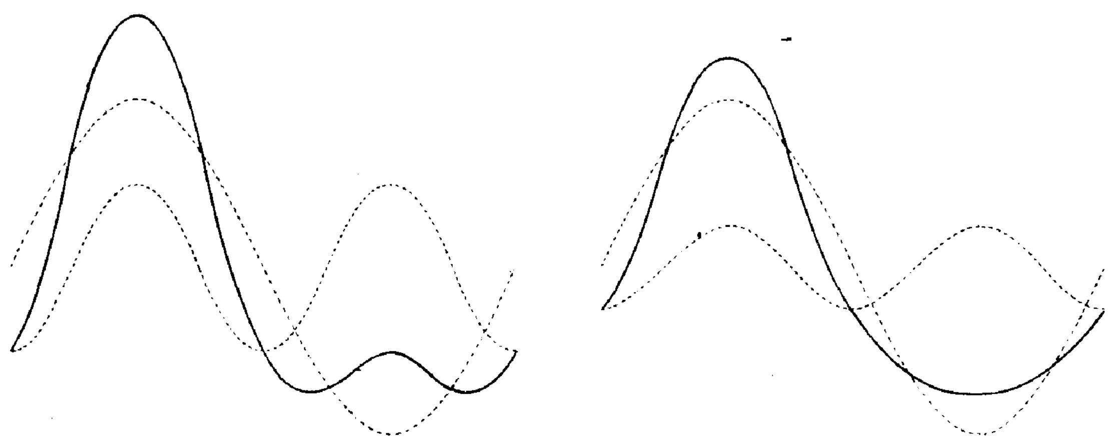
Class A.—An amplifier which is operated under conditions which ensure that the wave-form of the anode current variation is practically the same as that of the input grid-filament voltage. Under these conditions both the efficiency and power output are low. Class A amplifiers are used only for voltage amplification (both A.F. and R.F.) and audio-frequency power amplification in the output stage of R/T receivers, where absence of distortion is of greater importance than electrical efficiency.

Class B.—An amplifier which is operated with such negative grid bias that in the absence of any input signal voltage, the anode current is practically zero. The grid is then said to be biassed to "cut-off point." Grid current is usually allowed to flow during a portion of the cycle. The anode current wave-form is approximately a series of half sine waves, alternate half-cycles being suppressed. These amplifiers are frequently employed for the amplification of modulated radio-frequency voltages in R/T transmitters ; when adjusted for maximum output the efficiency is about 40 per cent.

Class C.—In this type of amplifier the grid is biassed to a point more negative than the cut-off voltage and therefore, when an input signal voltage is applied, anode current flows for less than one half-period. An efficiency of the order of 85 per cent. can be achieved in this manner. Class C amplifiers are used in both R/T and W/T transmitters. The simple C.W. transmitter described in Chapter IX is merely a special form of R.F. amplifier in which the grid excitation voltage is derived from the output of the valve. If operated under sinusoidal conditions without grid current, the conditions are those of class A amplification, while if operated at high efficiency they correspond to class B or class C according to the magnitude of the grid bias.

Fractional band width

3. In referring to the range of frequencies over which an amplifier is designed to operate, the terms "wide" and "narrow" bands are frequently employed. The sense in which these terms are used calls for some explanation. For example, a frequency band of 20 kilocycles per second is a wide band when dealing with audio-frequency amplification, but a narrow band so far as radio frequencies are concerned. The difficulty is removed if the term " band width " is understood to signify "fractional band width". If an amplifier operates uniformly over all frequencies between an upper limit f_2 and a lower limit f_1, it is said to respond to an absolute frequency band of width $f_2 - f_1$. The fractional band width is the quantity obtained by expressing the absolute width as a fraction of the geometric mean frequency, $\sqrt{f_1 f_2}$, and is therefore $\frac{f_2 - f_1}{\sqrt{f_1 f_2}}$. As an example, consider an amplifier which gives equal amplification of all frequencies between 32 and 20,000 cycles per second. The geometric mean frequency is 800 cycles per second and the fractional width 24·96, i.e. greater than unity. In contrast, take an amplifier designed to operate in the region of one megacycle per second, but which gives no appreciable amplification of frequencies 10,000 cycles above or below 1 Mc/s. The geometric mean frequency is practically equal to 1 Mc/s and the fractional band width of the amplifier is approximately $\frac{20{,}000}{1{,}000{,}000} = \cdot 02$. It will be noted that although the absolute band width is practically the same in each case (20,000 cycles per second) the fractional band width in the former example is very much greater



(a) Voltage applied to amplifier (b) Wave form of output voltage

FIG. 1, CHAP. XI.—Frequency distortion.

than in the latter. When the fractional band width is small, the amplifier is said to be selective and vice versa. As a rule, radio-frequency amplifiers are designed to be as selective as possible, while audio-frequency amplifiers are required to give equal amplification over a wide (fractional) band and are therefore unselective.

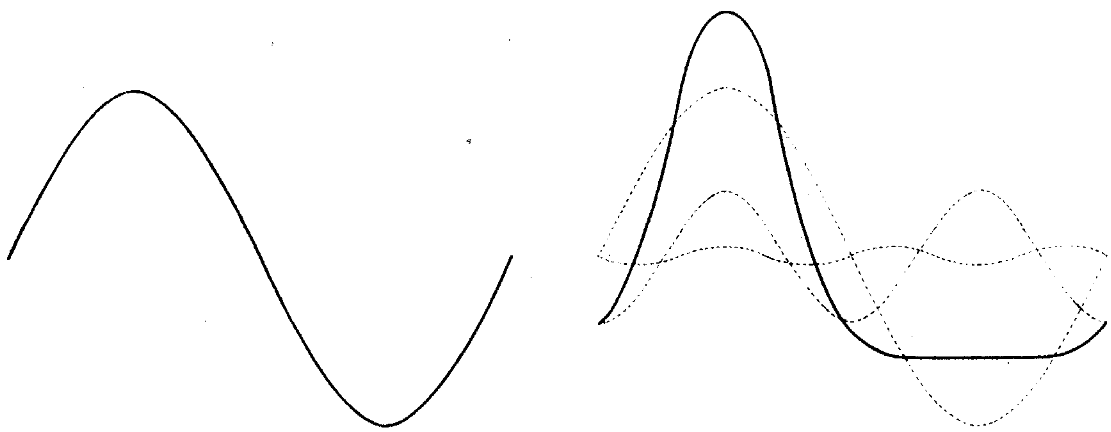
Distortion in amplifiers

4. An amplifier producing an output voltage which is an exact duplicate of the input voltage in every respect save magnitude may be called an ideal or distortionless amplifier ; although it

is not possible to construct such an amplifier a close approach to the ideal can be achieved by careful design. It is usual to distinguish between three types of distortion, namely :—

(i) *Frequency distortion.*—This term is applied to the unequal amplification of different frequencies. This is generally a desirable characteristic of a radio-frequency amplifier because it is then selective. In audio-frequency amplifiers, frequency distortion is usually undesirable, and its avoidance calls for considerable skill in design. The effect of this form of distortion is shown in fig. 1a, in which the original wave form has a second harmonic of amplitude one-half that of the fundamental. After amplification by a selective amplifier the wave-form may be that of fig. 1b, in which the amplitude of the second harmonic is only one-quarter of the fundamental.

(ii) *Amplitude distortion.*—The effect of amplitude distortion is illustrated in fig. 2 ; in this instance the input wave-form is assumed to be sinusoidal. After amplification the wave-form has become peaky and in fact contains second and third harmonics as well as the original wave-form. From figs. 1 and 2 it is seen that the effect of amplitude and frequency distortion may be very similar, at any rate in audio-frequency amplifiers. The difference lies rather in the cause



(a) Voltage applied to amplifier (b) Wave form of output voltage

FIG. 2, CHAP. XI.—Amplitude distortion.

than in the effect, and as a rule frequency distortion is caused by the nature of the circuits used in conjunction with the valve and amplitude distortion by the valve itself. Amplitude distortion is in fact the result of a non-linear relation between current and voltage, and must exist to some extent in all valve circuits, because neither the $I_a - V_g$ nor the $I_g - V_g$ characteristic is perfectly straight. Curvature of the $I_g - V_g$ curve is immaterial if the operating conditions are so chosen that grid current never occurs. This entails operating with considerable negative bias, and restricting the amplitude of the input voltage so that the grid potential never reaches the value at which grid current commences. The use of a high effective resistance in the anode circuit tends to reduce the curvature of the dynamic $I_a - V_g$ characteristic and therefore to reduce amplitude distortion. The steps taken to minimize amplitude distortion may therefore be summarized as below.

(*a*) Employ an anode circuit of high dynamic resistance.

(*b*) Apply sufficient H.T. voltage to ensure that an ample approximately straight portion of the $I_a - V_g$ curve exists, in the region of negative grid voltage.

(*c*) Adjust the grid bias to a value midway between the point at which the curvature of the characteristic becomes appreciable, and the point at which grid current starts to flow.

(*d*) Limit the input voltage so that the excursion of anode current is confined to the straight part of the curve.

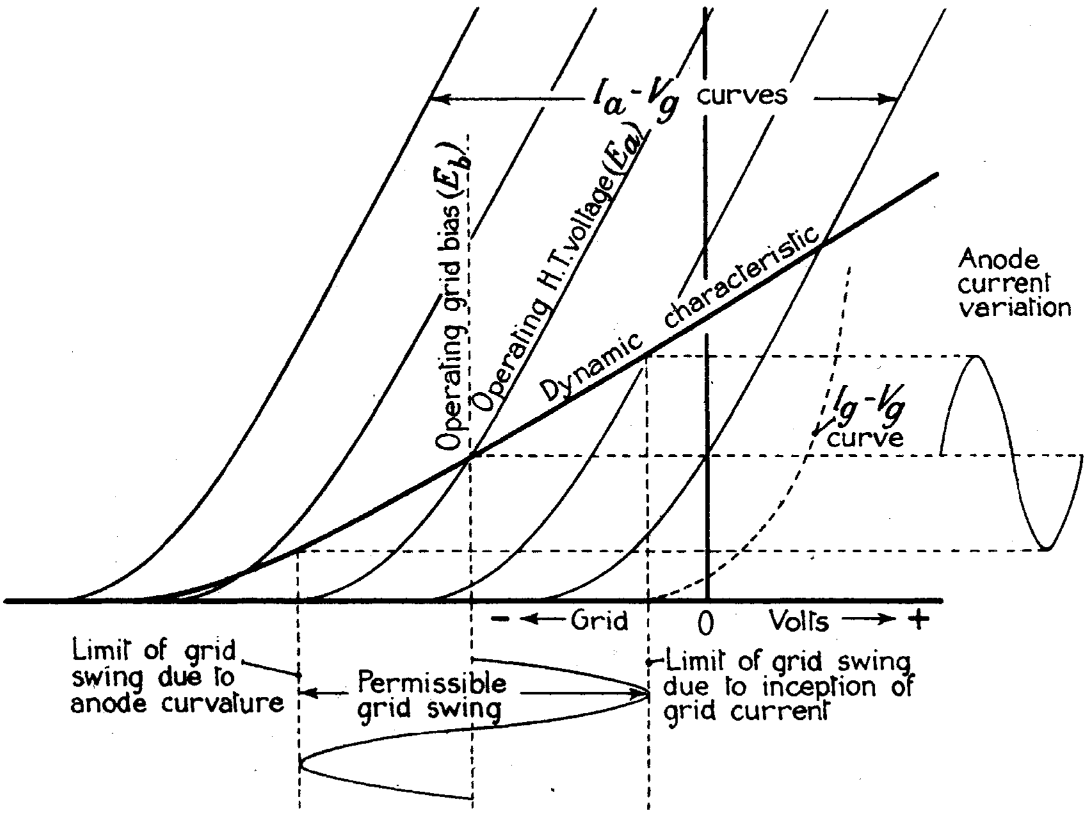


FIG. 3, CHAP. XI.—Operating conditions for distortionless amplification.

These operating conditions are shown in fig. 3. When they are fulfilled the wave-form of the anode current variation is practically identical with that of the grid-filament voltage variation. The class A amplifier may therefore be defined as one operated as indicated in (*a*), (*b*), (*c*) and (*d*) above.

(iii) *Phase distortion* results when the phase relationship between different frequency components is disturbed in such a manner that the wave-form of the output voltage differs from that of the input, although the relative amplitude of the various components is unchanged. The effect is illustrated in fig. 4, the original and distorted wave-form being again shown. The

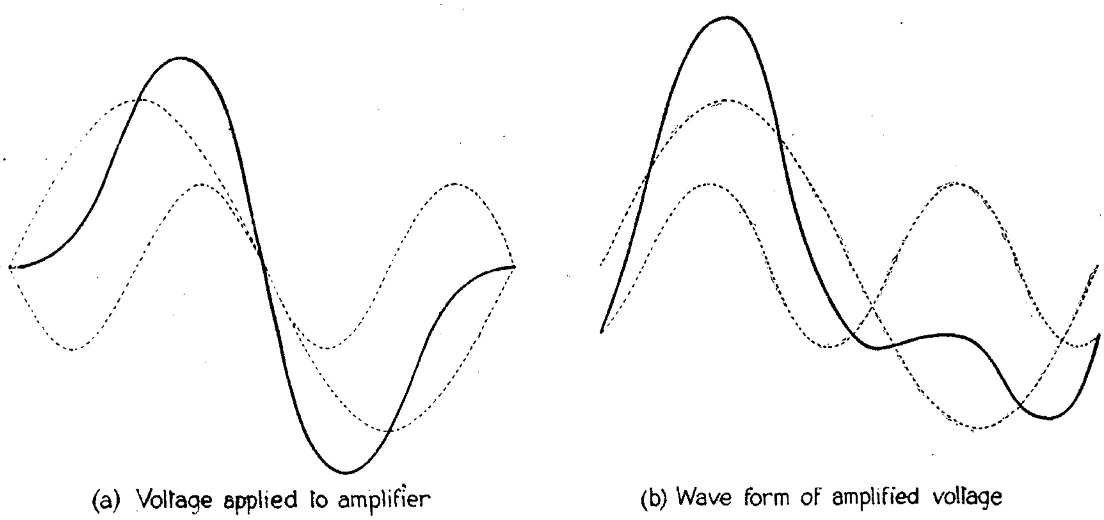


FIG. 4, CHAP. XI.—Phase distortion.

amplitudes of fundamental and second harmonic are as 2 : 1 in each case, but during amplification their relative phase has undergone a displacement of 90°. Phase distortion only occurs when the time taken for the signal to pass through the amplifier is comparable with the duration of the signal, and is therefore of no significance in ordinary reception.

Equivalent circuit of amplifier

5. The basic circuit of the triode amplifier is given in fig. 5a in which T is the triode, v_g the voltage of the signal to be amplified, E_b the voltage of the grid bias battery and E_a the voltage of the anode supply or H.T. battery. Z is the anode circuit load impedance, and for brevity is often referred to as the output circuit, while the circuit connected between grid and filament of the valve is called the input circuit. The variation of grid-filament voltage v_g gives rise to a corresponding variation of anode current, which must of necessity flow through the load impedance, and consequently a varying voltage v_a is set up across the latter. In a voltage amplifier the object is to obtain the largest possible voltage variation v_a. In a power amplifier, however, in addition to this voltage variation, an appreciable current variation is also required, so that the power dissipated in the output circuit, as a result of the grid-filament voltage variation v_g, shall be as large as possible. The variation of anode current produced by the application of a voltage v_g to the grid-filament path is exactly the same as would be produced by a voltage μv_g acting in

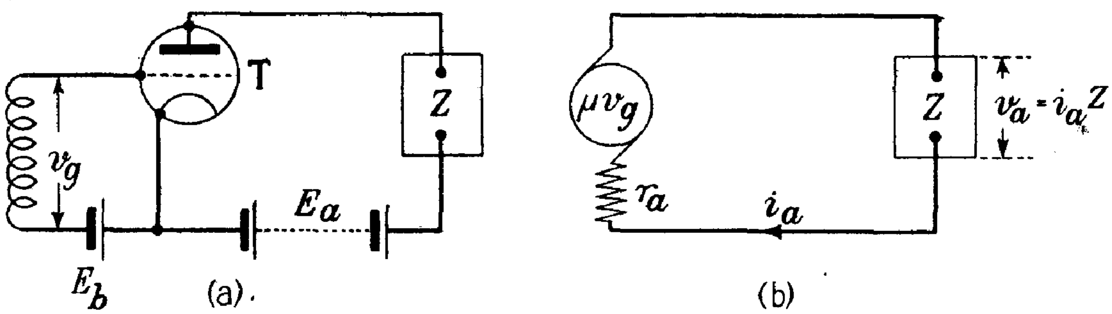


FIG. 5, CHAP. XI.—Basic circuit of amplifier, and equivalent circuit.

the anode circuit of the valve (Chap. VIII). The equivalent circuit of the valve amplifier is therefore as shown in fig. 5b. This equivalent circuit gives only those currents and P.D's which result from the application of the signal voltage, which are superimposed upon the steady or no-signal values of P.D. and current.

Voltage amplification factor

6. (i) The voltage amplification factor (V.A.F.) of a valve and an associated impedance in its anode circuit is the ratio of the voltage variation across the external impedance to the voltage variation between grid and filament of the valve. The anode circuit load impedance may be of any nature whatever, provided that it possesses finite conductivity at zero frequency, i.e. for direct current. This limitation merely signifies that a simple series condenser cannot be employed, because it is necessary to provide a complete conductive path for the steady component of anode current. In practice the anode load impedance may be a resistance, an inductive choke, or a tuned circuit consisting of inductance and capacitance in parallel. Further, an additional circuit may be coupled to the load impedance by any of the methods enumerated in Chapter VI.

(ii) In deriving the voltage amplification factor appropriate to any particular form of anode circuit, it is desirable to assume that no amplitude distortion will take place, and therefore that the following conditions are fulfilled.

(*a*) Ample filament emission is provided.

(*b*) The anode is maintained at a positive potential with respect to the filament during the whole cycle of applied grid voltage.

(*c*) The mean grid potential is sufficiently negative to ensure that no grid current flows.

(*d*) The $I_a - V_g$ characteristic is approximately straight over the whole portion to which the excursion of anode current extends.

(*e*) Unless the contrary is explicitly stated, the interelectrode capacitance is assumed to be negligibly small ; the effect of its finite magnitude will be indicated where this is of consequence in operation or design.

Ohmic resistance as load impedance

7. In the circuit shown in fig. 6a the load impedance consists of a non-inductive resistance of R ohms. If an alternating voltage $v_g = \mathcal{V}_g \sin \omega t$ is applied between grid and filament, its effect upon the anode current is exactly the same as would be produced if the grid potential

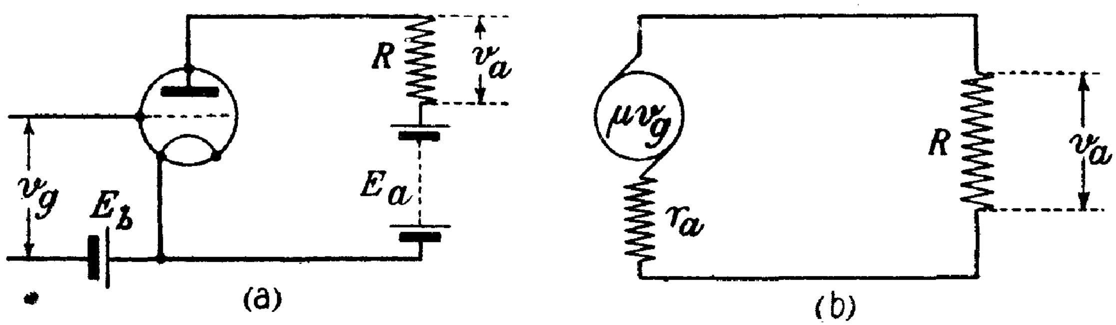


FIG. 6, CHAP. XI.—Resistance as anode load.

were maintained constant and an alternating E.M.F. $\mu \mathcal{V}_g \sin \omega t$ were introduced into the anode circuit (fig. 6b). The resulting variation of anode current will be

$$i_a = \frac{\mu \mathcal{V}_g \sin \omega t}{r_a + R}$$

The output voltage, v_a, of the amplifier is the P.D. across the load resistance R and is equal to $i_a R$, hence

$$v_a = \frac{R \mu \mathcal{V}_g \sin \omega t}{r_a + R}$$

With a sinusoidal variation of grid voltage, therefore, the output voltage v_a is also sinusoidal and the above equation may be written

$$v_a = \frac{R}{r_a + R} \mu v_g.$$

The V.A.F. is defined above as the ratio $\frac{v_a}{v_g}$ so that the V.A.F. of the circuit shown in fig. 6 is

$$\frac{v_a}{v_g} = \frac{R}{r_a + R} \mu.$$

Example.—A triode has an amplification factor of 20 and an r_a of 50,000 ohms. If a purely resistive impedance of 100,000 ohms is placed in the anode circuit, find the V.A.F.

$$\begin{aligned} \text{V.A.F.} &= \frac{R}{r_a + R} \mu \\ &= \frac{100{,}000}{50{,}000 + 100{,}000} \times 20 \\ &= 13\tfrac{1}{3}. \end{aligned}$$

8. An increase in the value of the load resistance will result in an increase in the V.A.F. ; for example if a resistance of 200,000 ohms is substituted in the circuit just considered, a V.A.F. of 16 is obtained. The limiting value of the V.A.F. is equal to the amplification factor of the valve, i.e. μ, but this value cannot be reached in practice. It is approached more closely as the ratio $\frac{R}{r_a}$ becomes larger and larger, as shown in fig. 7, in which the ordinate is $\frac{\text{V.A.F.}}{\mu}$ and the abcissa the ratio $\frac{R}{r_a}$, so that to obtain the V.A.F. with any particular valve, the ordinate must be multiplied by the appropriate value of μ. It may appear desirable to use the highest obtainable value of load resistance ; a practical limit to its magnitude is however imposed by the necessity of maintaining the anode at a positive potential of at least a few volts (e.g. 10 volts) above that of the filament. Since the anode D.C. resistance of the valve and the load resistance are in series, this limits the value of the latter to about $10\,r_a$. Reference to fig. 7 shows that the V.A.F. is then

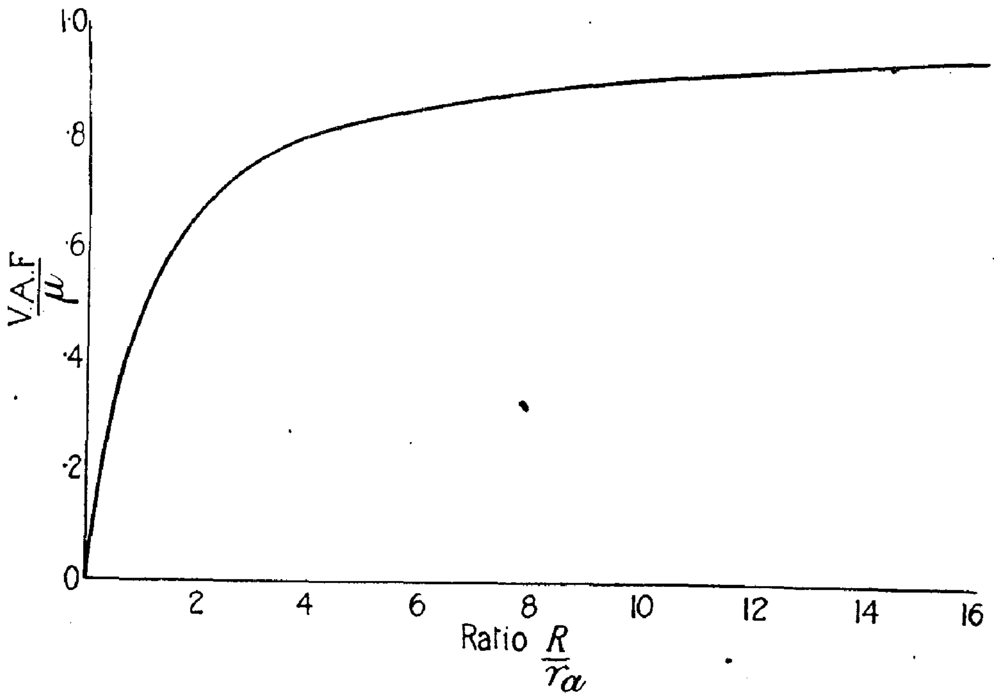


FIG. 7, CHAP. XI.—Effect of ratio $\frac{R}{r_a}$ upon V.A.F.

90 per cent. of the theoretical maximum value μ. From the expression given above it would also appear that the V.A.F. is independent of the frequency of the applied grid-filament voltage. In obtaining the V.A.F., however, the presence of stray capacitance was neglected, and it will presently be shown that the effect of such a capacitance, which effectively acts as a shunt upon the load resistance, is to cause serious reduction of the amplification obtainable at all frequencies above a few thousand cycles per second.

Inductance as load impedance

9. Fig. 8a shows a possible amplifier disposition in which the impedance Z of fig. 5 is constituted by an inductance of L henries and of negligible resistance, the equivalent circuit being given in fig. 8b. The mean operating potentials E_b and E_a are as in the previous example.

The application of a sinusoidal grid-filament voltage $v_g = \mathcal{V}_g \sin \omega t$ will result in a corresponding variation of anode current i_a, and

$$i_a = \frac{\mu\, v_g}{\sqrt{r_a^2 + (\omega L)^2}}.$$

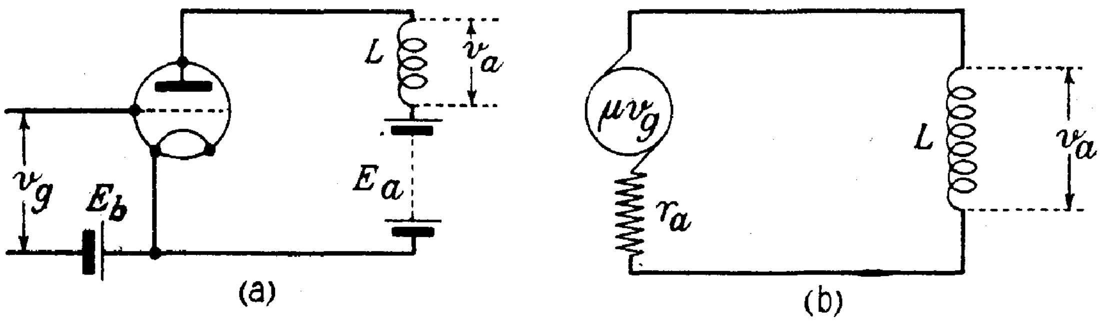


FIG. 8, CHAP. XI. Inductance as anode load.

The P.D. between the ends of the inductance, which is of course the output voltage v_a of the amplifier is $\omega L\, i_a$, or

$$v_a = \frac{\omega L}{\sqrt{r_a^2 + (\omega L)^2}}\, \mu\, v_g$$

and the V.A.F. is

$$\frac{v_a}{v_g} = \frac{\omega L}{\sqrt{r_a^2 + (\omega L)^2}}\, \mu.$$

Example.—In fig. 8a, if the valve has an amplification factor of 20 and an r_a of 20,000, while the load impedance is an inductance of 10 henries, find the V.A.F. when the frequency of the applied grid filament voltage is (i) 10 cycles per second, (ii) 100 cycles per second, (iii) 1,000 cycles per second.

(*a*) $\omega L = 2\pi \times 10 \times 10 = 200\pi = 628.$

$$\text{V.A.F.} = \frac{628}{\sqrt{20{,}000^2 + 628^2}} \times 20$$

$$\doteqdot \frac{628}{20{,}000} \times 20$$

$$\doteqdot \cdot 628$$

(*b*) $\omega L = 6{,}280.$

$$\text{V.A.F.} = \frac{6{,}280}{\sqrt{20{,}000^2 + 6{,}280^2}} \times 20$$

$$\doteqdot \frac{6{,}280}{21{,}000} \times 20$$

$$\doteqdot 6 \cdot 6$$

(*c*) $\omega L = 62{,}800.$

$$\text{V.A.F.} = \frac{62{,}800}{\sqrt{20{,}000^2 + 62{,}800^2}} \times 20$$

$$\doteqdot \frac{62{,}800}{66{,}400} \times 20$$

$$\doteqdot 18 \cdot 95.$$

With this form of load impedance, then, the V.A.F. increases as the frequency of the input voltage is increased, and approaches the limiting value μ as the load reactance, ωL, becomes larger and larger compared with the anode A.C. resistance r_a. The increase of V.A.F. obtained by an increase in the ratio $\frac{\omega L}{r_a}$ is shown in fig. 9 which should be compared with fig. 7. It is seen that no advantage is obtained by increasing the ratio $\frac{\omega L}{r_a}$ beyond about 5, ω being taken as 2π times the lowest frequency at which appreciable amplification is required. It should be noted that as the inductance is assumed to have zero resistance, there is no steady voltage drop between

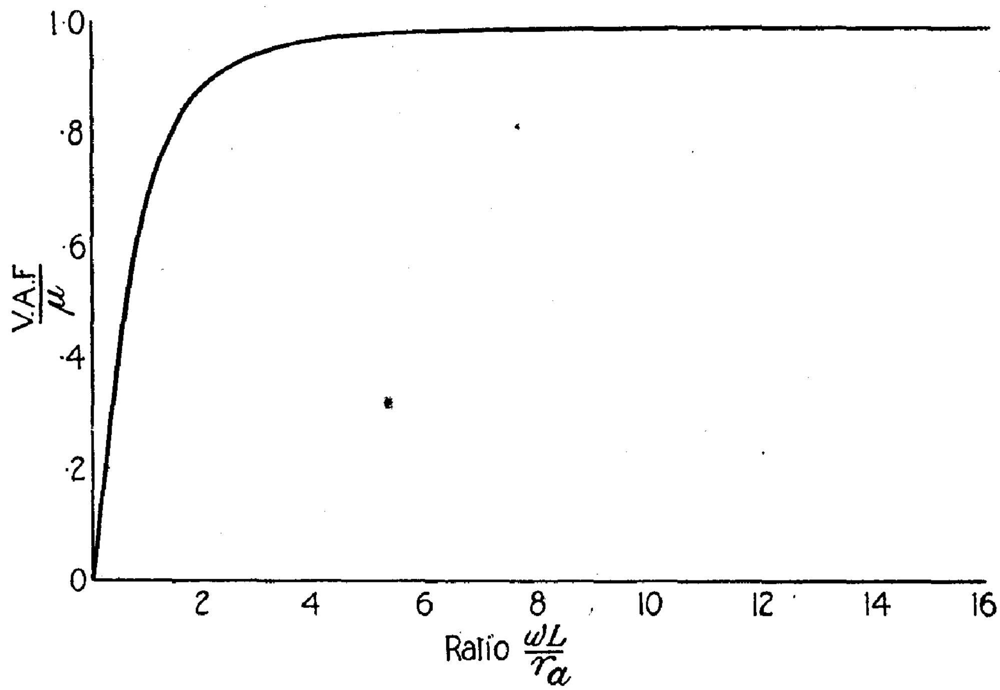


FIG. 9, CHAP. XI. Effect of ratio $\frac{\omega L}{r_a}$ upon V.A.F.

its ends, and the mean anode-filament P.D. is equal to the E.M.F. of the H.T. battery. In practice the resistance of the coil is always negligible compared with the anode-filament (D.C.) resistance of the valve.

Effect of stray capacitance

10. Before discussing the practical application of these and other forms of load impedance, it is desirable to study the effect of a stray capacitance in parallel with the inductance or resistance constituting the load; such a capacitance is always present, and its effective value may be of the order of 100 $\mu\mu F$. Let the stray capacitance be denoted by C_s, its reactance at a frequency $\frac{\omega}{2\pi}$ being $\frac{1}{\omega C_s}$. Taking the case of a resistance load, the anode circuit impedance,

Z_o, is then that of R and C_s in parallel. By methods explained in Chap. V it can be shown that

$$Z_o = \frac{R}{\sqrt{1 + \omega^2 C_s^2 R^2}} = \sqrt{X_o^2 + R_o^2}$$

$$X_o = \frac{\omega C_s R^2}{1 + (\omega C_s R)^2}$$

$$R_o = \frac{R}{1 + (\omega C_s R)^2}$$

and the V.A.F. becomes

$$\frac{Z_o}{\sqrt{(r_a + R_o)^2 + X_o^2}} \mu.$$

Example.—Assuming that $r_a = 50{,}000$ ohms, $\mu = 20$, $R = 100{,}000$ ohms, $C_s = 100\mu\mu F$, calculate the V.A.F. at the following frequencies, viz., (i) 800 cycles per second (ii) 16,000 cycles per second (iii) 800 kilocycles per second, the first being near the mean audio frequency, the second near the upper limit of audibility, and the third a radio frequency in the middle of the medium broadcast band.

At 800 cycles per second,

$$\omega = 5{,}000 \text{ (approx.)}$$

$$\omega C_s R = 5 \times 10^3 \times 10^2 \times 10^{-12} \times 10^5 = \cdot 05$$

$$(\omega C_s R)^2 = \cdot 0025$$

$$X_o = \frac{\cdot 05 \times 10^5}{1 \cdot 0025} = 4{,}988 \text{ ohms}$$

$$R_o = \frac{10^5}{1 \cdot 0025} = 99{,}750 \text{ ohms}$$

$$Z_o = 99{,}875 \text{ ohms}$$

$$\text{V.A.F.} = \frac{99{,}875}{\sqrt{(5 \times 10^4 + 9 \cdot 975 \times 10^4)^2 + 4{,}988^2}} \times 20$$

As X_o is so small compared to $r_a + R_o$ this is practically

$$\frac{9 \cdot 9875}{14 \cdot 975} \times 20$$

$$\text{V.A.F.} = 13 \cdot 3.$$

This result is of course identical with that found by assuming the stray capacitance to be negligible.

At 16,000 cycles per second,

$$\omega = 100{,}000$$

$$\omega C_s R = 10^5 \times 10^2 \times 10^{-12} \times 10^5 = 1$$

$$(\omega C_s R)^2 = 1$$

$$X_o = \frac{10^5}{2} = 50{,}000 \text{ ohms}$$

$$R_o = \frac{10^5}{2} = 50{,}000 \text{ ohms}$$

$$Z_o = \sqrt{50{,}000^2 + 50{,}000^2}$$
$$= 70{,}700 \text{ ohms}$$

$$\text{V.A.F.} = \frac{70{,}700}{\sqrt{(50{,}000 + 50{,}000)^2 + 50{,}000^2}} \times 20$$
$$= \frac{7 \cdot 07}{125} \times 20$$
$$= 12 \cdot 6.$$

At 800 kilocycles/sec.

$$\omega = 5 \times 10^6$$
$$\omega C_s R = 5 \times 10^6 \times 100 \times 10^{-12} \times 10^5 = 50$$
$$(\omega C_s R)^2 = 2{,}500$$
$$X_o = \frac{50 \times 10^5}{2{,}501} \doteqdot \frac{50 \times 10^5}{2{,}500} \doteqdot 2{,}000$$
$$R_o \doteqdot \frac{10^5}{2{,}501} \doteqdot \frac{10^5}{2{,}500} \doteqdot 40$$
$$\therefore Z_o = \sqrt{R_o^2 + X_o^2} \doteqdot 2{,}000$$
$$\text{V.A.F.} = \frac{2{,}000}{\sqrt{50{,}040^2 + 2{,}000^2}} \times 20$$
$$\doteqdot \frac{2{,}000}{50{,}000} \times 20$$
$$\doteqdot \cdot 8.$$

11. From the above example, the following general results may be deduced. The effect of stray capacitance upon the V.A.F. is negligible at the lower audio frequencies, but becomes of some importance at the higher audio frequencies. At that frequency for which the reactance of the stray capacitance $\left(\frac{1}{\omega C_s}\right)$ is equal to the joint resistance of the valve and load resistance in parallel, i.e. when

$$\frac{1}{\omega C_s} = \frac{R\, r_a}{R + r_a},$$

the V.A.F. is reduced to $\cdot 707\mu$, and falls off rapidly when the frequency is further increased. With the valve and circuit specified above, the corresponding frequency is about 480 kc/s. It may be taken as a general rule that with the valves at present employed, a resistive anode load will give no appreciable amplification at frequencies higher than about 500 kc/s.

Development of tuned anode circuit

12. (i) Now consider the effect of a similar capacitance in the case of the amplifier having an inductive anode load. It is at once apparent from the circuit diagram of fig. 10 that the capacitance and inductance together form a tuned circuit, the resonant frequency of which is equal to $\frac{1}{2\pi\sqrt{LC_s}}$ cycles per second. If the resistance were truly zero, the circuit would behave at this frequency as a perfect rejector, offering an infinitely high opposition to the flow of current, and the V.A.F. would reach the theoretical limiting value μ. As however some slight resistance must exist, the opposition of the parallel circuit at its resonant frequency is not infinite, but is equal to that of a purely resistive impedance of $\frac{L}{C_s R}$ ohms. This apparent resistance is termed the dynamic resistance of the circuit and is denoted by R_d. (Chapt. V.) In certain circumstances the stray capacitance may be deliberately augmented by connecting a condenser of capacitance C in parallel with the inductance; the resonant frequency is then equal to $\frac{1}{2\pi\sqrt{L\,(C + C_s)}}$ and the load resistance is $\frac{L}{(C + C_s)\,R}$ ohms. When a parallel L C

circuit is employed in this manner it is called a "tuned anode" circuit. The V.A.F. follows at once from the results obtained with a purely resistive anode load, i.e.

$$\text{V.A.F.} = \frac{R_d}{r_a + R_d}\mu$$

This expression is of course only valid when the parallel circuit L, $(C + C_s)$, is tuned to the input frequency. At frequencies higher than the resonant frequency, the circuit will offer both resistance and capacitive reactance, while at frequencies below the resonant frequency it will behave as an inductive resistance, its impedance being always less than the dynamic resistance under resonant conditions. Hence, an amplifier having an inductance in its anode circuit will always give greatest amplification at some particular frequency, depending upon the inductance

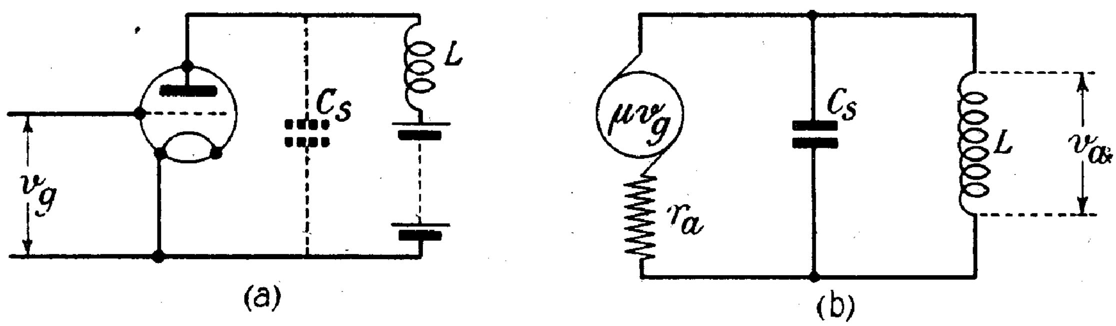


FIG. 10, CHAP. XI. Effect of stray anode-filament capacitance.

and the capacitance of the circuit, i.e. frequency distortion must exist in some degree. The amplifier will give a fairly even response over a wide frequency band, only if a large ratio of inductance to total capacitance is maintained.

(ii) The forms of load impedance just described form the basis of all amplifier designs and their application to practical receiving amplifiers will now be described.

RECEIVING AMPLIFIERS

13. The ultimate purpose of a radio receiver is to produce an audio-frequency variation of current in the windings of the telephone receiver, and a consequent emission of sound by the latter. In Chapter X it was shown that this invariably involves some form of rectification of the radio-frequency voltage, so that an audio-frequency variation of current will result. The purpose of amplification is to obtain either a greater signalling range for the same transmitter power, or a greater response, i.e. a louder sound, from the telephone receivers (for the same power and signalling range) or both. In general, amplification of the radio-frequency voltage, before rectification, will increase the signalling range, while amplification of the rectified output from the detector valve will increase the sound output from the telephones.

Radio-frequency and audio-frequency amplification

14. (i) In the early development of radio-frequency amplification it was thought desirable to aim at the production of aperiodic or semi-aperiodic amplifiers, which would amplify all frequencies, or a wide band of frequencies, equally well. The advantage gained by aperiodic amplification is the elimination of tuning controls, manipulation being confined to the adjustment of the aerial circuit and of the reaction control. The employment of a tuned amplifier of correct design endows the receiver with such enhanced selectivity that aperiodic R.F. amplification has fallen into complete disuse, and tuned radio-frequency amplifiers are now employed for the purpose of gaining selectivity, even if, in the absence of interference, the desired signalling range could be attained by a simple receiver of the type described in Chapter X.

(ii) A further advantage of radio-frequency amplification lies in the fact that for small input voltages, the output from any practical form of rectifier is approximately proportional to the square of the input voltage. If a certain small signal voltage is available, and is amplified, say, to four times its initial magnitude before being applied to the rectifier, the rectified output is sixteen times that which would be obtained by applying the signal voltage directly to the rectifier. The total R.F. amplification is however limited by certain factors to be discussed later, and in many cases the rectified signal may still be of insufficient amplitude to give the desired output of sound from the telephones. In a perfectly quiet room, a very faint telephone sound, such as a morse signal from a very distant transmitter, may prove to be quite readable, but it is necessary to increase the input to the telephone to an enormous degree in order to read a signal in noisy surroundings. In the reception of radio signals in an aeroplane, the high noise level renders it desirable to use the strongest signal which the telephone receivers will withstand without overload, and recourse to a considerable degree of audio-frequency amplification is necessary. As a rule, the audio-frequency amplifier is required to deal with a considerable frequency range, and should amplify equally well at all frequencies within this range, although for C.W. reception, a sharply tuned audio-frequency amplifier, or note selector, is sometimes employed. The actual volume of sound depends upon the supply of electric power to the telephone receivers and the final stage in the receiver must operate as a power amplifier. It must be noted that the term " power amplifier " carries no implication as to the amount of power supplied to the sound-producing device but is merely an indication of its mode of operation.

Multi-stage amplifiers

15. (i) Summarizing the above, then, it may be said that a typical radio receiver may consist of four portions, executing the following functions :—

(*a*) Radio-frequency voltage amplification, in order to ensure the maximum input grid voltage to the detector, and to achieve the highest possible degree of selectivity.

(*b*) Rectification or detection, which gives an audio-frequency output in response to a radio-frequency input voltage.

(*c*) Audio-frequency voltage amplification, by which the audio-frequency output voltage of the detector is increased in amplitude.

(*d*) Power amplification, giving the maximum transfer of power from the H.T. supply device to the telephone receiver, for a given input voltage from the preceding stage.

In addition, some form of heterodyne will be necessary for C.W. reception. This may be incorporated in the receiver either by the employment of sufficient reaction at some point in the radio-frequency portion (autodyne) or by an in-built separate heterodyne. Alternatively, an external heterodyne such as the syntonizer may be employed as described in the previous chapter.

(ii) In such a receiver each valve and its anode circuit is generally called a " stage " : for example, a receiver may have two radio-frequency amplifying stages, a detector stage, a stage of audio-frequency voltage amplification and a final power amplifying stage. It is usual to arrange that all the valve filaments of a multi-stage receiver are supplied from a single source, a common grid bias supply and common H.T. supply being also adopted. If these sources are either primary or secondary batteries the receiver is said to be " battery operated," while if arrangements are made to enable the supplies to be drawn from the power mains (either D.C. or A.C.) the receiver is said to be " mains operated." In the following discussion, battery operation is assumed unless otherwise stated ; mains operation is usually employed when a considerable sound output is required, as in public address systems and in broadcast receivers. In multi-stage amplifiers, the output circuit of one valve is the input circuit of its successor, and the valves are said to be coupled together owing to this dual function. Care must be taken not to confuse the term " inter-valve coupling " with the idea of coupling between tuned circuits as a means of transferring energy from one to the other. The purpose of inter-valve coupling is merely to apply the voltage developed by one valve to the grid and filament of the next, and energy transfer is generally to be avoided.

Audio-frequency amplification

16. Three forms of inter-valve coupling are in general use in A.F. circuits. These are:—

(i) Resistance-capacitance coupling.

(ii) Choke-capacitance coupling.

(iii) Transformer coupling.

(*a*) Series feed.

(*b*) Parallel feed.

Resistance capacitance coupling

17. (i) Referring to fig. 11a, when an alternating voltage v_{g1} is applied to the grid and filament of the valve T_1, an alternating voltage v_a is developed across the ends of the resistance *R*. It is desired to apply this voltage to the grid and filament of the following valve T_2. Now the upper end of the resistance is already connected to the filament of the valve T_2 through the H.T. battery, so that it is apparently only necessary to connect the anode of the valve T_1 to the grid of T_2. This connection has been inserted in the diagram as a dotted line. Unfortunately, the addition of such a connection would place the grid of T_2 at a positive potential, equal to the E.M.F. of the H.T. battery, with respect to its filament, a heavy grid current would therefore flow, and the valve would probably be destroyed; it would at least fail to function in the desired manner.

(ii) This effect is avoided by the insertion of a condenser C_g into the anode-grid connecting lead (fig. 11b). Under these conditions the grid is completely insulated from the filament, a

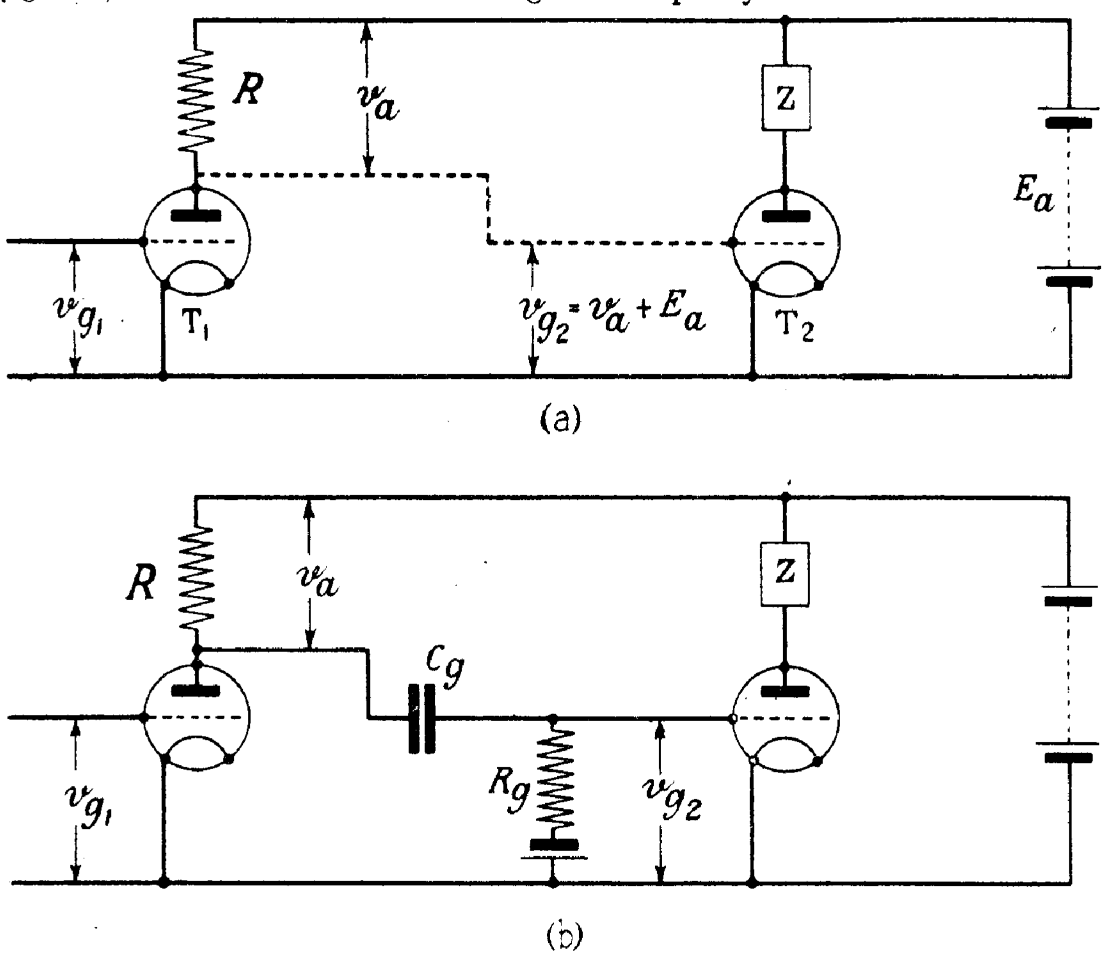


FIG. 11, CHAP. XI. Development of resistance-capacitance coupling.

condition which is not permissible, for the grid will inevitably collect a few electrons during each successive positive half-cycle of input voltage. Unless it is possible for these to return to the filament, the grid will assume an increasingly negative potential, ultimately reducing the anode current to zero. This phenomenon has already been referred to with regard to both detector and oscillator circuits, and the remedy in the present case is the same, namely the addition of a grid leak resistance R_g, which must be externally connected between the grid and filament of the valve. Any required biasing voltage may be applied by connecting a battery of the required voltage and polarity in series with the grid leak. In order that the voltage applied to the grid and filament of the succeeding valve shall be as large as possible, the grid leak should have a high resistance, e.g. ·1 to 1 megohm, and the capacitance of the grid condenser should also be large so that its reactance at the operating frequency is small compared with the resistance of the leak. As will be seen later, the values of C_g and R_g are chosen with regard to these and certain other considerations.

Stage gain

18. The "gain" of a stage of amplification is the ratio of the grid-filament signal P.D. v_{g2} of one valve to the input (grid-filament) voltage v_{g1} of the previous valve in the amplifier, and differs from the V.A.F. in that any voltage drop in the inter-valve coupling device is taken into consideration. In the present type of circuit the impedance of the grid condenser and leak (in series), forms a shunt upon the anode resistance R, and the V.A.F. is rather less than that calculated by the formula given above. At frequencies for which the reactance of the grid condenser is small compared with the resistance of the grid leak, the V.A.F. may be calculated approximately by assuming that the load resistance R_e consists of R and R_g in parallel, i.e.

$$R_e = \frac{R_g}{R + R_g} R.$$

The factor $\frac{R_g}{R + R_g}$ is obviously only of importance when R_g is not very much larger than R. Its effect is only mentioned because it forms one limit to the upper value which may usefully be adopted for the resistance R; as R_g will generally be not more than a megohm, the effective load resistance will be less than this even if R is very much greater. The stage gain is rather less than the V.A.F. because the grid condenser and leak together form a kind of alternating current potentiometer, supplying only a fraction of the P.D. across the anode load to the grid and filament of the succeeding valve, i.e.

$$v_{g2} = \frac{R_g}{\sqrt{R_g^2 + \left(\frac{1}{\omega C_g}\right)^2}} v_a.$$

At the frequency at which $R_g = \frac{1}{\omega C_g}$ the gain will be nearly $\frac{1}{\sqrt{2}}$ or 70 per cent. of the theoretical V.A.F. At the frequency which makes $\frac{1}{\omega C_g} = 2R_g$ the stage gain is nearly 50 per cent. of the V.A.F., while when the frequency is so high that $\frac{1}{\omega C_g}$ is less than $\frac{R_g}{3}$ the stage gain is within 5 per cent. of the V.A.F. The capacitance of the grid condenser depends upon the lowest frequency at which appreciable amplification is required, but should not be larger than necessary. The valves preceding the audio-frequency stages, i.e. R.F. amplifying valves and detector valve, are not absolutely steady in action but tend to set up a background of noise, which consists for the most part of very low frequency components. If the amplifier has negligible gain for frequencies below about 200 cycles per second, this background noise is reduced to a considerable extent. As an example, if we employ a grid condenser and leak of ·001 μF and 2 megohms respectively, the gain will be 70 per cent. of the V.A.F. at about 80 cycles per second, but will fall off rapidly at lower frequencies, being only 45 per cent. of the V.A.F. at 40 cycles per second. The actual amplification of frequencies below 200 cycles per second is shown in fig. 12 as a fraction of the V.A.F. and the reduction of low frequency noise due to preceding stages is seen to be considerable.

19. The extent to which the stage gain falls off at high frequencies is determined by the magnitude of the effective shunting capacitance, compared with the anode A.C. resistance of the valve and the circuit resistances R and R_g. The lower the values of these resistances the less will be the effect of the capacitance and if it is desired to avoid considerable fall in amplification

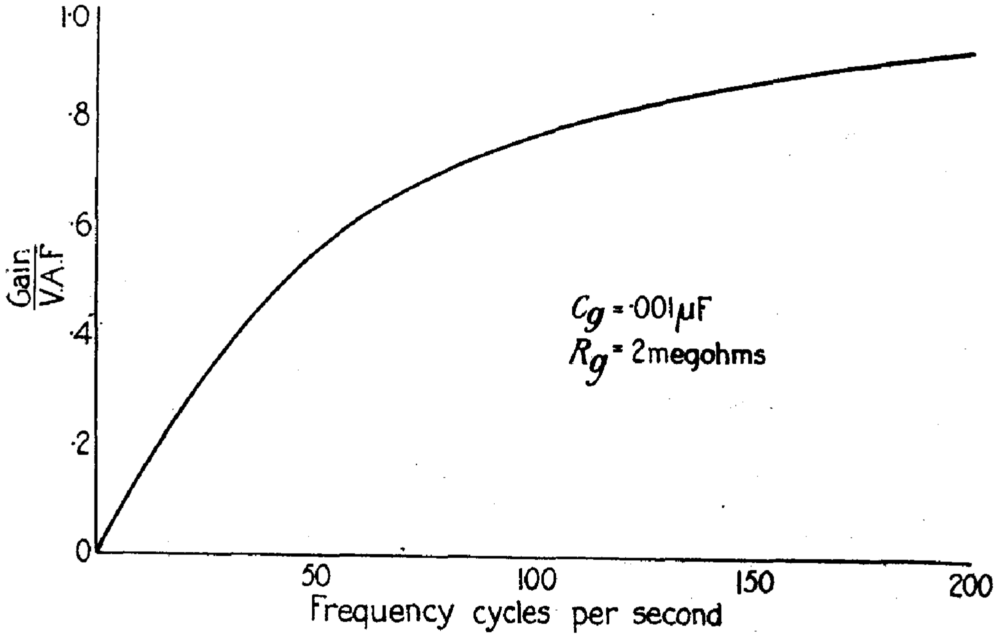


FIG. 12, CHAP. XI. Low-frequency cut-off due to grid condenser.

at the higher audio frequencies, it is necessary to employ a valve of low amplification factor, because such a valve usually has a low anode A.C. resistance. This in turn enables a V.A.F. approaching the limiting value μ to be achieved with a comparatively low anode resistance R and a smaller grid leak resistance than would otherwise be required. The effective shunt capacitance is approximately equal to $C_s + C_{af} + C_{gf} + C_{ag}(1 + A)$ where

C_s = stray capacitance between anode and filament of first valve.
C_{af} = anode-filament capacitance of first valve.
C_{gf} = grid-filament capacitance of second valve.
C_{ag} = anode-grid capacitance of second valve.
A = V.A.F. of the succeeding valve and its associated anode circuit.

Note.—A cannot exceed the amplification factor μ of the second valve.

Since each interelectrode capacitance (including that due to the valve holder) usually ranges from 5 to 10 $\mu\mu F$, it will be seen that the previous estimate of 100 $\mu\mu F$ for the total effective shunt capacitance is a fair average value. The effect of the last term, i.e. $C_{ag}(1 + A)$ is important, because it renders the gain of one stage dependent upon that of the succeeding valve. The overall gain of a number of amplifying stages can, therefore, only be accurately calculated by commencing with the final stage and working backwards, allowing for the effective load due to the shunting capacitance in each case.

20. An important practical point in the design and maintenance of this type of amplifier is the insulation resistance of the coupling condenser. Suppose that in fig. 11b

the insulation resistance of C_g is 20 megohms. If the anode resistance R is ·2 megohm and the grid leak 2 megohms, the H.T. battery voltage being say 222 volts, a direct current I_g will flow in the path R, C_g, R_g, its value being $\frac{222}{22 \cdot 2 \times 10^6}$ amperes or 10 microamperes. The P.D. between the ends of the grid leak will be $R_g\ I_g = (\cdot 2 \times 10^6)\ (10 \times 10^{-6}) = 2$ volts, and the grid will be positive with respect to the filament by this amount. Hence the grid bias voltage actually applied must be two volts greater than that indicated by a consideration of the $I_a - V_g$ curve of the valve. If the insulation resistance falls below the value given, an even greater positive bias will be applied to the grid, and this must be neutralized by the application of opposite bias. In practice, therefore, the insulation resistance of the grid condenser must be maintained at a high value, and moisture, dirt or dust will inevitably lead to a reduction in the amplification. Only high quality mica dielectric is suitable for grid condensers.

Choke-capacitance coupling

21. (i) Instead of placing a resistance in the anode circuit in order to develop the amplified voltage, an inductive coil may be used, as already shown. The coupling to the succeeding valve is made by means of a grid condenser as in the case of resistance-capacitance coupling. It has already been demonstrated that at the lower audio frequencies the effect of the shunt capacitance is negligible and the V.A.F. is practically equal to

$$\frac{v_a}{v_{g1}} = \frac{\omega L}{\sqrt{r_a^2 + (\omega L)^2}}\ \mu$$

The insertion of the grid condenser and leak will cause the stage gain to be less than this, for those frequencies at which the reactance of the grid condenser is comparable with the resistance of the leak, so that the gain is

$$\frac{v_{g2}}{v_{g1}} = \frac{R_g}{\sqrt{R_g^2 + \left(\frac{1}{\omega C_g}\right)^2}} \times \frac{\omega L \mu}{\sqrt{r_a^2 + (\omega L)^2}}, \text{ approximately.}$$

In practice, the inductance of the choke is so chosen that in conjunction with its own self-capacitance and the effective shunt capacitance previously alluded to, the anode circuit is a rejector for some frequency in the middle portion of the audio-frequency range, say 1,000 cycles per second. Allowing say 100 $\mu\mu F$ for the self-capacitance of the coil, the total capacitance may be of the order of 200 $\mu\mu F$. The inductance required to tune to 1,000 cycles per second may now be found from the formula

$$f = \frac{1}{2\pi\sqrt{L\ C.}}$$

$$\text{i.e. } L = \frac{1}{2^2\ \pi^2\ f^2\ C.} \quad (\text{N.B. } 2^2\ \pi^2 \doteqdot 40)$$

$$\text{or } L = \frac{10^{12}}{40 \times 10^6 \times 200} = 125 \text{ henries.}$$

At higher frequencies the amplification will fall off owing to the effect of the shunt capacitance and the overall response of the amplifier will be somewhat as shown in fig. 13.

(ii) This form of coupling is superior to resistance-capacitance coupling in that a somewhat higher stage gain can be achieved, and also because the steady voltage drop in the inductance is negligible, so that a lower H.T. supply voltage may be used. These advantages are however completely offset by the weight and cost of the inductance, and the serious reduction of amplification which occurs at the higher and lower ends of the audio-frequency range. Resistance-capacitance coupling is therefore generally preferred for service use, except where any particular stage is required to act as a note selector.

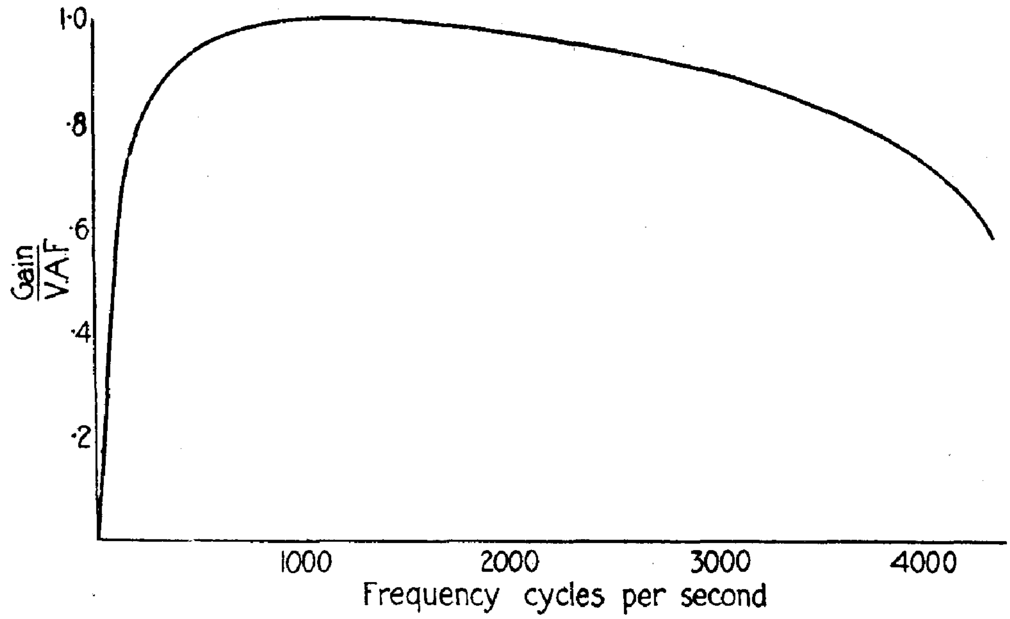


FIG. 13, CHAP. XI. Frequency response of choke-capacitance coupled amplifier.

Note selector

22. In C.W. reception the heterodyne beat note is under the control of the operator, and a high degree of selectivity can be attained by the use of a sharply tuned audio-frequency stage. The selectivity of a choke-capacitance coupled stage depends chiefly upon the design of the anode circuit impedance, an even response being obtained by a high $\frac{L}{C}$ ratio ; in a selective amplifier the inductance is reduced and the capacitance increased, the product $L\,C$ being so chosen that the anode circuit is resonant to the frequency at which the telephone receivers give maximum response, i.e. about 1,000 cycles per second. The circuit diagram is given in fig. 14. This is so arranged that when high audio-frequency selectivity is not desired, e.g. for R/T reception, the tuned circuit can be switched out and a resistance R_0 substituted.

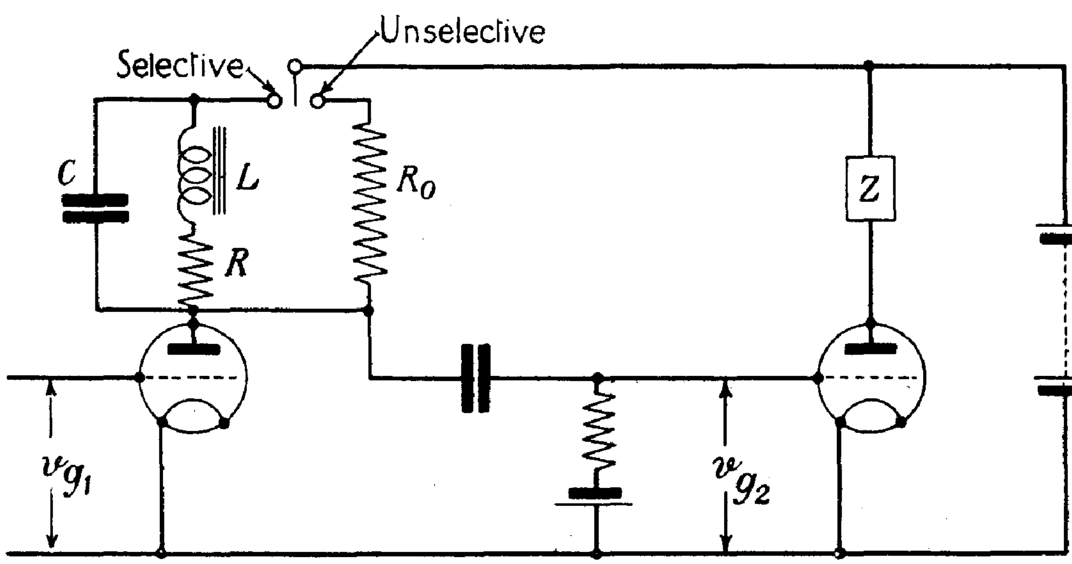


FIG. 14, CHAP. XI.—Note selector.

23. An expression showing the V.A.F. of a "tuned anode" amplifier at any frequency $\frac{\omega}{2\pi}$ may be derived as follows. The impedance operator of the inductive branch of the anode circuit is $R + j\omega L$, while that of the capacitive branch is $\frac{1}{j\omega C}$. The impedance operator of these branches in parallel is

$$z = \frac{\frac{1}{j\omega C}(R + j\omega L)}{R + j\left(\omega L - \frac{1}{\omega C}\right)}$$

If R is small compared with ωL this approximates closely to

$$z = \frac{L}{C\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}}$$

The current established in the anode circuit by a sinusoidal voltage v_{g1} is, in the notation of paragraph 64, Chapter V,

$$\mathbf{i}_a = \frac{\mu}{r_a + \frac{L}{C\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}}} \mathbf{v}_{g1} \qquad (1)$$

and the P.D. across the output circuit is

$$\mathbf{v}_a = \frac{\frac{L}{C\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}}}{r_a + \frac{L}{C\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}}} \mu\, \mathbf{v}_{g1}$$

$$= \frac{\frac{L}{C}}{r_a\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\} + \frac{L}{C}} \mu\, \mathbf{v}_{g1}$$

$$= \frac{\frac{L}{Cr_a}}{R + \frac{L}{Cr_a} + j\left(\omega L - \frac{1}{\omega C}\right)} \mu\, \mathbf{v}_{g1} \qquad (2)$$

The scalar value of the instantaneous output voltage is therefore

$$v_a = \frac{\frac{L}{Cr_a}}{\sqrt{\left\{\left(R+\frac{L}{Cr_a}\right)^2+\left(\omega L-\frac{1}{\omega C}\right)^2\right\}}}\mu\, v_{g1} \qquad (3)$$

and the V.A.F. is $\frac{v_a}{v_{g1}}$ or

$$\text{V.A.F.} = \frac{\frac{L}{Cr_a}}{\sqrt{\left\{\left(R+\frac{L}{Cr_a}\right)^2+\left(\omega L-\frac{1}{\omega C}\right)^2\right\}}}\mu \qquad (4)$$

At the resonant frequency $\omega L - \frac{1}{\omega C} = 0$ and the V.A.F. is equal to $\frac{R_d}{r_a + R_d}\mu$. The degree to which other frequencies are attenuated depends upon the ratio $\frac{L}{C}$ and upon the resistances R and r_a. To obtain the highest possible selectivity the circuit constants are so chosen that $\frac{L}{Cr_a} = R$ and under these conditions the V.A.F. at the selected frequency is equal to $\frac{\mu}{2}$. It should be appreciated that it is theoretically possible to achieve a considerably greater degree of note selectivity than can be usefully employed for C.W. reception, the difficulty of maintaining

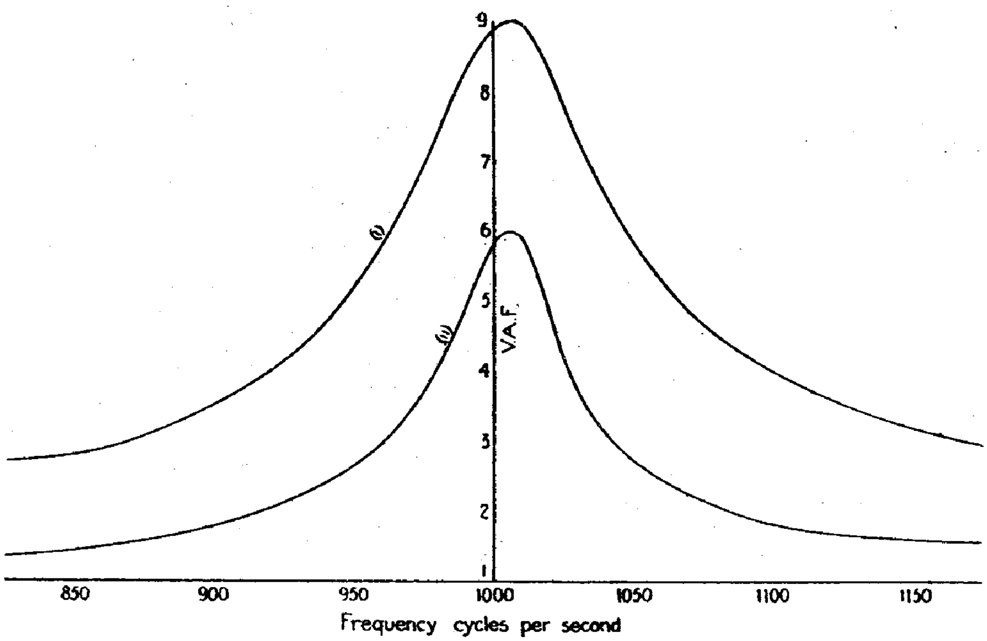


FIG. 15, CHAP. XI.—Response curves of note selectors.

a heterodyne note of constant frequency being a limiting factor. The curves in fig. 15 illustrate this. Curve (i) is the response curve of a note selector in which $L = \cdot 125$ henry, $C = \cdot 2\mu F$ and $R = 12\cdot 5$ ohms, the valve having an anode A.C. resistance of 12,500 ohms and an amplification factor of 10. Experience shows that suitable chokes for this purpose may be expected to have a resistance of about 100 ohms per henry, i.e. a magnification of 62·5 at 800 cycles per second, the latter frequency being usually adopted for purposes of standardization. Assuming that this magnification is obtainable at about 1,000 cycles per second, curve (ii) gives the theoretical optimum selectivity obtainable with the same valve. In this case L = ·03125 henries, $C = \cdot 8\mu F$ $R = 3\cdot 125$ ohms. It will be seen that the V.A.F. at the selected frequency is now $\frac{\mu}{2}$, but that non-resonant frequencies are greatly attenuated.

24. For the purpose of estimating the variation of signal strength with a given input, it is convenient to draw these response curves on the basis of " decibels below the standard response " against " cycles per second off resonance", using the relation

$$\left.\begin{array}{l}\text{decibels below the standard}\\ \text{response, at frequency } f\end{array}\right\} = 20\, log_{10}\left\{\frac{\text{V.A.F. at resonant frequency}}{\text{V.A.F. at frequency } f}\right\}$$

Thus, taking curve (i), at the resonant frequency, the V.A.F. is 8, and at 1,025 cycles per second —20 cycles per second off resonance—the V.A.F. is 6·75. $Log_{10}\, \frac{8}{6\cdot 75} = 0\cdot 07335$, and the response is 20 × ·07335 = 1·476 db. below that at resonance. Taking a number of points on curve (i) of fig. 15 in this way, we derive curve (i) of fig. 16. We see that, on the assumption that one signal strength on the arbitrary " audibility scale " corresponds to 6 db., a variation

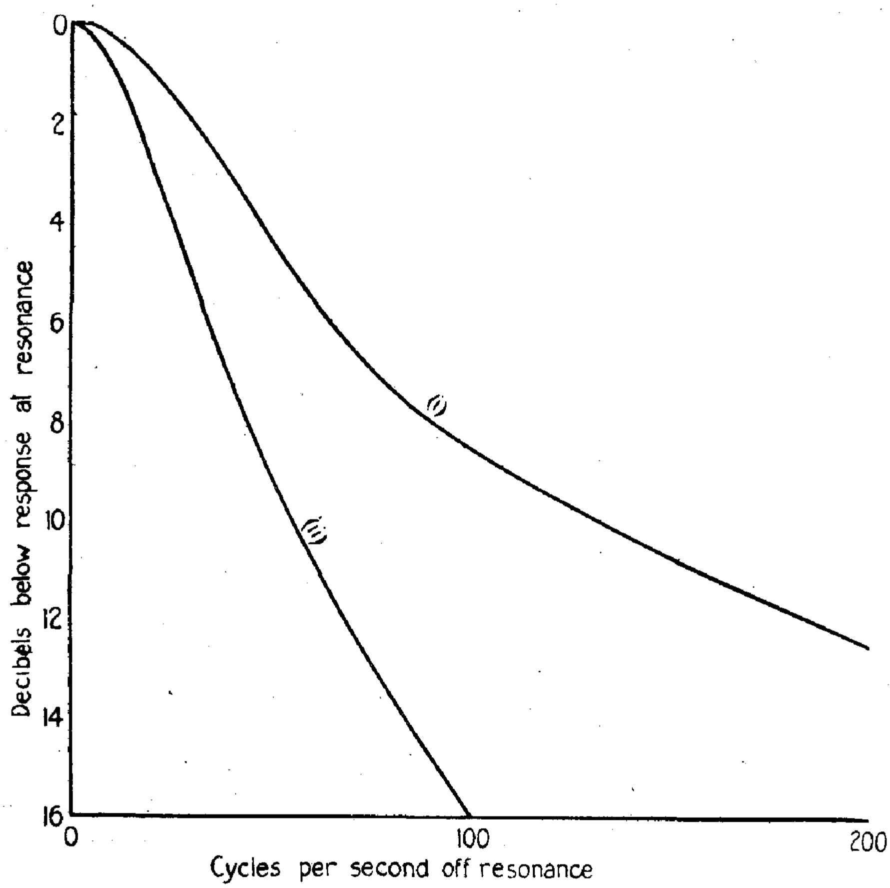


Fig. 16, Chap. XI.—Curves of **Fig. 15** plotted in decibels.

of 200 cycles per second in the heterodyne beat note will cause a variation of two units of signal strength. For this note selector to be of use to the receiving operator, the transmitter frequency must be controlled to within about 200 cycles per second, and preferably to within only 100 cycles per second. This necessitates a stability of 1 part in 10,000 in an operating radio-frequency of 1 Mc/s, even if the heterodyne oscillator is perfectly stable. Curve (ii) of fig. 16, which gives the response in db. corresponding to curve (ii) of fig. 15, shows that a fall of 6 db. will result if the frequency varies only about 30 cycles per second. This degree of stability can only be maintained at comparatively low radio frequencies.

25. To summarize the above, then, it may be said that curves (i) in figs. 15 and 16 respectively show the performance of the best note selector which is at present practicable for general purposes. Where it is possible to employ transmitters of very high frequency stability, i.e. in high power ground-to-ground communication on a single fixed frequency, note selectors of considerably better performance are theoretically possible. In super-heterodyne C.W. receivers (*see* paras. 81 *et seq.*) the selectivity of the intermediate frequency amplifier or amplifiers must be taken into account and it is found that a note selector having the response of curve (i) is entirely adequate.

Transformer coupling

26. In the transformer-coupled amplifier the anode circuit impedance is the primary winding of a transformer, the secondary voltage being applied to the grid and filament of the succeeding valve, as shown in fig. 17a. This type of amplifier has an advantage over the two previous types, in that the transformer itself may be designed to give a voltage step-up, while no grid condenser and leak is necessary. These advantages are offset to some extent by the greater weight and cost, while even an approach to uniform amplification over a wide frequency band can only be achieved by very careful design of the transformer, and by the choice of a suitable valve for use with it. An exact analysis of the transformer coupled audio-frequency amplifier, for the purposes of obtaining a general expression for the stage gain, is extremely laborious, and even when obtained the expression is difficult to interpret, owing to the large number of circuit constants which must be taken into account. It is however possible to derive comparatively simple expressions for the stage gain at low, medium and high audio frequencies respectively ; these are generally only in error by a few parts in one hundred.

27. Referring to fig. 17b, the equivalent circuit is seen to consist of an ideal transformer taking no magnetizing current and having no losses, so that it serves merely to increase the voltage v_a to $Tv_a = v_{g2}$. The magnetizing current is assumed to flow through the inductance L_p which is equal in magnitude to the actual primary inductance. Strictly, the resistance of the primary winding should also be included, in series with L_p. For the present purpose, however, both iron and copper losses may be represented by the resistance R_l, while $\frac{R_i}{T^2}$ represents the transferred input resistance of the succeeding valve (T_2 of fig. 17(a)). The joint resistance of R_l and $\frac{R_i}{T^2}$ may be denoted by R_e. Similarly C_e represents the whole of the effective capacitance between anode and filament of T_1. It is made up of several components, i.e. C_{af}, the inter-electrode capacitance of the valve T_1, including any distributed capacitance in parallel therewith, T^2C_s, where C_s is the distributed capacitance of the secondary winding plus the input capacitance of T_2, and $(T-1)^2\,C_m$ where C_m is the distributed capacitance between the primary and secondary windings. Thus

$$C_e = C_{af} + T^2C_s + (T-1)^2\,C_m$$

The inductance L_e represents the total leakage inductance of the transformer, transferred to the primary winding. At low audio frequencies, the reactance of C_e is very large. If it is considered to be infinite, the voltage v_a across the primary winding depends only upon the relative magnitudes of ωL_p and R_e ; if R_a is very much greater than ωL_p it may be entirely neglected, so that

$$i_a = \frac{\mu\, v_{g1}}{\sqrt{r_a^2 + (\omega L_p)^2}}$$

$$v_a = \frac{\omega L_p}{\sqrt{r_a^2 + (\omega L_p)^2}}\,\mu\, v_{g1}$$

$$\frac{v_{g2}}{v_{g1}} = \frac{Tv_a}{v_{g1}} = \frac{\omega L_p}{\sqrt{r_a^2 + (\omega L_p)^2}}\,T\,\mu. \quad (\text{"Formula A."})$$

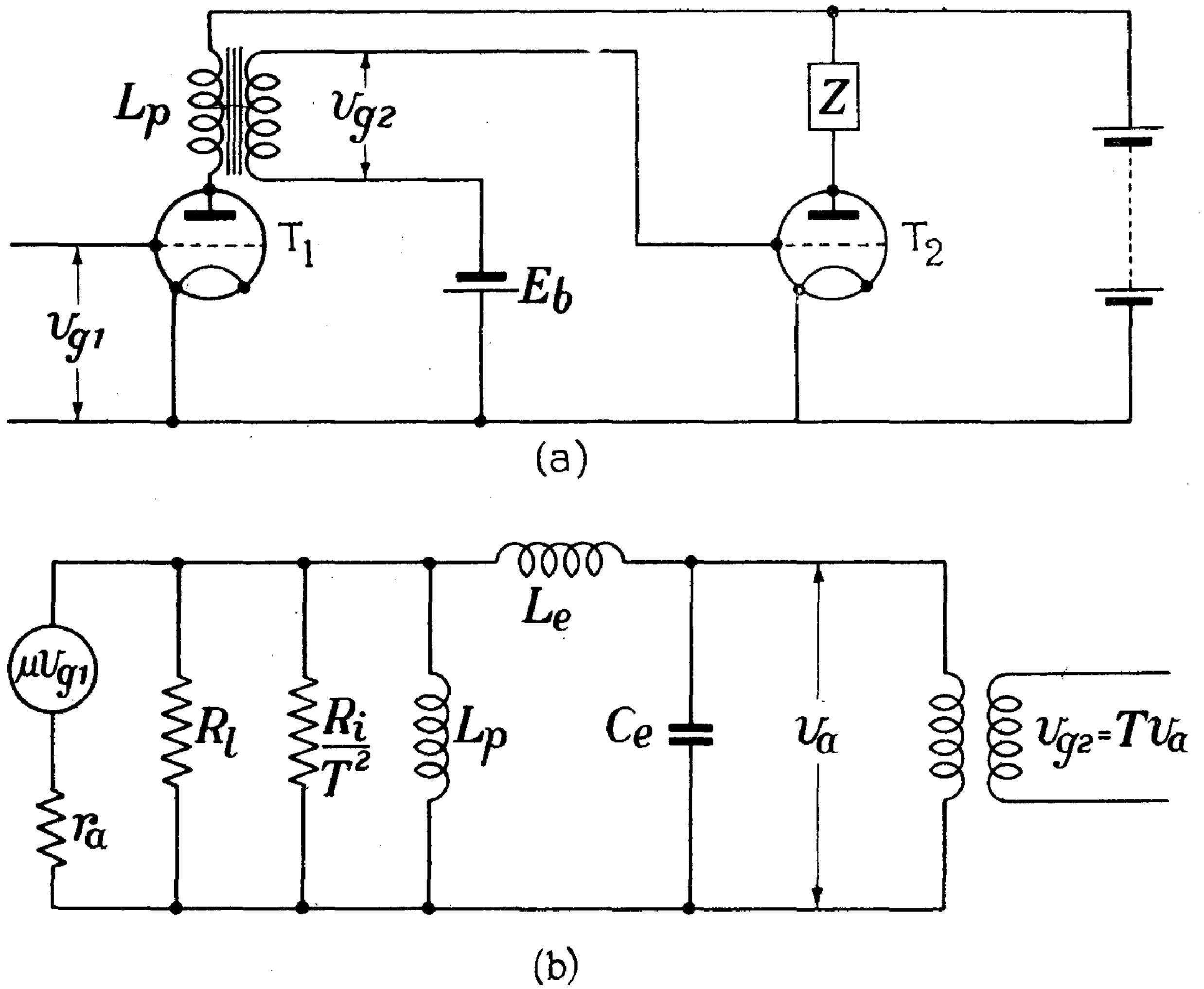


FIG. 17, CHAP. XI.—Transformer-coupled amplifier and equivalent circuit.

At medium audio frequencies, however, while the reactance of C_e is still large compared with R_e, the reactance ωL_p is large compared with R_e also, and the anode load impedance is to all intents and purposes the resistance R_e only. This is of course particularly true at the frequency at which $\omega L_p = \frac{1}{\omega C_e}$ (bearing in mind that L_e has no physical existence but is merely an effective inductance representing magnetic leakage). The stage gain in the region of this frequency will therefore be

$$\frac{v_{g2}}{v_{g1}} = \frac{R_e}{r_a + R_e} T\mu. \quad (\text{" Formula B."})$$

At frequencies appreciably higher than the resonant frequency of L_p, C_e, the current through L_p and R_e will be negligible. The leakage inductance and effective capacitance are then, to all intents and purposes, in series with r_a, and

$$i_a = \frac{\mu\, v_{g1}}{\sqrt{r_a^2 + \left(\omega L_e - \frac{1}{\omega C_e}\right)^2}}$$

$$\frac{v_{g2}}{v_{g1}} = \frac{\frac{1}{\omega C_e}}{\sqrt{r_a^2 + \left(\omega L_e - \frac{1}{\omega C_e}\right)^2}} T\mu. \quad (\text{" Formula C."})$$

28. Before illustrating the above principles by numerical examples, it is necessary to appreciate the magnitude of the quantities C_e, L_e, L_p and R_e, and to discuss the transformation ratio T. At first sight it would appear possible to obtain a very high stage gain merely by increasing T, but this is not so. At low audio frequencies a high V.A.F. can only be obtained by making L_p as large as possible. To obtain a large stage gain by virtue of a high value of T would therefore necessitate a corresponding increase in the number of secondary turns, since $T = \frac{N_2}{N_1}$. Such an increase would increase the distributed capacitance of the secondary, i.e would mean an increase in the value of C_s and also in the mutual capacitance C_m. An increase in T therefore increases C_e in greater proportion because $C_e = C_{af} + T^2C_s + (T-1)^2 C_m$. Alternatively, if T is increased by a decrease in the primary turns, C_e is still proportional to $T^2C_s + (T-1)^2 C_m$ and although C_s and C_m are not increased by a reduction of primary turns, the low-frequency response falls off owing to the reduction of L_p, while the medium-frequency response falls off owing to an increase in iron and copper losses, which is in effect a decrease of R_e. It should be noted that if there were no transformer losses R_e would be infinite and not zero. The transformation ratio therefore lies between limits of from 2 to 10, the higher values being only employed where uniformity of frequency response is unimportant. The capacitance C_e rarely exceeds 3,000 $\mu\mu F$; C_{af} will generally be of the order of 10 $\mu\mu F$, C_s will not usually exceed 200 $\mu\mu F$ and C_m 10 $\mu\mu F$. If $T = 2$,

$$C_e = 10 + 800 + 10 = 820\ \mu\mu F,$$

while if $T = 10$

$$C_e = 10 + 2{,}000 + 810 = 2{,}820\ \mu\mu F.$$

The effective resistance R_e usually lies between 10^5 and 5×10^5 ohms; it depends to some extent upon the transformation ratio and also upon the magnitude of the primary inductance. The effective inductance L_e depends upon the general design of the transformer but is usually about one per cent. of L_p.

29. In the following example the valve is assumed to possess the following constants, viz. $r_a = 10^4$ ohms, $\mu = 10$. The transformation ratio is $3 \cdot 16$ to 1. ($T^2 = 10$) and $R_e = 10^5$ ohms unless otherwise stated while $C_e = 1{,}500\ \mu\mu F$.

Example.—(i) Find the stage gain of a transformer-coupled amplifier at 200 cycles per second if the primary inductance is (*a*) 10 henries, (*b*) 50 henries.

$$(a)\ \omega L_p = 2\pi \times 200 \times 10$$
$$= 12{,}570 \text{ ohms.}$$

Since this is small compared with R_e the stage gain is, by formula A,

$$\frac{\omega L_p}{\sqrt{r_a^2 + (\omega L_p)^2}} T\mu$$
$$= \frac{1 \cdot 257 \times 10^4}{\sqrt{(10^4)^2 + (1 \cdot 257 \times 10^4)^2}} \times 3 \cdot 16 \times 10$$
$$= \frac{1 \cdot 257}{\sqrt{1 + 1 \cdot 257^2}} \times 31 \cdot 6$$
$$= 24 \cdot 8.$$

$$(b)\ \omega L_p = 2\pi \times 200 \times 50$$
$$= 62{,}800 \text{ ohms.}$$

By formula A,

$$\frac{v_{g2}}{v_{g1}} = \frac{6 \cdot 28 \times 10^4}{\sqrt{(10^4)^2 + (6 \cdot 28 \times 10^4)^2}} \times 31 \cdot 6$$
$$= \frac{6 \cdot 28}{\sqrt{41}} \times 31 \cdot 6$$
$$= 30 \cdot 6.$$

Since, however, ωL_p is of the same order as R_e, it is possible that formula B may give a more accurate result. By this method

$$\frac{v_{g^2}}{v_{g1}} = \frac{R_e}{r_a + R_e} T\mu$$

$$= \frac{10^5}{10^4 + 10^5} \times 31{\cdot}6$$

$$= 28{\cdot}6.$$

The latter result is probably more nearly correct than the former.

(ii) If the primary inductance is 50 henries, and the leakage reactance ·5 henry, find the stage gain at the frequency for which $\omega L_e = \frac{1}{\omega C_e}$.

By formula C,

$$\frac{v_{g2}}{v_{g1}} = \frac{\frac{1}{\omega C_e}}{\sqrt{r_a^2 + \left(\omega L_e - \frac{1}{\omega C_e}\right)^2}} T\mu$$

$$= \frac{1}{\omega C_e r_a} T\mu$$

$$= \frac{T\mu}{r_a}\sqrt{\frac{L_e}{C_e}}$$

$$= \frac{31{\cdot}6}{10^4}\sqrt{\frac{{\cdot}5 \times 10^{12}}{1{,}500}}$$

$$= 57{\cdot}6,$$

which is greater than $T\mu$. The frequency at which this occurs, i.e. $\frac{1}{2\pi\sqrt{L_e C_e}}$, is in this particular instance about 5,800 cycles per second.

30. It will be seen that the effect of the subsidiary resonance between L_e and C_e is to give increased amplification at the higher audio frequencies ; by careful attention to the relative magnitudes of r_a, L_p, C_e and L_e, the response curve may be made substantially flat up to 8,000 or 10,000 cycles per second. The effect of the magnitude of r_a is somewhat as shown in fig. 18. It will be seen that in order to obtain a high amplification at both ends of the frequency scale, r_a should be as low as possible. Since $\mu = r_a g_m$ and g_m is always made as high as possible, this implies that an even response over a wide frequency range is only obtained by using a valve of low amplification factor. For morse reception, the chief requirement of the transformer is a high turns ratio, giving a high amplification in the neighbourhood of 1,000 cycles per second ; reduced amplification of frequencies below about 800 or above 2,000 being an advantage rather than otherwise, for a certain degree of audio-frequency selectivity is then obtained. For reception of R/T, uniform amplification of the band covering from 400 to 2,000 cycles per second will give sufficient intelligibility, but for the reception of entertainment programmes it is usual to aim at even amplification of all frequencies from about 80 to 5,000, and an even wider band is desirable.

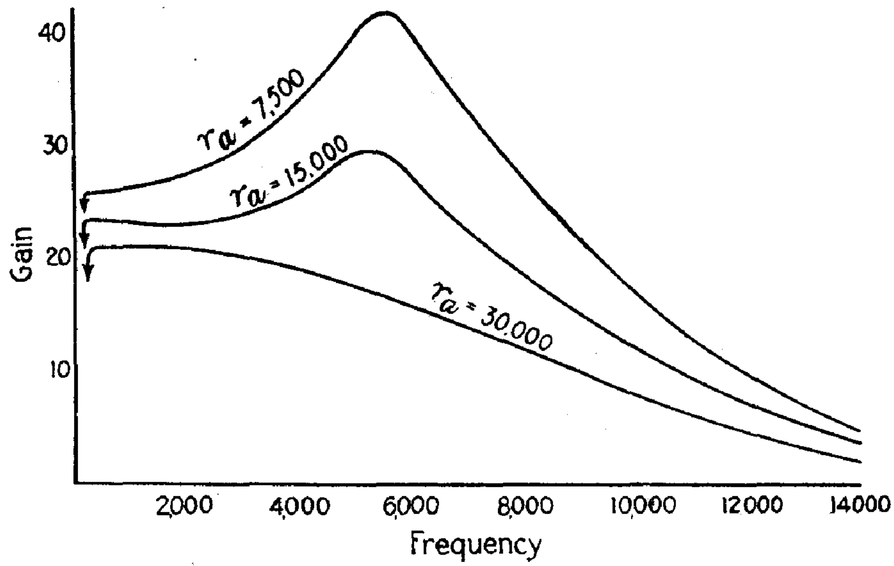


FIG. 18, CHAP. XI.—Effect of r_a upon high-note response.

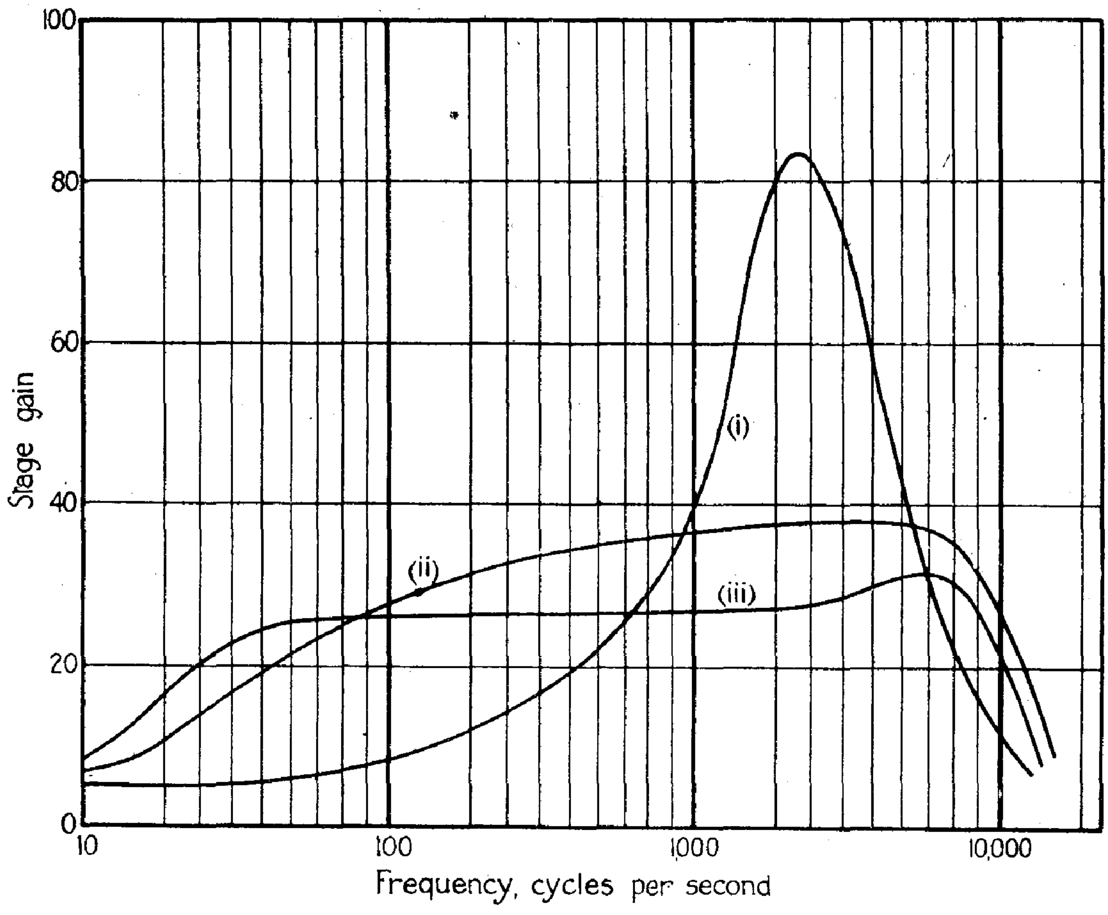


FIG. 19, CHAP. XI.—Response curves of different A.F. transformers.

31. These stringent requirements can only be met by designing the transformer with a very high primary inductance and small leakage reactance. For reasons already stated a high primary inductance is only achieved at the expense of a low turns ratio. Fig. 19 shows response characteristics of three different transformers. Curve (i) is that of a transformer perfectly suitable for C.W. reception, giving a high amplification only over the band 1,500 to 4,000 cycles. The primary inductance is 16 henries and the turns ratio 9/1. Curve (ii) is that of a transformer suitable for ordinary R/T communication, the amplification being practically even between 800 and 8,000 cycles. Its primary inductance is 10 henries and turns ratio 6/1. Curve (iii) is that of a transformer designed for reception of broadcast entertainment programmes. Its range of even amplification is practically from 60 to 9,000 cycles per second, the primary inductance being 70 henries and the turns ratio 3·5/1. It should be clearly understood that this comparatively high standard of reproduction is not necessary for the transmission of speech alone, i.e. for commercial or service communication, and may even be a disadvantage when electrical interference exists. The slight increase of gain between 5,000 and 9,000 cycles per second serves to compensate for the high frequency cut-off due to the selectivity of the pre-detector stages.

Parallel-feed transformer coupling

32. The primary inductance of a transformer is reduced by the presence of an appreciable steady flux in the core, such as is caused by the steady component of anode current. In some cases a circuit similar to fig. 20a is adopted in order to avoid this steady magnetization, the steady component of the anode current passing through the feed resistance R_f, and the alternating component through the circuit C, r, L_p. The circuit is then, in effect a combination of resistance and transformer coupling. The capacitance of the blocking condenser C is chosen with regard to the primary inductance, in such a manner that the circuit L_p, C, fig. 20b, is an acceptor for some low audio frequency, e.g. that at which $\omega L_p = r_a$. At this frequency, the opposition offered by the whole external circuit is that of r and R_f in parallel, and the anode current is

$$i_a = \frac{\mu\, v_{g1}}{r_a + \dfrac{rR_f}{r + R_f}}$$

Of this current, a fraction $\dfrac{R_f}{r + R_f}$ flows through the transformer primary, i.e.

$$i_L = \frac{\dfrac{R_f}{r + R_f}}{r_a + \dfrac{rR_f}{r + R_f}}\,\mu\, v_{g1}$$

$$= \frac{R_f}{r_a r + r_a R_f + rR_f}\,\mu\, v_{g1}$$

$$= \frac{1}{r_a\left(\dfrac{r}{R_f} + 1\right) + r}\,\mu\, v_{g1},$$

and the voltage v_a between the primary terminals is $\omega L_p\, i_L$ hence

$$\text{V.A.F.} \doteqdot \frac{v_a}{v_{g1}} = \frac{\omega L_p}{r_a\left(\dfrac{r}{R_f} + 1\right) + r}\,\mu$$

$$\doteqdot \frac{\omega L_p}{r_a}\,\mu,$$

and $\dfrac{v_{g2}}{v_{g1}} = \dfrac{\omega L_p}{r_a}\,T\mu.$

Since $\omega L_p = r_a$, the V.A.F. at the resonant frequency of the circuit $L_p\, C$ is very nearly equal to μ, and the stage gain equal to $T\mu$, whereas with the usual series connection it is $\frac{T\mu}{\sqrt{2}}$. Provided that the primary inductance L_p is of a high order, e.g. 50 henries, the low-frequency response is greatly improved by the adoption of parallel feed. In order to maintain a high amplification

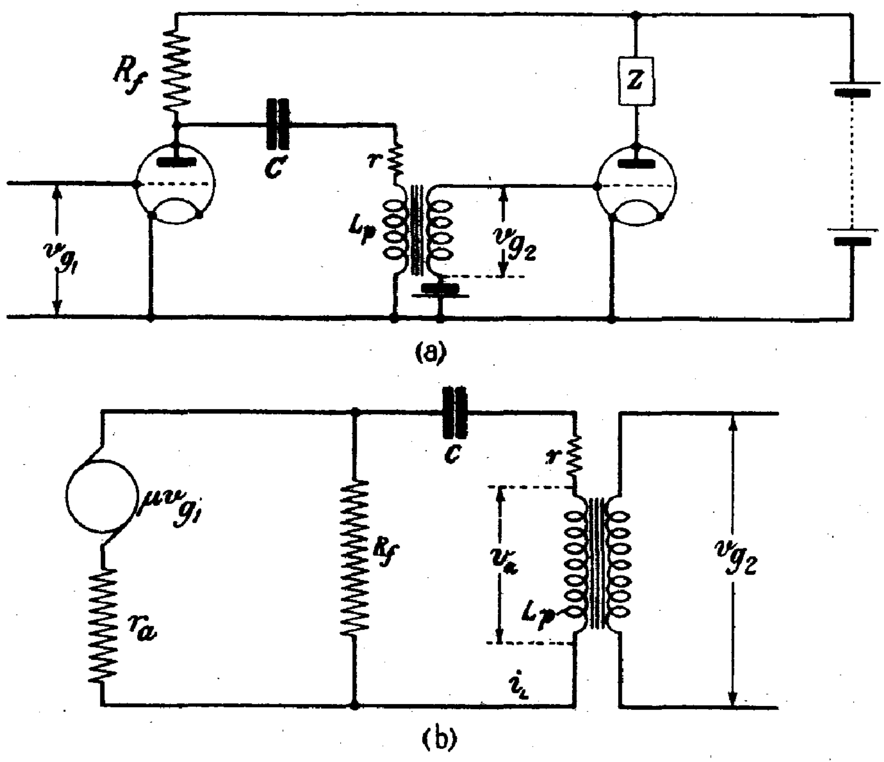


FIG. 20, CHAP. XI.—Parallel feed transformer coupling.

at frequencies above resonance, the feed resistance should be at least equal to $2\,r_a$. This necessitates an increase in the H.T. supply voltage, and consequently the arrangement is rarely adopted in battery-operated receivers.

Power amplification

33. The anode circuit load impedance in the final stage of an audio-frequency amplifier is the reproducing device, i.e. telephones or loudspeaker. Since the latter can only operate if supplied with electric power the final stage must function as a power amplifier, and may be adjusted to meet either of two requirements. These are (i) maximum power output for a given input voltage, or (ii) maximum undistorted power with the greatest input voltage that can be usefully employed. If it is desired to achieve the first condition, since the maximum power is to be developed irrespective of distortion, grid current may be allowed to flow during positive half-cycles of grid voltage, and the anode current excursion may be allowed to enter the lower curved portion of the characteristic curve of the valve. In either event, the load impedance should preferably be purely resistive, or if this is not practicable, the power factor should be as high as possible.

Conditions for maximum power output

34. The power developed in the output circuit will now be derived, first for maximum power with a given grid voltage. Assuming that the load is a dynamic resistance of R ohms, and the input voltage to be V_g (R.M.S.) the R.M.S. anode current is

$$I_a = \frac{\mu V_g}{r_a + R}$$

The output voltage V_a will be

$$RI_a = \frac{R}{r_a + R} \mu V_g,$$

and the power developed in the load resistance, is

$$P = \frac{R}{(r_a + R)^2} \mu^2 V_g^2.$$

For a given input voltage, the output power is directly proportional to $\frac{R}{(r_a + R)^2}$. This expression is a maximum when $R = r_a$ and the maximum possible output from a given input voltage V_g is

$$P_{max} = \frac{\mu^2 V_g^2}{4\, r_a} = \frac{\mu^2 \mathscr{V}_g^2}{8\, r_a}$$

where $\mathscr{V}_g$ is the peak value of the input voltage. The manner in which the power output varies for different values of the ratio $\frac{R}{r_a}$ is shown in fig. 21. The fall in output resulting from some slight mismatching is not serious, 90 per cent. of the maximum output being obtained when $R = \frac{r_a}{2}$ and also when $R = 2\, r_a$.

35. The power obtainable from an amplifier having a reactive load impedance of power factor $\cos \varphi$ will depend both upon the magnitude of the impedance and the power factor. This may be demonstrated as follows:—Let the amplifier have an anode load impedance $Z_o = \sqrt{R_o^2 + X_o^2}$, the anode A.C. resistance being denoted by r_a as usual. With an input voltage V_g (R.M.S.) the effective E.M.F. acting in the anode circuit is μV_g, and the resulting anode current I_a. The power dissipated in the load is $I_a^2 R_o = P$. Since

$$I_a = \frac{\mu\, V_g}{\sqrt{(r_a + R_o)^2 + X_o^2}}$$

$$P = \frac{\mu^2\, V_g^2\, R_o}{(r_a + R_o)^2 + X_o^2}$$

$$= \frac{R_o}{r_a^2 + 2 r_a\, R_o + R_o^2 + X_o^2} \mu^2\, V_g^2$$

$$= \frac{R_o}{r_a^2 + 2 r_a\, R_o + Z_o^2} \mu^2\, V_g^2$$

$$= \frac{\frac{R_o}{Z_o}}{r_a\left(\frac{r_a}{Z_o} + \frac{2\, R_o}{Z_o} + \frac{Z_o}{r_a}\right)} \mu^2\, V_g^2$$

Since $\frac{R_o}{Z_o}$ is the power factor of the load, i.e. $\cos\varphi$,

$$P = \frac{\cos\varphi}{\frac{r_a}{Z_o} + \frac{Z_o}{r_a} + 2\cos\varphi} \times \frac{\mu^2 V_g^2}{r_a}$$

For a constant power factor, this will be a maximum when $\frac{r_a}{Z_o} + \frac{Z_o}{r_a}$ is a minimum, or when $r_a = Z_o$. The power output is then

$$\frac{\cos\phi}{1 + \cos\varphi} \times \frac{\mu^2 V_g^2}{2r_a}.$$

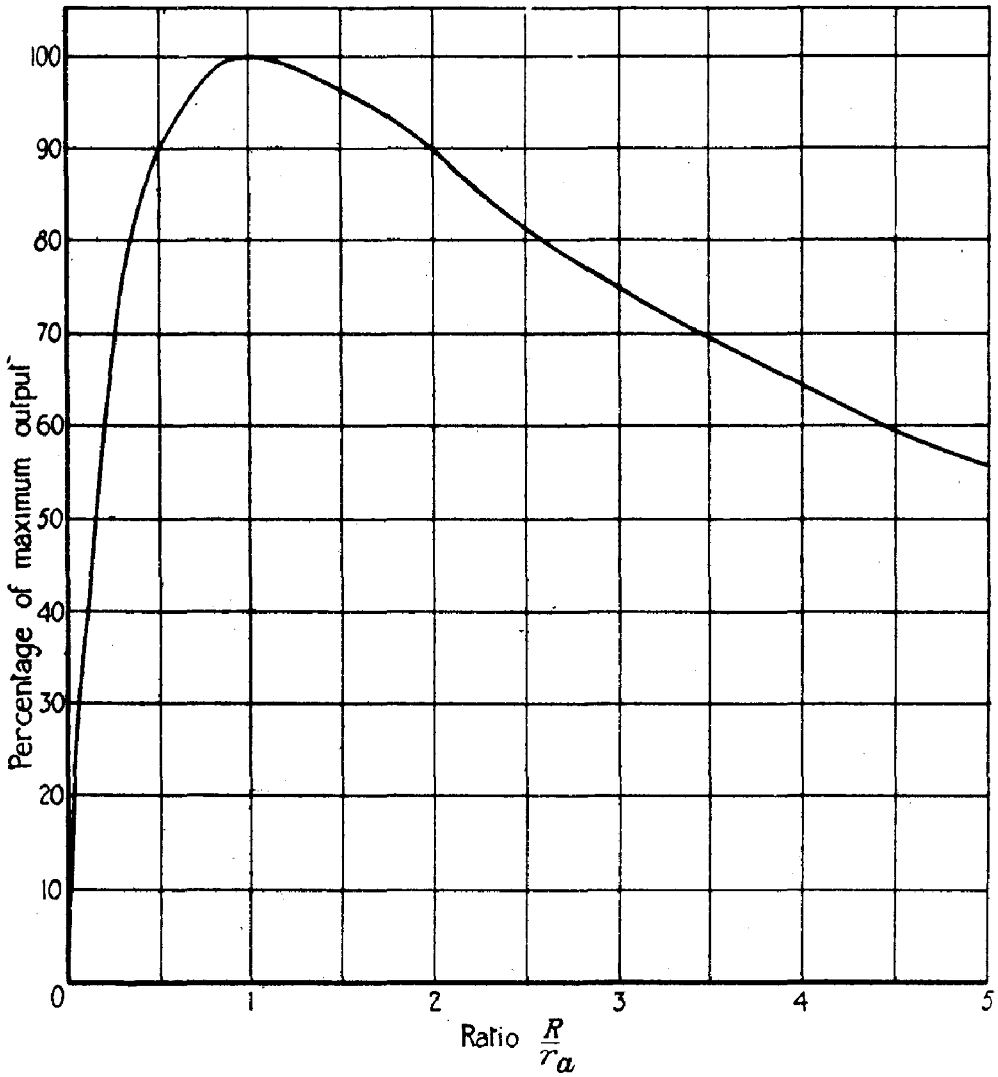


FIG. 21, CHAP. XI.—Variation of power output with ratio $\frac{R}{r_a}$.

For purely resistive loads, $\cos\varphi = 1$, and this expression becomes identical with that derived above.

Example.—Find the power output of a V.R.22 valve when the applied grid-filament voltage has a peak value of 3 volts at 1,000 cycles per second and the anode load is (i) purely resistive, 3,350 ohms, (ii) purely resistive, 13,400 ohms, (iii) an inductance of 1 henry and a resistance of 5,000 ohms, in series, (iv) the power output with optimum resistive load.

The constants of this valve are :—r_a = 6,700 ohms, μ = 16. The R.M.S. input is $\frac{3}{\sqrt{2}}$ volts, therefore $V_g^2 = \frac{9}{2} = 4\cdot5$.

$$P = \frac{R}{(r_a + R)^2} \mu^2 V_g^2$$

In case (i)

$$P = \frac{3{,}350}{(6{,}700 + 3{,}350)^2} \times 16^2 \times 4\cdot5$$
$$= \frac{1}{3{,}350 \times 99} \times 256 \times 4\cdot5$$
$$= \cdot0382 \text{ watts}$$
$$= 38\cdot2 \text{ milliwatts.}$$

In case (ii)

$$P = \frac{13{,}400 \times 16^2 \times 4\cdot5}{(6{,}700 + 13{,}400)^2}$$
$$= \frac{2 \times 256 \times 4\cdot5}{9 \times 6{,}700}$$
$$= \cdot0382 \text{ watts or } 38\cdot2 \text{ milliwatts as before.}$$

In case (iii)

$$Z_o = \sqrt{R_o^2 + X_o^2}$$
$$X_o = \omega L = 2\pi \times 1{,}000 \times 1 = 6{,}280 \text{ ohms}$$
$$R_o = 5{,}000 \text{ ohms}$$
$$Z_o = \sqrt{40 \times 10^6 + 25 \times 10^6}$$
$$= 1{,}000\sqrt{65}$$
$$= 8{,}060 \text{ ohms}$$
$$\frac{r_a}{Z_o} = \frac{6{,}700}{8{,}060} = \cdot831$$
$$\frac{Z_o}{r_a} = 1\cdot2$$
$$\cos\varphi = \frac{R}{Z_o} = \frac{5{,}000}{8{,}060} = \cdot62$$
$$P = \frac{\cos\varphi}{1 + \cos\varphi} \quad \frac{\mu^2 V_g^2}{2r_a}$$
$$= \frac{\cdot62}{1\cdot62} \times \frac{256 \times 4\cdot5}{13{,}400}$$
$$= \cdot0328 \text{ watts or } 32\cdot8 \text{ milliwatts.}$$

In case (iv)

$$P = \frac{\mu^2 V_g^2}{4r_a}$$
$$= \frac{256 \times 4\cdot5}{4 \times 6{,}700}$$
$$= \frac{18 \times 16}{6{,}700}$$
$$= \cdot043 \text{ watts or } 43 \text{ milliwatts.}$$

36. The operating conditions for maximum power output are illustrated in fig. 22 which gives the $I_a - V_a$ characteristics of a small power valve having an r_a of 5,000 ohms and a g_m of 2·4 milliamperes per volt, so that $\mu = 12$. The power dissipation of this valve is 600 milliwatts. In the valve oscillator, which is only a special form of amplifier, grid excitation is maintained so long as the H.T. supply and grid circuits are completed, so that the D.C. power input may be very much greater than that dissipated by the valve. In receiving amplifiers and certain amplifiers used in transmitters, the power input is maintained even when no grid excitation is applied and consequently the power input must never exceed the permissible dissipation of the valve itself.

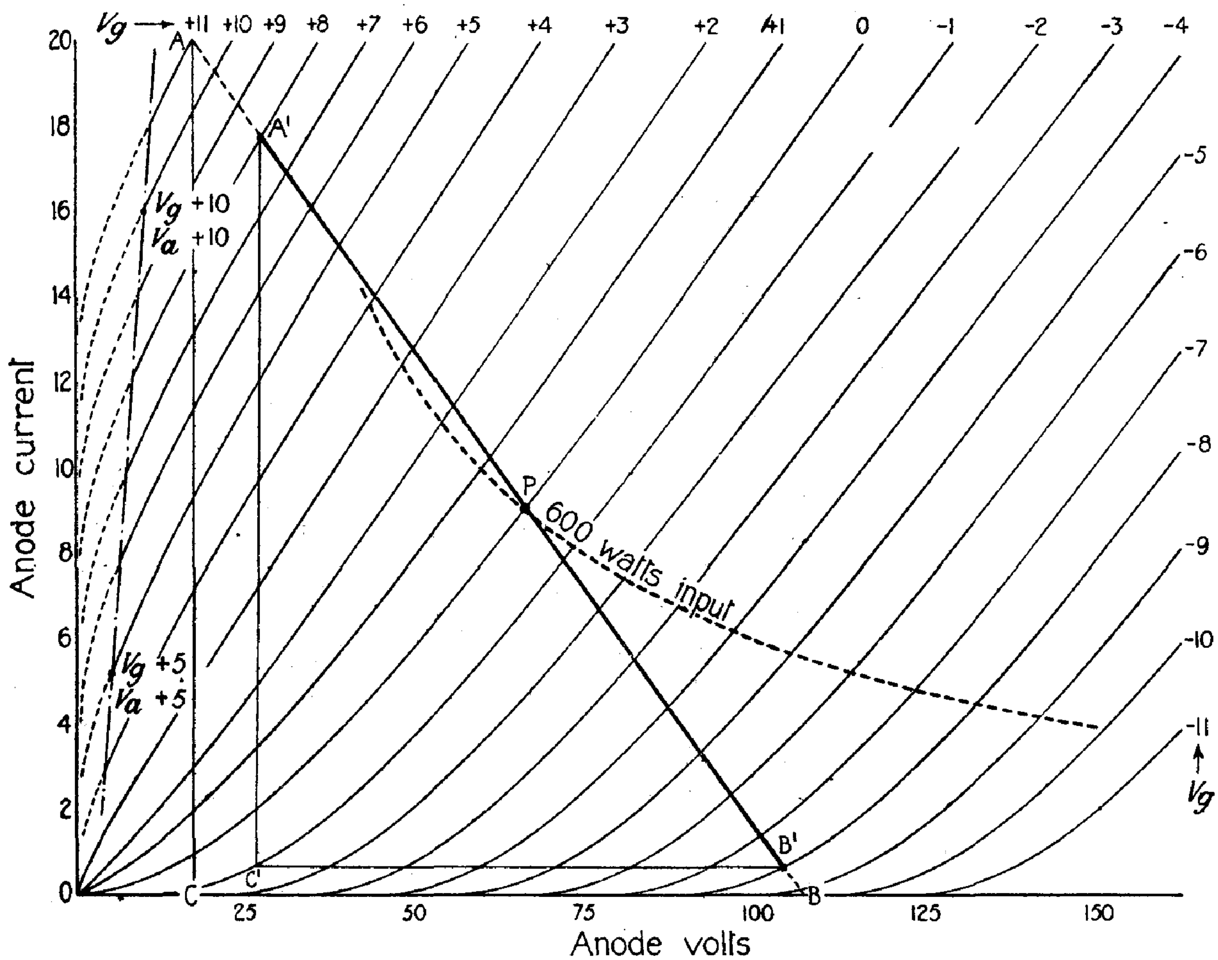


FIG. 22, CHAP. XI.—Operating conditions for maximum power output.

In the present instance then, the operating point P is located at the intersection of the curve $V_g = 0$ and the 600 watts power line. A load line AB corresponding to a resistance of approximately 5,000 ohms has also been drawn through the operating point. The power output is obviously dependent upon the grid swing.

37. In W/T receivers the maximum permissible swing is not subject to limitation in order to avoid distortion but nevertheless is restricted by the fact that the grid must never become positive with respect to the anode, otherwise secondary emission may occur at the grid and the valve will fail to function in the desired manner. The chain-dotted line passing through points such as ($V_g = +5$, $V_a = +5$), ($V_g = +10$, $V_a = +10$), marks the permissible limit of the positive part of the grid swing due to this factor. With this particular valve and load, it is seen that the peak value of the grid voltage must not exceed about 11 volts, and the anode current will be zero during a portion of the cycle, i.e. about the negative peak of the input grid voltage. It is not possible to develop a generally applicable expression giving the output and

efficiency under these conditions, but as an estimate, it is usual to take the output as equal to one quarter the area of the triangle ACB, i.e. to $\frac{AC \times CB}{8}$. In the figure the area of the rectangle AC × CB is 20 (milliamperes) × 90 (volts) or 1,800 milliwatts ; the output power is one-eighth of this or 225 milliwatts, and the efficiency $\frac{225}{600} \times 100$ or 37·5 per cent. The output wave-form will of course be considerably distorted.

38. If the input grid swing is limited to say 16 volts so that the load line is A'B' instead of AB, the distortion will be considerably reduced. Provided that the anode current does not reach zero at any point during the cycle, the ratio of second harmonic to fundamental may be estimated by measuring the length of the load line on each side of the operating point. The length of the positive half of the load line being A'P, and the negative half PB', the percentage of second harmonic distortion is

$$\frac{A'P - PB'}{2\,(A'P + PB')} \times 100$$

Example.—In the original of fig. 22 A'P = $5\frac{7}{16}$ inches, PB' = $5\frac{1}{4}$ inches. Find the percentage of second harmonic introduced.

Expressing A'P and PB' in sixteenths of an inch, this gives the percentage of second harmonic as

$$\frac{87 - 42}{2\,(87 + 42)} \times 100$$
$$= \frac{45}{2 \times 129} \times 100$$
$$= 17{\cdot}5 \text{ per cent.}$$

The power output under these conditions is $\frac{A'C' \times B'C'}{8}$, which is rather less than 180 milliwatts, and the efficiency nearly 30 per cent.

Maximum undistorted output

39. (i) If the second of the above conditions is to be fulfilled, grid current must not be allowed to flow during any portion of the cycle, and the anode current excursion must be confined to the practically straight portions of the $I_a - V_g$ curves. As in the previous instance the load impedance is preferably non-reactive. The curves shown in fig. 23 are somewhat idealized, but will serve to illustrate what is desirable. It is seen that the curves are straight lines except at the extreme foot, i.e. below $I_a = 1$ milliampere. To obtain maximum undistorted output, certain relations between the supply voltage, the grid bias and the load resistance must be fulfilled. The optimum load resistance is equal to $2r_a$. The proof of this is exactly as given in Chapter IX for the case of a sinusoidal oscillator operating without grid current.

(ii) Bearing in mind that grid current must be avoided and that the D.C. input must not exceed the permissible dissipation of the valve, the operating point P may be located as follows. Draw the minimum anode current line ST cutting off the curved portion of the characteristics. Erect a convenient voltage ordinate AB where A lies on the curve $V_g = 0$ and B on the line ST. Locate the point C, making CB = $\frac{1}{4}$ AB. From S, draw the straight line SC producing it to intersect the dissipation (i.e. input power) line at P. Then P is the desired operating point. The optimum load line is now easily found. Through P draw the current ordinate P P', intersecting the voltage ordinate AB at P'. Produce the line AB indefinitely in the upward direction, and locate the point D in such a manner that P'D = 2 P'C. Draw the line DP, producing it to meet the anode voltage base line at F. The required load line then lies upon DF. The maximum grid swing is defined by the intersection of this line with the characteristic $V_g = 0$ and its intersection with the minimum current line ST.

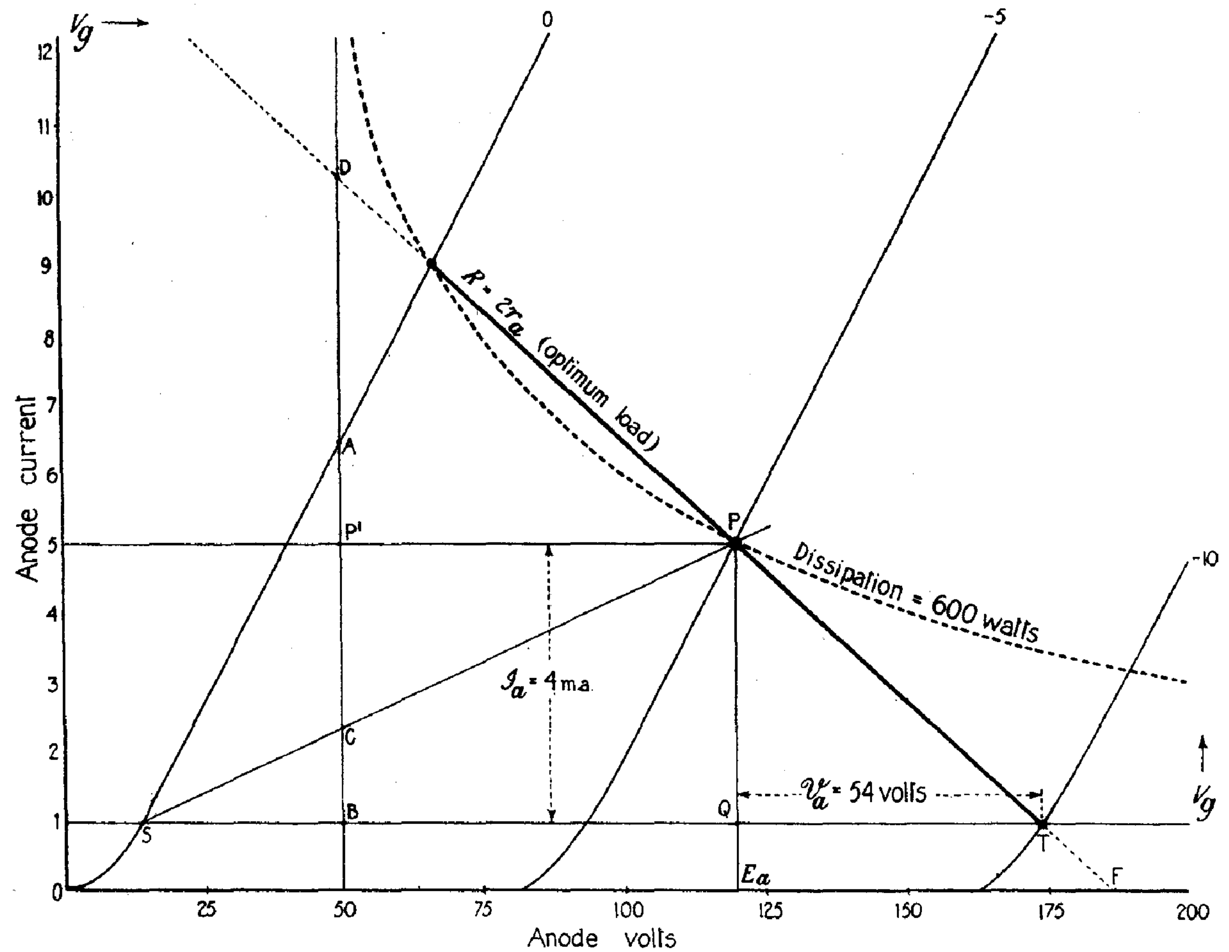


FIG. 23; CHAP. XI.—Theoretical operating conditions, for maximum output without distortion.

40. The co-ordinates of the operating point P give the correct negative grid bias and H.T. voltage for maximum undistorted output, and the peak input grid-filament voltage is equal to the grid bias. In fig. 25 this is 5 volts. The power output is equal to the area of the triangle P Q T. It is seen that PQ = $\mathcal{I}_a$ = 4 milliamperes, QT = $\mathcal{V}_a$ = 54 volts. The power output is $\frac{\mathcal{V}_a \mathcal{I}_a}{2} = \frac{54 \times 4}{2} = 108$ milliwatts. Alternatively, the same result may be derived algebraically from the relation

$$P = \frac{\mu^2 \mathcal{V}_g^2}{2} \times \frac{R}{(r_a + R)^2}$$

Since $R = 2r_a$

$$P = \frac{\mu^2 \mathcal{V}_g^2}{2} \times \frac{2r_a}{9r_a^2}$$

$$= \frac{\mu^2 \mathcal{V}_g^2}{9r_a}$$

$_a = 6{,}500$

$g_m = 2 \cdot 44$ milliamperes per volt

$\mu = 16$

$$\text{Hence} \quad P = \frac{16^2 \times 5^2}{9 \times 6{,}500} \text{ watts}$$
$$= \frac{64{,}000}{9 \times 65}$$
$$= 109 \text{ milliwatts}$$

which agrees very closely with that obtained by direct measurement on the curves of fig. 23. The efficiency is $\frac{109}{600} \times 100 = 18$ per cent. When an amplifier is adjusted to operate without distortion in this way, i.e. as a class A amplifier, its efficiency depends upon the degree of permissible distortion.

41. In practice, the $I_a - V_g$ curves are rarely so straight as in fig. 23 and we are more likely to meet curves like those in fig. 24. Here the straight portion of the characteristic $V_g = 0$ has

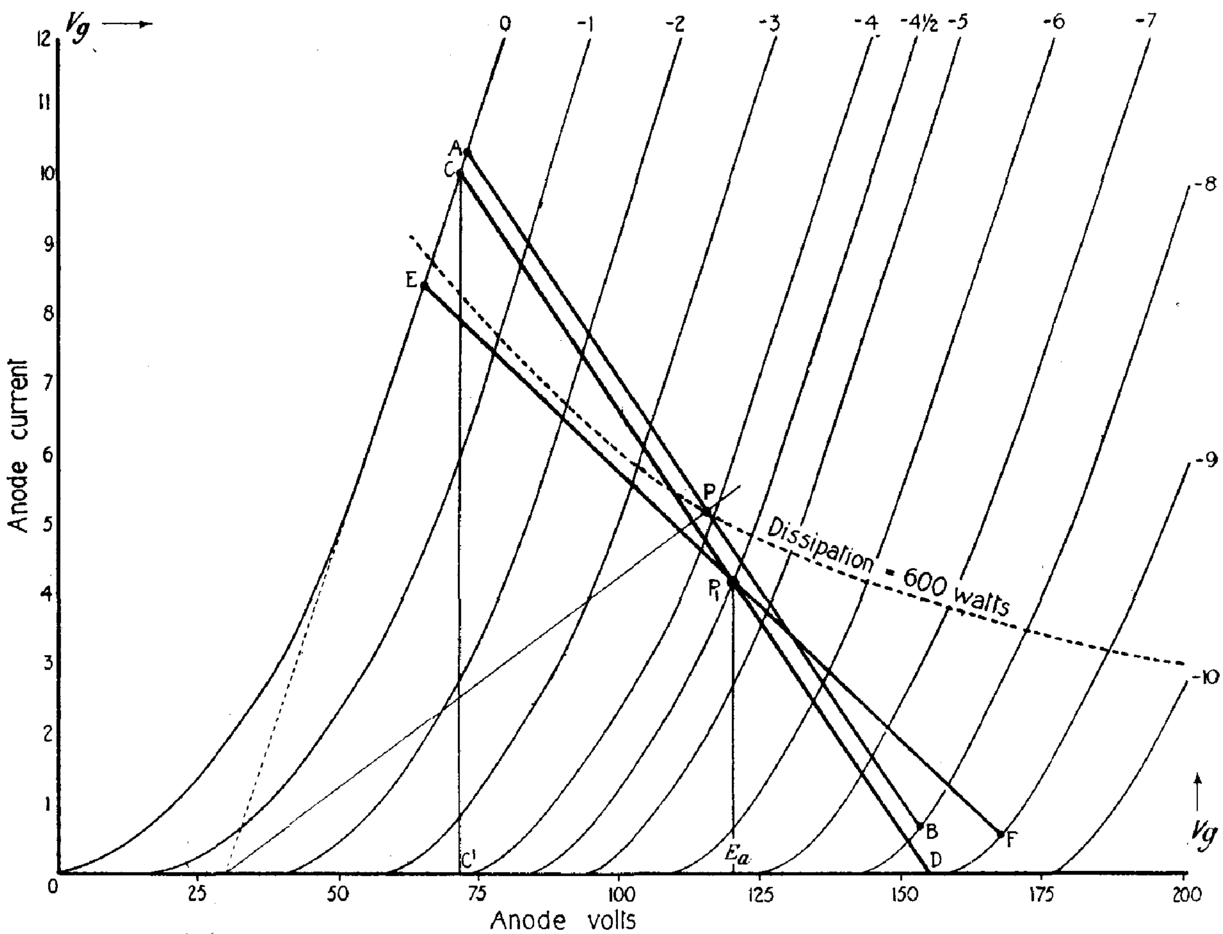


FIG. 24, CHAP. XI.—Practical operating conditions for minimum distortion.

been produced to meet the anode voltage base-line, and the previous construction then performed without drawing in the lower limit of anode current. The theoretical operating point would then be P and the correct load line would be A B. If battery bias only is available, it is usually possible to adjust it only in steps of 1·5 volts, and the H.T. voltage is also limited, e.g. to 120 volts. Consequently it is necessary to choose an operating point such as P_1—which is below the permissible input (600 watt) line—on the curve corresponding to the practicable bias nearest that indicated theoretically; in the present example this is $4\frac{1}{2}$ volts. The load line then becomes CD instead of AB, assuming that the full grid swing of 9 volts is to be applied. Before a final decision is made the percentage of distortion must be checked. In the original drawing $CP_1 = 7$ inches, $P_1D = 5$ inches and the percentage of distortion is

$$\frac{7 - 5}{2\ (7 + 5)} \times 100 = \frac{100}{12} = 8\tfrac{1}{3} \text{ per cent.}$$

The output power is equal to one-eighth of the rectangle CC^1, C^1D. Here $CC^1 = 10$ milliamperes, $C^1D = 84$ volts ; the output is 105 milliwatts and the efficiency 17·5 per cent. Now suppose the load resistance to be increased, giving the load line EF ($R = 12{,}500$ ohms). The output is then only 101 milliwatts and the efficiency about 16 per cent. The distortion is, however, reduced to about 4 per cent.

Figure of merit

42. (i) It has now been shown that the output of an audio-frequency amplifier must always be arrived at as the result of a compromise with regard to the permissible distortion. It is also apparent that to obtain an appreciable power output, a sufficient input grid-filament voltage and a large H.T. supply voltage must be available. Unless these requirements are met it is possible that a power valve may give a smaller output than one not specifically designed for this purpose. In other words, a general purpose triode is often used as an output valve when the input grid swing and H.T. supply voltage are insufficient to obtain increased output by the employment of a power valve.

(ii) It has also been shown that the power output of a given valve is proportional to $\frac{\mu^2}{r_a}$ and the latter is called the figure of merit for any particular valve. The figure of merit of certain valves used in the service is tabulated below.

Valve.	μ	r_a	*Figure of merit.*
V.R.12F	6	18,000	·002
V.R.19	8·5	5,000	·014
V.R.21	13·7	11,500	·016
V.T.20	8	3,700	·017
V.R.22	16	6,700	·038
V.T.23	5	2,000	·0125

The highest figure of merit in this series is that of the valve V.R.22 which is specifically designed as an output valve handling an input swing of about 9 volts. Valves such as the V.T.20 and V.T.23 will give a greater output, but only if a greatly increased grid swing is available. When it is necessary to provide a power output greater than about 100 milliwatts several alternatives present themselves, such as the use of power triodes in parallel or push-pull, a single pentode valve, or pentodes in push-pull. As these requirements are rarely called for except for R/T reception, they are dealt with in Chapter XII.

Use of output transformer

43. In most cases it is not convenient to design the output impedance especially to match one particular type of valve ; the ordinary telephone receiver is, for example, used in conjunction with several of the valves specified above in different service receivers. Fig. 21 shows that the ratio $\frac{R}{r_a}$ may vary from ·5 to 2 without causing an appreciable change in the power output, consequently, when the magnitude of the load impedance differs only slightly from the anode A.C. resistance of the valve, the load may be connected in series as has hitherto been assumed. Where the load impedance differs greatly from the anode A.C. resistance of the valve it is

necessary to use an output transformer. The connections are then as in fig. 25a, the primary winding being connected in series in the anode circuit and the load impedance to the secondary winding. If this load is Z ohms, and the transformation ratio of the transformer is T, the effective load transferred to the primary is $\frac{Z}{T^2}$ ohms, the power factor being practically unaltered. Hence, if the transferred primary load is to equal the anode A.C. resistance of the valve,

$$T = \sqrt{\frac{Z}{r_a}},$$

while for maximum undistorted output,

$$T = \sqrt{\frac{Z}{2\,r_a'}}.$$

44. When the output valve is of the small power type, e.g. the valve receiving V.R.22, or of course any type requiring a greater D.C. power supply, it is however inadvisable to connect the telephones or loudspeakers directly in circuit for two reasons. First, the comparatively

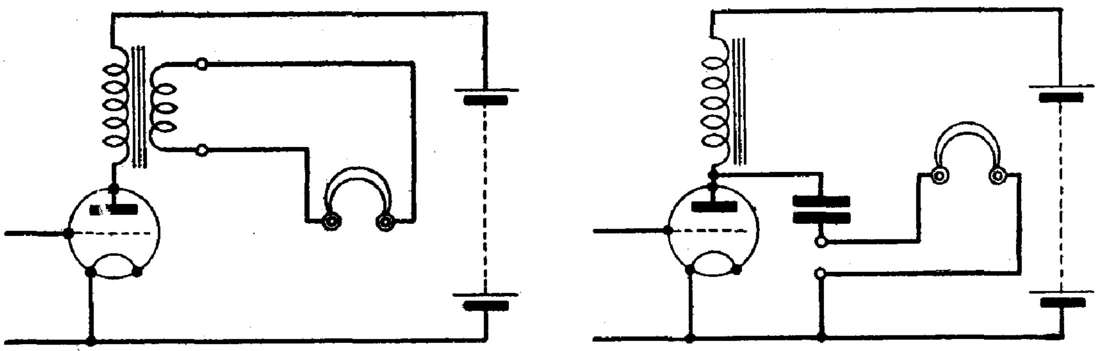


FIG. 25, CHAP. XI.—Typical output circuits for A.F. power amplifier.

large steady component of anode current may lead to overheating of the windings and subsequent breakdown; second, the steady flux caused by this current may be sufficient to saturate or depolarize the magnetic circuit, leading to an inferior response; third, as it is customary to insert the telephones in circuit by means of a plug and jack, the withdrawal of the telephone plug will cause a heavy induced E.M.F. in the anode circuit, which may lead to damage to valves or other components. Even if it is not necessary to use a step-up or step-down transformer for matching purposes, it is usual to use a 1/1 output transformer, which is usually an auto-transformer and is spoken of as the "output choke." As shown in fig. 25b, the telephones are connected across the ends of this inductance, a condenser of large capacitance being inserted in series with them, so that the steady anode current is carried by the choke only. Alternatively the choke and condenser may be said to form a filter, allowing direct current to pass through the choke but not through the telephones, and audio-frequency current to pass through the telephones, but only to a limited extent through the choke. For this filtering action to be efficient, the inductance of the choke should be of the order of ten times that of the telephones.

Radio-frequency amplification

45. The forms of intervalve coupling in general use for radio-frequency amplification are :—

(i) Tuned-anode capacitance coupling.

(ii) Tuned transformer coupling.

Resistance-capacitance coupling and a semi-aperiodic form of transformer coupling were formerly employed to some extent for amplification of frequencies below about 500 kc/s. The advantage of these, namely, fairly uniform amplification of a wide frequency band without the necessity for any tuning adjustment, is more than offset by their lack of selectivity.

Tuned-anode capacitance coupling

46. Fig. 26 shows a tuned aerial circuit, L_a C_a in which a signal voltage developed across the inductance L_a is applied to the grid and filament of the triode T_1. The anode circuit consists of an inductance L and capacitance C in parallel, the inevitable losses in this circuit being represented by the resistance R. The input voltage v_{g1} may be of the order of only a few millivolts. It has been shown earlier in this chapter that for signals of the frequency to which the anode circuit is tuned, the V.A.F. of the triode and its tuned anode circuit is

$$\frac{v_a}{v_{g1}} = \frac{R_d}{r_a + R_d}\,\mu$$

$$\text{where } R_d = \frac{L}{CR}.$$

The grid condenser C_g serves to apply this voltage to the grid of the succeeding valve T_2, the requirements being that its reactance at the operating frequency shall be small compared with

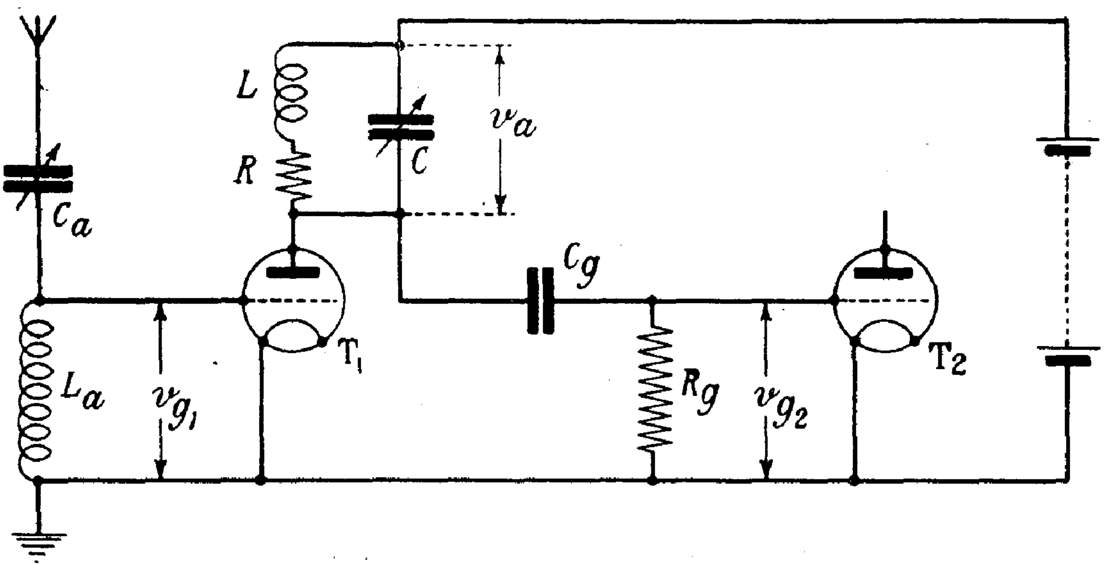


FIG. 26, CHAP. XI.—Tuned anode R.F. amplifier.

the impedance of the grid-filament path, and that its insulation resistance shall be extremely high. Both these conditions are easily fulfilled by a mica dielectric condenser having a capacitance of the order of ·0001 μF. The grid leak R_g may be from ·1 to 4 megohms depending in part upon the function of the valve T_2, which may be required to act as an additional R.F. amplifier or as a detector valve. The introduction of the grid leak will cause the stage gain to be somewhat less than the V.A.F. as calculated from the above expression, for its resistance must be considered to be in parallel with the dynamic resistance R_d. As however the latter is usually much less than the grid leak resistance, the effect is of minor importance. The input impedance of the following valve must also be taken into consideration, as will be shown later.

47. At frequencies other than the resonant frequency of the anode circuit, the V.A.F., and therefore the stage gain, is less than at the resonant frequency. Provided that the resistance

of the tuned anode circuit is small compared with the ratio $\frac{L}{C}$, the V.A.F. at any frequency $\frac{\omega}{2\pi}$ is given by the equation :—

$$\text{V.A.F.} = \frac{\frac{L}{Cr_a}}{\sqrt{\left(\frac{L}{Cr_a} + R\right)^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}}\,\mu.$$

(This expression is derived in paragraph 23).

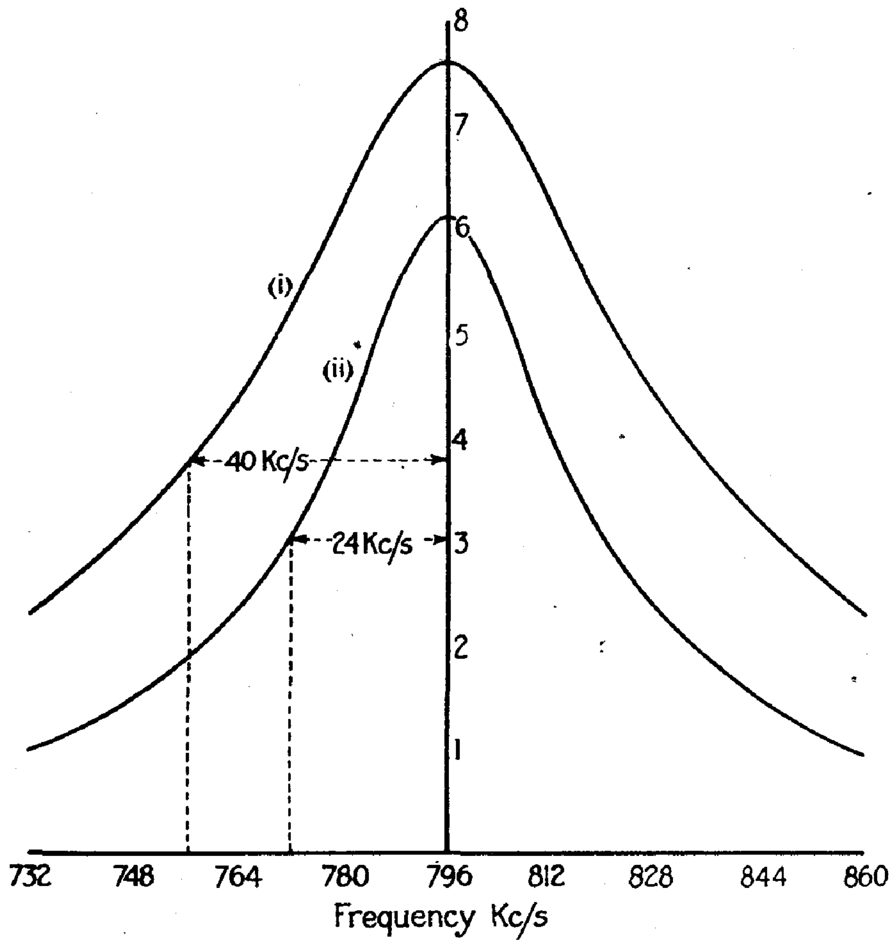


FIG. 27, CHAP. XI.—Response curves with different $\frac{L}{C}$ ratios.

At the resonant frequency of the circuit, $\omega L = \frac{1}{\omega C}$, and this expression simplifies to that already given. The tuned anode amplifier therefore possesses the property of selectivity, and the degree of discrimination is dependent chiefly upon the ratio $\frac{L}{C}$. A large value of $\frac{L}{C}$ gives high amplification at the resonant frequency, but the selectivity is poor, while a low ratio $\frac{L}{C}$ gives less amplification but greater selectivity. The stage gain of a certain tuned-anode capacitance-coupled amplifier in the region of the resonant frequency, 796 kc/s, is shown in fig. 27. The valve used has the following constants, viz. $r_a = 20{,}000$ ohms, $\mu = 10$. Curve (i) shows the V.A.F. when the inductance L is 160 μH, the capacitance $\cdot 00025$ μF, and the resistance 10 ohms, while curve (ii) shows the V.A.F. obtained by halving the inductance and doubling the capacitance the resistance being also reduced to 5 ohms. It will be observed that the magnification of the circuit is the same in each case, i.e. 80, but the selectivity of the second arrangement is greater than that of the first ; for example, with the circuit of large $\frac{L}{C}$ ratio the amplification of a signal 40 kc/s off resonance is one-half that of the desired signal, whereas with the circuit of lower $\frac{L}{C}$ ratio the same reduction is obtained if the undesired signal is only 24 kc/s off resonance. In Chapter V it is shown that the greatest selectivity is achieved when the dynamic resistance of the anode circuit is equal to the anode A.C. resistance of the valve, and the V.A.F. is then equal to $\frac{\mu}{2}$.

Tuned transformer coupling

48. This is an alternative kind of R.F. inter-valve coupling, which may take either of three forms, namely (*a*) with primary winding tuned to the frequency of the desired signal, (*b*) with the secondary winding tuned, or (*c*) with both windings tuned (fig. 28). The latter form is rarely,

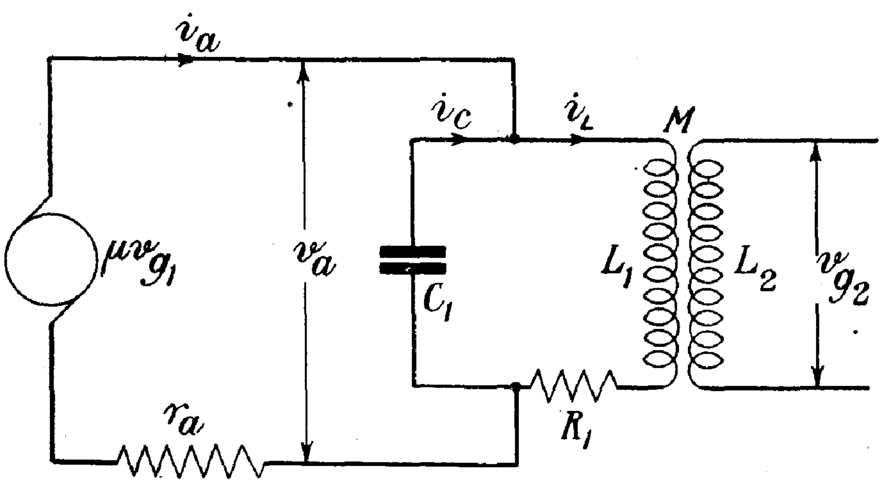
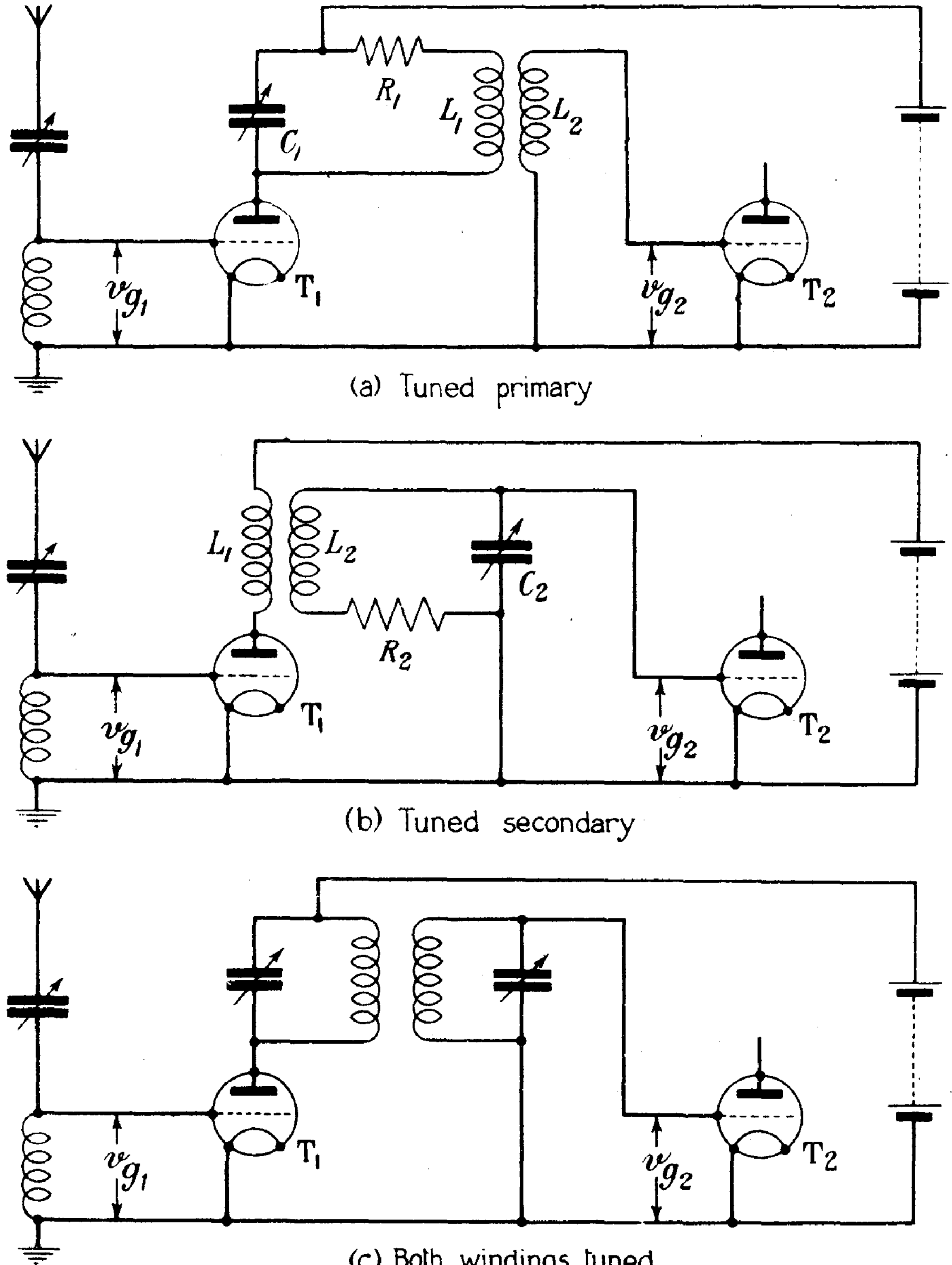


FIG. 29, CHAP. XI.—Equivalent circuit, "tuned primary" coupling.

if ever, adopted when it is necessary to vary the frequency to which the circuits are adjusted, but may be adopted in amplifying stages designed to operate at a single fixed frequency, such as the intermediate frequency stages of a supersonic heterodyne receiver (*see* paras. 81 *et seq.*). When an R.F. amplifying stage is required to be adjustable to any frequency within a fairly wide range, by means of a variable condenser or inductance, the choice lies between the first and second named arrangements. The stage gain and selectivity of the current shown in fig. 28b are superior to that of fig. 28a, and the former is almost universally employed. The relative selectivity and gain of the two arrangements may now receive a brief consideration.

"Tuned primary" circuit

49. The stage gain of this arrangement is easily derived, for reference to the equivalent circuit of the amplifier (fig. 29) shows that, provided the secondary winding is on open circuit,



TYPES OF R.F. TRANSFORMER COUPLING

FIG. 28
CHAP. XI.

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the voltage across the primary winding of the coupling transformer is, by analogy with the tuned anode circuit of fig. 26;

$$v_a = \frac{\dfrac{L_1}{C_1 r_a}}{\sqrt{\left(R_1 + \dfrac{L_1}{C_1 r_a}\right)^2 + \left(\omega L_1 - \dfrac{1}{\omega C_1}\right)^2}} \mu \, v_{g1} .$$

The primary current i_L will be very nearly $\frac{v_a}{\omega L_1}$ and the E.M.F. induced in the secondary winding is $\omega M i_L$ or $\frac{M}{L_1} v_a$. As the secondary is unloaded this is also the secondary terminal P.D., v_{g2}, hence

$$\frac{v_{g2}}{v_{g1}} = \frac{\dfrac{M}{C_1 r_a}}{\sqrt{\left(R_1 + \dfrac{L_1}{C_1 r_a}\right)^2 + \left(\omega L_1 - \dfrac{1}{\omega C_1}\right)^2}} \mu .$$

At the resonant frequency of the primary circuit $\omega L_1 = \frac{1}{\omega C_1}$ and

$$\frac{v_{g2}}{v_{g1}} = \frac{\dfrac{M}{C_1 r_a}}{R_1 + \dfrac{L_1}{C_1 r_a}} \mu .$$

If k is the coefficient of coupling between the windings of the transformer, $M = k\sqrt{L_1 L_2}$ and

$$\frac{v_{g2}}{v_{g1}} = \frac{\dfrac{k\sqrt{L_1 L_2}}{C_1 r_a}}{R_1 + \dfrac{L_1}{C_1 r_a}} \mu .$$

To simplify still further, multiply the numerator of the right-hand member by $\frac{L_1}{L_1} = 1$. It then becomes

$$\frac{k L_1 \sqrt{L_1 L_2}}{L_1 \, C_1 r_a} = k \sqrt{\frac{L_2}{L_1}} \times \frac{L_1}{C_1 r_a} .$$

Denoting $\sqrt{\frac{L_2}{L_1}}$ by T, and $\frac{L_1}{C_1 R_1}$, which is the dynamic resistance of the tuned primary circuit, by R_d, we have as a final expression, for the stage gain at the resonant frequency,

$$\frac{v_{g2}}{v_{g1}} = \frac{R_d}{r_a + R_d} k \, T \, \mu .$$

50. If the primary and secondary windings are of the same size and shape, for example, if they are wound side by side on a slotted former and the wire gauges so chosen that they occupy equal volumes of winding space, T is equal to the ratio $\frac{\text{Secondary turns}}{\text{Primary turns}}$. The product kT is the effective transformation ratio of the transformer and is less than T because the coefficient of coupling must be less than unity. It may appear possible to increase the stage gain without limit by increasing the ratio $\frac{L_2}{L_1}$, but this is not so, because such an increase must result in a corresponding reduction of k owing to the increased flux leakage between the windings. An

increase in the number of secondary turns will also lead to an increase in the distributed capacitance of the winding, and if the resonant frequency of the secondary winding approaches that of the primary, the stage gain is not given by the expression developed above.

"Tuned secondary" circuit

51. Turning now to the circuit given in fig. 28b and using the notation given in the equivalent circuit diagram (fig. 30), the stage gain is found to be

$$\frac{v_{g2}}{v_{g1}} = \frac{\left(\frac{k}{T}\right)^2 \frac{L_2}{C_2 r_a}}{\sqrt{\left[R_2 + \left(\frac{k}{T}\right)^2 \frac{L_2}{C_2 r_a}\right]^2 + \left(\omega L_2 - \frac{1}{\omega C_2}\right)^2}} \frac{T}{k}\mu$$

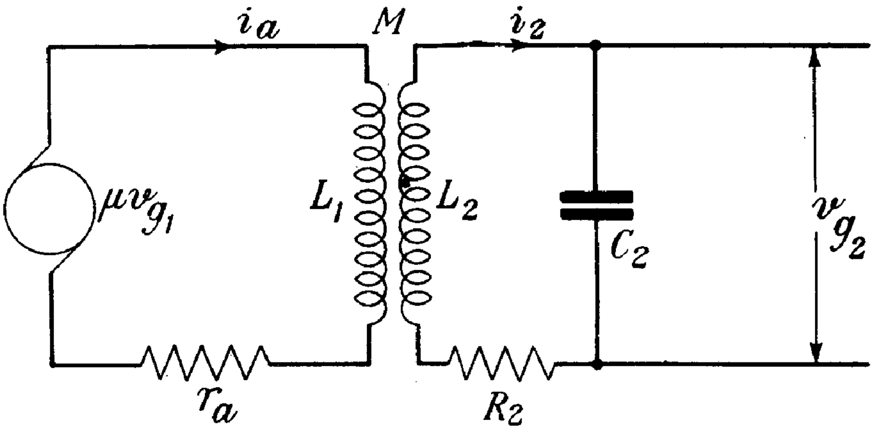


FIG. 30, CHAP. XI.—Equivalent circuit, "tuned secondary" coupling.

where k and T are as previously defined. At the resonant frequency $\omega L_2 - \frac{1}{\omega C_2} = 0$ and

$$\frac{v_{g2}}{v_{g1}} = \frac{\left(\frac{k}{T}\right)^2 R_d}{r_a + \left(\frac{k}{T}\right)^2 R_d} \frac{T}{k}\mu$$

where $R_d = \frac{L_2}{C_2 R_2}$. The effective transformation ratio is now $\frac{T}{k}$, whereas in the previous arrangement it was kT. This is because in the " tuned primary " circuit, the chief losses are due to the comparatively large circulating current in the primary. As the secondary circuit is assumed to be loss-free, an increase in the coupling coefficient k gives a larger secondary E.M.F. and terminal P.D. When the secondary winding is tuned, however, it becomes the seat of a circulating current and its losses are transferred to the primary circuit. The tighter the coupling, the larger is the equivalent primary resistance, and the smaller is the secondary E.M.F. and the P.D. at the condenser terminals. Maximum stage gain and selectivity are obtained when the transferred resistance $\left(\frac{k}{T}\right)^2 R_d$ is equal to r_a. For a given valve and secondary circuit, the optimum transformation ratio is found to be equal to $\sqrt{\frac{R_d}{r_a}}$ and the stage gain is then $\frac{\mu}{2}\sqrt{\frac{R_d}{r_a}}$.

Example.—In fig. 30 let $L_2 = 160\ \mu H$, C_2 be variable from $\cdot 00005\ \mu F$ to $\cdot 0005\ \mu F$, $r_a = 20{,}000$ ohms, $\mu = 10$. Assuming that R_2 is equal to 10 ohms over the whole tuning range and that a coupling coefficient of $\cdot 9$ can be achieved, find the primary inductance for optimum gain at 796 kc/s ($C_2 = \cdot 00025\ \mu F$), and the gain when $C_2 = \cdot 00005\ \mu F$, $\cdot 00025\ \mu F$ and $\cdot 0005\ \mu F$.

When $C_2 = \cdot 00025\ \mu F$, $R_d = \dfrac{L_2}{C_2 R_2} = 64{,}000$ ohms.

The optimum value of $\dfrac{T}{k}$ is $\sqrt{\dfrac{R_d}{r_a}} = \sqrt{\dfrac{64{,}000}{20{,}000}} = 1\cdot 79$

$$T = 1\cdot 79\, k = \sqrt{\frac{L_2}{L_1}}$$

$$\frac{L_2}{L_1} = 1\cdot 79^2\, k^2$$
$$= 3\cdot 2 \times \cdot 81$$
$$= 2\cdot 6$$

$$L_1 = \frac{L_2}{2\cdot 6}$$
$$= 61\cdot 5\ \mu H.$$

The gain at this frequency will be $\dfrac{\mu}{2}\sqrt{\dfrac{R_d}{r_a}}$ or

$$\frac{v_{g2}}{v_{g1}} = \frac{10}{2} \times 1\cdot 79$$
$$= 8\cdot 95.$$

When $C_2 = \cdot 00005\ \mu F$, $R_d = 320{,}000$ ohms.

$$\frac{v_{g2}}{v_{g1}} = \frac{\left(\frac{k}{T}\right)^2 R_d}{r_a + \left(\frac{k}{T}\right)^2 R_d}\ \frac{T}{k}\ \mu$$

$$\frac{T}{k} = 1\cdot 79$$

$$\frac{k}{T} = \cdot 558$$

$$\left(\frac{k}{T}\right)^2 = \cdot 312$$

$$\left(\frac{k}{T}\right)^2 R_d = \cdot 312 \times 320{,}000$$
$$= 100{,}000 \text{ ohms.}$$

$$\frac{v_{g2}}{v_{g1}} = \frac{100{,}000}{100{,}000 + 20{,}000} \times 1\cdot 79 \times 10$$
$$= 14\cdot 9,$$

and when $C_2 = \cdot 0005\ \mu F$, $R_d = 32{,}000$ ohms

$$\left(\frac{k}{T}\right)^2 R_d = \cdot 312 \times 32{,}000$$
$$= 100{,}000 \text{ ohms}$$

$$\frac{v_{g2}}{v_{g1}} = \frac{100{,}000}{100{,}000 + 20{,}000} \times 10$$
$$= 8\cdot 32.$$

52. For purposes of comparison, fig. 31a gives the respective response curves (stage gain against kilocycles per second off resonance) of (i) the transformer-coupled amplifier with tuned primary, and (ii) the same amplifier with tuned secondary. The resonant frequency is 796 kc/s, the valve has an r_a of 20,000 ohms and a μ of 10, while the constants of the tuned circuit are the same in each case, namely $L = 160\ \mu H$, $C = \cdot 00025\ \mu F$, $R = 10$ ohms. In fig. 31b, the relative gains are plotted in decibels with reference to an arbitrary level corresponding to a stage gain of 8. It is seen that at the resonant frequency the tuned secondary arrangement is the better by about 1·5 db. The tuned secondary attenuates a signal 34 kc/s off resonance by rather more than 12 db., equivalent to two units of signal strength on the operating scale,

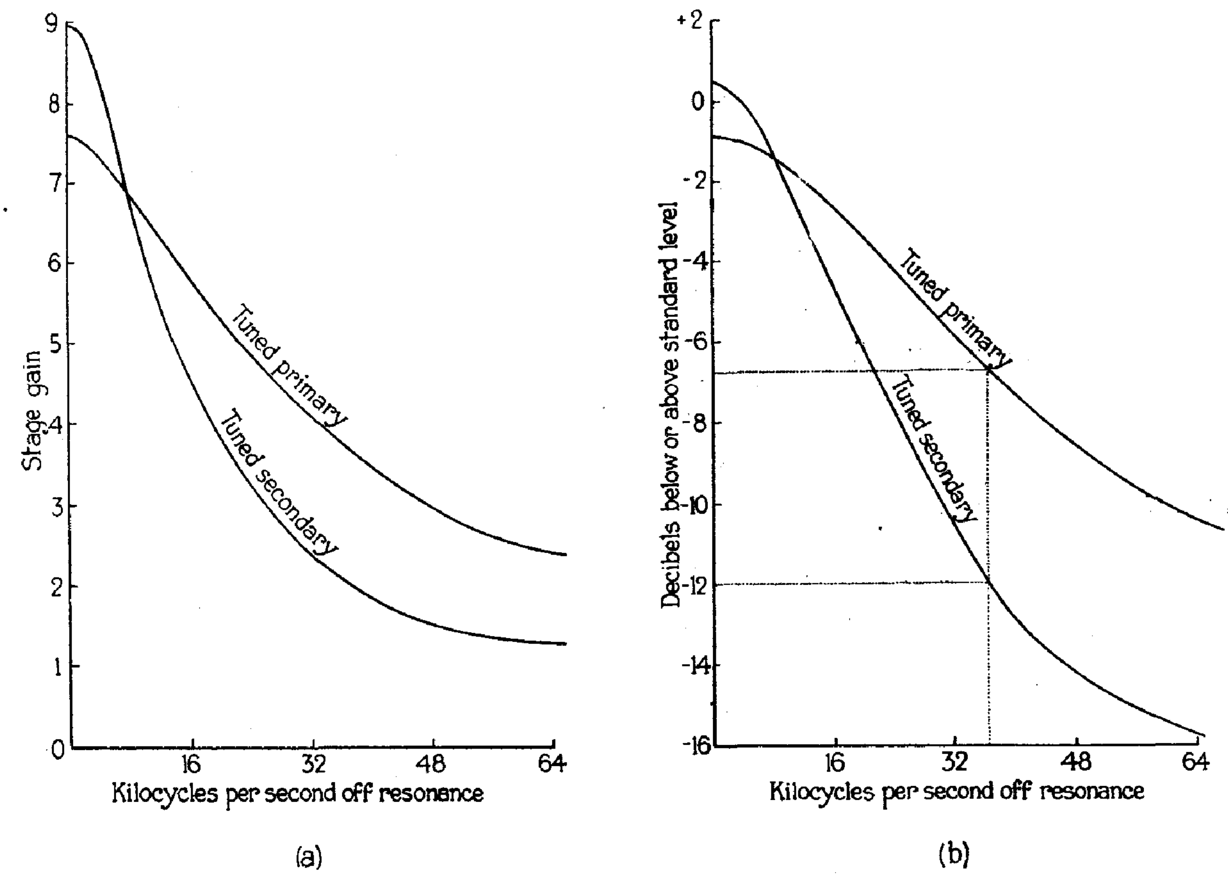


FIG. 31, CHAP. XI.—Comparison of selectivity of R.F. amplifying stages.

whereas the tuned primary circuit attenuates it by only 6 db. The relative selectivity of the two arrangements will be of the same order over the whole band covered by the tuning condenser, although the actual selectivity will be different, because the resistance of the tuned circuit increases with frequency.

INPUT ADMITTANCE

Components of input admittance

53. It has hitherto been assumed that if the grid and filament of a triode are connected to the two ends of an impedance, the effect is to add to the latter a small effective capacitance in parallel. In practice, the impedance added to the input circuit is of a more complex nature than this ; in discussing the various forms it may take it is convenient to refer to the input admittance of the triode, i.e. the reciprocal of its apparent impedance, measured between grid and filament.

This input admittance will be denoted by $Y_i = \frac{1}{Z_i}$ and in general consists of a susceptance $B_i = \omega C_i$ and a conductance $G_i = \frac{1}{R_i}$ in parallel, ω being 2π times the frequency of the applied E.M.F. as usual. The input conductance G_i consists of three portions, first the leakage conductance G_l due to imperfect insulation between grid and filament. The insulation resistance will rarely exceed 50 megohms and may be much less, e.g. if a grid leak resistance is fitted. The lower limit of the conductance is therefore of the order of 2×10^{-8} siemens (mho). Second, a conductance due to the flow of an electron convection current between the filament and grid of the valve, commonly referred to as grid current. This conductance will be denoted by g_g and the corresponding resistance by r_g. Third, a conductance G_m due to the finite admittance Y_{ag} of the grid-anode path of the valve. The input susceptance B_i consists of two portions, first, that due to the grid-filament capacitance C_{gf} and second, that due to the finite admittance of the grid-anode path ; this will be denoted by B_m.

The grid-anode admittance

54. (i) This consists of a conductance $G_{ag} = \frac{1}{R_{ag}}$ and a susceptance $B_{ag} = \omega C_{ag}$ in parallel, and serves in effect to couple the anode and grid circuits of the valve. It is therefore responsible for a transference of energy from the anode to the grid circuit or vice versa, according to the operating conditions. The phenomena which are directly due to this coupling are collectively referred to as the Miller effect. In ordinary triodes the grid-anode capacitance usually lies between 2 and 10 $\mu\mu F$, while the resistance R_{ag} rarely exceeds 50 megohms. For theoretical purposes it is often convenient to consider this resistance to be infinite, but the effect of its finite value is of importance at audio frequencies. It will be seen later that the input conductance due to the Miller effect may in certain circumstances be of negative sign. In Chapter X it was shown that if the anode circuit of a simple receiver is coupled to the tuned input circuit by mutual inductance, an effective resistance is transferred to the input circuit, and further that this transferred resistance may be either of positive sign, tending to increase the damping, or of negative sign, tending to reduce the damping, according to the sign of the mutual inductance. Since the grid-anode capacitance acts as a coupling between the input circuit and the anode load impedance its effect is also to transfer an effective resistance to the input circuit. With a resistive or capacitive load, the transferred resistance is positive, but if the load offers inductive reactance, the transferred resistance may be negative.

(ii) For ease of reference the notation explained above is collected in the following table.

Z_i = total input impedance $= \frac{1}{Y_i}$.

Y_i = total input admittance $= \sqrt{G_i^2 + B_i^2}$.

B_i = total input susceptance $= B_{gf} + B_m$.

G_i = total input conductance $= G_l + G_m + g_g$.

B_{gf} = susceptance of grid-filament path $= \omega C_{gf}$.

G_l = leakage conductance of grid-filament path.

g_g = internal grid-filament conductance of valve $= \frac{1}{r_g}$.

G_m = effective input conductance due to Miller effect.

B_m = effective input susceptance due to Miller effect.

Y_{ag} = admittance of grid-anode path $= \sqrt{G_{ag}^2 + B_{ag}^2}$.

G_{ag} = conductance of grid-anode path $= \frac{1}{R_{ag}}$.

B_{ag} = susceptance of grid-anode path $= \omega C_{ag}$.

In subsequent paragraphs, the following notation will also occur.

G_d = dynamic conductance of input circuit alone.
G'_d = total dynamic conductance of input circuit = $G_d + G_i$.
χ = magnification of input circuit alone.
χ' = effective magnification of input circuit.
L_o = effective inductance of anode circuit load.
R_{do} = dynamic resistance of tuned anode circuit.

Positive and negative reaction effects

55. (i) The effect of the input conductance of the valve upon the grid-filament circuit is most easily shown by numerical examples. Suppose the circuit to consist of an inductance of 160 μH, a capacitance of ·00025 μF and a resistance of 10 ohms, so that its dynamic resistance R_d is 64,000 ohms and its dynamic conductance G_d is $1{\cdot}5625 \times 10^{-5}$ siemens. If the valve has a positive input conductance $G_i = 10^{-5}$ siemens, the total conductance between grid and filament is

$$\begin{aligned} G'_d &= G_d + G_i \\ &= (1{\cdot}5625 + 1) \times 10^{-5} \\ &= 2{\cdot}5625 \times 10^{-5} \text{ siemens.} \end{aligned}$$

The effective dynamic resistance of the input circuit is therefore reduced from R_d to R'_d, where

$$\begin{aligned} R'_d &= \frac{10^5}{2{\cdot}5625} \text{ ohms} \\ &= 39{,}000 \text{ ohms} \end{aligned}$$

and the effective magnification is correspondingly less. Since $R_d = \dfrac{\omega^2 L^2}{R}$ and the magnification is $\chi = \dfrac{\omega L}{R}$, $\chi = \dfrac{R_d}{\omega L}$. The magnification of the tuned circuit alone is therefore $\dfrac{64{,}000}{800} = 80$, but when the input conductance of the valve is connected in parallel, the magnification is reduced to

$$\begin{aligned} \chi' &= \frac{R'_d}{\omega L} \\ &= \frac{39{,}000}{80} \\ &= 48{\cdot}8 \text{ (approx.).} \end{aligned}$$

For a given induced E.M.F. the effect of the input conductance in this particular instance, is to reduce the grid-filament P.D. in the ratio of 80 to 48·8.

(ii) Now suppose the input conductance to be of the same magnitude, namely 10^{-5} siemens, but of negative sign. The effective conductance between grid and filament then becomes

$$\begin{aligned} G'_d &= G_d + G_i \\ &= (1{\cdot}5625 - 1)\ 10^{-5} \\ &= 5{\cdot}625 \times 10^{-6} \text{ siemens,} \end{aligned}$$

and the effective dynamic resistance is increased to 177,600 ohms. The circuit magnification becomes $\chi' = 222$, so that for a given induced E.M.F., the grid-filament P.D. is nearly 2·8 times that obtained when the input conductance is zero. This increase of circuit magnification, giving rise to an increased signal strength for the same induced E.M.F., is thus responsible for the phenomenon known as regenerative amplification.

Effect of grid current

56. (i) In fig. 32a, e_g is the instantaneous E.M.F. of a source of alternating voltage of internal impedance Z, and T a triode valve, the grid of which is maintained at such a mean negative potential E_b with respect to the filament, that no grid current flows during any portion of the cycle. These operating conditions are shown in fig. 32b. In these circumstances no current will be supplied by the source, and the alternating voltage V_g between grid and filament will be equal to the E.M.F. of the source, i.e. $v_g = e_g$; the triode may therefore be said to impose no load upon the generator.

(ii) Now suppose the bias voltage to be removed and grid current to flow during the whole of the cycle, the operating conditions being those of fig. 32c. The electron stream between filament and grid will now be subjected to an alternating voltage and a pulsating movement will be superimposed upon the mean steady progression of the electrons. Work must be done by

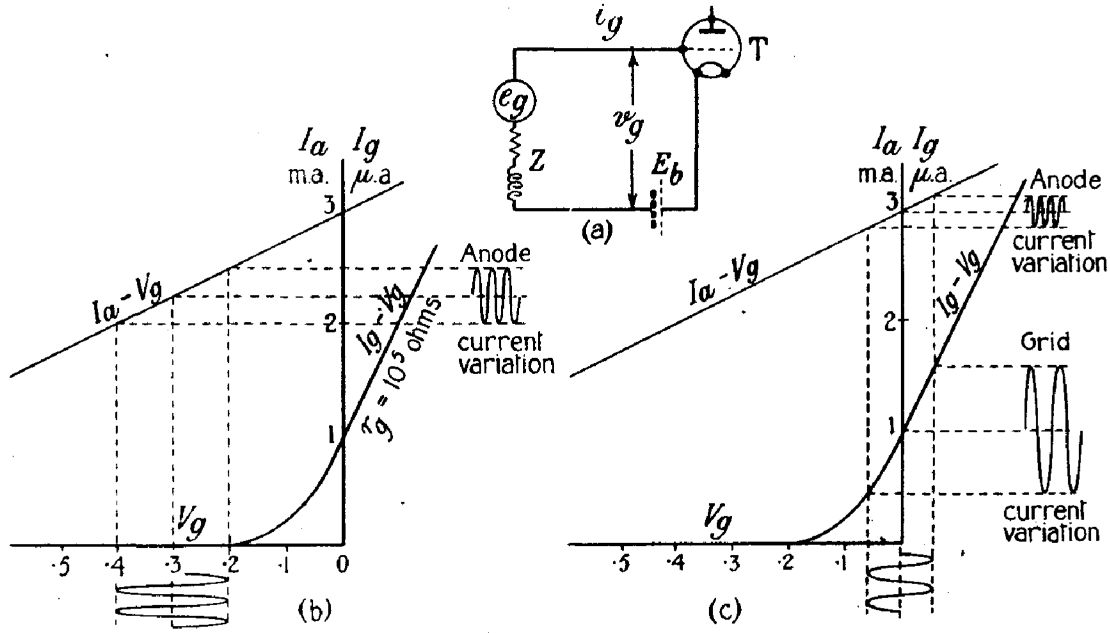


FIG. 32, CHAP. XI.—Effect of grid current.

the source in order to cause this pulsation, and its power output will be equal to the rate at which work is performed. The pulsation of the electrons within the valve must result in a similar motion of the electrons in the external circuit, so that an alternating current i_g (I_g, R.M.S.) may be considered to exist. There is therefore an internal voltage drop, i_gZ, in the generator, and the grid-filament P.D. will be less than the E.M.F., i.e. $v_g = e_g - i_gZ$. Electrical energy is converted into heat owing to the motion of the electrons, and the power expended in the valve is equal to $I_g^2 r_g$ watts, where $r_g = \frac{1}{g_g}$ and g_g is the slope of the $I_g - V_g$ curve. In fig. 32 r_g is approximately equal to 10^5 ohms, hence if the other components of the input conductance are zero, the effect of this grid current upon a tuned input circuit in which $L = 160\ \mu H$, $C = \cdot 00025\ \mu F$, $R = 10$ ohms, is precisely as calculated in paragraph 55.

(iii) If the operating conditions are intermediate between those of fig. 32b and fig. 32c so that grid current flows during only a portion of the cycle, the input conductance will not be constant, and the wave-form of the grid-filament P.D. will not be identical with that of the E.M.F.

of the source. This must occur to some extent in all cases where grid current is allowed to flow, since the $I_g - V_g$ characteristic always possesses considerable curvature. In general then, it may be said that in addition to the reduction in grid-filament voltage, amplitude distortion must arise if grid current is permitted.

Effect of grid-anode admittance

57. (i) In dealing with the portion $Y_m = \sqrt{G_m^2 + B_m^2}$, of the input admittance which is due to the finite impedance of the grid-anode path, it is convenient to assume that the operating conditions are those of fig. 32a, i.e. that no grid current flows. The circuit diagram of the amplifier is then as given in fig. 33a in which the grid-filament voltage is $v_g = \mathcal{V}_g \sin \omega t$, its

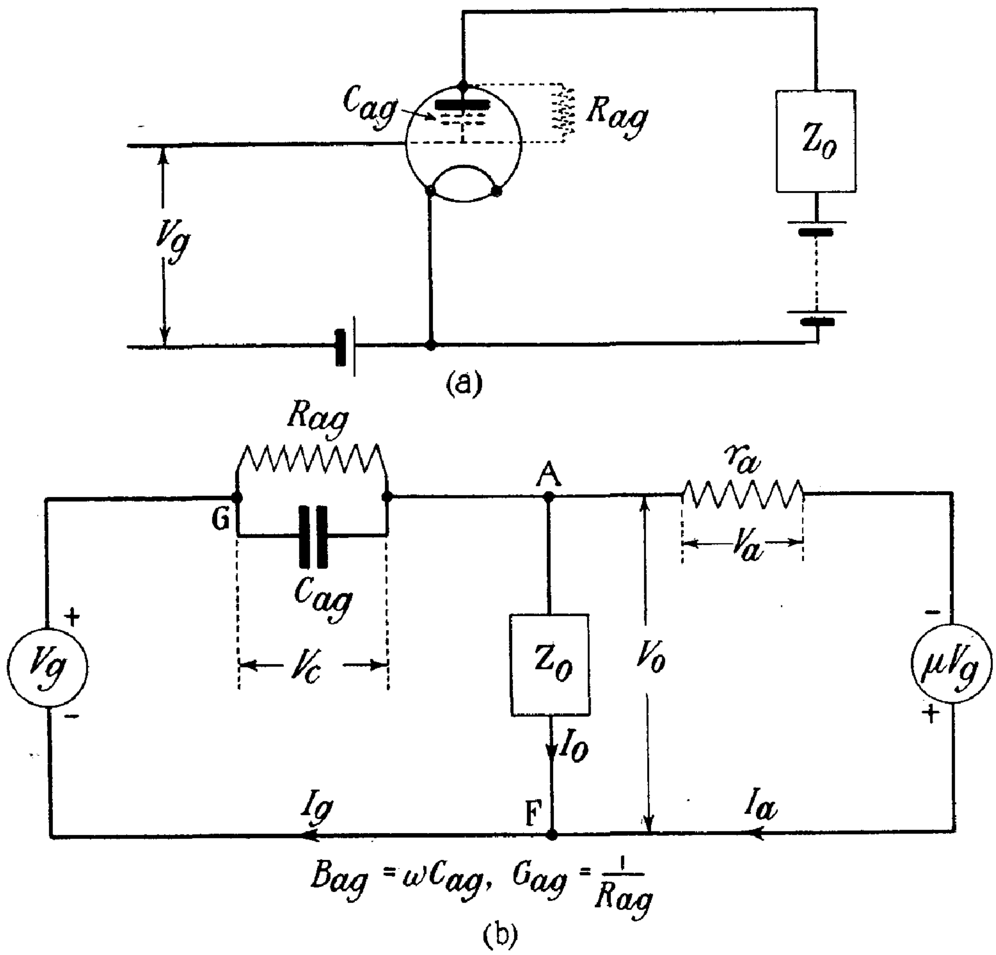


FIG. 33, CHAP. XI.—Anode-grid admittance of amplifier and equivalent circuit.

R.M.S. value being V_g. The load impedance is denoted by Z_o, the grid-anode capacitance by C_{ag}, and the grid-anode conductance by $G_{ag} = \frac{1}{R_{ag}}$. The electrical equivalent of the circuit is given in fig. 33b in which the input voltage is V_g and an amplified voltage μV_g is assumed to act in series with the anode A.C. resistance r_a. It is now obvious that owing to the admittance Y_{ag} the input voltage is able to supply current directly to the load impedance, while the equivalent generator μV_g is able to supply current to the input circuit.

(ii) In certain numerical examples, the triode will be assumed to possess the following constants, viz. :—

$$\mu = 10$$
$$r_a = 2 \times 10^4 \text{ ohms}$$
$$C_{ag} = 5 \times 10^{-12} \text{ Farad.}$$
$$R_{ag} = 5 \times 10^7 \text{ ohms } (G_{ag} = 2 \times 10^{-8} \text{ siemens}).$$

This will be referred to as the typical triode.

(iii) When B_{ag} is very much larger than G_{ag}, i.e. at frequencies above about 8,000 cycles per second, G_{ag} may be ignored ; at frequencies in the region of 600 cycles per second, B_{ag} is approximately equal to G_{ag} and both must be taken into account. At frequencies below about 400 cycles per second, the grid-anode admittance may be approximately represented by its conductance only. The current flowing through the generator V_g and admittance Y_{ag} will be referred to as the grid current, but must not be confused with the true grid-filament current previously discussed. Its instantaneous value is i_g and its R.M.S. value I_g. The corresponding anode current is i_a and the current through the load impedance i_o, their respective R.M.S. values being I_a and I_o. The conventional signs (+ and —) on the respective generators are inserted merely to show their relative phase.

58. From fig. 33 certain deductions may be made by inspection. If G_{ag} is zero, and the load impedance is a loss-free acceptor circuit for the applied frequency, $I_a = \frac{\mu V_g}{r_a}$ and $I_g = \omega C_{ag} V_g$, i.e. the input admittance Y_m is equal to B_{ag}. If the anode load is of this nature, then, the input admittance is purely capacitive and imposes no damping upon the input circuit. On the other hand, if the load impedance is a loss-free rejector circuit, the load current I_o is zero and $I_g = I_a = \dfrac{(\mu + 1) V_g}{\sqrt{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2}}$.

Under these conditions

$$Y_m = \frac{\mu + 1}{\sqrt{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2}} = \frac{(\mu + 1) B_{ag}}{\sqrt{g_a^2 + B_{ag}^2}} g_a.$$

Y_m is easily resolved into its conductive and susceptive components,

$$G_m = \frac{(\mu + 1) r_a}{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2} = \frac{(\mu + 1) B_{ag}^2}{g_a^2 + B_{ag}^2} g_a$$

and

$$B_m = \frac{(\mu + 1) \frac{1}{\omega C_{ag}}}{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2} = \frac{(\mu + 1) g_a^2}{g_a^2 + B_{ag}^2} B_{ag}.$$

Example.—Under the above conditions, find the input conductance and susceptance of the typical triode at 796 kc/s ($\omega = 5 \times 10^6$).

$$B_{ag} = \omega C_{ag} = 5 \times 10^6 \times 5 \times 10^{-12}$$
$$= 25 \times 10^{-6}$$
$$\frac{1}{\omega C_{ag}} = 4 \times 10^4$$
$$G_m = \frac{(\mu + 1)\, r_a}{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2}$$
$$= \frac{11 \times 2 \times 10^4}{(2 \times 10^4)^2 + (4 \times 10^4)^2}$$
$$= 1{\cdot}1 \times 10^{-4} \text{ siemens}$$
$$B_m = \frac{(\mu + 1)\, \frac{1}{\omega C_{ag}}}{r_a^2 + \left(\frac{1}{\omega C_{ag}}\right)^2}$$
$$= \frac{11 \times 4 \times 10^4}{20 \times 10^8}$$
$$= 2{\cdot}2 \times 10^{-4} \text{ siemens.}$$

Finite, purely resistive load

59. If, as before, we assume the grid-anode conductance to be zero, and the anode load is a pure resistance, r_o, the values of G_m and B_m are given by the following expressions.

$$G_m = \frac{A + 1}{Z^2}\, R = \frac{(A + 1)\, B_{ag}^2}{G^2 + B_{ag}^2}\, G$$
$$B_m = \frac{A + 1}{Z^2}\, \frac{1}{B_{ag}} = \frac{(A + 1)\, G^2}{G^2 + B_{ag}^2}\, B_{ag}$$
$$\text{where } A = \frac{r_o}{r_a + r_o}\, \mu$$
$$R = \frac{r_o\, r_a}{r_a + r_o} = \frac{1}{G}$$
$$Z^2 = R^2 + \left(\frac{1}{\omega C_{ag}}\right)^2$$

It will be observed that the above equations are of the same form as for an infinite load, but are modified by the substitution of the V.A.F. (A) for the amplification factor (μ), and the conductance (G) of the valve and load in parallel for that of the valve alone.

60. The effects of a resistive load are shown by a vector diagram in fig. 34. The loads on the two sources of supply are shown separately in fig. 34a. Looking into the circuit from the generator V_g the load impedance is that of the capacitance C_{ag} in series with the parallel combination of r_a and r_o, i.e. R. The current through C_{ag} due to this voltage is

$$I_1 = \frac{V_g}{\sqrt{R^2 + \left(\frac{1}{\omega C_{ag}}\right)^2}}$$

In the vector diagram (fig. 34b), I_1 is leading on V_g by nearly 90°, i.e. B_{ag} is assumed to be very much smaller than G. From the point of view of the generator μV_g, the load is composed of the resistance r_a in series with the parallel combination of r_o and C_{ag}. The current due to the voltage μV_g is denoted by I_2, its components being I_c flowing through C_{ag} and I_o flowing through the load resistance r_o ; I_2 will lead on μV_g by an angle much less than 90°, because I_c will lead on μV_g by 90° and I_o will be in phase with μV_g. We are chiefly concerned with the component I_c, which combines with I_1 to give the total current I_g through the capacitance C_{ag}. Since I_g is the vector sum of I_c and I_1, it leads on V_g by an angle θ which is less than 90°. It must have a component $I_g \cos \theta$ in phase with V_g and the power dissipated in the input circuit will be $I_g V_g \cos \theta$. This power is always small because in all practical amplifiers B_{ag} is small compared with G so that the angle θ is very nearly 90°. The effect of the grid-anode susceptance is therefore merely to increase the input capacitance by an amount which is less than $(A + 1)\ C_{ag}$. At very low audio frequencies, the admittance G_{ag} of the grid-anode path may be of considerable

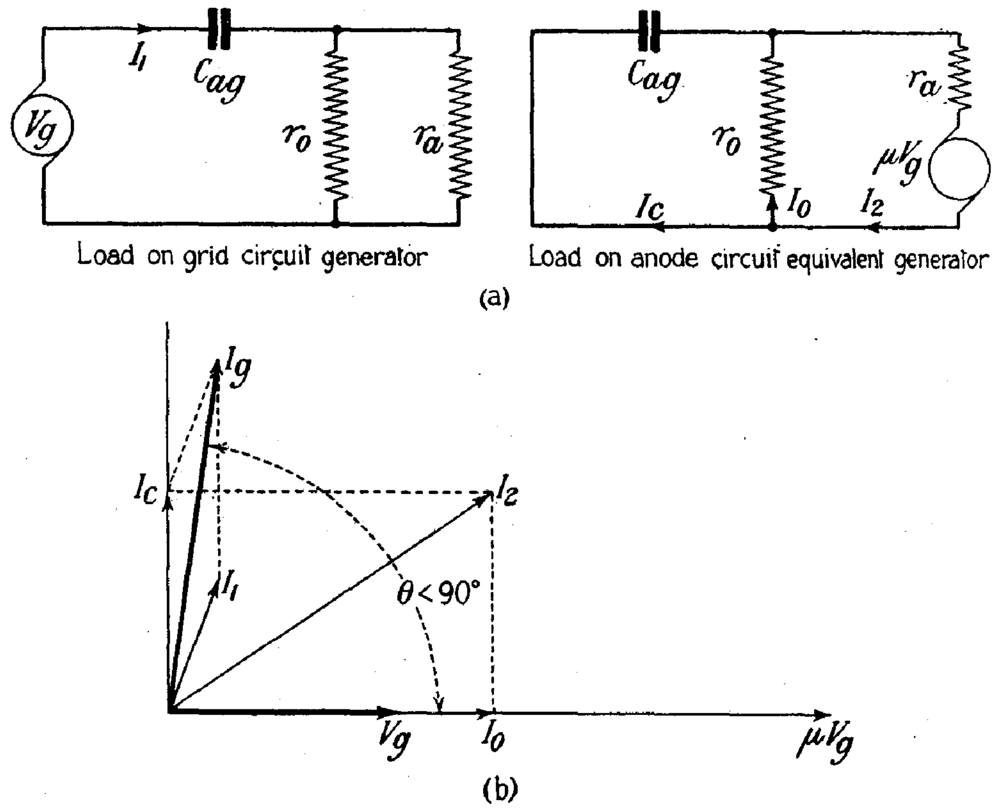


Fig. 34, Chap. XI.—Effect of finite, resistive load.

importance. When B_{ag} is very small compared with G_{ag} the input susceptance B_m is negligible, while the input conductance G_m approaches the value $(A + 1)\ G_{ag}$. Thus at 157 cycles per second, the grid-anode susceptance of the typical triode is $B_{ag} = 2\pi \times 157 \times 5 \times 10^{-12} = 5 \times 10^{-9}$ and its conductance G_{ag} is 2×10^{-8}, i.e. $G_{ag} = 4\ B_{ag}$. Let $r_o = r_a$, so that $A = \frac{\mu}{2} = 5$, $A + 1 = 6$. Then $G_m = 6 \times 2 \times 10^{-8} = 1{\cdot}2 \times 10^{-7}$ siemens. Thus the input conductance is equivalent in effect to a leak of about 8 megohms. This may not appear to be serious, but the effect may be appreciable if G_{ag} is allowed to assume a high value owing to an accumulation of dirt or moisture between grid and anode connections.

Capacitive load

61. If the load impedance offers capacitive reactance, e.g. if it consists of a resistance and a condenser in parallel, the input admittance becomes that of a positive conductance and a capacitive susceptance in parallel. The vector diagram for a purely capacitive load is given in fig. 35. Taking μV_g, the total voltage acting in the anode circuit, as the datum vector, the anode current I_a leads on μV_g by an angle less than 90°. The anode-filament P.D. V_a will be equal in magnitude to $I_a r_a$ and will be in phase with I_a. The P.D. V_o across the load will be equal to $\mu V_g - V_a$, vectorially, and the load current I_o will lead on the voltage V_o by 90°. The grid current I_g is equal to $I_o - I_a$ vectorially, and is therefore less than 90° ahead of V_o, i.e. very much less than 90° ahead of V_g. As a result, the input admittance has a large

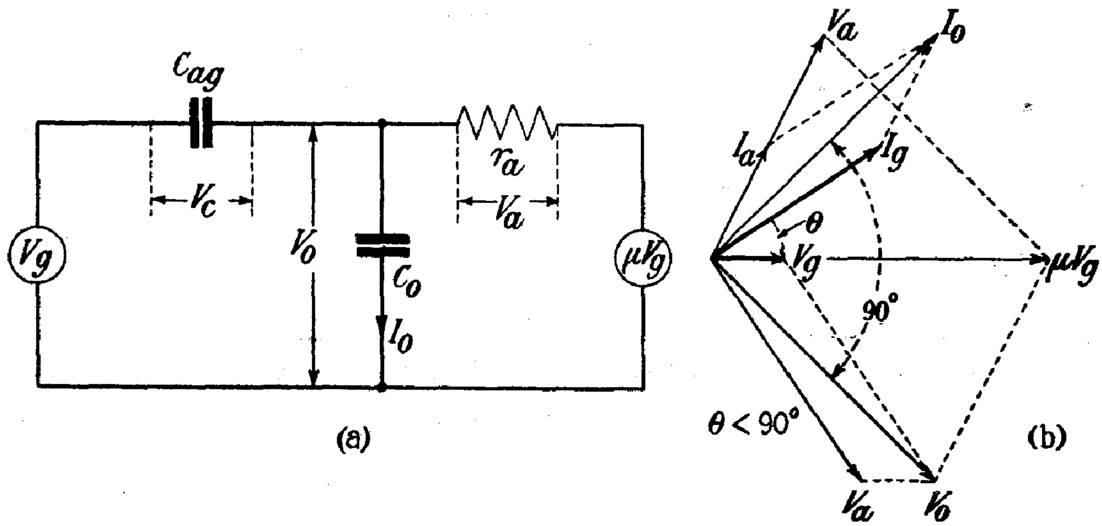


FIG. 35, CHAP. XI.—Effect of finite, capacitive load.

conductance component, and the valve may, in certain circumstances, impose a very heavy load upon the input circuit. The maximum value of the input conductance (assuming G_{ag} to be zero) is

$$G_m \text{ (max.)} = \frac{\mu}{2} B_{ag}$$

and is obtained with a load reactance equal in magnitude to r_a.

Inductive load

62. When the anode load offers inductive reactance, the input conductance may become negative in sign, and the manner in which this comes about may again be explained by means of a vector diagram. Referring to fig. 36 and taking the vector μV_g as datum, the anode current I_a will lag on this by some angle less than 90°, and the anode-filament P.D., $V_a = I_a r_a$, will be in phase with I_a. The P.D. V_o across the load will be equal to $\mu V_g - V_a$, vectorially, while the load current I_o lags on V_o by nearly 90°. The grid current I_g is equal to $I_a - I_o$, vectorially, while the grid-anode P.D. V_c is the vector sum of V_g and V_o. Provided that G_{ag} is negligible, therefore, I_g leads on the voltage V_c by 90°. Since V_c must lead on V_g, I_g must lead on V_g by an angle θ greater than 90°. As the power input to the grid circuit is $V_g I_g \cos \theta$, and $\cos \theta$ is negative if θ lies between 90° and 270°, it follows that under these conditions the "generator" V_g is receiving power from the circuit instead of supplying power. The maximum negative value of the input conductance (assuming G_{ag} to be zero) is

$$G_m \text{ (max.)} = -\frac{\mu}{2} B_{ag}$$

and is obtained with a load reactance equal in magnitude to r_a.

63. Whereas with a capacitive load the input conductance is invariably positive, with an inductive load it may be positive or negative. There is in fact, a kind of resonance effect between the effective load inductance L_o and the grid-anode capacitance C_{ag}; at frequencies for which G_{ag} is negligible compared with B_{ag},

$$G_m = \frac{\dfrac{\omega^2 L_o^2}{r_a}(\mu + 1) - \dfrac{L_o\,\mu}{C_{ag}\,r_a}}{\left(\dfrac{L_o}{C_{ag}\,r_a}\right)^2 + \left(\omega L_o - \dfrac{1}{\omega C_{ag}}\right)^2}\,. \qquad (1)$$

The resonance effect occurs when

$$\omega^2 = \frac{1}{L_o\,C_{ag}\left(1 + \dfrac{1}{\mu}\right)}\,, \qquad (2)$$

and at this frequency $G_m = 0$. At lower frequencies G_m is negative and at higher frequencies is positive. The magnitude of the input conductance at audio frequencies may be illustrated numerically as follows.

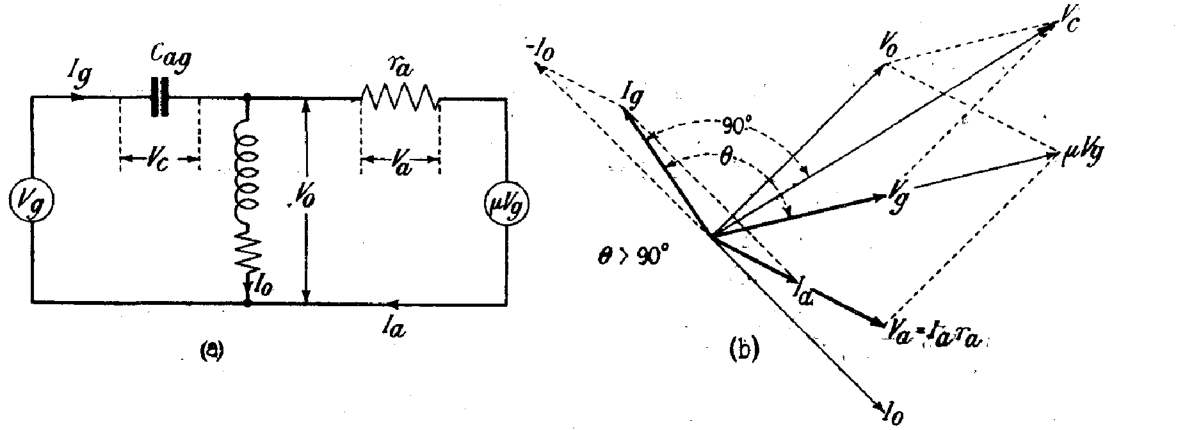


FIG. 36, CHAP. XI.—Effect of finite inductive load.

Example.—The anode load of a typical triode is an inductance of 50 henries. Assuming G_{ag} to be negligible, find the input conductance when $\omega = 5 \times 10^3$, $\omega = 5 \times 10^4$ and $\omega = 7 \times 10^4$.

(i)

$$\begin{aligned}
\omega &= 5 \times 10^3 \\
\omega L_o &= 5 \times 10^3 \times 50 = 2{\cdot}5 \times 10^5 \\
\frac{(\omega L_o)^2}{r_a}(\mu + 1) &= 6{\cdot}25 \times 10^{10} \times 11 \times 5 \times 10^{-5} \\
&\doteq 3{\cdot}44 \times 10^7 \\
\frac{L_o}{C_{ag}\,r_a} &= 50 \times 2 \times 10^{11} \times 5 \times 10^{-5} \\
&= 5 \times 10^8 \\
\frac{\mu L_o}{C_{ag}\,r_a} &= 5 \times 10^9 \\
\frac{1}{\omega C_{ag}} &= 2 \times 10^{-4} \times 2 \times 10^{11} \\
&= 4 \times 10^7 \\
\omega L_o - \frac{1}{\omega C_{ag}} &= 2{\cdot}5 \times 10^5 - 4 \times 10^7 \\
&\doteq -4 \times 10^7.
\end{aligned}$$

It is apparent that the positive term in the numerator is negligible compared with the negative term, and that the denominator is practically equal to $\left(\frac{L_o}{C_{ag} r_a}\right)^2$. Hence

$$G_m = -\frac{\frac{\mu L_o}{C_{ag} r_a}}{\left(\frac{L_o}{C_{ag} r_a}\right)^2}$$

$$= -\mu \frac{C_{ag} r_a}{L_o}$$

$$= -10 \times \frac{1}{5 \times 10^8}$$

$$= -2 \times 10^{-8} \text{ siemens.}$$

(ii)

$$\omega = 5 \times 10^4$$

$$\omega L_o = 2{\cdot}5 \times 10^6$$

$$\frac{(\omega L_o)^2}{r_a}(\mu + 1) = 3{\cdot}44 \times 10^9$$

$$\frac{L_o}{C_{ag} r_a} = 5 \times 10^8, \quad \frac{\mu L_o}{C_{ag} r_a} = 5 \times 10^9 \text{ as before.}$$

$$\frac{1}{\omega C_{ag}} = 4 \times 10^6$$

$\left(\omega L_o - \frac{1}{\omega C_{ag}}\right)^2$ is again negligibly small compared with $\left(\frac{L_o}{C_{ag} r_a}\right)^2$

$$G_m = \frac{(3{\cdot}44 - 5)\,10^9}{25 \times 10^6}$$

$$= -{\cdot}624 \times 10^{-8} \text{ siemens.}$$

(iii)

$$\omega = 7 \times 10^4$$

$$\omega L_o = 3{\cdot}5 \times 10^6$$

$$\frac{(\omega L_o)^2}{r_a}(\mu + 1) = 6{\cdot}7 \times 10^9$$

$\frac{L_o}{C_{ag} r_a}$ and $\frac{\mu L_o}{C_{ag} r_a}$ remain as before, and the denominator is to all intents and purposes equal to 25×10^{16}

$$G_m = \frac{(6{\cdot}7 - 5)\,10^9}{25 \times 10^{16}}$$

$$= +{\cdot}684 \times 10^{-8} \text{ siemens.}$$

64. At audio frequencies, no matter what the nature of the load may be, the input admittance consists of a capacitive susceptance, approximately equal to $(A + 1)\,B_{ag}$, and a conductance, nearly always positive. Its magnitude is governed largely by the conductance G_{ag} of the grid-anode path, which has hitherto generally been assumed to be negligible. For practical purposes, it may be said that the input conductance G_m is positive if G_{ag} exceeds $\frac{B_{ag}}{2}$,

which is nearly always true at audio frequencies. Under these conditions the magnitude of G_m is approximately equal to $\left(1 + \frac{A^2}{\mu}\right) G_{ag}$. At radio frequencies, the effect of the anode-grid conductance is negligible, and the input conductance is very nearly that given by equation 1, paragraph 63. Its order of magnitude will now be illustrated with reference to the typical triode.

Example.—The anode load of the typical triode is an inductance of 200 μH. Find the input conductance when $\omega = 5 \times 10^6$.

$$\omega L_o = 5 \times 10^6 \times 200 \times 10^{-6} = 10^3$$

$$\frac{(\omega L_o)^2}{r_a}(\mu + 1) = 10^6 \times 11 \times 5 \times 10^{-5}$$

$$= 5\cdot5 \times 10^2$$

$$\frac{L_o}{C_{ag}\, r_a} = 200 \times 10^{-6} \times 2 \times 10^{11} \times 5 \times 10^{-5}$$

$$= 2 \times 10^3$$

$$\frac{\mu L_o}{C_{ag}\, r_a} = 2 \times 10^4$$

Obviously $\frac{\mu L_o}{C_{ag}\, r_a}$ is very much larger than $\frac{(\omega L_o)^2}{r_a}(\mu + 1)$.

$$\frac{1}{\omega C_{ag}} = \frac{1}{5 \times 10^6 \times 5 \times 10^{-12}} = 4 \times 10^4$$

$$\left(\omega L_o - \frac{1}{\omega C_{ag}}\right)^2 \doteqdot \left(\frac{1}{\omega C_{ag}}\right)^2 = 8 \times 10^8$$

and is large compared with $\frac{L_o}{C_{ag}\, r_a}$.

Hence the input conductance is approximately

$$G_m = -\frac{2 \times 10^4}{8 \times 10^8}$$

$$= -\ 2\cdot5 \times 10^{-5} \text{ siemens.}$$

The corresponding input shunt resistance is $-\ 4 \times 10^4$ ohms.

Self-oscillation due to negative input conductance

65. The effect of this negative input conductance upon the input circuit depends upon the dynamic conductance of the circuit itself. If the latter is that considered in paragraph 55, having a dynamic conductance of $1\cdot5625 \times 10^{-5}$ siemens, the total conductance will be

$$G'_d = G_d + G_i = (1\cdot5625 - 2\cdot5) \times 10^{-5}$$

$$= -\ \cdot9375 \times 10^{-5} \text{ siemens.}$$

The total conductance between grid and filament is therefore negative. Although this condition may exist momentarily, it cannot persist. Suppose it to be true at the instant of switching on the H.T. supply to the valve, bearing in mind that the mere acceleration of electrons consequent upon this action must give rise to an oscillation at some frequency or other. The valve will then supply power to the input circuit and, since the effective resistance of the latter is momentarily negative, it will set up a large circulating current and a corresponding grid-filament P.D. This large grid swing will give rise to grid current and the input conductance will become less negative, while since the D.C. resistance of the input circuit cannot be zero, some mean negative grid bias will be developed. The operating point of the dynamic $I_a - V_a$ curve will therefore move to a point of greater curvature, i.e. where the anode A.C. resistance is greater. Thus the valve itself adjusts the various operating parameters in such a manner that, although an oscillatory current is maintained in the input circuit, the total grid-filament conductance becomes zero, while the power dissipated in the valve and its associated circuits is exactly equal to that supplied by the H.T. source ; the grid voltage swing is then maintained at the magnitude which it possessed at the instant of stabilization. This, briefly, is the mechanism by which an oscillation is maintained as a result of the capacitive coupling between grid and anode.

Input conductance with tuned anode

66. Even with an inductive load, however, the input conductance is not necessarily negative. The input conductance is positive if

$$\omega^2 > \frac{1}{L_o C_{ag}\left(1 + \frac{1}{\mu}\right)}$$

as already explained. It is of interest to find the magnitude of the load inductance which will make the input conductance positive ; e.g. at 796 kc/s the input conductance is positive if

$$L_o > \frac{1}{\omega^2 C_{ag}\left(1 + \frac{1}{\mu}\right)}$$

$$> \frac{10^{12}}{25 \times 10^{12} \times 5 \times 1{\cdot}1},$$

i.e. if $L_o > 7{,}260\ \mu H.$

All practical radio-frequency amplifiers may be reduced to an equivalent tuned anode circuit by the methods of paras. 49 *et seq.* and at the resonant frequency of the tuned circuit the load is purely resistive. At frequencies very close to resonance, however, the load becomes highly reactive—inductive if the circuit is tuned to a frequency higher than that applied and vice versa. Thus, if the anode circuit has an inductance L of 200 μH in parallel with a capacitance C of $\cdot 0002\ \mu F$, and negligible resistance, its reactance at its resonant frequency is infinite ; at a slightly lower frequency, e.g. $\omega = 4{\cdot}9 \times 10^6$, its reactance is $X_o = \frac{\omega L}{1 - \omega^2 LC}$ ohms, and is inductive if the sign of this quantity is positive. The reactance at $\omega = 4{\cdot}9 \times 10^6$ is

$$X_o = \frac{4{\cdot}9 \times 200 \times 10^6 \times 10^{-6}}{1 - 4{\cdot}9^2 \times 10^{12} \times 200 \times 10^{-6} \times {\cdot}0002 \times 10^{-6}}$$

$$= \frac{980}{1 - {\cdot}96}$$

$$= 24{,}500 \text{ ohms},$$

and is positive. It may therefore be represented by an equivalent inductance L_o, where

$$L_o = \frac{24{,}500}{\omega}$$

$$= 4{,}900\ \mu H$$

With the typical triode, therefore, the input conductance at this frequency would be approximately $-\frac{\mu C_{ag} r_a}{L_o}$ or $-\frac{1}{200}$ siemens and if the input circuit is tuned to the frequency corresponding to $\omega = 4{\cdot}9 \times 10^6$ the mere switching on of the power supply is sufficient to cause self-oscillation as explained in paragraph 65.

67. On the other hand, if the applied frequency is slightly higher than the resonant frequency of the anode circuit, e.g. if $\omega = 5 \cdot 1 \times 10^6$ the load reactance is approximately of the same magnitude as in the case just discussed, but of positive sign. The effective dynamic resistance of the input circuit is then less than 200 ohms. The following example is illustrative of the conditions which may exist at very high radio frequencies.

Example.—(i) The typical triode is required to amplify a signal at the frequency corresponding to $\omega = 4 \cdot 9 \times 10^7$, the equivalent tuned anode having the constants $L = 20\ \mu H$, $C = 20\ \mu\mu F$, $R = 10$ ohms. Find the input conductance.

The resonant frequency of the tuned anode is $\frac{5 \times 10^7}{2\pi}$, and its reactance at the applied frequency is

$$X_o = \frac{\omega L}{1 - \omega^2 LC}$$

$$= 24{,}500 \text{ ohms}$$

$$L_o = 490\ \mu H$$

$$\frac{(\omega L_o)^2}{r_a}(\mu + 1) = (2 \cdot 45 \times 10^4)^2 \times 11 \times 5 \times 10^{-5}$$

$$= 330{,}000$$

$$\frac{L_o}{C_{ag}\, r_a} = 4 \cdot 9 \times 10^{-4} \times 2 \times 10^{11} \times 5 \times 10^{-5}$$

$$= 4{,}900$$

$$\mu \frac{L_o}{C_{ag}\, r_a} = 49{,}000.$$

$$\frac{1}{\omega C_{ag}} = \frac{2 \times 10^{11}}{4 \cdot 9 \times 10^7}$$

$$= 4{,}070$$

$$\omega L_o - \frac{1}{\omega C_{ag}} \doteqdot 2 \times 10^4$$

$$G_m = \frac{(330 - 49)\, 10^3}{(\cdot 49 \times 10^4)^2 + (2 \times 10^4)^2}$$

$$= \frac{281}{4 \cdot 24 \times 10^5}$$

$$= 6 \times 10^{-4} \text{ siemens,}$$

which is equivalent to a leak of 1,667 ohms across the input circuit.

(ii) If the aerial circuit has an inductance L_A of 40 μH and a resistance R_A of 100 ohms find the overall gain of the aerial and amplifying stage.

The dynamic resistance of the aerial circuit is $\frac{\omega^2 L_A{}^2}{R_A} \doteqdot 40{,}000$ ohms, and its magnification, $\chi \doteqdot 20$. The total dynamic conductance of the input circuit is

$$G'_d = \frac{1}{4 \times 10^4} + \frac{6}{10^4}$$

$$= \frac{6 \cdot 25}{10^4}$$

$$R'_d = \frac{10^4}{6 \cdot 25} = 1{,}600 \text{ ohms.}$$

The effective magnification of the input circuit is therefore

$$x' = \frac{R'_d}{\omega L} = \frac{1{,}600}{5 \times 10^7 \times 40 \times 15^6}$$
$$= \cdot 8.$$

The V.A.F. of the amplifier alone is

$$\frac{R_{do}}{R_{do} + r_a} \mu$$

where R_{do} is the dynamic resistance of the anode circuit and is equal to 100,000 ohms.

$$\text{V.A.F.} = \frac{100{,}000 \times 10}{120{,}000}$$
$$= 8 \cdot 7$$

and the overall gain only $\cdot 8 \times 8 \cdot 7 \doteqdot 7$ which is of course much less than the magnification of the aerial circuit.

68. The introduction of an amplifier stage may therefore actually reduce the voltage available at the detector. It is often thought that the valve fails to act as an amplifier at high frequencies, but the above example shows that this is not the case. The valve performs its amplification in the manner predicted by the theory discussed in paras. 45 *et seq.*, but fails to give an overall gain owing to the damping it imposes upon its input circuit. If however the anode circuit is very slightly detuned so that its resonant frequency is very slightly higher than the signal frequency, its input conductance becomes sufficiently negative to maintain the input circuit in oscillation. In practice it is impossible so to adjust the anode circuit that stable amplification is obtained, unless the input and output circuits are of very low dynamic resistance. In these conditions, although the amplifier gain may be appreciable, the overall gain is negligible. An amplifier in which stability is attained by the employment of a high ratio of capacitance to inductance, giving low dynamic resistance and negligible overall gain, is often referred to as a buffer amplifier.

Use of screen-grid valves

69. (i) Unless special precautions are adopted, then, it is almost impossible to maintain a tuned radio-frequency triode amplifier stage in a non-oscillatory condition. Although at first sight it would appear possible to adjust both input and output circuits to the same frequency, so that they offer dynamic resistance only, and thus to ensure a positive input conductance, this is not so in practice. One complication which arises is that the two circuits are capacitance-coupled as in fig. 30, Chapter VI, and therefore possess two resonant frequencies. If the circuit constants are (L_1, C_1), (L_2, C_2), and $L_1 \times C_1 = L_2 \times C_2 = LC$ we have

$$f_1 = \frac{1}{2\pi\sqrt{LC}}$$

and to this frequency both circuits are practically non-reactive. The other resonant frequency is

$$f_2 = \frac{1}{2\pi\sqrt{LC\left(1 + \frac{C_{ag}}{C_1} + \frac{C_{ag}}{C_2}\right)}}$$
$$= \frac{f_1}{\sqrt{\left(1 + \frac{C_{ag}}{C_1} + \frac{C_{ag}}{C_2}\right)}}$$

which is obviously less than f_1, hence both input and output circuits offer inductive reactance at this frequency. The input conductance will therefore be negative, and if the positive grid-filament conductance is sufficiently low the valve will maintain oscillations at this frequency.

(ii) Owing partly to its low amplification factor, but chiefly to the difficulty of preventing self-oscillation, the triode valve is rarely used for radio-frequency amplification in modern receivers. Either the screen-grid valve, or one of its later developments, e.g. the variable-mu screen-grid valve and the radio-frequency pentode, is invariably adopted. The variable-mu valve is chiefly used where it is necessary to adopt some form of automatic radio-frequency gain control, and is again referred to in Chapter XII. The object of the radio-frequency pentode is to avoid the effects of secondary emission and so enable a larger anode voltage swing than is permissible with screen-grid valves. It is therefore capable of giving a greater output voltage than its prototype. The variable-mu pentode combines both the above features and is largely used in broadcast receivers. The gain and stability of the screen-grid valve will now be dealt with, the discussion being equally applicable to valves of later development.

Gain and stability

70. The principal features and characteristic curves of the screen-grid valve were discussed in chapter VIII. It was there stated that the introduction of the screening electrode reduces the effective grid-anode capacitance to only a fraction of a micro-microfarad and the input admittance is therefore very much less than that of a triode. Under normal operating conditions, namely with a screen voltage of about two-thirds the anode voltage, and with the grid swing limited to rather less than two volts, the equivalent circuit (and therefore the method of computing stage gain) is identical with that of a triode amplifier. The S.G. valve is not inherently stable in operation as is sometimes thought, for although its anode-grid capacitance is low it is not zero, and there is an upper limit to the magnitude of the dynamic load resistance R_{do} on this account. The input conductance becomes negative at a frequency very near to resonance, its minimum value being approximately

$$G_m = -\frac{\omega C_{ag} A}{2} = -\frac{\omega C_{ag}}{2} \times \frac{g_m}{\frac{1}{R_{do}} + \frac{1}{r_a}}$$

Instability ensues if the total grid-filament conductance is zero, so that if G_d is the admittance of the input circuit in the absence of the valve the limiting condition for perfect stability is clearly

$$G_d + G_m = 0.$$

Inserting the minimum value of G_m

$$G_d - \frac{\omega C_{ag}\, g_m}{2\left(\frac{1}{R_{do}} + \frac{1}{r_a}\right)} = 0$$

$$G_d\left(\frac{1}{R_{do}} + \frac{1}{r_a}\right) = \frac{\omega C_{ag}\, g_m}{2}$$

If for simplicity it is assumed that the input and output circuits are of similar design, so that $G_{do} = \frac{1}{R_{do}} = \frac{1}{R_d}$,

$$\frac{1}{R_d}\left(\frac{1}{R_d} + \frac{1}{r_a}\right) = \frac{\omega\, C_{ag}\, g_m}{2}.$$

This is the fundamental formula for use in calculations on stability. As in practice r_a is always very much larger than R_d, it is permissible to use a further approximation; provided $\frac{1}{R_d} >> \frac{1}{r_a}$, $\frac{1}{R_d}\left(\frac{1}{R_d} + \frac{1}{r_a}\right) \doteqdot \frac{1}{R_d^2}$, and

$$\frac{1}{R_d^2} = \frac{\omega C_{ag}\, g_m}{2}$$

The maximum permissible dynamic resistance with which the circuit will remain stable is therefore

$$R_d \text{ (max.)} = \sqrt{\frac{2}{\omega C_{ag}\, g_m}}.$$

71. As an example of the use of this equation, we may calculate the maximum permissible dynamic resistance at 796 kc/s ($\omega = 5 \times 10^6$) for a typical battery-operated S.G. valve ($g_m = 1$ milliampere per volt, $r_a = 300{,}000$ ohms, $C_{ag} = \cdot 005\ \mu\mu F$.)

$$R_d \text{ (max.)} = \sqrt{\frac{2}{5 \times 10^6 \times 5 \times 10^{-15} \times 10^3}}.$$

$$= 283{,}000 \text{ ohms.}$$

As the frequency is increased the permissible value of R_d becomes smaller, e.g. at $\omega = 5 \times 10^7$ it is only 90,000 ohms. Such high dynamic resistances are not ordinarily obtainable in the absence of regenerative amplification from the succeeding stage, and neglecting this complication the above assumption, $R_d << r_a$ invariably applies. The V.A.F. may then be reduced to a very simple expression, i.e.

$$\text{V.A.F.} = \frac{\mu R_d}{r_a} = R_d\, g_m$$

and with the maximum permissible value of R_d

$$\text{V.A.F.} = g_m \sqrt{\frac{2}{\omega C_{ag}\, g_m}} = \sqrt{\frac{2\, g_m}{\omega C_{ag}}}.$$

At $\omega = 5 \times 10^6$ the maximum V.A.F. is about 200 and at $\omega = 5 \times 10^7$ is about 70. In practice the gain seldom approaches such high values without a tendency to instability, owing to the effects of other forms of coupling between input and output circuits. In any event it is absolutely essential that the magnetic and electrostatic screening shall be of a very high order if gains comparable with the theoretical maxima are sought. The selectivity of the S.G. amplifier is theoretically higher than that of a triode with an identical anode circuit, because the damping due to the anode A.C. resistance of the valve is much less. Compared with the tuned anode circuit, tuned transformer coupling does not improve the selectivity to the same extent as in triode amplification for the same reason. In R/T reception the improvement in selectivity is only achieved when the input circuit has a high degree of initial selectivity, owing to an effect known as cross-modulation which will receive attention in Chapter XII.

Multi-stage amplifiers

72. An important consideration in the design of tuned radio-frequency amplifiers is the increase of selectivity obtained by using two or more stages in cascade. Curve (i) of fig. 37 shows the selectivity of a single tuned anode stage, and corresponds with curve (i) of fig. 27 ; instead of being plotted as gain against frequency, however, it is shown as "decibels below response at resonance" against "kilocycles off resonance," and it is seen that for equal input voltages a signal 66 kc/s off resonance is attenuated by 6 db. With two such stages, and no regenerative inter-stage coupling, the same reduction is obtained when the interference is 42 kc/s off resonance (curve ii), and when 32 kc/s off resonance in the case of three similar stages (curve iii). If the gain at resonant frequency is A, the overall gain of n stages is theoretically equal to A^n, although in practice it is usually less owing to the effect of the input admittance of each valve upon the gain of the preceding stage. It is usual to provide some means of reducing the amplification in order that when the initial field strength is high, the detector valve may not be overloaded ; if the original selectivity is to be maintained this control must operate in some manner other than the introduction of damping into the oscillatory circuits. When screen-grid valves are used, the gain is practically equal to $R_d\, g_m$, and g_m varies with the mean grid bias to a

greater extent than with triodes. It is therefore possible to control the amplifier gain by applying negative grid bias, and this arrangement is fairly satisfactory in W/T amplifiers, although for R/T reception, it is not entirely so. Alternatively, a variation of screen potential may be employed for gain control. The disadvantages of both these methods, when used in R/T receivers, are dealt with in Chapter XII.

73. The circuit diagram of a typical receiver designed in accordance with the foregoing principles, is given in fig. 38. Two stages of radio-frequency amplification are employed in order that considerable selectivity may be achieved. The inductances L_a, L_b and L_c are of equal value, and are tuned by the condensers C_a, C_b, C_c, which are mounted on a common axis and are capable of simultaneous variation by a single tuning control. Such condensers are said to be "ganged". Controllable reaction is obtained by the variable condenser C_r, the maximum capacitance of which may be only a few micro-microfarads. This is virtually in parallel with

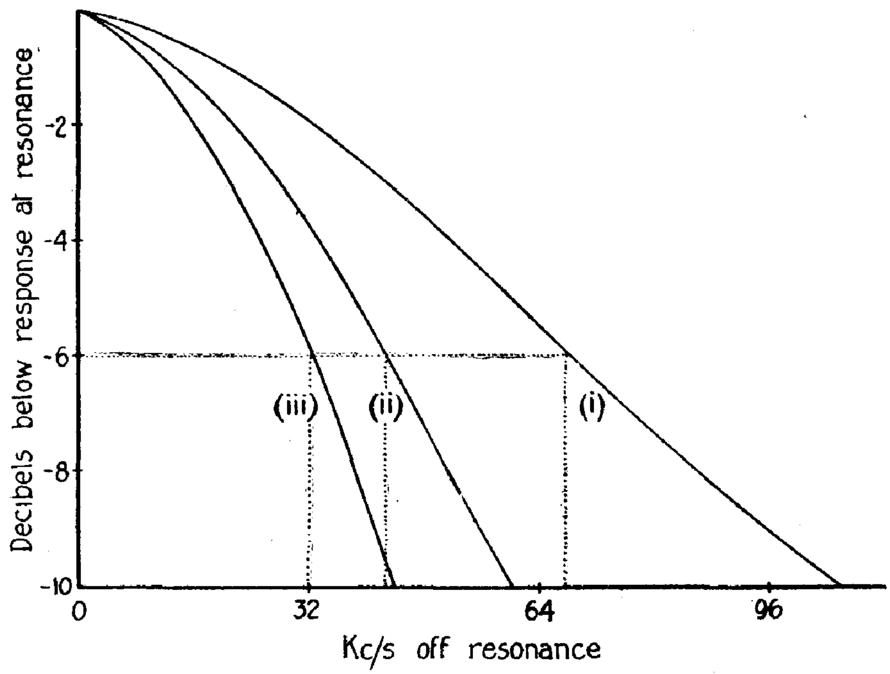


FIG. 37, CHAP. XI.—Selectivity of several R.F. stages in cascade.

the grid-anode capacitance of the second R.F. amplifying valve. The third valve operates as a cumulative grid rectifier, and is transformer-coupled to a stage of audio-frequency amplification, a similar stage being interposed between this and the final valve, which operates as a power amplifier. The gain is controlled by the potentiometer R_s, by which the screen potential of the amplifying valves may be adjusted between limits of about 40 to 80 volts.

Self oscillation in multi-stage amplifiers

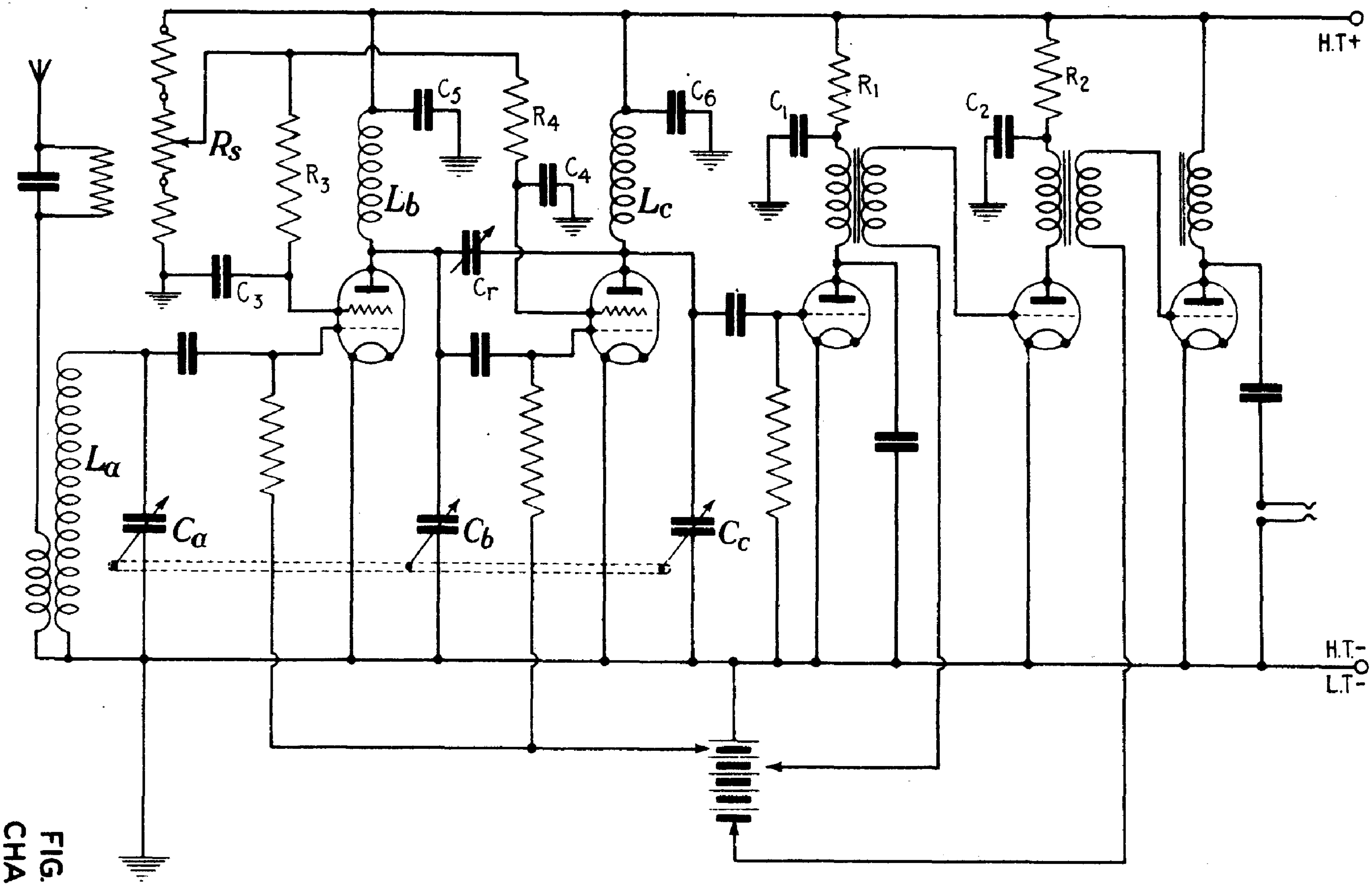
74. In addition to the effects caused by grid-anode capacitance and fortuitous inductive coupling between stages, one important form of inter-stage coupling is that caused by the employment of a single H.T. supply for all valves in the receiver. The internal resistance of a new 120-volt battery may be only about 20 ohms, but during use this increases and may reach several hundred ohms before the battery is discarded. This resistance is common to all the anode circuits in the receiver, and may give rise to transfer of energy between them, the result being to cause either positive or negative reaction effects, both of which are undesirable. The frequency of the oscillation caused by positive reaction may be from one to several thousand cycles per second ; when of the order of from 4 to 8 cycles per second the resulting noise emitted is often referred to as "motor-boating". It must be realized that the unwanted coupling cannot

be reduced to any extent by the use of a condenser in parallel with the H.T. battery, e.g. at 10 cycles per second a 10 $\mu\mu F$ condenser has a reactance of $1 \cdot 6 \times 10^4$ ohms, which is very large compared with the internal resistance of the battery, and in any event is still common to all circuits of the amplifier. Each A.F. anode circuit is therefore given its own by-pass or decoupling condenser; in fig. 38 these are denoted by C_1 and C_2. The greater portion of the varying component of anode current is confined to the capacitive path by the insertion of decoupling resistances R_1 R_2. Each decoupling resistance may be from 500 to several thousand ohms, the larger values being used when a certain fall of voltage is permissible, e.g. in the detector and early A.F. stages, while the decoupling condensers are usually of from ·5 to 2 microfarad. Radio-frequency stages are usually sufficiently decoupled by the use of a condenser only, provided it is inserted immediately adjacent to the anode load inductance as shown in fig. 38. Decoupling is also necessary in the screen circuits of the R.F. stages when these are fed from a single potentiometer, owing to the large common resistance which then exists. In fig. 38 these resistances and condensers are indicated as R_3, R_4, C_3, C_4. When the H.T. supply is drawn from either D.C. or A.C. mains, the effective internal resistance of the supply device may be many hundred ohms, and decoupling is an essential feature of such receivers, while if the grid bias is provided by any device offering a high impedance, decoupling is also necessary in the grid circuits.

Methods of tuning

75. (i) In the circuit diagrams illustrating the various types of radio-frequency amplifier, the various circuits have been drawn on the assumption that for a given frequency band—usually referred to as the tuning range of a particular inductance—the latter is of fixed value, the frequency adjustment being performed by means of a variable condenser. Where an amplifier is required to cover a wide frequency range, e.g. from 3 Mc/s to 100 kc/s, it is necessary to use a number of inter-changeable coils each of which, in conjunction with the tuning condenser, will cover a definite portion of the total frequency band. For example, let us consider a tuned-anode amplifier in which the capacitance of the tuning condenser (including all distributed capacitance in parallel therewith) may be varied from ·00005 to ·0005 μF. An 80 μH coil will then cover a frequency range of from 2,500 to 796 kc/s, a 160 μH coil from 1,780 to 560 kc/s, a 320 μH coil from 1,250 to 398 kc/s and a 640 μH coil from 890 to 280 kc/s. To cover the frequency range from 2,500 to 280 kc/s therefore, we actually require only 2 coils, of 80 μH and 640 μH respectively. The provision of an intermediate coil of 160 μH will however allow the tuning condenser to be used near its mean value over a greater portion of its range and will tend to give a greater approach to uniform selectivity.

(ii) An alternative method of tuning is to employ a fixed capacitance for each frequency band and to adjust the value of the inductance in order to tune to the desired frequency. The most convenient method of attaining this end is by varying the magnetic properties of the coil, and is usually referred to as permeability tuning. Although the practical application was attempted many years ago really satisfactory methods of permeability tuning have only been developed in recent years. The chief object in view is to maintain the ratio $\frac{L}{R}$ at a high and appreciably constant value over the whole tuning range of any particular coil. Since a fixed capacitance is associated with the latter, this also implies a nearly constant stage gain and an approach to uniform selectivity. The practical realization of this ideal depends upon the introduction of a core of comparatively low reluctance, which is composed of iron dust bound with a phenolic compound. The introduction of iron in any shape or form in radio-frequency tuning circuits is, however, fraught with considerable difficulty. The basic idea is, of course, to alter the resonant frequency of the circuit, but the earlier attempts were found to introduce such heavy losses that the circuit became practically aperiodic. In the modern method of permeability tuning the core is made of iron dust of a high purity. It must be very finely divided



TYPICAL RECEIVER

FIG. 38
CHAP. XI

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and is generally produced by the chemical reduction of iron oxide. The diameter of the particles is only of the order of 5 × 10^{-6} inch. Each individual particle must be separately insulated from all others, and this necessitates the use of a special varnish which has high viscosity and yet is capable of forming a very thin film under pressure of the order of 20 tons per sq. in. The particles are afterwards incorporated with phenolic resin to form a substance which resembles solid iron in appearance and actually contains more than 90 per cent. of iron, its specific gravity being of the order of 5, as against 7·8 for solid iron.

Amplifier noise

76. In any form of receiving circuit it will be found that some sound output exists under all circumstances, in addition to that caused by the desired signal or by electrical interference. This is often termed " amplifier noise". A portion of the noise is the result of such obvious causes as induction from power circuits, run-down accumulators, exhausted H.T. and grid bias batteries, faulty contacts, defective components and mechanical vibration of valve electrodes. The remedy in most of these instances is in the hands of the operator, and suitable steps for the elimination of electrical interference have been suggested in the previous chapter. Some valves are more " microphonic " than others, filament tension being one factor in this respect. Specially sprung valve holders are a palliative usually adopted. When all these causes are non-existent, there is always a certain amount of noise which cannot be eliminated because it arises from the actual mechanism of the valve itself. There are three distinct types of spontaneous fluctuation of anode current in any amplifier, which are known as (i) the " shot " effect, (ii) the " flicker " effect, and (iii) the temperature effect. The existence of these effects imposes a limit to the amplification of very small voltages.

The shot effect results from the fact that the anode current is actually a stream of discrete particles rather than the continuous flow generally envisaged. The presence of a space charge tends to smooth out any irregularity in the rate at which electrons arrive at the anode, and it is therefore necessary to provide ample emission from the cathode if the shot effect is to be maintained at a low level.

The flicker effect is probably chiefly caused by the occasional emission of positive ions from the cathode, and to ionisation of residual gas molecules. Whereas the shot effect occurs to the same degree irrespective of the operating frequency, the flicker effect is of greatest importance in audio-frequency amplification, particularly below 1 kc/s. Improvements in the cathode and vacuum may eventually reduce the flicker effect to a negligible amount.

The temperature effect is at once the best understood in theory and the most serious in practice. It is due to the random motion of the electrons within any conductor as briefly described in Chapter I. For engineering purposes it is usual to say that an electron current flows when a large number of electrons are simultaneously moving in the same direction, but in reality every movement of an electron constitutes a current. Although in the absence of an applied E.M.F. the average current is zero, the R.M.S. current may be comparatively large, because even though the average velocity along the conductor is zero, the electrons themselves move about at random in the conductor with a velocity of about 12 × 10^6 cm. per second. It may also be noted that various electrons or groups of electrons set up voltages at one frequency and other groups at different frequencies, hence the total energy is distributed throughout the frequency spectrum. If the conductor is a portion of the input circuit of an amplifier, the voltage set up in the anode circuit will depend chiefly upon its selectivity. As an example of the magnitude of this effect, it has been calculated that if the amplifier gives effective amplification over a 5 kc/s band, and the input resistance is a ·5 megohm grid leak, the noise level produced is equivalent to that produced by an input voltage of 6·4 micro-volt. The total amplifier noise determines the minimum voltage which can be usefully amplified. In the case just cited it is obvious that no advantage will be obtained by using this amplifier to increase the strength of a signal which is initially less than 6·4 micro-volt, for the noise and the signal will be amplified to the same degree and there will be no increase in the ratio of signal to noise.

Amplification of high and very high frequencies

77. The gain of a stage of amplification is chiefly dependent upon the possibility of designing an anode circuit impedance comparable with or greater than the anode A.C. resistance of the amplifying valve, and also upon the input admittance, for if the latter is low it has been shown that the input voltage may be reduced to such an extent that the overall gain is actually negative, i.e. a loss. The amplification obtained at high frequencies (with conventional valves and circuits) is always very low, because the input conductance of the succeeding valve shunts the output circuit in the above manner. The input resistance of a cumulative grid rectifier following a R.F. stage may be only 5,000 ohms, and no matter how large the dynamic resistance of the preceding output circuit may be, the effect is to reduce the anode load of the amplifying valve to less than the input resistance of the detector. Little advantage is to be gained by the use of screen-grid valves or R.F. pentodes in such circumstances, as is shown below.

A triode, having an r_a of 20,000 and an amplification factor of 20, works into a resistive load of 5,000 ohms. Find the V.A.F.

$$\text{V.A.F.} = \frac{R_a}{r_a + R_a}\,\mu$$

$$= \frac{5{,}000}{25{,}000} \times 20$$

$$= 4.$$

If the above valve is replaced by a screen-grid valve, $r_a = 300{,}000$, $\mu = 300$, the V.A.F. is

$$\frac{v_a}{v_g} = \frac{5{,}000}{305{,}000} \times 300$$

$$= \frac{5 \times 300}{305}$$

$$= 4{\cdot}9.$$

The overall gain will be higher than appears from the latter figure because the input admittance of the S.G. valve will be greater than that of the triode and so will not damp the input circuit to the same extent. This is not a serious disadvantage of the triode, however, because reaction may be used to counteract the damping due to the valve. The chief disadvantage of reaction is the difficulty of very smooth control at such high frequencies. The reduction of inter-electrode capacitance in modern R.F. pentodes has made R.F. amplification with conventional circuits feasible up to about 10 Mc/s.

78. By means of special circuits and valves (e.g. the magnetron, chapter IX) it is possible to produce electro-magnetic waves of exceedingly high frequency ; in this connection it will be recalled that about 300 Mc/s is the limit above which ordinary reaction methods fail to generate oscillations because the duration of a single cycle is comparable with the average time taken by the individual electrons to travel from cathode to anode. Ordinary valves also fail to function as amplifiers and rectifiers at such high frequencies for the same reason, while the following factors are contributory to this failure :—

(i) The inter-electrode capacitance of the valve is so large that the ratio $\frac{C}{L}$ is too great to achieve any considerable dynamic resistance.

(ii) The inductance of the connections between the external contacts and the actual electrodes is considerable, and results in an appreciable reactive drop.

(iii) The inductance referred to in (ii) acts in conjunction with the inter-electrode capacitance to form undesired tuned rejector circuits.

(iv) The R.F. resistance of the valve is increased by the presence of these inductive and capacitive paths.

Acorn valves

79. In any given valve design, if all the physical dimensions are reduced in a common ratio, there will be no change in mutual conductance, anode current and amplification factor, for given operating potentials. On the other hand the inter-electrode capacitance, lead-in inductance and time of electron transit are all reduced in direct proportion to the reduction ratio. This principle may be expressed by the statement that for optimum performance the linear dimensions of all apparatus should be proportional to the wavelength. At low frequencies, however, adherence to this principle would lead to inconvenient dimensions with little improvement in

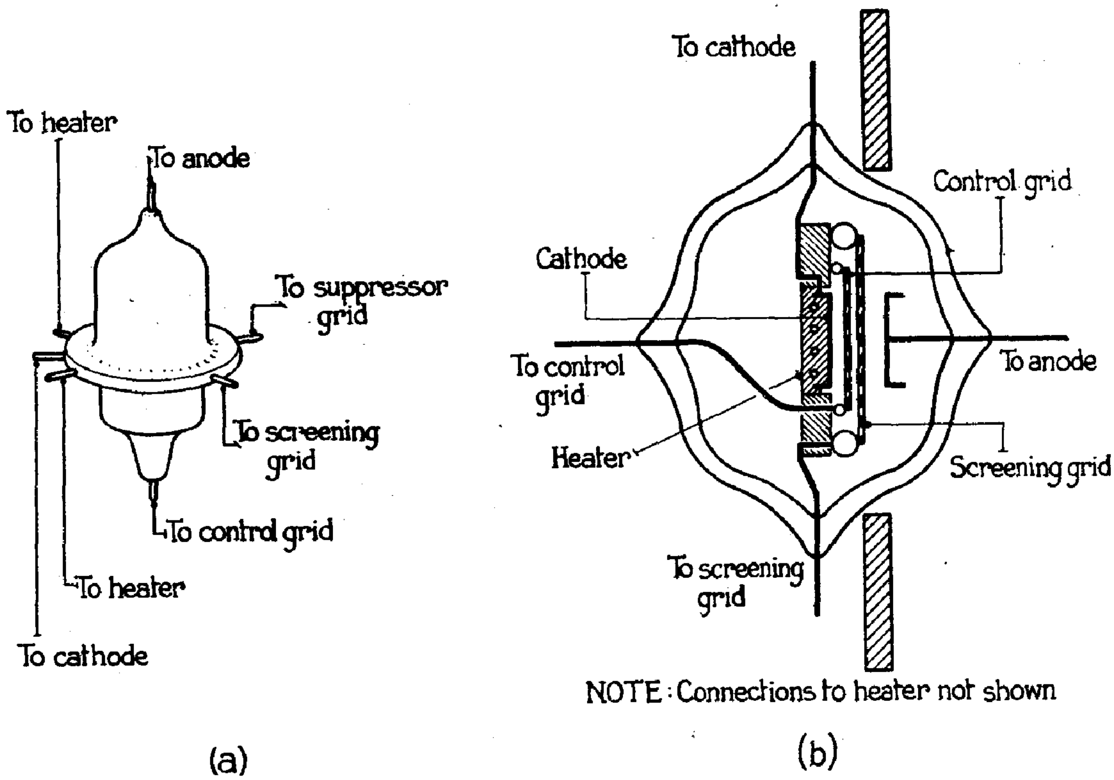


FIG. 39, CHAP. XI.—(a) Acorn R.F. pentode, actual size.
(b) Arrangements of electrodes and connecting leads in acorn tetrode.

performance, but it has been used to produce valves giving appreciable amplification at frequencies up to 600 Mc/s. As the lower limit of useful application of say a standard R.F. pentode is about 5 metres, approximately the same performance can be obtained at ·5 metre if all the dimensions of the valve are reduced to one-tenth the normal. Valves produced according to this theory are often referred to as " acorn " valves, the term giving a good idea of the actual size of the valve. Fig. 39a is a perspective drawing of an R.F. pentode of this type, actual size, and also shows the manner in which connections are made to the electrodes, while the latter point is further illustrated in the enlarged sectional drawing of the screen-grid valve of this type (fig. 39b). It will be noted that the electrodes are so arranged and supported that no " pinch " is necessary. The cathode is indirectly heated, but as the current consumption is quite low (about equal to that of a small power valve) the heating current may be derived from a secondary battery. No valve of this type has so far been standardized for service use. Different makers use different heater voltages and slightly different arrangements of lead-in pins. If valves of this kind are used for experimental purposes low-loss valve holders of porcelain or other efficient dielectric should be used, but on no account should connections be soldered directly to the valve, as the glass may soften and allow air to enter the interior.

80. Conventional circuits may be employed for R.F. amplification, detection and the production of oscillation up to about 600 Mc/s, provided great care is taken in the arrangement of circuits, the use of shortest possible connecting leads, etc. Where by-pass condensers are used, these should be of quite small capacitance, and the inductance of their connectors reduced to the lowest possible limit. The principle of reduction of linear dimensions should be applied as far as possible to circuit components such as coils and condensers. For example, a coil of 5 turns of 25 s.w.g. copper (diameter, ·02 inch) ·125 inch in length and of the same diameter, has an inductance of ·0587 μH, and a resistance of ·37 ohms, at a frequency of 600 Mc/s. The capacitance required to tune this to the stated frequency is 1·2 $\mu\mu F$ and the dynamic resistance of such a circuit is 132,000 ohms, which is higher than is ordinarily obtainable at medium frequencies. Its magnification is of the order of 200. These theoretical values are undoubtedly much higher than are practically attainable. For example, the eddy current loss due to the proximity of both metallic and dielectric materials must be very serious and its magnitude incalculable. Nevertheless it is possible to achieve dynamic resistances and magnifications comparable with those attainable at medium frequencies.

THE SUPER-HETERODYNE RECEIVER

Principle of operation

81. The principle of heterodyne reception of C.W. signals is discussed in Chapter IX, and the supersonic heterodyne or super-heterodyne receiver operates upon similar principles. If a C.W. signal of 1,000 kc/s is to be received by the ordinary heterodyne method, the local oscillator may be set to say 1,001 or 999 kc/s, so that heterodyne beats occur 1,000 times per second. After rectification the output of the detector valve possesses (amongst many others) a component having a frequency of 1,000 cycles per second, which is then amplified as necessary and so caused to operate telephone receivers or loudspeaker at the desired power level. If the frequency of the local oscillator differs from that of the incoming signal by more than 1 kc/s the telephonic response is of higher pitch, and with a sufficiently large frequency difference will be above the limit of audibility as shown in fig. 14, Chapter X. It is important to observe, however, that although no audible sound is produced, the anode current of the detector valve still contains a component having a frequency equal to the number of beats per second. Thus, if the signal frequency is 1,000 kc/s and the local oscillator set to 800 kc/s or 1,200 kc/s, the difference is 200 kc/s and the anode current of the detector valve contains a 200 kc/s component. An output circuit tuned to this frequency will then be set into oscillation and becomes, virtually, the source of a signal having a frequency which is entirely controlled by the adjustment of the local oscillator ; such a combination of rectifier and local oscillator is said to function as a frequency-changer. It is also important to note that although the local oscillation is continuously maintained, the new frequency is only produced when a signal is received. To avoid confusion the oscillator used to produce this new radio-frequency will be referred to as the R.F. oscillator.

82. The frequency to which the incoming signal is transferred is called the intermediate or supersonic frequency and is often abbreviated to S.F. A super-heterodyne receiver therefore consists of the following :—

(i) Radio (signal) frequency circuits, preceding the frequency-changing valve, or valves.

(ii) The R.F. oscillator circuits.

(iii) Intermediate or supersonic frequency amplifier circuits.

(iv) Intermediate or supersonic frequency rectifying valve ; this is also referred to as the second detector.

(v) Audio-frequency amplifying circuits.

In its original form some means of rectification was invariably incorporated in the frequency-changer, and this stage is also sometimes referred to as the first detector. The modern frequency-changing valve performs a somewhat complicated operation upon the signal and the local oscillation, and is therefore sometimes called a " mixing " valve. By whatever name it may be known, it must be fully realized that its function is something more than a mere algebraic addition of the two voltages or currents.

83. If the original signal is of modulated wave-form, the output of the frequency-changer will also be modulated and the output of the second detector will have an audio-frequency wave-form similar to the modulation envelope of the original signal. During the reception of a C.W. signal, the frequency-changer will give an output of unvarying amplitude, and a second heterodyne oscillator is necessary in order to obtain an audio-frequency output from the second detector. This may be an entirely separate oscillator, or alternatively, one stage of the S.F. amplifier may be operated in a self-oscillatory state. The former is generally to be preferred

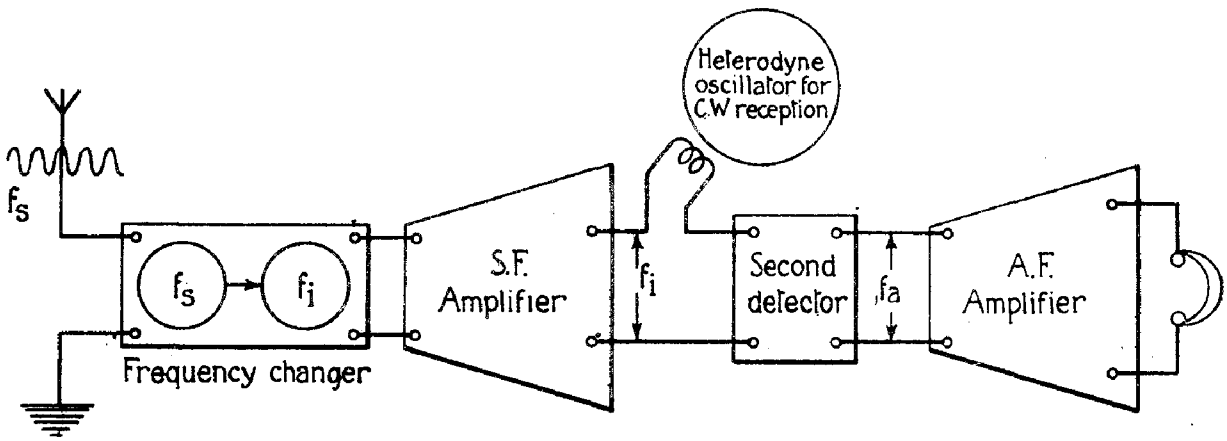


FIG. 40, CHAP. XI.—Schematic diagram of super-heterodyne receiver.

for several reasons, among which are, first, the amplitude of the local oscillation is more easily controlled ; second, the overall gain of the S.F. amplifier is greater, because if one stage is operated in a stage of self-oscillation it cannot contribute appreciably to the overall gain. The complete super-heterodyne receiver is shown schematically in fig. 40. For brevity, the oscillator required for C.W. reception will be referred to as the C.W. oscillator.

84. The advantages of the super-heterodyne receiver over the tuned radio-frequency amplifier are :—

(i) However high the signal frequency may be, the actual amplification is performed at the intermediate frequency, which is very much lower. The difficulties associated with the amplification of high radio frequencies are therefore overcome. This statement is not entirely applicable to the first S.F. amplifier of the double super-heterodyne receiver, which is dealt with later.

(ii) By suitable design and adjustment of the frequency-changer the same intermediate frequency is always obtained, and the S.F. amplifier may therefore be designed to amplify only a band extending from five to ten kc/s each side of the nominal S.F. Once they have been properly set up, these stages require no tuning adjustment whatever.

(iii) The overall selectivity is greater than that of the tuned radio-frequency amplifier.

The various forms of frequency-changer will be described briefly, before discussing the factors governing the choice of an intermediate frequency and the selectivity.

Single-triode frequency-changer

85. This is the simplest type, and the action of all others is easily followed if its operation is understood. The circuit diagram is given in fig. 41. The triode maintains the aerial circuit in oscillation by means of the reaction coupling between the coils L_1 L_2. Suppose a C.W. signal of 1,000 kc/s is to be received, and the S.F. amplifier stages to be tuned to 200 kc/s. The triode must then maintain the aerial circuit in oscillation at either 1,000 + 200 = 1,200 kc/s or 1,000 − 200 = 800 kc/s, and the circuit is, therefore, out of resonance with the incoming signal. This must cause a slight reduction in signal strength, which however is negligible at such high

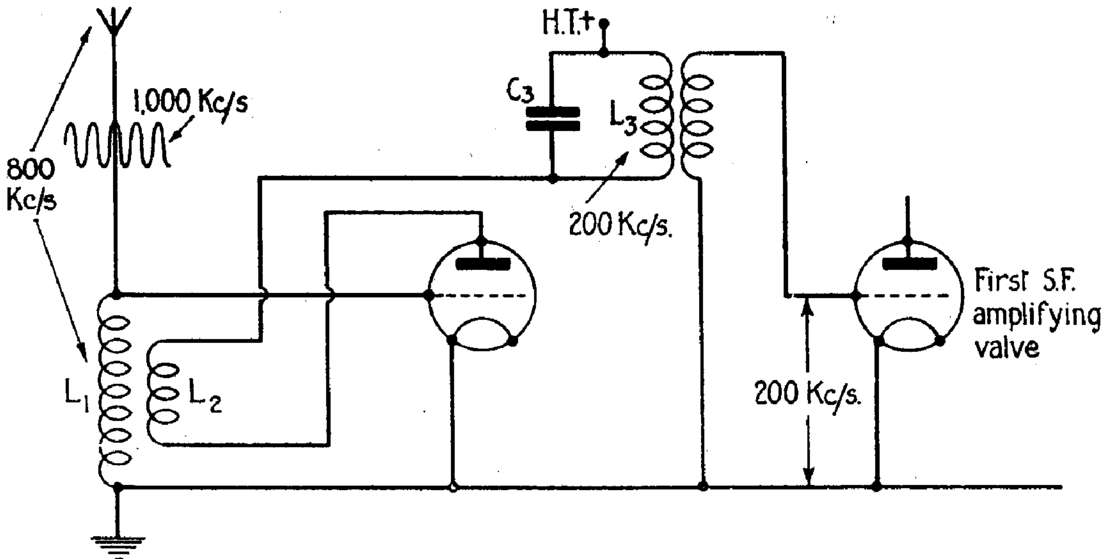


FIG. 41, CHAP. XI.—Single-triode frequency-changer.

frequencies. The amplitude of the oscillatory voltage between grid and filament, in the absence of any signal, must be fairly large, i.e. of the order of ten volts ; the resulting heterodyne beats will resemble those of fig. 42a. The $I_a - V_g$ curve of a suitable valve is shown in fig. 42b, with this voltage applied to the grid. The negative edge of the beat envelope has no effect upon the anode current, because it is located beyond the cut-off point on the grid voltage base-line, and the anode current variations may be considered to be due entirely to the positive edge. In this respect the valve functions in a manner somewhat similar to a lower-anode-bend rectifier, hence the term " first detector." The anode current varies in a complex manner, possessing components of 1,000 kc/s and 800 (or 1,200) kc/s, as well as 1,000 + 800 (or 1,000 + 1,200) kc/s. The important component however is that caused by the beat envelope, a.b.c.d.e ; this corresponds to an anode current component of 1,000 − 800 (or 1,200 − 1,000) kc/s, i.e. having a frequency of 200 kc/s. The output circuit (L_3 C_3, fig. 41) is tuned to this frequency, and will set up an appreciable oscillatory current in the tuned circuit. The resulting voltage across the coil L_3 will therefore be applied to the S.F. amplifier, and will eventually affect the telephones in the usual manner.

Conversion conductance

86. Although the action is complicated by the rectification process, the action of the frequency-changer is analogous to that of a radio-frequency amplifier, in that a signal voltage $v_s = \mathcal{V}_s \sin \omega_s t$ causes an anode current variation i_a where $i_a = g_c \mathcal{V}_s \sin \omega_i t$, $\frac{\omega_i}{2\pi}$ being the intermediate frequency. The constant g_c, i.e. the ratio $\frac{\text{intermediate frequency component of anode current}}{\text{signal frequency input voltage}}$ (in the absence

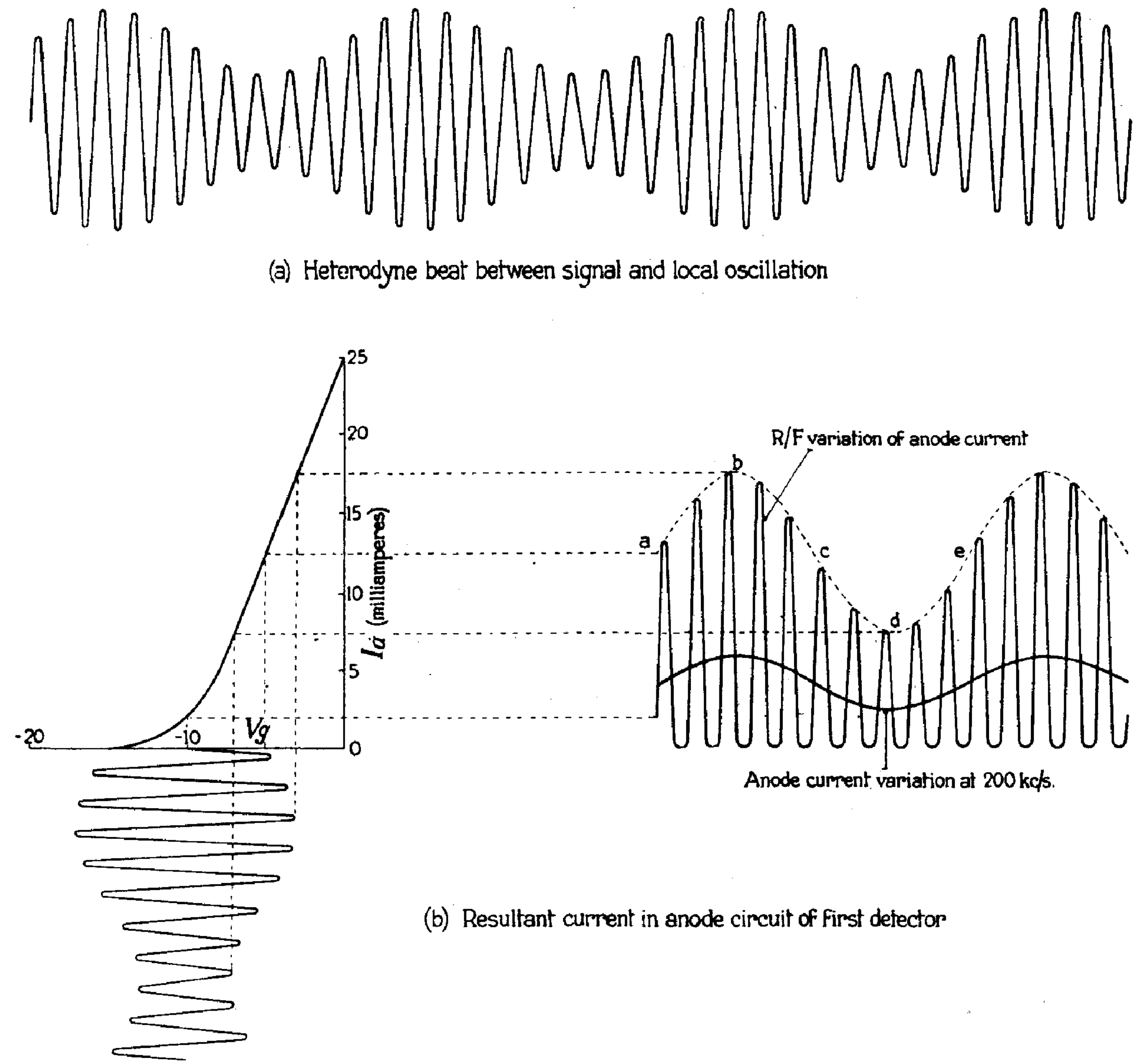


FIG. 42, CHAP. XI.—Operation of single-triode frequency-changer.

of any anode load), is termed the conversion conductance of the valve. It is analogous to the mutual conductance of the valve when used as an amplifier. When the anode circuit contains an S.F. dynamic load R_d, the relation between i_a and v_s is

$$i_a = \frac{g_c}{1 + \frac{R_d}{r_c}} v_s$$

where r_c is the effective anode A.C. resistance over the operating portion of the $I_a - V_g$ curve; r_c is sometimes called the conversion resistance of the valve.

87. In fig. 42b, the local oscillation has an amplitude of 5 volts and the amplitude $\mathscr{V}_s$ of the signal voltage is 2 volts, the amplitude of the beat voltage envelope varying between 7 and 3 volts. The anode circuit load is assumed to be absent, and the S.F. component of anode

current (shown in heavy line) has an amplitude of 1·5 milliamperes. The mutual conductance g_m is 2·5 milliamperes per volt, and the conversion conductance g_c is 1·5 milliamperes ÷ 2 volts = ·75 milliamperes per volt. In this particular instance $g_c = \frac{g_m}{3}$. The conversion conductance of a given valve depends upon the grid bias and the amplitude of the local oscillation as well as upon the shape of the $I_a - V_g$ curve. Under linear rectification conditions, to which fig. 42 approximates, g_c approaches $\frac{g_m}{\pi}$ as a limiting value. With a square law rectifier g_c is rarely greater than $\frac{g_m}{4}$. Owing to the presence of the local oscillation, the conversion resistance of the valve is of the nature of a dynamic resistance rather than of the static nature associated with the usual meaning of the term "A.C. anode resistance." It is possible to measure r_c by plotting an $I_a - V_a$ curve while an alternating voltage is applied between grid and filament, its amplitude being equal to that of the local oscillation under the intended operating conditions ; r_c is usually about 2 r_a.

Conversion factor

88. In the first detector of a super-heterodyne receiver the quantity corresponding to the V.A.F. of an ordinary amplifier is the conversion factor (C.F.), which is defined as the ratio of S.F. voltage across the output impedance to the signal voltage applied to the grid-filament path. From ordinary triode theory it is easily seen that the C.F. of a frequency-changer is given by the equation

$$\text{C.F.} = \frac{g_c\, r_c\, R_d}{R_d + r_c}$$

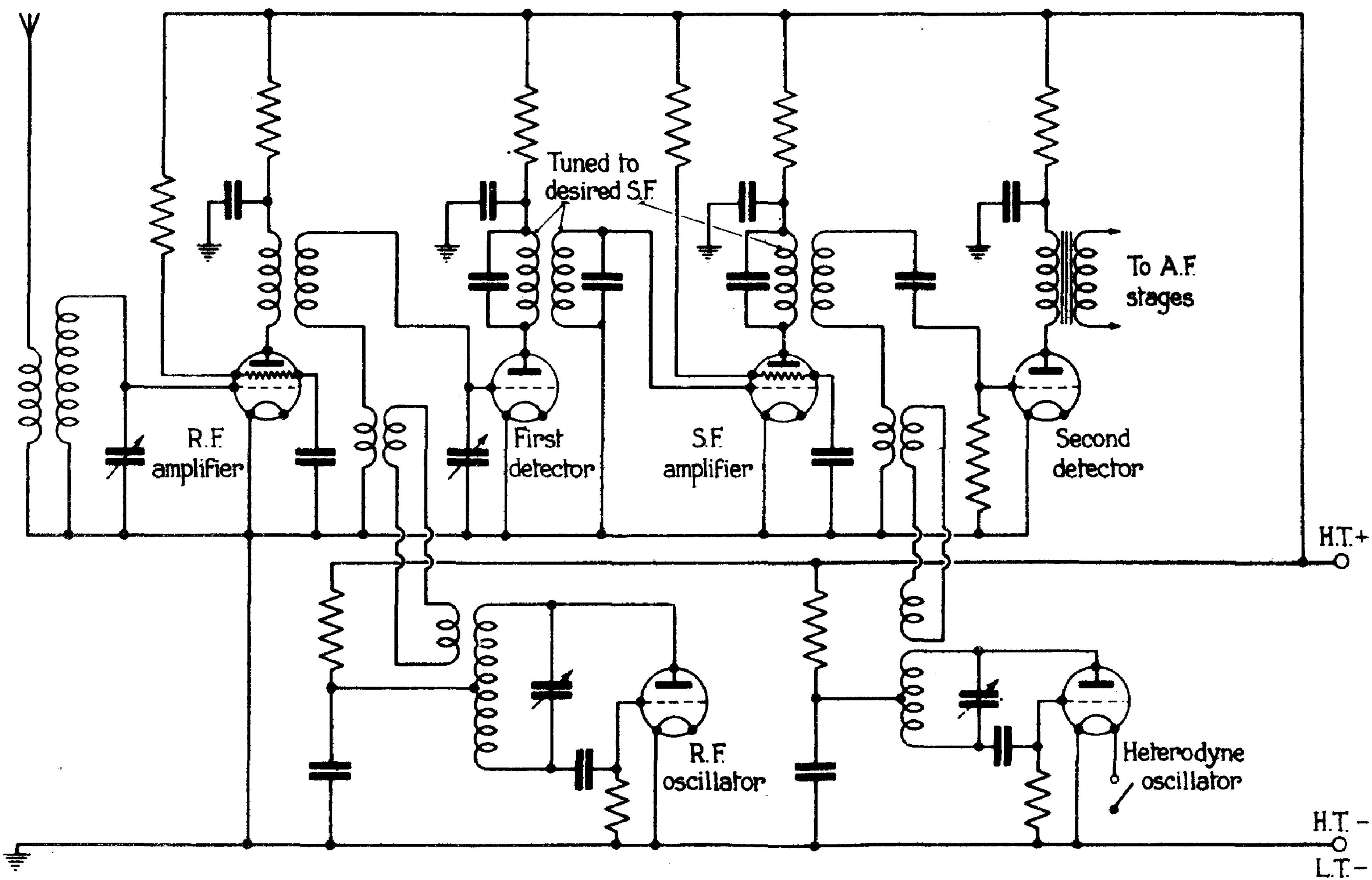
where R_d is the effective dynamic resistance of the load at the supersonic frequency. The product $g_c\, r_c$ is analogous to $g_m\, r_a$ ($= \mu$) in ordinary amplifier calculations. Obviously the conversion gain of various forms of transformer coupling can be dealt with by introducing k and T^2 (coupling factor and inductance ratio) as in paragraphs 48 *et seq.*, the conversion gain being $\frac{v_{g2}}{v_{g1}}$.

Disadvantages of single-triode frequency-changer

89. The single-triode frequency-changer suffers from several disadvantages, the principal being :—

(i) The aerial circuit is in oscillation and therefore radiates energy, causing interference with neighbouring receivers. This can be overcome to some extent by the use of a buffer stage between the aerial and the input to the frequency-changer, which however increases the number of tuning controls.

(ii) The operating conditions, e.g. H.T. voltage and grid bias, must be a compromise between the values best suited for the production of oscillation and those appropriate to a lower anode-bend detector. For maximum conversion conductance, it is desirable that the grid should be biased nearly to cut-off point (but not so negative that oscillations cannot be initiated) and that the grid swing of the local oscillation should not extend into the grid current region. In ordinary oscillators however, the local oscillation builds up to an amplitude which is limited chiefly by damping due to the flow of grid current, so that the conversion conductance must be less than the theoretical maximum. Further, the amplitude of the local oscillation will vary to a considerable extent with the tuning of the input circuit, and the sensitivity of the receiver will vary accordingly. In a certain receiver of this type it was found experimentally that the amplitude of the R.F. oscillation varied between 14 and 30 volts as the aerial tuning was changed from 6 to 3 Mc/s.



SUPER-HETERODYNE WITH TWO-TRIODE FREQUENCY CHANGER

FIG. 43
CHAP. XI

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(iii) The receiver as described above is equally effective for two frequencies. Suppose the oscillating aerial circuit to be tuned to 800 kc/s, and the S.F. amplifier to 200 kc/s. Then signals of 600 kc/s and 1,000 kc/s both set up 200,000 heterodyne beats per second and therefore both give rise to an output of the correct intermediate frequency. This is called "image-frequency interference" and can be eliminated by the introduction of suitable tuned circuits between the aerial and the frequency-changer. These are referred to as the pre-selector circuits and may of course be incorporated in a buffer stage.

The two-triode frequency-changer

90. (i) The first of the disadvantages of the single-triode frequency-changer can be overcome by generating the local oscillation in a separate triode oscillator. Fig. 43 gives a possible arrangement in which a signal frequency amplification stage is inserted before the frequency-changer in order to discriminate between the wanted and the image frequency. The R.F. oscillator circuit is coupled to the first detector by means of a link circuit, to reduce so far as is possible any alteration in the tuning of one circuit by the adjustment of the other. The output circuit of the first detector is tuned to the desired intermediate frequency. A screen-grid valve or radio-frequency pentode may be used as the first detector instead of a triode. A modification to this circuit utilizes a buffer or isolator stage between the oscillator and the detector (fig. 44). A screen-grid valve, variable-mu screen-grid valve, or radio-frequency pentode is preferably used in this position, and the buffer then acts as a coupling device for the transfer of energy from the oscillator to the input circuit of the first detector, but not in the opposite direction. Variation of the tuning control of one circuit then has only a negligible effect upon the tuning of the other.

(ii) It will be observed that in both the frequency-changers described above the action is similar to that of the ordinary heterodyne reception of a C.W. signal. The two voltages are first added, giving a heterodyne beat, and the latter is rectified to produce an oscillatory current of the beat frequency. This type of frequency change is sometimes called the additive method.

Multi-electrode frequency-changers

91. The introduction of the mains operated valve led to the trial of a new technique which is referred to as cathode injection. It was first applied to the screen-grid valve but soon led to the development of the triode-pentode, consisting of two entirely separate electrode assemblies in a single envelope, but with a common cathode. The triode section is caused to maintain the R.F. oscillation in a parallel feed circuit, and the incoming signal is applied between the cathode and the control grid of the pentode section. The oscillatory voltage across the grid coil of the triode is virtually in series with the signal voltage and the whole valve functions as an additive frequency-changer, rectification being still required to produce an oscillatory current of intermediate frequency.

The Pentagrid or heptode

92. The cathode injection system has been supplanted by what is called electron coupling in which the action at one electrode is in effect multiplied by that at another. The anode current then contains, amongst other components, a modulation product of the form $A \sin \omega_s t \sin \omega_o t$ where $\frac{\omega_o}{2\pi}$ and $\frac{\omega_s}{2\pi}$ are the frequencies of the R.F. oscillation and incoming signal respectively. Thus both sum $(\omega_s + \omega_o)$ and difference $(\omega_s - \omega_o)$ components may be expected to exist; the output circuit of the valve is tuned to the latter, and the intermediate frequency amplifier will therefore operate upon the difference frequency. The prototype of one class of electron coupled frequency-changer is the pentagrid or heptode (fig. 45a) which is in effect a triode and screen-grid valve in series. This valve has a cathode, five grids, and an anode, and its I_a—V_a curve exhibits the "negative slope" region peculiar to the screen-grid valve. The grids are referred to by numbers from the cathode outwards, the grid G_1 being used for the purpose of oscillator voltage control, while the grid G_2 acts as the R.F. oscillator anode. As the latter is perforated it allows the electron stream—which pulsates at the oscillator frequency—to penetrate to the outer electrodes. This electron stream, at the point of emergence from the screening grid G_3

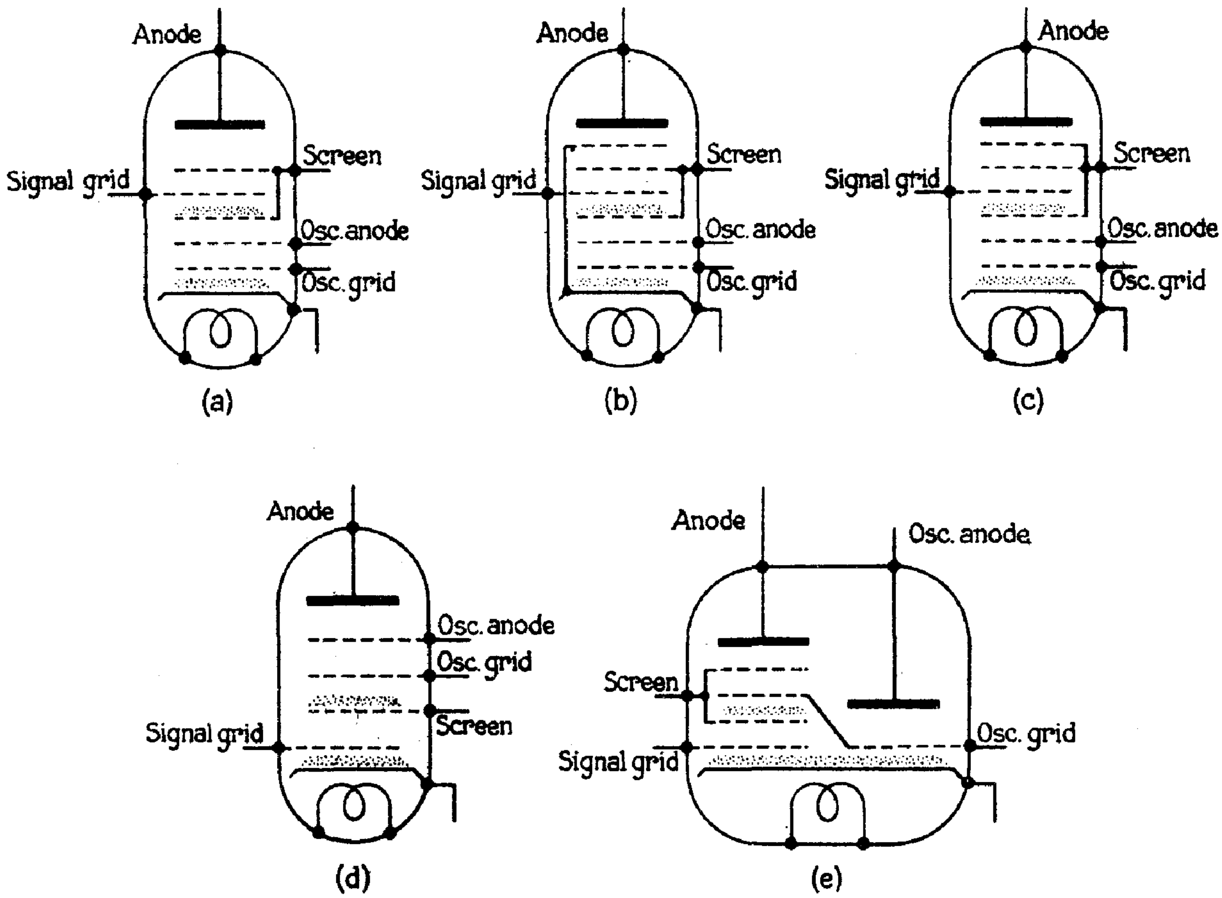


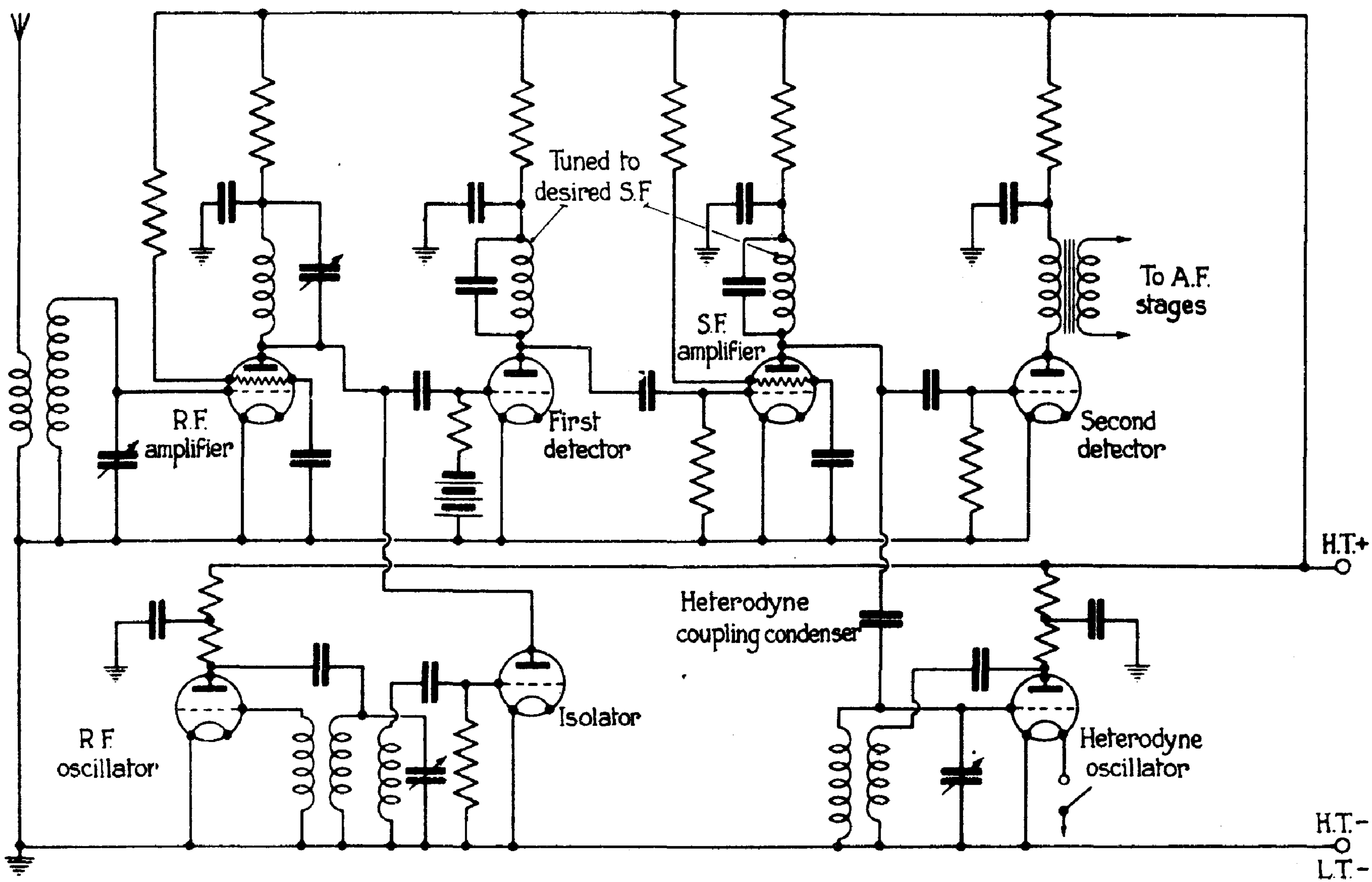
FIG. 45, CHAP. XI.—Frequency-changer valves. (a) pentagrid. (b) suppressor-grid octode. (c) accelerator-grid octode. (d) hexode. (e) triode-hexode.

is referred to as the virtual cathode of the tetrode formed by the remaining electrodes. The grid G_4 is the signal voltage control grid, while the grids G_3 and G_5 are connected internally and screen the control grid from the oscillator anode G_2 and from the true anode. The various circuits are connected to the valve as in fig. 46a. If the R.F. oscillator were inoperative, the emission of the virtual cathode would be constant and the mutual conductance $\left(\frac{dI_a}{dV_{g4}} = g_{m4}\right)$ of the tetrode portion also constant. When the R.F. oscillator is generating a frequency $\frac{\omega_o}{2\pi}$ the virtual emission is varied in a sinusoidal manner and the mutual conductance of the tetrode varies correspondingly, becoming in effect

$$g'_{m4} = \left(\frac{dI_a}{dV_{g4}}\right) = K\, g_{m4} \sin \omega_o t_o$$

For a small sinusoidal change $v_{g4} = \mathscr{V}_{g4} \sin \omega_s t$, in the potential of the signal control grid, the corresponding change in anode current is

$$\begin{aligned} i_a &= g'_{m4}\, v_{g4} \\ &= K\, g_{m4} \sin \omega_o t \times \mathscr{V}_{g4} \sin \omega_s t \\ &= \frac{K}{2}\, g_{m4}\, \mathscr{V}_{g4} \left[\cos (\omega_o - \omega_s)t - \cos (\omega_o + \omega_s)\, t\right] \\ &= g_c\, \mathscr{V}_{g4} \left[\cos (\omega_o - \omega_s)\, t - \cos (\omega_o + \omega_s)\, t\right] \end{aligned}$$



SUPER-HETERODYNE WITH ISOLATOR STAGE

FIG. 44
CHAP. XI

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Hence g_c cannot be greater than $\frac{g_{m4}}{2}$ because K cannot exceed unity. The output circuit is tuned to the difference frequency $\frac{\omega_o - \omega_s}{2\pi}$ and supplies excitation to the S.F. amplifier as in the types of frequency-changer previously discussed.

93. It will be observed that the anode current contains a difference component even if g'_{m4} is independent of the anode voltage and the grid bias hence the frequency changing action is not dependent upon the curvature of the static characteristics and the difference frequency is not produced by a rectifying action as in an additive frequency changer. In practice, the curvature of the characteristics may lead to the production of a S.F. by rectification of the beats resulting from the interaction of harmonics of the signal and any interfering frequency which may be present. The conversion gain of the pentagrid is considerably greater than that obtained with the single-triode or two-triode arrangement, but is of the same order as is obtained by employing a tetrode or radio-frequency pentode as the first detector, together with a separate R.F. oscillator.

The octode

94. The octode functions in a similar manner to the pentagrid, the principal difference being the introduction of an additional screening electrode. In the suppressor-grid octode fig. 45b, this screen is at cathode potential, giving the valve the $I_a - V_a$ curves of a radio-frequency pentode in that the region of negative slope is eliminated. It is therefore, capable (under certain conditions) of giving a larger output than the pentagrid. In certain designs, however, the screening electrodes are all maintained above cathode potential as in fig. 45c, that nearest the anode being called the accelerator grid. The $I_a - V_a$ curves then exhibit a region of negative slope. The chief advantage of the octode is that automatic volume control is more effective, and simpler in application, than in the pentagrid. Its conversion resistance is also higher, giving a slightly higher conversion gain for an equal conversion conductance, while the selectivity of the output circuit is somewhat greater, owing to the smaller damping effect of the valve resistance.

95. The development of other types of frequency-changer, such as the hexode and the triode-hexode, is bound up with the requirements peculiar to broadcast receivers, where the necessity for a single tuning control renders it necessary to gang the R.F. oscillator and signal frequency circuits. In most circumstances, this entails that the R.F. oscillator shall operate on the higher of the two possible frequencies. At high and very high frequencies the input impedance of the frequency-changer is also of importance. The tuned circuit of the R.F. oscillator is coupled to the input circuit by the inter-electrode capacitance, and since the two circuits are not tuned to the same frequency, the input impedance of the valve will have either a positive or a negative resistance component. When the R.F. oscillator frequency is above that of the signal, as in most commercial super-heterodynes, the tuned circuit of the oscillator is, in effect, a capacitive load upon the input circuit and the input resistance of the valve is positive, imposing damping upon the input circuit. As a result of this damping, the effective conversion conductance of the pentagrid falls off at frequencies higher than about 2 Mc/s, and that of the octode above about 5 Mc/s. Where the oscillator is set to the lower frequency, the effective conversion conductance increases at the higher frequencies.

The hexode

96. In both heptode and octode the R.F. oscillator control grid is that nearest the cathode, i.e. G_1. The prototype of a second type of frequency-changer is the hexode (fig. 45d), which possesses a cathode, four grids and an anode. In this valve the signal voltage is applied between G_1 and the cathode, G_2 is a perforated electrode which is maintained at a positive potential and serves both as a screen and as an accelerating electrode. The grids G_3 and G_4 are the R.F. oscillator grid and anode respectively. In operation, a virtual cathode is formed on the outside

of G_2, and during reception of a signal its emission will pulsate at signal frequency. This emission will be modulated by the R.F. oscillation and on arrival at the anode the electron stream will possess a component of difference frequency as in the pentagrid. The hexode suffers from the disadvantage that automatic gain control is difficult to achieve. This is because the R.F. oscillator derives its electron supply from the virtual cathode. If a heavy negative bias is applied to the signal grid, the virtual cathode becomes non-existent or at any rate of such a low electron density that the R.F. oscillation is no longer maintained. The effect of the inter-electrode capacitance coupling upon the input impedance of the valve is very much less in the hexode than in the pentagrid or octode. In effect, this is due to the relative potentials of the adjacent electrodes. Taking the octode, we see that only a single screen separates the input grid from the R.F. oscillator anode, the potential of the latter being above that of the screen. In the hexode, the signal grid and R.F. oscillator anode are separated by two electrodes, viz. the screen G_2, at a fairly low positive potential, and the R.F. oscillator grid G_3, which has a mean potential negative to the cathode. A prima-facie examination would therefore lead to the conclusion that the inter-electrode coupling between the tuned oscillator circuit and the input circuit is less than in the octode. Actually, this is found to be the case, although the actual explanation is much more complicated, involving as it does the electro-dynamic effects due to the varying electron density of the virtual cathode.

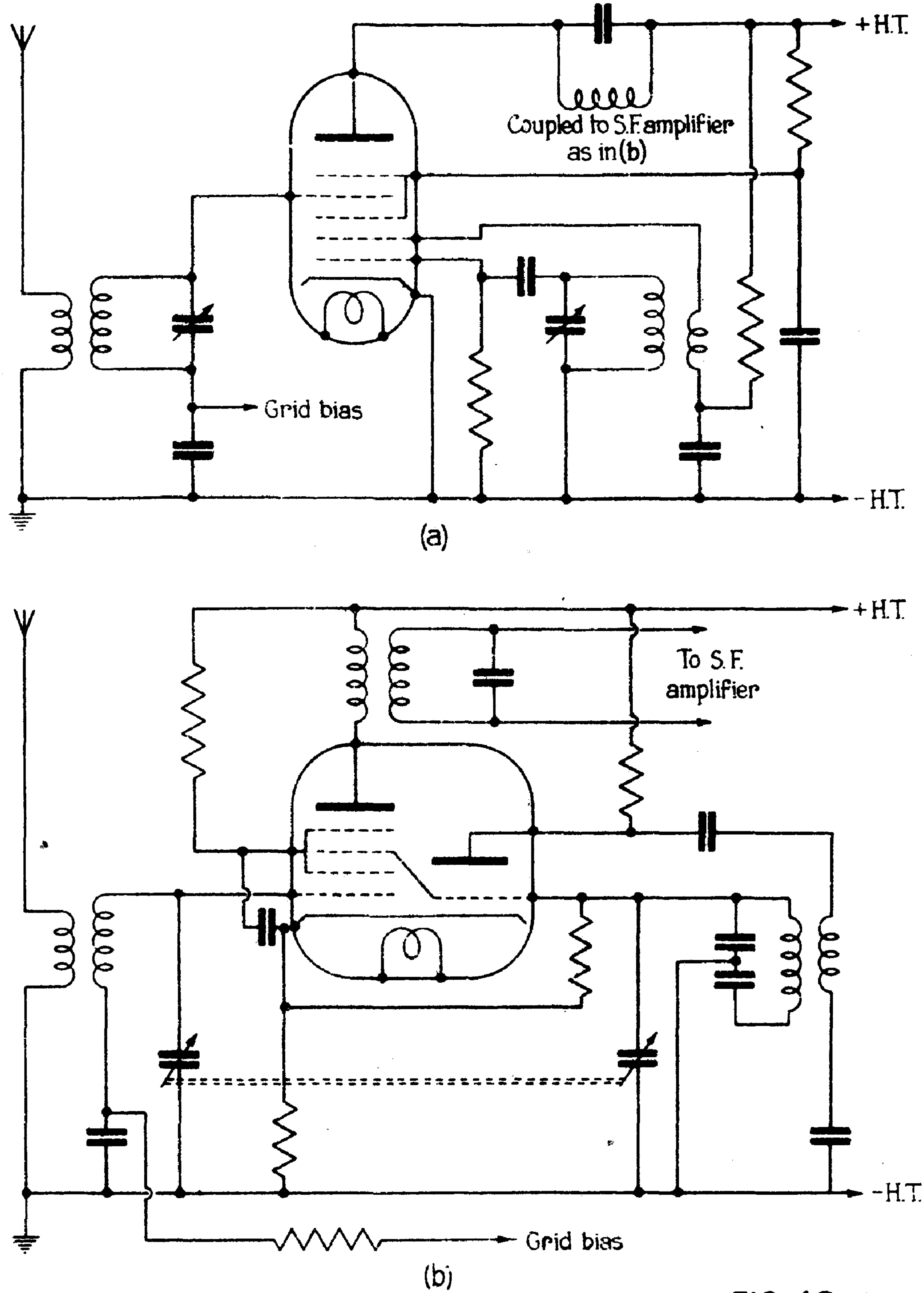
The triode-hexode

97. (i) This valve (fig. 45e) consists of a hexode and a triode in a single envelope, the triode portion operating as an R.F. oscillator. The circuit connections are shown in fig. 46b. In the hexode portion, G_1 is the signal voltage grid, G_2 and G_4 are linked together to form a screening electrode surrounding G_3 which is called the injector grid and is in direct connection with the R.F. oscillator grid. The tuned circuit of the R.F. oscillator is now almost entirely decoupled (both electro-statically and electro-dynamically) from the input circuit, and the input impedance is very little affected by the R.F. oscillator at frequencies below about 100 Mc/s.

(ii) Most of the electron-coupled frequency-changers suffer from a tendency for the R.F. oscillation to pull into step with the signal frequency when the signal grid bias is varied by the automatic gain control. This is again only of importance at high frequencies where the percentage difference between signal and oscillator frequencies is very small. The electro-dynamic properties of the virtual cathode are again involved in a complete explanation, but bearing in mind that the frequency of a valve-maintained oscillation depends to some extent upon the anode A.C. resistance r_a of the valve, the effect is to all intents and purposes due to the variation of the value of r_a with grid bias. It is therefore not encountered in a triode-hexode frequency-changer, in which the electron current of the triode portion is in no way affected by the variation of signal grid bias since this operates only upon the hexode portion.

Choice of R.F. oscillator frequency

98. Except in the case of the single-triode frequency-changer (the input circuit of which is tuned to the signal frequency, plus or minus the predetermined supersonic frequency) the signal frequency circuits are preferably ganged as in the tuned radio-frequency amplifier. Where the receiver is to be operated by unskilled personnel, arrangements are often made to gang the R.F. oscillator tuning also. At first sight this does not appear possible because the oscillator frequency should theoretically differ from the signal frequency by the correct amount, i.e. the supersonic frequency, at all points in the tuning range. It is found, however, that if the minimum capacitance of the R.F. oscillator condenser is set up by means of a small pre-set condenser, in parallel, and its maximum capacitance slightly decreased by the insertion of a fairly large pre-set condenser, in series, the required frequency difference can be maintained at the two ends and at the middle of the tuning range ; elsewhere the actual supersonic frequency will not be quite correct. It will be noticed that the effective frequency ratio between the minimum and the maximum condenser setting will be less than that of the variable condenser alone. If the latter will give a 3 : 1 frequency ratio, the presence of the added pre-set condensers will reduce the effective ratio to only about 2·75 : 1. In practice this method of achieving what is



TYPICAL FREQUENCY CHANGERS

FIG. 46
CHAP. XI

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called correct " tracking " necessitates that the oscillator frequency shall be higher than that of the signal. As an example, let us suppose the supersonic frequency to be 100 kc/s and the desired frequency range, with a given set of coils, to be 550 to 1,500 kc/s. The oscillator must then tune either from 650 to 1,600 or from 450 to 1,400 kc/s, i.e. over a frequency ratio of 2·4 or 3·1 to 1 respectively. The latter figure entails a variation of capacitance of 1 to 10, e.g. from ·00005 to ·0005 μF. The former requires a variation of only 1 to 5·5, e.g. from ·00008 to ·00044 μF. When this method of ganging is adopted, therefore, the R.F. oscillator invariably operates on the higher frequency.

Selectivity—the S.F. amplifier

99. The high degree of selectivity obtainable in a super-heterodyne receiver has led to its adoption for the reception of frequencies of the order of 200 kc/s and below, in addition to the high and very high frequencies for which it was originally intended. This selectivity is chiefly dependent upon the circuits of the S.F. amplifier. Consider the reception of I.C.W. signals on

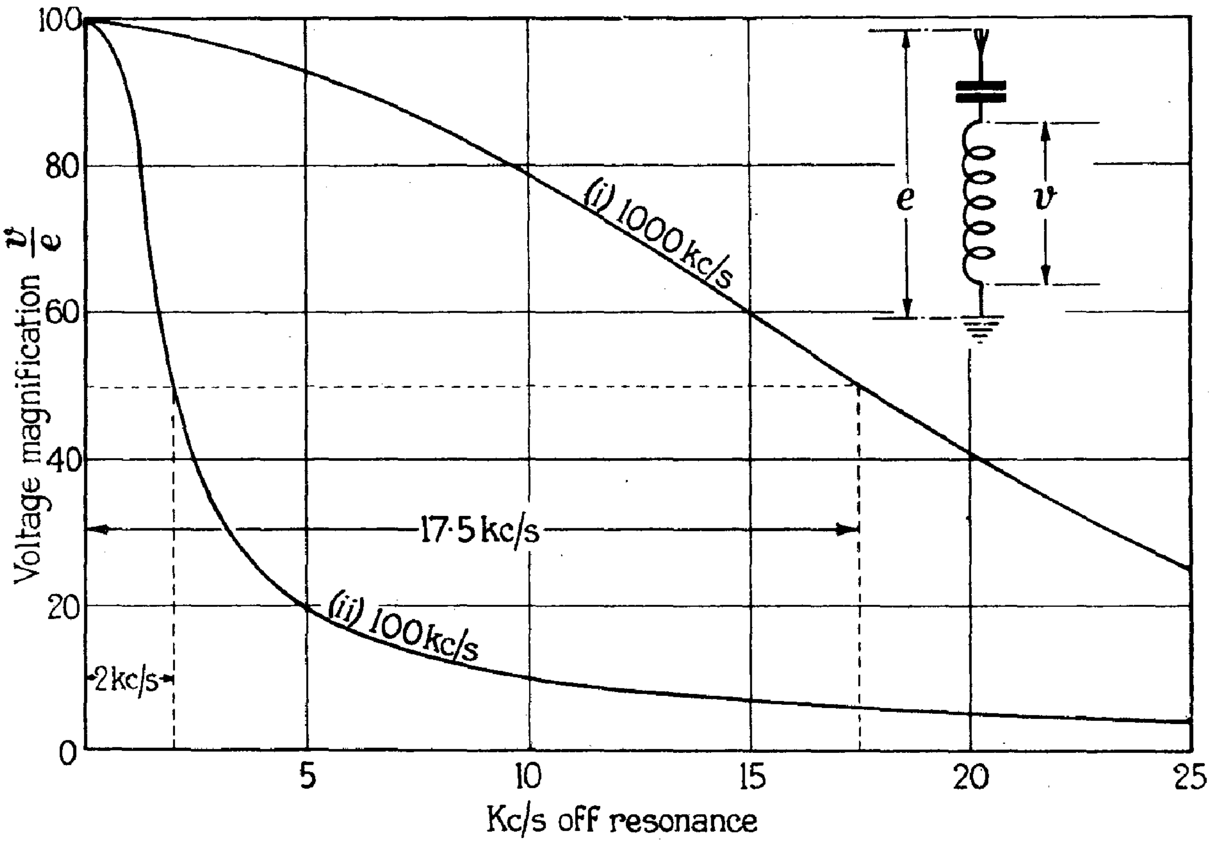


FIG. 47, CHAP. XI.—Comparative selectivity of tuned circuits of equal χ.

1,000 kc/s, first with a simple non-regenerative receiver, the single tuned circuit having a magnification of 100, and the response characteristic given in curve (i) of fig. 47. Assuming equal field strengths, a signal 17·5 kc/s off resonance will give an input voltage to the detector one-half that at resonance and all signals within a band of 35 kc/s may therefore be expected to cause interference. If, however, a similar type of receiver is used on 100 kc/s, and the circuit magnification is the same, namely 100, the response characteristic is that shown in curve (ii), and the same degree of interference is experienced only over a band of 4 kc/s. Now consider the reception of three signals of equal field strength, the frequencies being 990, 1,000 and 1,010 kc/s, by a super-heterodyne having a S.F. of 100 kc/s. The first local oscillator may be adjusted to 1,100 kc/s producing 100,000 beats per second from the 1,000 kc/s signal, and a current of the correct supersonic frequency. This signal will be fully amplified in the S.F. stages. The 990 kc/s signal will give a S.F. current of 1,100 — 990 = 110 kc/s and the 1,010 kc/s signal a S.F. current of 1,100 — 1,010 = 90 kc/s. Both the undesired signals are now 10 kc/s, or 10 per cent. out of resonance and will receive negligible amplification. This form of discrimination is referred to as adjacent channel selectivity.

Types of interference peculiar to super-heterodyne

100. (i) Again, if a supersonic frequency of 50 kc/s is chosen, the local oscillator being adjusted to 1,050 kc/s, the interfering signals will give rise to S.F. currents of 60 and 40 kc/s respectively, which are 20 per cent. out of resonance, and will be attenuated to an even greater extent than in the previous case. From this it would appear that a low supersonic frequency should be chosen. If too low, however, another form of interference is liable to occur ; for example, suppose an interfering signal to have a frequency of 1,050 kc/s. By the above reasoning it should give no heterodyne beat with the local oscillator and produce no S.F. response, but actually it will form heterodyne beats with the desired oscillation, and these give rise to an S.F. of 1,050 — 1,000 = 50 kc/s, which is, of course, fully amplified. As this interference only occurs when both transmitters are radiating, it gives rise to a peculiar " mush " which renders reception extremely difficult. When a high supersonic frequency is chosen, this form of interference is reduced, because the frequency causing it is further removed from that of the desired signal and the preliminary tuned circuits, or pre-selector stages, are competent to deal with it ; e.g. if the S.F. is 200 kc/s, this form of interference would be caused only by a signal on 1,200 kc/s (or 800 kc/s), 16 per cent. out of resonance. Unless its field strength is considerable, such a signal will be considerably attenuated before it reaches the frequency-changer.

(ii) A form of interference peculiar to the super-heterodyne is that caused by the harmonics of undesired signals. This may be shown briefly as follows :—

	Kc/s.		Kc/s.
Desired signal	1,000		
Supersonic frequency	100		
Local oscillator	1,100	or	900
Undesired signal	1,150		
Supersonic current due to undesired signal	50	or	250

Neither of the latter will give appreciable interference. Now consider the local oscillator and undesired signal to possess a considerable second harmonic content.

	Kc/s.		Kc/s.
Local oscillator, second harmonic	2,200	or	1,800
Undesired signal, second harmonic ..	2,300	or	2,300
Supersonic current due to undesired signal	100	or	500

With the local oscillator set to a higher frequency than the desired signal, the second harmonic interference will be fully amplified.

(iii) Yet another form of interference results if two signals, the respective frequencies of which are equal to the desired signal plus and minus one-half the S.F., are simultaneously received. The rectification of this combination obviously results in a current of the supersonic frequency. Other complicated reactions often occur, e.g. between the desired signal frequency and harmonics of the supersonic frequency, if the latter are allowed to be transferred to the circuit preceding the first detector. The remedy in this case is adequate screening of signal, R.F. oscillator, and S.F. circuits.

(iv) Signals of the supersonic frequency may be received in the aerial and transferred to the S.F. amplifier by electro-magnetic or electro-static coupling. This interference is reduced by careful screening and by the provision of a high degree of selectivity in the R.F. stages.

101. To sum up, the choice of an intermediate frequency is a compromise between the following factors :—

(*a*) Given an adequate degree of selectivity in the pre-selector stages, image-frequency interference is reduced as the S.F. is increased. The possibility of interference by two stations, the frequency difference of which is equal to the supersonic frequency, is also reduced by a high S.F.

(*b*) Low-order harmonics of the S.F. should not fall in the frequency band to be received, e.g. a S.F. of 500 would be most unsuitable for reception of 1,000 kc/s.

(*c*) It is not possible to receive signal frequencies equal, or nearly equal to the S.F. By suitable switching arrangements the S.F. stages only may be used as an ordinary (tuned radio-frequency) amplifier for this frequency band.

(*d*) The difficulty of obtaining a high degree of selectivity and amplification increases with an increase of S.F.

The double super-heterodyne

102. (i) A little consideration of the above summary will show that for reception of high radio frequencies of the order of say 20 Mc/s it is difficult or impossible to achieve a high degree of image frequency discrimination by the ordinary super-heterodyne circuit. Thus with a S.F. of 100 kc/s and an oscillator frequency of 20 Mc/s the receiver deals with 20·1 and 19·9 Mc/s

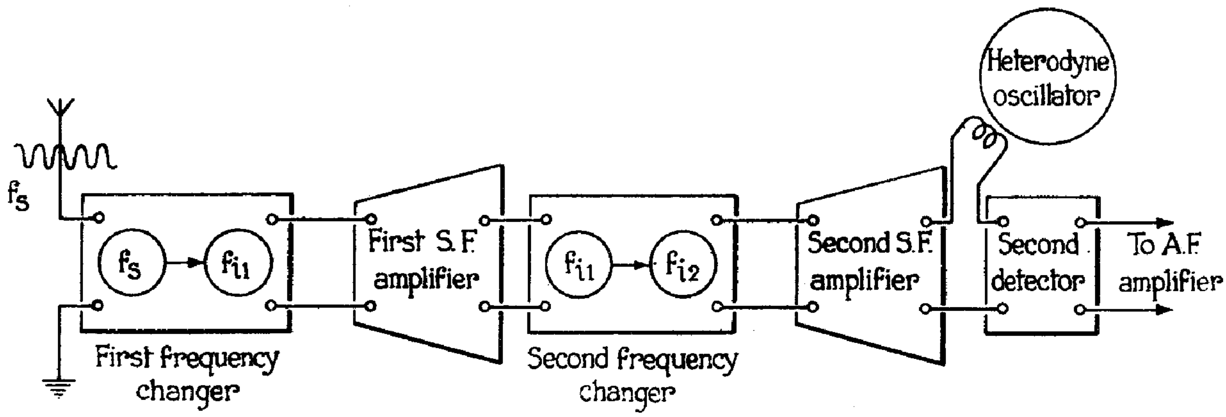


FIG. 48, CHAP. XI.—Super-heterodyne with double frequency change.

equally well, unless the pre-selector stages are capable of giving a very high degree of discrimination against a signal only one per cent. off resonance. If the S.F. is appreciably increased, so that a single pre-selector stage is competent to deal with the image frequency, the adjacent channel selectivity, i.e. that of the intermediate frequency amplifier, will be correspondingly reduced. The difficulty may be overcome by using what is called a double super-heterodyne, the schematic diagram of which is given in fig. 48. This in effect is an ordinary super-heterodyne receiver in which the pre-selector, instead of being a tuned radio-frequency amplifier, is another super-heterodyne receiver with a comparatively high S.F. This S.F. must of course fall within a frequency band of which reception is not required. As an example, suppose it is desired to cover a frequency band of from 20 Mc/s to 150 kc/s except for the region 1,500 to 2,000 kc/s. The first S.F. will therefore naturally fall in the latter band. The lower the frequency the better is the image frequency discrimination. Since however, broadcast transmission may cause direct interference on 1,500 kc/s a little margin must be allowed and 1,700 kc/s is found to be suitable. A single tuned circuit preceding the first frequency-changer will then be found to give adequate discrimination against the image frequency, e.g. for a signal of 20 Mc/s the image frequency and signal frequency differ by $\frac{2 \times 1 \cdot 7}{20} \times 100 = 17$ per cent. Even at such a high frequency as this it is possible to provide an aerial circuit magnification sufficient to attenuate the image frequency by about 12 db. At lower signal frequencies the image frequency attenuation is even better. The second S.F. will be dictated by the same considerations as in the single super-heterodyne, viz. it must give adequate discrimination against adjacent channel interference. Let us assume that the chosen frequency is 100 kc/s. The first S.F. amplifier must have sufficient selectivity to act as a pre-selector to the second frequency-changer. Bearing in mind that no matter what the signal frequency may be, the input to the first S.F. amplifier is at 1,700 kc/s, the second R.F. oscillator may be set to either 1,600 or 1,800 kc/s. If the former is chosen the desired signal, 1,700 kc/s, and any interference on 1,500 kc/s, will be fully amplified by the second S.F. amplifier. The first S.F. amplifier must therefore be sufficiently selective to suppress any 1,500 kc/s interference which it may receive from the first frequency-changer valve. The second S.F. amplifier may be provided with variable selectivity for R/T and W/T reception.

(ii) As might be expected, such a receiver must be very carefully designed with a view to reducing image interference and that due to the signal frequency plus or minus one-half the S.F. There are other problems peculiar to the employment of a double frequency change. The number of possible interfering harmonics is reduced by operating all the oscillators (i.e. first R.F., second R.F., and C.W. heterodyne) at the higher of their two possible frequencies, but in practice it is preferable to operate the second oscillator at the lower one. The reason is as follows. When the receiver is operating on 200 kc/s, the first oscillator is on 1,900 kc/s. It is difficult to decouple the various circuits so thoroughly that no beats occur between first and second oscillators. If such a beat does occur with the second R.F. oscillator on 1,800 kc/s, its frequency is 1,900 — 1,800 = 100 kc/s, which gives rise to a continuous signal in the second S.F. amplifier. With the second oscillator set at 1,600 kc/s this difficulty does not arise.

RADIO-FREQUENCY POWER AMPLIFIERS

103. In most modern transmitting equipment the oscillator operates at a low power level and steps are taken to ensure that the frequency is as nearly constant as possible, having regard to the conditions under which the transmitter is to be employed. The low-power, constant-frequency oscillation is amplified by successive stages until the required power level is reached, the final output circuit being the transmitting aerial with its feeder line and impedance-matching device. In such a transmitter each stage is a power amplifier, and the efficiency of conversion from direct to oscillatory power is of great importance. Where it is possible to provide a sufficiently high anode supply voltage, the power output is limited only by (i) the permissible dissipation, (ii) the filament emission and (iii) the efficiency, of the valve. For high efficiency of power conversion it is necessary to operate in such a manner that anode current flows for not more than one-half the duration of each cycle, and the oscillatory anode-filament P.D. must be nearly equal to the supply voltage. The load impedance of a radio-frequency power amplifier invariably consists of an oscillatory circuit which is tuned to the frequency of the input (grid-filament) voltage.

Angle of current flow

104. It is often convenient to speak of the duration of the anode current impulses in terms of the electrical angle during which anode current flows. In the class B amplifier, the grid is given such negative bias that the anode current, in the absence of an oscillatory grid-filament P.D., is reduced to zero, or very nearly so. The application of an oscillatory P.D. between grid and filament then causes anode current to flow during the positive half-cycles of grid voltage only ; the current is therefore said to flow for 180 degrees. These anode current impulses are, to all intents and purposes, half sine waves, and the operating conditions are very similar to those of the high-efficiency oscillator discussed in Chapter IX, except that the grid bias voltage must be maintained at a constant value irrespective of the magnitude of the grid-filament oscillatory voltage. Bias cannot therefore be obtained by means of a condenser and leak resistance. In the class C amplifier, the grid is biased to beyond the point of anode current

cut-off, and anode current flows for less than 180 degrees. It is important to observe that in the class B amplifier, the electrical angle is the same no matter what the amplitude of the grid-filament voltage may be, whereas in the class C amplifier the electrical angle generally varies with the amplitude of the grid-filament voltage.

Limiting conditions

105. Before proceeding to an approximate derivation of the power output and efficiency of these amplifiers, it is necessary to appreciate certain practical limitations which may be imposed. Possibly the most important of these is the filament emission under average working conditions. In some cases this is given by the valve manufacturer, but the following figures may be taken as a working approximation where precise information is not available. As an average during its useful life a pure tungsten filament may be expected to give from 5 to 10 milliamperes for each watt expended in filament heating. The exact figure to be employed depends upon the size—and therefore the cost—of the valve. A small and comparatively inexpensive valve may be run with a higher emission than a large one, for although its life will be shorter the cost of replacement is much less. The following empirical rule may be taken as a rough guide :—

$$I_e = P_F\,(10 - \cdot 005\, P_L).$$

where I_e is the filament emission in milliamperes.

P_L is the permissible dissipation in watts.

P_F is the power expended in filament heating, in watts.

According to this rule, a 30-watt valve will give nearly 10 milliamperes per "filament watt", while a 1,000-watt valve will give only 5 milliamperes per "filament watt." The thoriated tungsten filament, such as is used in the V.T. 25 valve, will give a momentary or "flash" emission of from 20 to 40 milliamperes per "filament watt", but in operation the anode current must not be allowed even to approach the flash emission value, otherwise the valve will deteriorate rapidly. It may therefore be assumed that such a filament will give about 15 to 20 milliamperes per "filament watt." Coated filaments are employed in certain power amplifying valves of ratings up to about 50 watts. Their working emission is about 60 to 80 milliamperes per filament watt. A second limitation which arises is the magnitude of the supply voltage, which may be restricted by the compactness of a given installation, and its consequent small margin of insulation resistance, or by the weight and volume of the H.T. generator or other supply device. The maximum supply voltage for a particular valve is generally given on the maker's label, and should not be exceeded.

Class B amplifier

106. The behaviour of a class B amplifier may be approximately determined by a consideration of the current and voltage relations during operation. It is always necessary to make certain assumptions, and the different conclusions reached by various writers are chiefly due to differences in the stipulated conditions. It will be convenient to commence by assuming that the valve possesses ideal characteristics, the notation used being as follows.

I_e = filament emission, milliamperes.
I_p = peak value of anode current impulse.
I_{av} = average anode current = D.C. component.
I_{gp} = peak value of grid current impulse.
$\mathcal{I}_a$ = amplitude of the fundamental component of anode current.
$\mathcal{V}_a$ = amplitude of oscillatory P.D. across load impedance.
E_a = anode supply voltage.
E_b = magnitude of grid bias voltage.
E_g = maximum permissible positive grid voltage.
E_o = minimum anode-filament P.D.

$\mathcal{V}_g$ = amplitude of grid excitation voltage.
R_d = dynamic resistance of load.
R_e = virtual load resistance $= \frac{R_d}{2}$.
μ = amplification factor of valve.
r_a = anode A.C. resistance of valve.
P_i = power input.
P_o = power output.
P_L = permissible dissipation of valve.
P_d = power actually dissipated.
P_g = grid driving power.

107. As already stated (Chapter IX) if I_p is the peak value of the semi-sinusoidal anode current impulse, its average value is $\frac{I_p}{\pi}$, and the amplitude of the fundamental (input frequency) oscillatory component will be $\frac{I_p}{2}$. The amplitudes of the respective harmonics are comparatively small and are of little importance since the load impedance is a resonant circuit and is tuned to the input frequency. The fundamental component of the anode current is therefore the only one which can set up a considerable P.D. across the load impedance, and this P.D. is to all intents and purposes sinusoidal in spite of the impulsive nature of the anode current. As the oscillatory current and P.D. are proportional to the grid-filament voltage, the amplifier is said to be linear. The anode-filament P.D. is the difference between the supply voltage E_a and the instantaneous P.D. across the load. At the instant when the anode current reaches its peak value I_p, the anode-filament P.D. is $E_o = E_a - \mathcal{V}_a$. The oscillatory voltage acting in the anode circuit at this instant is $\mu \mathcal{V}_g - \mathcal{I}_a R_d$ and the anode current reaches the value

$$I_p = \frac{\mu \mathcal{V}_g - \mathcal{I}_a R_d}{r_a}$$

Since $\mathcal{I}_a = \frac{I_p}{2}$,

$$I_p r_a = \mu \mathcal{V}_g - \frac{I_p R_d}{2}$$

$$= \mu \mathcal{V}_g - I_p R_e$$

$$I_p = \frac{\mu \mathcal{V}_g}{r_a + R_e}$$

The following special cases will now be considered :—

(i) The grid is not to be allowed to become positive with respect to the filament at any point in the cycle.

(ii) The grid is allowed to become positive with respect to the filament, but the effect of the resulting grid current upon the output and efficiency will be neglected.

Power relations without grid current

108. The conditions under (i) above are applicable when the power amplifier immediately follows a valve-driven master-oscillator (i.e. one without crystal or tuning fork control). If grid current is allowed to flow, the effective resistance of the " master " oscillatory circuit will vary during each cycle and will cause frequency variation. Even if the frequency is stabilized by

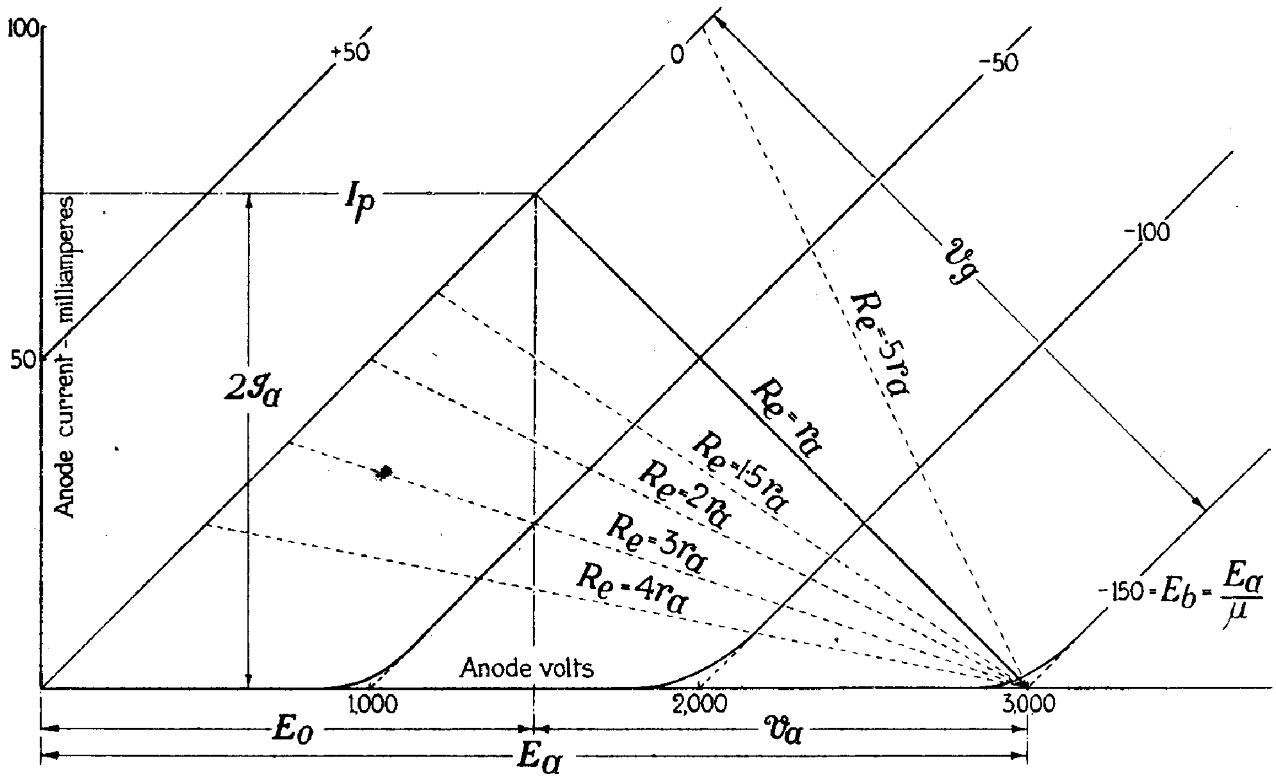


FIG. 49, CHAP. XI.—Class B amplifier; operation without grid current.

some form of mechanical oscillator this variation of load is to be avoided if possible. Referring now to fig. 49, the operating conditions are found as follows. The grid bias voltage is $-E_b = -\frac{E_a}{\mu}$, and the peak excitation will be numerically equal to the grid bias, i.e., $\mathcal{V}_g = \frac{E_a}{\mu}$.

Then

$$I_p = \frac{E_a}{r_a + R_d}$$

$$\mathcal{J}_a = \frac{E_a}{2\, r_a + R_a}$$

and the power delivered to the load is

$$P_o = \frac{\mathcal{J}_a^2 R_d}{2} = \frac{E_a^2 R_d}{2\,(2\, r_a + R_d)^2}$$

$$= \frac{E_a^2 R_e}{4\,(r_a + R_e)^2}$$

This is a maximum when $R_e = r_a$ and the maximum possible power output is $\frac{E_a^2}{16 r_a}$ watts. The power input P_i is equal to the product of the supply voltage and the average anode current.

The latter is $\frac{I_p}{\pi}$ so that

$$P_i = \frac{E_a I_p}{\pi}$$

$$= \frac{E_a^2}{\pi (r_a + R_e)}$$

and the efficiency η is

$$\frac{P_o}{P_i} = \frac{R_e E_a^2}{4 (r_a + R_e)^2} \times \frac{\pi (r_a + R_e)}{E_a^2}$$

$$= \frac{\pi}{4} \frac{R_e}{r_a + R_e}.$$

Thus the efficiency improves as the load impedance is increased, approaching the limiting value, 78·54 per cent., when R_e is very much larger than r_a. When, as in the present instance, the $I_a - V_a$ characteristics of the valve are available, the output and efficiency for any given load can be readily computed directly from the curves. In fig. 49 assuming that E_a (*max.*) $= 3{,}000$ volts, a load line representing a virtual resistance $R_e = r_a$ has been drawn in solid line, intersecting the curve $V_g = 0$ at $I_a = 75$, $V_a = 1{,}500$. Hence $I_p = 75$ milliamperes and $\mathcal{V}_a = 1{,}500$ volts. The input, output and efficiency may now be calculated directly by means of the relations

$$P_o = \frac{\mathcal{V}_a I_p}{4}$$

$$P_i = \frac{E_a I_p}{\pi}$$

$$\eta = \frac{P_o}{P_i} \times 100, \text{ per cent.}$$

Other load lines have also been drawn in dotted line, and the results derived from these are plotted in fig. 50. It is seen that the output rises to a maximum when $R_e = r_a$, but is very little reduced if $R_e = 2\,r_a$, while the efficiency increases continuously towards its limiting value as R_e increases. For many purposes it is desirable to aim at a maximum value of the product ηP_o which gives the best compromise between output and efficiency. This quantity has also been plotted in fig. 50 and it is seen that a value of R_e in the region of from $1{\cdot}5\,r_a$ to $2{\cdot}5\,r_a$ satisfies this requirement.

Power when grid current is permitted

109. The output can be considerably increased by permitting grid current to flow, although some distortion of the anode current wave-form will be introduced. For the present, we shall continue to assume that the valve possesses ideal characteristics. Maximum output will obviously be obtained when both the anode current and the load P.D. have the greatest possible amount of variation. A further limitation must however be introduced, in that the anode-filament P.D. must never be allowed to fall below the instantaneous grid-filament P.D., i.e. the grid must not be allowed to become positive with respect to the anode. If this limitation is not observed, the secondary electrons emitted by the anode will be attracted to the grid and the grid current will be excessive. In order to allow for a reasonable amount of grid current, it may be taken that the peak anode current I_p must not exceed 80 per cent. of the filament emission. Even if the grid current reaches an instantaneous value of $\cdot 2\ I_p$ the filament will then be able to supply the total current demanded by both electrodes. This total current is usually referred to as the space current. Referring now to fig. 51, the chain-dotted line has been drawn to indicate the positive limit of the grid swing—and therefore the limiting value of the minimum anode-filament P.D.—from the following considerations. Suppose the

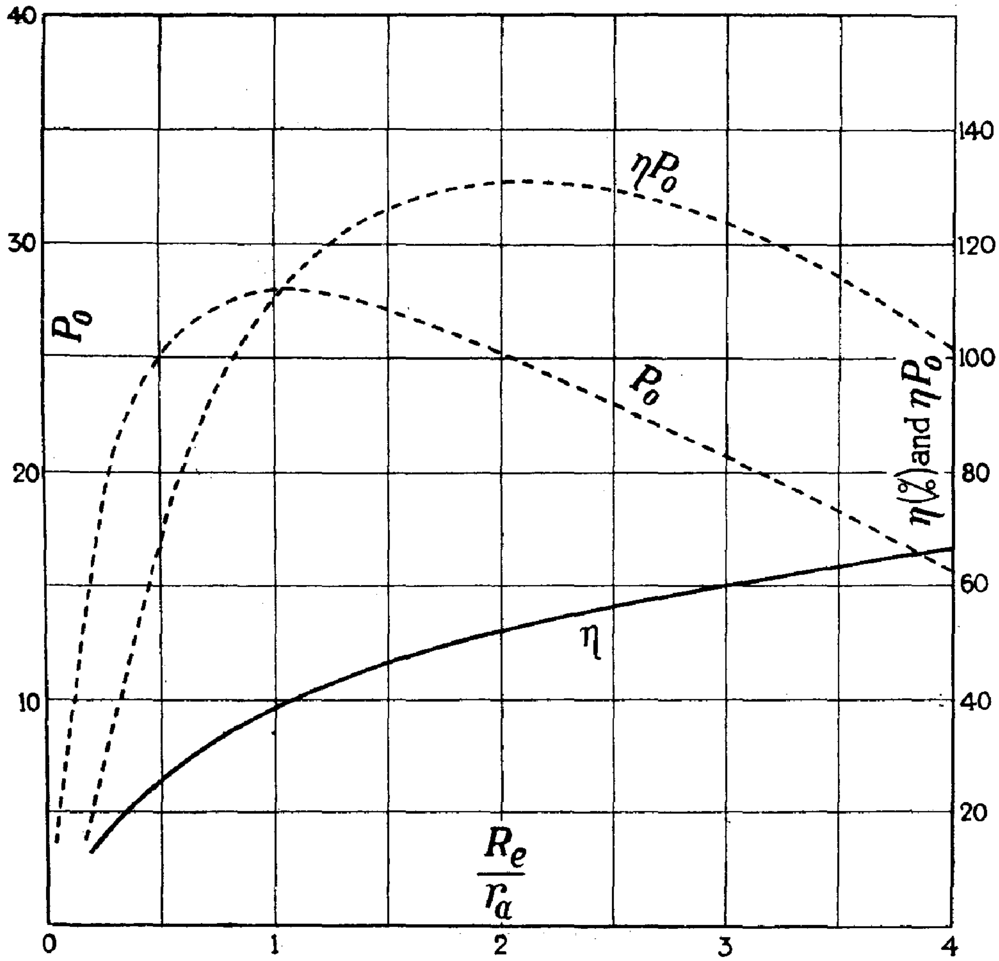


FIG. 50, CHAP. XI.—Output and efficiency of class B R.F. amplifier, for various ratios of $\frac{R_e}{r_a}$ and without allowing grid to become positive.

grid to be 200 volts positive with respect to the filament, then the anode must also be at least 200 volts positive, thus locating the point a. Similarly, when the grid is 100 volts positive, the anode filament P.D. must not be less than 100 volts, locating the point b. The chain-dotted line is drawn through these points. It is easily seen that for a given supply voltage E_a the greatest excursions of anode current and load P.D. will be achieved by locating the peak anode current I_p at its maximum permissible value, viz. ·8 I_e and operating on the load line corresponding to a virtual resistance

$$R_e = \frac{E_a - E_o}{I_p}$$

The value of E_o follows from the equation to the chain-dotted line. With the ideal characteristics postulated, the relation between I_a, V_a, and V_g is

$$I_a r_a = V_a + \mu V_g$$

When $V_a = E_o$, V_g is also equal to E_o and $I_a = I_p$, so that we may write

$$I_p r_a = E_o + \mu E_o$$

$$E_o = \frac{r_a}{\mu + 1} I_p,$$

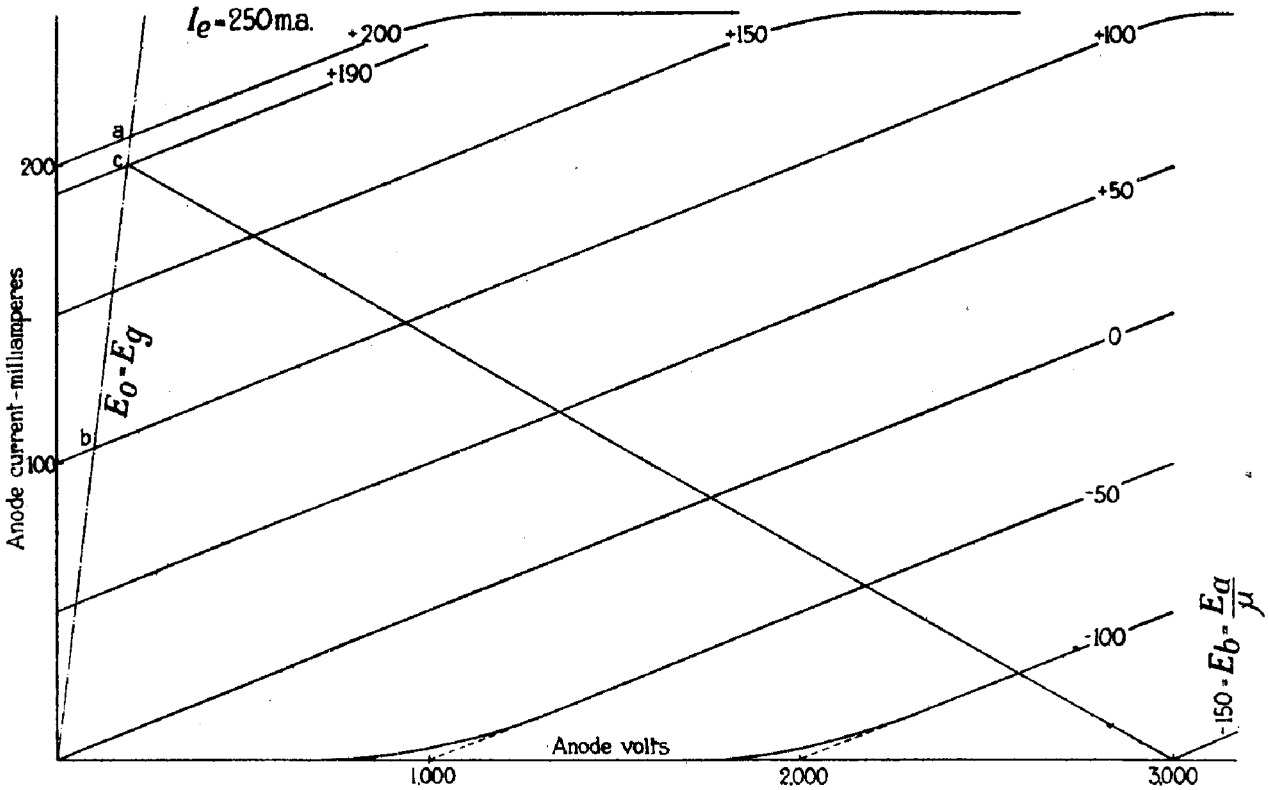


FIG. 51, CHAP. XI.—Class B amplifier; limiting E_o line when grid current is permitted.

and the particular value of virtual load resistance corresponding to the load line is therefore

$$R_e = \frac{E_a}{I_p} - \frac{r_a}{\mu + 1}$$

In fig. 51, I_p = 200 milliamperes, E_o = 190 volts.

$$R_e = \frac{3{,}000 - 190}{\cdot 2}$$
$$= 14{,}050 \text{ ohms.}$$

The output power is

$$P_o = \frac{E_a - E_o}{4} I_p$$
$$= \frac{E_a I_p}{4} - \frac{I_p^2 r_a}{4(\mu + 1)},$$

and in the particular example

$$P_o = \frac{3{,}000 \times \cdot 2}{4} - \frac{\cdot 2 \times \cdot 2 \times 20{,}000}{4 \times 21}$$
$$= 150 - 9\cdot 5$$
$$= 140\cdot 5 \text{ watts.}$$

This may be verified directly from fig. 51, in which $E_a - E_o = 2{,}810$ volts, so that $P_o = \frac{2{,}810 \times \cdot 2}{4} = 140\cdot 5$ watts as calculated.

The input power is

$$P_o = \frac{E_a I_p}{\pi}$$

$$= \frac{3{,}000 \times \cdot 2}{\pi}$$

$$= 191 \text{ watts,}$$

and the efficiency 73·5 per cent. The actual power dissipated is only 50·5 watts.

The grid excitation is

$$\mathscr{V}_g = \frac{E_a}{\mu} + E_g, \text{ and } E_g = E_o$$

$$\therefore \mathscr{V}_g = \frac{E_a}{\mu} + \frac{I_p r_a}{\mu + 1}.$$

In the example, $\frac{E_a}{\mu} = 150$, $\frac{I_p r_a}{\mu + 1} = 190$, and therefore $\mathscr{V}_g = 340$ volts. This again may be verified in Fig. 51, from which it is seen that the grid must reach + 190 volts from a bias of − 150 volts, so that the peak value of the oscillatory grid voltage must be 190 + 150 = 340 volts as calculated. Finally, if the actual radio-frequency resistance R of the output circuit is known, the circulating current $\mathscr{I}_o$ in this circuit is found from the relation

$$P_o = \mathscr{I}_o^2 R$$

$$\text{or } \mathscr{I}_o = \sqrt{\frac{P_o}{R}}.$$

Operation with limited power input

110. In certain circumstances it may be necessary to restrict the input power, e.g. to an amount not exceeding P_L, the permissible dissipation of the valve. Unless this condition is fulfilled a failure in the output circuit of the amplifier may result in damage to the valve. Provided that a sufficient grid swing is available, the greatest output and highest efficiency is obtained by operating with the maximum permissible anode supply voltage. The peak anode current is then given by the equation

$$I_p = \frac{\pi P_L}{E_a}$$

and the power output by

$$P_o = \frac{(E_a - E_o)\, I_p}{4}$$

$$= \frac{E_a I_p}{4} - \frac{I_p^2 r_a}{4(\mu + 1)}$$

$$= \frac{\pi}{4} P_L - \frac{r_a}{4(\mu + 1)} \times \frac{\pi^2 P_L^2}{E_a^2}.$$

The latter expression gives the power output in terms of the supply voltage and valve data and may therefore be used for calculation when the actual characteristics are not available.

For example, taking the valve previously used for illustrative purposes, suppose that P_l is not to exceed P_L, i.e. 100 watts. If $E_a = 3{,}000$ volts as before,

$$I_p = \frac{\pi \times 100}{3{,}000}$$
$$= 104{\cdot}6 \text{ milliamperes.}$$
$$P_o = \frac{314}{4} - \frac{20{,}000}{84} \times \frac{\pi^2 \times 10^4}{9 \times 10^6}$$
$$= 78{\cdot}5 - 2{\cdot}6$$
$$= 75{\cdot}9 \text{ watts}$$
$$E_o = \frac{I_p r_a}{\mu + 1}$$
$$= 100 \text{ volts.}$$

The virtual load resistance is

$$R_e = \frac{\mathcal{V}_a}{I_p}$$
$$= \frac{3{,}000 - 100}{104{\cdot}6}$$
$$= 27{,}700 \text{ ohms.}$$

To obtain this output, we must provide a grid excitation

$$\mathcal{V}_g = \frac{E_a}{\mu} + E_o$$
$$= 150 + 100$$
$$= 250 \text{ volts.}$$

Operation when limited by permissible dissipation

111. In some high-power ground station installations, arrangements are made for the valve or valves to be put out of action by an automatic device in the event of any failure which would cause a dangerous increase of anode current. The valves may then be operated in such a manner that they dissipate the power corresponding to their actual rating, although this is generally only possible if the valves and circuits are capable of withstanding a very high voltage. It is necessary to ensure that the anode is always at a potential considerably above that of the grid. Fig. 52 gives the $I_a - V_a$ characteristics of a 1,000-watt valve having an emission current of 1,200 milliamperes, a μ of 50 and an r_a of 20,000 ohms. The thin solid line O a represents the theoretical lower limit of the anode voltage swing, assuming that E_o must be not less than 5 E_g. For ideal characteristics,

$$I_a r_a = V_a + \mu V_g$$

and when $I_a = I_p$, $V_a = E_o$ and $V_g = E_g = \frac{E_o}{5}$.

$$\therefore \quad I_p = \frac{E_o}{r_a}\left(1 + \frac{\mu}{5}\right).$$

In general, if it is stipulated that $E_g = \frac{E_o}{n}$, $I_p = \frac{E_o}{r_a}\left(1 + \frac{\mu}{n}\right)$.

The actual characteristics are considerably curved in the region of low anode voltage, and the actual limiting line is that drawn in chain-dotted line. It is a close approximation to the theoretical one and the above equation may be used for practical calculation with negligible error. The greatest output will again be obtained with the highest permissible value of I_p. The dissipation is equal to the difference between the power input and output, i.e.

$$P_d = \frac{E_a I_p}{\pi} - \frac{\mathcal{V}_a I_p}{4}$$

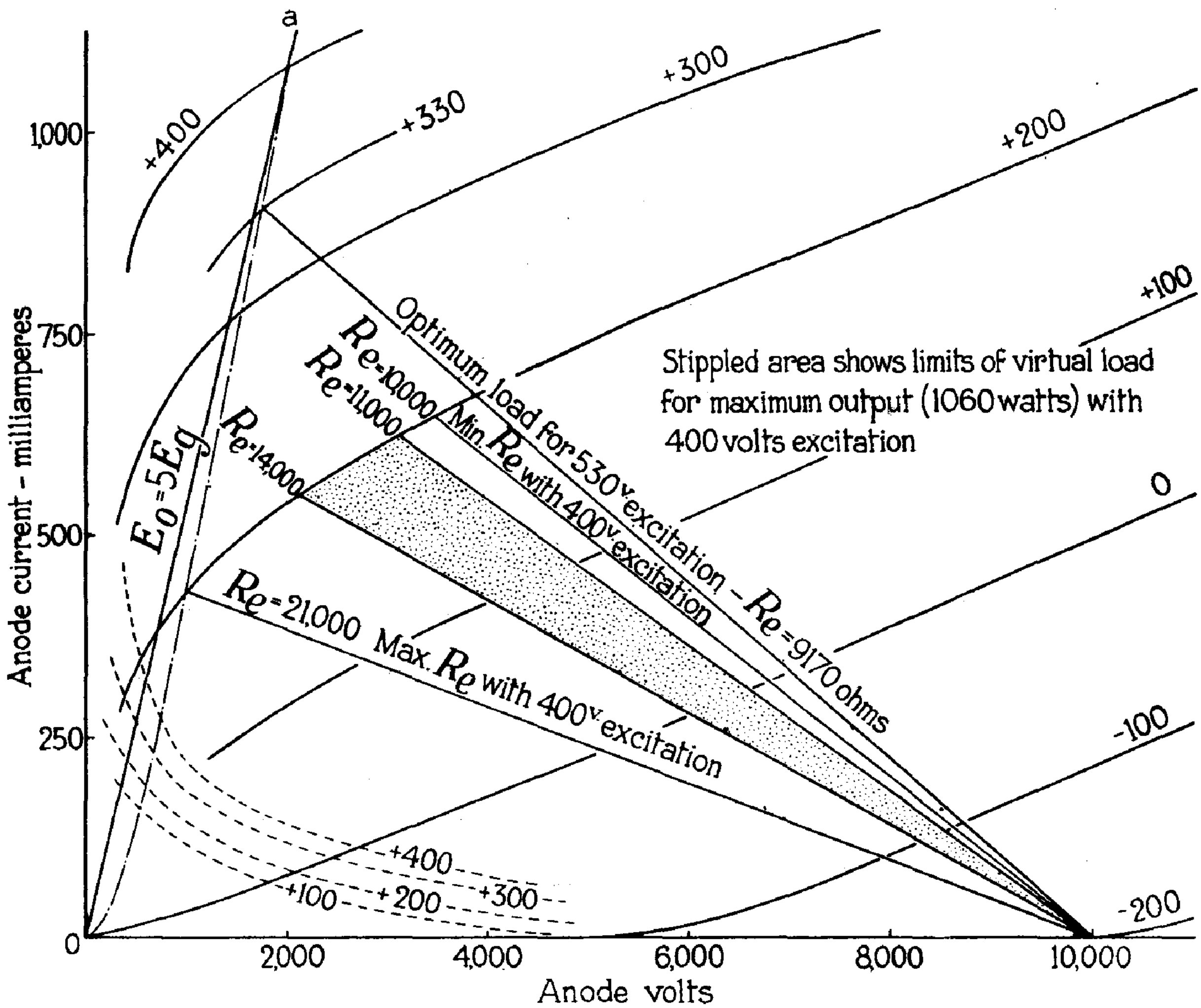


FIG. 52, CHAP. XI.—Operation of class B amplifier under different limiting conditions

If the valve is to dissipate its maximum power, $P_d = P_L$, while as it is also required to give maximum output, the virtual load will be approximately equal to r_a. Since $E_a = \mathcal{V}_a + E_o$,

$$\mathcal{V}_a I_p \left(\frac{1}{\pi} - \frac{1}{4}\right) + \frac{I_p^2 \, r_a}{\pi\left(1 + \frac{\mu}{n}\right)} = P_L$$

$$\frac{1}{\pi} - \frac{1}{4} = \cdot 0684,$$

$$\cdot 0684 \, \frac{\mathcal{V}_a}{I_p} + \frac{r_a}{\pi\left(1 + \frac{\mu}{n}\right)} = \frac{P_L}{I_p^2}$$

$$\frac{\mathcal{V}_a}{I_p} = R_e = r_a$$

$$I_p = \sqrt{\frac{P_L}{\left\{\cdot 0684 + \frac{1}{\pi\left(1 + \frac{\mu}{n}\right)}\right\} r_a}}$$

With the valve under consideration, $\mu = 50$, $r_a = 20{,}000$, $P_L = 1{,}000$ and we have decided upon $n = 5$.

$$\therefore I_p = \sqrt{\frac{1{,}000}{(\cdot 0684 + \cdot 028)\ 20{,}000}}$$

$$= \cdot 716 \text{ ampere.}$$

The operating conditions are therefore as follows:—

$$\mathcal{V}_a = I_p R_e = \cdot 716 \times 20{,}000$$

$$= 14{,}320 \text{ volts.}$$

$$E_o = \frac{I_p r_a}{1 + \frac{\mu}{n}} = \frac{14{,}320}{11}$$

$$= 1{,}300 \text{ volts.}$$

$$E_a = 15{,}620 \text{ volts.}$$

If such a high supply voltage is actually employed, the insulation of the valve, and of certain portions of the circuit, will be called upon to withstand a potential difference of $E_a + \mathcal{V}_a = 29{,}940$ volts. If this is permissible, the power input will be

$$P_i = \frac{E_a I_p}{\pi} = \frac{15{,}620 \times \cdot 716}{\pi}$$

$$= 3{,}550 \text{ watts,}$$

and the power output

$$P_o = \frac{\mathcal{V}_a I_p}{4} = \frac{14{,}320 \times \cdot 716}{4}$$

$$= 2{,}560 \text{ watts.}$$

The actual dissipation is therefore

$$P_d = 990 \text{ watts}$$

and the efficiency

$$\eta = \frac{P_o}{P_i} = 72 \text{ per cent.}$$

The required grid excitation is

$$\mathcal{V}_g = \frac{E_o}{n} + \frac{E_a}{\mu}$$

$$= 260 + 332 \cdot 5$$

$$= 592 \cdot 5 \text{ volts.}$$

112. Actually, the maximum permissible anode supply voltage for the valve having the characteristics of fig. 52 is only 10,000 volts. The permissible peak current may be derived from the relation previously used, viz.,

$$P_L = \frac{E_a I_p}{\pi} - \frac{\mathcal{V}_a I_p}{4}$$

or

$$E_a I_p \left(\frac{1}{\pi} - \frac{1}{4}\right) = P_L - \frac{I_p^2 r_a}{4\left(1 + \frac{\mu}{n}\right)}$$

The permissible peak anode current is found by inserting the appropriate values of E_a, r_a, P_L, and n. With the valve of fig. 52, if $E_a = 10{,}000$ volts, we have

$$\cdot 0684 \times 10{,}000\, I_p = 1{,}000 - 455\, I_p^2$$

$$455\, I_p^2 + 684\, I_p - 1{,}000 = 0$$

$$I_p = 910 \text{ milliamperes.}$$

$$E_o = \frac{\cdot 91 \times 20{,}000}{11}$$

$$= 1{,}650 \text{ volts.}$$

$$\mathscr{V}_a = 10{,}000 - 1{,}650$$

$$= 8{,}350 \text{ volts}$$

$$R_e = \frac{\mathscr{V}_a}{I_p}$$

$$= \frac{8{,}350}{\cdot 91}$$

$$= 9{,}170 \text{ ohms.}$$

$$P_i = \frac{\cdot 91 \times 10{,}000}{\pi}$$

$$= 2{,}900 \text{ watts}$$

$$P_o = \frac{\cdot 91 \times 8{,}350}{4}$$

$$= 1{,}900 \text{ watts}$$

$$P_d = 2{,}900 - 1{,}900 = 1{,}000 \text{ watts} = P_L$$

$$\eta = \frac{1{,}900}{2{,}900}$$

$$= 65 \cdot 5 \text{ per cent.}$$

$$\mathscr{V}_g = \frac{E_o}{n} + \frac{E_a}{\mu}$$

$$= 330 + 200$$

$$= 530 \text{ volts.}$$

Grid driving power

113. The curves shown in the bottom left-hand corner of fig. 52 are the $I_g - V_a$ characteristics for different values of V_g. With their aid it is possible to form an estimate of the power which must be expended in driving the grid positive as in the previous example. It will be seen that at the instant when the anode current reaches the value I_p the grid current also reaches its peak value I_{gp}, about 158 milliamperes. The average grid current may be taken as $\frac{I_{gp}}{2}$, and the grid driving power is approximately equal to one-half the product of the average grid current and the peak excitation, i.e.

$$P_g = \frac{I_{gp}\, \mathscr{V}_g}{4}$$

CHAPTER XI.—PARA. 114

Under the operating conditions of the previous paragraph,

$$P_g = \frac{\cdot 158 \times 530}{4}$$

$$= 21 \text{ watts.}$$

It will now be seen that one reason for placing a fairly high limiting value upon E_o, i.e. $E_o = 4\ E_g$ or $5\ E_g$ rather than $E_o = E_g$, is to reduce the power required for the grid drive. Thus, suppose that the valve of fig. 52 is operated under the latter conditions, with such a supply voltage and anode load that $I_p = 875$ milliamperes when $E_g = E_o = 400$ volts. The grid will at the same instant call for 400 milliamperes, so that the total space current should be 1,275 milliamperes. The filament is unable to supply this, but will give only $\frac{875}{1{,}275} \times 1{,}200 = 824$ milliamperes to the anode circuit and $\frac{400}{1{,}275} \times 1{,}200 = 376$ milliamperes to the grid. The power output will therefore be less than that calculated without taking the grid current into account, while the grid driving power will have to be considerably increased, i.e. if the grid excitation is 530 volts as before, to 50 watts.

Output with limited excitation

114. If both the supply voltage and the available excitation are limited, the load resistance for maximum output may be considerably less than r_a, and its value somewhat critical. This is due to the curvature of the characteristics and is most easily appreciated by laying down a set

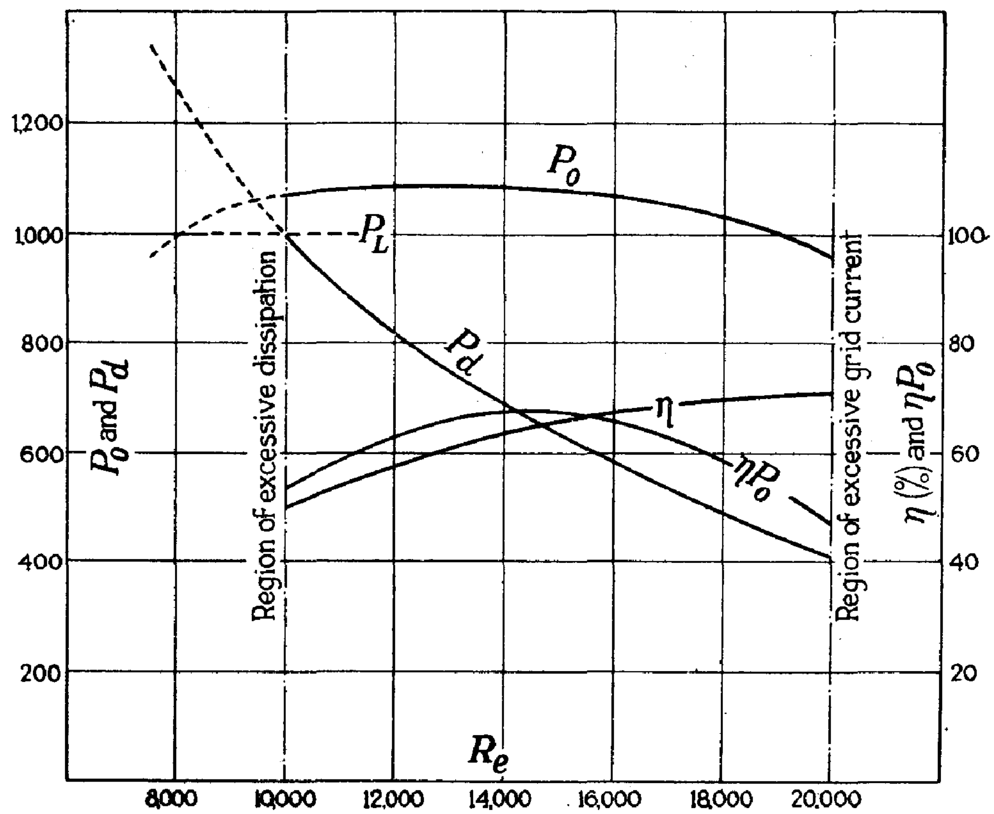


FIG. 53, CHAP. XI.—Output and efficiency of class B R.F. amplifier, with limited grid excitation, for various values of virtual load resistance.

of conditions and calculating the output with different values of R_e. Taking the valve characteristics of fig. 52 let $E_a = 10{,}000$ volts, $E_g \not> \frac{E_o}{5}$ and suppose that $\mathcal{V}_g$ cannot exceed 400 volts. Then one end of the virtual load line will be located at $I_a = 0$, $V_a = E_a = 10{,}000$, and the other will lie somewhere on the curve $V_g = +\ 200$. Owing to the curvature in the region of low anode voltage, it is impossible to operate with a virtual load greater than r_a (approximately) unless either the excitation is reduced or the grid allowed to become more positive than $\frac{E_o}{5}$. On the other hand if a comparatively low virtual load is adopted, i.e. less than $\frac{r_a}{2}$, the dissipation becomes excessive and the output again falls off. The optimum virtual load is approximately equal to the anode A.C. resistance of the valve, measured at the point where the load line reaches the $I_a - V_a$ curve corresponding to the positive limit of grid voltage, but this point can only be located by trial and error. The manner in which the power output varies with the load resistance, for a peak excitation of 400 volts and an anode voltage of 10,000 volts, is shown in fig. 53. It is seen that if R_e falls below 10,000 ohms the permissible dissipation will be exceeded, that values of R_e between 10,000 and 16,000 ohms give substantially the same output, and that the product ηP_o is a maximum when R_e is about 15,000 ohms. On the whole, it appears desirable to operate under the latter conditions, since the actual power dissipated in the valve is then much below the permissible value.

Anode current wave-form

115. It has hitherto been assumed that in a class B amplifier the wave-form of the anode current impulse is truly a half sine wave, but this is seldom if ever so. The degree of departure depends upon the amount by which the actual valve characteristics differ from those of an ideal valve having the same constants. The wave-form of the impulse set up by a sinusoidal excitation voltage can be derived from the $I_a - V_a$ characteristics in the following manner. Referring to fig. 54, in which the curves from $V_g = +\ 5$ to $V_g = -\ 3$ are given, the appropriate virtual load line is first drawn, and this is then divided into a number of parts equal to the number of curves, i.e. eight. From the point 0, the curves are lettered A B C . . . . H, and the equal divisions on the load line a b c . . . . h. If all eight curves from $V_g = -\ 2$ to $V_g = +\ 5$ were equally spaced, a sinusoidal excitation voltage would set up a semi-sinusoidal anode current

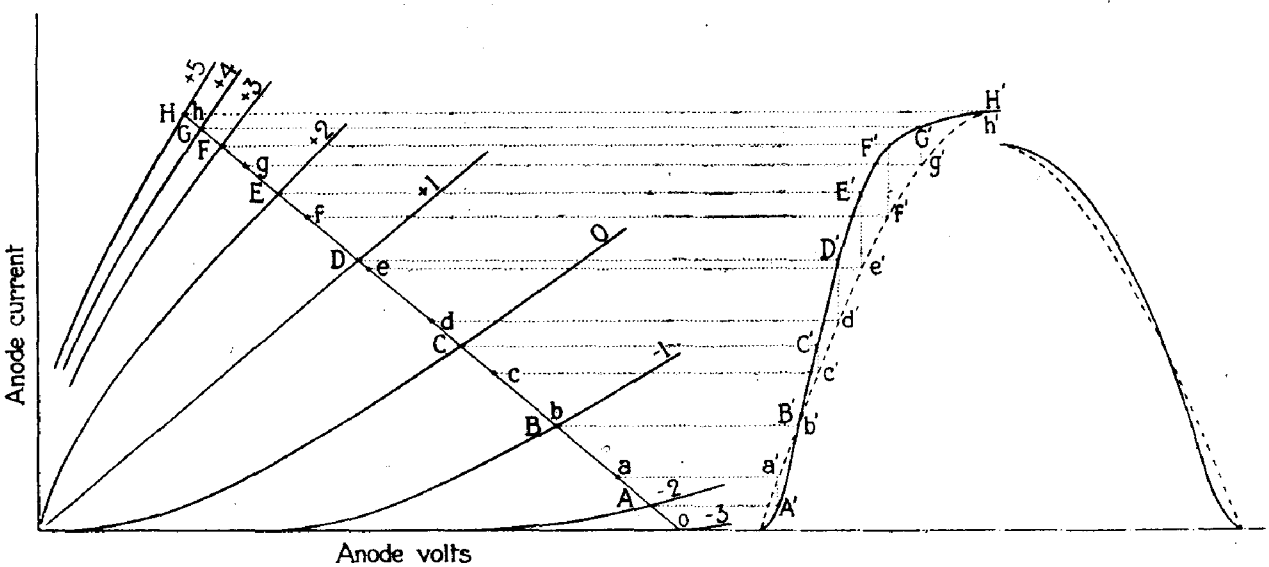
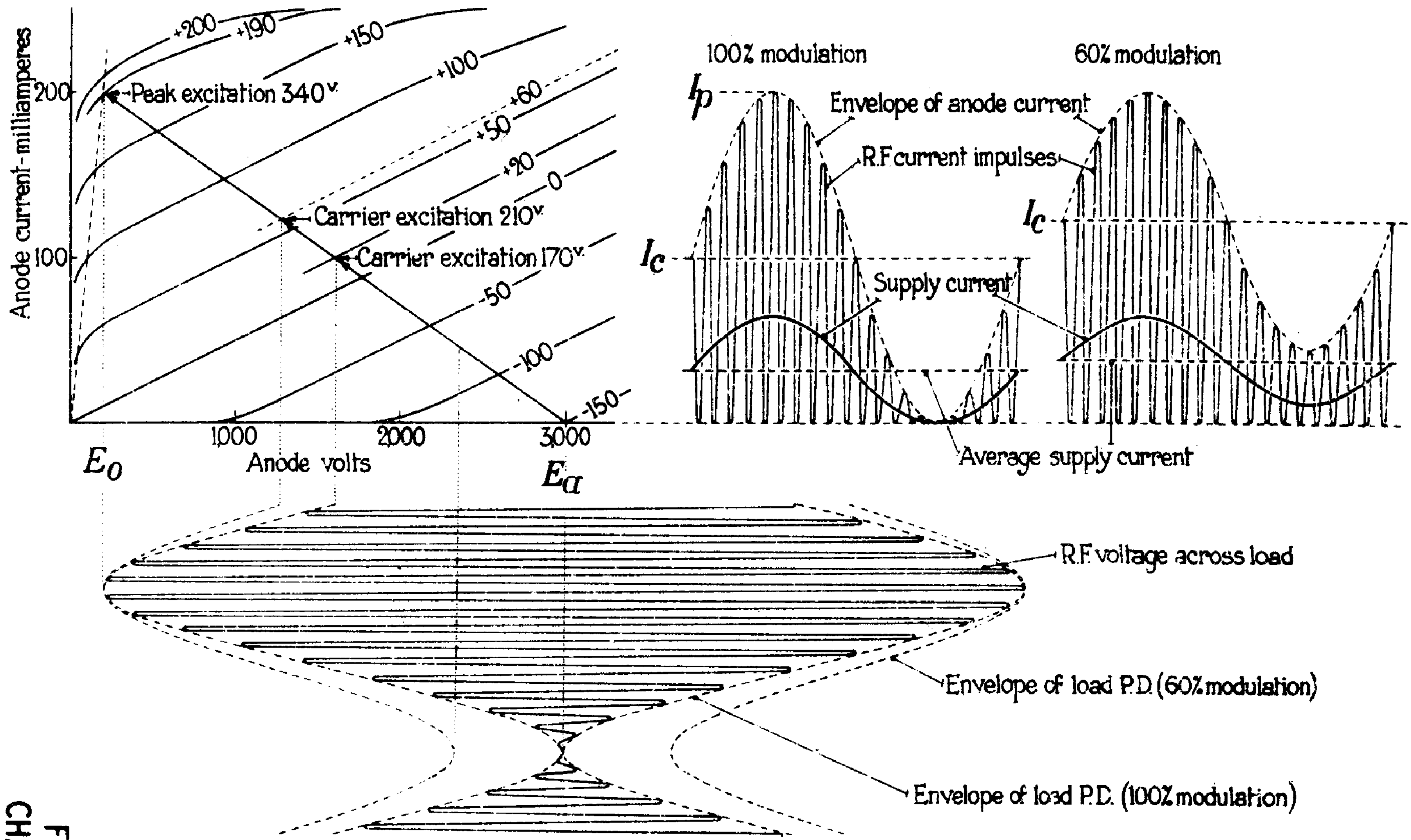


FIG. 54, CHAP. XI.—Distortion due to curvature of characteristics.

impulse, as shown immediately to the right of the characteristics, the points a′ b′ c′ . . . . h, being the projections of the points a b c . . . . h, as indicated by the dotted lines. As it is, however, the grid is initially biased to — 3 volts, and on reaching — 2 volts, instead of rising to the value corresponding to a, the current rises only to A. The point A′ which is obtained by projecting A until it lies vertically below a′ is a point on the current wave-form. When the grid voltage rises to — 1 volt, the points B and b coincide, as also do their projections B′ and b′. Similarly the projection C′ of the point C lies vertically above the projection c′ of the point c. The construction of the current wave-form A′ B′ C′ . . . . H′ is obtained by repetition of the process described. It is seen that the wave is rather more flat-topped than a true sine curve. In practice a class B amplifier is rarely operated under conditions giving such a high degree of distortion as this ; the characteristics corresponding to $V_g = +3$, $V_g = +4$ and $V_g = +5$ are deliberately drawn close together in order to exaggerate the effect for illustrative purposes. The grid is also shown as being biased slightly beyond the cut-off voltage, giving rise to distortion of the lower portion of the current wave. If the grid swing were limited in such a manner that the grid reached a maximum positive potential of only 3 volts, the current wave-form would be that shown in the extreme right of the diagram, with the corresponding half sine wave in dotted line for comparison. This is more closely representative of practical operating conditions.

Amplification of modulated waves

116. The principal use of the class B radio-frequency power amplifier is for the amplification of modulated waves. Although the wave-form of the anode current impulse is not identical with that of the grid excitation voltage, the envelopes of the anode current and load P.D. will be nearly so provided that the operating conditions are chosen in such a manner that the dynamic characteristic is sensibly straight. This implies that the dynamic resistance of the load should be as large as the other conditions will allow, and as already shown, R_d should if possible be not less than $2\,r_a$ and preferably of the order of $4\,r_a$. It must, however, be observed that a high dynamic load resistance can be profitably employed only if a very high supply voltage is available, and the grid excitation must also be correspondingly increased. If the relation between anode current and excitation voltage were truly linear, the average anode current would remain constant, irrespective of the depth of modulation, and the reading of a milliammeter in the anode circuit is therefore an indication of the degree with which linearity is approached. In discussing modulated waves we are chiefly interested in (i) the carrier conditions and (ii) the conditions at the peak of the modulation cycle. It is convenient to denote quantities such as peak anode current, load P.D., etc., under the former conditions by I_p, $\mathcal{V}_a$, etc., and the corresponding quantities at the modulation peak by I'_p, $\mathcal{V}'_a$, etc. Assuming that provision is to be made for 100 per cent. sinusoidal modulation, the peak anode current I'_p when completely modulated will be found as above, but the peak anode current I_p of the unmodulated carrier will be equal to $\frac{I'_p}{2}$. The load P.D. will also be only one-half its greatest permissible value, hence the input power P_i with unmodulated carrier, will be $\frac{E_a I_p}{\pi} = \frac{E_a I'_p}{2\pi}$, and the carrier output power P_o will be $\frac{(E_a - E_o) I_p}{4} = \frac{(E_a - E_o) I'_p}{8}$. For example, referring to fig. 55, with an anode supply voltage of 3,000 volts and a virtual load of 14,050 ohms, the peak value of the anode current under unmodulated conditions will be 100 milliamperes, and the input power $\frac{E_a I_p}{\pi} = \frac{3{,}000 \times \cdot 1}{\pi} = 95 \cdot 5$ watts. The carrier power output will be $\frac{2{,}810 \times \cdot 1}{8} = 35 \cdot 125$ watts, and the efficiency $\frac{35 \cdot 125}{25 \cdot 5} = 36 \cdot 7$ per cent. Comparing these results with those of para. 109, it will be observed that the power output is one-quarter and the efficiency one-half of that obtained



OPERATION OF LINEAR AMPLIFIER

FIG. 55
CHAP. XI

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with the same supply voltage and load impedance, but operating with the greatest permissible grid swing, i.e. 340 volts. The grid excitation required to produce this carrier power will of course be only 170 volts.

117. If the wave is to be modulated to a depth of less than unity, the carrier power may be increased. Thus if the load resistance and supply voltage remain as before, viz. $R_e = 14{,}050$ ohms, $E_a = 3{,}000$ volts, and the excitation under carrier conditions is increased to 210 volts, $I_p = 125$ milliamperes. The load P.D. will now be 1,760 volts and the power output $\frac{1{,}760 \times \cdot 125}{4} = 55$ watts, while the power input is $\frac{3{,}000 \times \cdot 125}{\pi} = 119$ watts. The carrier efficiency is now 46 per cent. Under these conditions the greatest anode current change which can be achieved, without departure from linearity, is 75 milliamperes. The corresponding depth of modulation is easily found ;

$$I'_p = 125 + 75 = 200 \text{ milliamperes}$$

$$\mathcal{I}'_a = \frac{I'_p}{2} = 100 \text{ milliamperes}$$

$$I_p = 125 \text{ milliamperes}$$

$$\mathcal{I}_a = 62 \cdot 5 \text{ milliamperes}$$

$$\begin{aligned} K &= \frac{\mathcal{I}'_a - \mathcal{I}_a}{\mathcal{I}_a} \\ &= \frac{100 - 62 \cdot 5}{62 \cdot 5} \\ &= \cdot 6, \end{aligned}$$

corresponding with a depth of 60 per cent.

118. Since, if the amplifier is truly linear, the average current during modulation is the same as when modulation is not occurring, the power input is the same no matter what the depth of modulation may be, whereas the output power during modulation is greater than under unmodulated conditions. Thus when modulation is taking place, the power converted into heat is greater than during unmodulated periods but the power output is greatly increased and the efficiency is increased. For this reason, the dynamic resistance of the load should always be adjusted by an alteration of anode tapping point (or rearrangement of the tappings on the pulse coil, where fitted) with an unmodulated grid excitation corresponding to the desired carrier amplitude. In order to obtain a high value for the product ηP_o the virtual resistance of the load should be at least equal to r_a and preferably greater than $1 \cdot 5\, r_a$. Care must be taken however, not to exceed the maximum permissible supply voltage, which must increase with the load resistance in order to maintain the desired output.

The class C amplifier

119. The class C amplifier is employed for the following purposes :—

(i) In all the amplifier stages of W/T transmitters. In the stage immediately following the master-oscillator it is preferable to avoid running into grid current, in order to reduce the load on the master-oscillator and to maintain it at a constant magnitude during the whole oscillatory cycle.

(ii) As the modulated stage in a frequency-controlled R/T transmitter. The amplifier may be modulated by imposing an audio-frequency variation either upon the anode voltage or upon the grid-filament P.D., the former being referred to as an anode-modulated amplifier, and the latter as a grid-modulated amplifier. The physical aspects of the process of modulation are dealt with in Chapter XII.

(iii) As a frequency multiplier, the amplified output being usually of double or treble the input frequency. The stage is then called a frequency-doubler or frequency-trebler.

Class C W/T amplifier

120. As already stated, in the class C amplifier the grid is biased to beyond cut-off point and the anode current flows for less than 180 electrical degrees; the actual angle is generally between 120° and 160°. It is convenient to use one-half of this angle for purposes of calculations and in the following discussion, the symbol φ will be used to denote one-half the angle of anode current flow; φ will be referred to as the "operating angle". For example, in a class B amplifier $\varphi = 90°$. It is easily seen that the quantities $\mathcal{V}_g$, E_g, E_b, E_a and φ are interdependent. In all practical cases the maximum anode supply voltage is fixed by practical considerations, while the minimum anode-filament P.D. E_o and most positive grid potential E_g must be decided as in any other form of amplifier. Thus it is usually necessary to decide upon a tentative value of φ and find the relations between $\mathcal{V}_g$, E_b and φ.

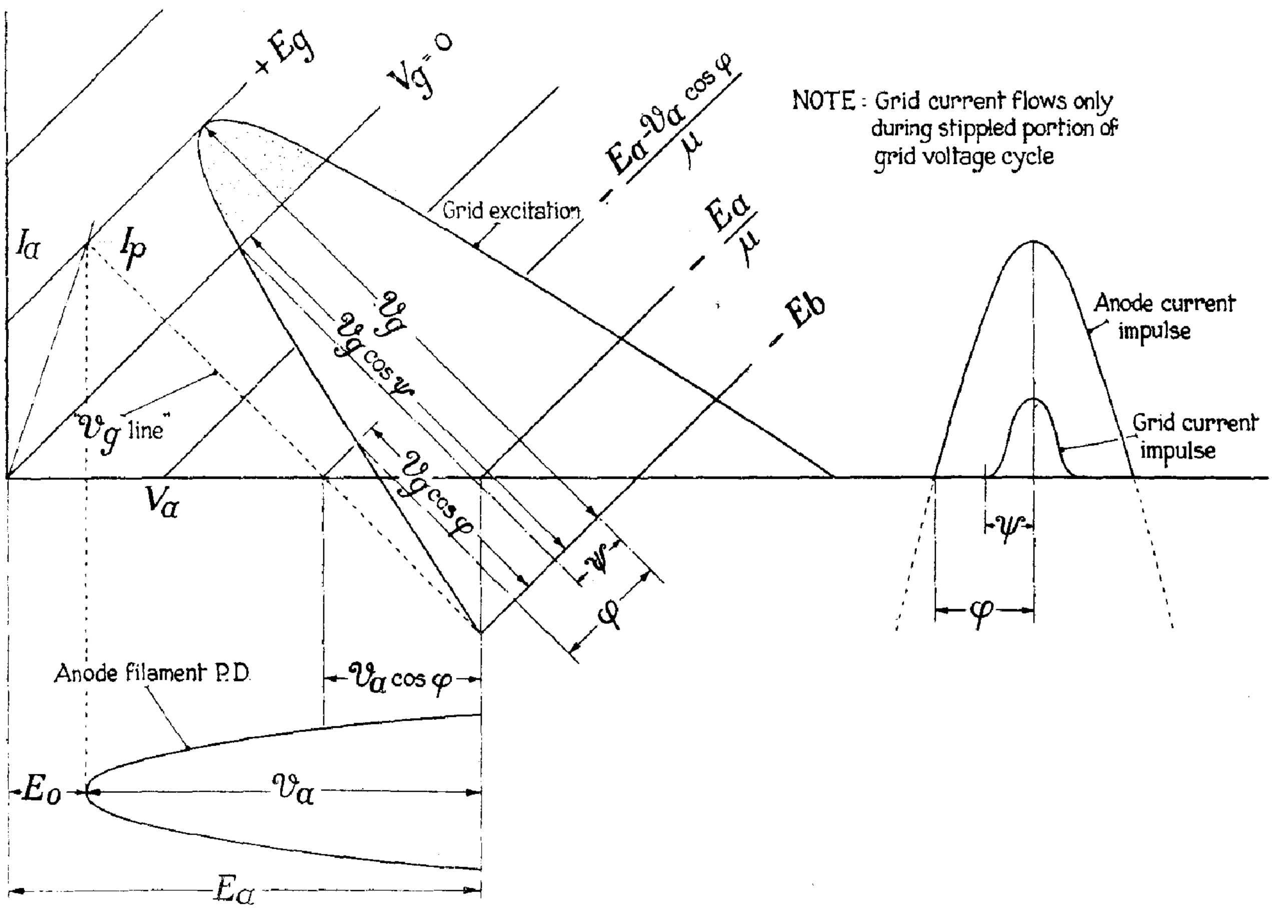


FIG. 56, CHAP. XI.—Relation between various operating voltages in class C amplifier.

Referring to fig. 56, the anode supply voltage being E_a and the minimum anode-filament P.D. E_o, the load P.D. will be $\mathcal{V}_a = E_a - E_o$. It is seen that anode current will not commence to flow until the instantaneous grid voltage has arisen from $-E_b$ to $-E_b + \mathcal{V}_g \cos\varphi$. The oscillatory load P.D. v_a rises as the grid becomes less negative and at the instant when anode current commences, $v_a = \mathcal{V}_a \cos\varphi$. At this instant the grid potential is $\dfrac{E_a - \mathcal{V}_a \cos\varphi}{\mu}$. Assuming that the grid may be allowed to swing positive up to the value E_g volts therefore, the required excitation is

$$\mathcal{V}_g = E_g + \frac{E_a - \mathcal{V}_a \cos\varphi}{\mu} + \mathcal{V}_g \cos\varphi$$

or

$$\mathcal{V}_g = \left[E_g + \frac{E_a - \mathcal{V}_a \cos\varphi}{\mu}\right]\frac{1}{1 - \cos\varphi}.$$

The magnitude of the required bias voltage is

$$
\begin{aligned}
E_b &= \mathscr{V}_g - E_g \\
&= \left[E_g + \frac{E_a - \mathscr{V}_a \cos\varphi}{\mu}\right] \frac{1}{1-\cos\varphi} - E_g \\
&= E_g \frac{\cos\varphi}{1-\cos\varphi} + \frac{E_a}{\mu}\left(\frac{1}{1-\cos\varphi}\right) - \frac{(E_a - E_o)\cos\varphi}{\mu\,(1-\cos\varphi)} \\
&= \frac{E_a}{\mu} + \left(E_g + \frac{E_o}{\mu}\right)\frac{\cos\varphi}{1-\cos\varphi}.
\end{aligned}
$$

This relation is of fundamental importance in the theory of class C amplification. The sign of E_b is of course negative.

Power output and efficiency

121. Since the anode current flows for less than 180°, the anode current impulses are no longer half sine waves, so that the fundamental component (H_1) of the anode current will be less than $\frac{I_p}{2}$ and the average anode current I_{av} less than $\frac{I_a}{\pi}$. It is possible to calculate the ratios $\frac{I_{av}}{I_p} = \alpha$ and $\frac{\mathscr{I}_a}{I_p} = \beta$ for various operating angles. The results are exhibited in graphical form in fig. 57, in which three pairs of H_1 curves are given in solid line. Those marked (1) are calculated for ideal characteristics, i.e. assuming that $r_a I_a = V_a + \mu V_g$. Similarly the curves marked (1·5) refer to characteristics obeying the law $r_a I_a = (V_a + \mu V_g)^{1\cdot5}$ and those marked (2) to characteristics obeying the law $r_a I_a = (V_a + \mu V_g)^2$. The latter are also of use in determining the grid driving power with greater accuracy than has hitherto been attempted. With the aid of these curves the power input, output and efficiency can be determined by means of the relations

$$
\begin{aligned}
P_i &= I_{av}\, E_a \\
&= \alpha\, I_p\, E_a \\
P_o &= \frac{\mathscr{I}_a \mathscr{V}_a}{2} \\
&= \frac{\beta\, I_p\, \mathscr{V}_a}{2} \\
\eta &= \frac{P_o}{P_i}.
\end{aligned}
$$

122. To illustrate the method of calculating the performance of a class C amplifier we may take the 1,000-watt valve previously discussed. Allowing for grid current we have already found that with the limiting value of E_o set at not less than 5 E_g the maximum permissible peak anode current is about 900 milliamperes, while the maker stipulates that E_a shall not exceed 10,000 volts. Assuming that an operating angle of 75° is chosen and that the first power law is obeyed, $\alpha = \cdot265$, $\beta = \cdot45$. The maximum permissible input is therefore

$$
\begin{aligned}
P_i &= \alpha\, I_p\, E_a = \cdot265 \times \cdot9 \times 10{,}000 \\
&= 2{,}380 \text{ watts,}
\end{aligned}
$$

and the minimum value of E_o is found directly from the valve characteristics to be 1,800 volts.

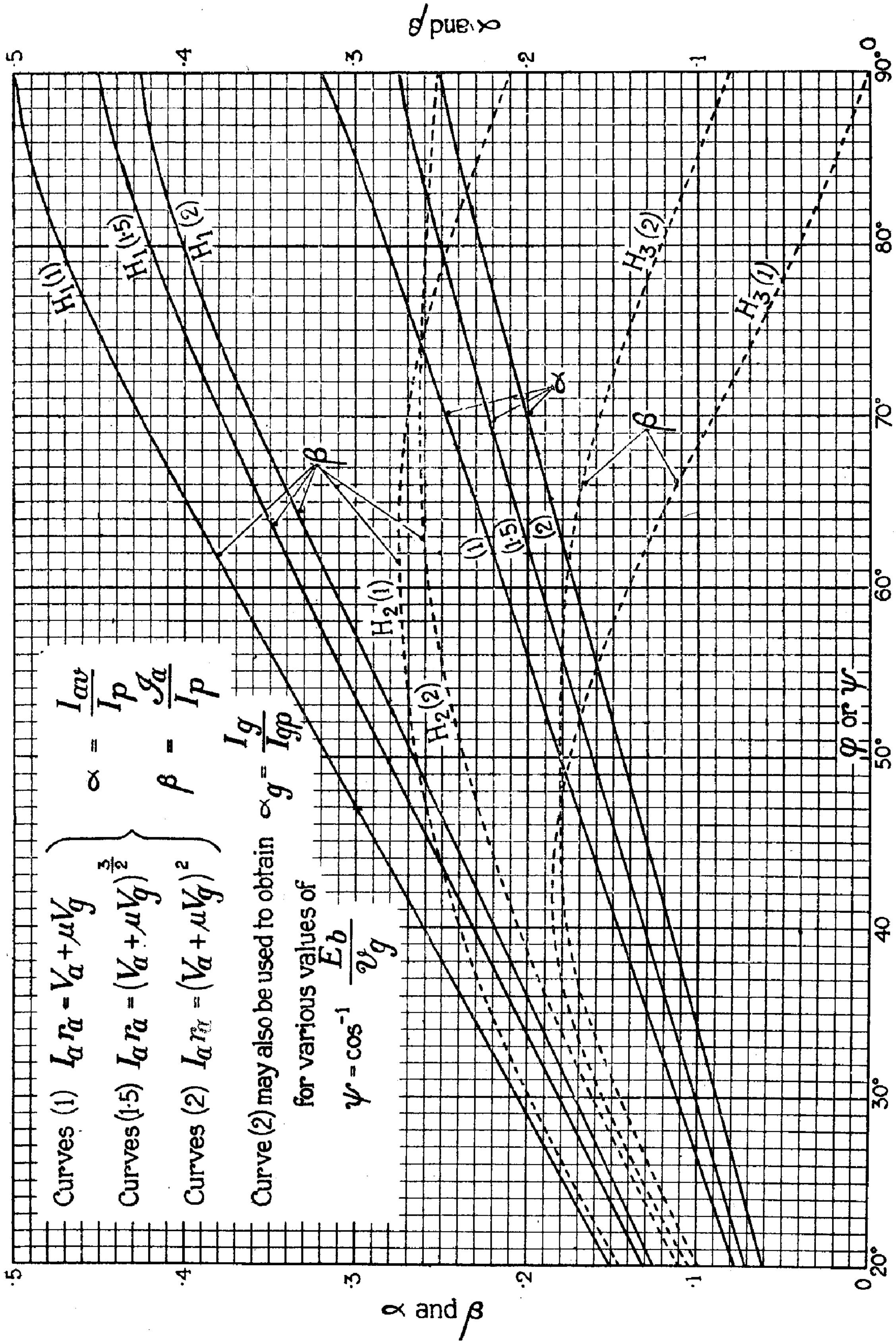


FIG. 57, CHAP. XI.—Values of α and β for power amplifier calculations.

Hence $\mathscr{V}_a$ will be 8,200 volts.

$$P_o = \frac{\mathscr{I}_a \mathscr{V}_a}{2} = \frac{\beta I_p \mathscr{V}_a}{2}$$
$$= \cdot 225 \times \cdot 9 \times 8{,}200$$
$$= 2{,}050 \text{ watts.}$$
$$P_d = P_i - P_o$$
$$= 330 \text{ watts.}$$
$$\eta = 86 \text{ per cent.}$$

The appropriate load impedance R_d is equal to

$$\frac{\mathscr{V}_a}{\mathscr{I}_a} = \frac{\mathscr{V}_a}{\beta I_p} = 40{,}700 \text{ ohms.}$$

Since $\cos 75° \doteqdot \cdot 26$ and $E_g = \frac{E_o}{5}$, the required grid bias will be

$$E_b = \frac{10{,}000}{50} + \left(\frac{1{,}800}{5} + \frac{1{,}800}{50}\right) \frac{\cdot 26}{1 - \cdot 26},$$
$$= 200 + 139$$
$$= 339 \text{ volts.}$$

The grid excitation necessary to obtain this output is

$$\mathscr{V}_g = \left(\frac{1{,}800}{5} + \frac{10{,}000 - 8{,}200 \times \cdot 26}{50}\right) \frac{1}{1 - \cdot 26}$$
$$= 705 \text{ volts.}$$

It is of interest to repeat the calculations on the assumption that the characteristics obey the $\frac{3}{2}$—power law. From curves $(1 \cdot 5)$ of fig. 57, if $\varphi = 75°$, $\alpha = \cdot 234$ and $\beta = \cdot 4$. Hence for the same value of I_p and supply voltage E_a,

$$P_i = \cdot 234 \times \cdot 9 \times 10{,}000$$
$$= 2{,}106 \text{ watts,}$$
$$P_o = \cdot 2 \times \cdot 9 \times 8{,}200$$
$$= 1{,}476 \text{ watts,}$$
$$P_d = 630 \text{ watts,}$$
$$\eta = 70 \text{ per cent.}$$

The load impedance under these conditions will be

$$\frac{\mathscr{V}_a}{\beta I_p} = \frac{8{,}200}{\cdot 4 \times \cdot 9} = 22{,}800 \text{ ohms.}$$

Grid driving power

123. Referring again to fig. 56 it is seen that grid current commences when $\mathscr{V}_g \cos \psi = E_b$ and therefore flows during an interval corresponding to an operating angle ψ, where

$$\cos \psi = \frac{E_b}{\mathscr{V}_g}.$$

In the previous example therefore

$$\cos \psi = \frac{339}{705} = \cdot 48,$$

$$\psi = 61°.$$

Note that this angle is smaller than φ. The grid current impulse is in fact considerably more peaky than a sine curve ; the peak grid current I_{gp} can be found from the $I_g - V_a$ characteristics and its average value I_g is approximately given by

$$I_g = a_g \, I_{gp}$$

where a_g is taken from the square law curve (2) of fig. 57, using the angle ψ instead of φ. In the present example I_{gp} = 158 milliamperes and $a_g = \cdot 175$, the grid driving power being

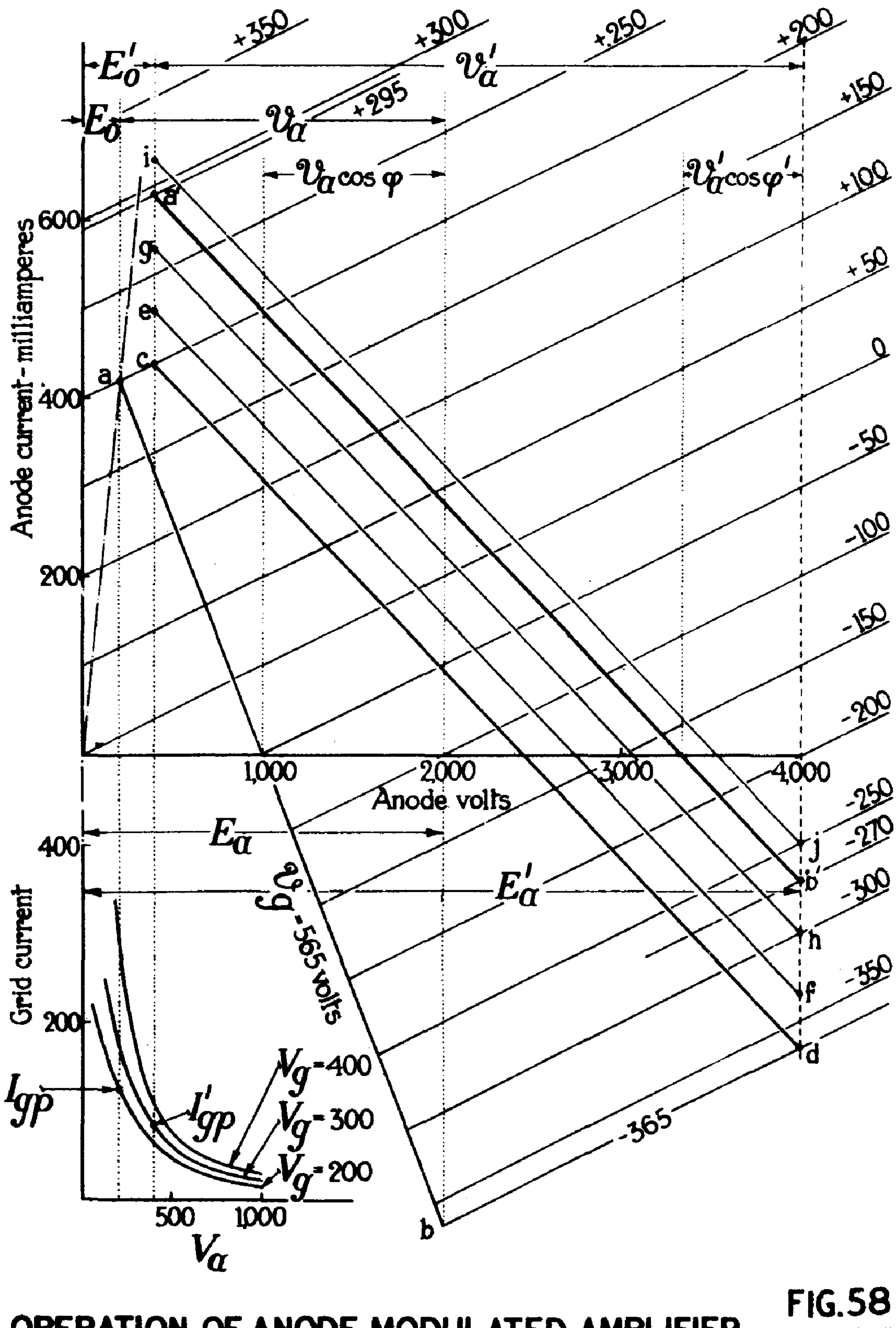
$$P_g = a_g \, I_{gp} \, \mathcal{V}_g$$
$$= \cdot 175 \times \cdot 158 \times 705$$
$$= 19 \cdot 5 \text{ watts.}$$

The anode-modulated amplifier

124. The operation of this amplifier resembles that of the W/T amplifier in that the grid excitation is derived from a master-oscillator (or a buffer amplifier) and is of constant amplitude during the whole of the audio-frequency cycle. The anode voltage is the sum of the supply voltage E_a and the audio-frequency voltage set up in the speech choke, the amplitude of the latter voltage being dependent upon the intensity of the sound input. The greatest audio-frequency voltage which can be handled without serious distortion possesses an amplitude equal to the anode supply voltage, thus the total H.T. voltage varies between zero and $2E_a$ during the audio-frequency cycle. In order to obtain distortionless modulation the relation between currents and voltages must be linear, i.e. when the supply voltage is doubled, the P.D. across the load impedance, and the fundamental component of anode current, must also rise to twice the values under unmodulated conditions. Averaged over an audio-frequency cycle however, the supply voltage and current remain nearly constant because the superimposed variations are practically symmetrical about their mean values. For sinusoidal modulation to a depth of unity the power input must be 1·5 times the carrier input, and the output power must also increase by 50 per cent., in order to supply the side-band power. The average efficiency does not vary to any extent and therefore the power dissipated during sinusoidal modulation will be about 50 per cent. greater than under carrier conditions.

125. In general the anode-modulated amplifier requires a much higher grid bias than the W/T amplifier, because the grid bias must be allowed to vary during the modulation cycle. It is desirable that this variation should not be too large a fraction of the total bias, and therefore the latter should be as large as possible. It follows that the operating angle should be considerably smaller than in the W/T amplifier.

126. An idea of the possibilities with regard to any particular valve is most easily obtained by a study of the operation, first under carrier conditions, and afterwards at the peak of the audio-frequency cycle. The process will be illustrated with reference to the valve of which the ideal characteristics are given in fig. 58. The valve constants are, $r_a = 10{,}000$ ohms, $\mu = 20$, $I_e = 800$ milliamperes, $P_L = 60$ watts. The carrier conditions will be first investigated, the notation used being as in previous paragraphs. In a low-power valve of this kind we may allow the anode-filament P.D. to fall to not less than the maximum positive grid voltage, and the chain-dotted line represents its lower limit, $E_o = E_g$. As a W/T amplifier the maximum permissible supply voltage is 3,000 volts, but as an anode-modulated amplifier it is assumed



OPERATION OF ANODE MODULATED AMPLIFIER

FIG. 58
CHAP. XI

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to be only 2,000 volts. If then we also stipulate that E_o shall not be less than 10 per cent. of the supply voltage, since $E_a = 2{,}000$ volts, $E_o = 200$ volts. The value of I_p follows from the equation to the chain-dotted line, i.e.

$$I_p = E_o \frac{\mu + 1}{r_a}$$

$$= \frac{200 \times 21}{10{,}000}$$

$$= 420 \text{ milliamperes.}$$

It is now necessary to adopt a trial operating angle which will give a sufficiently large output with a high efficiency and a dissipation not exceeding $\frac{2}{3} P_L$. A value of φ in the neighbourhood of 60° is indicated in order that the carrier bias may be considerably negative, i.e. of the order of three times the "cut-off" bias. The latter is equal to $-\frac{E_a}{\mu}$ or -100 volts. Actually the line a b, representing the excursion of the excitation voltage, was drawn from a through the point $V_a = 1{,}000$, $I_a = 0$, intersecting the E_a line at $V_g \doteq -365$. Then

$$\mathcal{V}_a \cos\varphi = 1{,}000$$

$$\mathcal{V}_a = 1{,}800$$

$$\cos\varphi = \frac{1{,}000}{1{,}800}$$

$$= \cdot 554,$$

so that $\varphi = 56°$ (nearly).

A trial calculation of the input, output and efficiency then shows whether this bias and operating angle is satisfactory. From fig. 57 for $\varphi = 56°$, $\alpha = \cdot 2$, $\beta = \cdot 348$

$$I_{av} = \alpha I_p$$

$$= \cdot 2 \times 420$$

$$= 84 \text{ milliamperes.}$$

$$\mathcal{I}_a = \beta I_p$$

$$= \cdot 348 \times 420$$

$$= 146 \text{ milliamperes.}$$

$$P_i = I_{av} \; E_a$$

$$= \cdot 084 \times 2{,}000$$

$$= 168 \text{ watts.}$$

$$P_o = \frac{\mathcal{V}_a \mathcal{I}_a}{2}$$

$$= \frac{146 \times 1{,}800}{2}$$

$$= 131 \cdot 4 \text{ watts.}$$

$$P_d = P_i - P_o$$

$$= 36 \cdot 6 \text{ watts,}$$

which is rather less than $\frac{2}{3} P_L$ and may be considered satisfactory. The efficiency is

$$\eta = \frac{P_o}{P_i}$$
$$= \frac{131 \cdot 4}{168}$$
$$= 78 \text{ per cent.},$$

and the grid excitation required for these conditions is

$$\mathscr{V}_g = E_b + E_g = 365 + 200 = 565 \text{ volts.}$$

127. In order to obtain 100 per cent. distortionless modulation the audio-frequency voltage must, during its cycle, vary the anode voltage of the modulator valve between 0 and $2E_a$. At the peak of the cycle, therefore, $E'_a = 2E_a$; it is also necessary that $\mathscr{V}'_a = 2\mathscr{V}_a$ and $\mathscr{I}'_a = 2\mathscr{I}_a$; in other words the instantaneous power output at the peak must be four times the carrier output. It follows that $E'_o = 2E_o$. The grid excitation, being derived from the previous stage, will remain constant at 565 volts. The slope of the line representing the excitation—which may be called the "$\mathscr{V}_g$-line"—is easily found, since one end must lie on the vertical through $E'_a = 4{,}000$, and it must extend over a range of 565 volts. Several trial $\mathscr{V}_g$-lines are drawn in the diagram. The first, (c d) was inserted in order to deduce what would occur if the grid bias were held rigidly constant. It crosses the anode voltage base line at $V_a = 2{,}450$ volts, so that $\mathscr{V}'_a \cos \varphi' = 1{,}650$ volts. It follows that $\cos \varphi' = \frac{1{,}650}{3{,}600} = \cdot 457$, $\varphi' = 63°$. Referring to fig. 57, for $\varphi' = 63°$, $\beta = \cdot 388$. On this $\mathscr{V}_g$-line the value of I'_p is 440 milliamperes, therefore $\mathscr{I}'_a = \cdot 388 \times 440 = 172$ milliamperes, which is much less than $2\mathscr{I}_a$. Thus an operating angle of 63°, at the audio-frequency peak, will not give distortionless modulation. Successive trials with the $\mathscr{V}_g$-lines e f and g h gave the following operating angles and values of $\mathscr{I}'_a$:—

line e f, $I_p' = 500$, $\varphi' = 70°$, $\mathscr{I}'_a = 212$ milliamperes;
line g h, $I_p' = 570$, $\varphi' = 74°$, $\mathscr{I}'_a = 254$ milliamperes;
line i j, $I'_p = 670$, $\varphi' = 82°$, $\mathscr{I}'_a = 322$ milliamperes.

The latter is fairly close to the desired value and, after a further trial, too close to the final value to be shown, the line a′ b′ was chosen. Here $I'_p = 630$, $E'_b = 270$, $\mathscr{V}'_a \cos \varphi' = 650$, $\cos \varphi' = \cdot 18$ and $\varphi' = 79°$. From fig. 57, $\beta' = \cdot 468$ and therefore $\mathscr{I}'_a = \cdot 468 \times 630 = 295$ milliamperes. This is quite near to the desired value (292 milliamperes) and the remainder of the calculations may be completed. For $\varphi' = 79°$, $\alpha' = \cdot 27$, and therefore

$$I'_{av} = \alpha' I'_p = 27 \times 630$$
$$= 170 \text{ milliamperes}$$

$$P'_i = I'_{av} E'_a$$
$$= \cdot 170 \times 4{,}000$$
$$= 680 \text{ watts}$$

$$P'_o = \frac{\mathscr{I}'_a \mathscr{V}'_a}{2}$$
$$= \frac{\cdot 295 \times 3{,}600}{2}$$
$$= 532 \text{ watts.}$$

$$P'_d = 148 \text{ watts,}$$

i.e. four times the dissipation under carrier conditions. The output P'_o is also equal to $4P_o$ within the limits of the arithmetical accuracy. Thus the modulator will be practically distortionless if the bias is allowed to vary between -365 and -270 volts during the period in which the audio-frequency voltage is additive to the supply voltage. During the other half-cycle the grid will run appreciably more negative than 365 volts, but the operating angle becomes vanishingly small at the negative peak of the modulation cycle ; consequently the peak current may theoretically have any value whatever, and the average current will still fall to practically zero. As a result, the variation of $\bar{I}_{av}$ is nearly symmetrical about the carrier value and no appreciable fluctuation of mean anode current will be observed if the operating conditions are correct.

Grid driving power and bias variation

128. The grid driving power for each of the two above conditions may be found as in the W/T amplifier. Under carrier conditions it is seen from the $I_g - V_a$ curves that $I_{gp} = 125$ milliamperes. As $\cos\ \psi = \dfrac{E_b}{\mathcal{V}_g} = \dfrac{365}{565} = \cdot 645$, $\psi = 50°$, and reference to fig. 57 shows that $\alpha_g = \cdot 143$, so that $I_g = \cdot 143 \times 125 = 17\cdot 8$ milliamperes. The grid driving power will be

$$\begin{aligned} P_g &= I_g \mathcal{V}_g \\ &= \cdot 0178 \times 565 \\ &= 10\cdot 1 \text{ watts.} \end{aligned}$$

At the peak of the audio-frequency cycle, $I'_{gp} = 75$ milliamperes, and $\cos\ \psi = \dfrac{270}{565} = \cdot 477$, $\psi = 62°$. Hence $\alpha'_g = \cdot 18$ and $I'_g = \cdot 18 \times 75 = 13\cdot 5$ milliamperes. The grid driving power will therefore be only

$$\begin{aligned} P'_g &= I'_g \mathcal{V}_g \\ &= \cdot 0135 \times 565 \\ &= 7\cdot 7 \text{ watts.} \end{aligned}$$

If the grid bias is to be derived from a condenser with a leak resistance of R_g ohms, we have

$$\begin{aligned} E'_g &= R_g\, I'_g = 270 \text{ volts} \\ E_g &= R_g\, I_g = 365 \text{ volts,} \end{aligned}$$

and it is of interest to verify whether these conditions are both satisfied. We have

$$\begin{aligned} R_g &= \frac{270}{I'_g} \\ &= \frac{270}{\cdot 0135} \\ &= 20{,}000 \text{ ohms,} \end{aligned}$$

and also

$$\begin{aligned} R_g &= \frac{365}{I_g} \\ &= \frac{365}{\cdot 0178} \\ &= 20{,}600 \text{ ohms.} \end{aligned}$$

The values of R_g so derived agree to well within the accuracy with which the peak values of grid current can be read off the curves. This is not always so, and it may be found desirable to use a combination consisting of battery (or automatic) grid bias, which is of constant magnitude throughout the audio-frequency cycle, together with a smaller bias voltage derived from a condenser and leak resistance. Such expedients are only necessary where a high degree of fidelity is of primary importance.

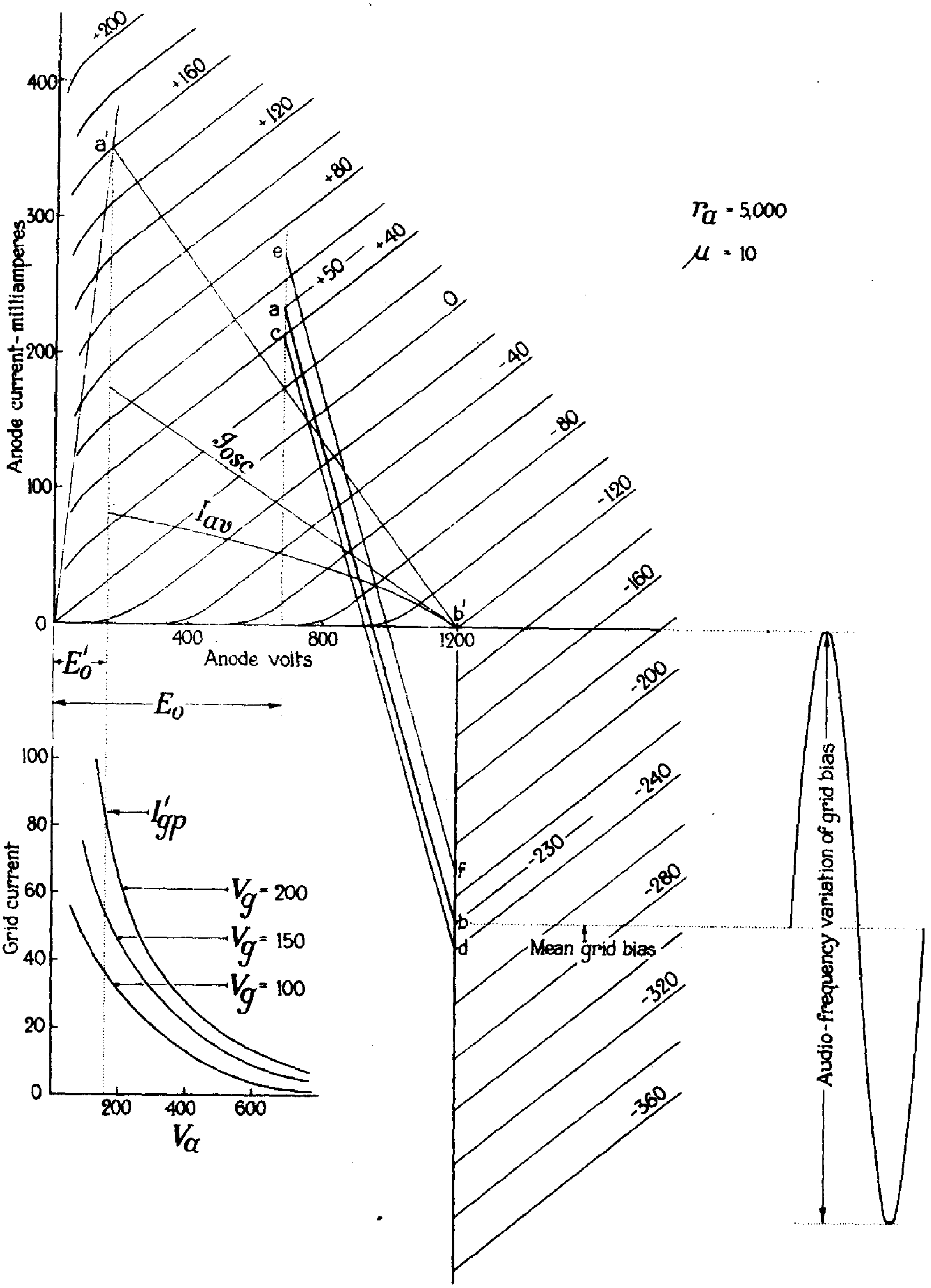
The grid-modulated amplifier

129. In the grid-modulated amplifier the grid-filament P.D. consists of (i) the grid bias voltage, (ii) the radio-frequency excitation derived from the previous stage, e.g. a master oscillator or buffer amplifier, and (iii) an audio-frequency voltage derived from the secondary winding of the microphone transformer, either directly or by means of a sub-modulator valve. The latter voltage, in effect, causes a cyclical variation of grid bias. The operation of this form of modulator is discussed briefly in Chapter XII ; it is here only proposed to deal with the possibility of an approach to distortionless modulation, and the power relations. As before, a concrete example will be taken, and fig. 59 shows the ideal characteristics of a valve having the following constants, namely, $r_a = 5{,}000$ ohms, $\mu = 10$, $P_L = 60$ watts, $I_e = 450$ milliamperes, V_a (max.) $=$ 1,200 volts. The chain-dotted line represents the lower limit of the anode-filament P.D. assuming that it must not be less than the grid-filament P.D. Whereas in the anode-modulated amplifier the operating angle is maintained very nearly constant, grid bias modulation necessitates a cyclical variation of operating angle as the actual bias changes. For 100 per cent. modulation this variation is from 0° to 90°, and when only the radio-frequency excitation is applied, i.e. under carrier conditions, the operating angle is generally very near to 60°.

130. Let us now investigate the conditions under which 100 per cent. modulation may be obtained. Under carrier conditions, the grid will be biased to a value considerably more negative than the cut-off voltage $\frac{E_a}{\mu}$, the exact value being as yet unknown. At the positive peak of the audio-frequency cycle the grid-filament P.D. may be carried exactly to " cut-off." If then we find the greatest permissible values of I_p and E_g, the required radio-frequency excitation is easily found. These values of I_p and E_g are I'_p and E'_g respectively. The current I'_p is found as in earlier discussions, i.e. $I'_p = \cdot 8\ I_e = \cdot 8 \times 450 = 360$ milliamperes. At the instant when the anode current reaches this value $V_g = E'_g$ and $V_a = E'_o = E'_g$. For ease of location, we may therefore let $I'_p = 350$ milliamperes, and the upper end of the line representing the radio-frequency excitation—the $\mathscr{V}_g$-line—will be at the point a′, where $E'_g = E'_o =$ 160 volts. As the grid bias is to be equal to $\frac{E_a}{\mu}$ at the instant of peak modulation, the operating angle φ will be 90°, so that the other end of the $\mathscr{V}_g$-line will be at the point b′.

131. The following data regarding the conditions at the peak of the modulation cycle are now known.

$$I'_p = 350 \text{ milliamperes}$$
$$\varphi' = 90°$$
$$\left.\begin{array}{l}\alpha' = \cdot 3184 \\ \beta' = \cdot 5\end{array}\right\} \text{from fig. 57}$$
$$E'_o = 160 \text{ volts}$$
$$E'_g = 160 \text{ volts}$$
$$E'_b = \frac{E_a}{\mu} = 120 \text{ volts}$$
$$\mathscr{V}'_a = E_a - E'_o = 1{,}040 \text{ volts}$$
$$\mathscr{I}'_a = \beta' I_p = 175 \text{ milliamperes}$$
$$I'_{av} = \alpha' I_p = 110 \text{ milliamperes}$$
$$\mathscr{V}_g = E'_g + E'_b = 280 \text{ volts.}$$



OPERATION OF GRID MODULATED AMPLIFIER
(K = 1)

FIG. 59
CHAP. XI

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From these we may calculate the instantaneous power input P'_i, output P'_o and efficiency η'. Before doing so, however, the corresponding data will be found for the carrier condition. For 100 per cent. distortionless modulation, we must have

$$\mathscr{I}_a = \frac{\mathscr{I}'_a}{2} = 87 \cdot 5 \text{ milliamperes}$$

$$\mathscr{V}_a = \frac{\mathscr{V}'_a}{2} = 520 \text{ volts}$$

$$E_o = E_a - \mathscr{V}_a = 680 \text{ volts},$$

and φ must be about 60°. The upper end of the $\mathscr{V}_g$-line must lie somewhere upon the vertical through E_o, and its lower end somewhere upon the vertical through E_a. The amplitude $\mathscr{V}_g$ of the radio-frequency voltage is still 280 volts. From these considerations the trial $\mathscr{V}_g$-line c d was drawn, intersecting the anode voltage base line at $V_a = 930$ volts. We then have

$$\mathscr{V}_a \cos \varphi = 1{,}200 - 930 = 270$$

$$\cos \varphi = \frac{270}{520} = \cdot 52$$

$$\varphi = 58\tfrac{1}{2}°$$

$$\beta = \cdot 36.$$

At the point c, we find $I_p = 215$ milliamperes ;

$$\mathscr{I}_a = \beta I_p = 77 \cdot 5 \text{ milliamperes},$$

which is too small. Similarly the $\mathscr{V}_g$-line e f intersects the base line at $V_a = 1{,}000$ volts.

$$\mathscr{V}_a \cos \varphi = 200$$

$$\cos \varphi = \frac{200}{520} = \cdot 384$$

$$\varphi = 67\tfrac{1}{2}°$$

$$\beta = \cdot 41$$

$$I_p = 275 \text{ milliamperes}$$

$$\mathscr{I}_a = 116 \text{ milliamperes},$$

which is too large. Now take an intermediate such as a′ b′, intersecting the base line at $V_a = 950$ volts.

$$\mathscr{V}_a \cos \varphi = 250$$

$$\cos \varphi = \frac{250}{520} = \cdot 48$$

$$\varphi = 61°$$

$$\beta = \cdot 375$$

$$I_p = 235 \text{ milliamperes}$$

Also $\alpha = \cdot 215$ and $I_{av} = 51$ milliamperes.

On this line

$$\mathscr{I}_a = \beta I_p$$
$$= 88 \text{ milliamperes},$$

which is practically the desired value. Hence we see that the mean bias must be that given by the intersection of this line and the vertical through E_a, and is $-$ 230 volts, so that the amplitude of the audio-frequency grid-filament voltage must be $230 - 120 = 110$ volts. At the negative peak of its cycle, the grid bias will swing down to -340 volts, and as the excitation voltage is 280, the grid will reach $+60$ volts. Provided, however, that the output circuit has a high magnification, i.e. low damping, the oscillatory load P.D. will swing almost equally above and below its carrier value, and consequently the operating angle at the point under consideration will be very small indeed. Since for small values of φ, both β and α also become very small, it may be taken that both $\mathcal{I}_a$ and I_{av} will fall to zero at this instant. The amplitude of the oscillatory component of anode current is shown in fig. 59 as $\mathcal{I}_{osc}$, and the average current during the audio-frequency cycle as I_{av}.

132. The power relations will be as follows.

Peak of modulation cycle :—

$$
\begin{aligned}
P'_i &= I'_{av} E_a \\
&= \cdot 110 \times 1{,}200 \\
&= 132 \text{ watts} \\
P'_o &= \frac{\mathcal{I}'_a \mathcal{V}'_a}{2} \\
&= 175 \times 520 \\
&= 91 \text{ watts} \\
P'_d &= 132 - 91 \\
&= 41 \text{ watts} \\
\eta' &= \frac{91}{132} = 69 \text{ per cent.}
\end{aligned}
$$

Unmodulated carrier :—

$$
\begin{aligned}
P_i &= I_{av} E_a \\
&= \cdot 051 \times 1{,}200 \\
&= 61 \cdot 2 \text{ watts} \\
P_o &= \frac{\mathcal{I}_a \mathcal{V}_a}{2} \\
&= \cdot 088 \times 260 \\
&= 22 \cdot 8 \text{ watts} \\
P_d &= 61 \cdot 2 - 22 \cdot 8 \\
&= 38 \cdot 4 \text{ watts} \\
\eta &= \frac{22 \cdot 8}{61 \cdot 2} \\
&= 37 \text{ per cent.}
\end{aligned}
$$

It will be observed that the supply current rises and falls in a nearly symmetrical manner about its value I_{av} at the carrier condition. It follows that the average power input during operation is approximately equal to P_i, i.e. about 60 watts. During 100 per cent. sinusoidal modulation, the output power increases by the amount necessary to generate the required side-frequencies, i.e. from about 23 to about 34 watts. The average efficiency will be rather higher than that achieved with the unmodulated carrier, but cannot be expected to exceed 60 per cent. When the depth of modulation is less than 100 per cent., the efficiency is correspondingly decreased.

Grid driving power

133. The grid driving power at the modulation peak is found from the $I_g - V_a$ curves as before, taking the conditions at the audio-frequency peak. The peak grid current I'_{gp} will be about 84 milliamperes, and $\cos\ \psi = \frac{E'_b}{\mathscr{V}_g} = \frac{120}{280} = \cdot 43$, hence $\psi \doteqdot 65°$. Taking $\alpha'_g = \cdot 185$ from the square law curve of fig. 57, $I'_g = \cdot 185 \times 84 = 15 \cdot 5$ milliamperes, and

$$\begin{aligned} P_g &= I'_g \mathscr{V}_g \\ &= \cdot 0155 \times 280 \\ &= 4 \cdot 35 \text{ watts.} \end{aligned}$$

It must be particularly noted that this power is necessarily supplied by the audio-frequency source. The average power is of course somewhat smaller than this, but the calculation shows that the microphone transformer, or sub-modulator valve as the case may be, must be capable of delivering a peak current of 84 milliamperes, and yet not cause appreciable distortion during the majority of the cycle when it is very lightly loaded. A permanent resistance shunt across the audio-frequency input terminals, although increasing the load on the source, will tend to reduce distortion.

Automatic grid bias

134. The mean grid bias of a grid-modulated amplifier is often obtained by means of a resistance R_b in series with the negative H.T. supply lead. If this resistance is to carry only the anode current of the amplifier itself, its value is easily calculated. Continuing the previous example, we have found that the mean bias should be 230 volts. As the average direct current under carrier conditions is 51 milliamperes, we have

$$\begin{aligned} E_b &= R_b\, I_{av} \\ R_b &= \frac{230}{\cdot 051} \\ &= 4{,}500 \text{ ohms.} \end{aligned}$$

Unless certain precautions are taken, the bias will vary during the modulation cycle, e.g. at the positive peak it will be $R_b\, I'_{av} = 4{,}500 \times \cdot 175 = 790$ volts, and at the negative peak will be zero. It is therefore necessary to shunt the resistance by a condenser of such a value that its reactance X_c at the lowest audio-frequency is small compared to R_b. Suppose we require to transmit audio frequencies down to 200 cycles per second ($\omega = 1{,}250$), and X_c is to be not more than $\cdot 1\ R_b$ at this frequency.

$$\begin{aligned} X_c &= \frac{1}{\omega C} = \cdot 1\ R_b \\ \frac{1}{\omega C} &= 450 \\ C &= \frac{1}{450 \times 1{,}250} \text{ farad} \\ &= 1 \cdot 8 \text{ microfarad.} \end{aligned}$$

If, however, the resistance is called upon to carry the anode current of other stages, the variation of grid bias is not so pronounced. Suppose the transmitter to consist only of a master-oscillator and an amplifying stage as above, and the master-oscillator to require a mean anode current of 50 milliamperes. The total current flowing through R_b being denoted by I, we have

At the positive modulation peak $I = \cdot 175 + \cdot 05 = \cdot 225$ amperes.

At the carrier $I = \cdot 050 + \cdot 050 = \cdot 1$ ampere.

At the negative modulation peak $I = 0 + \cdot 050 = \cdot 05$ ampere,

and the required bias resistance will be approximately 2,300 ohms ; if no shunt condenser is fitted the bias will then vary from 520 volts at the positive to 115 volts at the negative peak. A shunt condenser will of course reduce this variation to an inconsiderable amount, nevertheless, where possible battery bias is preferable to automatic bias.

High-mu triode as grid-modulated amplifier

135. In recent years, a coated-filament type of power triode has been developed, ratings up to about 30 watts being available, and these are coming into use in the output stages of low power transmitters. Since with this construction it is possible to obtain a large mutual conductance, a high amplification factor may be obtained with a fairly low anode A.C. resistance ; such valves are sometimes referred to as high-mu (i.e. high mutual conductance) triodes. Owing partly to the comparatively small clearance between the electrodes, and partly to the tendency of the coating to disintegrate under a high electrical stress, the maximum permissible H.T. voltage is usually not more than 600 volts. The ideal characteristics of a valve of this type will be used to illustrate the operation of the grid-modulated amplifier under the conditions giving a maximum depth of modulation of the order of 60 per cent. The data for this valve are as follows

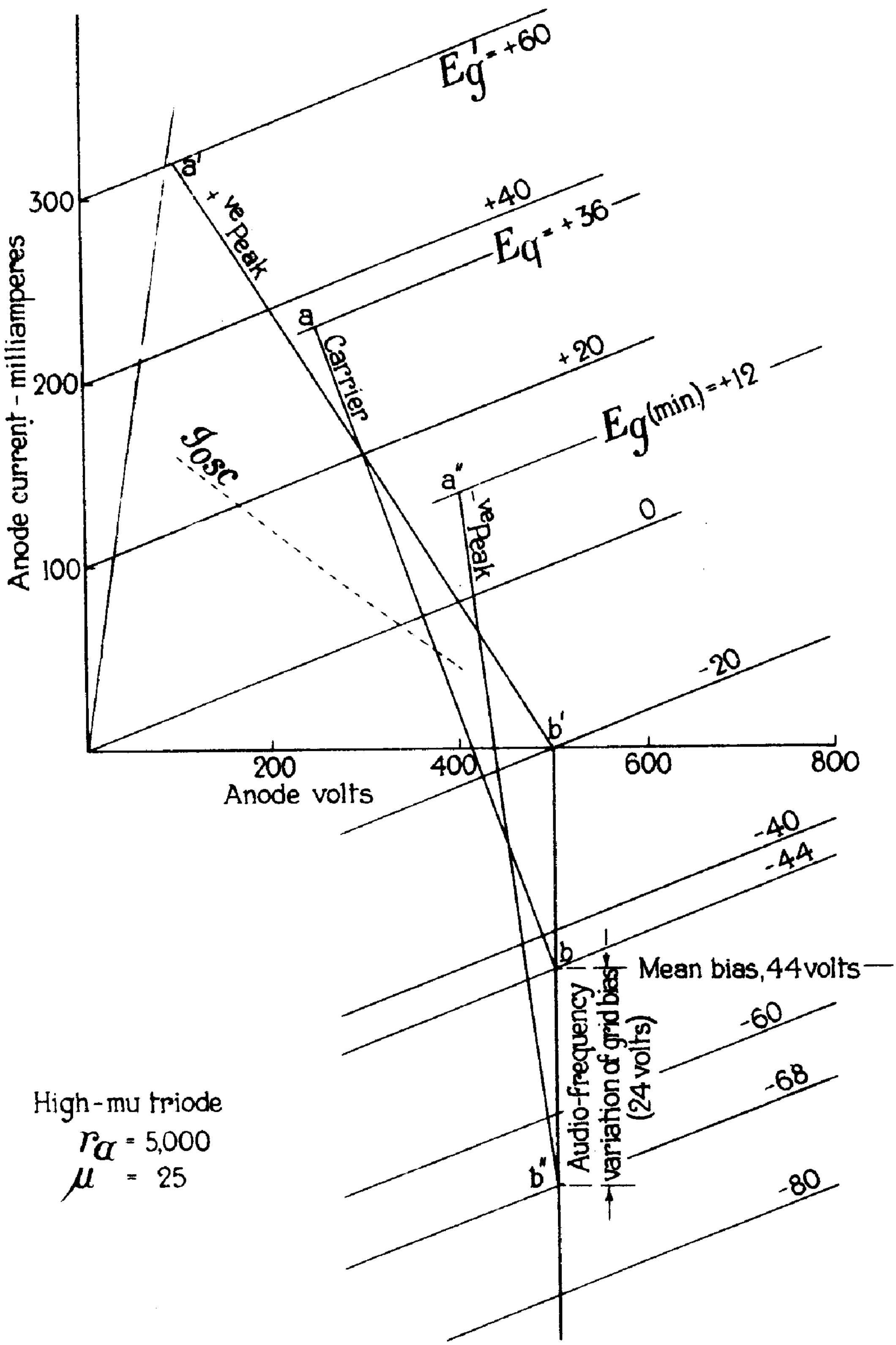
$$I_e = 400 \text{ milliamperes}$$
$$P_L = 20 \text{ watts}$$
$$E_a \text{ (max.)} = 500 \text{ volts}$$
$$r_a = 5{,}000 \text{ ohms}$$
$$\mu = 25$$
$$\frac{\mu^2}{r_a} = \cdot 125$$

At the present stage of development these valves are somewhat prone to trouble due to secondary emission, if the anode-filament P.D. is allowed to fall too low. In the diagram (fig. 60) the chain-dotted line is the limit for E_o on the assumption that E_g is not to exceed 1.5 E_o. Then for $K = \cdot 6$, the $\mathcal{V}_g$-lines shown have been derived thus. Since $I_e = 400$ milliamperes we have, at the positive modulation peak, $I'_p = \cdot 8\ I_e = 320$ milliamperes. The point a′ on the "+ve peak" $\mathcal{V}_g$-line a′ b′ has been located at $I'_p = 320$, $E'_g = 60$, rounding off E'_o to 100 volts instead of 90, and $\mathcal{V}'_a = E_a - E'_o = 400$ volts. For this condition, $\varphi = 90°$ and the grid bias $E'_b = -20$, so that a radio-frequency excitation of $\mathcal{V}_g = 80$ volts is required. By methods previously explained it is easily found that $P'_i = 51$ watts, $P'_o = 32$ watts, $P'_d = 19$ watts, $\mathcal{I}'_a = 160$ milliamperes. Under carrier conditions, for 60 per cent. modulation we must have

$$\mathcal{I}_a = \frac{\mathcal{I}'_a}{1 + K} = \frac{\mathcal{I}'_a}{1 \cdot 6} = 100 \text{ milliamperes}$$

$$P_o = \frac{P'_o}{(1 + K)^2} = 12 \cdot 5 \text{ watts}$$

$$\mathcal{V}_a = \frac{\mathcal{V}'_a}{1 \cdot 6} = 250 \text{ volts.}$$



OPERATION OF GRID MODULATED AMPLIFIER (K = 0·6)

FIG. 60
CHAP. XI

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After a few trials, the "carrier" $\mathcal{V}_g$-line a b was located, giving $I_p = 230$ milliamperes, $\varphi = 71°$, $\alpha = \cdot 25$, $\beta = \cdot 43$. $\mathcal{I}_a = \cdot 43 \times 230 = 99$ milliamperes, which is quite near the required value. Then $P_i = 28\cdot 8$ watts, $P_o = 12\cdot 4$ watts, $P_d = 16\cdot 4$ watts and $\eta = 43$ per cent. On the "—ve peak" $\mathcal{V}_g$-line a″ b″ we have I_p (min.) $= 140$ milliamperes, $\varphi = 50°$, $\alpha = \cdot 18$, $\beta = \cdot 31$, $\mathcal{I}_a$ (min.) $= 43$ milliamperes, rather higher than the required value. The variation of $\mathcal{I}_a$ during the audio-frequency cycle is shown in dotted line as $\mathcal{I}_{osc}$; it is to all intents and purposes linear. It is seen that an audio-frequency variation of grid bias of 24 volts peak will be required, with a mean bias of -44 volts. This variation may be obtained directly from the secondary of a microphone transformer, without the aid of a sub-modulator stage. The correct load impedance will be $R_d = \frac{\mathcal{V}_a}{\mathcal{I}_a} = 2{,}500$ ohms.

Neutralization

136. In the operation of radio-frequency power amplifiers it is essential to ensure that no energy is transferred from the output to the input circuit, otherwise the amplifier may become unstable, tending to produce self-oscillation as described in para. 65. Such transfer of energy may be prevented by arranging the circuit in such a manner that an amount of negative reaction

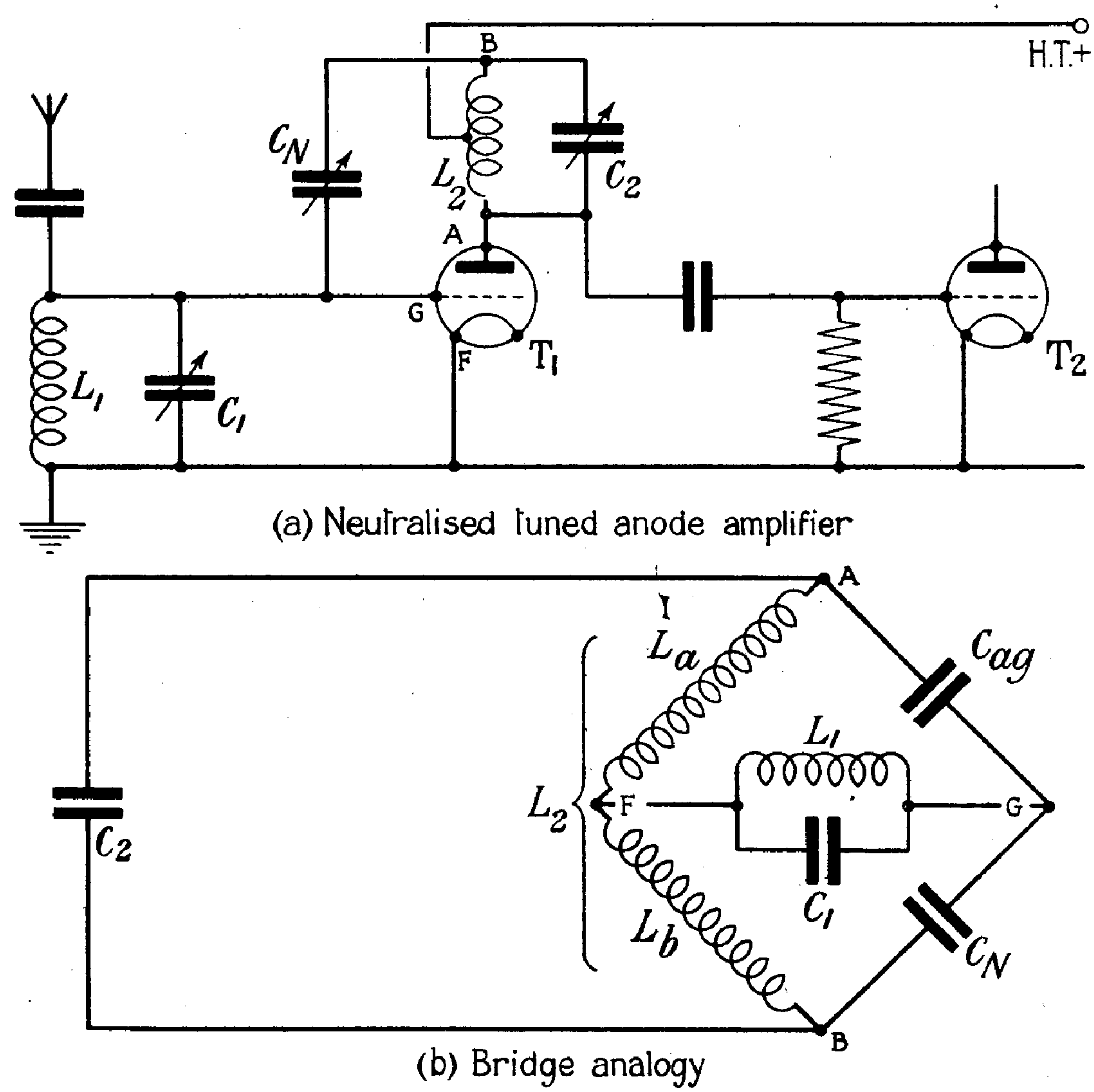


FIG. 61, CHAP. XI.—Principle of neutralization.

is introduced, the magnitude of which is exactly equal to the positive reaction which causes the instability. A truly neutralized circuit operates equally well in the opposite sense, i.e. if the valve tends to transfer energy from the input to the output circuit the neutralizing device will apply positive reaction of such magnitude as to prevent this energy transference. Thus, when a valve is truly neutralized, its input admittance is purely capacitive and neither increases nor decreases the damping of the input circuit. This principle was first applied in triode receiving amplifiers, but the introduction of the screen-grid valve has caused its entire disappearance in this field. Nevertheless it is convenient to refer to this application, and a typical circuit is given in fig. 61 in which a tuned aerial circuit $L_1 C_1$ supplies a signal voltage to the grid and filament of the triode T_1. The output circuit of the latter is a tuned anode $L_2 C_2$, the positive H.T. supply being connected to a centre tap on the coil L_2. Provided that the capacitance to earth of the two ends of the coil is the same, equal and opposite P.D's must be developed between opposite ends of the inductance and the earthed filament. The grid-anode capacitance C_{ag} is then balanced by connecting a neutralizing condenser C_N between the upper end (B) of the inductance L_2 and the grid. Since the ends (A, B) of the coil are at equal and opposite potentials with respect to the filament, transference of energy between input and output circuits will be entirely prevented by making the capacitance C_N equal to C_{ag}.

137. Provided that the grid-anode admittance is purely susceptive, the equivalent circuit of the amplifier is analogous to the wheatstone bridge described in Chapter I, and the circuit of fig. 60a has been re-drawn (fig. 60b) in order to show the analogue. It is assumed that the P.D. to be neutralized is that of the anode tuning condenser C_2, which is therefore regarded as the supply voltage to the bridge. The inductance L_2 is split into two portions L_a L_b, the midpoint being connected to the filament F, so far as oscillatory currents are concerned. The neutralizing condenser C_N and the grid-anode capacitance C_{ag} are in series between the points A and B, the centre point being connected to the grid G. The input circuit $L_1 C_1$ is connected between G and F and from the bridge point of view may be regarded as analogous to the galvanometer in fig. 16, Chapter I. It is easily seen that the points F and G will be at the same potential if

$$\frac{L_a}{L_b} = \frac{C_N}{C_g},$$

i.e. the balance is independent of frequency. Unfortunately, however, this ideal state of affairs rarely exists in practice, owing chiefly to the presence of conductive paths in parallel with the various reactances. The effect of these shunts is to render the effective values of L_a, L_b, C_{ag}, C_n, dependent upon the frequency, so that it becomes necessary to re-neutralize whenever the frequency is changed appreciably. In a receiving amplifier, however, it is highly desirable to keep the number of controls as low as possible and before the introduction of the screen-grid valve a great deal of ingenuity was expended in the endeavour to maintain a constant neutralizing adjustment over a wide frequency range.

138. In the application of this principle to radio-frequency power amplifiers, it may be regarded as almost axiomatic that complete neutralization over a very wide frequency band is only practicable where weight and space are of no account whatever. These circumstances, however, are only applicable to large ground station installations, where the transmitters are generally operated on a spot frequency for very long periods. On the infrequent occasion of a change of operating frequency, the little extra time required for re-neutralization is of no importance, and there is no necessity to introduce extra complication in an endeavour to maintain a single neutralization adjustment over a wide frequency band. Aircraft transmitters are usually adjusted to a few spot frequencies on the ground (using an artificial aerial) and the tuning and neutralizing adjustments tabulated. Only a slight adjustment of amplifier tuning and neutralization is then required when the aircraft is airborne.

Screen-grid power amplifiers

139. In recent years the screen-grid valve has been developed for use as a radio-frequency power amplifier, both for linear or class B amplification, and as a class C amplifier in W/T transmitters. The screening of this valve is less complete than in the receiving type, and the grid-anode capacitance is by no means negligible, although much smaller than that of the corresponding power triode. The screen-grid power valve is chiefly used in very high frequency transmitters, where it is difficult to obtain an accurate neutralizing adjustment with the power triode. It is usually necessary to incorporate the usual neutralizing arrangements, but it is often found possible to obtain a setting of the neutralizing condenser which will maintain a balance over a very wide frequency range.

Frequency multipliers

140. A valve frequency multiplier is essentially an amplifier operating under conditions which lead to a high degree of amplitude distortion, so that the output contains harmonics of the frequency applied to the grid. The distortion may be produced by the curvature of either the $I_g - V_g$ or the $I_a - V_g$ characteristics, the former being generally used when low power is required with a high multiplication, e.g. for the purpose of radio-frequency measurements. In high and very high frequency transmitters it is usual to attain the desired multiplication by a number of frequency doubling or trebling stages, and the anode current curvature is utilized. The amplitude of any harmonic higher than the third is generally too small to be employed economically. The anode current wave-form of an ideal class B amplifier contains only even harmonics, but the output of a class C amplifier contains both even and odd harmonics, and this type is generally used. The anode circuit is of course tuned to the harmonic frequency and not to the frequency of the input, consequently the circuit has no tendency to self-oscillation and does not require neutralization. If the $I_a - V_a$ characteristics of the valve are available, its performance as a frequency doubler or tripler can be calculated in a manner similar to that adopted when its output circuit is tuned to the input frequency, the curves shown in dotted line in fig. 57 being used for this purpose. They give the ratios

$$\beta_2 = \frac{\mathcal{I}_2}{I_p}$$

and

$$\beta_3 = \frac{\mathcal{I}_3}{I_p}$$

where $\mathcal{I}_2$ and $\mathcal{I}_3$ are the amplitudes of the second and third harmonics, respectively, of the fundamental component of anode current. The desired harmonic is of course selected by suitably tuning the anode circuit, and the load P.D, V_a will be, to all intents and purposes, sinusoidal and of the selected harmonic frequency.

141. Before giving typical numerical examples the general trend of these curves should be observed. Take the curve $H_2(1)$ giving the values of β_2 for first-power-law characteristics. The output power is equal to $\frac{\beta_2}{2} I_p (E_a - E_o)$ and the input to $\alpha I_p E_a$. It follows that, for a given ratio of E_o to E_a, the efficiency is directly proportional to $\frac{\beta_2}{\alpha}$, and the product ηP_o directly proportional to $\frac{\beta_2^2}{\alpha}$. By actual point-to-point calculation for various values of φ_2, it is found that for large output and high efficiency the operating angle should be between 40° and 70°, maximum output being achieved when $\varphi_2 \doteqdot 65°$. The efficiency increases as φ_2 is reduced and ηP_o is a maximum when $\varphi_2 \doteqdot 60°$. Similar considerations may be applied to the curve giving the values of β_3 against φ_3, which is applicable to the frequency trebler; in this case, the output

is a maximum (for first-power-law characteristics) when $\varphi_3 \doteqdot 40°$, and the product ηP_o a maximum with a rather smaller operating angle. It is seen then that where β_2 or β_3 is very nearly constant over a range of values of φ, the smaller operating angle is preferable in order to reduce the power dissipation and increase the efficiency. It will however be found that a small operating angle necessitates a large grid excitation, and a consequent increase in grid driving power.

Power output and efficiency of frequency doubler

142. To illustrate the method of calculating the power relations, we shall take the ideal characteristics of fig. 62, the permissible dissipation being 250 watts and the maximum supply voltage 5,000 volts. Let $E_a = 4{,}000$ volts, $E_o = E_g = 300$ volts, and $I_p = 630$ milliamperes; if the valve is to operate as a frequency doubler, reference to the curve H_2 (1) of fig. 57 shows that the maximum value of β_2 is achieved with an operating angle $\varphi_2 = 65°$. We then have

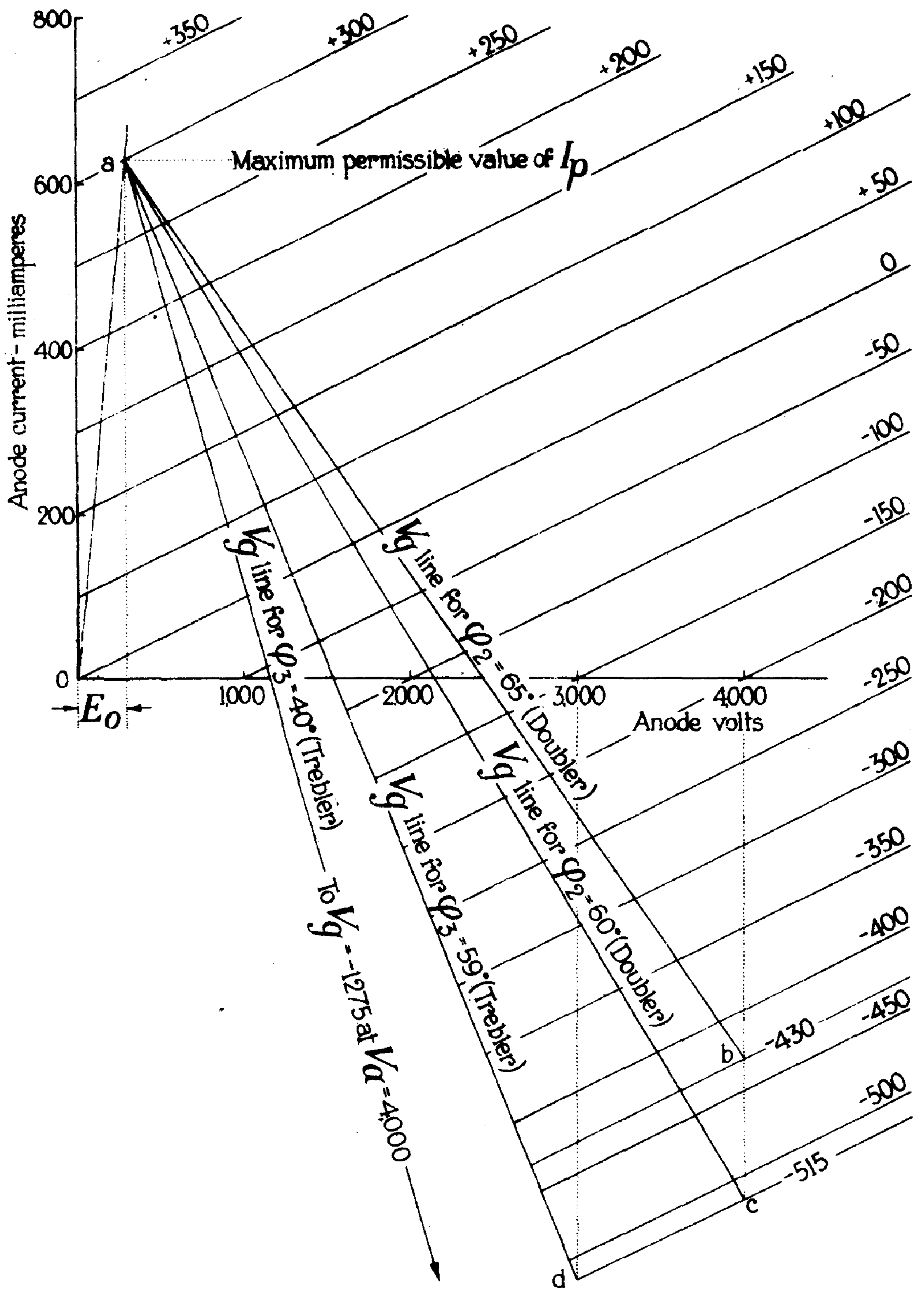
$$\begin{aligned}\mathcal{V}_a &= E_a - E_o = 3{,}700 \text{ volts}\\ \varphi_2 &= 65°\\ \cos\varphi_2 &= \cdot 423\\ \mathcal{V}_a \cos\varphi_2 &= \cdot 423 \times 3{,}700\\ &= 1{,}560 \text{ volts.}\end{aligned}$$

The $\mathcal{V}_g$-line a b has been drawn through the anode voltage base line at $4{,}000 - 1{,}560 = 2{,}440$ volts, intersecting the vertical through E_a at a point giving $E_b = 430$ volts. Hence the required excitation $\mathcal{V}_g$ is $430 + 300 = 730$ volts. If this can be provided we may proceed to find the power relations.

$$\begin{aligned}P_i &= a\, I_p\, E_a\\ &= \cdot 23 \times \cdot 63 \times 4{,}000\\ &= 580 \text{ watts}\\ P_o &= \frac{\beta_2}{2} I_p \mathcal{V}_a\\ &= \frac{\cdot 275 \times \cdot 63 \times 3{,}700}{2}\\ &= 322 \text{ watts}\\ P_d &= 258 \text{ watts}\\ \eta &= 55 \text{ per cent.}\end{aligned}$$

Repeating the calculation for an operating angle $\varphi_2 = 60°$, we find

$$\begin{aligned}\mathcal{V}_a &= 3{,}700\\ \cos\varphi_2 &= \cdot 5\\ \beta_2 &= \cdot 274\\ a &= \cdot 212\\ P_i &= \cdot 212 \times \cdot 63 \times 4{,}000\\ &= 535 \text{ watts}\\ P_o &= \frac{\cdot 274 \times \cdot 63 \times 3{,}700}{2}\\ &= 320 \text{ watts.}\\ P_d &= 215 \text{ watts}\\ \eta &= 60 \text{ per cent.}\end{aligned}$$



VARIOUS OPERATING CONDITIONS FOR FREQUENCY MULTIPLICATION

FIG. 62
CHAP. XI

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