CHAPTER 13

AUDIO FREQUENCY POWER AMPLIFIERS

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SECTION 1 : INTRODUCTION

(i) Types of a-f power amplifiers (ii) Classes of operation (iii) Some characteristics of power amplifiers (iv) Effect of power supply on power amplifiers.

(i) Types of a-f power amplifiers

An interesting analysis of possible types of a-f power amplifiers is given in Ref. G1. However, for practical applications we may sub-divide the arrangements as under :

On basis of form of loading—
(a) Plate loaded (the normal arrangement).
(b) Cathode loaded (i.e. cathode follower).
(c) Combined plate and cathode loaded (e.g. the McIntosh amplifier—see Fig. 13.50D and the Acoustical QUAD amplifier, Refs. H4, H6).

On basis of coupling to load—
(a) Transformer coupled.
(b) Choke-capacitance coupled (shunt feed).

On basis of load connection—
(a) Single-ended output (i.e. one valve).
(b) Push-pull output.
(c) Parallel output.
(d) Push-pull parallel output.
(e) Single-ended push-pull (Refs. E32, H1, H2).

On basis of excitation—
(a) Single input to grid.
(b) Push-pull input to grids.
(c) Single input to one grid, with other valve excited from common plate or cathode circuit—see Sect. 6(viii).
(d) Grounded-grid single input.*
(e) Grounded-grid push-pull input*.

*Not normally used with a-f amplifiers.

On basis of input coupling—
(a) Resistance-capacitance.
(b) Choke-capacitance.
(c) Transformer.
(d) Direct coupling.

On basis of type of valve—
(a) Triode.
(b) Pentode.
(c) Beam power amplifier.

On basis of use of feedback (see Chapter 7)—
(a) Without feedback.
(b) Negative voltage feedback.
(c) Negative current feedback.
(d) Cathode follower.
(e) Combined positive and negative feedback.

On basis of Class of Operation—see (ii) below.

On basis of input power required—
(a) No grid input power.
(b) With grid input power.

(ii) Classes of operation

Class A operation* is the normal condition of operation for a single valve, and indicates that the plate current is not cut off for any portion of the cycle.

Limiting Class A push-pull operation* is operation such that one valve just reaches plate current cut-off when the other reaches zero bias.

Class AB operation* indicates overbiased conditions, and is used only in push-pull to balance out the even harmonics.

Class B operation* indicates that the valves (which are necessarily in push-pull) are biased almost to the point of plate current cut-off.

The numeral " 1 " following A or AB indicates that no grid current flows during any part of the cycle, while " 2 " indicates that grid current flows for at least part of the cycle. With Class B operation the " 2 " is usually omitted since operation with grid current is the normal condition.

(iii) Some characteristics of power amplifiers

Amplifiers incorporating **negative feedback** are covered in Chapter 7. The design of **output transformers** is covered in Chapter 5 Sect. 3.

For a limited supply voltage Class A_1 gives the lowest **power output** with given valves, while Class AB_1 and Class AB_2 give successively higher outputs. Pentodes and beam power amplifiers give greater power output than triodes under the same conditions. Negative feedback does not affect the maximum power output.

The **power plate efficiency** is the ratio of the audio frequency power output to the d.c. plate and screen power input. It is least for Class A_1, and increases progressively with Class AB_1, AB_2 and B. It is less for Class A_1 triodes than for Class A_1 pentodes or beam power amplifiers.

The **sensitivity**† is normally taken as the ratio of milliwatts output to the square of the r.m.s. grid voltage. Pentodes and beam power amplifiers have considerably greater sensitivity than triodes. Class AB_1 or any push-pull operation decreases the sensitivity. Amplifiers with grid current require power in the grid circuit ; sensitivity cannot be quoted for such types except for the whole section including the driver valve.

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*See also definitions in Sect. 5(i)B.

†The writer favours the use of the alternative form

$$\text{Sensitivity} = \frac{\sqrt{\text{power output in milliwatts}}}{\text{grid voltage r.m.s.}}$$

which gives values from slightly over 1 (e.g. type 45) to nearly 50. This form has the advantage that, for the same power output, the value is proportional to the voltage gain. This value is the square root of the " sensitivity " as normally measured.

Another alternative form is the ratio of milli-watts output to the square of the peak grid voltage.

The **effective plate resistance** (or output resistance) of a power amplifier is an important characteristic when the load is a loudspeaker. The optimum value of plate resistance depends upon the loudspeaker, but in the majority of cases the optimum is about one fifth of the load resistance for the best frequency response—lower values give heavier loudspeaker damping but a loss of bass response. Feedback amplifiers with very high feedback factors normally have a very low value of output resistance as the result of negative voltage feedback ; in some cases the output resistance is purposely raised to a more suitable value by some device such as bridge feedback—see Chapter 7 Sect. 1(iv). The general question of the optimum plate resistance is covered in Chapter 21 Sect. 1(ii), while loudspeaker damping is covered in Chapter 21 Sect. 3.

Critical load resistance—Pentodes are much more critical than triodes as regards the effects of variation from the optimum value of load resistance—this holds with or without feedback in both cases. See Sect. 2(iv), Sect. 3(viii), Sect. 5(ii), Sect. 6(ii), and summary Chapter 21 Sect. 1(iii).

Distortion—Single Class A_1 triodes are usually operated with 5% second harmonic distortion at maximum output, while the third and higher order harmonics are small under the same conditions. All published data for such valves are based on 5% second harmonic unless otherwise specified. With push-pull Class A_1 triodes the even harmonics are cancelled and only small third and higher order odd harmonics remain. Push-pull class A_1 triode operation is regarded as providing the best fidelity obtainable without the use of feedback.

As the bias is increased towards Class AB_1 operation the odd harmonic distortion increases only slightly until cut-off is just reached during the cycle (i.e. up to Limiting Class A operation), beyond which point a kink appears in the linearity (transfer) characteristic, and the distortion is more displeasing to the listener than is indicated by the harmonic distortion.

Power pentodes operated under Class A_1 conditions on a resistive load may have very slight second harmonic distortion but from 7% to 13% total distortion. This is largely third harmonic with appreciable higher order harmonics. When operated into a loudspeaker load the harmonic distortion is much more severe at low and high frequencies due to the variation of loudspeaker impedance with frequency. Negative feedback may be used to reduce distortion at all frequencies.

With a load of varying impedance, such as a loudspeaker, there is a selective effect on the harmonic distortion. For example, if the impedance of the load is greater to a harmonic than to the fundamental, the harmonic percentage will be greater than with constant load resistance equal to that presented to the fundamental. See Sect. 11(iii) and Fig. 13.54, also Sect. 2(iv) for triodes, Sect. 3(viii) for pentodes.

Owing to the fact that the dominant harmonic with power pentodes is the third, there is very little reduction of distortion due merely to push-pull operation. If, however, the load resistance per valve is decreased, the effect is to increase the second harmonic per valve (which is cancelled out in push-pull) and to decrease the third harmonic, and thus to improve the fidelity.

Some of the distortion occurring with Class AB_2 or Class B operation is due to the effect of grid current on the input circuit. The design of such amplifiers is treated more fully later in Sections 7 and 8.

Normally the harmonic distortion is stated for full power output, but the **rate of increase** is also of importance. Second harmonic distortion (Class A_1 triodes or beam power tetrodes) increases more or less linearly from zero to full power. Third harmonic distortion in pentodes increases less rapidly at first, and then more rapidly as full output is approached. Higher order odd harmonics show this effect even more markedly.

Beam power amplifiers in Class A_1 have considerable second harmonic, but less third and higher order harmonics. When operated in push-pull the second harmonic is cancelled, and the total harmonic distortion on a constant resistive load is small. On a loudspeaker load, however, the same objections apply as for pentodes, and negative feedback is necessary in all cases where good fidelity is required.

For further information on fidelity and distortion see Chapter 14.

In calculating the **frequency response** of an output stage, it should be noted that the inductance of the output transformer varies with the applied signal voltage.

When the d.c. plate current remains nearly constant at all output levels the **regulation of the power supply** is not important as regards the output power, provided that it is adequately by-passed—see (iv) below. With Class AB_1 operation there is a greater variation in current drain from zero to maximum signal, and improved regulation is required in the power supply in order to avoid loss of power and increased distortion. Class AB_2, and particularly Class B amplifiers, require extremely good power supply regulation owing to the large variations in current drain.

The use of **self bias** (cathode bias) reduces the variation of plate current due to change of signal level, and frequently enables less expensive rectifier and filter systems to be used, although in some cases the output may be slightly reduced and the distortion slightly increased as a result. Self bias cannot be used with Class AB_2 or Class B operation.

Parasitic oscillation in the power stage is sometimes encountered, either of a continuous nature or only under certain signal conditions. High-mutual-conductance valves are particularly liable to this trouble, which may be prevented by one or more of the expedients listed below. Class AB_2 or Class B amplifiers sometimes suffer from a negative slope on portion of the grid characteristic; this may sometimes be recognised by a " rattle " in the loudspeaker. Improvement in most cases may be secured by the use of one or more of the following expedients:

A small condenser from each plate to earth.
A condenser from each grid to earth (with transformer input only).
Series stopping resistors in grid, screen and plate circuits, arranged as close as possible to the valve.
Improved layout with short leads.
Input and output transformers with less leakage inductance.

In addition to these expedients, it is usually helpful to apply negative feedback from the plate of the output valve to its grid circuit or to the cathode of the preceding stage.

A Class AB_2 or Class B amplifier requires a **driver stage** and (usually) coupling transformer in addition to the final stage. These, together with the additional cost due to the good regulation power supply, should all be considered in calculating the total cost. It is desirable to consider the whole combination of driver valve, driver transformer and push-pull power stage as forming the power amplifier, and the input voltage to the driver will generally be comparable with that required by a single power pentode.

When fixed bias is required for a class AB_1, AB_2 or B amplifier, this may be obtained from a battery or from a separate rectifier and filter. In order to reduce the cost, back-bias with the addition of a heavy bleeder resistance is often used. Some variation in bias is inevitable with this arrangement, and a loss of power output and an increase of distortion will result. The additional cost of the power supply and filter needed to handle the total current of valves and bleeder must also be considered.

Pentodes and beam power amplifiers may be used as triodes by connecting plate and screen. If the suppressor is brought out to a separate pin it may be connected to the plate. Care should be taken to avoid operating such a valve at a plate voltage higher than the maximum rated screen voltage unless special " triode " maximum ratings are available.

The load impedance with triode power amplifiers may be open-circuited (if necessary) without any serious effects; it should never be short-circuited. The load impedance with pentodes may be short-circuited only if the valve is being operated considerably below its maximum plate dissipation; it should never be open-circuited unless an appreciable degree of negative voltage feedback is being used.

(iv) Effect of power supply on power amplifiers

The usual by-pass capacitor connected across the power supply is large enough to act as a reservoir and maintain practically constant voltage, except at low frequencies.

The analysis of a power supply shunted by a condenser and supplying sine-wave signal current does not appear to have been published.

An analysis has been made (Ref. B6) for the simpler case where the signal current is of rectangular form. The minimum frequency of rectangular waveform for a specified variation of supply voltage over the half-cycle is given by

$$f_r = \frac{K(R_L/R_s) + 1}{4.606KR_L C \log_{10} A} \qquad (1)$$

where $K = I_{b0}/(I_{b0} + I_p)$

R_s = resistance of voltage source

$R_L = E_{b0}/I_{b0}$

C = shunt capacitance

$$A = \frac{(1 - K)}{(1 - K) - [K(R_L/R_s) + 1]x/100}$$

E_{b0} = direct voltage across C

I_{b0} = direct plate current of valve

I_p = peak signal current through valve plate circuit

and x = fall of supply voltage between beginning and end of positive half-cycle, expressed as a percentage of E_{b0}.

The only approximation in eqn. (1) is that the voltage across the condenser at the commencement of the positive half-cycle is taken as being E_{b0}.

Eqn. (1) has been applied to the following example—

$E_{b0} = 250$ V, $I_{b0} = 40$ mA, $I_p = 30$ mA, $C = 16\mu$F, $x = 2$,

to give these results :

When R_L/R_s =	5	10	20	30	35
then f_r =	170	157	127	96	49 c/s

This example is a severe one since a rectangular waveform is more severe than a sine waveform with the same peak current, while the value of x only allows 2% drop in voltage over the half-cycle. If 5% voltage drop is permissible, f_r becomes 39 c/s for $R_L/R_s = 10$. The importance of good power-supply regulation for satisfactory performance at low frequencies is demonstrated.

This effect does not occur to any appreciable extent with push-pull Class A amplifiers, and is not so pronounced with Class AB amplifiers as it is with single-ended Class A or with Class B.

Amplifiers should always be tested for frequency response and distortion at maximum power output, as well as for frequency response at a lower level.

SECTION 2 : CLASS A SINGLE TRIODES

(i) Simplified graphical conditions, power output and distortion (ii) General graphical case, power output and distortion (iii) Optimum operating conditions (iv) Loudspeaker load (v) Plate circuit efficiency and power dissipation (vi) Power sensitivity (vii) Choke-coupled amplifier (viii) Effect of a.c. filament supply (ix) Overloading (x) Regulation and by-passing of power supply.

(i) Simplified graphical conditions, power output and distortion

The basic circuit of a single Class A_1 triode is Fig. 13.1. The load resistance (R_2) is normally connected to the secondary of a transformer (T) whose primary is connected in the plate circuit of the valve. The load resistance (R_1) presented to the valve is given (see Chapter 5 Sect. 1) by

$$R_1 = (N_1/N_2)^2 R_2 \qquad (1)$$

where N_1/N_2 = transformer turns ratio, primary to secondary. In the ideal case, the primary and secondary windings of T may be assumed to have zero resistance, and the d.c. plate current I_b to remain constant under all operating conditions.

The valve characteristics are shown by Fig. 2.22 in which the loadline AQB has a slope $- 1/R_L$ and passes through the operating point Q. In this example, $E_b = 250$ volts and $E_c = - 10$ volts, and Q is vertically above $E_b = 250$ volts, because the full E_b voltage is applied between plate and cathode.

The maximum grid swing which can be used is 20 volts, that is from A ($E_c = 0$) to B ($E_c = -20$ volts) since the $E_c = 0$ curve is the border of the grid current region. Actually a slight grid current usually flows at zero bias, but this is generally neglected.

FIG. 13.1

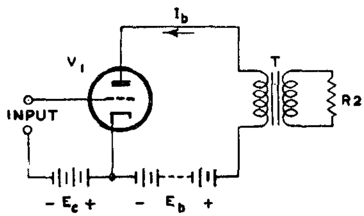


Fig. 13.1. *Basic circuit of a Class A_1 triode with transformer-coupled load.*

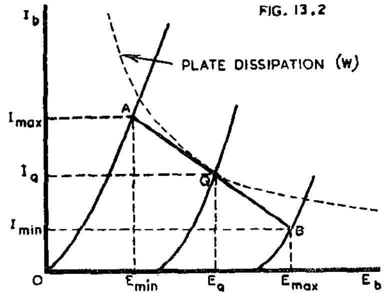


Fig. 13.2. *Loadline applied to triode plate characteristics to determine power output and distortion.*

The **power output** may be calculated from a knowledge of the maximum and minimum voltages and currents along the loadline, assuming a sinewave input voltage (Fig. 13.2):

$$\text{Power Output} \approx \tfrac{1}{8}(E_{max} - E_{min})(I_{max} - I_{min}) \quad (2)$$
$$\approx \tfrac{1}{8}R_L(I_{max} - I_{min})^2 \quad (3)$$
$$\approx \tfrac{1}{8}(E_{max} - E_{min})^2/R_L \quad (4a)$$
$$\approx \tfrac{1}{2}E_b I_Q(1 - E_{min}/E_b)(1 - I_{min}/I_b) \quad (4b)$$

Equations 2, 3 and 4 give the exact fundamental frequency power output when the odd harmonic distortion is negligible. The power output of 5% second harmonic frequency is only 1/400 of the fundamental power output, which is negligible.

The load resistance corresponding to the loadline is

$$R_L = (E_{max} - E_{min})/(I_{max} - I_{min}) \quad (5a)$$
$$= (E_b/I_b)(1 - E_{min}/E_b) \quad (5b)$$

The percentage second harmonic distortion

$$= \frac{\tfrac{1}{2}(I_{max} + I_{min}) - I_Q}{I_{max} - I_{min}} \times 100 \quad (6)$$

$$= \frac{AQ - QB}{2(AQ + QB)} \times 100 \quad (7a)$$

$$= \frac{AQ/QB - 1}{2(AQ/QB + 1)} \times 100 \quad (7b)$$

The equation (7b) has been plotted graphically in Fig. 13.3.

The **voltage gain** of a power amplifier is given by

$$M = (E_{max} - E_{min})/(2E_{c1})$$
$$= 2.82\sqrt{P_0 R_L} \quad (8)$$

where P_o = power output in watts.

The peak values of the fundamental and second harmonic components of the signal plate current are:

$$\text{Fundamental} \quad I_{h1} = \tfrac{1}{2}(I_{max} - I_{min}) \quad (9)$$

$$\text{Second harmonic} \quad I_{h2} = \tfrac{1}{4}(I_{max} + I_{min} - 2I_Q) \quad (10)$$

If $AQ/QB = 11/9 = 1.22$, then the second harmonic distortion will be 5%. This

FIG. 13.3

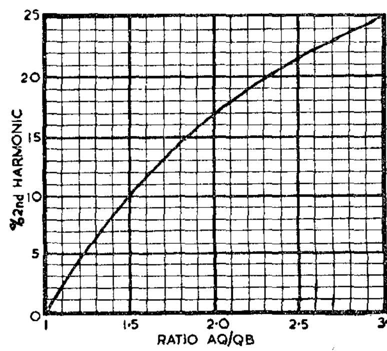
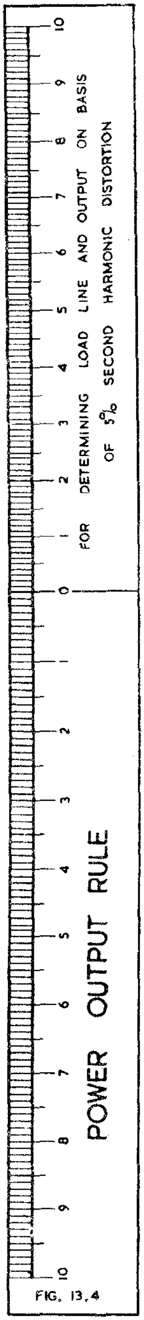


Fig. 13.3. *Curve of second harmonic distortion plotted against the ratio AQ/QB for the two portions of the loadline (Fig. 13.2).*

is the principle of "**5% distortion rule**" reproduced in Fig. 13.4. The rule has each division to the left of 0 a length of 11/9 or 1.22 of the length of a corresponding division to the right of 0. It may be made with each left-hand division 11 millimetres and each right-hand division 9 millimetres. Each of these divisions may be divided into 10 equal subdivisions. It may be used by placing the "0" of the distortion rule at any likely operating point and tilting the rule gradually until the reading on the rule corresponding to the zero bias curve is the same as the reading corresponding to the curve of twice the grid bias at the operating point. AQ/QB will then be 11/9 and the second harmonic distortion corresponding to AQB as a loadline will be 5%.



There are also methods for the graphical determination of any degree of second harmonic distortion. One of these was originated by Espley and Farren (Ref. B1) and is illustrated in Fig. 13.5; it must be drawn on transparent material. It is applied to the curves in a manner similar to the 5% distortion rule, except that a loadline is first drawn in, and then the Harmonic Scale is moved so that DBC is parallel with the loadline. For example, with the loadline EFG the distortion is 20% while with E′FG the distortion is 10%. The scale may be constructed on the basis of:

2nd Harmonic Distortion %	0	5	10	15	20
DB/BC	1.0	1.22	1.5	1.86	2.33

If the valve is to be operated with a known plate voltage, the usual procedure is to take as the operating point (Q) the intersection of the vertical line through E_b with one of the E_c curves such that Q is either on, or below, the "maximum plate dissipation curve" (Fig. 13.2). If the latter is not included in the published curves, it may readily be plotted over a small range. This dissipation curve only affects the operating point Q, and there is no harm if the loadline cuts the curve. In general, if the plate voltage is fixed, a triode gives greatest output when it is operated at the maximum permissible plate current, with the limit of $I_b = \frac{1}{4}E_b/r_p$ (see eqn. 22).

If there is no predetermined plate voltage, a triode gives increasing power output as the plate voltage is increased, even though the plate current is limited in each case by the dissipation.

(ii) General graphical case, power output and distortion

In practice there are several additional factors which should be taken into account.

(A) Resistance of transformer primary

Provision may be made for this resistance by giving the line QE_b a slope of $-1/R'$ where R' is the resistance of the transformer primary or choke (Fig. 13.6). By this means the plate voltage of point Q is less than E_b by the value $R'I_Q$. This does not affect the slope of the loadline, which is determined solely by the effective impedance of the load reflected across the primary of the transformer. It is assumed that the final filter condenser has sufficient capacitance to supply the varying current over each cycle without appreciable change in voltage.

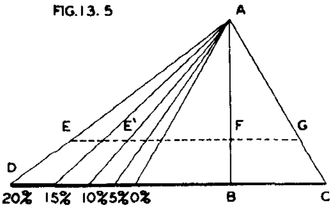
(B) Effect of primary inductance

The reactance of the transformer primary gives an elliptical loadline [see Chapter 2 Sect. 4(vi)]. This may be reduced to any desired extent by increasing the inductance. Obviously this effect is only serious at the lowest signal frequencies; suitable values of inductance are given in Chapter 5 Sect. 3(iii)c.

Fig. 13.4. 5% distortion rule for use in calculating the power output of triodes.

(C) Effect of regulation of power supply

This has only a slight effect on Class A triodes, provided that the final filter condenser has sufficient capacitance to supply the varying current over each cycle without appreciable change in voltage.



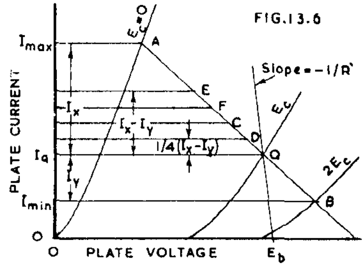


Fig. 13.5. *Alternative form of distortion rule for application to triode loadlines* (*Ref. B*1).
Fig. 13.6. *Plate characteristics of triode exaggerated to show rectification effects.*

(D) Rectification effects

Owing to the second harmonic component of the signal frequency plate current, the average plate current under operating conditions is greater than with zero signal. The increase in plate current is given by

$$\Delta I_b = \tfrac{1}{4}(I_{max} + I_{min} - 2I_Q) \qquad (11)$$

which is also equal to the peak value of the second harmonic current (I_{h2}) as shown by eqn. (10). This may be put into the form

$$\Delta I_b = \tfrac{1}{4}(I_A - I_B) \qquad (12)$$

where $I_A = I_{max} - I_Q$ and $I_B = (I_Q - I_{min})$.

This is illustrated in Fig. 13.6 where AE = QB and EQ = AQ − QB. Obviously AQ/QB $\propto I_A/I_B$. The line EQ is then divided into four equal parts ; C is the centre point of AB and D is the centre point of CQ. The plate current of point D is greater than that of point Q by the amount $\tfrac{1}{4}(I_A - I_B)$ which is equal to ΔI_b. Point D is the only point on the loadline which fulfils the conditions regarding current, while point Q is the only point which similarly fulfils the conditions regarding voltage. The condition is therefore an impossible one, and the loadline must shift upwards until the point of average current lies on the line QE_b.

Unfortunately, as the loadline moves upwards, the relationship between I_A and I_B changes, and the simplest procedure is to draw a second loadline A′B′ (Fig. 13.7) parallel to AB, then to determine its points E′F′C′D′ as for the original loadline. Points D and D′ are joined by a straight line DD′, and the intersection of this with QE_b gives point D″ which must be on the working loadline. Through D″ draw a loadline parallel to AQB, and A″D″B″ is then the maximum signal dynamic loadline. The only error is through assuming that DD′ is a straight line, whereas it is actually the locus of point D ; the error is small if DD′ is short.

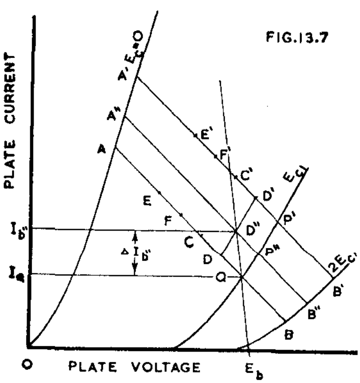


Fig. 13.7. *Method for graphical determination of " shifted " loadline due to rectification.*

It is evident that the change from the no-signal quiescent point Q to the maximum signal dynamic loadline A″B″ will be a gradual process. As the signal increases, so will the loadline move from AQB to A″P″B″. At intermediate signal voltages the loadline will be intermediate between the two limits. The average plate current for zero signal will be I_Q but this will rise to I_b'' at maximum signal.

It will be observed that D″ does not correspond to the intersection of the loadline and the static bias curve. Point P″ is not the quiescent point (which is Q) but may be described as " the point of instantaneous zero signal voltage on the dynamic loadline." Point P″ must therefore be used in the calculations for harmonic distortion at maximum signal. The loadline A″D″P″B″ provides the data necessary for the calculation of power output, second harmonic distortion, and average direct current. All these will, in the general case, differ from those indicated by the loadline AQB.

Summary :

1. Power output is calculated from the loadline A″B″ in the usual manner.
2. Second harmonic distortion (per cent) at maximum signal = (2D″P″/A″B″) × 100.
3. Average d.c. current at maximum signal $= I_b'' = I_Q + \Delta I_b'' =$ current for point D″.

The graphical method above is accurate, within the limits of graphical construction, but rather slow. Approximate results may be calculated from the original loadline by the equations :

$$I_b'' \approx I_Q + \tfrac{1}{4}(I_{max} + I_{min} - 2I_Q)(1 + R_L/r_p) \qquad (13)$$
$$\Delta I_b'' \approx \tfrac{1}{4}(I_{max} + I_{min} - 2I_Q)(1 + R_L/r_p) \qquad (14)$$

Alternatively the rise in current may be calculated from the second harmonic distortion (H_2) and power output :

$$\Delta I_b'' \approx \sqrt{2}H_2\sqrt{P_0/R_L}(1 + R_L/r_p) \qquad (15)$$

In another form, the rise in current may be calculated from the harmonic distortion, I_Q and I_{min} :

$$\Delta I_b'' \approx (I_Q - I_{min})(1 + R_L/r_p)[H_2/(1 - 2H_2)] \qquad (16)$$

or less accurately by

$$\Delta I_b'' \approx (0.78\, I_Q)(1 + R_L/r_p)[H_2/(1 - 2H_2)] \qquad (17)$$

(on the assumption that $I_{min} = 0.22\, I_Q$).

Eqn. (17) is interesting, since it does not involve any data beyond those normally published. If $H_2 = 0.05$ it may be reduced to

$$\Delta I_b'' \approx 0.043\, I_Q(1 + R_L/r_p) \qquad (18)$$

Eqns. (13) to (18) give results slightly lower than the graphical method.

For rectification effects with cathode bias see below.

(E) Cathode bias

Cathode bias loadlines may be drawn on the mutual characteristics [Chapter 2 Sect. 4(v)] but the position is complicated by the rise in plate current caused by rectification. The simplest approach is to assume a voltage between plate and cathode (E_b), determine the plate current under maximum signal conditions as for fixed bias (I_b'') and then to calculate the cathode bias resistor (R_k) from the equation $R_k = E_c/I_b''$. It is then necessary to check for plate dissipation at no signal ; this may be done by the method of Fig. 2.27, drawing a cathode loadline having the correct slope, and noting the plate current (I_{b1}) at which the cathode loadline intersects the mutual characteristic for the desired value of E_b. The dissipation with no signal input is then $E_b I_{b1}$. There is a slight error through assuming that E_b remains constant from zero to maximum signal, but this is usually negligible. Alternatively, the maximum value of I_{b1} may be calculated from the maximum dissipation, and marked on the corresponding E_b curve ; through this point a cathode loadline may be drawn whose slope will indicate the minimum permissible value of cathode bias resistor.

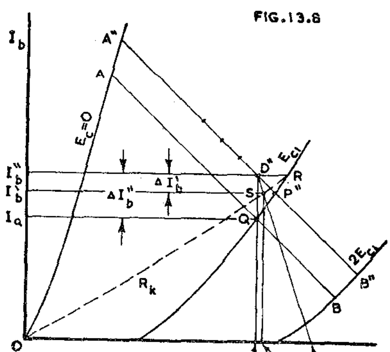


Fig. 13.8. Determining change of plate current with fixed and cathode bias, having operating conditions identical at maximum signal.

Alternatively, cathode loadlines may be drawn on the plate characteristics, but they will be slightly curved [for method see Chapter 12 Sect. 2(iii)].

Rectification effects with cathode bias may be determined entirely from the plate characteristics (Fig. 13.8). The loadline corresponding to maximum signal conditions is A″B″ (as for Fig. 13.7) while point P″ is the point of instantaneous zero signal voltage. The maximum signal plate current is I_b'' as for fixed bias. For the sake of comparison, the zero signal loadline AQB for fixed bias has also been added. From D″ a straight line is then drawn to E_{bb} where E_{bb} is the total supply voltage across both valve and R_k; this line will have a slope of $-1/R_k$.

The cathode bias loadline must pass through the point R on the E_{c1} curve at which the plate current is I_b'', because $R_k = E_{c1}/I_b''$. The R_k loadline also passes through O as shown; it intersects the sloping line through E_{bb} at point S with plate current I_b'. This point S is the quiescent operating point with cathode bias and has a plate voltage E_{b2} which is slightly greater than E_{b1}.

The change in average plate current from no signal to maximum signal is given by $\Delta I_b'$ for cathode bias and $\Delta I_b''$ for fixed bias. It is obvious that $\Delta I_b'$ is always less than $\Delta I_b''$. More specifically,

$$\frac{\Delta I_b'}{\Delta I_b''} \approx \frac{r_p}{R_k} \cdot \frac{E_{c1}}{E_{b1}} \qquad (19A)$$

As a typical example, for type 2A3, $E_b = 250$ volts, $E_{c1} = -45$ volts, $r_p = 800$ ohms, $R_k = 750$ ohms; $\Delta I_b'/\Delta I_b'' \approx 0.2$.

In words, the change of plate current with fixed bias is five times that with cathode bias.

If the value of the cathode bias resistor and the total supply voltage ($E_{bb} = E_b + E_c$) are known, the following procedure may be adopted (Fig. 13.9A).

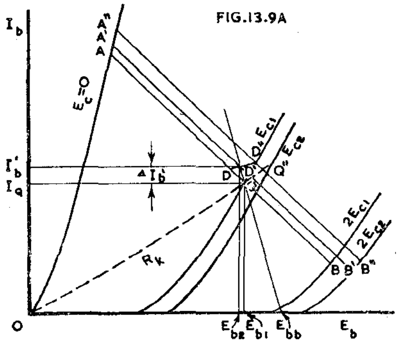


Fig. 13.9A. Determining position of loadline and rise in plate current, given plate supply voltage ($E_{bb} = E_{b1} + E_{c1}$) and cathode bias resistor only.

1. Draw the cathode bias loadline (R_k) on the plate characteristics.
2. Draw a straight line through E_{bb} with a slope of $-1/R_k$.
3. Take the point of intersection (Q) and draw a vertical line to the horizontal axis. This will give the quiescent plate-to-cathode voltage E_{b1}, and Q will be the quiescent operating point, with grid bias $-E_{c1}$.
4. Through Q draw a loadline with the desired slope from $E_c = 0$ to $E_c = -2E_{c1}$ —this will be AQB. Mark the point of average plate current (D).
5. Select the adjacent characteristic curve ($-E_{c2}$) and arbitrarily select point Q″ having slightly higher plate current than Q. Through Q″ draw a parallel loadline A″Q″B″ from $E_c = 0$ to $E_c = -2E_{c2}$. On this loadline mark the point of average plate current (D″).
6. Join DD″, and mark its point of intersection (D′) with the sloping line through E_{bb}. Through D′ draw a parallel loadline A′B′ which will be the maximum signal loadline. Obviously B′ will be intermediate between B and B″, and its position may be fixed approximately by joining BB″ and making B′ the point of intersection with the new loadline. The rise in plate current from the quiescent condition (I_Q) to the maximum signal condition (I_b') is given by $\Delta I_b'$.

If it is desired to make allowance for the transformer primary resistance, the slope of $D''E_{bb}$ in Fig. 13.8 should be $-1/(R' + R_k)$ where R' = primary resistance.

In general, any single Class A triode may be operated either with fixed or cathode bias as desired. The maximum value of grid circuit resistance frequently depends on the source of bias. When it is permissible to use a value of cathode bias resistor which provides the same grid bias as required for fixed bias, the load resistance, distortion, and power output will be identical. When it is necessary, on account of plate dissipation, to use a higher value of R_k, then the conditions in the two cases will be different.

Cathode by-passing—It is important for the by-pass condenser to be sufficiently large to maintain the bias voltage constant over each cycle—any fluctuation leads to increased distortion and loss of power output. When operating at low levels it is usually sufficient to ensure that the reactance of the by-pass condenser, at the lowest frequency to be amplified, does not exceed one tenth of the resistance of the cathode bias resistor.

At high operating levels it is necessary to ensure that the direct voltage from B + to cathode remains substantially constant over each half-cycle. This may be treated in the same manner as the by-passing of the power supply—see Sect. 1(iv) and Eqn. (1).

The network comprising the resistive elements R_s and R_k and the capacitive elements C and C_k may be replaced by its equivalent network comprising C_1 in parallel with R_1, connected from B+ to cathode. In the special case when $C/C_k = R_k/R_s$ we have $R_1 = R_s + R_k$ and $C_1 = CC_k/(C + C_k)$, but the general case is more involved. Eqn. (1) of Sect. 1 may then be applied directly.

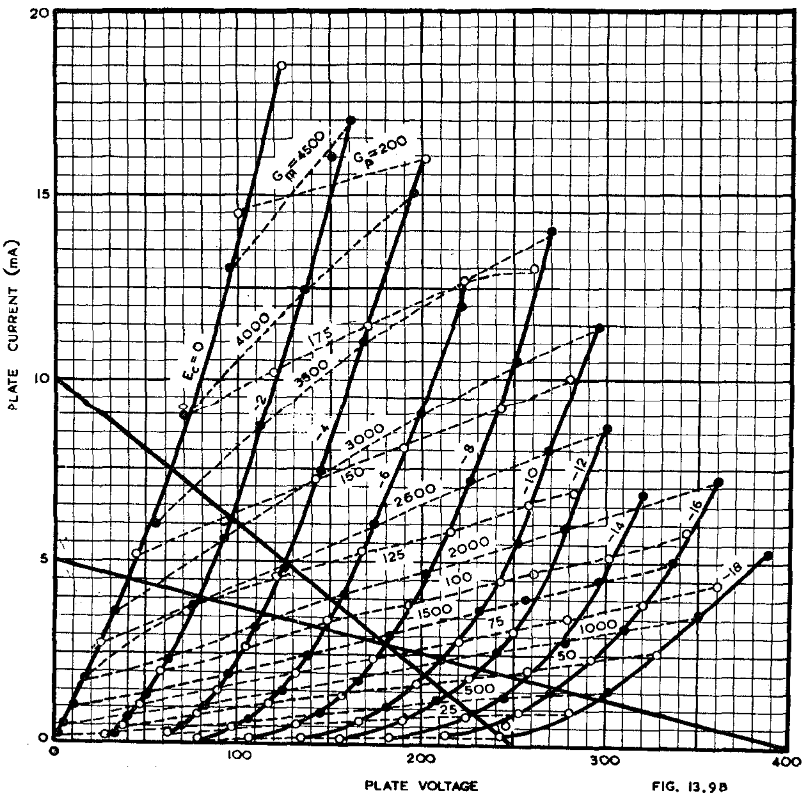


Fig. 13.9B. *Plate characteristics type* 6J5 *with* g_m *and* g_p *curves and loadlines (Ref. B5).*

(F) Alternative method using g_m and g_p curves

$$A = - g_m R_L/(1 + g_p R_L) \tag{19B}$$

where A = amplification with small input signal
g_m = mutual conductance
and g_p = plate conductance = $1/r_p$.

For predominant second harmonic distortion :

$$A' = \tfrac{1}{2}(A_1 + A_2) \tag{19C}$$

Second harmonic distortion per cent = $25(A_1 - A_2)/(A_1 + A_2)$
where A' = amplification with maximum input signal
A_1 = amplification at maximum positive excursion
and A_2 = amplification at maximum negative excursion.

The values of A_1 and A_2 may be calculated, using eqn. (19B), from the values of g_m and g_p corresponding to the relevant points on the loadline which may be derived from Fig. 13.9B for type 6J5. Alternatively, the values of g_m and g_p may be derived by measurement or graphically at the points of maximum positive and negative excursion on the loadline. By this means it is possible to calculate amplification and distortion with high signal inputs from measurements made at low inputs.

This method may be used for cathode followers by using eqn. (19D) in place of (19B) :

$$A = g_m R_k/[1 + (g_m + g_p)R_k] \tag{19D}$$

It may also be used for degenerative amplifiers by using eqn. (19E) :

$$A = - g_m R_L/[1 + (g_m + g_p)R_k + g_p R_L] \tag{19E}$$

References B5, B7, B8, H17.

(iii) Optimum operating conditions

The preceding graphical treatment enables the optimum load resistance to be determined to give maximum power output for limited distortion, provided that valve curves are available. It is frequently desirable to be able to make an approximate calculation without going to so much trouble. The optimum load resistance is a function of the operating conditions (see Refs. A). The following treatment is based on Ref. A14, and relates to " ideal " (linear) valve characteristics.

Case 1 : Grid current and distortion zero, fixed signal input voltage, no limitations on plate current or voltage

See loadline AQB in Fig. 13.10 in which the characteristics are parallel and equidistant straight lines. The loadline may be placed anywhere between the $E_c = 0$ curve and the E_b axis. Maximum power output will be obtained when $R_L = r_p$, that is when the slope of AQB is equal in magnitude to the slope of the characteristic curves. The power output with any load resistance is given by

$$P_0 = \frac{(\mu E_g)^2 R_L}{(r_p + R_L)^2} \tag{20}$$

where E_g = r.m.s. grid input voltage.

Since there is no advantage in having the loadline in the position shown, it may be transferred to the position A′Q′B′ where it has the lowest possible grid bias and plate voltages. Under these conditions the " plate circuit efficiency "* is 16.6% when $R_L = r_p$.

Case 2 : Grid current and distortion zero, fixed plate voltage, no limitations on plate current or signal input voltage

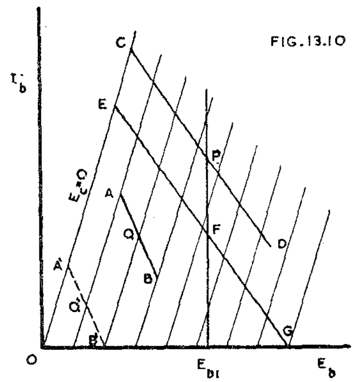
Let E_{b1} be the fixed plate voltage. The loadline may be CPD in Fig. 13.10 where C is the intersection with the $E_c = 0$ curve which is the boundary of the grid current region. PD must equal CP for zero distortion. Maximum power output is obtained when the loadline is adjusted so that its lower extremity touches the E_b axis, as with EFG, and when $R_L = 2r_p$. Under these conditions the plate circuit efficiency* is 25%.

The maximum power output with any load resistance is given by

$$P_0 = \frac{\tfrac{1}{2}E_{b1}{}^2 R_L}{(2r_p + R_L)^2} \tag{21}$$

See also equations (31) to (34) inclusive.

*See Section 2(v).



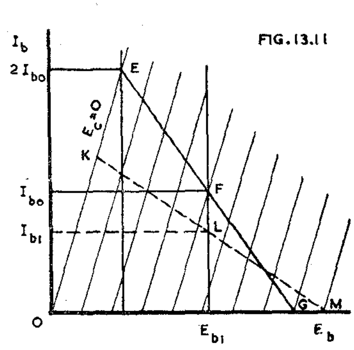


Fig. 13.10. "*Ideal*" *plate characteristics of triode to illustrate choice of optimum loadline.*

Fig. 13.11. "*Ideal*" *plate characteristics of triode with fixed plate voltage to illustrate choice of optimum loadline.*

Case 3 : Grid current and distortion zero, fixed plate voltage and maximum plate current, no limitation on signal input voltage

Let E_{b1} = fixed plate voltage
and I_{b1} = maximum plate current.

It is always advantageous to operate the valve with the maximum plate voltage, which is here regarded as being fixed. There is, however, an optimum value of plate current, and any increase or decrease in plate current results in lower power output. The optimum plate current is given by

$$I_{b0} = \tfrac{1}{4}E_{b1}/r_p \qquad (22)$$

where $R_L = 2r_p$
as illustrated in Fig. 13.11 with loadline EFG.

If the maximum plate current is higher than I_{b0}, then it is preferable to operate the valve at I_{b0} so as to avoid loss of power output and waste of current.

If the maximum plate current is less than I_{b0}, it is necessary to increase the grid bias to reduce the plate current to I_{b1} (point L). One possible loadline is KLM where KL = LM and for this condition

$$R_L = (E_{b1}/I_{b1}) - 2r_p \qquad (23)$$

$$P_0 = \tfrac{1}{2}(E_{b1} - 2I_{b1}r_p)I_{b1} \qquad (24)$$

and the plate circuit efficiency is given by

$$\eta = \tfrac{1}{2}(1 - 2r_pI_{b1}/E_{b1}) \qquad (25)$$

The load resistance for this condition is always greater then $2r_p$.

Case 4 : Grid current and distortion zero, fixed plate dissipation, no limitation on plate voltage, plate current or signal input voltage

The power output under these conditions is given by eqn. (24) which may be put into the form

$$P_0 = \tfrac{1}{2}(P_{pm} - 2I_{b1}{}^2r_p) \qquad (26)$$

where P_{pm} = maximum plate dissipation.

Obviously the power output continues to increase as E_{b1} is increased. The plate circuit efficiency approaches 50% as E_{b1} is made very large. The value of load resistance is given by eqn. (23).

Case 5 : Grid current and distortion zero, fixed plate dissipation, fixed minimum instantaneous plate current

Owing to the existence of " bottom bend " curvature in actual valve characteristics, it is necessary to fix a minimum value of instantaneous plate current if distortion is to be reasonable (Fig. 13.12).

It is here assumed that the portions of the characteristics above $I_{b\,min}$ are straight, and that operation is restricted to the linear region. $E_{b\,min}$ is the plate voltage at which the extended straight portion of the characteristic cuts the axis.

Applying this procedure to each of the previous cases, we have :

(1) No change—$R_L = r_p$.

(2) No change—$R_L = 2r_p$.

(3) $P_0 = \frac{1}{2}[E_{b1} - (2I_{b1} - I_{b\,min})r_p - E_{b\,min}][I_{b1} - I_{b\,min}]$ (27)

The optimum value of load resistance is given by

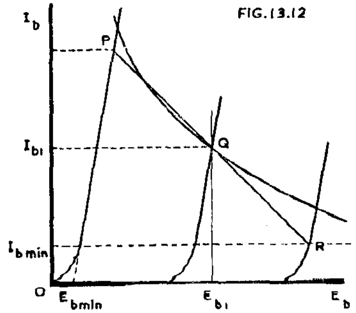
$R_L = (E_{b1}/I_{b1}) - 2r_p + (E_{b1}/I_{b1} - r_p - E_{b\,min}/I_{b\,min})I_{b\,min}/(I_{b1} - I_{b\,min})$ (28)

The plate circuit efficiency is given by

$\eta = \frac{1}{2}(1 - 2r_pI_{b1}/E_{b1} + r_pI_{b\,min}/E_{b1} - E_{b\,min}/E_{b1})(1 - I_{b\,min}/I_{b1})$ (29)

(4) The optimum value of load resistance for maximum power output is given by

$R_L = (4I_{b1}/I_{b\,min} - 1)r_p + E_{b\,min}/I_{b\,min}$ (30)



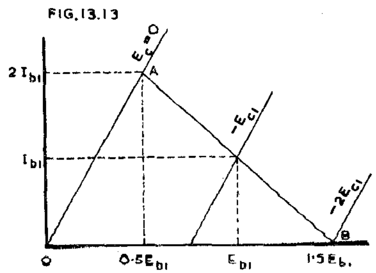


Fig. 13.12. *Plate characteristics of imaginary triode with "ideal" characteristics above $I_{b\,min}$ and curved characteristics below this line.*

Fig. 13.13. *Ideal triode characteristics with $R_L = 2r_p$.*

Summary of operating conditions

(A) Ideal characteristics

If there is no limitation on plate current, the optimum load $R_L = 2r_p$ and the load-line is as Fig. 13.13.

Grid bias : $E_{c1} = 0.75E_{b1}/\mu$ (31)

Plate current : $I_{b1} = E_{b1}/4r_p$ (32)

Max. power output : $P_0 = E_{b1}^2/16r_p$ (33)

Plate circuit efficiency : $\eta = 25\%$ (34)

(B) Practical characteristics

The optimum plate current for ideal characteristics is given by

$I_{b0} = \frac{1}{4}E_{b1}/r_p$

and in practice is slightly higher.

Grid bias for optimum plate current :

$E_{c1} = 0.75(E_b - E_t)/\mu$ (35)

where E_t = plate voltage at intersection of E_b axis and tangent to $E_c = 0$ curve at current I_{max},

and E_{c1} = grid bias for d.c. filament operation (add $E_f/2$ for a.c. operation).

Usual values of E_t are between 30 and 45 volts. Examples are :

Type 2A3 E_t = 35 volts approx.

Type 45 E_t = 40 volts approx.

Note : At voltages above about 180 volts, the grid bias will have to be increased to reduce the plate dissipation.

When a triode is being operated at the optimum plate current, the load resistance should be approximately $2r_p$. Fig. 13.14 shows the variation in power output, and second harmonic distortion, of a typical small power triode indicating that maximum power output occurs at slightly less than $2r_p$. The percentage **third harmonic distortion** is usually between one third and one tenth of the second harmonic distortion up to 5% of the latter ; the ratio decreases as the load resistance is increased.

When a triode is being operated at a plate current which is less than the optimum, then R_L will be greater than $2r_p$ for maximum power output. Refer to Case (3), eqns. 23, 24 and 25.

(iv) Loudspeaker load

A loudspeaker (see Chapters 20 and 21) presents a load impedance which is neither purely resistive nor constant. At most frequencies it causes an elliptical loadline [see Chapter 2 Sect. 4(vi)] with the shape of the ellipse varying widely over the audio frequency range. All that can be said here is that the elliptical loadline results in higher distortion and lower power output than a purely resistive load.

The variation in impedance is almost entirely an increase above the nominal (400 c/s) impedance. As a result, with constant signal voltage applied to the grid, the distortion and the power output decrease as the load resistance is increased (see Fig. 13.14).

A triode applies nearly constant voltage across the load impedance. This is a standard condition of test for a loudspeaker, and some models of loudspeakers are designed to operate under these conditions (see Chapters 20 and 21). A triode is almost the ideal output stage for a loudspeaker load when looked at from the load point of view, with or without feedback.

See also Sect. 3(viii) for pentodes, Sect. 11(iii) and Fig. 13.54, and Chapter 21 Sect. 1.

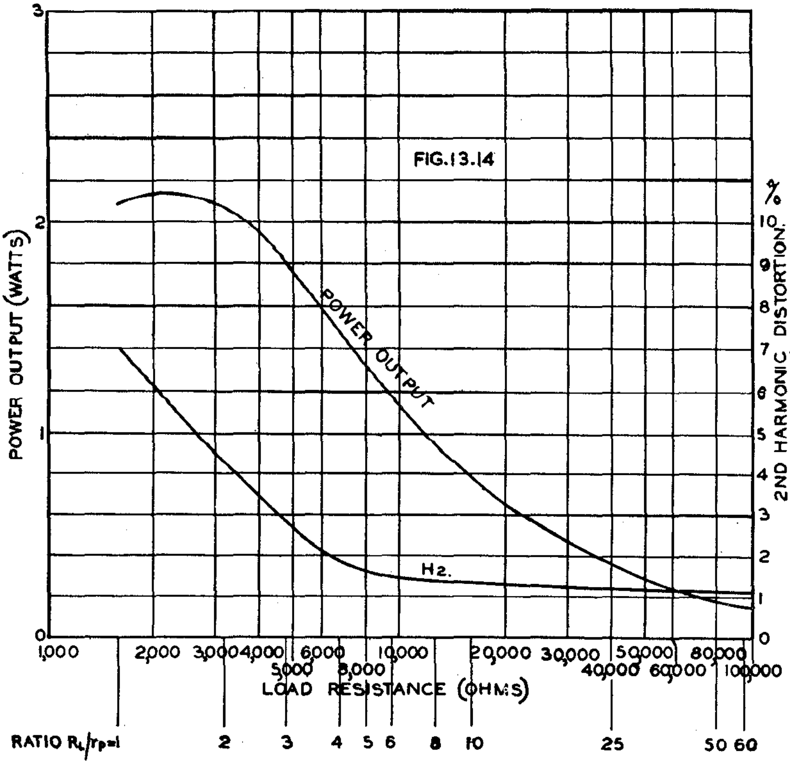


Fig. 13.14. Power output and distortion of type 45 triode plotted against load resistance. The curves were derived graphically, with allowance for shifting loadline. The operating conditions are : $E_b = 250$ volts, $E_c = -50$ volts, peak grid signal voltage 50 volts, $I_b = 34$ mA (zero signal), $r_p = 1610$ ohms.

(v) Plate circuit efficiency and power dissipation

The plate circuit efficiency is defined as the ratio of the maximum signal frequency power output to the d.c. power input under these conditions. (Owing to rectification, the d.c. power input under maximum signal conditions is usually greater than with zero signal input).

The plate circuit efficiency cannot exceed 50% in any Class A amplifier, and usually does not exceed 25% with Class A triodes. The plate circuit efficiency in an ideal Class A triode is given by

$$\eta = R_L/(2R_L + 4r_p) \qquad (36)$$

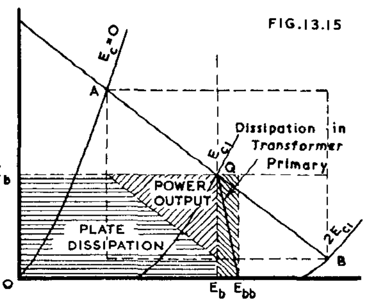
The power relationships in a Class A triode are shown in Fig. 13.15 on the assumption that no second harmonic or rectification effects are present, and that fixed bias is employed. Allowance is made for the power loss in the d.c. resistance of the transformer primary. The plate-to-cathode voltage is E_b, but the total supply voltage is E_{bb}.

The plate dissipation in a Class A triode is greatest with zero input signal. If there are no rectification effects, the power input remains constant and the power output is simply power transferred from heating the plate to useful output. With normal rise of plate current for 5% second harmonic distortion, the plate dissipation is more nearly constant, but it is still sufficient to base the design on the zero-signal condition.

The quiescent operating point may be limited by one or more of the following:

1. Maximum plate dissipation.
2. Maximum plate voltage.
3. Maximum plate current.

Fig. 13.16 is an example including all three. The dissipation curve may be plotted from the relationship $I_b = P_0/E_b$. The horizontal line AB is the maximum plate current, while the vertical line CD is the maximum plate voltage. The quiescent operating point must not be outside the area ABCD.



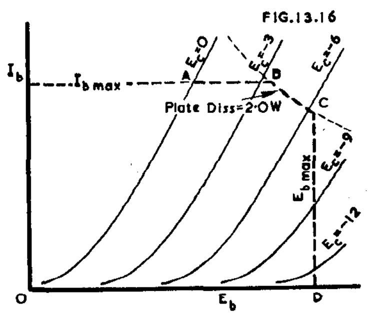


Fig. 13.15. *Power output and dissipation relationships in a Class A triode.*
Fig. 13.16. *Plate characteristics of small battery operated power triode with limits of maximum plate dissipation, maximum plate voltage and maximum plate current.*

(vi) Power sensitivity

A triode suffers from low power sensitivity when compared with pentodes and beam power amplifiers, but the difference becomes smaller when a higher degree of negative feedback is applied to the latter. In any case, it is usually a simple matter to provide sufficient gain in the voltage amplifier.

Maximum power sensitivity is obtained when $R_L = r_p$, although this limits the maximum power output for a fixed value of distortion.

See Sect. 1(iii) for the form in which the sensitivity is expressed.

(vii) Choke-coupled amplifier

The choke-coupled amplifier of Fig. 13.17 may be used as an alternative to the transformer-coupled arrangement of Fig. 13.1, and the foregoing discussion may be applied exactly on the understanding that

1. the choke inductance and resistance correspond to the transformer primary inductance and resistance,
2. the condenser C offers negligible impedance to signal frequencies compared with R_1.

(viii) Effect of a.c. filament supply

The curves of power triodes are usually drawn for the condition with d.c. filament supply and with the negative filament terminal regarded as the cathode. When a.c. is applied to the filament, roughly the same performance is achieved by increasing the grid bias by $E_f/2$. The increase is taken as 1.5 volts for a 2.5 volt filament.

FIG.13.17

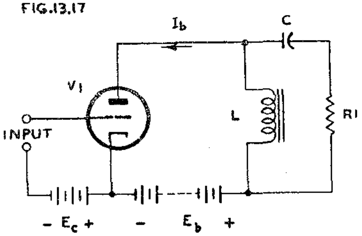


Fig. 13.17. *Choke-coupled amplifier.*

FIG.13.18

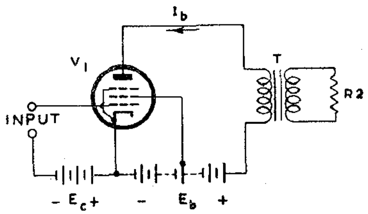


Fig. 13.18. *Basic circuit of power pentode.*

(ix) Overloading

A triode has no " cushioning " effect such as occurs with a pentode when the peak signal voltage approaches the bias voltage. There is therefore no warning of the impending distortion which occurs when grid current flows. The distortion may be minimized by the use of a low-plate-resistance triode in the preceding stage, a low grid circuit resistance or a grid stopper resistor (10 000 ohms is normally satisfactory, but values from 1000 to 50 000 ohms have been used).

(x) Regulation and by-passing of power supply

The regulation and by-passing of the power supply affect the minimum frequency which can be handled satisfactorily at full power output—see Sect. 1(iv).

SECTION 3 : CLASS A MULTI-GRID VALVES

(i) Introduction (ii) Ideal pentodes (iii) Practical pentodes—operating conditions (iv) Graphical analysis—power output and distortion (v) Rectification effects (vi) Cathode bias (vii) Resistance and inductance of transformer primary (viii) Loudspeaker load (ix) Effects of plate and screen regulation (x) Beam power valves (xi) Space charge tetrodes (xii) Partial triode operation of pentodes.

(i) Introduction

Multi-grid valves include pentodes, beam power amplifiers and similar types, and space-charge valves. All of these have a family resemblance, in that they have higher plate circuit efficiencies and greater power sensitivities than triodes, but on the other hand they have higher distortion, particularly odd harmonic distortion. Their advantages over triodes generally outweigh their disadvantages, particularly as the distortion may be minimized by negative feedback, and they are almost exclusively used in ordinary commercial radio receivers. Reference should be made to Chapter 7 for amplifiers incorporating negative feedback.

The basic circuit of a power pentode with transformer-coupled load is Fig. 13.18. A choke-capacitance coupled load may also be used, as for a triode (Fig. 13.17). The only essential difference from a triode is that provision must also be made for a constant voltage to be applied to the screen. In some cases the screen is operated at the same voltage as the plate, so as to avoid a separate screen voltage supply.

In a practical circuit the screen, the plate return and the grid return should each be by-passed to the cathode.

(ii) Ideal pentodes

An ideal pentode is one having infinite plate resistance, a 90° angular knee and equally spaced characteristic curves (Fig. 13.19). The zero bias characteristic is OAC, the $-E_{c1}$ characteristic is ODE, while the $-2E_{c1}$ characteristic is OB. The operating point is Q, the optimum loadline is AQB where AQ = QB and the distortion is zero.

The following may readily be derived—

Optimum $R_L = E_b/I_b$ (1)

Then for optimum load resistance :

$P_0 = \frac{1}{2}E_bI_b$ (2)

D.C. power input $= E_bI_b$ (3)

Plate circuit efficiency $= \frac{1}{2}E_bI_b/E_bI_b = 50\%$ (4)

$E_{c1} = E_{c2}/2\mu_t$ (5)

where E_{c2} = screen voltage

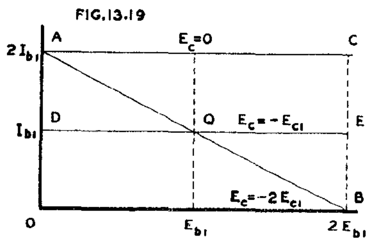
and μ_t = triode amplification factor $= \mu_{g1g2}$.

For any value of load resistance :

$P_0 = \frac{1}{2}I_b^2R_L$ when R_L is less than E_b/I_b (6)

$P_0 = \frac{1}{2}E_b^2/R_L$ when R_L is greater than E_b/I_b (7)

(It is assumed here that the signal voltage is reduced for other than optimum R_L in order to avoid distortion.)



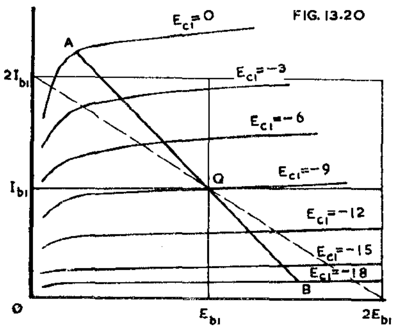


Fig. 13.19. *Plate characteristics of ideal pentode.*

Fig. 13.20. *Plate characteristics of typical pentode compared with ideal characteristics.*

(iii) Practical pentodes—operating conditions

A practical pentode has characteristics of the form of Fig. 13.20 on which has been superimposed the "ideal characteristics" for the same plate voltage and current.

(A) The **load resistance** corresponding to the loadline AQB is slightly lower than the ideal value E_b/I_b. A good general rule for all pentodes is

$R_L \approx 0.9\ E_b/I_b$ (8)

which is usually correct within ± 10% for maximum power output. Eqn. (8) is safe to use under all circumstances, although an adjustment may be made experimentally or graphically to secure the best compromise between distortion and power output.

(B) The **power output** is always less than $\frac{1}{2}E_bI_b$, and may be calculated graphically [see (iv) below].

(C) Owing to rectification effects, the direct **plate current** may rise or fall from zero signal to maximum signal [see (v) below].

(D) The **plate circuit efficiency** is usually between 28% and 43% for total harmonic distortion less than 10%. Of course, the screen current is wasted, so that the total plate + screen circuit-efficiency is lower. The plate dissipation is equal to the d.c. watts input less the signal power output. The screen dissipation is equal to the d.c. screen input.

(E) **Screen current** : In a pentode, if the control grid bias is kept constant and only the plate voltage varied, the total cathode current (plate + screen) will remain nearly constant, decreasing slightly as the plate voltage is reduced down to the knee of the curve. Below this plate voltage the screen current increases more rapidly until zero plate voltage is reached, at which point the screen current is a maximum (Fig. 13.21). It is evident therefore that if a dynamic loadline cuts the zero bias curve below the knee, the screen current will rise rapidly and the screen dissipation may be exceeded. The average maximum-signal screen current may be calculated from the approximation

$$I_{c2\,av} \approx \tfrac{1}{4}I_A + \tfrac{1}{2}I_Q \qquad (9)$$

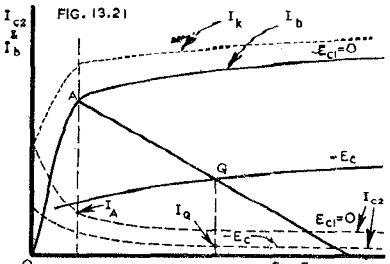
where I_A = screen current at minimum plate voltage swing and zero bias (point A),
and I_Q = screen current at no signal and normal bias.

The screen dissipation is therefore P_{g2} where

$$P_{g2} = E_{c2}(\tfrac{1}{4}I_A + \tfrac{1}{2}I_Q) \qquad (10)$$

The variation of screen current with change of control grid voltage is such that the ratio between plate and screen currents remains approximately constant provided that the plate voltage is considerably higher than the knee of the curve. This ratio may be determined from the published characteristics.

A pentode, or beam power tetrode may be used as an amplifier with the plate voltage in the region of the knee of the curve of the $E_c = 0$ characteristic, that is somewhere about one fifth of the screen voltage, provided that care is taken to keep both plate and screen dissipations within their ratings. This gives operation very similar to that of a triode with a non-critical and low value of load resistance, which may be useful in some special applications (Ref. C6). This method of operation is not here described in detail.



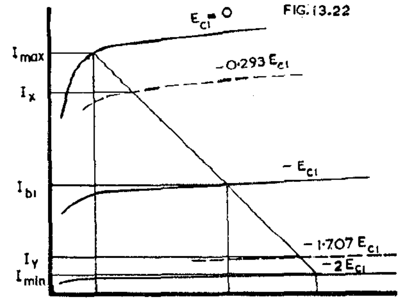


Fig. 13.21. *Plate, screen and cathode current characteristics of pentode.*
Fig. 13.22. *Calculation of pentode power output and distortion using the five selected ordinate method.*

(F) The screen should be supplied from a voltage source of good regulation ; a voltage divider, whose "bleed" current is very much higher than the maximum screen current under dynamic conditions, is the most common source when the screen voltage is lower than the plate voltage.

(G) The **voltage gain** is determined as for a triode [Section 2(i) eqn. 8].

(H) If **oscillation or parasitics** are experienced, a grid stopper resistance of about 500 ohms may be used. A small mica condenser from the plate terminal to earth is often beneficial. The bias resistor and its electrolytic by-pass condenser may be by-passed by a 0.001 μF mica condenser. The shells of metal valves should be earthed directly. The length of leads should be as short as possible, particularly between the valve terminals and their by-pass condensers.

With type 807, a stopper resistor of 100 ohms (non-inductive) may be connected directly to the screen terminal of the valve, and a by-pass condenser (0.01 μF mica) taken from the end of the resistance remote from the screen, directly to earth.

(I) **Overload characteristics**—Owing to the crowding together of the characteristics at and below the knee of the curves, a pentode tends to give a smoother overload characteristic than a triode. There is no point at which the distortion begins sharply, but rather a gradual flattening effect on the top, or top and bottom, of the signal current wave. Pentodes or beam power amplifiers with negative feedback lose most of this " cushioning effect," and more closely resemble triodes as regards overload. The distortion due to grid current may be minimized by a low impedance (triode) preceding stage and a fairly low value of grid resistor ; this will then leave the distortion due to the characteristics as the predominant feature, and provide a reasonably smooth overload.

(iv) Graphical analysis—power output and distortion

The choice of loadline for a pentode is usually made by firstly selecting the screen voltage and then selecting a convenient plate voltage. The third step is to note the grid voltage which allows a small plate current to flow, and to divide this voltage by 2 to obtain the grid voltage for the working point—thus determining the quiescent operating point Q. A scale is then swung around Q (a small pin is helpful as a pivot) until the two parts of the loadline are equal. If the scale is calibrated in millimetres, the 10 cm. calibration may be held at Q and the two parts compared directly by eye. When the two parts are equal, the second harmonic distortion will be zero, and the loadline is a first approximation for maximum power output.

In the case of some pentodes, and particularly beam power amplifiers of the 6L6 class, the loadline for zero second harmonic is obviously far from the knee of the curve In such a case the loadline should be taken to the knee of the curve, if sharply defined.

To determine the exact loadline for maximum power output, it is necessary to take several angles, and to calculate the power output for each. The final choice of loadline is usually a compromise between power output and distortion ; in such a case the load resistance is always less than the one giving maximum power output.

Load resistances higher than that giving maximum power output are to be avoided because they give greater harmonic distortion, less power output and higher screen current. It is better to err on the low side, particularly with nominal loudspeaker loads.

Owing to the presence of appreciable percentages of third and higher order odd harmonics, the formulae for triodes are not suitable for pentodes. There are several methods for pentodes, the choice depending on whether third harmonic only is required, or third and higher harmonics. Reference should be made to Chapter 6 Sect. 8(iii) " Graphical Harmonic Analysis."

The treatment here is based on a working loadline. If the amount of second harmonic is very small, the rise of plate current due to rectification may be neglected. If there is an appreciable amount of second harmonic, it is necessary to use the method of Sect. 3(v) for drawing the corrected loadline ; the formulae below should then be applied to the corrected loadline.

(A) **Five selected ordinate method** (For second and third harmonics only)

This method makes use of the plate currents corresponding to grid voltages -0.293 and $-1.707\ E_{c1}$ in addition to the three currents used with Class A_1 triodes. If these are all available, the loadline of Fig. 13.22 is all that is required—if not, it will be necessary to plot the dynamic characteristic as in Fig. 13.24.

$$P_0 = [I_{max} - I_{min} + 1.41\ (I_X - I_Y)]^2 R_L/32 \tag{11}$$

$$R_L = (E_{max} - E_{min})/(I_{max} - I_{min}) \tag{12}$$

$$\%\ \text{2nd harmonic} = \frac{I_{max} + I_{min} - 2I_{b1}}{I_{max} - I_{min} + 1.41\ (I_X - I_Y)} \times 100 \tag{13}$$

$$\%\ \text{3rd harmonic} = \frac{I_{max} - I_{min} - 1.41\ (I_X - I_Y)}{I_{max} - I_{min} + 1.41\ (I_X - I_Y)} \times 100 \tag{14}$$

where I_X = plate current at $-0.293\ E_{c1}$
I_Y = plate current at $-1.707\ E_{c1}$.

Eqns. (11) to (14) are exact provided that there is no harmonic higher than the third. The power output is that for fundamental frequency only, but the power output for 10% harmonic distortion is only 1% of the fundamental power output.

The presence of third harmonic distortion has a flattening effect on both the positive and negative peaks, thus increasing the power output for a limited value of $I_{max} - I_{min}$. For example, 10% third harmonic reduces the value of I_{max} by 10% while it increases the total power output by 1%. If there were no distortion, and the value of I_{max} were maintained at 90%, the power output would be only 81%. Thus the ratio of power output with and without 10% third harmonic distortion is 101/81 = 1.25, on condition that $I_{max} - I_{min}$ is kept constant. In simple language, for the same plate current swing, 10% third harmonic distortion increases the total power output by 25%. This will be modified by the presence of fifth and higher harmonics [see iv(D) below ; Sect. 7(iii)].

(B) **Five equal-voltage ordinate method** (Espley)

For second, third and fourth harmonics only.

The previous method requires characteristics for -0.293 and $-1.707\ E_{c1}$, which are often not directly available. The following method is usually more convenient and is exact provided that all harmonics above H_4 are zero (see Fig. 13.23 for symbols).

$$P_0 = (I_0 - I_{2\cdot 0} + I_{0\cdot 5} - I_{1\cdot 5})^2 R_L/18 \tag{15}$$

$$\%\ \text{2nd harmonic} = \frac{3(I_0 - 2I_{1\cdot 0} + I_{2\cdot 0})}{4(I_0 + I_{0\cdot 5} - I_{1\cdot 5} - I_{2\cdot 0})} \times 100 \tag{16}$$

$$\%\ \text{3rd harmonic} = \frac{(I_0 - 2I_{0\cdot 5} + 2I_{1\cdot 5} - I_{2\cdot 0})}{2(I_0 + I_{0\cdot 5} - I_{1\cdot 5} - I_{2\cdot 0})} \times 100 \tag{17}$$

$$\%\ \text{4th harmonic} = \frac{(I_0 - 4I_{0\cdot 5} + 6I_{1\cdot 0} - 4I_{1\cdot 5} + I_{2\cdot 0})}{4(I_0 + I_{0\cdot 5} - I_{1\cdot 5} - I_{2\cdot 0})} \times 100 \tag{18}$$

$$\text{D.C. plate current} = \frac{1}{6}(I_0 + 2I_{0\cdot 5} + 2I_{1\cdot 5} + I_{2\cdot 0}) \tag{19}$$

A third harmonic scale may be prepared for reading the third harmonic percentage (Ref. A14 pp. 71-72).

(C) **Seven equal-voltage ordinate method** (Espley)

This is exact for harmonics up to the sixth, provided that higher harmonics are zero. It is sometimes more convenient than the five ordinate method, when there are no -0.5 and $-1.5\ E_{c1}$ curves, even when the higher harmonics are of no interest.

The symbols have the same significance as in Fig. 13.23 with the subscript indicating the grid voltage.

$$P_0 = (167I_0 + 252I_{0\cdot 33} - 45I_{0\cdot 67} + 45I_{1\cdot 33} - 252I_{1\cdot 67} - 167I_{2\cdot 0})^2 R_L/819\,200 \tag{20}$$
$$H_2\% = 25(559I_0 + 486I_{0\cdot 33} - 1215I_{0\cdot 67} + 340I_{1\cdot 0} - 1215I_{1\cdot 33} + 486I_{1\cdot 67} + 559I_{2\cdot 0})/I \tag{21}$$
$$H_3\% = 250(45I_0 - 36I_{0\cdot 33} - 63I_{0\cdot 67} + 63I_{1\cdot 33} + 36I_{1\cdot 67} - 45I_{2\cdot 0})/I \tag{22}$$
$$H_4\% = 450(17I_0 - 42I_{0\cdot 33} + 15I_{0\cdot 67} + 20I_{1\cdot 0} + 15I_{1\cdot 33} - 42I_{1\cdot 67} + 17I_{2\cdot 0})/I \tag{23}$$
$$H_5\% = 4050(I_0 - 4I_{0\cdot 33} + 5I_{0\cdot 67} - 5I_{1\cdot 33} + 4I_{1\cdot 67} - I_{2\cdot 0})/I \tag{24}$$
$$H_6\% = 2025(I_0 - 6I_{0\cdot 33} + 15I_{0\cdot 67} - 20I_{1\cdot 0} + 15I_{1\cdot 33} - 6I_{1\cdot 67} + I_{2\cdot 0})/I \tag{25}$$

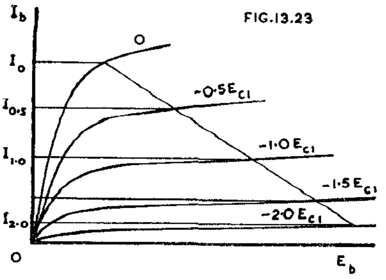
where $I = 167I_0 + 252I_{0\cdot 33} - 45I_{0\cdot 67} + 45I_{1\cdot 33} - 252I_{1\cdot 67} - 167I_{2\cdot 0}$.

(D) **Eleven selected ordinate method**

To use this method, it is first necessary to plot the loadline on the plate characteristics, and then to transfer it to a dynamic mutual characteristic (Fig. 13.24) from which the required values of plate current may be derived.

$$P_0 = \tfrac{1}{2}(0.5I_0 - 0.5I_{2\cdot 0} + I_{hm3} - I_{hm5})^2 R_L \tag{26}$$
$$I_{h1} = 0.5I_0 - 0.5I_{2\cdot 0} + I_{hm3} - I_{hm5} = \text{fundamental} \tag{27}$$
$$I_{h2} = \tfrac{1}{4}(I_0 + I_{2\cdot 0} - 2I_{1\cdot 0}) \tag{28}$$
$$I_{h3} = 0.167(2I_{0\cdot 5} + I_{2\cdot 0} - I_0 - 2I_{1\cdot 5}) \tag{29}$$
$$I_{h4} = \tfrac{1}{8}(I_0 + 2I_{1\cdot 0} + I_{2\cdot 0} - 2I_{0\cdot 3} - 2I_{1\cdot 7}) \tag{30}$$
$$I_{h5} = 0.1(2I_{0\cdot 7} + I_0 + 2I_{1\cdot 8} - 2I_{0\cdot 2} - 2I_{1\cdot 3} - I_2) \tag{31}$$

Percentage second harmonic $= (I_{hm2}/I_{hm1}) \times 100$



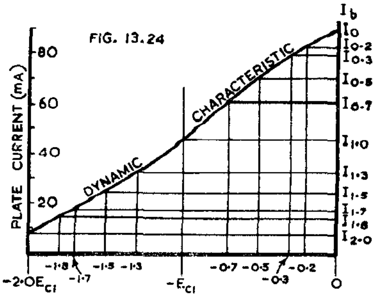


Fig. 13.23. *Calculation of pentode power output and distortion using the five equal-voltage ordinate method.*
Fig. 13.24. *Calculation of pentode power output and distortion using the eleven selected ordinate method.*

Percentage third harmonic $= (I_{hm3}/I_{hm1}) \times 100$ etc.

Note that all currents, including I_{hm1}, I_{hm2} etc., are peak values. The exact values of grid voltage are 0, 0.191, 0.293, 0.5, 0.691, 1.0, 1.309, 1.5, 1.707, 1.809 and 2.0. The approximate values are, however, sufficiently accurate for most purposes.

Eqn. (26) shows that third harmonic distortion adds to the power output, while fifth harmonic subtracts from the power output as calculated for distortionless conditions. See also (iv)A above ; Sect. 7(iii).

For theoretical basis see Chapter 6 Sect. 8(iii).

(v) Rectification effects

The general effects are the same as for a triode except that it is sometimes possible with a pentode to have zero second harmonic and no loadline shift. In all other cases it is necessary to determine the corrected loadline before applying harmonic analysis or calculating power output. The method of deriving the corrected loadline is the same for pentodes as for triodes [Sect. 2(ii)D]. For example see Fig. 13.28.

With a triode, the loadline always shifts upwards into such a position that the distortion is less than it would otherwise be. With a pentode, the loadline shifts upwards when the load resistance is lower than a certain value, and shifts downwards when the load resistance is higher than this value. When a pentode loadline shifts downwards it causes increased distortion or decreased power output.

(vi) Cathode bias

The operation of cathode bias is the same as with triodes [Section 2(ii)E] except that the total current flowing through the cathode bias resistor is the sum of the plate and screen currents [Chapter 2 Section 4(v)]. If the screen is supplied from a voltage divider which is returned to the cathode, the bleed current must be added to the cathode current in calculating R_k. This tends to stabilize the bias voltage. Even in cases where there is no rise of plate current at maximum signal, there will always be a rise in screen current [Section 3(iii)E].

When a tentative value of R_k has been determined for maximum signal conditions, it is necessary to check both plate and screen dissipations at zero signal. If the plate and screen voltages are equal, the simplest method is the use of the " triode " mutual characteristics, if available [Chapter 2 Sect. 4(v)].

If there is a " bleed " current through R_k, the procedure is shown in Fig. 13.25 where the triode mutual characteristic is shown in the upper part, with OA representing the bleed current in the lower part. The cathode loadline is drawn from A (instead of from 0) with a slope of $-1/R_k$ where R_k is the value determined for maximum signal conditions.

The no-signal grid bias is $-E_{c1}$ and the cathode current $(I_b + I_{c2})$ is I_{k1}. The dissipation on plate and screen is given by

$$P_p = I_{k1}E_b m/(m+1)\ ;\ P_{g2} = I_{k1}E_{c2}/(m+1) \qquad (32)$$

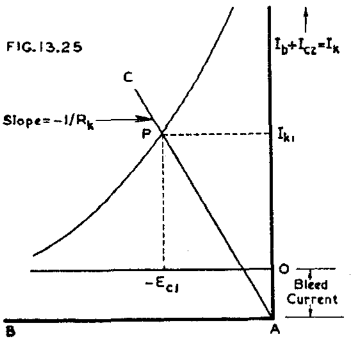
where m = ratio of plate to screen currents $= I_b/I_{c2}$.

The rectification effects with cathode bias may, like triodes (Fig. 13.8) be determined from the plate characteristics but an approximation is involved since $D''E_{bb}$ must have a slope of $-m/R_k(m+1)$ where $m = I_b/I_{c2}$. The value of I_{c2} may be taken as that for maximum signal, and a slight error will then be introduced when deriving the zero signal condition. The cathode bias loadline should, ideally, be drawn as two loadlines—one for zero signal, and the other for maximum signal. For most purposes, however, the maximum signal cathode loadline may be used for both conditions. The zero bias lines are, in each case, above the lines for maximum signal, and the difference in slope is of the order of 5% to 10%.

In general, any Class A_1 pentode may be operated either with fixed or cathode bias, as desired, except for the limitations on the maximum grid circuit resistance [see Sect. 10(i)]. It occasionally happens that a condition is permissible only with fixed bias, owing to the rise in plate dissipation at zero signal.

For back bias, further details regarding fixed bias, and grid circuit resistance see Sect. 10.

For cathode by-passing see Sect. 2(ii).



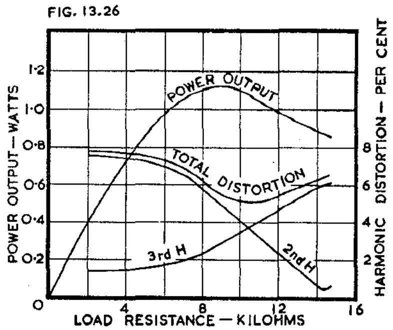


Fig. 13.25. *Cathode loadline of pentode with additional bleed current passing through the cathode resistor, plotted on mutual characteristics.*

Fig. 13.26. *Power output and harmonic distortion of typical pentode plotted against load resistance. Valve type 6AK6,* $E_b = E_{c2} = 180$ *V,* $E_{c1} = -9$, *peak signal* $= 9$ *volts.*

(vii) Resistance and inductance of transformer primary

The **resistance** of the transformer primary may be allowed for in the same way as with triodes [Sect. 2(ii)A]. If R' is the resistance of the primary, and R_k is the cathode bias resistance, then the slope of the line QE_b in Fig. 13.6 should be $-1/(R' + R_k)$.

The **transformer primary inductance** may be based on Chapter 5 Sect. 3(iii)c.

For the same load impedance and the same high frequency attenuation, pentodes may have higher transformer **leakage inductance** than triodes if frequency response is the only criterion. However, owing to the distortion with reactive loads at high output levels, it is very desirable to maintain the leakage inductance as low as practicable, particularly in push-pull amplifiers.

(viii) Loudspeaker load

A pentode is, unfortunately, critical in its load resistance for both maximum power output and distortion. At low operating levels the output power rises steadily as the load resistance is increased up to the value $R_L = r_p$ provided that the grid input voltage is maintained constant (the low operating level is to avoid overloading under any conditions). A loudspeaker has pronounced impedance peaks at the bass resonant frequency and at high audio frequencies. When a loudspeaker is operated at a low level, the acoustical output is accentuated at the bass resonant frequency and at high audio frequencies. The latter may be reduced to any desired degree by a shunt filter—a resistance R in series with a capacitance C. If R is variable, the combination is the simplest form of tone control. If R is fixed, typical values are :

$R = 1.3R_L$; $C = 0.025\mu F$ for $R_L = 5000$ ohms.

This does not affect the rise of impedance at the bass resonant frequency, which is a function of the loudspeaker design and the type of baffle—see Chapter 20 Sect. 2(iv) and Sect. 3.

At maximum signal voltage the conditions are somewhat different (Fig. 13.26). The power output in this case reaches a maximum at $R_L = 9000$ ohms. The second harmonic reaches a minimum (practically zero) at about $R_L = 14\,000$ ohms, and then rises steadily ; actually it undergoes a change of phase near $R_L = 14\,000$ ohms. The third harmonic rises all the way from zero to the limit of the graph. Minimum " total distortion " occurs at $R_L = 10\,000$ ohms, which is the published typical load, being a close approach to maximum power output. In this particular case the load resistance for zero second harmonic is not that for maximum power output.

In the case of a loudspeaker load, the load resistance may rise from the nominal value to (say) 6 or 8 times this value ; all the variation is in the upwards direction. If full signal voltage is maintained for all frequencies, the distortion will be very severe and the maximum power output will be reduced at low and high frequencies. The only methods of minimizing the trouble are the use of a loudspeaker and baffle with less prominent impedance peaks, and the use of negative voltage feedback (see Chapter 7) or reduced signal voltage on the grid. A pentode, operating well below its nominal power output, is capable of giving reasonable fidelity even on an ordinary loudspeaker load. In a normal radio receiver, a power pentode with a nominal maximum power output of 4 or 5 watts can give reasonable fidelity up to somewhat over 1 watt, but it has the advantage of being capable of delivering its full power output when the distortion can be tolerated.

The effect of these high impedance loads, which are here assumed to be purely resistive for the purpose of illustration, is shown by the beam power amplifier plate characteristics in Fig. 13.27. The effect may be minimized by reducing R_L slightly below the optimum value.

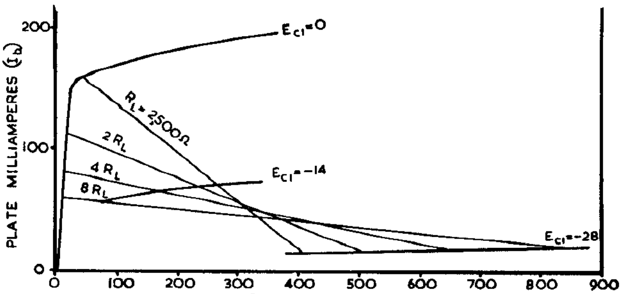


FIG. 13.27

Fig. 13.27. *Plate characteristics of* 6L6 *or* 807 *beam tetrode with loadlines of optimum resistance also twice, four times and eight times optimum. The loadlines have been corrected for the rectification effect.*

The impedance of a loudspeaker is, however, far from being resistive (see Chapters 20 and 21), having a reactive component varying with frequency, which must be combined with the shunt reactance of the transformer primary at low frequencies and that of the shunt capacitance from plate to earth at high frequencies. The combined reactive components increase the distortion and reduce the power output [see also Chapter 5 Sect. 3(iii)c and Chapter 2 Sect. 4(vi) ; Ref. C4].

The published values of power output apply to highly efficient output transformers. The available power output from the secondary of a normal power transformer is equal to η times the published value, where η is the efficiency [see Chapter 5 Sect. 2(ii) and Sect. 3(vi)]. Typical efficiencies are from 70% to 95% (depending on the price class) for well-designed transformers.

(ix) Effects of plate and screen regulation

The internal resistances of plate and screen supply sources cause a reduction in power output. If the regulation of the power supply is such that the rise of plate and screen currents causes a decrease of 1% in all the electrode voltages, the decrease in power output (by the use of conversion factors) will be approximately 2.5%. This will be modified by the shape of the characteristics and rectification effects, and the only accurate method is the graphical one outlined below. See also remarks on push-pull—Sect. 6(iii).

In addition, the regulation and by-passing of the power supply also affect the minimum frequency which can be handled satisfactorily at full power output—see Sect. 1(iv).

Graphical method (Ref. B4).

Let R_1 = internal resistance of plate supply source,
R_2 = internal resistance of screen supply source,
ΔI_b = increase in plate current from zero to maximum signal, with constant electrode voltages,
ΔI_{c2} = increase in screen current, with constant electrode voltages,
ΔE_b = change in plate voltage,
ΔE_{c1} = change in screen voltage,
E_b' = plate voltage at maximum signal,
and E_{c2}' = screen voltage at maximum signal.

Then $E_b' = E_b - \Delta E_b$ (33) $E_{c2}' = E_{c2} - \Delta E_{c2}$ (34)
where $\Delta E_b = R_1 \Delta I_b$ (35) $\Delta E_{c2} = R_2 \Delta I_{c2}$ (36)

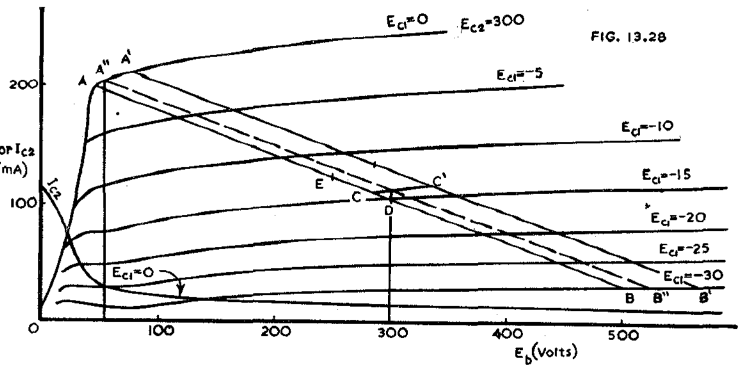


Fig. 13.28. *Plate characteristics of 6L6 or 807 illustrating method of correcting the loadline for rectification. The corrected loadline is A″ B″.*

On the plate characteristics (Fig. 13.28) AB is the uncorrected loadline with C as its average current point, A′B′ is a parallel loadline with C′ as its average current point, A″B″ is the corrected loadline. The average current may be determined from the point on A″B″ which has a plate voltage 300 volts ; from this the rise of current ΔI_b may be calculated. The value of ΔI_{c2} may be determined from eqn. (9), ΔE_{c2} may be calculated by eqn. (36) and E_{c2}' by eqn. (34). In the example of Fig. 13.28 let it be assumed that the plate, screen and grid voltages are all reduced proportionally, F_e* being 0.9 :

$E_b = E_{c2} = 300$ volts $E_b' = E_{c2}' = 270$ volts
$E_{c1} = -15$ volts $E_{c1}' = -13.5$ volts

The procedure is :

1. Plot a new curve for $E_{c1} = 0$, $E_{c2}' = 270$ volts, by the use of conversion factors*. This is drawn on Fig. 13.29. For $F_e = 0.9$, $F_i = 0.86$.
2. Plot a curve for $E_{c1}' = -13.5$, $E_{c2}' = 270$ volts similarly. This will be the -15 volt curve (for $E_{c2} = 300$ volts) with the current ordinates multiplied by the factor 0.86.

*For conversion factors see Chapter 2 Sect. 6 and Fig. 2.32A.

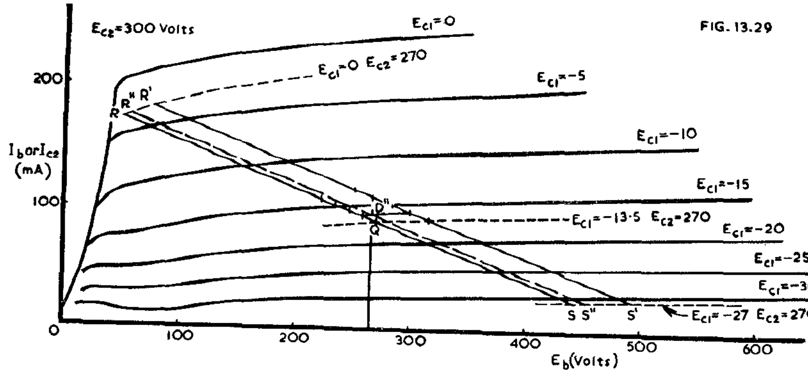


Fig. 13.29. *Plate characteristics of* 6L6 *or* 807 *giving graphical method for deriving the effect of internal resistance in the voltage supply source.*

3. Plot a curve for $E_{c1}' = -27$, $E_{c2}' = 270$ volts. This will be the -30 volt curve (for $E_{c2}' = 300$ volts) with the current ordinates multiplied by 0.86.
4. Mark the point Q at $E_b = 270$, $E_{c2} = 270$, $E_{c1} = -13.5$ volts.
5. Through Q draw the uncorrected loadline RQS.
6. Draw the corrected loadline R″D″S″ by the same method as previously (Fig. 13.28).
7. Determine the rise of plate current from Q to D″ (intersection of R″S″ with $E_b = 270$ volts). This should be approximately the same as derived in the first case (ΔI_b). If there is an appreciable error, some adjustment should be made.

(x) Beam power valves

Beam power amplifier valves, otherwise known as output tetrodes, are in two principal classes. The first class includes most of the smaller valves, which have characteristics so similar to pentodes that they may be treated as pentodes in all respects. The second class includes types such as 6L6, 807 and KT66, which differ from pentodes principally in having sharper " knees " to their plate characteristics, more second harmonic but less third harmonic distortion. The optimum load resistance is more critical than with ordinary power pentodes.

The screen currents of many types of beam power valves, due to variations in grid alignment, may have considerably greater tolerances than in pentodes—up to ± 100% in some instances—and screen dropping resistances should not be used unless recommended by the valve manufacturer. The screen supply should normally be obtained from a voltage divider.

The distortion and power output of type 6L6 beam power valve are plotted against load resistance in Fig. 13.30. The second harmonic is 9.6% at the rated load resistance, the third harmonic only 2.4% and all higher harmonics negligible. At lower load resistances the second harmonic rises, although not seriously, the third harmonic decreases steadily, and all higher harmonics are negligible—the overall effect being quite satisfactory. At higher load resistances the performance is not good, and the overall effect is roughly the same as with a pentode. See Chapter 7 for the effect of negative feedback.

Reference C3.

(xi) Space charge tetrodes

A space charge tetrode is actually a triode operating on a virtual cathode provided by the thermionic cathode and the inner (space charge) grid. It is capable of low distortion, even lower than a triode, but the power output for a given d.c. power input is less than that of a triode, owing to the power taken by the space-charge grid.

See Refs. C1, C2.

(xii) Partial triode operation of pentodes ("ultra-linear" operation)

When the screen and plate of a pentode are being operated at the same voltage, pentode operation is obtained when the screen is connected to the B+ end of the output transformer primary, while triode operation is obtained when the screen is connected to the plate end of the primary. Any desired intermediate condition can be obtained by connecting the screen to a suitable tap on the primary. In such intermediate condition the valve operates as a pentode having negative feedback applied to the screen, with a section of the load impedance common to both electrodes, and minimum high level distortion with push-pull operation is usually obtained when the tapping point is about 43% of the total primary turns (Refs. C7, H5, H6).

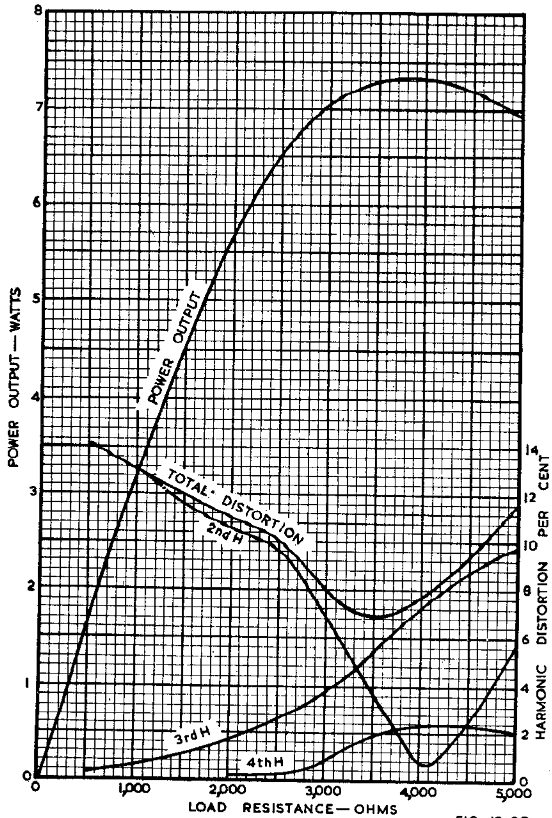


FIG. 13.30

Fig. 13.30. Power output and distortion of type 6L6 beam power amplifier plotted against load resistance, for $E_b = E_{c2} = 250$ volts, $E_{c1} = -14$ volts, peak signal 14 volts.

SECTION 4 : PARALLEL CLASS A AMPLIFIERS

Any two Class A amplifier valves may be connected in parallel, with suitable provision for their correct operation, to provide double the power output. It is assumed that these have identical characteristics (the normal manufacturing tolerances have only a very slight effect). The load resistance will be half that for one valve. The distortion will be the same as for one valve. The input voltage will be the same as for one valve. The total plate current will be twice that for one valve. The effective plate resistance will be half that for one valve.

Parasitics are likely to occur, particularly with high-slope valves. Precautions to be taken include the following :

1. The two valves should be placed closely together, with very short leads between grids and plates (and screens in the case of pentodes).
2. A grid stopper should be connected directly to one or both grids. It is usually cheaper—and just as effective—to have one grid stopper of, say, 200 ohms than two stoppers each of 100 ohms.
3. Screen stoppers (50 or 100 ohms for each screen) are very helpful, particularly with types 6L6 or 807 [see Sect. 3(iii)H].

N.B. Plate stoppers are less helpful, are wasteful of power, and are generally unnecessary.

The advantages of parallel operation lie principally in the elimination of the phase-splitter or input transformer required with push-pull operation. The disadvantages are :

1. The necessity for handling the heavy direct plate current. This necessitates either a special output transformer or a choke (say 20 henrys) with parallel feed to the output transformer.
2. The higher distortion—this is not serious if negative feedback is used, and in any case is no worse than that of a single valve.
3. The attenuation of lower frequencies at maximum power output due to the limited size of the by-pass capacitor. This effect is also a function of the plate supply regulation—see Sect. 1(iv).

Parallel operation may be used with a cathode-follower stage, permitting the use of two smaller valves with lower plate voltage in place of one valve with higher plate voltage, and thereby reducing the difficulties of grid excitation.

SECTION 5 : PUSH-PULL TRIODES CLASS A, AB_1

(i) Introduction (ii) Theory of push-pull amplification (iii) Power output and distortion (iv) Average plate current (v) Matching and the effects of mismatching (vi) Cathode bias (vii) Parasitics.

(i) Introduction

(A) Fundamental principles of push-pull operation

The fundamental circuit of a push-pull power amplifier is Fig. 13.31. A balanced (push-pull) input voltage must be applied to the three input terminals, and a balanced (push-pull primary) output transformer must be connected to the two plates with its centre-tap connected to E_{bb} +. For the best results the input voltage must be exactly balanced, the valves must have identical characteristics and the output transformer must be exactly balanced between the two sections of the primary, with perfect coupling between them. Under these conditions, any even harmonics introduced by the valves will be cancelled, but the odd harmonics will not be affected ; the flux in the core due to the d.c. plate currents would be zero.

FIG. 13.31

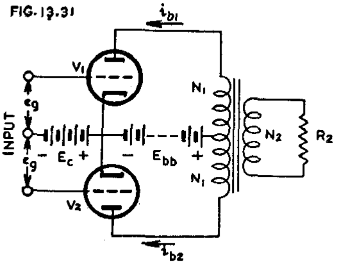


Fig. 13.31. Fundamental circuit of push-pull power amplifier.

The load resistance R_2 is connected across the secondary, and the reflected resistance* across the whole primary is

$$R_L = 4R_2(N_1^2/N_2^2)$$

*See Chapter 5 Sect. 1(ii).

and the reflected resistance across half the primary is

$R_L'' = \frac{1}{2}R_L = 2R_2(N_1^2/N_2^2)$

where N_1 = turns in half primary winding

and N_2 = turns in secondary winding.

If valve V_2 in Fig. 13.31 were removed from its socket, the load resistance effective* on V_1 would then be

$R_L' = \frac{1}{4}R_L = R_2(N_1^2/N_2^2)$

which is half the load resistance on V_1 under push-pull conditions. This is the condition which occurs when one of the valves reaches plate current cut-off.

If the output transformer were replaced by two separate transformers, one from each plate to E_{bb} +, the even harmonics would be cancelled but some advantages of push-pull operation would be lost. The whole principle of push-pull operation is based on the assumption that the two plates are always exactly 180° out of phase with one another. This is achieved by an " ideal " output transformer (see Chapter 5 Sect. 1) even if one of the valves is not operating.

If two valves are to be operated in push-pull, the plate and grid voltages may be the same as for single valve operation, the plate-to-plate load resistance may be twice and the plate currents twice the respective values for a single valve ; under these conditions the power output will be exactly twice that for a single valve. On the other hand, higher power output is obtainable, with some increase in third harmonic distortion, merely by decreasing the plate-to-plate load resistance without any change in grid bias (Ref. E7). Still higher power output is obtainable by increasing the grid bias and the signal input voltage, also by increasing the plate voltage to its maximum value.

(B) Classes of operation

A **Class A amplifier** is an amplifier in which the grid bias and alternating grid voltages are such that the plate current of the output valve or valves flows at all times. The suffix 1 indicates that grid current does not flow during any part of the input cycle.

A **Class AB_1 amplifier** is an amplifier in which the grid bias and alternating grid voltages are such that the plate current in any specific valve flows for appreciably more than half, but less than the entire, input cycle. The suffix 1 has the same meaning as with Class A.

A very useful operating condition is the borderline case between Class A and Class AB_1, that is when the plate current just reaches the point of cut-off—this is called **Limiting Class A_1 operation.**

Class A_1 triodes may be operated from poor regulation power supplies without serious loss of power output owing to the comparatively small rise of current at maximum signal. Class AB_1 triodes require good regulation of the power supply. Class AB_1 also requires very tight coupling between the two halves of the transformer primary.

Automatic bias control has been developed (Refs. A24, A25) whereby the operation is pure Class A_1 with small input voltages, with the bias automatically increasing with the input voltage to provide firstly Class AB_1 and finally Class AB_2 operation. See also Sect. 11(ii).

(C) Important features

Class A_1 triodes may be operated in push-pull from power supplies having poor regulation without any serious loss of power output, owing to the comparatively small rise of current at maximum signal. In addition, with normal by-passing, the regulation of the power supply does not affect the minimum frequency that can be handled satisfactorily at full power output.

Good regulation of the power supply is necessary with Class AB operation, on both these counts.

With Class AB operation, the current in one half of the output transformer is zero for part of the cycle—this causes a very rapid rate of change of current at the cut-off point, tending to cause parasitics due to the leakage inductance of the transformer. This effect may be minimized by the use of a transformer with low leakage induct-

*See Chapter 5 Sect. 1(ii).

ance—see Chapter 5 Sect. 3(iii)c—or by reducing the rate of change of current. The latter is accomplished in limiting Class A operation owing to the avoidance of the sharp bend or " discontinuity " in the characteristic at the cut-off point. Valves of the 6L6 or 807 class, when connected as triodes, have a slower rate of cut-off than normal triodes, and are therefore particularly adapted to Class AB_1 operation, which merges closely into limiting Class A.

The use of a resistive network to pass a steady current through the primary of the transformer does nothing to reduce the rate of change of current, although it may help in damping out any parasitics which may occur.

(ii) Theory of push-pull amplification

It is assumed that the input voltage is sinusoidal, that the operation is Class A_1, that the output transformer is ideal having no resistance or leakage reactance, and that the input and output voltages are balanced. The valves are assumed to be perfectly matched. For circuit and conditions see Fig. 13.31.

(A) The plate-current grid-voltage characteristic of any valve may be expanded into an infinite series—

$$i_{b1} = a_0 + a_1e_g + a_2e_g^2 + a_3e_g^3 + a_4e_g^4 + a_5e_g^5 + \ldots\ldots \quad (1)$$

where i_{b1} = instantaneous plate current,

e_g = instantaneous grid signal input voltage,

a_0 = plate current at zero signal,

and a_1, a_2, a_3, etc. are coefficients.

The instantaneous fundamental power output is given by

$$P_0 = (a_1e_g)^2R_L \quad (2)$$

where R_L = load resistance on one valve.

In a push-pull amplifier the second valve V_2 has a grid voltage opposite in polarity to V_1:

$$i_{b2} = a_0 - a_1e_g + a_2e_g^2 - a_3e_g^3 + a_4e_g^4 - a_5e_g^5 + \ldots\ldots \quad (3)$$

where i_{b2} = instantaneous plate current of V_2,

a_0 = plate current of V_2 at zero signal,

and a_1, a_2, a_3 etc. have the same meanings as in eqn. (1).

Eqn. (3) has been derived from eqn. (1) merely by making e_g negative, so that e_g^2 becomes positive and e_g^3 becomes negative.

As the plate currents are in phase opposition to the output load, the net flux-producing current is

$$i_d = i_{b1} - i_{b2}$$
$$= 2a_1e_g + 2a_3e_g^3 + 2a_5e_g^5 + \ldots\ldots \quad (4)$$

where i_d = net flux-producing current in N_1 turns, i.e. one half the primary winding,

and all currents are instantaneous values.

The d.c. components and all the even harmonic terms are seen to have been cancelled, only the fundamental and odd harmonic terms remaining.

The total plate current from E_{bb} is given by

$$i_t = i_{b1} + i_{b2}$$
$$= 2a_0 + 2a_2e_g^2 + 2a_4e_g^4 + \ldots\ldots \quad (5)$$

from which it will be seen that the d.c. components plus the even harmonic terms are present. Thus the total supply current will only remain constant when there is no even harmonic distortion.

(B) The effect of hum in the plate and grid supply voltages

If the plate supply voltage E_{bb} in a distortionless Class A amplifier is changed slightly, both plate currents will change together, and the net flux-producing current (i_d) will be unchanged. Hum from the supply voltage would not therefore appear in the output. The effect of hum from the grid bias supply will be similar.

In a practical amplifier, owing to the curvature of the valve characteristics, the effect may be analysed as follows, commencing with hum in the grid bias supply voltage. Assume that there is a hum voltage $E_h \cos pt$ in series with E_c, and that the input signal is $E_g \cos qt$.

The input voltage to V_1 is $E_h \cos pt + E_g \cos qt$.
The input voltage to V_2 is $E_h \cos pt - E_g \cos qt$.
Substituting these values in equation (1) we have

$$\begin{aligned} i_{b1} &= a_0 + a_1(E_h \cos pt + E_g \cos qt) + a_2(E_h \cos pt + E_g \cos qt)^2 + \ldots\ldots \\ &= a_0 + a_1E_h \cos pt + a_1E_g \cos qt + a_2E_h{}^2 \cos^2 pt + a_2E_g{}^2 \cos^2 qt + \\ &\quad 2a_2E_hE_g \cos qt \cos pt + \ldots\ldots \end{aligned}$$

Similarly

$$\begin{aligned} i_{b2} &= a_0 + a_1(E_h \cos pt - E_g \cos qt) + a_2(E_h \cos pt - E_g \cos qt)^2 + \ldots\ldots \\ &= a_0 + a_1E_h \cos pt - a_1E_g \cos qt + a_2E_h{}^2 \cos^2 pt + a_2E_g{}^2 \cos^2 qt - \\ &\quad 2a_2E_hE_g \cos qt \cos pt + \ldots \end{aligned}$$

Therefore

$$\begin{aligned} i_d &= i_{b1} - i_{b2} = 2a_1E_g \cos qt + 4a_2E_hE_g \cos qt \cos pt \\ &= 2a_1E_g \cos qt[1 + 2(a_2/a_1)E_h \cos pt] \qquad (6) \end{aligned}$$

Eqn. (6) has the form of a carrier ($2a_1E_g \cos qt$) modulated to the depth $2(a_2/a_1)E_h$ by a hum frequency $\cos pt$.

The effect of hum in the plate supply voltage is similar to its effect on the grid bias supply, that is to say the hum frequency modulates the signal frequency. It should be noted that these modulation components appear in single-ended amplifiers to the same extent, but in combination with the fundamental and harmonics of the ripple, which latter are absent with push-pull operation.

Summary—Push-pull operation tends always to reduce the effects of hum in either the grid bias or plate supply voltage.

(C) Effects of common impedance

The fact that no fundamental component is found in the total plate current i_t (eqn. 5) prevents any fundamental signal voltage from being fed back to earlier stages as the result of a common impedance in the plate voltage source. As the even harmonics are fed back, each will again result in higher order harmonics, so that there can be no instability caused by feedback around a push-pull stage.

General deductions

1. Because the d.c. components of the plate currents cancel each other, no steady flux is maintained in the core of the output transformer.
2. Because the even harmonics are zero, the limit placed on single-ended amplifiers no longer applies. It is usual to design push-pull amplifiers for maximum power output without primarily considering the odd harmonics; when distortion is objectionable, this may be reduced—at the expense of power output—by increasing the load resistance.

The effects of the regulation and by-passing of the power supply are covered in Sect. 5(i).

(D) Application to characteristic curves—Composite characteristics

From eqn. (4) the net flux-producing current in one half of the primary winding is

$$i_d = i_{b1} - i_{b2} \qquad (7)$$

Thus a "composite" characteristic may be drawn for the fixed grid voltage $-E_c$ by subtracting the currents in the two valves. At the quiescent plate voltage E_{bb}, both valves draw the same plate current and therefore $i_d = 0$. The composite characteristic must therefore pass through the point $E_b = E_{bb}$, $i_d = 0$. At other plate voltages the value of i_d is given by $i_{b1} - i_{b2}$ when the plate voltage of one is increased

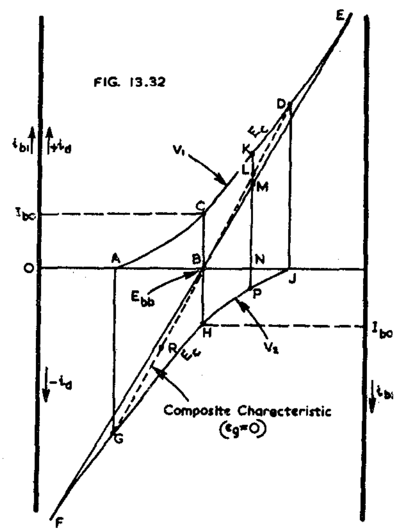


Fig. 13.32. The derivation of the composite characteristic for matched push-pull valves.

by the same voltage that the other is decreased. This may be applied graphically as shown in Fig. 13 32 ; the upper half includes the E_c characteristic for V_1 (ACDE) while the lower half includes the E_c characteristic for V_2 but inverted and placed left to right (FGHJ). Point C on the V_1 characteristic is the quiescent operating point, and point B corresponds to the plate voltage E_{bb}. The V_2 characteristic is placed so that H, the quiescent operating point, comes below C ; then since BC = BH = I_{b0} the point B is on the composite characteristic. At any other plate voltage N, the point L on the composite characteristic is found by subtracting PN from KN, giving LN as an ordinate. Valve V_2 cuts off at point J, so that the amount to be subtracted from the V_1 ordinate is zero, giving D (and its opposite number G) as points on the composite characteristic. The composite characteristic is therefore FGRBLDE which may be compared with the straight line FBE. Sudden bends occur at D and G, but the portion between D and G is fairly straight ; the latter includes the whole Class A operating region. It is obvious that the non-linearity of the composite characteristic becomes worse as the quiescent operating point is moved towards the foot of the characteristic—that is as I_{b0} becomes less. Thus we have the Class A condition (including limiting Class A) with nearly straight composite characteristics, the Class AB_1 condition which includes the kinks at the points of plate-current cut-off, and finally the Class B condition with quite considerable non-linearity.

The composite characteristic of Fig. 13.32 is that for zero signal input voltage (e_g = 0). Other composite characteristics may be drawn by a somewhat similar method, except that the ($E_c + e_g$) characteristic of V_1 must be combined with the ($E_c - e_g$) characteristic of V_2 to give the $+ e_g$ composite characteristic. For example, if $E_c = -60$ volts, we may take e_g in increments of 10 volts, giving :

e_g	0	+10	+20	+30	+40	+50	+60	volts
$E_c + e_g$	−60	−50	−40	−30	−20	−10	0	volts
$E_c - e_g$	−60	−70	−80	−90	−100	−110	−120	volts

A family of composite characteristics is shown in Fig. 13.33 in which E_{bb} = 300 volts and $E_c = -60$ volts, with values of e_g in accordance with the table above. The

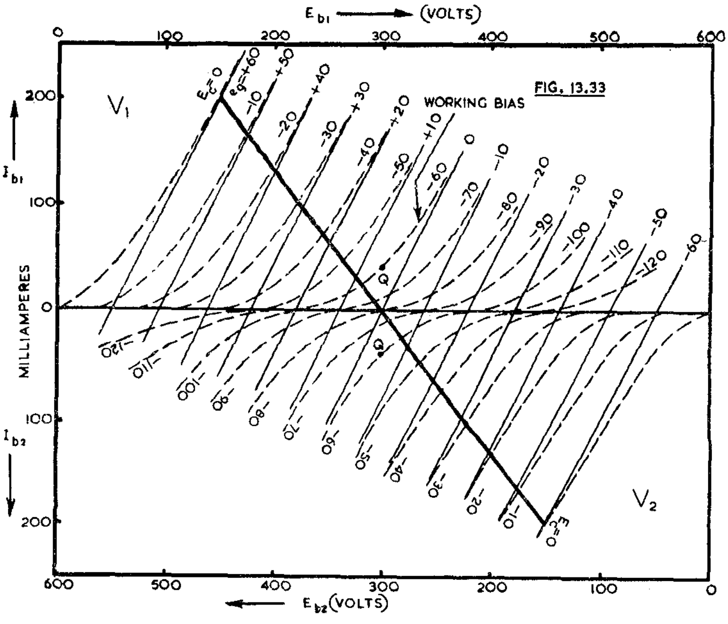


Fig. 13.33. Family of composite characteristics for two type 2A3 triodes.

composite operating point is $e_g = 0$ and $E_b = E_{bb} = 300$ volts. The composite loadline is a straight line through this point, with a slope corresponding to $R_L' = \frac{1}{4}R_L$, where R_L = load resistance plate to plate. We may therefore imagine a composite valve, taking the place of both V_1 and V_2, working into half the primary winding with the other half open-circuited. This composite valve will have a plate resistance (r_d) as indicated by the slope of the composite characteristic, which value is approximately half that of one valve at the quiescent operating point (r_{po}). For this reason the slope of the composite characteristic changes slightly with the grid bias.

Maximum power output is obtained from the composite valve when its load resistance is equal to its plate resistance

i.e. when $R_L' = R_L/4 = r_{p0}/2$ or $R_L = 2r_{p0}$ (8)

On the composite characteristics therefore, maximum power output is obtained when the slope of the loadline is the negative of the slope of the composite characteristics.

Owing to the good linearity of the composite characteristics for Class A_1 operation, and the freedom from limitations in the vertical direction, **elliptical loadlines** may be accommodated with less distortion than with any other method. **Negative voltage feedback** makes such an amplifier practically distortionless for any type of load, resistive or reactive, of any value of impedance ; the only limitation is regarding grid current.

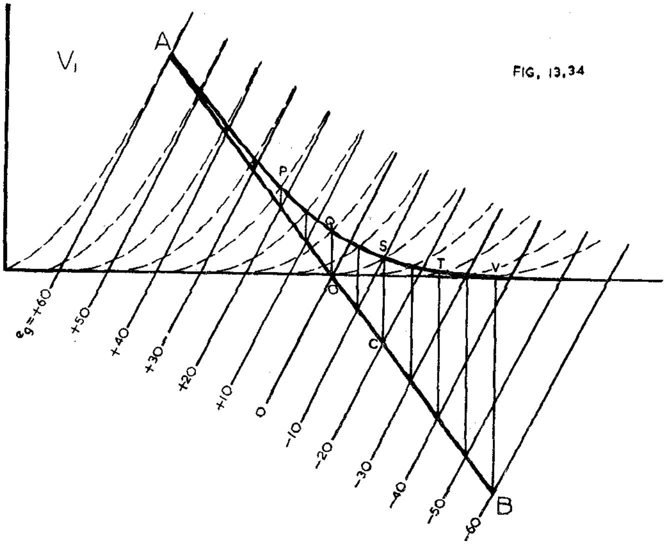


Fig. 13.34. *Method of deriving the loadline on an individual valve (AQV) from the composite characteristic AOB.*

The loadline on one valve (e.g. V_1) is determined as Fig. 13.34 in which the composite characteristics, the composite loadline AB, and the individual V_1 characteristics are the same as in Fig. 13.33. At every intersection of a composite characteristic with the composite loadline (e.g. point C) draw a vertical line (e.g. CS) to cut the corresponding V_1 characteristic. Then join these points APQSTV etc. with a smooth curve, which is the loadline on a single valve, Q being the quiescent operating point. It is obvious that this loadline is curved, although it is less curved with Class A than with Class AB_1 operation.

Equivalent circuit for push-pull amplifier

There are various forms which an equivalent circuit might take, but the one adopted here (due to Krauss) has some special advantages for the purpose (Fig. 13.35). The two valves are assumed to have equal constant amplification factor (μ), and plate resistances (r_{p1} and r_{p2}) which are functions of the plate current.

From equation (4) we have

$i_d = i_{b1} - i_{b2}$.

We may write

$$i_{b1} = I_{b0} + \Delta i_{b1} \text{ and } i_{b2} = I_{b0} - \Delta i_{b2} \tag{9}$$

where I_{b0} = quiescent plate current of either valve

Δi_{b1} = change of plate current in valve V_1

and Δi_{b2} = change of plate current in valve V_2.

Combining (4) and (9),

$$i_d = \Delta i_{b1} + \Delta i_{b2} \tag{10}$$

in which the varying components of i_{b1} and i_{b2} add so far as i_d is concerned. Let the varying components be i_{p1} and i_{p2}, thus leading to the equivalent circuit of Fig 13.35 in which two generators each developing a voltage μe_g through their internal resistances r_{p1} and r_{p2}, are effectively in parallel to supply the load current i_d through the load resistance R_L'. All quantities except R_L' and μ are instantaneous values.

FIG. 13.35

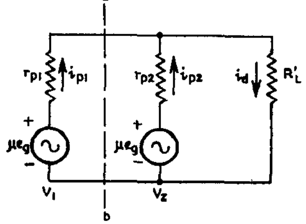


Fig. 13.35. Equivalent circuit which may be used for deriving certain impedance and current relationships.

The impedance seen by V_1 is the impedance of the circuit to the right of the line *ab*, which is called r_{ab}.

It may be shown (Ref. E7) that

$$r_{ab} = R_L'(1 + r_{p1}/r_{p2}) \tag{11}$$

The dynamic plate resistance of the composite valve may be expressed

$$r_d = \Delta e_{b1}/i_d \tag{12}$$

From (10) $i_d = \Delta i_{b1} + \Delta i_{b2}$.

$$\text{Now } \Delta i_{b1} = \Delta e_{b1}/r_{p1} \tag{13}$$

$$\text{and } \Delta i_{b2} = \Delta e_{b2}/r_{p2} \tag{14}$$

$$\text{From (12), (13), (14), } r_d = r_{p1}r_{p2}/(r_{p1} + r_{p2}) \tag{15}$$

This indicates that the plate resistance of the composite valve at any instant is equal to the parallel combination of the individual plate resistances of V_1 and V_2. For Class A operation r_d is very nearly constant, so that

$$r_d \approx r_{p0}/2 \tag{16}$$

where r_{p0} = plate resistance of V_1 or V_2 at the quiescent operating point.

Now if $R_L' = r_d$ (the condition for maximum power output)

$$R_L' = r_d \approx r_{p0}/2 \tag{17}$$

The load impedance seen by a single valve (V_1 or V_2) in Class A operation is given by

$$\text{[From (11), (15), (17)], } r_{ab} \approx r_{p1} \tag{18}$$

so that each valve at any instant is working into a load resistance approximately equal to its own plate resistance.

(iii) Power output and distortion

It has been shown in the preceding subsection that maximum power output is obtained from two matched valves in push-pull when the load resistance from plate to plate (R_L) is equal to 4 times the plate resistance of the imaginary composite valve (r_d) or approximately twice the plate resistance of one of the valves at the quiescent operating point (r_{p0}). This value of R_L may be regarded as the minimum value, since any decrease would cause loss of power output, increased distortion, and high peak currents. In some circumstances it is found desirable to increase R_L—even though this reduces the power output—thereby reducing the odd harmonic distortion, the peak currents and plate dissipation.

The value of load resistance to provide maximum power output may be determined approximately from the plate characteristics of one valve (Fig. 13.36). Since the $E_c = 0$ characteristic approximately follows the 3/2 power law, it may be shown (Ref. E5) that maximum power output occurs when the loadline intersects the $E_c = 0$ curve at $0.6E_{bb}$. The plate current at this point is I_{bm} and the other values are :

$R_L = 1.6E_{bb}/I_{bm}$ plate to plate (19)

$P_0 = 0.2E_{bb} \cdot I_{bm}$ for 2 valves (20)

If it is desired to determine the optimum value of R_L in a particular case, several half-loadlines may be drawn as in Fig. 13.36 but radiating from B. The power output for each may be calculated from the expression

$P_0 = \frac{1}{8}I_{bm}{}^2R_L$ (21)

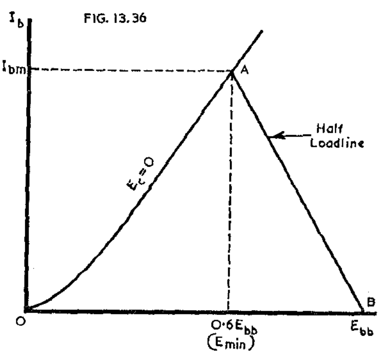
or $P_0 = \frac{1}{2}I_{bm}(E_{bb} - E_{min})$. (22)

Eqns. (21) and (22) may be used with reasonable accuracy for Class AB_1, since the third harmonic distortion is usually less than 3%. A useful rule is to multiply the power output, as indicated by these equations, by the factor $10\,000/(100 - H_3\%)^2$ to obtain a close approximation to the actual power output. Values of this factor for various third harmonic percentages are given below :

H_3	1%	2%	3%	5%	7%	10%
Factor	1.02	1.04	1.06	1.11	1.15	1.23

It is here assumed that fifth and higher order odd harmonics are negligible.

The loadline slope is unaffected by the grid bias, but the two extremities are slightly affected—as the grid bias is decreased, the point A (which is really the intersection of the loadline with the composite characteristic) moves slightly towards B. Thus with Class A_1 the output will be slightly less than indicated by eqns. (20), (21) and (22).



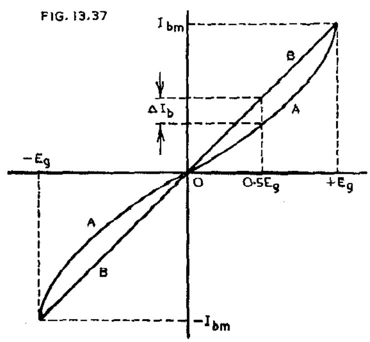


Fig. 13.36. Method of deriving the approximate power output and load resistance of a Class AB_1 amplifier from the characteristics of a single valve. The method may also be used with poorer accuracy for Class A_1.

Fig. 13.37. Calculating third harmonic distortion with balanced push-pull. Curve A is the dynamic characteristic.

The dissipation at maximum signal may be calculated from the product of the plate supply voltage E_{bb} and the average plate current [see (iv) below], minus the power output. The normal procedure is firstly to select a grid voltage such that the plate dissipation is slightly below the maximum rating at zero signal, then to adjust the loadline so that it approaches as closely as possible the value for maximum power output, without exceeding the plate dissipation limit at maximum signal.

Simple method for calculating third harmonic distortion

This method is only accurate in the absence of all distortion other than third harmonic ; it is a close approximation under normal conditions provided that all even harmonics are zero.

The procedure is to draw the loadline on the plate characteristics, then to transfer this to the mutual characteristics in the form of a dynamic characteristic (Fig. 13.37).

Curve A is the dynamic characteristic while B is a straight line joining the two ends of A and passing through O. ΔI_b is the difference in plate current between curve A and line B at one half of the peak grid voltage. The percentage of third harmonic is given by

$$\left(\frac{2}{3}\Delta I_b \times 100\right) \Big/ \left(I_{bm} - \frac{2}{3}\Delta I_b\right) \tag{23}$$

or $$\left(\frac{2}{3}\Delta I_b \times 100\right) / I_{bm} \text{ approximately.} \tag{24}$$

Calculating up to fifth harmonic distortion

The usual method is to transfer from the loadline to a dynamic characteristic, then to proceed using the " eleven selected ordinate method " (Sect. 3(iv)D). This gives the power output and harmonics up to the fifth. If the balance is good, only the third and fifth harmonics are likely to be required ; the fifth is very much less than the third harmonic for Class A operation.

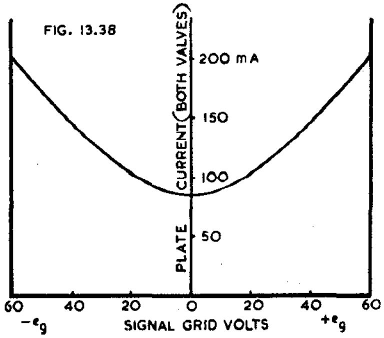


Fig. 13.38. Plate current (both valves) plotted as a function of the signal grid voltage.

(iv) Average plate current

The average plate current with maximum signal input is always greater than under quiescent conditions—slightly greater for " single valve " conditions, more so for limiting Class A_1, and considerably greater for Class AB_1.

The average plate current may be calculated approximately by the expression

$$I_b \approx I_{b0} + \tfrac{1}{4}(I_{max} + I_{min} - 2I_{b0}) \tag{25}$$

provided that the plate current does not actually cut off (Ref. E1).

More generally, and more accurately, the average plate current may be determined by adding the plate currents of the two valves instead of subtracting them as for the composite characteristics. The total plate current may then be plotted as in Fig. 13.38 as a function of the signal grid voltage. In order to find the average current it is generally most convenient to take equal angle increments over the cycle, for example, every 10° as shown in Fig. 13.39. The plate current should then be noted at each point corresponding to 10° increase in angle over the whole 360°. The average plate current is then the average of these individual values.

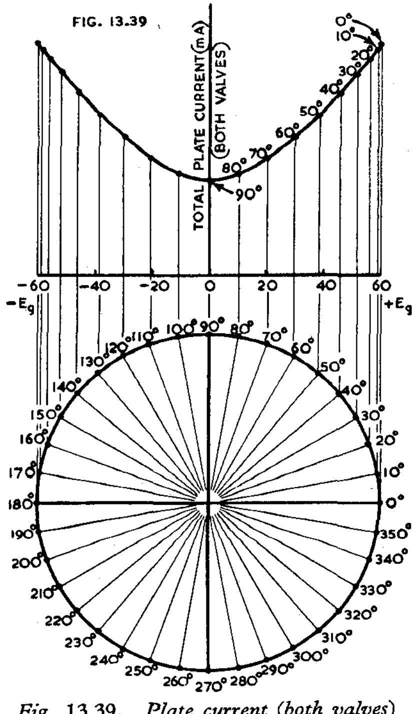


Fig. 13.39. Plate current (both valves) with 10° angular increments.

In order to reduce the amount of work involved in this calculation, use may be made of the fact that each quadrant (90°) is similar. It is very easy to make an error in this calculation and the following method of obtaining the average from the plate current curve over one quadrant is therefore given.

This method is illustrated in Fig. 13.40 which shows the total plate current for both valves over one-quarter of a cycle. Grid voltages are shown as fractions of the peak grid voltage. The plate currents corresponding to grid voltages of 0, 0.17, 0.34, 0.5, 0.64, 0.77, 0.87, 0.94, 0.98 and 1.0 times the peak voltage are shown as I_0, $I_{0.17}$ etc. The average plate current (I_{av}) is then given by

$$I_{av}=\tfrac{1}{9}(\tfrac{1}{2}I_0+I_{0.17}+I_{0.34}+I_{0.5}+I_{0.64}+I_{0.77}+I_{0.87}+I_{0.94}+I_{0.98}+\tfrac{1}{2}I_{1.0}). \qquad (26)$$

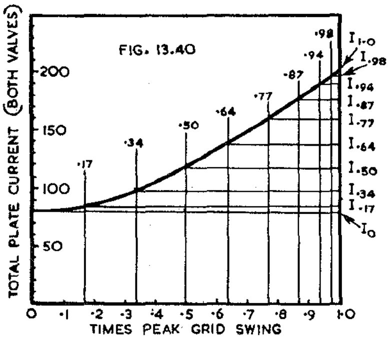


Fig. 13.40. Plate current (both valves) plotted over one quadrant for the calculation of average plate current.

(v) Matching and the effects of mismatching

Matching is the process of selecting valves for satisfactory push-pull operation. Most valve manufacturers are prepared to supply, at some additional cost, valves which have been stabilized and matched for the application required.

It is important to specify the conditions of operation when ordering matched valves, as unless the valves are stabilized and matched under conditions similar to those in which the valves are to be used, they will drift apart during life or may even be mismatched initially if the matching conditions are unsuitable.

All valves intended for matching should be operated for at least 50 hours under similar conditions to those under which the valves are intended to be operated in the amplifier.

The matching technique varies with the class of operation for which the valves are intended.

Matching valves for class A_1 service is not as critical as for valves intended for class AB_1, class AB_2 or class B. For class A_1 service it is usually sufficient to match for zero signal plate current only.

Triodes for class AB_1, AB_2 or class B service should be checked at a number of points on the plate current grid bias curve. The points usually taken are (a) zero signal condition (b) a bias corresponding to the maximum permissible plate dissipation. The plate currents so measured should agree at all points within 2%. Triodes intended for class AB_2 or class B service should also be matched for amplification factor.

When matching tetrodes and pentodes it is usually sufficient to match for zero signal plate current and power output.

Even with perfect initial matching, valves are likely to drift apart during life and in critical applications it is desirable to provide some means of balancing the plate currents of the valves in the equipment. This may take the form of separate bias resistors in the case of self-bias or adjustable bias supplies in the case of fixed-bias applications. Pentodes and tetrodes may also be balanced by adjusting the screen voltages.

It is important that matched valves should never be run, even momentarily, at dissipations or ratings in excess of those recommended by the valve manufacturers as such treatment will render the valves unstable and destroy the matching.

Effects of mismatching with Class A_1 triodes

In Class A_1 push-pull triodes, a considerable degree of mismatching between the valves is permissible without serious effects, provided that the valves are being operated under single valve conditions as regards grid bias. There will be only a slight effect on the maximum power output or the odd harmonic distortion, but there will be some second harmonic distortion and some out-of-balance flux-producing current in the transformer. The second harmonic distortion will normally be small, particularly if the plate-to-plate load resistance is not much less than $4r_{p0}$, and for most purposes it may be neglected with valves of ordinary tolerances. The maximum out-of-balance plate current should be provided for in the design of the output transformer and this additional cost should be compared with the alternative additional cost of using valves which have been stabilized and matched for quiescent plate current only. There is normally no real need for matching for any other characteristic.

In order to demonstrate the effects of abnormal mismatching, composite characteristics have been drawn in Fig. 13.41 for two valves of entirely different types. It has been shown by Sturley (Ref. E27) that the method of deriving composite characteristics also holds with mismatching.

Type	2A3	45
Amplification factor	4.2	3.5
Mutual conductance	5250	2175 μmhos
Plate resistance	800	1610 ohms

There is a difference of 12% in amplification factor, while there is a ratio exceeding 2 : 1 for the other characteristics. The selected operating conditions are : E_b = 250 volts, E_c = – 50 volts, R_L = 5800 ohms (plate-to-plate). It will be seen that the composite characteristics (dashed lines) are not quite parallel, although they are very nearly straight. On account of the unmatched condition, rectification occurs, leading to a shift of the loadline, and the corrected loadline may be derived by the method of Sect. 2(ii)D and Fig. 13.7. The second harmonic distortion on the corrected loadline is only 5%.

FIG. 13.41

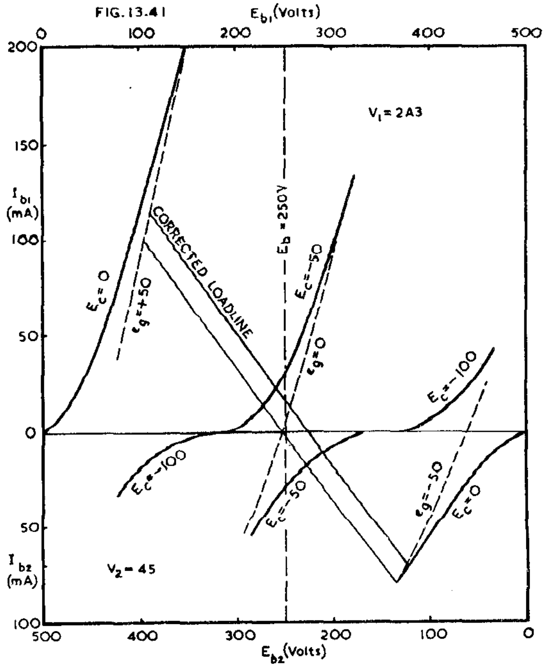


Fig. 13.41. Composite characteristics for two different valves (2A3 and 45) in Class A_1 push-pull. E_{bb} = 250 volts, E_c = –50 volts, R_L = 5800 ohms plate-to-plate.

The fact that two valves so different in characteristics can be used in Class A_1 push-pull to give what is generally classed as good quality (5% second harmonic) indicates the wide latitude permissible, provided that the bias is retained at the value for a single valve and also that the load resistance is not much less than the sum of the single valve loads (or 4 times the average plate resistance).

For Class A_1 triodes it has been suggested (Ref. E8) that the following are reasonably satisfactory limits in conjunction with ordinary commercial tolerances for valves :

	Max. unbalance
Signal input voltages on the grids of V_1 and V_2 :	5%
Phase unbalance at high and low frequencies :	
Quadrature component =	3%

These are easily achieved by attention to the phase splitter or other source (see Chapter 12 Sect. 6). When testing for balance with a C.R.O. on the grids of V_1 and V_2, it is important to reverse the connections of the B.F.O. when changing from one grid to the other.

(vi) Cathode bias

In an accurately balanced push-pull Class A_1 amplifier there is no point in by-passing the common cathode bias resistor, since there is no fundamental signal current flowing through it. In Class A_1 amplifiers which are not accurately balanced, there will be some degeneration, and it is usual to by-pass the cathode bias resistor, although this is not essential.

In Class AB_1 amplifiers it is essential to by-pass the cathode bias resistor.

Provision may be made, with cathode bias, for balancing the plate currents provided that they do not differ too seriously. One excellent arrangement is incorporated in Fig. 7.44 which may be used, with the necessary adjustments in the values of the resistances, for triodes, pentodes or beam power amplifiers.

The value of cathode bias resistance may be determined, in the same way as for a single valve, on the basis of the maximum signal total plate current and desired bias voltage. This will give the same performance as fixed bias, but it is necessary to check for plate dissipation at zero signal [for method see Sect. 2(ii)E]. If the dissipation at zero signal is too great, it will be necessary to increase the bias resistance. This will introduce a tendency to change from Class A_1 to AB_1, which may be undesirable ; it may be minimized by the use of a bleed resistor to pass current from E_{bb} to the cathode, and thence through the bias resistor, thus giving an approach towards fixed bias. Alternatively, the load resistance may be increased, thereby reducing the maximum signal plate current and grid bias ; this will also reduce the power output.

Cathode bias causes a smaller change in average plate current from no signal to maximum signal than fixed bias. This permits a poorer regulation power supply than may be used with fixed bias. However, the regulation and by-passing of the power supply also affect the minimum frequency which can be handled satisfactorily at full power output—see Sect. 1(iv).

Changes in effective gain occur in Class AB_1 amplifiers employing cathode bias, during heavy low frequency **transients**, which add to the distortion measured under steady conditions.

(vii) Parasitics

Parasitic oscillations in the plate circuit may occur with Class AB_1 operation when the plate current is cut off for an appreciable part of the cycle, as a result of the transformer leakage inductance and the rapid rate of change of current at the cut-off point—see Sect. 5(i)C. They may usually be cured by the use of a *RC* network shunted across each half of the primary of the output transformer—see Sect. 7(i)—and, if necessary, by the use of a transformer with lower leakage inductance—see Chapter 5 Sect. 3(iii)c.

Parasitics in the grid circuit are not usually troublesome except when the valves are driven to the point of grid current flow. Grid stoppers up to 50 000 ohms are often used with both Class A and AB_1 operation to give a smoother overload without parasitics.

SECTION 6: PUSH-PULL PENTODES AND BEAM POWER AMPLIFIERS, CLASS A, AB_1

(i) Introduction (ii) Power output and distortion (iii) The effect of power supply regulation (iv) Mismatching (v) Average plate and screen currents (vi) Cathode bias (vii) Parasitics (viii) Phase inversion in the power stage (ix) Extended Class A (x) Partial triode ("ultra-linear") operation.

(i) Introduction

Push-pull pentodes follow the same general principles as triodes (see Sect. 5) although there are some special features which will here be examined. Provision must be made for the supply of voltage to the screens, and for it to be maintained constant with respect to the cathode. For economy, the screens are often operated at the same voltage as the plates. If the screens are to be operated at a lower voltage, the alternatives are the use of a separate power supply, or a low resistance voltage divider. The McIntosh Amplifier may also be used for Class A or AB_1 operation—see Sect. 8. Quiescent push-pull pentodes are covered in Sect. 7(vii).

(ii) Power output and distortion

Composite characteristics may be drawn as for triodes, on the assumption that the screen voltage is maintained constant, but they will not be straight (Fig. 13 42). The loadline will pass through the point ($E_b = E_{bb}, I_b = 0$) and the two knees of the characteristics ($e_g = \pm E_{c1}$).

In practice it is not necessary to draw the composite characteristics, if we are only interested in power output. In limiting Class A_1 or in AB_1 one valve reaches cut-off, so that the $E_{c1} = 0$ characteristic is the same as the e_g characteristic. Even with ordinary Class A_1 the error due to the approximation is small. The loadline may therefore be drawn (*AB* in Fig. 13.43) and the plate-to-plate load resistance will be 4 times that indicated by the slope of *AB*. With valves of the 6L6 class, the third

FIG. 13.42

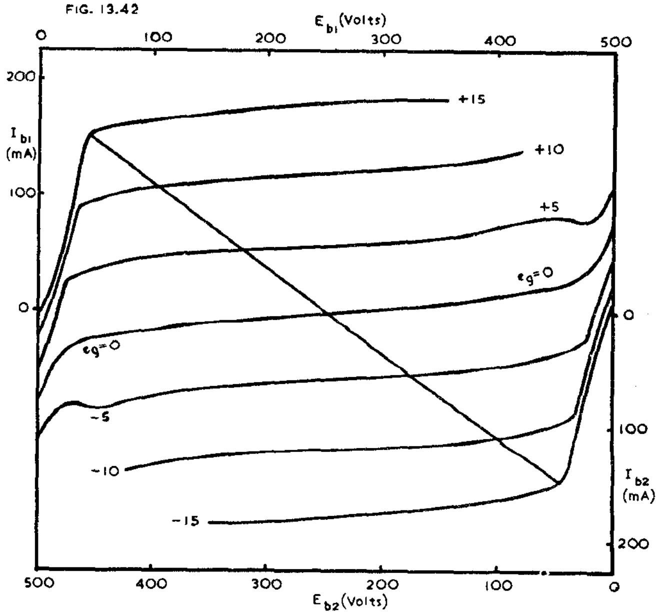


Fig. 13.42. Composite characteristics for push-pull Class A_1 beam power amplifiers type 6L6, $E_{bb} = 250$ volts, $E_{c2} = 250$ volts, $E_{c1} = -15$ volts.

harmonic is so small that its effect on the power output may be neglected, so that

$$P_o = \tfrac{1}{2} I_{bm}(E_{bb} - E_{min}) \quad (1)$$

which is the same as for triodes. With pentodes the power output will be somewhat higher than indicated by eqn. (1) owing to the third harmonic distortion. In general, the third harmonic distortion is slightly less than half that with a single valve, owing to the effect of the lower load resistance. If it is desired to calculate the harmonic distortion, it will be necessary to plot at least portion of the composite characteristics.

The effect of a higher load resistance is to increase rapidly the odd harmonic distortion, while the effect of a lower load resistance is to decrease the power output. It is therefore advisable, with a loudspeaker load, to adopt a nominal impedance rather less than the value for maximum power output.

The power output of Class A push-pull pentodes is only slightly greater than twice that for a single valve.

FIG. 13.43

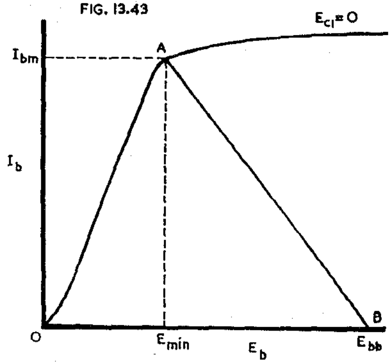


Fig. 13.43. *Calculation of power output using $E_{c1} = 0$ characteristic of one valve only.*

(iii) The effect of power supply regulation

As with the case of a single valve the screen is more sensitive to voltage changes than the plate. If neither screen nor plate is being operated at its maximum rating, the simplest procedure is to adjust the voltages as desired for maximum signal and allow them to rise with no signal. Alternatively, if it is desired to obtain maximum power output, both screen and plate may be adjusted to their maximum ratings with no signal, and allowed to fall with increasing signal. If it is desired to calculate the maximum power output, the procedure for a single valve [Sect. 3(ix)] may be followed, except that in this case there is no loadline shift caused by rectification.

The regulation of a common plate and screen power supply will not have a serious effect on the minimum frequency which can be handled satisfactorily at full power output with Class A push-pull operation, provided that a reasonably large by-pass capacitor is used. If separate plate and screen supplies are used the screen supply regulation is very much more important than the plate supply regulation, with Class A operation.

With Class AB_1 operation, the regulation of both plate and screen supplies should be good.

(iv) Matching and the effects of mismatching

Matching is covered generally in Sect. 5(v). It is important to match the valves under the operating conditions in the amplifier.

The effects of mismatching with pentodes are more serious than with triodes. If no care is taken in matching, or in the design of the output transformer, it is possible for the distortion to be higher than with two valves in parallel. The advantages of push-pull operation will only be obtained in proportion to the care taken to achieve correct balance, particularly with regard to the quiescent plate currents and the signal input voltages. As with triodes, Class AB_1 is more sensitive to mismatching than Class A_1.

(v) Average plate and screen currents

The average plate current for Class A_1 may be calculated approximately as for triodes [Sect. 5(iv), Eqn. 25]. In all other cases, the composite characteristics are required, following the same method as for triodes.

The average screen current may be calculated by the same method as for single pentodes [Sect. 3(iii)E].

(vi) Cathode bias

See Sect. 5(vi) as for push-pull triodes, except that the screen dissipation must also be checked.

(vii) Parasitics

See Sect. 5(vii) as for push-pull triodes, also Sect. 3(iii)H.

(viii) Phase inversion in the power stage

In the interests of economy, push-pull is sometimes used in the output stage without a prior phase inverter. All such methods—except the Cathamplifier—have inherently high distortion, and some have serious unbalance between the two input voltages.

(A) **Phase inverter principle** (Fig. 13.44)

The grid of V_2 is excited from the voltage divider R_3R_4 across the output of V_1. $R_3 + R_4$ must be very much greater than the load resistance (say 50 000 ohms). R_5 and R_6 are grid stoppers. All other components are normal. R_k may be by-passed if desired.

FIG. 13.44

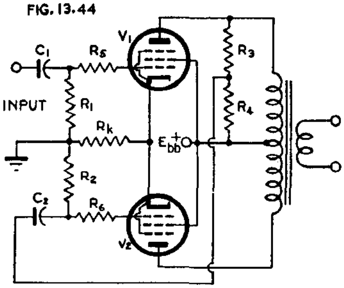


FIG. 13.45

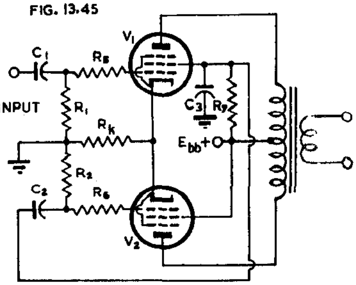


Fig. 13.44. *Push-pull circuit using phase inversion in the power stage.*
Fig. 13.45. *Push-pull circuit using screen resistance coupling from V_1 to the grid of V_2.*

The signal voltage on the grid of V_2 must first pass through V_1 where it is distorted, then through V_2 where it will be distorted again. Thus the second harmonic will be the same as for a single valve, and the third harmonic will be approximately twice the value with balanced push-pull. The balance, if adjusted for maximum signal, will not be correct for low volume, owing to the third harmonic " flattening."

(B) **Screen resistance coupling** (Fig. 13.45)

This is a modification of (A) being an attempt to obtain from the screen a more linear relationship than from the plate. No comparative measurements have been published. R_7 may be about 1500 ohms for type 6V6-GT or 2500 for type 6F6-G, with $E_b = E_{c2} = 250$ volts—the exact value should be found experimentally ; C_3 may be 0.002 μF. For better balance an equal screen resistor might be added for V_2. Ref. E10.

(C) **Common cathode impedance** (Fig. 13.46)

R_1 and R_2 in series provide a common cathode coupling impedance [see Chapter 12 Sect. 6(vi)]. R_2 may have a value of, say, 1000 ohms to give an approach towards balance, but necessarily must carry the plate currents of both valves—say, 70 or 80 mA—and will have a voltage of, say, 70 to 80 with a dissipation around 6 watts. Care should be taken to avoid exceeding the maximum heater-cathode voltage rating.

See Reference E23.

FIG. 13.46

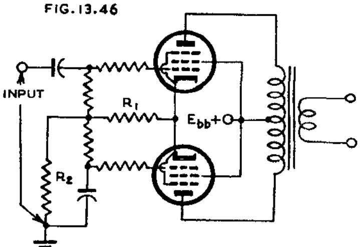


Fig. 13.46. *Push-pull circuit with common cathode impedance coupling.*

(D) The Parry " Cathamplifier "

The basic circuit is Fig. 13.46A, the two cathodes being coupled by a centre-tapped a-f transformer, whose secondary winding excites the grid of V_2. A theoretical analysis is given in Ref. E29, while some practical designs are in Ref. E30.

For balance, $\dfrac{N_3}{N_1 + N_2} = \dfrac{1 + g_m R/2}{g_m R}$ and $N_1 = N_2$

where g_m = mutual conductance of V_2.

Distortion is reduced by the factor $T(2T - 1)$

where $T = N_3/(N_1 + N_2)$.

Note that T should normally be slightly greater than 1.

Gain is reduced by the factor $T(T - 0.5)$.

The common cathode resistor R_0 helps to reduce unbalance.

In practice, R is made variable (say 100 ohms total) so as to permit the amplifier to be balanced experimentally. One method is to connect a valve voltmeter across R_0, and to adjust R for minimum reading.

Instability may occur if R is too small.

A modified circuit is Fig. 13.46B in which the centre-tapped primary of T_1 is not necessary.

Fig. 13.46C permits both a.c. and d.c. balancing.

Fig. 13.46D keeps the circulating screen current out of the cathode circuit and so maintains the ratio between plate and screen currents at the negative voltage peak swing. Resistors R_1 are to prevent coupling from cathode to cathode through the screen by-pass condensers ; their values should be low—say 100 to 250 ohms each.

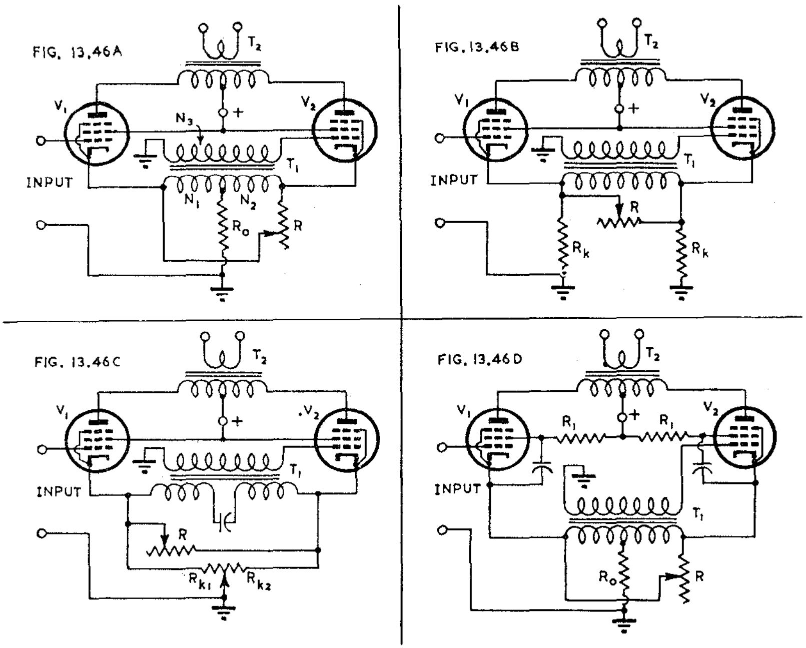


Fig. 13.46A. *Basic circuit of Parry Cathamplifier, (B) Modified circuit, (C) With both a.c. and d.c. balancing, (D) Keeps circulating screen current out of cathode circuit (Ref. E30).*

(ix) Extended Class A

Extended Class A is the name given to a push-pull amplifier using a triode and a pentode in parallel on each side. The amplifier operates entirely on the push-pull triodes at low levels, with the pentodes (or beam power amplifiers) cut off; at high levels the output is mainly from the pentodes. Consequently there is some curvature in the linearity (" transfer ") characteristic at the transition point. The total dissipation is, however, only about one third of that of a Class A amplifier with the same maximum power output (Ref. E31).

This principle might also possibly be applied to any Class AB or Class B amplifier merely to avoid current cut-off in the transformer and its resultant parasitics (Ref. E13) unless one with very low leakage inductance is used.

(x) Partial triode (" ultra-linear ") operation

See page 570.

SECTION 7: CLASS B AMPLIFIERS AND DRIVERS

(i) Introduction (ii) Power output and distortion—ideal conditions—Class B_2 (iii) Power output and distortion—practical conditions—Class B_2 (iv) Grid driving conditions (v) Design procedure for Class B_2 amplifiers (vi) Earthed-grid cathode coupled amplifiers (vii) Class B_1 amplifiers—quiescent push-pull.

(i) Introduction

A Class B amplifier is an amplifier in which the grid bias is approximately equal to the cut-off value, so that the plate current is approximately zero when no signal voltage is applied, and so that the plate current in a specific valve flows for approximately one half of each cycle when an alternating signal voltage is applied.

Class B amplifiers are in two main groups—firstly Class B_1 (otherwise known as quiescent push-pull, see below) in which no grid current is permitted to flow, secondly Class B_2 (generally abbreviated to Class B) in which grid current flows for at least part of the cycle.

Class B_2 amplifiers have inherently high odd harmonic distortion, even when the utmost care is taken in design and adjustment. This distortion frequently has a maximum value at quite a low power output, making this type of amplifier unsuitable for many applications. They are also comparatively expensive in that a driver valve and transformer together with two output valves form an integral part of the stage, and together give only the same order of sensitivity as a pentode.

The one outstanding advantage of a Class B_2 amplifier is in the very high plate-circuit efficiency, although the current drawn by the driver stage should be included. The principal applications are in battery-operated amplifiers, public address systems and the like.

The grid bias must be fixed (either battery or separate power supply) and special high-mu triodes have been produced to permit operation at zero bias to avoid the necessity for a bias supply. In the smaller sizes, twin triodes are commonly used. For the best results, accurate matching of the two valves is essential. If they are being operated at a negative bias it is possible to match their quiescent plate currents by adjusting the bias voltages separately. If the dynamic characteristics are not matched, a difference of 10% in the two plate currents, measured by d.c. milliammeters under operating conditions, will produce roughly 5% second harmonic distortion (Ref. E18). The matching of valves is covered in Sect. 5(v).

Well regulated plate, screen (if any) and bias supplies (if any) are essential.

The rate of change of current in each half of the output transformer at the plate current cut-off point is considerable, often resulting in parasitics

Parasitics in the grid circuit may be eliminated by the use of a driver transformer having low leakage inductance [Chapter 5 Sect. 3(iii)a] and, if necessary, by connecting a small fixed condenser from each grid to cathode—a typical value is 0.0005 μF—see (iv) below.

Parasitics in the plate circuit may be eliminated by the use of an output transformer with low leakage inductance—see Chapter 5 Sect. 3(iii)c—together with a series resistance-capacitance network connected across each half of the primary or alternatively from plate to plate.

Typical values for connection across each half of primary :

Load resistance (p-p)	6000	10 000	12 000	ohms
R	3300	5600	6800	ohms
C	0.05	0.05	0.05	μF

Typical values for connection from plate to plate :

Load resistance (p-p)	6000	10 000	12 000	ohms
R	6800	12 000	15 000	ohms
C	0.03	0.02	0.02	μF

Note—The McIntosh Amplifier may also be used for Class B operation—see Sect 8(iii).

(ii) Power output and distortion—ideal conditions—Class B_2

(Circuit Fig. 13.31 ; characteristics Fig. 13.36.)

If the input voltage is sinusoidal, the output transformer ideal, the plate characteristics equidistant straight lines, and the valves biased exactly to cut-off, the operating conditions will be—

Power output (total) : $P_0 = \frac{1}{2}I_{bm}(E_{bb} - E_{min})$ (1)

or $P_0 = \frac{1}{8}I_{bm}{}^2R_L = 2(E_{bb} - E_{min})^2/R_L$ (2)

where I_{bm} = maximum (peak) plate current of either valve

E_{min} = minimum plate voltage of either valve

and R_L = load resistance plate-to-plate.

Load resistance (plate-to-plate) : $R_L = 4(E_{bb} - E_{min})/I_{bm}$ (3)

Load resistance per valve : $R_L{}' = R_L/4 = (E_{bb} - E_{min})/I_{bm}$ (4)

Maximum power output is obtained when $R_L = 4r_p$

i.e. when $E_{min} = 0.5E_{bb}$

where r_p = plate resistance of one valve.

Average plate current (each valve) : $I_b = I_{bm}/\pi \approx 0.318I_{bm}$ (5)

Power input from plate-supply : $P_b = 2E_{bb}I_b \approx 0.637E_{bb}I_{bm}$ (6)

Plate circuit efficiency : $\eta = (1 - E_{min}/E_{bb}) \times 0.785$ (7)

Plate dissipation : P_p = d.c. power input − power output

$= I_{bm}(0.137E_{bb} + 0.5E_{min})$ (8)

Power output in terms of plate dissipation and plate circuit efficiency :

$P_0 = P_p\eta/(1 - \eta)$ (9)

The current in the secondary of the transformer will be sinusoidal, there being no distortion. The plate current in each valve will have the form of a rectified sine wave.

The slope of the composite characteristic is half that for Class A operation, so that the slope of the composite plate resistance will be twice that for Class A.

(iii) Power output and distortion—practical conditions—Class B_2

The practical treatment is based on the use of the composite characteristics [Sect. 5(ii)D]. If the valves are biased completely to cut-off, each half of the composite characteristic is identical with the individual valve characteristic, and a considerable degree of non-linearity occurs in the middle region (Fig. 13.47A). As a result, the greatest distortion of the loadline occurs with fairly small signals. For any particular characteristic there is one value of grid bias beyond which the distortion increases rapidly with increase of bias—there is a small quiescent plate current at this point (Fig. 13.47B). Even in this case the whole of the composite characteristic, except the small middle portion, is identical with the single valve characteristic.

It is therefore practicable to calculate the **power output** from the characteristics of a single valve (e.g. Fig. 13.48). The power output as calculated by eqns. (1) and (2) is modified by the presence of harmonics and should be multiplied by the factor

$F = (1 + H_3 - H_5 + H_7 - H_9 + H_{11})^2$ (10)

where H_3 = third harmonic distortion (i.e. I_3/I_1),

and H_5 = fifth harmonic distortion (i.e. I_5/I_1), etc.

The various harmonics are unpredictable, but at maximum power output H_3 usually predominates, and a purely arbitrary approximation is to take

$F = (1 + 0.6H_3)^2.$

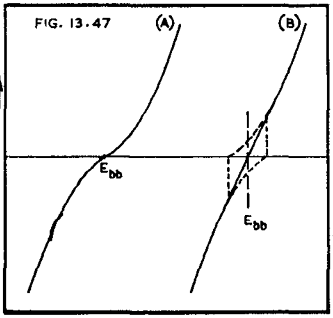
Usual values of H_3 at maximum power output vary from 0.05 to 0.10, so that F varies from 1.06 to 1.12, a reasonable design value being 1.08. Thus, from eqn. (2),

$$P_0 \approx 0.135 I_{bm}{}^2 R_L \qquad (11)$$

Alternatively, *the total power output may be calculated from a knowledge of the* average and quiescent currents of one valve, with an error not exceeding 9% (see eqns. 12 and 13 below)—

$$P_0 \approx 2.47(I_b - 0.25 I_{b0})^2 R_L \qquad (12)$$

The more accurate method is to use the composite characteristics (which are the same as the single valve characteristics of Fig. 13.48 over the range concerned) and to apply the " eleven selected ordinate method " [Sect. 3(iv)D] for the determination of power output and distortion. Although harmonics higher than the fifth are appreciable, it is difficult to calculate them graphically with any degree of accuracy.



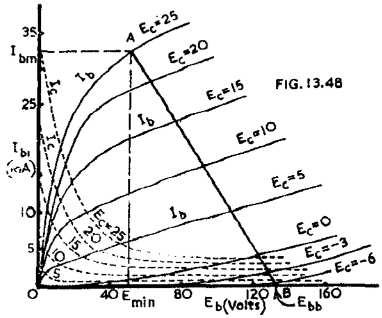


Fig. 13.47. *Composite characteristics of Class B amplifiers (A) biased to cut-off (B) biased to the point of minimum distortion.*

Fig. 13.48. *Plate and grid characteristics of a typical small battery-operated Class B triode* (1J6-G).

The **load resistance** may be calculated by eqn. (3) or (4).

Maximum power output is usually achieved when the slope of the loadline is numerically equal to the slope of the individual valve characteristic corresponding to the peak grid voltage, at its point of intersection with the loadline. A different loadline is usually necessary for each value of peak grid voltage. Each loadline should be checked for peak current, average current, plate dissipation and grid driving power.

When any loadline has been established as permissible, a higher load resistance may be used with perfect safety without any further checking. This will require less grid driving power, and will draw less plate current, but will give a lower power output; the distortion will be roughly unchanged, and the plate circuit efficiency will be higher.

The **average plate current** may be calculated by the accurate method of Sect. 5(iv) and eqn. (26). Alternatively, with an accuracy within 3% for third harmonic not exceeding 10% (Ref. E11)—

(for each valve) $I_b \approx 0.318 I_{bm} + 0.25 I_{b0}$ (13)

Eqn. (13) may also be put into the form—

$$I_{bm} \approx 3.14 I_b - 0.785 I_{b0} \qquad (14)$$

The average power input from the plate supply is given by

$$P_b = 2E_{bb} I_b \approx E_{bb}(0.637 I_{bm} + 0.5 I_{b0}). \qquad (15)$$

Plate circuit efficiency : $\eta = P_0 / 2E_{bb} I_b$ (16)

[Usual values of plate circuit efficiency are from 50% to 60%.]

Plate dissipation : $P_p \approx E_{bb}(0.637 I_{bm} + 0.5 I_{b0}) - P_0$ (17)

With a plate circuit efficiency of 60%, the power output is 1.5 times the total plate dissipation, whereas with Class A operation and a plate circuit efficiency of 25%, the power output is 0.25 times the plate dissipation. Thus, if in both cases the plate dissipation is the only limiting factor, six times more output may be obtained from the same valves in Class B than in Class A.

(iv) Grid driving conditions

Grid current characteristics are provided for use with valves suitable for Class B amplification (e.g. Fig. 13.48). It is usual practice to select a likely peak positive grid characteristic and loadline, then to calculate the grid driving power and power output. This may be repeated for several other loadlines, and the final choice will be made after considering all the relevant features.

In the case of a particular loadline such as AB in Fig. 13.48, the peak grid current may be determined by noting the intersection of the vertical through A and the grid current curve (shown with a broken line) corresponding to the E_c characteristic which passes through A. This is shown in Fig. 13.49 where $-E_c$ indicates the fixed grid bias and the positive grid current commences to flow at approximately $E_c = 0$. The peak grid current is I_{gm} and the peak signal grid voltage is E_{gm}, corresponding to a positive grid bias voltage E_{cm}. The peak grid input power is then given by

FIG. 13.49

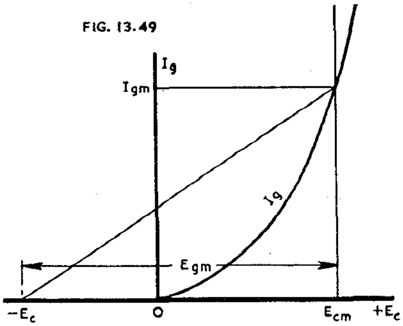


Fig. 13.49. *Typical grid mutual characteristic illustrating Class B operating conditions.*

$$P_{gm} = E_{gm} I_{gm} \tag{18}$$

and the minimum grid input resistance is given by

$$r_{g\,min} = E_{gm}/I_{gm} \tag{19}$$

[Note : The minimum variational grid resistance is derived from the slope of the I_g curve—it is always less than $r_{g\,min}$.]

The driver valve has to supply this peak power, plus transformer losses, into a half-secondary load varying from infinity at low input levels to $r_{g\,min}$ at the maximum input. The basic driver circuit is shown in Fig. 13.50A, where V_1 is the driver valve and T_1 the step-down transformer with a primary to half-secondary turns ratio N_1/N_2. In practice, the transformer has losses, and it may be represented by the equivalent circuit Fig. 5.9 (omitting C_W).

The efficiency is usually calculated (or measured) at 400 c/s where r_1, r_2 and R_0 are the principal causes of loss. An efficiency of from 70 to 80% at peak power is typical of good practice. The voltage applied to the grid of V_2, provided that the iron losses are small, is approximately [Chapter 5 Sect. 2(ii)].

$$E_{g2} \approx E_p \eta (N_2/N_1) \tag{20}$$

where η = percentage efficiency $\div$ 100,
and E_p = voltage applied across primary.

The load presented to V_1 is roughly as calculated for an ideal transformer [Chapter 5 Sect. 2(ii)] driving the same grid.

At high audio frequencies, the effects of L_1 and L_2 (Fig. 5.9) become appreciable, causing an additional loss of voltage during the time of grid current flow, and a tendency towards instability. The instability is brought about by a negative input resistance which some valves possess over a portion of the grid characteristic [Chapter 2 Sect. 2(iii)]. It is therefore important to reduce the leakage inductance of the transformer to the lowest possible value. Fortunately the bad effects of any remaining leakage inductance may be minimized by connecting a small condenser across each half of the secondary so as to resonate at a frequency about 1.5 times the highest frequency to be amplified (Ref. E18). The condenser and leakage inductance form a half-section of a simple two-element constant k type low-pass filter [Chapter 4 Sect. 8(vii)].

At low audio frequencies the primary inductance L_0 causes loss of gain as with any a-f transformer. It should be designed for an unloaded secondary (Chapter 5 Sect. 3) because it is unloaded for part of the cycle. Secondary loading resistances should be avoided with Class A triode drivers (Refs. E17, E18).

In order to avoid excessive distortion due to the non-linear grid current characteristic, the effective impedance looking backwards from the grid-cathode terminals towards the driver, must be small compared with the effective input impedance of the valve. The " looking backwards " impedance has as its components (Fig. 5.9) :

" Driver resistance " $R' = r_2 + (N_2/N_1)^2(r_{p1} + r_1)$ (21)

" Driver inductance " $L' = L_2 + (N_2/N_1)^2 L_1$ (22)

It is usual to restrict the driver resistance R' to a value less than 0.2 $r_{g\,min}$ for the most favourable conditions with high-mu valves and with limited grid drive, where $r_{g\,min}$ does not differ seriously from the minimum variational grid resistance ; in other cases it should be less than one fifth of the minimum variational grid resistance (Refs. E2, E12). For minimum distortion, R' should be made as low as practicable. Hence the transformer usually has a step-down ratio from primary to half-secondary.

The " driver inductance," if large and uncompensated, produces an effect like a faint high pitched hiss or scratch that rises and falls with the signal (Ref. E18).

The effect of the varying load resistance on the driver valve is shown by the curved loadline in Fig. 13.50B. The horizontal portion corresponds to the conditions without grid current flow, while the slope at any other point corresponds to the variational grid resistance at that point. The broken line joining the two extremities has a slope corresponding to a resistance of $r_{g\,min}(N_1/N_2)^2$; this should not be less than $4r_{p1}$. The driver valve should preferably be operated near its maximum rated plate voltage and dissipation.

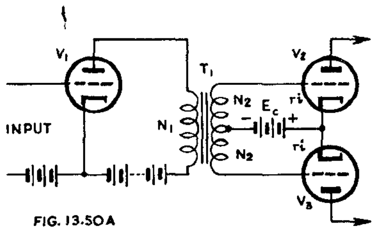


Fig. 13.50A. *Fundamental driver and Class B (or AB) amplifier circuit.*

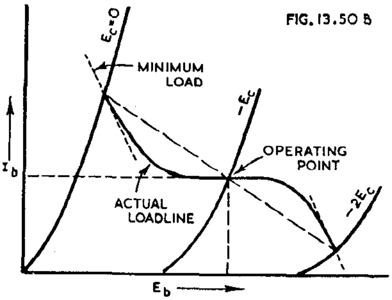


Fig. 13.50B. *Loadline on a typical triode driver.*

As the output valves are driven harder, the driver valve is called upon to provide increased power into a decreased minimum grid resistance ; this necessitates a greater step-down ratio in the transformer. The extreme limit is when the " diode line " is reached. As a guide which may be used for a first trial, the minimum plate voltage may be taken as twice the "diode line" voltage at the peak current level. The diode line is shown in many characteristic curves, and is the envelope of the characteristics—it is the line where the grid loses control, in other words the minimum plate voltage for any specified plate current no matter how positive the grid may be.

The cathode follower makes an excellent driver for all forms of amplifiers drawing grid current, having very low effective plate resistance and low distortion—see Chapter 7 Sect. 2(i).

A cathode follower has been applied as a transformerless driver, one 6SN7-GT twin triode being used as a push-pull driver to the grids of two 6L6 valves giving an output of 47 watts (Ref. E22). However, this is not usually the most efficient arrangement—see Chapter 7 Sect. 1(ii).

(v) Design procedure for Class B_2 amplifiers

The major difficulty is in keeping the electrode voltages stabilized.

The following procedure applies to the case where the voltages are stabilized and the plate voltage and desired power output are known.

1. Assume a value of plate-to-plate load resistance R_L. Draw a loadline with a slope corresponding to $R_L/4$ on the characteristics for one valve, through the point $E_b = E_{bb}$, $I_b = 0$.
2. Select a value of grid bias to give the minimum distortion of the composite characteristic (Fig. 13.47).
3. Determine the quiescent plate current I_{b0} of one valve.
4. Calculate the peak plate current from the approximation :

$$I_{bm} \approx \sqrt{7.4P_0/R_L} \tag{23}$$

5. Calculate the average power input from the plate supply, using eqn. (15).
6. Calculate the plate dissipation from equation (17).
7. Check the plate dissipation and the peak plate current to see that they do not exceed the maximum ratings.
8. Determine the maximum positive grid voltage E_{cm} from the loadline and I_{bm}.
9. Determine the peak grid current I_{gm} corresponding to extreme point of the loadline.
10. Calculate the peak signal grid voltage : $E_{gm} = E_{cm} + E_c$.
11. Calculate the peak grid input power : $P_{gm} = E_{gm}I_{gm}$.
12. Calculate the minimum input resistance : $r_{g\,min} = E_{gm}/I_{gm}$.
13. Repeat steps 1 to 12 for several other values of R_L and select what appears to be the best compromise, with a view to the lowest driving power
14. Assume a reasonable peak power transformer efficiency—say 70%. This value is used in the following step.
15. Select a driver valve with a maximum power output at least $0.9P_{gm}$ under typical operating conditions (this makes allowance for the higher load resistance).
16. Select a driver load resistance twice the typical value or at least four times the plate resistance.
17. Draw the assumed (straight) driver loadline on the driver plate characteristics and determine the maximum available peak signal voltage E_{pm}.
18. Determine the transformer ratio from primary to half-secondary :
    $N_1/N_2 = \eta E_{pm}/E_{gm}$.
19. Calculate the " driver resistance " from eqn. (21) using values for r_1 and r_2 indicated* from similar transformers. These resistances are usually less than 1900 and 350 ohms respectively for 10 watt amplifiers, or 2700 and 500 ohms respectively for low power battery amplifiers.
20. Check the " driver resistance " to see that it is less than $0.2r_{g\,min}$ for high-mu valves or $0.1r_{g\,min}$ for other cases (this assumes that $r_{g\,min}$ is approximately twice the minimum variational grid resistance).

*Alternatively the driver transformer may be tentatively assumed to have 10% primary and 10% secondary copper losses, and 10% iron losses. On this basis, $r_2 = r_{g\,min}/7$ and $r_1 \approx r_{g\,min}(N_1^2/N_2^2)/9$.

(vi) Earthed-grid cathode-coupled amplifiers

A cathode-coupled output stage, with earthed grids, has the advantages of a much more nearly constant input resistance than a Class B stage, but it requires more driver power for the same output. It is more suited to high-power amplifiers or modulators than to conventional a-f amplifiers (Ref. E15). The increase in plate resistance due to feedback is disadvantageous.

(vii) Class B_1 amplifiers—Quiescent push-pull

Class B_1 amplifiers do not draw any grid current whatever, but in other respects resemble Class B_2 amplifiers. Either triodes or pentodes may be used, the latter being more popular—they are used as " quiescent push-pull pentodes " in battery receivers. The two valves (or units) must be very accurately matched, and an adjustable screen voltage is desirable. A high-ratio step-up input transformer is used to supply the high peak signal voltage, nearly equal to the cut-off bias. Both the input and output transformers should have low capacitance and leakage inductance. The output transformer should have its half-primary inductance, and its leakage inductance, accurately balanced. Condensers (0.002 to 0.005 μF) are usually connected across each half primary (Ref. E4).

SECTION 8 : CLASS AB_2 AMPLIFIERS

(i) Introduction (ii) Bias and screen stabilized Class AB_2 amplifier (iii) McIntosh amplifier.

(i) Introduction

Class AB_2 amplifiers closely resemble Class B_2 amplifiers, but the valves are biased as for Class AB_1 operation. Consequently they are less critical with regard to matching, and the distortion which occurs in the plate circuit is much the same as for Class AB_1. The plate circuit efficiency is intermediate between Class AB_1 and B_2 operation —typical values for triodes are from 40 to 48%. The variation in plate current from zero to maximum signal is less than with Class B operation, being of the order of 1 : 2.

The matching of valves is covered in Sect. 5(v).

Pentodes and beam power amplifiers may be used quite successfully in Class AB_2 operation.

Fixed bias is essential.

Type 6L6 or 807 beam power amplifiers may be used in Class AB_2 with plate circuit efficiencies of about 65% (or 61% including screen losses). The peak grid input power does not exceed 0.27 watt for power outputs from 30 to 80 watts. The total harmonic distortion under ideal laboratory conditions is 2% at maximum signal and less than 2% at all lower output levels (this assumes a low-distortion driver stage).

Well regulated plate, screen and bias power supplies are essential.

Parasitics in plate and grid circuits—see Sect. 7(i).

References E20, E21.

(ii) Bias and screen stabilized Class AB_2 amplifier (Ref. E26)

Fig. 13.50C gives a circuit with 807 valves operated in Class AB_2 with stabilization of the bias and screen supply provided by the driver valve. Two high tension sources are used, obtained from separate windings on the power transformer, and neither needs good regulation, so that condenser input filters can be used.

FIG. 13.50C

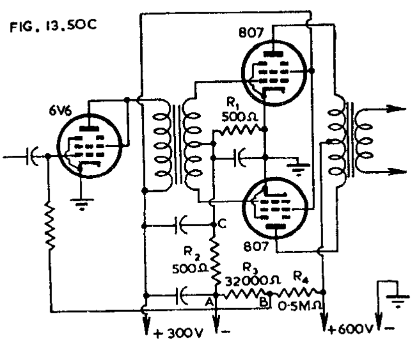


Fig. 13.50C. *Class AB_2 amplifier with stabilized bias and screen voltages (Ref. E26).*

The circuit offers three types of compensation, the combined effect being to allow the output valves to draw a minimum of plate current in the no signal condition, although still being capable of delivering an output of 60 watts for small distortion. The three types of compensation are

(a) As the 807 plate voltage falls, due to increasing plate current with signal input, the screen voltage is increased and the bias reduced.

(b) When the 807's are driven into grid current their bias is increased due to the current in the 500 ohm resistor

R_1. This is minimized by a d.c. negative feedback circuit giving several times reduction of the effect, and is offset by a corresponding increase in 807 screen voltage.

(c) Any tendency for the 807 screen voltage to fall and for the bias voltage to increase with rising 807 screen current due to signal input is minimized by the same negative feedback circuit.

Referring to Fig. 13.50C, the voltage divider from the positive side of the 600 volt supply to the negative side of the 300 volt supply has values to provide a suitable bias for the triode-operated 6V6 so that it draws about 45 mA with a standing plate voltage of 250 V. A decrease in the 807 plate voltage causes the 6V6 grid to become more negative, decreasing its plate current which also flows through R_1 and so decreases the 807 bias. The low voltage power supply is deliberately given poor regulation with resistive filtering so that the decrease in 6V6 plate current appreciably increases its plate voltage and thus the 807 screen voltage.

The 807 grid current at large output levels increases the negative potential at C and also at A and B and thus increases the 6V6 bias. This reduces the 6V6 plate current which makes point C less negative and minimizes the increase in 807 bias, and at the same time increases the 807 screen voltage thus off-setting the bias increase.

The 807 screen current rises with increasing signal input and this tends to decrease the screen voltage and increase the 807 bias. However the increased bias is also applied to the 6V6, reducing its plate current and thus tending to restore the 807 bias and screen voltages to their original values.

Between no output and full output the 6V6 plate current falls from 45 mA to 25 mA but the valve is still well able to provide the small power required to drive the 807's.

An additional feature of the circuit is that a very large input reduces the ability of the driver to overload the output valves, and after full output is reached very effective limiting is provided.

(iii) McIntosh Amplifier (Ref. E28)

With any push-pull amplifier in which each valve is cut-off during portion of the cycle, some form of quasi-transient distortion tends to occur at a point in each cycle at the higher audio frequencies (Ref. E13). This distortion is caused by the leakage reactance of the primary of the output transformer, which cannot be reduced sufficiently by conventional transformer design technique. A completely new approach to the problem is made by the McIntosh amplifier which incorporates special types of driver transformer and output transformer, together with many other novel features. The basic principles are indicated in Fig. 13.50D in which both driver transformer T_1 and output transformer T_2 have two windings wound together in a bifilar manner so that the coupling between them is almost unity. It is claimed that it is practicable to wind coils with a ratio of primary inductance to leakage inductance better than 200 000 to 1, whereas conventional transformers do not nearly reach the minimum requirement (for low distortion) of 80 000 to 1. This type of transformer is cheaper to wind than a sectionalized winding as used in conventional high quality transformers.

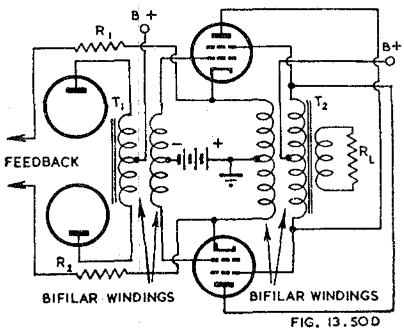
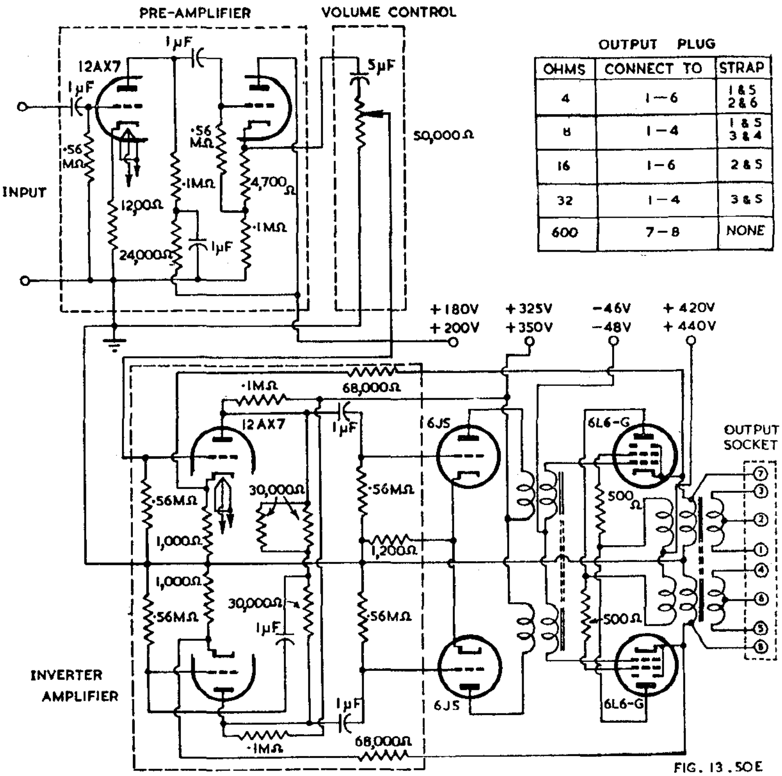


Fig. 13.50D. *Basic principles of McIntosh Amplifier* (*Ref.* E28).

Each output valve works into two primary sections, one in its plate circuit and the other in its cathode circuit, but these have practically unity coupling. The effective number of primary turns for each valve is equal to the total turns for each of the bifilar windings. The output transformer should therefore be designed to have a total impedance on each of its primaries equal to one quarter of the plate-to-plate load impedance.

In the case of the amplifier of Fig. 13.50E the plate-to-plate load imped-



OHMS	CONNECT TO	STRAP
4	1—6	1 & 5 2 & 6
8	1—4	1 & 5 3 & 4
16	1—6	2 & 5
32	1—4	3 & 5
600	7—8	NONE

Fig. 13.50E. *Circuit diagram of complete* 50 *watt McIntosh Amplifier* (*Ref.* E28).

ance is 4000 ohms and the total impedance of each primary winding is 1000 ohms ; the impedance from each cathode to earth is only 250 ohms. These low impedances reduce the effects of capacitive shunting and thereby improve the high frequency performance.

The voltage from each screen to cathode is maintained constant by the unity coupling between the two halves of the bifilar windings, no screen by-pass capacitor being required. This arrangement, however, has the limitation that it can only be used for equal plate and screen voltages.

The driver transformer T_1 makes use of the same bifilar winding method adopted in the output transformer. The primary impedance in this design is above 100 000 ohms from 20 to 30 000 c/s, while the response of the whole transformer is within 0.1 db from 18 to 30 000 c/s. These high performances are made necessary by the inclusion of this transformer in the second feedback path of the whole amplifier.

The method of loading the output stage, with half the load in the plate and half in the cathode circuit, provides negative feedback as a half-way step towards a cathode follower. Additional feedback is achieved by connecting suitable resistors between the cathodes of the output valves and the cathodes of the phase inverter stage. The complete amplifier (Fig. 13.50E) has a typical harmonic distortion of 0.2% from 50 to 10 000 c/s, rising to 0.5% at 20 c/s and 0.35% at 20 000 c/s, at an output level of 50 watts. The frequency response under the same conditions, measured on the secondary of the transformer, is level from 20 to 30 000 c/s, − 0.4 db at 10 c/s and − 0.3 db at 50 000 c/s. The phase shift is zero from 50 to 20 000 c/s, − 10° at 20 c/s and + 4° at 50 000 c/s.

The output resistance is one tenth of the load resistance, thus giving good damping and regulation.

These same principles may be applied to any type of push-pull output whether triode or pentode, with any class of operation, A, AB_1, AB_2 or B.

The high power output of the McIntosh amplifier as described above is due to the operation of the screens at voltages greater than 400 volts, which is very considerably in excess of the maximum rating of 270 volts (design centre) for type 6L6 or 300 volts (absolute) for type 807. It is unfortunate that the McIntosh amplifier is limited to operation with equal plate and screen voltages, but these should always be within the maximum ratings for the particular valve type.

See also Refs. E31, G1.

The principle of combined plate and cathode loading is applied in some high-fidelity Class A amplifiers, e.g. Acoustical QUAD, Refs. H4, H6.

SECTION 9: CATHODE-FOLLOWER POWER AMPLIFIERS

The principles of cathode followers have been covered in Chapter 7 Sect. 2(i). A cathode follower may be used either as driver for a Class B or AB_2 stage, or as the output stage itself.

A cathode follower forms almost an ideal driver stage, having very low plate resistance and distortion, although it requires a high input voltage. It is commonly used, either singly or in push-pull, in high power a-f amplifiers where the distortion must be reduced as much as possible. If parallel-feed is used, the hum is reduced by the factor $1/(\mu + 1)$; see Chapter 7 Sect. 2(ix) Case 4. A cathode follower driver stabilizes the a-f signal voltage, but does not stabilize the grid bias.

Cathode follower output stages introduce serious problems, and are not suitable for general use, even though their low plate resistance and low distortion appear attractive. The difficulty is in the high input voltage which is beyond the capabilities of a resistance-coupled stage operating on the same plate supply voltage. Two methods are practicable, either a step-up transformer in the plate circuit of a general purpose triode, or a resistance-coupled amplifier with a plate supply voltage about 3 times the plate-cathode voltage of the cathode follower. In order to take advantage of the low distortion of the cathode follower, the preceding stage should also have low distortion. A general purpose triode is to be preferred to a pentode or high-mu triode with resistance coupling, and it may have an unbypassed cathode resistor.

One practical amplifier which has been described in the literature (Ref. F3) uses 700 volts supply voltage to the 6SN7 penultimate stage and eight 6V6-GT valves in push-pull parallel operation in the cathode follower output stage. Negative feedback is used from the secondary of the output transformer, and the damping on the loudspeaker is as high as practicable. However, the total harmonic distortion at 50 c/s is over 1% at 8 watts output, and 1.7% at 20 watts. The high output voltage which must be delivered by the resistance-coupled penultimate stage thus shows its effect on the distortion, even though the plate supply voltage has been increased to a dangerously high value.

SECTION 10: SPECIAL FEATURES

(i) Grid circuit resistance (ii) Grid bias sources (iii) Miller Effect (iv) 26 volt operation (v) Hum.

(i) Grid circuit resistance

A maximum value of grid circuit resistance is usually specified by the valve manufacturer, a higher value being usually permitted with cathode (self) bias than with fixed bias. The reason for the latter is that cathode bias provides increased bias as the plate current rises, and so gives a degree of protection against "creeping" plate current. This effect is due to the combined gas and grid emission currents which flow through the grid-cathode circuit—see Chapter 2 Sect. 2(iii), also Chapter 3 Sect. 1(v)d, and Chapter 12 Sect. 2(iii)E, (iv)B, (iv)C, (iv)D ; Sect. 3(iv)C, (v).

Typical values (e.g. type 6V6-GT) are :

With cathode bias	0.5 megohm (maximum)
With fixed bias	0.1 megohm (maximum).

If the heater is operated, even for limited periods, more than 10% above its average rated voltage, the grid circuit resistance must be reduced considerably ; this holds in automobile receivers.

If partial self-bias operation (" back bias ") is used, the maximum value of grid circuit resistance may be found from the relation :

$$R_{gm} = R_{gf} + P(R_{gs} - R_{gf}) \qquad (1)$$

where R_{gf} = max. grid circuit resistance (fixed bias)
R_{gs} = max. grid circuit resistance (self bias)
and P = ratio of cathode current in output valve to the total current flowing through the bias resistor.

In cases where the maximum value of grid resistance for a particular application is not specified, the procedure of Chapter 3 Sect. 1(v)d should be followed, using eqn. (6) for pentodes or tetrodes, or eqn. (7) for triodes. The value of maximum reverse grid current (ΔI_{c1}) should be obtained from the specifications or from the valve manufacturer ; failing this, a value of 2μA may be tentatively assumed for valves of the 6F6, 6V6 class, and 4μA for valves having higher cathode currents.

If either the value of R_{g1} for cathode bias, or that for fixed bias is specified, the value for the other may be calculated from Chapter 3 Sect. 1 eqns. (8) to (13).

(ii) Grid bias sources

Fixed bias is normally obtained from a separate power source with rectifier, filter and load resistance. A typical circuit is Fig. 13.51 in which V_1 is the usual rectifier, V_2 is the bias rectifier and R_4 the bias load resistance. V_2 may be any half-wave indirectly-heated rectifier ; if the V_2 heater is operated from a common heater winding, V_2 should be a type capable of withstanding a high voltage between heater and cathode. Alternatively a shunt diode may be used (see Chapter 30 Sect. 6). If the output stage draws positive grid current, as in Class AB$_2$ or overloaded Class A operation, the grid current flowing through the bias load resistance increases the bias voltage. It is therefore advisable to design this to be as low as practicable. The effect of negative grid current on the bias is usually negligible.

FIG. 13.51

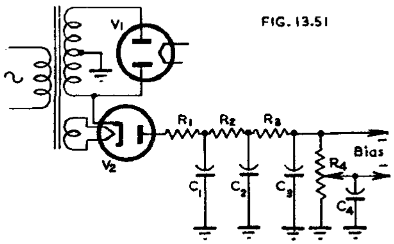


Fig. 13.51. *Method of obtaining fixed bias using half of the transformer secondary (plate) winding. Typical values of resistors are :* $R_1 = 2500$, $R_2 = 25\,000$, $R_3 = 15\,000$, $R_4 = 3000$ *ohms ;* $C_1 = 8\mu F$, $C_2 = 16\mu F$, $C_3 = C_4 = 50\mu F$. *Alternatively* R_2 *may be replaced by a choke.*

Voltage stabilized grid bias supplies are sometimes used (see Chapter 33).

It is sometimes convenient to have two plate voltage supplies, one to provide for the plates of the output stage only, the other for the earlier stages together with the screens of the output stage and grid bias. The second supply is loaded by a heavy current voltage divider (say 150 mA total drain) tapped near the negative end and the tapping connected to the cathodes of the output stage ; one section provides grid bias, and the other section provides the positive potentials.

Back bias is intermediate between cathode bias and fixed bias, so far as its constancy is concerned. With this arrangement, the whole return current of a receiver or amplifier is passed through a resistor between the cathodes (which are generally earthed) and $- E_{bb}$ (Fig. 13.52). The value of R_1 is given by E_{c1}/I where E_{c1} is

FIG. 13.52

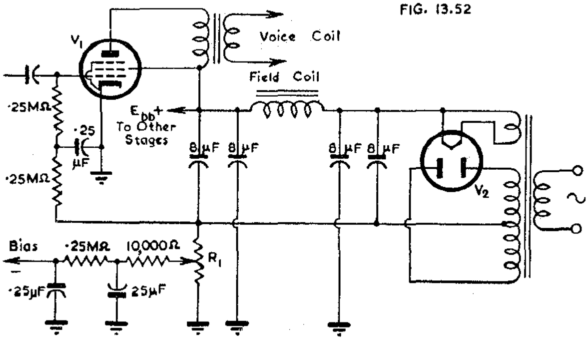


Fig. 13.52. *Circuit of portion of a receiver using back bias. V_1 is the power amplifier, V_2 is the rectifier.*

the bias required by V_1 and I is the total current passed through R_1. Bias for earlier stages is obtainable from a tapping or tappings on R_1. A hum filter is necessary for each separate bias system. [Note that the curved plate of an electrolytic condenser represents the negative terminal ; while on a paper condenser the curved plate represents the outside electrode.] Back bias works best in large receivers where the total current is much greater than the cathode current of the power amplifier.

An alternative form having fewer components is shown in Fig. 13.53. Here the field coil is in the place of R_1 in Fig. 13.52. The bias on V_1 is equal to the voltage drop across the field coil multiplied by $R_2/(R_2 + 0.5 \text{ megohm})$. The filtering in this circuit is not so complete as in Fig. 13.52.

FIG. 13.53

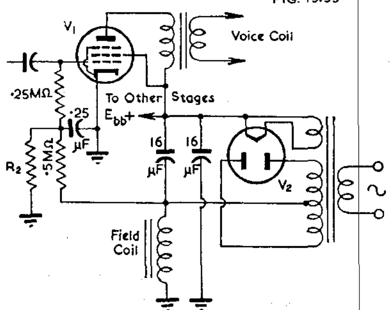


Fig. 13.53. *Alternative form of back bias, with the speaker field coil in the negative lead.*

(iii) Miller Effect

The Miller Effect has been introduced in Chapter 12 Sect. 2(xi). If the output valve is type 6V6-GT under typical operating conditions, $A = 17$ and $C_{gp} = 0.7$ $\mu\mu$F, then the additional input capacitance, due to C_{gp} alone, with a resistive load is $18 \times 0.7 = 12.6$ $\mu\mu$F. If the valve is a triode the effect is more pronounced. For example, with type 2A3 under typical operating conditions $A = 2.9$ and $C_{gp} = 16.5$ $\mu\mu$F ; the additional input capacitance is $3.9 \times 16.5 = 64$ $\mu\mu$F.

(iv) 26 volt operation

Standard power valves give very limited power output with plate and screen both at 26 or 28 volts. Special types (e.g. 28D7) have been developed to give higher power output under these conditions.

(v) Hum from plate and screen supplies

The hum of amplifiers with and without feedback is covered in Chapter 7 Sect. 2(ix)C.

Power amplifiers not using feedback are here summarized briefly, using the conditions for types 2A3 and 6V6 as in Chapter 7.

Triode with (a) conventional output transformer—
Relative hum voltage = 0.85
(b) parallel feed (series inductor)—
Relative hum voltage = 0.024

Pentode with conventional output transformer
and (a) common plate and screen supplies—
Relative hum voltage = 2.05
(b) screen perfectly filtered—
Relative hum voltage = 0.09

Pentode with parallel feed (series inductor)
(a) common plate and screen supplies
Relative hum voltage = 1.78
(b) screen perfectly filtered
Relative hum voltage = 0.179

See also Chapter 31 Sect. 4(ii)—Effect of the output valve on hum originating in the plate supply voltage.

SECTION 11 : COMPLETE AMPLIFIERS

(i) Introduction (ii) Design procedure and examples (iii) Loudspeaker load.

(i) Introduction

While every amplifier must be built up of individual stages, the purpose of the design is to produce a complete amplifier having specified characteristics in regard to overall gain, maximum power output, frequency range, output resistance, distortion, hum, noise level and special features such as overload characteristics and tone control. A designer's job is to produce the required results at the minimum cost within any limitations imposed by space or by the availability of valve types.

(ii) Design procedure and examples

The correct procedure is to commence with the loudspeaker, and then to work backwards through the amplifier. For example, after investigation of the loudspeaker and speaker transformer efficiency (see Chapter 20 Sect. 6) it should be possible to estimate the maximum power output required from the output valve(s). Then the output valve(s) should be selected allowing for the possible use of feedback. Then the preceding stage should be designed to work into the known following grid resistor and effective input capacitance with a satisfactory frequency response and distortion level. Finally, the input stage should be designed to give the required hum and noise performance. If, after a first calculation, the overall gain is too high, it may be reduced by one of many expedients such as increasing (or adding) negative feedback on the output stage, removing a cathode by-pass, changing a voltage amplifier stage from pentode to triode operation, or changing one or more valves to a lower gain type. In large amplifiers a pre-set second volume control may be used, to be adjusted in the laboratory.

The following information is a general guide in the design of typical amplifiers
A = public address (speech)
B = typical radio receiver (a-f amplifier only)
C = good quality radio receiver or amplifier
D = high fidelity receiver or amplifier

Performance applies to amplifier only (resistance load).

Category	A	B	C	D	
Frequency range	200-5 000	100-5 000	50-10 000	40-15 000	c/s
Attenuation at limits	3	3	2	1	db
Total harmonic distortion*	10	10	2	<1	%
Output resistance	—	—	$<R_L/2$	$<R_L/10$‡	
Noise level†	—	−25	−25	−30	dbm
Hum level (120 c/s) **	—	−15 to −30	−35 to −40	−40 to −45	dbm
Intermodulation distortion*— r.m.s. sum (voltage ratio 4 : 1)	40	40	8	<4	%

*at maximum power output.

**For 3% loudspeaker efficiency. In all cases the hum should not be audible under listening conditions. See Chapter 31 Sect. 4(iii) for hum.

†For room volume 3000 cubic feet, 3% loudspeaker efficiency—see Chapter 14 Sect. 7(v). For measurement of noise see Chapter 19 Sect. 6.

‡Or as required to give the desired degree of damping at the bass resonant frequency.

See also Chapter 14 Sect. 12(i) for high fidelity.

With careful design and the use of negative feedback it is possible to make the total harmonic distortion of the amplifier less than 0.1% and the intermodulation distortion less than 0.5%. This has been achieved in the circuit of Fig. 7.44 and in the Williamson amplifier on which it is based (the new version of the Williamson amplifier is given in Figs. 17.35 B, C, D, E, F, G), also in the Leak amplifier. See Refs. F1, F4, F5, H4, H15, H16.

An example of category C is given in Figs. 7.42 and 7.43.

Undoubtedly one of the best and most versatile single-ended amplifiers incorporating negative feedback (and, if desired, also tone control) is the circuit Fig. 7.33. The feedback should be as high as permissible, being usually limited by the sensitivity required by a pickup. The only serious limitation is that V_1 cannot be a combined second detector and amplifier.

The **Lincoln Walsh amplifier** with automatic bias control (E24, E25) is quoted as having intermodulation distortion of 0.2% at 5 watts, 0.6% at 10 watts and 1.7% at 25 watts, the power in each case being the low frequency power only (50 c/s). The valves used in this amplifier are type 6B4-G (similar to 2A3).

The **McIntosh 50 watt Class AB_2 amplifier** (Fig. 13.50E) has harmonic distortion less than 0.5% from 20 to 30 000 c/s, with exceptionally flat frequency response and low phase shift—see Sect. 8(iii).

References to complete amplifiers: Refs. (F) and (H).

(iii) Loudspeaker load

Most amplifiers are designed to work into a constant resistive load, whereas a loudspeaker presents a far from constant impedance which is largely inductive or capacitive at almost all frequencies within its range. This feature is covered from the loudspeaker angle in Chapter 20 Sects. 1, 2 and 7. See also Sect. 2(iv) for triodes and Sect. 3(viii) for pentodes.

A simple 2 stage amplifier with a triode output stage is shown in Fig. 13.54, together with the fundamental, second, third and fourth harmonics (see Chapter 20 for comments). An obvious conclusion is that all amplifiers should be tested for output and distortion while delivering full power into a loudspeaker load. While highly desirable, this requires a large sound-proof room for all except low levels. An alternative which is highly recommended is to construct a dummy load which has the same impedance as the loudspeaker at 400 c/s, at 10 000 c/s, and at the bass resonant frequency, based on the circuit of Fig. 20.3. The value of $(R_0 + R_1)$ in Fig. 20.3 should be equal to the measured value of the working impedance of the loudspeaker at the bass resonant frequency as described in Chapter 20 Sect. 2(iv).

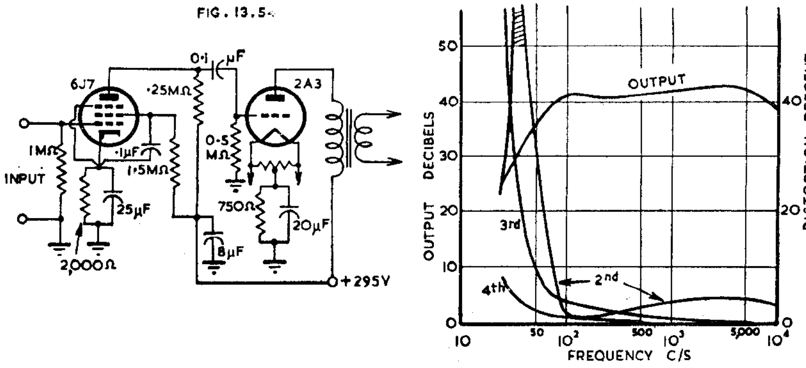


Fig. 13.54. *Harmonic distortion of amplifier having type* 6J7 *pentode coupled to* 2A3 *triode loaded by a loudspeaker on a flat baffle. The distortion was measured by a wave analyser connected across the voice coil. The bass resonance frequency is* 70 *c/s.*

SECTION 12 : REFERENCES

(A) REFERENCES TO OPTIMUM LOAD RESISTANCE AND DISTORTION

A1. Brain, B. C. " Output characteristics of thermionic amplifiers " W.E. 6.66 (March 1929) 119.
A2. Notthingham, W. B. " Optimum conditions for maximum power in Class A amplifiers " Proc. I.R.E. 29.12 (Dec. 1941) 620.
A3. Editorial " Optimum conditions in Class A amplifiers " W.E. 20.233 (Feb. 1943) 53.
A4. Hadfield, B. M. (letter) " Optimum conditions in Class A amplifiers," W.E. 20.235 (April 1943) 181.
A5. Sturley, K. R. " Optimum load, Ra or 2Ra " ? W.W. 50.5 (May 1944) 150.
A6. Sturley, K. R. (letter) " Optimum conditions in Class A amplifiers " W.E. 20.235 (April 1943) 181.
A7. Bradshaw, E. (letter) " Optimum conditions in Class A amplifiers " W.E. 20.237 (June 1943) 303.
A8. Benham, W. E. (letter) " Optimum conditions in Class A amplifiers " W.E. 20.237 (June 1943) 302.
A9. Gladwin, A. S. (letter) " Optimum conditions in Class A amplifiers " W.E. 20.240 (Sept. 1943) 436.
A10. Date, W. H. (letter) " Optimum load," W.W. 50.7 (July 1944) 222.
A11. Good, E. F. (letter) " Optimum load " W.W. 50.8 (Aug. 1944) 252.
A12. Hughes, E. " Optimum valve load—unified treatment for different operating conditions " W.W. 51.8 (Aug. 1945) 246.
A13. Foster, H. G. " Load conditions in Class A triode amplifiers " Electronic Eng. 19.227 (Jan. 1947) 11.
A14. Sturley, K. R. (book) " Radio Receiver Design " Part II pp. 56-64 (Chapman and Hall, London, 1945).
A15. Warner, J. C., and A. V. Loughren " The output characteristics of amplifier tubes " Proc. I.R.E. 14.6 (Dec. 1926) 735.
A16. Kilgour, C. E. " Graphical analysis of output tube performance " Proc. I.R.E. 19.1 (Jan. 1931) 42.
A17. Lucas, G. S. C. " Distortion in valve characteristics " W.E. 8.98 (Nov. 1931) 595. Letter P. K. Turner 8.99 (Dec. 1931) 660.
A18. Nelson, J. R. " Calculation of output and distortion in symmetrical output systems," Proc. I.R.E. 20.11 (Nov. 1932) 1763.
A19. Ferris, W. R. " Graphical harmonic analysis for determining modulation distortion in amplifier tubes " Proc. I.R.E. 23.5 (May 1935) 510.
A20. Mouromtseff, I. E., and H. N. Kozanowski " A short-cut method for calculation of harmonic distortion in wave modulation " Proc. I.R.E. 22.9 (Sept. 1934) 1090.
A21. Hutcheson, J. A. " Graphical harmonic analysis " Elect. 9.1 (Jan. 1936) 16.
A22. Harries, J. H. O. " Amplitude distortion " W.E. 14.161 (Sept. 1937) 63.
A23. Kemp, P. " Harmonic analysis of waves up to eleventh harmonic " Electronic Eng. 23.284 (Oct. 1951) 390.
See also References to Graphical Harmonic Analysis Chapter 6 Sect. 9(C).

(B) GENERAL REFERENCES TO CLASS A AMPLIFIERS

B1. Espley, D. C., and L. I. Farren, " Direct reading harmonic scales," W.E. (April 1934) 183.
B2. R.C.A. " Application Note on design of audio systems employing type 2A3 power amplifier triodes " No. 29 (Dec. 29, 1933).
B3. R.C.A. " Application Note on short-cut method for determining operating conditions of power output triodes " No. 42 (Sept. 5, 1934).
B4. R.C.A. " Application Note on use of the plate family in vacuum tube power output calculations " No. 78 (July 28, 1937)—Reprinted Radiotronics 80 (Oct. 11, 1937) 72.
B5. Pullen, K. A. " The use of G curves in the analysis of electron-tube circuits " Proc. I.R.E. 37.2 (Feb. 1949) 210.
B6. Langford-Smith, F., and W. J. Steuart " Effect of power supply on power amplifiers "—unpublished analysis.
B7. Pullen, K. A. " G curves in tube circuit design " Tele-Tech 8.7 (July 1949) 34 ; 8.8 (Aug. 1949) 33.
B8. Pullen, K. A. " Use of conductance, or G, curves for pentode circuit design " Tele-Tech 9.11 (Nov. 1950) 38.

(C) REFERENCES TO MULTI-GRID AMPLIFIERS

C1. Brian, W. S. " Experimental audio output tetrode " Elect. 20.8 (Aug. 1947) 121.
C2. Pickering, N. " Space-charge tetrode amplifiers " Elect. 21.3 (Mar. 1948) 96.
C3. Schade, O. H. " Beam power tubes " Proc. I.R.E. 26.2 (Feb. 1938) 137.
C4. Jonker, J. L. H. " Pentode and tetrode output valves " Philips Tec. Com. No. 75 (July 1940) 1.
C5. " Advantages of space-charge-grid output tubes " (Review of paper by N. C. Pickering) Audio Eng. 31.9 (Oct. 1947) 20.
C6. Hadfield, B. M. " Amplifier load impedance reduction," Tele-Tech 7.5 (May 1948) 33. See also B5.
C7. Hafler, D., & H. I. Keroes "An ultra-linear amplifier" Audio Eng. 35.11 (Nov. 1951) 15.

(D) REFERENCES TO PARALLEL AMPLIFIERS

D1. Jones, F. C. " Parallel tube high fidelity amplifiers " Radio 29.10 (Oct. 1945) 27.
D2. Jones, F. C. " Additional notes on the parallel tube amplifier " Radio (June 1946) 26.

(E) REFERENCES TO PUSH-PULL AMPLIFIERS

E1. Thompson, B. J. " Graphical determination of performance of push-pull audio amplifiers " Proc. I.R.E. 21.4 (April 1933) 591.
E2. Barton, L. E. " High audio output from relatively small tubes " Proc. I.R.E. 19.7 (July 1931) 1131.
E3. M.I.T. Staff (book) " Applied Electronics " (John Wiley and Sons, New York ; Chapman and Hall, London, 1943) pp. 433-448, 548-558.
E4. Sturley, K. R. " Radio Receiver Design " Part 2 (Chapman and Hall, London, 1945) pp. 84-97.
E5. " Push-pull triode amplifiers " Radiotronics No. 79 (Sept. 1937) 64 ; No. 80 (October 1937) 78.
E6. Adorjan, P. " Power amplifier design," Radio Eng. 16.6 (June 1936) 12.
E7. Krauss, H. L. " Class A push-pull amplifier theory " Proc. I.R.E. 36.1 (Jan. 1948) 50.
E8. Cocking, W. T. " Push-pull balance " W.W. 53.11 (Nov. 1947) 408.
E9. R.C.A. Application Note No. 54 " Class AB operation of type 6F6 tubes connected as pentodes," (Dec. 1935).
E10. Williams, W. N. " Novel 10 watt amplifier for P.A." Radio and Hobbies, Australia (April 1942) 37.
E11. Heacock, D. P. " Power measurement of Class B audio amplifier tubes " R.C.A. Rev. 8.1 (Mar. 1947) 147.
E12. Barton, L. E. " Application of the Class B audio amplifier to a-c operated receivers " Proc. I.R.E. 20.7 (July 1932) 1085.
E13. Sah, A. Pen-Tung " Quasi transients in Class B audio-frequency push-pull amplifiers," Proc. I.R.E. 24.11 (Nov. 1936) 1522.
E14. Nelson, J. R. " Class B amplifiers considered from the conventional Class A standpoint " Proc. I.R.E. 21.6 (June 1933) 858.
E15. Butler, F. " Class B audio-frequency amplifiers " W.E. 24.280 (Jan. 1947) 14.
E16. R.C.A. Application Notes Nos. 5, 14, 18, 33 and 54 giving operating data on types 79, 53, 19, 800, 54 respectively.
E17. Barton, L. E. " Recent developments of the Class B audio- and radio-frequency amplifiers " Proc. I.R.E. 24.7 (July 1936) 985.
E18. McLean, T. " An analysis of distortion in Class B audio amplifiers " Proc. I.R.E. 24.3 (March 1936) 487.
E19. Strafford, F. R. W. " Join-up distortion in Class B amplifiers," W.E. 12.145 (Oct. 1935) 539.
E20. R.C.A. Application Note No. 40 " High power output from type 45 tubes " (June 1934).
E21. Reich, H. J. " Theory and applications of electron tubes " (McGraw-Hill, 2nd edit. 1944).
E22. Greenwood, H. M. " Cathode-follower circuits " Q.S.T. 29.6 (June 1945) 11, especially Fig. 20.
E23. Amos, S. W. " Push-pull circuit analysis—cathode-coupled output stage " W.E. 23.269 (Feb. 1946) 43.
E24. Edinger, A. " High-quality audio amplifier with automatic bias control " Audio Eng. 31.5 (June 1947) 7. Based on patent by Lincoln Walsh.
E25. Minter, J. " Audio distortion and its causes " Chapter 6. " Standard FM Handbook " (FM Company, Great Barrington, Mass., U.S.A. 1st edit. 1946)—gives intermodulation test results on Lincoln Walsh amplifier.
E26. Patent application, inventors C. G. Smith, W. Storm, E. Watkinson, assignee Philips Electrical Industries of Aust. Pty. Ltd.
E27. Sturley, K. R. " Push-pull A.F. amplifiers—load curves for Classes A, B and C conditions " W.E. 26.313 (Oct. 1949) 338.
E28. McIntosh, F. H., and G. J. Gow " Description and analysis of a new 50-watt amplifier circuit " Audio Eng. 33.14 (Dec. 1949) 9.
E29. Parry, C. A. " The cathamplifier " Proc. I.R.E. Aust. 11.8 (Aug. 1950) 199.
Parry, C. A. " The cathamplifier " Proc. I.R.E. 40.4 (April 1952) 460.
E30. " The Cathamplifier Handbook " (Mingay Publishing Co., Sydney, Australia).
E31. Sterling, H. T. " Extended Class A audio " Elect. 24.5 (May 1951) 101.
See also A13, A18, A20, B5, C3.
E32. Peterson, A., and D. B. Sinclair " A single-ended push-pull audio amplifier " Proc. I.R.E. 40.1 (Jan. 1952) 7.

(F) REFERENCES TO COMPLETE AMPLIFIERS

F1. Williamson, D. T. N. " Design for a high-quality amplifier " (1) Basic requirements : alternative specifications, W.W. 53.4 (April 1947) 118 ; (2) Details of final circuit and its performance, W.W. 53.5 (May 1947) 161.
F2. Baxandall, P. J. " High-quality amplifier design—advantages of tetrodes in the output stage " W.W. 54.1 (Jan. 1948) 2 ; Correction 54.2 (Feb. 1948) 71.
F3. Gibson, W. F., and R. Pavlat " A practical cathode-follower audio amplifier " Audio Eng. 33.5 (May 1949) 9.
F4. Williamson, D. T. N. " High-quality amplifier—new version " W.W. 55.8 (Aug. 1949) 282 ; 55.10 (Oct. 1949) 365 ; 55.11 (Nov. 1949) 423.
F5. " The Williamson Amplifier," Reprint of articles published in the Wireless World (Iliffe and Sons Ltd., London).

(G) GENERAL REFERENCES

G1. Sulzer, P. G. "A survey of audio-frequency power-amplifier circuits" Audio Eng. 35.5 (May 1951) 15.

Additional references H will be found in the Supplement commencing on page 1475.