# The Technique of Sound Reproduction

*Theory and Practice*

# AMPLIFIERS

# The Technique of Sound Reproduction

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# AMPLIFIERS

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# CONTENTS

# EDITOR'S INTRODUCTION

*High Fidelity—by which we mean the recording and reproduction of sound with maximum faithfulness to the original—has been in existence for only a decade or two. Yet in that time it has become an absorbing hobby for thousands of technical and musical enthusiasts. In countless homes it provides the key to unlimited musical enjoyment in the form of today's high quality gramophone records, tape records and the VHF/FM broadcasts.*

*During the same period, high fidelity has risen in importance as a subject of study at Technical Colleges where it is variously listed under Audio, Electronic, Broadcast and Telecommunication Engineering.*

*The decision to produce this series of books on the Technique of Sound Reproduction in such a way as to make them add up to a comprehensive Manual of High Fidelity in six volumes arose from the following considerations:*

*1. To treat this wide subject adequately in a single book and at a level suited to the needs of both the technical student and the keen amateur would require a work of unmanageable proportions.*

*2. The practice of assembling high fidelity equipment in separate components conveniently allows the student and the amateur to study or work separately at these six aspects—Acoustics, Amplifiers, Loudspeakers, Disc, Tape and Radio.*

*3. In this age of specialization, we were presented with the seeming paradox that six expert specialist authors were easier to find and brief than a single polyhistor of high fidelity techniques.*

*Accordingly, and before one word was written, the Editor was able to hold a series of meetings with the authors, each of whom is an expert in his particular field. This procedure has ensured that the technical level is uniform throughout the series, and that the volumes dovetail together to provide complete coverage while nevertheless taking their place in the literature as individual works in their own right.*

*Special attention has been given to terminology. Each book includes a Glossary of Terms so that anyone possessing the complete series has access to a sectionalized dictionary and reference to the whole subject of the Technique of Sound Recording and Reproduction.*

*JOHN BORWICK*

# PREFACE

IN this book I have attempted to describe, in simple terms, the theory and practice of the audio frequency amplifier of the type comprised under the popular description 'high fidelity amplifier'.

The aim has been to deal with the subject from simple theory up to the development of the circuits most often encountered in an understandable, concise and practical way. The work is not intended to be comprehensive nor a text-book and is directed to the user of this type of equipment who wishes to have some understanding of its construction, the enthusiast and the amateur designer.

At the date of publication, the vast majority of audio amplifiers use the thermionic valve although a few transistorised amplifiers are making their appearance. At present the main advantage accruing from the use of transistors in this field is that amplifiers can be made smaller and lighter, not a matter of great significance in fixed sound reproducing equipment. One considerable advantage of a transistor output stage, however, is that it can be designed so that a loudspeaker can be connected without the necessity of using a matching transformer. Further development in the manufacture of transistors and the reduction in their cost during the next few years, will doubtless make available low priced audio amplifiers with a standard of performance superior to that of the valve amplifier. At that time, the chapter on transistors may require expansion in the form of a second volume.

Valve application reports for some of the most popular valve types have been gathered together in Appendix II and I believe that these will be found to be a useful scource of information on amplifier performance and design. All illustrations are based on current valve types and data sheets from the appropriate application reports have been used for this purpose by courtesy of Mullard Ltd.

The author wishes to thank Mullard Ltd. and The General Electric Co. Ltd. of England for permission to reproduce application reports.

H. LEWIS YORK

# INTRODUCTION TO ELECTRONICS

THE subject of high fidelity amplifiers, or more correctly audio frequency amplifiers, is covered by a branch of electronics generally known as audio frequency engineering. To understand this subject it is therefore necessary to have some knowledge of the fundamentals of electronics. The outline given in this chapter is of necessity over-simplified, but it is hoped that it will nevertheless be sufficient to render what follows understandable.

### Electrons

Electronics is concerned with the controlled motion of electrons. All substances may be considered to be composed of very small particles called atoms, of which electrons are a component part. The atom has a central nucleus around which is gathered a number of smaller particles called electrons. The electrons are held in the proximity of the nucleus by a mutual attraction described as an electrical charge; the nucleus is designated positive + and the electrons negative −. The number of electrons in an atom produce a total negative charge which equals the positive charge of the nucleus, so that the atom is normally neutral. Some of the electrons are in a continual state of migration from one atom to another, due to thermal energy, but on an average there is no net migration in any particular direction unless an externally applied electrical stress is present.

In the case of materials called *conductors* of electricity, there is a large number of electrons in migration. If at any particular point there is an excess of electrons, the balance of charge is negative; if there is a deficiency of electrons, the balance of charge is positive. The most commonly used conducting materials are copper, brass and silver. Atoms in which there is little migration of electrons

constitute substances which are called *insulators* and common examples are glass, mica, rubber, paper and ceramic. The degree to which a material resists a stress upon its electrons to migrate, is related to its *specific resistance*; thus conductors have a low resistance and insulators a high resistance.

A battery is a source of electricity. It has two terminals at which, by electrochemical means, the electrical charge is unbalanced. Therefore at the positive terminal the atoms have a deficiency of electrons, and at the negative terminal the atoms have an excess of electrons. If these two terminals are joined by a conductor, e.g. a length of copper wire, the atoms in the wire adjacent to the negative terminal commence to receive electrons, but readily pass on electrons to nearby atoms. Thus a current of electrons flows through the conductor to the positive terminal of the battery where the positive charge, or deficiency of electrons, is made good. This flow will continue until the unbalance in the battery is restored to neutrality, when the battery is discharged and the flow ceases.

The flow through the conductor depends upon the specific resistance of the material and its dimensions. The unit of resistance is the ohm. The pressure which the excess of electrons exerts upon the conductor is called the voltage of the battery (measured in volts) and the rate at which the electrons flow through the conductor, i.e. the number of electrons in a given period of time, is the current, the unit being the ampère. These three electrical units are related by Ohm's Law which states that the current ($I$) is equal to the voltage ($V$) divided by the resistance of the conductor ($R$). Given any two of these quantities, the third can always be calculated, viz.

$$I = V/R, V = IR \text{ and } R = V/I.$$

### Resistance

In electronics, electrons are required to flow and move about the various parts of the circuit via conductors, the resistance of which should be as low as conveniently possible. When the flow is required to be controlled, components with the required amount of resistance are introduced into the circuit and these components are called resistors. Any resistor is indicated in diagrams by the letter $R$ but when its ohmic value is specified numerically, the number is followed by the sign $\Omega$. Thus a resistor of 10 ohms is indicated by 10 $\Omega$. Resistors used in audio amplifiers normally have values between 1 ohm and 10,000,000 ohms. Resistors between 1,000 and 1,000,000 ohms are specified in multiples of 1,000, indicated by the letter $k$,

so that 1,000 ohms becomes 1 kΩ, 4,700 ohms becomes 4·7 kΩ and 680,000 ohms becomes 680 kΩ. Resistors of one million ohms and over are specified in multiples of one million, indicated by the letter M, so that 1,000,000 ohms becomes 1 MΩ and 10,000,000 ohms becomes 10 MΩ.

The nominal resistance value of a resistor is indicated by a system of coloured bands commencing from one end. The colour coding is as follows:

TABLE 1.1   *Colour Coding of Resistors*

| Colour code | | Number of noughts | |
|---|---|---|---|
| Black | 0 | Black | None |
| Brown | 1 | Brown | 0 |
| Red | 2 | Red | 00 |
| Orange | 3 | Orange | 000 |
| Yellow | 4 | Yellow | 0,000 |
| Green | 5 | Green | 00,000 |
| Blue | 6 | Blue | 000,000 |
| Violet | 7 | | |
| Grey | 8 | *Tolerance* | |
| White | 9 | | |
| | | None | 20% |
| | | Silver | 10% |
| | | Gold | 5% |

| 10% and 20% resistors | | | | | |
|---|---|---|---|---|---|
| Brown | Black | **10** | Orange | Orange | **33** |
| Brown | Red | **12** | Orange | White | **39** |
| Brown | Green | **15** | Yellow | Violet | **47** |
| Brown | Grey | **18** | Green | Blue | **56** |
| Red | Red | **22** | Blue | Grey | **68** |
| Red | Violet | **27** | Grey | Red | **82** |

The first two bands indicate the first and second significant figures and the third the decimal multiplier, i.e. the number of noughts to be added. The fourth band indicates the percentage tolerance about the nominal resistance value. The above table gives the first two significant figures for all 10% resistors, the figures in heavy type being also available in 20% tolerance.

Let us consider a circuit in which two resistors $R_1$ and $R_2$ are connected end to end and the free ends applied across a battery.

13

This connection of the resistors is called a *series* connection (Fig. 1.1). A current will flow, but it will be limited by the total resistance of $R_1$ and $R_2$. The voltage across the pair is the battery voltage, so that the voltage across each resistor must be less than the battery voltage,



FIGURE 1.1

Resistors in series. The current flowing $I = V/R$, where $R$ is the total resistance. The power in each resistor is $W = I^2R$.

but such that the voltage across $R_1$ plus the voltage across $R_2$ equals the battery voltage. The ratio of the voltages across the resistors is in fact equal to the ratio of the ohmic values of the resistors. The voltage of the battery is therefore divided into $R_1$ plus $R_2$ parts, and $R_1$ parts appear across $R_1$: i.e. $\dfrac{R_1}{R_1 + R_2}$ parts of $V$. Similarly the voltage

FIGURE 1.2

The resistors $R_1$ and $R_2$ form a potential divider so that $V_0 = V_I \times \dfrac{R_2}{R_1 + R_2}$
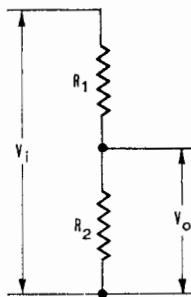


across $R_2$ is $\dfrac{R_2}{R_1 + R_2}$ $V$. (Fig. 1.2). This circuit is referred to as a potential divider and is repeatedly encountered in audio amplifier circuitry. The current flowing in the circuit may be calculated by the application of Ohm's Law $I = V/R$. The *power* in the circuit is

14

measured in *watts* and is equal to *VI*. The power may also be calculated by the alternatives $W = I^2 R$ *or* $W = \dfrac{V^2}{R}$ (from Ohm's Law).

If the two resistors $R_1$ and $R_2$ are both connected across the battery, the connection is called a *parallel* connection (Fig. 1.3). In this case the voltage across each resistor is the voltage of the battery and the current flowing in each resistor is $\dfrac{V}{R}$. The total current in the circuit is the current through $R_1$ plus the current through $R_2$. The total effective resistance of resistors in parallel is

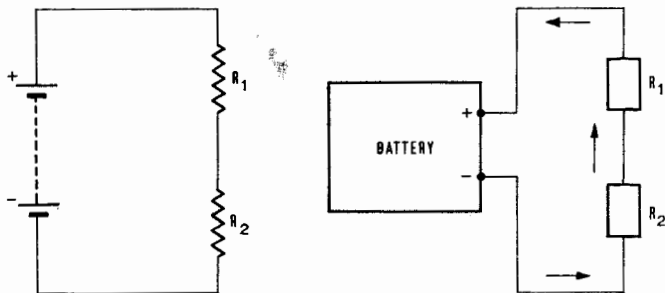$$R = \frac{1}{\dfrac{1}{R_1} + \dfrac{1}{R_2} + \dfrac{1}{R_3}} \text{ etc.}$$
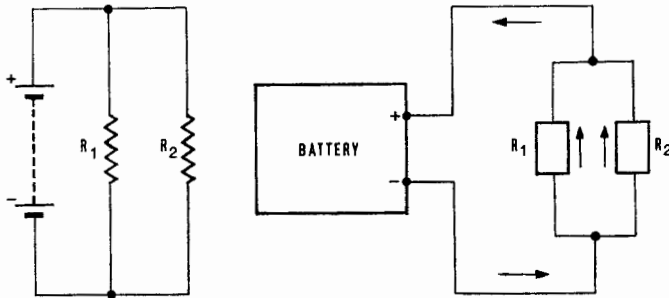


FIGURE 1.3

Resistors in parallel. The current flowing $I = V/R$, where $R$ is the total resistance. The power in each resistor is $W = V^2/R$.

When only two resistors are concerned, this becomes

$$\frac{R_1 \times R_2}{R_1 + R_2}$$

The power in each resistor is again $VI$ or $I^2R$ or $\dfrac{V^2}{R}$.

A resistor can be constructed in a special form so that a sliding connection can be made at any point along its length. When there are connections to only one end of the resistor and to the sliding connection, the component is called a variable resistor; when there are connections to each end of the resistor and to the sliding connection, the component is called a potentiometer (Fig. 1.4). If the ends of a potentiometer are connected across a battery, the voltage at the slider relative to one end can be varied from the full voltage of the battery
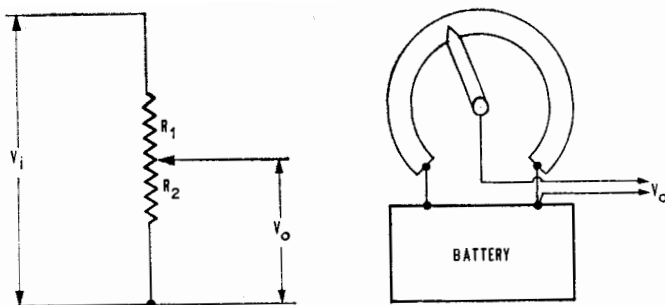
to zero. A potentiometer used in this way in an audio amplifier is frequently referred to as a volume control.

## Alternating Current

It has been stated that the voltage at the terminals of a battery is due to electrochemical action, and it is normally assumed that this voltage is steady and unvarying with respect to time. Hence the current which flows when a resistor is connected across a battery is called *direct current* (d.c.). Another method of producing electricity is by a generating machine known as an alternator, and in this case the voltage is not constant but is continually changing from zero to a peak maximum and back to zero. Alternate maxima are reversed, so that an excess of electrons is first at one terminal of the supply and then at the other. A complete sequence of change is called a *cycle* and the number of cycles which occur in each second is referred to as
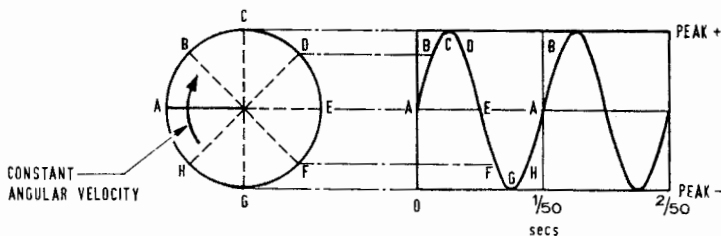


FIGURE 1.5

50 c/s sine wave. Voltage plotted against time.

the *frequency, f* in cycles per second (c/s) or thousands of cycles per second (kc/s) (Fig. 1.5). The change in voltage generated by a machine of this kind follows the *sine law*, the voltage change being

16

rapid around zero and slow around peak. In other words, the voltage output varies with time in the same way as the speed of movement of a pendulum bob. The rate of change of voltage as it passes through zero is determined by the peak amplitude and the frequency of the sine wave. The rate of change decreases until at the peak value of the waveform it is zero.

If the peak value of the sine wave alternating voltage is divided by $\sqrt{2}\,(= 1 \cdot 414)$ the *root mean square* or r.m.s. value is obtained, so that peak voltage equals r.m.s. voltage times $\sqrt{2}$. It is sometimes convenient to remember that dividing by $1 \cdot 414$ is the same thing as multiplying by $0 \cdot 707$, so that r.m.s. equals $0 \cdot 707 \times$ peak. The peak and r.m.s. values of an alternating current (a.c.) are similarly related (Fig. 1.6). The electrical power in an a.c. circuit is again measured
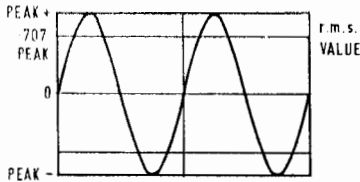


FIGURE 1.6

The r.m.s. value of a sine wave is peak $\times$ $0 \cdot 707$. Peak value is r.m.s. $\times$ $1 \cdot 414$.

in watts and equals $I$ (r.m.s.) $\times$ $V$ (r.m.s.), assuming that the circuit is composed of pure resistance only.

## Capacitance

If two metal plates are brought close together and are connected to the terminals of a battery, part of the positive and negative charges at the battery terminals will be transferred to the plates. If the plates are separated by an insulator, this will not conduct electrons but will strongly resist the strain placed upon it. When the connection is first made, there will be a sudden flow of electrons forming a high current. But as the plates charge, the current will fall away until the plates have accepted all the electrons they can and the current flow will be zero. If the *capacitor* so formed is disconnected from the battery, it will remain charged until some leakage from one plate to the other restores neutrality (Fig. 1.7).

The capacitance of this arrangement increases as the size of the plates is increased and as the plates are brought closer together, and is also dependent upon the nature of the insulator between them. The capacitance of this component is measured in *Farads* (F). In audio amplifiers, capacitors are usually found of values from one millionth of a millionth of a Farad, i.e. one micro-microfarad (or

B A

one picofarad written $\mu\mu F$ or pF) to one hundred millionth of a Farad, i.e. one hundred microfarads (written 100 $\mu F$). If the pressure exerted by the battery is increased, a point will be reached where the
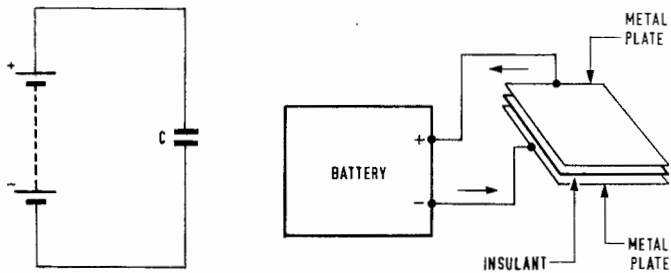


FIGURE 1.7

When a capacitor is connected to a battery, a charging current flows until the capacitor is charged.

insulator can no longer withstand the strain placed upon it and will break down, with a heavy flow of electrons destroying the capacitor. Capacitors are therefore marked with the maximum working voltage at which they are intended to operate.

It has been mentioned that when a capacitor is first connected to a battery, a high current flows and when the capacitor is fully charged, the current drops to zero. If a capacitor is connected to an a.c. supply, when the rate of change of the voltage is high, the current is also high. At peak voltage the rate of change is instantaneously zero,
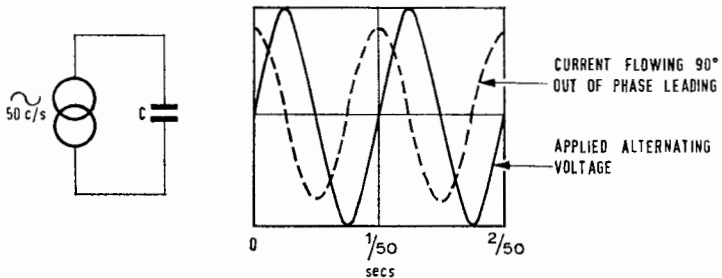


FIGURE 1.8

When a capacitor is connected to an A.C. supply, the current flow is 90° out of phase with the voltage and is said to lead the voltage.

so that the current is then also zero. The current in the circuit therefore depends upon the size of the capacitor, the r.m.s. value of the applied voltage and upon the rate of change of the alternating voltage,

18

i.e. the frequency. If the frequency is low, the current is also relatively low (Fig. 1.8).

A cycle is considered to be made up of 360°. A sine voltage increases from zero to peak in a quarter of a cycle or 90°. In the case of a capacitor connected to such a supply, the current is at minimum when the voltage is at maximum; so that the current and the voltage are 90° 'out of phase'. It follows that the capacitor controls the flow of current in this circuit and thus behaves like a resistor the value of which is dependent upon frequency. To distinguish this apparent resistance from that of a resistor, it is referred to as *reactance*. This is measured in ohms, but indicated by $X$ instead of $R$. The reactance of a capacitor is thus related to its capacitance $C$,
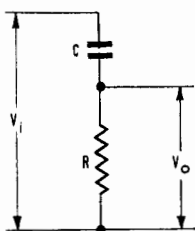


FIGURE 1.9
The capacitor $C$ and the resistor $R$ form a potential divider when $Vi$ is alternating. $V_0$ is dependent upon frequency and is

$$V_i \times \frac{R}{\sqrt{X_c^2 + R^2}}$$

in Farads, and to the frequency of the applied voltage. It can be calculated from the formula

$$X_c = \frac{1}{2\pi f C}$$

and since $2\pi \simeq 6 \cdot 28$, $X_c \simeq \dfrac{1}{6 \cdot 28 f C}$.

When capacitors are connected in parallel, the total capacitance is $C1 + C2 + C3$, etc. When capacitors are connected in series the total capacitance is

$$\frac{1}{\frac{1}{C_1}} + \frac{1}{\frac{1}{C_2}} + \frac{1}{\frac{1}{C_3}}, \text{ etc.}$$

If only two series capacitors are concerned, the total capacitance can be found as

$$\frac{C_1 \times C_2}{C_1 + C_2}.$$

Consider the case of a capacitor and resistor connected in series across an a.c. supply. This forms a potential divider similar to that previously considered, but with one resistor replaced by a capacitor (Fig. 1.9). The voltage across either the resistor or the capacitor will

19

now depend upon the frequency of the applied voltage; it is proportional to the ratio which the resistance or the reactance bears to the sum of the resistance and reactance of the pair. Due to the 90° phase difference, the sum of the resistance and reactance cannot be determined by simple addition, as in the case of the two resistors. So the resistance and reactance must be added in the form $\sqrt{X_c^2 + R^2}$. The proportion of the applied voltage appearing across the capacitor in this type of circuit is therefore

$$\frac{X_c}{\sqrt{X_c^2 + R^2}}$$

and the voltage across the resistor is

$$\frac{R}{\sqrt{X_c^2 + R^2}} V_i.$$

As the reactance of the capacitor will be low at high frequencies, the voltage across the resistor in this type of divider circuit will be greatest when the frequency is high and least when the frequency is low. Alternatively, to maintain the voltage across the resistor as high as possible at a given frequency, the reactance of the capacitor must be as low as possible, i.e. the capacitor must be large.
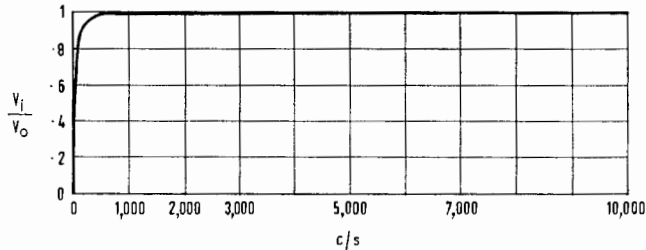


FIGURE 1.10

The voltage across the resistor $R$ in Fig. 1.9 as a proportion of $V_i$, over the frequency range 20 to 10,000 c/s, when $C = 0\cdot01$ μF, $R = 220$ k.

The way in which the voltage across the resistor in Fig. 1.9 varies with frequency may be found by calculating the value of $\sqrt{X_c^2 + R^2}$ over the relevant frequency range. Take as example, $C = 0\cdot01$ μF, $R = 220$ kΩ and $f = 20$ c/s to 10 kc/s. $X_c$ is found for a number of frequencies over the range, and $\sqrt{X_c^2 + R^2}$ calculated. The voltage appearing across $R$ is then

$$\frac{R}{\sqrt{X_c^2 + R^2}} V_i. \text{ (See Table 1.2.)}$$

20

These results may be plotted in graphical form as in Fig. 1.10 in which the frequency and voltage scales are linear. The frequency of
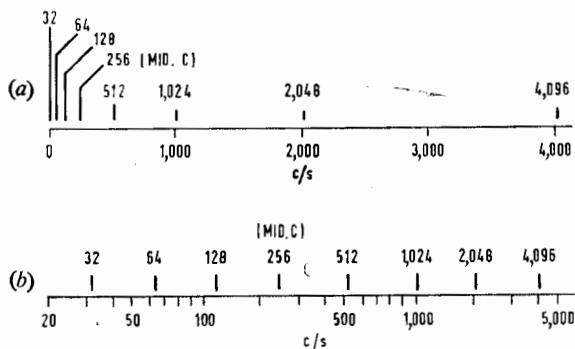


FIGURE 1.11

(a) Linear frequency scale showing position of note C in each octave. Middle C at 256 c/s.

(b) The same information with a logarithmic frequency scale. The spacing between octaves is now equal.

the musical note called middle C is 256 c/s (or 261 c/s at concert pitch). The frequency of the C one octave above middle C is 512 c/s and the C one octave, below 128 c/s. To conform to this relationship, the frequency scale must be drawn on a logarithmic basis (Fig. 1.11).

The intensity of sound is measured in *bels* and one-tenth of a bel, one decibel (dB), is about the smallest intensity change which the ear can detect. If the power is increased from $P_1$ to $P_2$, the change in

TABLE 1.2   *Proportion of input across voltage resistor*

| Selected f | $X_c$ | $\sqrt{X_c^2 + R^2}$ | $\dfrac{R}{\sqrt{X_c^2 + R^2}}$ | dB |
|---|---|---|---|---|
| 20 | 795,000 | 825,000 | 0·27 | −11 |
| 50 | 318,000 | 387,000 | 0·57 | − 5 |
| 100 | 159,000 | 271,000 | 0·81 | − 2 |
| 200 | 79,500 | 234,000 | 0·94 | −0·5 |
| 500 | 31,800 | 222,000 | 0·99 | −0·1 |
| 1,000 | 15,900 | 220,000 | 1 | 0 |
| 2,000 | 7,950 | 220,000 | 1 | 0 |
| 5,000 | 3,180 | 220,000 | 1 | 0 |
| 10,000 | 1,590 | 220,000 | 1 | 0 |
| 20,000 | 795 | 220,000 | 1 | 0 |

21
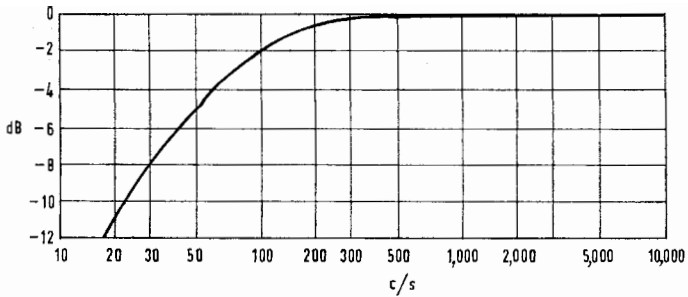
FIGURE 1.12

The information given in Fig. 1.10 when the frequency scale is logarithmic and the voltage ratio is expressed in decibels (dB).

dB is $10 \log \dfrac{P_2}{P_1}$. For the same value of resistance, power and voltage or current are related by Ohm's Law, so that when a voltage or current change occurs the change in dB is $20 \log \dfrac{V_2}{V_1}$, or $20 \log \dfrac{I_2}{I_1}$. When the power, voltage or current decrease, the change in decibels is in a negative direction (Fig. 1.12). (See Appendix 1.)

In amplifier circuits, this variation of voltage with frequency may be put to use in the form of a tone control. Where there must be no loss at a particular frequency, it is necessary to choose appropriate values for $R$ and $C$.

**Inductance**

Another common component is the *inductor*, which often consists of
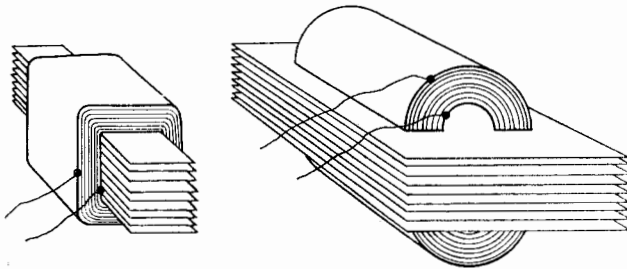


FIGURE 1.13

An inductor formed by winding a bundle of wire on special steel strips. The usual form is shown on the right in which the core has two windows to accommodate the wire, increasing the available inductance.

a length of wire wound upon a pile of strips or laminations of special steel (Fig. 1.13). When the ends of the wire are connected to the

22

terminals of a battery, electrons flow through the wire, but the effect of the turns on the strip steel core is to cause a magnetic field to be built up around the component. The flow of electrons in the wire is reduced whilst the magnetic field is being built up, but, once the field is fully formed, the flow is that which would be expected due to the resistance of the wire alone (Fig. 1.14). If the voltage of the battery is reduced, the decrease in the flow of electrons does not result as quickly as would be expected by virtue of the resistance of the wire alone, due to the collapse of the magnetic field. The *inductance* of this component is measured in *Henrys* (H) and increases with the number of turns of wire and according to the amount and nature of the core material. In audio amplifiers, inductors of one-thousandth of a Henry, i.e. one millihenry (1 mH) to about 50 Henrys (50 H) are found.

When inductors are connected in series, the total inductance is $L1 + L2 + L3$, etc. When inductors are connected in parallel, the total inductance is

$$\frac{1}{\dfrac{1}{L_1}} + \frac{1}{\dfrac{1}{L_2}} + \frac{1}{\dfrac{1}{L_3}} \quad \text{etc.}$$
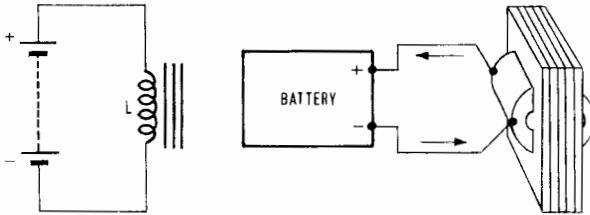


FIGURE 1.14

When an inductor is connected to a battery a current flows which is less than that due to the resistance of the wire alone until an external magnetic field is fully formed.

When only two inductors are concerned, the total inductance can be found as

$$\frac{L_1 \times L_2}{L_1 + L_2}.$$

The action of an inductor in an a.c. circuit is thus opposite to that of a capacitor. The inductor has a reactance similar to the capacitor but in the opposite sense, i.e. the reactance increases as the frequency rises and is given by the formula $X_L = 2\pi fL$ or $6 \cdot 28 fL$ (Fig. 1.15). The forms $X_c$ and $X_L$ are used to distinguish between capacitive and inductive reactance. Reactance combined with resistance is called *impedance*.

23
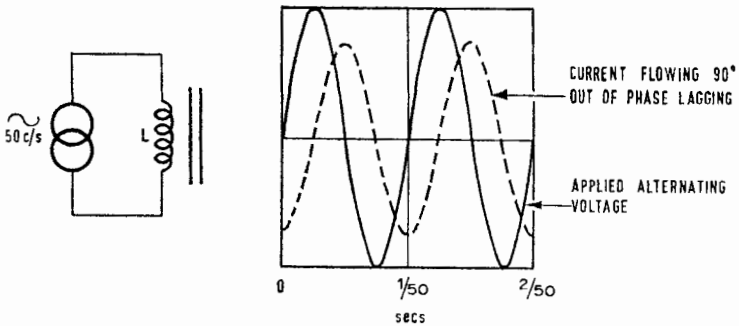
FIGURE 1.15
When an inductor is connected to an A.C. supply, the current flow
is 90° out of phase with the voltage and is said to lag the voltage.

If an inductor and resistor are connected in series across an alternating supply, the voltage across one element will be proportional to the ratio of the resistance or reactance to the sum of the resistance and reactance, as in the case of the capacitor. Thus the voltage across the resistor in such a circuit will be

$$\frac{R}{\sqrt{X_L{}^2 + R^2}} V_i$$

and will decrease as the frequency rises (Fig. 1.16).



FIGURE 1.16

The inductor $L$ and the resistor $R$ form a potential divider when $V_i$ is alternating. $V_0$ is dependant upon frequency and is

$$V_i \times \frac{R}{\sqrt{X_L^2 + R^2}}$$
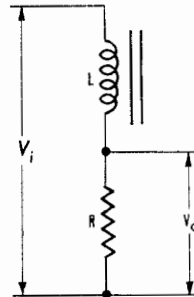
**The Transformer**

If the winding on the inductor is split into two sections and one of the sections, called the *primary*, is connected to an a.c. supply, a similar alternating current will be induced into the second or *secondary* winding. This is due to the magnetic flux linkage between the two windings of the component, which is now called a *transformer* (Fig. 1.17). The alternating voltage which appears at the connections to the secondary winding depends directly upon the ratio of the number of turns in the two windings, assuming a perfect transformer, i.e. one with no losses. The secondary voltage can

24

therefore be made greater or less than the primary voltage and several secondaries can be used with one primary.

Thus if the primary is connected to a 240 V a.c. supply, and the ratio of the primary to a secondary is 40 : 1, the voltage across this secondary will be 6 V. If the ratio of the primary to another secondary is 1 : 1·5, the voltage across this secondary will be 360 V.
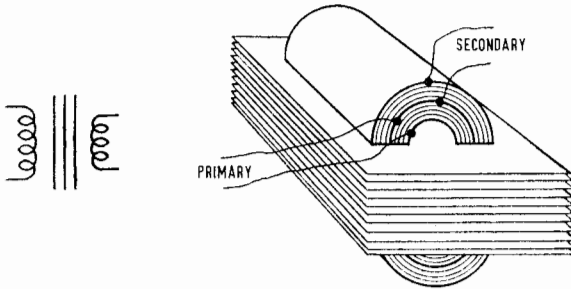


FIGURE 1.17
An inductor with two separate windings forming a transformer.
A number of secondaries can be associated with the primary.

The power in the primary circuit, $Wp$, must equal the total power in all the secondary circuits, $Ws$, if there is no loss in the transformer. Assuming one secondary winding, the secondary voltage, $Vs$, is $nVp$, where $n =$ turns ratio. Now from Ohm's Law $W = \dfrac{V^2}{R}$, so that $Wp = Ws$ may be written

$$\frac{Vp^2}{Rp} = \frac{(nVp)^2}{Rs} \quad \therefore \quad \frac{Rs}{Rp} = n^2 \text{ (Fig. 1.18)}.$$

The reflected impedance of the primary circuit in the secondary is therefore the primary impedance times the square of the turns ratios when there is a step up, or the primary impedance divided by the square of the turns ratio, when there is a step down.



REFLECTED
IMPEDANCE
1,500 Ω

15 Ω

10 : 1
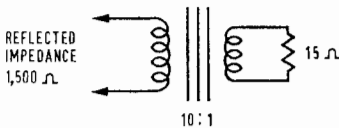
FIGURE 1.18
The reflected impedance in one transformer winding is proportional to the square of the turns ratio.

**Parallel Resonant Circuit**

If a capacitor and inductor are connected in parallel, the arrangement becomes a resonant circuit. The circuit is resonant at the frequency where the reactances are equal. That is, $X_c = X_L$, from which $f$, in

25

cycles per second, is found to be $\frac{1}{2\pi\sqrt{LC}}$ (Fig. 1.19). If this circuit is connected to a battery, the capacitor will charge and the inductor draw current. When the battery is disconnected, the capacitor will discharge through the inductor which will build up a magnetic field. When the capacitor has discharged sufficiently, the inductor field will collapse and electrons will flow back into the capacitor which will recharge. In the absence of losses, an alternating flow would continue indefinitely.

In any practical circuit, there will be losses. Thus a capacitor will have resistance due to the presence of the insulating material, although a high-quality capacitor will have a resistance measured in megohms. An inductor will have resistance due to the winding, and
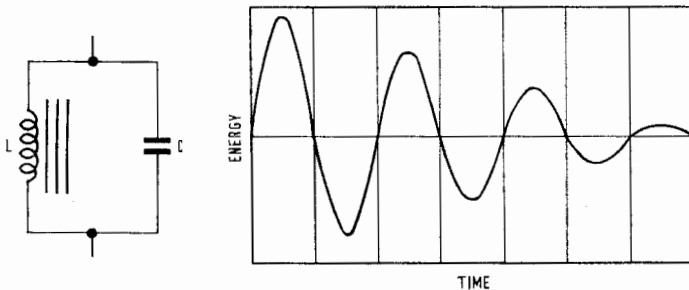


FIGURE 1.19

Inductor and capacitor in parallel forming a resonant circuit (left).
A damped sine wave diagram of the flow of current in a resonant circuit (right).

this will be of significant value even in a high quality component. Due to these losses, the alternating current in a reasonant circuit will not continue indefinitely, but will die away in much the same way that a balance wheel coupled to a hair spring will gradually become stationary when there is no energy in the main spring. The flow of current in a resonant circuit with losses, when represented graphically, is called a damped sine wave (Fig. 1.19).

The time taken for the flow to cease depends upon the resistance in the circuit, and the energy factor, $Q$, is

$$2\pi \times \frac{\text{peak energy storage}}{\text{energy dissipated per cycle}}.$$

In audio frequency amplifiers, resonance must be avoided for true reproduction, so that inductors are not normally included in signal circuits. In one case only is it difficult to avoid introducing

26

inductance, the special case of the output transformer (see page 55). The unwanted effects due to resonance can sometimes be reduced by adding resistance to the circuit, so that the time taken for the damped sine wave to reach zero is short. In other words, additional damping is introduced into the resonant circuit, or the $Q$ is reduced.

For certain purposes (e.g. a signal generator for testing) a continuously oscillating circuit is required, and this can be achieved by supplying energy to a resonant circuit equal to the losses. In the case of an output transformer in an audio amplifier, it is difficult to avoid supplying energy to a resonant circuit within the transformer, and in certain circumstances, continuous oscillation can occur, usually at a frequency above audibility. If an amplifier is in a state of continuous oscillation it cannot correctly perform the function for which it is designed.

### The Function of an Amplifier

Sound may be considered as a complex combination of sine waves in the frequency range 20 c/s to 20,000 c/s and these may be translated into electrical counterparts by means of a microphone. To recreate sound from electrical signals, the signals are made to operate a loudspeaker system. The output from a microphone is not sufficiently powerful to do this in direct connection with a loudspeaker, and the function of the audio amplifier in a high fidelity system is to accept small signals and present them in a form suitable for the operation of a loudspeaker. The signals from the microphone may have been conveyed over a distance by a cable or a radio transmitter and receiver system. They may have been stored by recording them on a gramophone record or magnetic tape, to be recreated as and when required by means of a gramophone pickup or magnetic tape head.
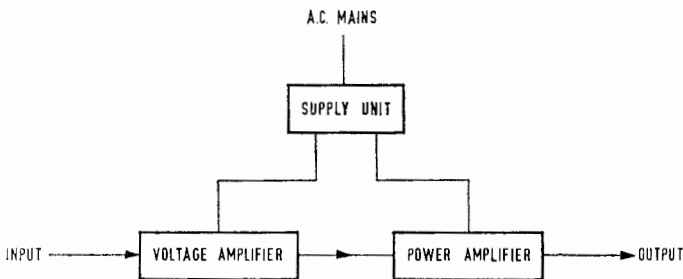


FIGURE 1.20

Block diagram of an audio frequency amplifier, showing the three main sections, namely the supply unit, the voltage amplifier and the power amplifier.

27

The amplifier is designed to accept the type of signal which a microphone, radio tuner, gramophone pickup or tape replay head will provide, and to generate the power which a loudspeaker system will require for its operation.

One means of generating the power required to operate a loudspeaker is to use a thermionic valve capable of producing the voltage and current changes in the loudspeaker which will enable it to develop the required volume of sound. This type of valve, used singly or in pairs, forms a power output stage in an amplifier and requires a relatively high controlling signal voltage. Between the small input signal and the output stage, therefore, voltage amplifying stages are used. Within these stages there are usually placed frequency correcting and tone controlling networks and a volume control. It is convenient and cheap to supply the working power, which the voltage amplifiers and output valves require for their operation, from the house a.c. mains supply, suitable supplies being derived via a supply unit incorporated in the amplifier.

The audio amplifier may therefore be considered to be composed of three sections:

1. A voltage amplifier section which will control and amplify the voltage of the input signals to a magnitude sufficient to operate the power output stage.

2. A power output stage which will operate the loudspeaker.

3. A supply unit which will convert the house a.c. mains to a form suitable for operation of the voltage and output amplifying stages.

These three sections are indicated in Fig. 1.20, and will be considered individually and then together in the form of a complete audio amplifier.

# VOLTAGE AMPLIFICATION

THE basic device by means of which voltage amplification is obtained is the thermionic valve. A valve is a device through which the current flow may be controlled by the application of a control voltage. If a valve is connected in series with a resistor across a battery or other source of voltage, the voltage drop across the resistor will depend upon the current in the circuit. If an alternating control voltage is now applied to the valve, the valve current will also alternate and hence the voltage drop across the resistor will vary in a manner dictated by the alternating control voltage. If the voltage change across the resistor is greater than the control voltage applied to the valve, voltage amplification has been achieved.

## The Diode Valve

In a thermionic valve, the current is due to a flow of electrons through the device. A supply of electrons is obtained by heating a special
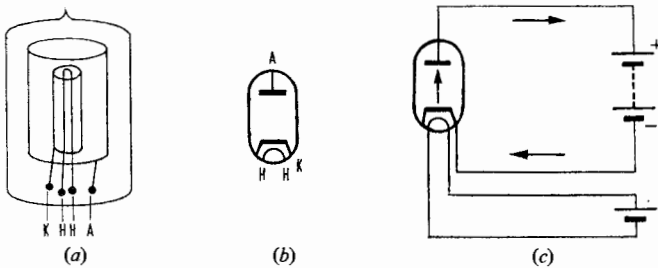


FIGURE 2.1

A diode valve showing (a) construction in evacuated glass tube, (b) the symbol used in circuit diagrams, (c) electron flow when the anode is made positive with respect to cathode.

substance in a vacuum. In practice, the vacuum is formed in a glass tube in the centre of which is placed a small electric heater wire which can be brought to the required temperature by the application

of an external supply voltage. Around this heater is placed a tube
called the cathode, to which the special material has been applied
(Fig. 2.1).

When the heater becomes hot, the cathode temperature is raised
due to radiation and electrons are released at the surface of the
cathode. In the absence of any attraction, the electrons form a cloud
around the cathode, referred to as the space charge. Another tube,
called the anode, is placed around the cathode, and if the anode is
made positive with respect to the cathode, it will attract electrons
from the space charge and current will flow through the valve. The
space charge will be maintained by further emission of electrons from
the cathode but, as the anode is made more positive, more and more
electrons are drawn off the space charge until a point is reached when
the anode is taking electrons as quickly as they can be emitted by the
cathode. A further increase in anode voltage will then produce no
further increase in anode current, and the valve is said to be saturated.
This valve with its two electrodes, cathode and anode, is called a
*diode*. It is not capable of providing amplification but is used in
audio amplifiers in the power supply unit. (see page 84).

**The Triode Valve**
If a spiral of fine wire is now placed between the cathode and anode,
this spiral (called the control grid) can be made to control the flow
of electrons from cathode to anode (Fig. 2.2). If the anode is main-
tained at a steady positive voltage and a battery is connected between
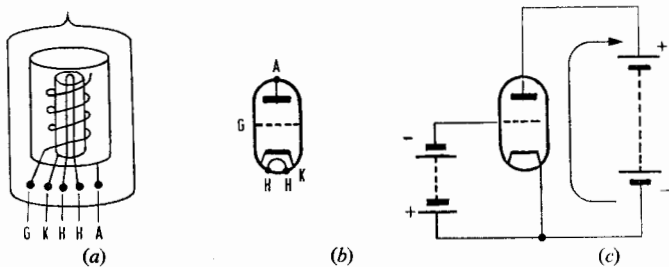


FIGURE 2.2

A triode valve showing (a) construction, (b) symbol, (c) electron flow
when the anode is made positive with respect to cathode. The flow
is controlled by making the grid negative with respect to cathode.

the control grid and the cathode so that the control grid is negative
with respect to cathode, the effect will be to reduce the attraction of
electrons from the space charge to the anode. This will reduce the
electron flow and hence the current in the anode circuit. As the

30

control grid is made more negative, the electron flow is progressively reduced until the point is reached where it is negligible. At this point, making the control grid more negative cannot significantly reduce the electron flow and the valve is then said to be 'cut off'. This valve with three electrodes—cathode, anode and control grid—is called a *triode*.

## Amplification

As the current through a triode can be controlled by the voltage applied to the control grid, amplification can be obtained by means of
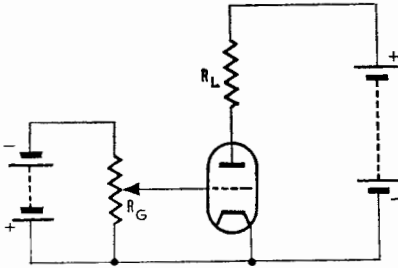
FIGURE 2.3

The addition of $R_L$ in the anode circuit, produces a circuit in which an alternating voltage at the grid produces an amplified alternating change of voltage at the anode.

the series circuit previously described. It will be observed that, when the voltage at the control grid is made more negative relative to the cathode, the current through the valve decreases. Therefore the voltage drop in the resistor is less, so that the anode becomes more positive. The output voltage changes are, therefore, in the opposite sense to those of the input voltage changes, and the anode is said to be 180° out of phase with the grid. If the grid is made positive with respect to the cathode, it will itself attract electrons which will flow in the grid circuit. This is undesirable, so the grid is not normally allowed to become less negative than about 1 volt and is maintained at a greater negative voltage than this by the application of a steady negative voltage called *grid bias*.

Consider the circuit of Fig. 2.3; $R_L$ is the load resistor and the negative grid voltage can be varied by means of the potentiometer $R_G$. Taking a typical case, $R_L$ might be 20 kΩ, the h.t. supply voltage 200 V and, when the grid is −3 V with respect to cathode, the current which flows might be 3 mA. The voltage drop across $R_L$ is $RI$ or 20,000 × 0·003, i.e. 60 V, so that the anode voltage is 200 − 60 = 140 V. If the grid voltage is changed to −2 V, there will be an increase in anode current depending upon the characteristics of the valve. If the current were to rise to 4 mA, the voltage drop across $R_L$

31
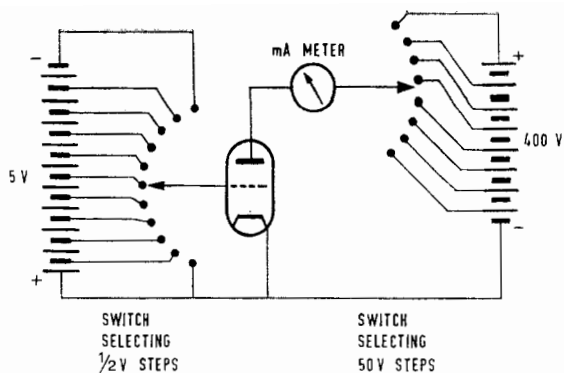
FIGURE 2.4

A circuit enabling the anode current to be measured for various negative voltages at the grid and positive voltages at the anode.

would increase to 80 V and the anode would then be at 120 V. Similarly, if the grid were made −4 V, and the anode current dropped to 2 mA, the anode would change to 160 V. A change of 1 V at the grid has therefore brought about a change of 20 V at the anode, and a voltage amplification (sometimes called a 'gain') of 20 has been achieved.

**Triode Characteristics**

If a valve is operated with, say, 200 V at its anode, and −3 V at its grid, and the grid is then made one volt more negative, i.e. −4 V, the current through the valve will drop. This drop in current can be compensated by increasing voltage at the anode. The additional voltage at the anode required to restore the current to the original value is called the *amplification factor* ($\mu$) of the valve. The internal resistance of the valve is called its *anode impedance* ($r_a$), and the change in current in the anode circuit with constant anode voltage brought about by a change of 1 volt at the grid is called the *mutual conductance* ($g$) in milliamps per volt (mA/V). These three characteristics are related so that $g = \dfrac{\mu}{r_a}$.

By means of a circuit such as that of Fig. 2.4, the current which flows when the grid is maintained at a given negative voltage may be measured for various anode voltages. From such measurements, a set of anode characteristic curves such as Fig. 2.5 may be constructed.

Applying Ohm's Law, the three valve characteristics may be obtained from Fig. 2.5. Thus in the region of anode volts = 200 V,

32

anode current $= 1$ mA and grid volts $= -1 \cdot 5$ V, the valve impedance, $V/I$, is found to be $\dfrac{235 - 163}{1 \cdot 8 - 0 \cdot 7}$ k$\Omega = 72/1 \cdot 1 = 65 \cdot 5$ k$\Omega$.

The amplification factor can be read straight off and is $235 - 139$ for 1 V grid change, i.e. 96. Similarly, the mutual conductance is $2 \cdot 05 - 0 \cdot 57$ mA for 1 V grid change, i.e. $1 \cdot 42$ mA/V.

The relationship $g = \mu/r_a$ can be used to test the accuracy of the figures so obtained. Thus 96/65,500 gives $1 \cdot 46$ mA/V which closely agrees with the figure for mutual conductance found as above. It should be noticed that the characteristics vary according to the position selected for the working point, and also that the figures obtained will not necessarily agree with those given in the valve application reports, which are usually for an anode voltage of 100 V and zero grid volts. The difference will not be great, and the report figures indicate the general nature of the valve, and enable valve types to be compared.
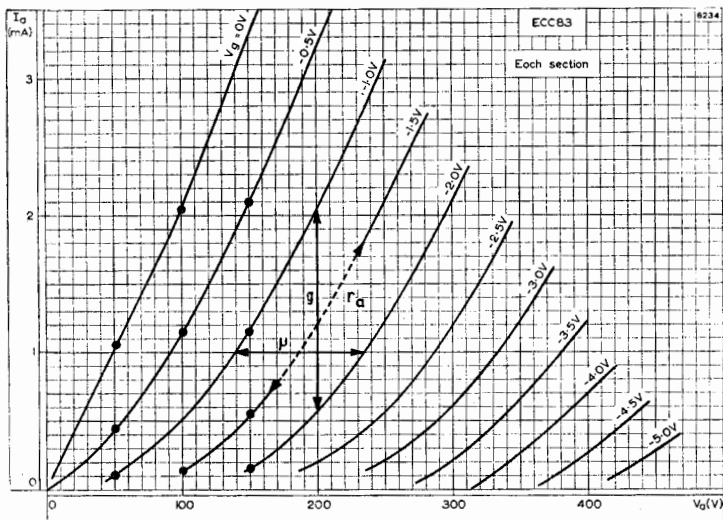


<p style="text-align:center">FIGURE 2.5</p>

The measurements taken by means of the circuit of Fig. 2.4 plotted to form a set of anode characteristic curves.

## Equivalent Circuit

The valve may be considered as a generator of $\mu$ times the input voltage $E$. As stated, the internal resistance of the valve $r_a$ and the load resistor $R_L$ form a potential divider, so that the output across

C A

the load is $E_0 = E_i\mu \dfrac{R_L}{r_a + R_L}$ and the stage gain will be:

$$\frac{E_0}{E_i} = \mu \frac{R_L}{r_a + R_L} \quad \text{(Fig. 2.6)}.$$

FIGURE 2.6
Equivalent circuit of the triode amplifier of Fig. 2.3 from which it can be shown that amplification $A = \dfrac{E_0}{E_i} = \dfrac{\mu R_L}{R_L + r_a}$.
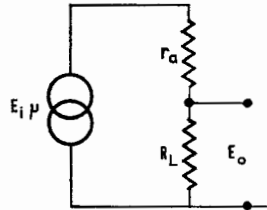


It will be apparent that the larger the value of $R_L$, the more nearly will the amplification of the stage approach the amplification factor $\mu$ of the valve itself. In practice, the value of $R_L$ is limited as there will be a standing voltage drop across it and, if this were made too great, the battery voltage would have to be increased to an impractically high figure. The value assigned to $R_L$ is also affected by other matters concerned with distortion and frequency response, which will be fully dealt with later.

**Automatic or Cathode Bias**
Whilst the grid bias voltage can be obtained by the connection of a battery between cathode and grid, as previously described, in practice it is convenient to avoid this by using cathode bias, sometimes called automatic bias. To make the grid more negative than the cathode, we can instead make the cathode more positive than the grid. The cathode can be made positive with respect to the grid by including a resistor in the connection between the cathode and the negative terminal of the h.t. supply, to which point the grid circuit is also returned. We then have three elements in the circuit, the anode load, the valve and the cathode resistor $R_k$. If an alternating voltage is applied to the grid circuit, part of the available output voltage will now appear across the cathode resistor, so that the output at the anode will be reduced by this amount. The resistance of the cathode resistor will normally be small compared with that of the anode resistor, so that the reduction in output at the anode will be small. In some special cases however, the cathode resistor is of a higher value than that required for correct biassing and then the reduction in the output at the anode becomes significant.

34

## Current Feedback

If the standing current increases for any reason, e.g. due to an increase in the h.t. supply, the current through the cathode resistor and hence the bias would rise and offset the current increase. The cathode resistor therefore gives 'automatic' biasing. In the circuit under consideration, the effective input voltage is that which appears between the grid and the cathode. If the grid is made more negative, the current through the valve will drop and hence the standing bias voltage at the cathode will drop also. This will once more have the effect of increasing the current flow, thus partially off-setting the effect of the applied negative voltage at the grid. This effect is referred to as *current feedback* and can be avoided by placing a large capacitor across the cathode resistor, which charges to the value of the cathode bias and maintains this steady irrespective of alternating changes in current. It will be apparent that, if the current changes are slow, i.e. the applied alternating voltage is a low frequency, the ability of this capacitor to maintain a steady bias is diminished. Consequently, as the frequency of the applied alternating voltage is reduced, the output voltage at the anode will also be reduced. It is therefore necessary to choose a value for the cathode bypass capacitor which will be effective down to the lowest frequency at which the stage is required to operate.

It has been mentioned that the variation in the current passing through the valve is limited by saturation and cut off. If the grid swings sufficiently negative and positive for these effects to occur, the output voltage will not be a true amplified copy of the input voltage. This effect is referred to as distortion. Even before the saturation and cut-off points are reached, distortion will occur due to the non-linear characteristic which all valves exhibit. Distortion from this cause can be reduced by limiting the grid swing to the relatively linear portion of the valve characteristic. In high fidelity amplifiers, each stage is required to introduce minimum distortion. In some cases this means that the input voltage must be severely restricted. The performance of a valve stage can be determined by a graphical method, using the anode characteristic of the valve. From this the amplification and distortion can be predicted with sufficient accuracy for most practical purposes.

# GRAPHICAL ANALYSIS

THE operation of a resistance coupled valve amplifier stage can be understood and ascertained by the use of the anode current/anode voltage characteristic, usually referred to as the *anode characteristic*. This method will be described with reference to a commercially available valve, the 12AX7 (ECC83) which is a double-triode valve. This valve has two similar valves built into one glass envelope for convenience. The graphical analysis is equally applicable to either triode section.

The simple valve stage is as shown in Fig. 3.1, where an anode load resistor is inserted between the anode of the valve and the h.t. positive supply line.

**Load Line**

Fig. 3.2 is the anode characteristic of the valve and upon this may be drawn a line which represents the selected anode load. For this



FIGURE 3.1

The load resistor $R_L$ enables the triode valve to amplify, i.e. $E_0$ exceeds $E_i$.
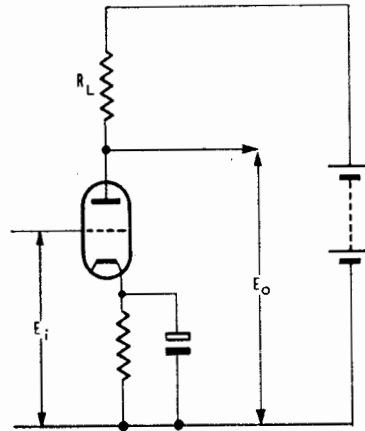
particular valve, inspection and experience show that in the ordinary case the load will not be lower than 47 kΩ or higher than 220 kΩ.

In the first instance, consider a load of 100 kΩ, and a supply voltage of 300 V. If the valve could be made zero resistance, the resistor would drop 300 V and draw 3 mA, giving point X on the O
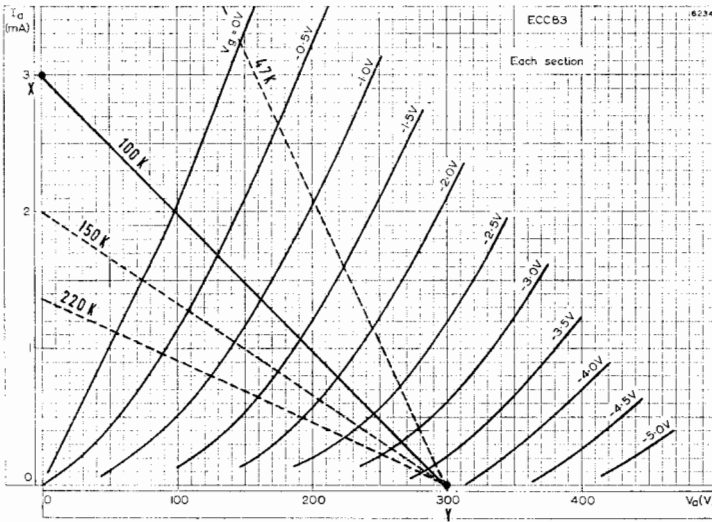
FIGURE 3.2

The load resistance $R_L$ can be represented by a line $XY$ drawn on the anode characteristic.

volt line. If the valve could be made infinite resistance, the current would be zero, giving point Y at 300 volts on the O current line. joining X to Y gives the 100kΩ load line. Fig. 3.2.

**Bias**

The working point of the valve can be found by inspection, and will be where the applied bias voltage line crosses the load line. The grid should not be allowed to become less negative than −1 V, and it can be seen that the valve cuts off when the grid voltage approaches −4 V. The peak to peak input swing cannot, therefore, exceed 3 V, or 1½ V peak. The best working point for this valve stage will therefore be at about −2 V and, at the point where the −2 V grid line crosses the load line, the anode voltage is seen to be 220 V and the standing current 0·8 mA.

When a sine wave input is applied to the grid, the grid voltage will change about the bias point, and points where the grid lines cross the load line show the state of affairs for each 0·5 V peak grid change.

37

The following table can be completed by reading off the anode voltage corresponding to 0·5 V changes in the grid voltage.

Table 3.1 shows that when the input swings 1 V positive, the anode swings −58 V negative. The stage gain is therefore 58, and the anode is 180° out of phase with the grid. It will be observed that,

TABLE 3.1   *Readings from Anode Characteristic*

| Grid voltage | Change in grid voltage | Anode voltage | Change in anode voltage |
|---|---|---|---|
| −1·0 V | +1·0 V | 162 V | −58 V |
| −1·5 V | +0·5 V | 191 V | −29 V |
| −2·0 V | 0   V | 220 V | 0 V |
| −2·5 V | −0·5 V | 246 V | +26 V |
| −3·0 V | −1·0 V | 268 V | +48 V |

when the grid swings 1 V negative, the anode swings 48 V positive. As the anode voltage changes for equal and opposite grid voltage changes are not equal, it is apparent that the valve is introducing distortion. This can be more clearly seen if the grid and anode voltages are plotted in the form of a *dynamic characteristic* (Fig. 3.3).
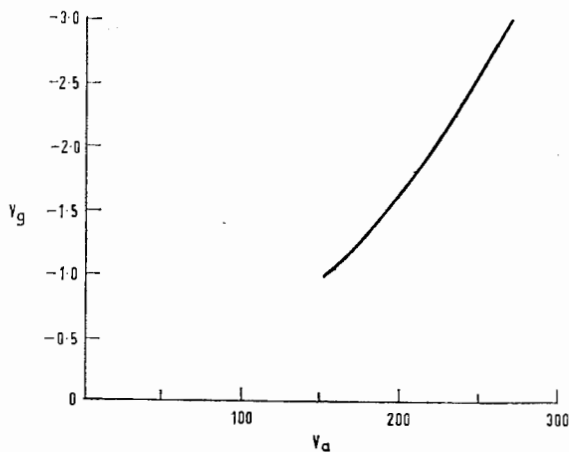


FIGURE 3.3

The anode voltages corresponding to grid voltages along the load line *XY* of Fig. 3.2. This dynamic characteristic exhibits curvature which indicates that the output voltage $E_0$ will be distorted.

## Dynamic Characteristic

The dynamic characteristic shows that, if the grid swing is small, the valve will work over the relatively straight portion of the dynamic characteristic. But, as the grid swing increases, more and more of the curved portion of the dynamic characteristic is covered and hence distortion increases. The distortion produced by a triode is mainly *second harmonic distortion*, and the percentage can be estimated by the expression $\dfrac{\text{difference in swings}}{2 \times \text{sum of swings}} \times 100$. In the example, when the input is 1 V peak, the distortion is $\dfrac{58 - 48}{2\,(58 + 48)} \times 100 = 4 \cdot 7\%$.
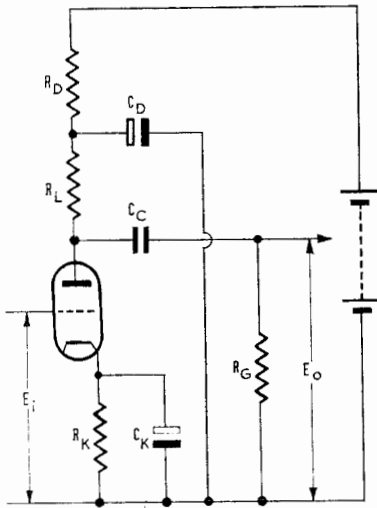


FIGURE 3.4
Complete circuit of practical voltage amplifier stage with decoupling $R_D C_D$, cathode biasing $R_k$ and $C_k$ and coupling to following stage $C_C R_G$.

The performance of the stage when the load resistance is 47, 150 and 220 kΩ can be investigated in the same way, using the appropriate load line, shown dotted on the anode characteristic.

A practical amplifier stage will usually include a cathode bias resistor $R_k$ and a de-coupling resistor $R_D$, and its output will be taken off via a capacitor and resistor network $C_C R_G$. These components will be fully discussed later. The complete circuit is shown in Fig. 3.4. With regard to $R_G$, it is sufficient for the moment to notice that as far as the signal is concerned, assuming $C_C$ to be of negligible reactance, $R_G$ is effectively in parallel with $R_L$, so the dynamic load in this valve stage is $\dfrac{R_L \times R_G}{R_L + R_G}$. Taking a typical case, $R_D$ might be 47 kΩ, $R_L$

39

100 kΩ, $R_G$ 470 kΩ and the h.t. supply line 300 V. The grid bias will depend on the value of $R_k$ and, if $R_k$ is 1 kΩ, the bias will be 1 V when the current is $\dfrac{1}{R_k} = \dfrac{1}{1000}$ A = 1 mA (from Ohm's Law). Similarly, the bias will be 1·5 V when the current is 1·5 mA, 2 V when the current is 2 mA, etc, so that a 1 kΩ bias line may be drawn through these points. Bias lines for 1·5 kΩ, 2·2kΩ and 3·3kΩ may be drawn similarly, as they are likely standard values (see Fig. 3.5). The d.c. load line must be drawn for all resistors in the circuit
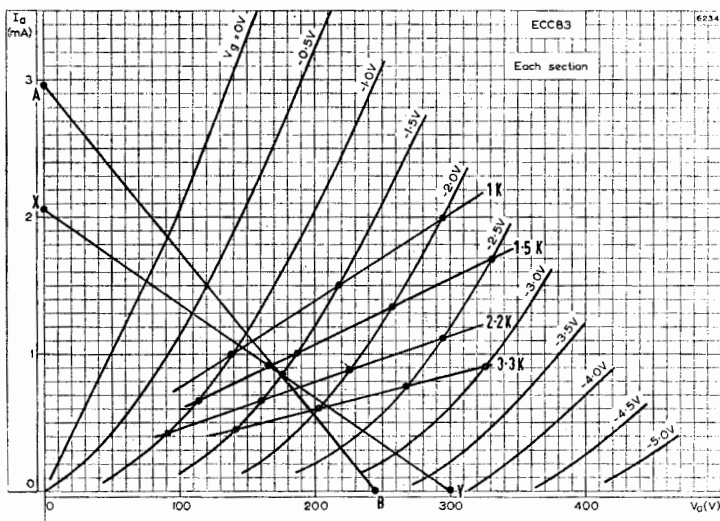
FIGURE 3.5

Characteristic with load line $XY$ and lines representing cathode bias resistors of 1 kΩ, 1·5 kΩ, 2·2 kΩ and 3·3 kΩ. The working point is where the load line crosses the selected bias resistor line. For a working point of − 1·5V grid bias, 175 V anode voltage and 0·84 mA anode current, the bias resistor is $\dfrac{1·5}{·84 \times 10^3} = 1·78$ kΩ. The grid resistor of a following stage reduces the dynamic load which is represented by AB drawn through the working point.

and will be of value equal to $R_D + R_L + R_k$. In the example, $R_D + R_L = 47$ kΩ + 100 kΩ = 147 kΩ and, compared with this, any likely value for $R_k$ is negligible. As before, point Y is at 300 V and point X is at $\dfrac{300}{147,000}$ A = 2·04 mA.

If the input voltage to the stage does not exceed 0·5 V peak,

40

inspection shows that a bias resistor of $2 \cdot 2 \, \text{k}\Omega$ would give a bias of about $1 \cdot 7 \, \text{V}$.

**Distortion**

The performance of the stage can conveniently be estimated by taking $-1 \cdot 5 \, \text{V}$ as the working point, so that the $-1 \, \text{V}$ and $-2 \, \text{V}$ grid lines can be used to read off anode voltage and current.

The dynamic load is $R_L$ in parallel with $R_G$ and equals $\dfrac{100 \times 470}{100 + 470}$ $= 83 \, \text{k}\Omega$. At $-1 \cdot 5 \, \text{V}$ the working anode voltage is $175 \, \text{V}$ and, to drop $175 \, \text{V}$ to zero when the load resistance is $83 \, \text{k}\Omega$, the current change must be $\dfrac{175}{83{,}000} \, \text{A} = 2 \cdot 1 \, \text{mA}$. The working current is $0 \cdot 84 \, \text{mA}$ and must be added, so that point A is at $2 \cdot 94 \, \text{mA}$. Joining point A and the working point gives the dynamic load line, A — B.

A table can be completed, as before, to enable the performance of the stage to be estimated.

TABLE 3.2  *Readings from Dynamic Load Line*

| Grid voltage | Change in grid voltage | Anode voltage | Change in anode voltage |
|---|---|---|---|
| $-1 \cdot 0 \, \text{V}$ | $+0 \cdot 5 \, \text{V}$ | $149 \, \text{V}$ | $-26 \, \text{V}$ |
| $-1 \cdot 5 \, \text{V}$ | $0 \quad \text{V}$ | $175 \, \text{V}$ | $0 \quad \text{V}$ |
| $-2 \cdot 0 \, \text{V}$ | $-0 \cdot 5 \, \text{V}$ | $199 \, \text{V}$ | $+24 \, \text{V}$ |

The anode voltage change for $0 \cdot 5 \, \text{V}$ peak grid change is $25 \, \text{V}$ peak average, so that the stage gain is approximately 50. The percentage second harmonic distortion is $\dfrac{26 - 24}{2(26 + 24)} \times 100 = 2 \cdot 0 \%$.

**Cathode bypass**

The cathode resistor has been ignored in this example except so far as it determines the working point of the stage. It has previously been mentioned that this resistor must be bypassed by a large capacitor, say $100 \, \mu\text{F}$, to avoid current feedback. The effect of omitting this capacitor, so that current feedback operates, can be demonstrated by calculating the voltage which will appear across the cathode

resistor. This voltage will equal $IR$, and the performance with current feedback may be found by constructing the following table:

TABLE 3.3  *Performance with Current Feedback*

| Grid voltage | (A) Grid voltage change | Current mA | Change in current | (B) Cathode voltage change with 2·2 kΩ resistor | (A + B) Effective input change | Anode voltage | Change in anode voltage |
|---|---|---|---|---|---|---|---|
| −1·0 | +0·5 | 1·15 | 0·31 | 0·68 | +1·18 | 149 | −26 |
| −1·5 | 0 | 0·84 | 0 | 0 | 0 | 175 | 0 |
| −2·0 | −0·5 | 0·55 | 0·29 | 0·64 | −1·14 | 199 | +24 |

It will be seen that the effect of current feedback is to reduce the stage gain since, without feedback, a negative output swing of 26 V is produced by a positive input swing of 0·5 V, whereas with feedback the input required to produce the same output becomes 1·18 V. The gain reduction factor is $1·18/0·5 = 2·36$, and the distortion will be reduced by approximately the same factor. Current feedback also increases the apparent output impedance of the stage, an effect which will be mentioned again when feedback is discussed in more detail.

It will be apparent that in some cases a small error in the estimation of voltage or current values will lead to an incorrect assessment of the performance of a stage and the graphical method should be used with caution in this respect. It does however give extremely useful information regarding the operating conditions of a stage.

# TETRODES AND PENTODES

Two types of thermionic valve have been considered, the diode, with two electrodes (cathode and anode), and the triode, with three electrodes (cathode, anode and grid). A number of further electrodes can be introduced but, in audio engineering, the other types of valve most usually employed are the tetrode, with four electrodes, and the pentode, with five electrodes.

### The Tetrode

The additional electrodes usually take the form of spiral wire grids, inserted between the control grid and anode. In the case of the tetrode, the fourth electrode is called the *screen grid* and is maintained at a positive potential relative to the cathode, but less positive than the anode. As the potential of the screen grid will have an effect on
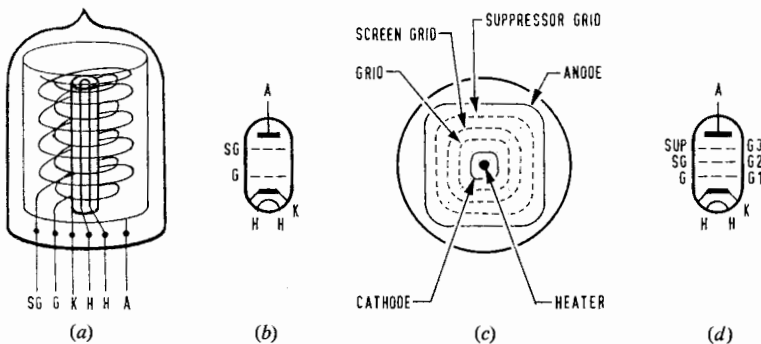


FIGURE 4.1

(a) Construction of a tetrode valve. (b) Symbol. (c) Construction of Pentode valve. (d) Symbol.

electron flow similar to that already described for the control grid, the potential of these two electrodes controls the current flow through the valve. The number of electrons passing through the control grid

43

is determined by the grid voltage and these pass on towards the screen grid. However, due to their velocity and the open construction of the screen grid wires, few electrons are collected at the screen grid, the bulk passing through to be collected by the anode.

### The Pentode

If the anode voltage is allowed to become lower than the screen grid voltage, secondary electrons emitted from the anode are collected by the screen grid. This causes the anode current to fall, producing a knee or kink in the anode current curve. This may be avoided by a fifth electrode, called the *suppressor grid*. This is placed between the screen grid and the anode and is maintained at cathode potential, thus repelling secondary electrodes which return to the anode, smoothing out the kink in the anode curve. The control grid, screen grid and suppressor grid are often referred to as grid 1, grid 2 and grid 3 respectively.

Tetrode valves are seldom used in voltage amplifier stages, so that only the pentode need be considered here. It will be observed from the examination of the anode characteristic of a typical pentode, the EF86, that the anode voltage has little effect upon the anode current over the working range (see Fig. 4.3). Thus, when the screen grid (g2) is maintained at 140 V with a grid bias of $-2 \cdot 5$ V, the anode current corresponding to an anode voltage of 100 is 2 mA. If the anode voltage is increased to the 300 V, the anode current rises to only $2 \cdot 08$ mA. This indicates that anode impedance of the valve is high,

i.e. $\dfrac{300 - 100}{2 \cdot 08 - 2} \times 1000 = 2 \cdot 5 \, \text{M}\Omega$.

### Pentode as Voltage Amplifier

If a pentode is used in a voltage amplifier stage, it is apparent that the valve impedance will have negligible effect upon the stage gain. Consider the formula $A = \dfrac{\mu R_L}{r_a + R_L}$. When $r_a$ is much higher than $R_L$, $R_L + r_a$ approximates closely to $r_a$ and the formula can be re-written $A = \dfrac{\mu R_L}{r_a}$. Now $\dfrac{\mu}{r_a} = g$, so that $A \simeq g R_L$. The mutual conductance, $g$, can be found from the anode characteristic as before. For the EF86, when $V_a = 100$ V, $V_{g2} = 140$ V and $V_{g1} = -2$ V, the current is $2 \cdot 87$ mA and when $V_{g1}$ is $-3$ V, the current is $1 \cdot 4$ mA; so that $g = 1 \cdot 47$ mA/V. If the load resistance is 100 k$\Omega$, the gain will be approximately 147 and, if the load resistor is 200 k$\Omega$, the gain becomes 294. It has been shown that the amplification to be

44

expected from a high gain triode such as the ECC83 is around 50, so that the amplification from a pentode stage can be several times that of a triode stage.

The practical circuit of a pentode voltage amplifier stage is shown in Fig. 4.2. The resistor $R_D$ and the capacitor $C_D$ decouple the stage from the h.t. supply, and the anode and cathode circuits are similar to the triode circuit of Fig. 3.4. The screen grid supply is obtained from the decoupled supply by means of a resistor $R_{SG}$. The capacitor $C_{SG}$ maintains the screen voltage at a steady value irrespective of
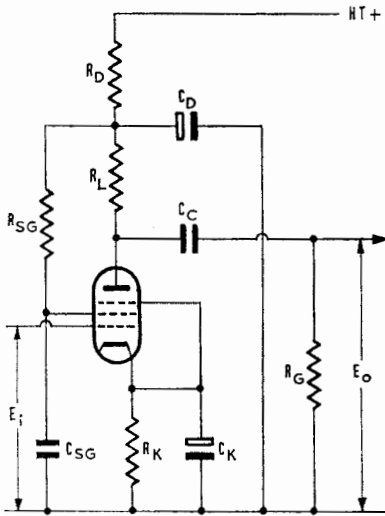


FIGURE 4.2
Complete circuit of pentode amplifier stage. The screen is supplied via resistor $R_{SG}$ and decoupled by $C_{SG}$.

current changes, when a signal is applied to the grid. The suppressor grid is connected to the cathode.

The graphical analysis of the performance of the pentode amplifier may be made as for the triode. Taking the EF86 as an example, a battery or h.t. supply of 300 V, after decoupling, is assumed. If $R_L$ is taken to be 100 kΩ and the screen operated at 140 V, a load line can be drawn upon the characteristic (see Fig. 4.3). A suitable operating point is seen to be at $V_A = 100$ V, when $V_{g1} = -2\cdot5$ V and $I_A = 2$ mA. Now the ratio of the anode to screen current in a pentode valve is relatively constant and, for the EF86, when the screen is at 140 V this ratio is 5 : 1. The screen current in the example will therefore be $0\cdot4$ mA and the value of the screen resistor will be that required to drop 300 V to 140 V at $0\cdot4$ mA, i.e. 400 kΩ. The anode and screen current derive from the cathode, so that the
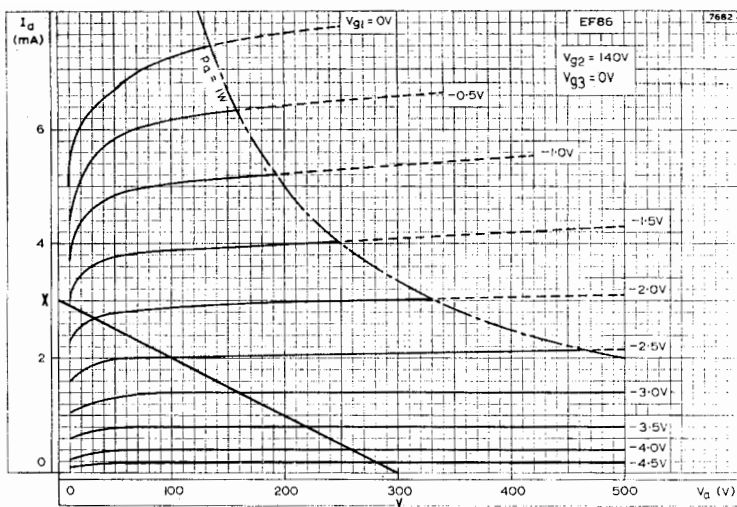
FIGURE 4.3

Anode characteristic of pentode valve when screen voltage is 140 V.
The anode load line XY, represents 100 kΩ.

cathode current will be $2\cdot4$ mA and the cathode resistor required
to give a bias of $2\cdot5$ V will be about 1 kΩ. Both the screen and cathode
resistors require adequate bypass capacitors.

### Dynamic Characteristic

The output coupling resistor $R_G$ is effectively in parallel with $R_L$
and, if $R_G$ is assumed to be 330 kΩ, the dynamic load is 77 kΩ
and a line can be drawn representing the dynamic load line. Using
this line, the following table can be constructed and the dynamic
characteristic of Fig. 4.4. drawn.

TABLE 4.1   *Readings from Dynamic Load Line*

| Grid voltage | Change in grid voltage | Anode voltage | Change in anode voltage |
|---|---|---|---|
| $-1\cdot5$ V | $+1\cdot0$ V | 11 V | $-$ 92 V |
| $-2\cdot0$ V | $+0\cdot5$ V | 43 V | $-$ 60 V |
| $-2\cdot5$ V | 0 V | 103 V | 0 V |
| $-3\cdot0$ V | $0\cdot5$ V | 153 V | $+$ 50 V |
| $-3\cdot5$ V | $-1\cdot0$ V | 201 V | $+$ 98 V |
| $-4\cdot0$ V | $-1\cdot5$ V | 236 V | $+133$ V |

46

**Distortion**

Examination of this dynamic characteristic shows that the curvature, and hence the distortion, is different from that of the triode shown in Fig. 3.3. The triode distortion is mainly second harmonic, and the calculation of distortion for the triode took account of this harmonic only. The distortion of the pentode includes the second and higher harmonics, i.e. third, fourth, fifth, etc. Fortunately the amount of harmonic content beyond the fifth is relatively small and, to a first approximation, calculation of the second and third harmonic content only will suffice. This can be determined by finding the anode voltage corresponding to 0·707 of the positive and negative input swings and applying the following formulae:

Second harmonic distortion

$$= \frac{\text{difference in peak swings}}{\text{sum of peak swings} + 1\cdot41 \, (\text{sum of } 0\cdot707 \text{ swings})} \times 100 \, \%$$

Third harmonic distortion

$$= \frac{\text{sum of peak swings} - 1\cdot41 \, (\text{sum of } 0\cdot707 \text{ swings})}{\text{sum of peak swings} + 1\cdot41 \, (\text{sum of } 0\cdot707 \text{ swings})} \times 100 \, \%$$

Referring to the example of Fig. 4.4, the required figures are:

| Grid swings | Anode swings |
|---|---|
| +1·0 V | −92 V |
| +0·707 V | −77 V |
| 0 | 0 |
| −0·707 V | +71 V |
| −1·0 V | +98 V |
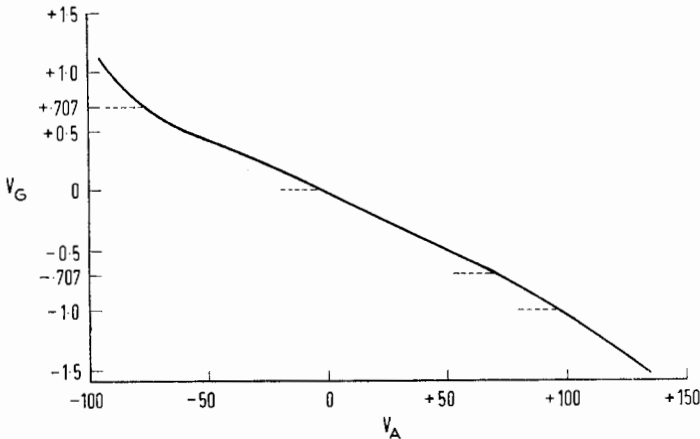


FIGURE 4.4

Dynamic characteristic constructed from Table 4.1. The curvative is different to that of the triode characteristic of Fig. 3.3.

The percentages of second and third harmonic content are ascertained as follows:

$$\text{2nd harmonic} = \frac{6}{190 + 1\cdot41\,(148)} \times 100 = \frac{600}{398} = 1\cdot5\%$$

$$\text{3rd harmonic} = \frac{190 - 1\cdot41\,(148)}{190 + 1\cdot41\,(148)} \times 100 = \frac{1800}{398} = 4\cdot5\%$$

It will be seen that for an input of 1 V peak, the third harmonic distortion is greater than the second. Reference can be made to the data given in the application reports for typical pentode distortion figures. Page 200 relates to the EL84, and it will be seen that at small inputs the second harmonic distortion predominates. However, as the input is increased, third harmonic distortion increases rapidly, while the second harmonic tends to decrease. This decrease in second harmonic can be anticipated by examination of Fig. 4.4, where the deviation from a straight line can be seen to reduce towards the extremes of the anode voltage swings.

### Cathode Bypass

Current feedback due to omission of the cathode bypass capacitor will have the effect of reducing stage gain and distortion, as in the case of the triode (see page 42). In the case of the pentode, the increase of apparent valve impedance due to current feedback will not be significant, as the impedance is already of very high value.

# ADDING STAGES

IT has been shown that stage gain increases as the amplification factor of the valve and its anode load are increased. In practice, the value of the anode load resistor is limited due to the h.t. supply voltage drop which a high value would produce. It is not possible to construct a triode valve which has a very high amplification factor and is satisfactory from other points of view. The amplification which can be obtained from a single valve stage is therefore limited, and so it is often necessary to use more than one stage.

When one stage follows another, the output voltage at the anode of the first stage is applied to the grid of the second stage, where it becomes the input voltage to that stage. This signal is then amplified again by the second stage and the output voltage appears at its anode. Thus if the first stage gives a gain of 50 and the second a gain of 40, the total gain will be $50 \times 40 = 2,000$ and an input of 5 mV at the grid of the first stage will produce 10 V at the anode of the second.

## Resistance-Capacity Coupling

Apart from the effect of the cathode bypass capacitor, no mention has been made of the frequency response of an amplifier stage, but this has to be further investigated when two stages are employed. The anode of the first valve cannot conveniently be connected directly to the grid of the second valve, as this would upset the biasing requirements of the valve. As the signal voltage at the anode of the first stage is alternating, the anode may be coupled to the grid of the following stage by means of a capacitor. So that this grid may still be negative with respect to cathode, a resistor must be connected from the grid to earth, called the *grid resistor*.

The maximum value of this resistor is stated by the valve manufacturers. In the case of voltage amplifier valves, the grid resistor must normally not exceed 2 M$\Omega$, and for large output valves, 250 k$\Omega$. As the grid is held negative, it does not normally attract electrons;

D *A*

but if the input swing is such as to cause the grid to become positive, the grid will collect electrons and grid current will flow in the grid resistor, affecting the operation of the valve. If the value of the grid resistor is high, electrons will accumulate at the grid, preventing correct operation until they leak away to the cathode. A low value grid resistor will assist in avoiding this 'blocking' effect, should the input occasionally exceed the design maximum.

### Low Frequency Response

The coupling capacitor and the grid resistor form a divider across the output of the preceding stage, and the proportion of the output appearing at the following grid will depend upon the frequency of the signal. The reactance of the coupling capacitor must be low, compared with the resistance of the grid resistor, down to the lowest frequency at which the circuit is required to be effective (see Fig. 1.12). At the frequency where the reactance of the capacitor is numerically equal to the value of the grid resistor, the loss is 3dB.

When the h.t. supply is obtained via a supply unit from the a.c. mains, it may have 50 and 100 c/s components. These give rise to 'hum' in the signal, the proportion appearing in the output depending upon the ratio of the anode load to the valve impedance. A high resistance load and a low impedance valve will therefore produce the least hum from this cause.

Hum from this source can be reduced by placing between the anode load and the h.t. supply a decoupling resistor and capacitor, the resistance and reactance forming a potential divider to the hum 'signal'. The inclusion of the resistor will have the effect of reducing the working h.t. supply, due to voltage drop, and so the value of the decoupling resistor must be limited. The value of the capacitor, however, can be made as high as the practical limits of size and cost will permit.

It has been assumed that the h.t. supply source is of negligible resistance. In practice, this is not the case and any resistance in the supply will form an extension of the anode load resistance. A portion of the output signal will therefore appear in the supply resistance, which is common to all stages. There is a phase difference between the outputs of a two-valve amplifier, so that the parts of the signal developing in the common supply resistance will tend to cancel. In a three-stage amplifier, however, the parts due to the first and third stages will be in phase, and instability can result. The decoupling of the h.t. supply will assist in avoiding this instability, but it must be borne in mind that the reactance of the decoupling capacitor

increases as the frequency decreases, so that decoupling cannot be fully maintained at very low frequencies where instability can still occur. In a single stage, the gain will tend to rise at low frequencies as anode decoupling becomes less effective, and as the finite value of the cathode decoupling capacitor causes a loss in gain due to cathode feedback, in some cases it is permissible to arrange for these effects to cancel, thus maintaining the response down to a lower frequency.

**High Frequency Response**
The amplification of two stages at high frequencies is limited by capacitances which are present in the circuit, but are not indicated by the circuit diagram. The first valve may be said to form a capacitor due to the proximity of its anode and cathode, the value depending upon the physical dimensions and spacing of these electrodes. Similarly the grid and cathode of the second valve form a capacitor. In a practical amplifier, the valve holder and the wiring of the electrodes to external components will introduce small capacitances which are collectively referred to as stray capacitances. There is a further capacitance due to the anode and grid of the second valve, the capacitance of which is increased by the gain of the stage and so may be of considerable significance in high gain valves except pentodes. This apparent increase in the capacitance is called 'Miller Effect', the capacitance being $C_{g-a} (1 + A)$ where $A$ is the stage gain.

These various capacitances added together are referred to as the 'shunt' capacitances $(C_s)$ and have the effect of forming a divider with the first valve impedance, its load resistor and the following grid resistor all in parallel $(R_p)$. The reactance of $C_s$ must be high compared with $R_p$ up to the highest frequency at which the circuit is required to be effective. At the frequency where the reactance of $C_s$ is numerically equal to $R_p$, the loss is 3 dB. If the required high frequency response cannot be obtained with a particular valve, it may be necessary to employ a lower impedance type and operate with a low value load resistor.

**A Practical Two-Stage Amplifier**
The design of a practical two-stage resistance-capacity coupled amplifier may now be attempted, and the performance assessed. The circuit is that of Fig. 5.1, comprising two similar triode stages of the type described in Chapter 3 and illustrated in Fig. 3.4.

The two valves are the two sections of an ECC83 (12AX7) valve, contained in the same envelope. The pin connections at the valve base are shown in Appendix II, page 185.

It will be assumed that the h.t. supply line is 350 V and that the decoupling resistors $R_D$ are 68 kΩ. A graphical analysis shows that when $R_k = 2 \cdot 2$ kΩ the anode current is $0 \cdot 75$ mA, so that the voltage drop in $R_D$ is $68 \times 0 \cdot 75 \simeq 50$ V. If the grid resistor $R_G$ is 470 kΩ, the first stage is that discussed on pp. 40, 41, where the stage gain was found to be 50, with 2% second harmonic distortion. The gain of the second stage will depend upon the value of any resistance shunted across the output. If the second stage were resistance-capacity coupled to a further valve with a grid resistor of 470 kΩ, its gain would also
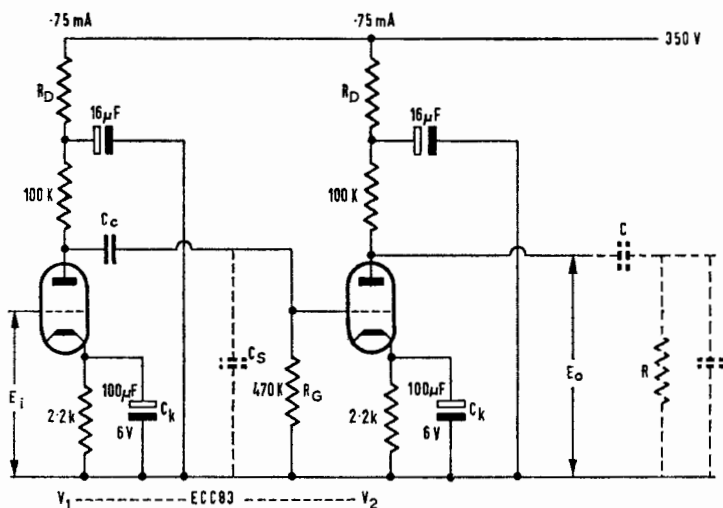


FIGURE 5.1

Complete circuit of two stage ECC83 voltage amplifier. The low frequency response depends upon thè coupling capacitor $C_C$ and the high frequency response upon the shunt capacitor $C_S$. The response will be affected if $V_2$ is coupled via the capacitor $C$ and there is shunt capacitance at the output.

be 50, and the total for the two stages would be 2,500. For 2% distortion, the input to the second state is limited to $0 \cdot 5$ V, so that the maximum output is $0 \cdot 5$ V $\times 50 = 25$ V, and the maximum input to the first stage, $0 \cdot 5/50 = 10$ mV.

If $C_D$ and $C_k$ are large enough, their effect upon the low frequency response down to 30 c/s can be ignored. Suitable values would be $C_D = 16$ μF and $C_k = 100$ μF. The low frequency response will depend upon the coupling components $C_c$, $R_G$. Now $R_G$ is 470 kΩ, so if the reactance of $C_c$ is 470 kΩ at 30 c/s, the loss at this frequency will be 3 dB, which is adequate for the highest quality reproduction.

Reference to Appendix I, page 167, shows that a value of $0\cdot01$ $\mu$F is suitable for $C_c$. If the second stage were resistance-capacity coupled and $C$ was $0\cdot01$ $\mu$F and $R$ 470 k$\Omega$, the low frequency loss would be doubled, i.e. 6 dB at 30 c/s, or 3 dB at 60 c/s.

The high frequency response depends upon the shunt capacitance $C_S$ which includes the anode to grid capacitance of $V_2$ increased by the stage gain (the Miller effect). The value of this capacitance, (Appendix II, page 185), is $1\cdot6$ pF, and the stage gain is 50, so that the capacitance due to Miller effect is $1\cdot6 \times 51 = 81\cdot6$ pF. Total $C_S$ is therefore:

| | |
|---|---|
| Miller effect | 81·6 |
| $V_1\,C_{a-k}$ | 0·3 |
| $V_2\,C_{g-k}$ | 1·6 |
| Stray capacitances | 10·0 (average figure) |
| Total $C_S$ | 93·5 pF |

The value of $R_p$ (page 51) is

$$\cfrac{1}{\cfrac{1}{r_a} + \cfrac{1}{R_L} + \cfrac{1}{R_G}} = \cfrac{1}{\cfrac{1}{65,000} + \cfrac{1}{100,000} + \cfrac{1}{470,000}} = 36\cdot5\text{ k}\Omega$$

The high frequency response will therefore be 3 dB down at the frequency at which the reactance of $93\cdot5$ pF is $36\cdot5$ k$\Omega$.

Reference to Appendix I, page 164 shows that for the nearest capacitance, 100 pF, this frequency would be between 30 and 50 kc/s.

More accurately $X_c = 36,500$ or $\dfrac{1}{6\cdot28 f \times 100 \times 10^{-12}} = 36,500$

so that $f = \dfrac{1}{6\cdot28 \times 100 \times 10^{-12} \times 36,500} = 43\cdot7$ kc/s.

Any shunt capacitance across the output stage will cause an additional loss at high frequencies. If this shunt capacitance were also about 100 pF and $R_G$ was 470 k$\Omega$, the overall loss would be 6 dB at $43\cdot7$ kc/s, or 3 dB at $21\cdot85$ kc/s.

# POWER AMPLIFIERS

IN the previous chapters attention has been directed to the voltage change at the anode of a valve in an amplifier stage. It was observed that voltage change is due to current change in the anode load. *Power* is therefore developed, being the product of the r.m.s. voltage change and the r.m.s. current change. The current change in a voltage amplifier stage is relatively small, so that the power in the load is low and in any event is of no particular significance.

**Loudspeaker Load**

The most commonly used loudspeaker is the moving coil type. This is constructed by winding turns of wire upon a hollow former which is freely suspended in an annular gap in a block of magnetic material, a high flux magnetic field being maintained across the gap.

If an alternating current is made to flow in the coil, the coil is driven in and out of the gap due to the magnetic field, and if it is attached to a cone diaphragm, sound waves are set up in the air. The size of the coil which can be constructed in a practical loudspeaker determines its resistance and, in the case of high fidelity loudspeakers, this is usually about 12 $\Omega$. The inductance of the coil, together with the attached diaphragm and the suspension arrangement, introduces a reactive element which varies with the frequency of the alternating current in the coil. So it is the *impedance* of the speaker which is generally specified. This is usually 15 $\Omega$ at about 400 c/s and increases as the frequency falls below and rises above 400 c/s due to the factors referred to above. The matter is further complicated due to the fact that the reactance is inductive at high frequencies and capacitive at low frequencies (down to cone

resonance). For the present it will be assumed that the loudspeaker forms a non-reactive load of 15 Ω.

There are two reasons why it is not possible to connect the ends of the loudspeaker coil between the h.t. supply and the anode of a valve to form a load. 1. The steady current drawn by the valve would cause the coil to move and be permanently displaced in the magnetic gap. 2. No practical valve can develop the required power in a load as low as 15 Ω, which would require a current of 1 A to give 15 W.

### Output Transformer

These difficulties are removed by the use of a coupling transformer The primary of this transformer is connected between the h.t. positive line and the anode of the valve, and the loudspeaker coil is connected to the secondary winding. The load impedance seen by the valve will depend upon the square of the turns ratio of the primary and secondary windings, and is given by the expression $R_L = n^2 \times 15$ ohms (see page 25).

The performance of an output stage can be assessed by the graphical method used for the voltage amplifier, although the complicated nature of the transformer and loudspeaker load make the case more difficult. The valve is required to develop power in the load, so that the analysis must take account of the current change in addition to the voltage change, to arrive at the power ($W = VI$). The best value for the load resistance will be that which permits maximum power to be obtained with minimum distortion. The value may be fairly critical and is called the optimum load.

### Single Output Valve

Taking the case of the EL84, a typical output pentode, the maximum h.t. supply voltage for single-valve operation is 250 V for anode and screen grid ($g_2$). The load will be presented by a transformer and it will be assumed that the secondary is loaded by a 15Ω resistor. The transformer ratio will then be $n = \sqrt{\dfrac{R_L}{15}}$. Ignoring the voltage drop due to the resistance of the transformer primary winding, the standing voltage at the valve anode and screen will be 250 V and the current will depend upon the control grid ($g_1$) bias, normally obtained by the cathode resistor method. The anode characteristic shows that, when $V_a = V_{g2} = 250$ V, the anode current must be limited to 48 mA for a maximum anode rating of 12 W, and this is obtained by a control grid voltage of $-7 \cdot 3$ V.

The valve data also shows that, at this working point, the screen current will be 5·5 mA. As the anode current and screen current originate from the cathode, the cathode current will be 53·5 mA.

FIGURE 6.1

Circuit of output pentode stage with cathode biasing. The load resistance depends upon the ratio of the output transformer and is 15 $n^2$ ohms, where $n$ is the transformer turns ratio. The input voltage $E_i$ will cause voltage and current changes in the transformer primary, and hence power will be developed in the 15 Ω resistor, connected to the secondary.
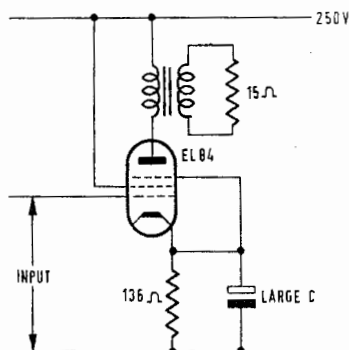


For a bias voltage of −7·3 V, the cathode bias resistor will be $\frac{7·3}{53·5} \times 1,000 = 136$ Ω. The circuit is shown in Fig. 6.1.

**Optimum Load**

Using this working point, several trial load lines may be drawn as in Fig. 6.2. It is apparent that if the load resistance is high, the current change will be small, and if it is low, the voltage change will be small. In either case maximum power in the load will not be produced. If the grid is not allowed to become less negative than −1 V, the posi-

TABLE 6.1   *Application of Single EL 84*

| Case | Anode V @ −1·0 V | Anode I @ −1·0 V | Anode V @ −13·6 V | Anode I @ −13·6 V |
|------|------------------|------------------|-------------------|-------------------|
| 1    | 77 V             | 118 mA           | 355 V             | 4·5 mA            |
| 2    | 35 V             | 92·5 mA          | 455 V             | 5·0 mA            |
| 3    | 22 V             | 72·5 mA          | 630 V             | 5·5 mA            |

tive input swing will be 6·3 V, so that the corresponding negative swing will be to −13·6 V. For the three load lines drawn, the anode voltage swings and the current swings can be noted.

These are peak to peak voltage and current swings, which must be halved to give the peak swing and then divided by $\sqrt{2}$ to give the

56

r.m.s. value from which the power may be calculated. The power will then be:

$$\frac{\text{Max. voltage swing} - \text{Min. voltage swing}}{2 \times \sqrt{2}}$$

$$\times \frac{\text{Max. current swing} - \text{Min. current swing}}{2 \times \sqrt{2}}$$

$$= \frac{(\text{Max V} - \text{Min. V})\,(\text{Max I} - \text{Min I}).}{8} \; (\text{I in amps})$$

For the three cases quoted above, the power output is found to be 3·92 W, 4·59 W and 1·29 W respectively.

The load line of Case 2 produces the highest power output and can be found as $R = \dfrac{V}{I} = \dfrac{250}{100 - 48} \times 1{,}000 = 4{,}800\ \Omega$. Further investigation of a series of load lines around this area can be carried out to determine the optimum load, bearing in mind 2nd and 3rd
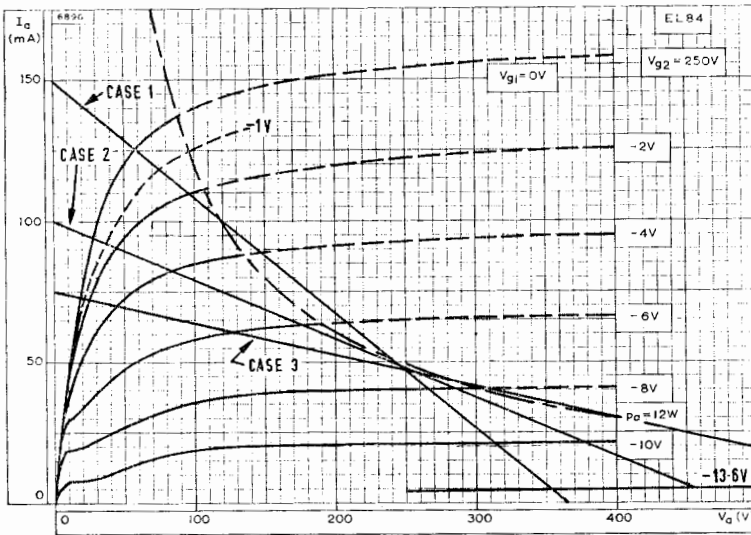
Anode characteristic of EL84 output pentode with load lines for $V_a = V_{g2} = 250$ V. The voltage gain and power output can be found for three values of load.

harmonic distortion calculated by the method previously described for voltage amplifiers, page 47.

Such an investigation involves a considerable amount of work, but as a valve is only likely to be used under conditions which will

produce the best performance, the valve manufacturers publish full data with optimum load figures, a selection of these being contained in Appendix II.

### Reactive Load

The complicated nature of a transformer and loudspeaker load has been referred to, and the load presented will not be resistive except
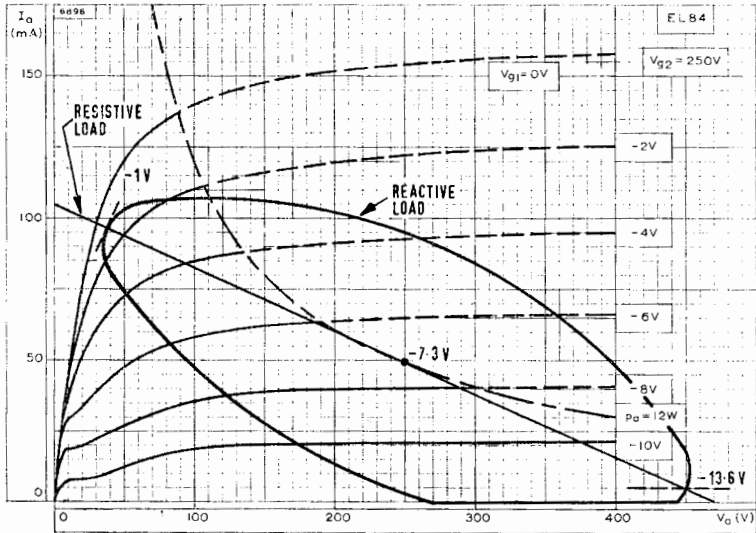


FIGURE 6.3

When the load has a reactive component the load 'line' is an ellipse. Distortion is caused unless the input is restricted or the value of the load increased, i.e. by increasing the transformer turns ratio.

over a restricted frequency range. When the load is reactive, the load 'line' becomes an ellipse as in Fig. 6.3 and the performance is affected. It is apparent that, as the minor axis of the ellipse increases, there is a possibility of the valve operating under cut-off conditions over a portion of the cycle, causing considerable distortion. This can be avoided by shortening the major axis, i.e. reducing the input, which reduces power output.

### Frequency Response

The voltage amplifier stage feeding the output stage will be designed to produce the required input voltage with low distortion; and the frequency response of the voltage amplifier will take into account the input capacitance of the output valve and the value of its grid

58

resistor. The frequency response of the output stage is dependent upon the design of the output transformer. The primary winding of the transformer is an inductor (measured in Henrys), so that its reactance will depend upon frequency, being lowest at the lowest frequency. For 3 dB loss at the lowest frequency required, the reactance of the transformer primary inductance when the winding is carrying the standing anode current, must equal the value of the load resistance $R_L$, and the valve impedance $r_a$ in parallel. The high frequency response will depend upon the leakage inductance of the transformer and also upon the capacitance of windings one to the other, and to the frame which is normally earthed. The leakage inductance will cause a loss of 3 dB when its reactance is numerically equal to $r_a$ and $R_L$ in series.

The output transformer will normally be purchased from a specialist manufacturer and the primary inductance and leakage inductance will be stated. The transformer itself gives rise to distortion and the core should be of adequate size and good quality material. The windings should be of low d.c. resistance to avoid losses. This component is of a complicated nature and unless it is properly constructed, a good performance cannot be obtained. Triode, tetrode or pentode valves may be used in the output stage.



FIGURE 6.4

Triode, tetrode and pentode valves may be used in the power output stage, with transformer connections as shown.

## Push-Pull Operation

The power output of the output stage can be increased, and distortion reduced, by using two valves in push-pull. When two valves are so connected, distortion due to even order harmonics cancels in
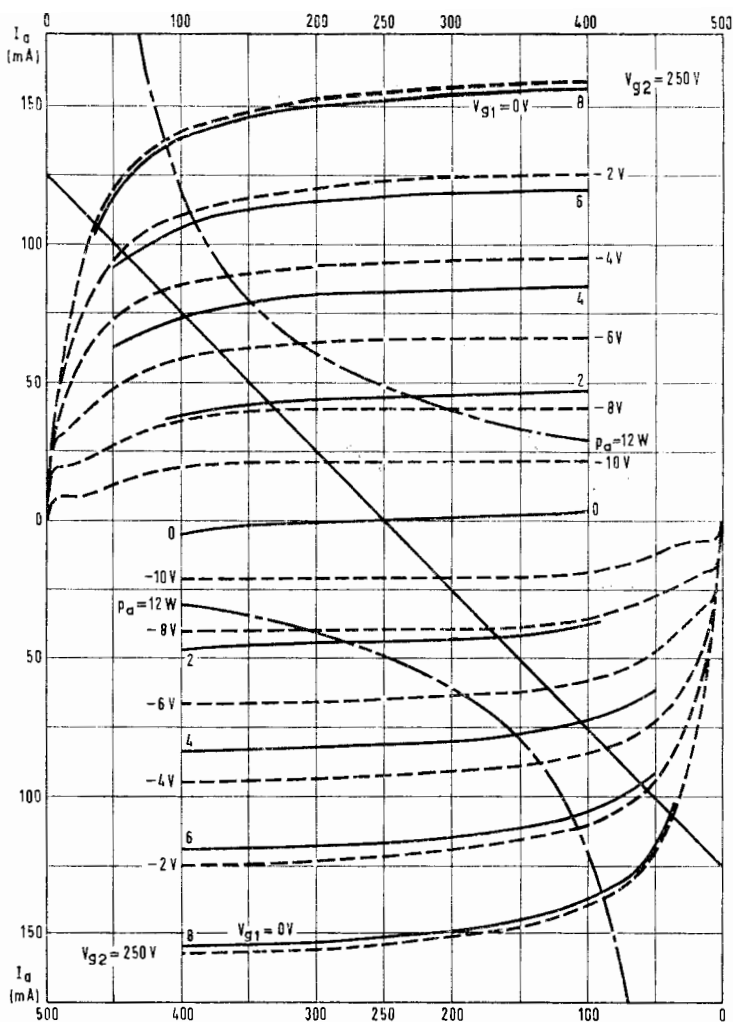
**FIGURE 6.5**

The solid lines show the composite anode characteristic for two
EL84 valves in push-pull, at a h.t. supply voltage of 250 V. Anode
to anode load line 8 kΩ.

the output transformer. This can be appreciated by the construction
of a composite anode characteristic which shows the nature of an
equivalent single valve.

A composite characteristic for two EL84 valves, with an h.t.
supply of 250 V, is shown in Fig. 6.5. The straightening and more
equal spacing of the grid voltage lines indicates that distortion will

be reduced. The power output with a resistive load can be calculated from the composite characteristic, but again the valve manufacturers recommend operating conditions in detail. Another big advantage of push-pull operation is that the flux in the transformer created by the standing current of the two valves is zero, due to opposite winding senses of the two halves of the primary. This enables a high value to be obtained for the primary inductance (Fig. 6.6).

### Ultralinear Operation

A variation of push-pull is the distributed load circuit, of which 'ultralinear' connection is an example. In this case, the screen grids of the valves are not connected direct to the h.t. line, but are connected to taps on the output transformer primary winding. This connection introduces a degree of negative feedback into the output stage, and has the further advantage of reducing the power at which the screen grids operate.

The performance of an ultralinear stage depends on the percentage of the primary turns included in the screen grid circuit. When the
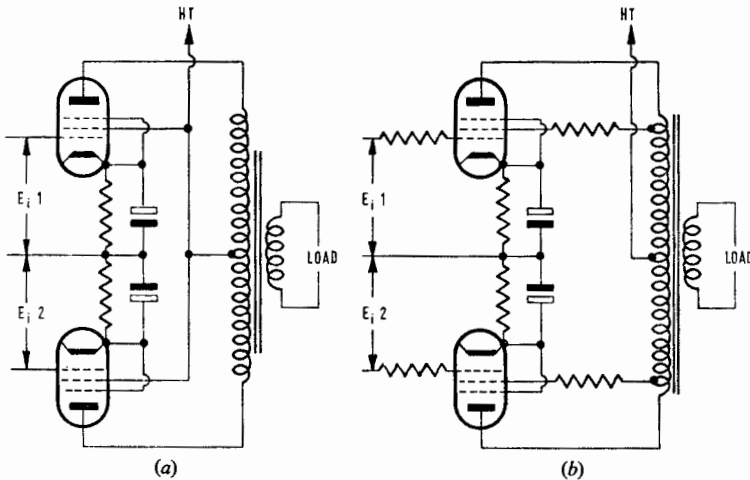


FIGURE 6.6

(a) Circuit of push-pull pentode output stage. In some cases it is permissable to connect the cathodes and use a common bias resistor without a bypass capacitor.

(b) Circuit of 'ultralinear' output stage in which the screens are connected to taps on the output transformer primary. The resistors in the screen and grid circuits are 'stoppers'.

screen is tapped at h.t. positive, the valve operates as a normal tetrode or pentode, and if the screen is moved to the opposite end of the primary, so that it is connected to the anode, the valve then

61

operates as a triode. (The valve data for pentodes usually gives data for triode connection.) When the screen is mid-way between h.t. positive and the anode, the distributed load effect modifies the performance, and the valve tends to have the efficiency of a tetrode or a pentode with the lower impedance of a triode.

Efficiency and performance can be compromised by selecting a suitable tap, although considerations of output transformer design also enter into the matter. In practice, the percentage of primary turns included in the screen grid circuit are nominal standards of 10%, 20% or 43%, the last-named being regarded as the closest approach to triode operation which retains an adequate degree of tetrode or pentode efficiency (see page 231).

Here again the design of the output transformer is of paramount importance as leakage reactance between the tapped portions and the other windings, and the various capacitances between the sections will affect the performance.

### Cathode Biasing

The output valves are normally biased by the use of a resistor in the cathode circuit which must be adequately bypassed unless current feedback is required. In push-pull, both cathodes may be biased by individual cathode resistors and capacitors or by using a common cathode resistor. In the latter case the resistor carries the current of both valves, so that its value should be halved. The signal currents will be equal and opposite and will cancel in the cathode circuit, so that no bypass capacitor is required if perfect balance is maintained. In most cases the common cathode resistor is bypassed when the output valves are large, and in some cases individually bypassed cathode resistors are recommended.

### Class B Operation

The distortion cancelling effect of push-pull connection can be used to increase the output of the stage by overbiasing. The peak negative grid swing will then cause the valve to approach cut-off, but at this time the other valve of the pair will receive a positive swing so that it is operating over the linear part of the characteristic, thus maintaining the output. Carried to its extreme, the condition is called Class B, when on alternate half-cycles one of the valves is virtually cut off. In this case, cathode biasing cannot be used and the grids must be negatively biased from a separate supply, the operation then being 'Class B fixed bias'. A compromise is also possible, retaining cathode bias, often called Class AB1.

Stray capacitance in the push-pull circuit can form resonant circuits with inductances which may receive sufficient energy to maintain oscillation (see page 26). The frequency of oscillation is high, and the tendency to oscillate can be reduced by adding resistors in series with the grid and screen so that the valve capacitances bring about a loss of high frequency gain. These resistors are called 'stoppers'.

When two valves are used in push-pull, separate inputs are required for each valve, these inputs being 180° of phase, but otherwise balanced. As the voltage amplifier section of the amplifier produces a single output, a 'phase splitting' device has to be employed (see Chapter 8).

# NEGATIVE FEEDBACK

IT has been shown in Chapter 3 that when the bias voltage for a valve is developed in an unbypassed cathode resistor, a signal voltage will also be developed in this resistor. This signal voltage subtracts from the input voltage, and hence the effective gain of the stage is reduced. When the gain is reduced in this way it is described as reduction of gain by *current negative feedback*, because voltage is generated by the flow of output current in a resistor.

Efforts are usually made to avoid current feedback of this kind since not all of its effects are beneficial in audio frequency amplifiers. However, *voltage negative feedback* is very commonly used to improve performance, particularly in power amplifiers, and some account of the principles involved will be given here.

Initially the discussion will be limited to the one case of series voltage negative feedback. As the term implies, in this case some proportion of the output voltage is fed back and placed in series with the input voltage, so that the true input to the amplifier is the difference between these two voltages.

### Benefits of Negative Voltage Feedback

Before proceeding to a more detailed discussion of the effects of negative feedback, it should be emphasized that the advantages are gained at the expense of overall amplification. The advantages may be summarized as:

1. Amplifier performance becomes much less dependent upon changes in valve characteristic and power supply voltage.

2. Frequency response is improved, i.e. variation of amplifier gain with frequency is greatly reduced.

3. Non-linear distortion is reduced, i.e. curvature in dynamic characteristics (Fig. 3.3) is effectively 'straightened out'.

4. Output impedance is reduced, thus the output voltage for a given input voltage will be less dependent upon variations in load impedance.

5. Input impedance is increased, thus reducing the load which the amplifier will impose on any preceding signal source.

Hum and electronic noise are also reduced but, except in very special cases, the signal is reduced by the same factor, consequently the significant quantities of the signal-to-noise ratio and signal-to-hum ratio are unaffected.

### Feedback and Frequency Response

It is not easy to give a clear statement of the physical processes in negative feedback, since what happens in one portion of the circuit
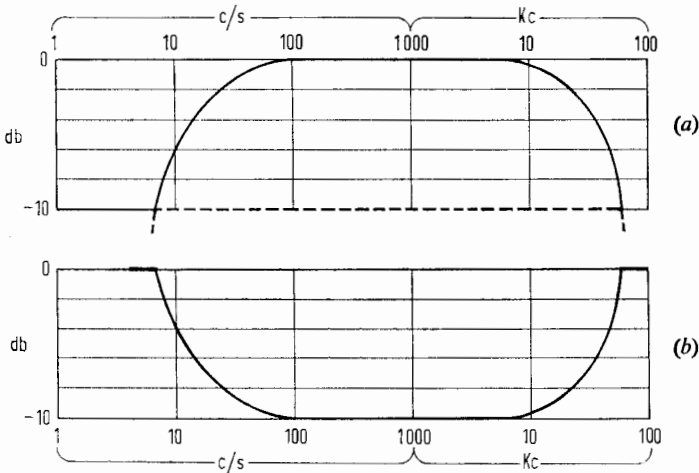


FIGURE 7.1

(a) Frequency response of amplifier without feedback.

(b) If the input could be made to vary with frequency according to this response curve, the overall response would be that shown by the broken line in (a).

As the 'shape' of the variation of input required is the inverse of the output response, a proportion of the output can be fed back to the input in opposite phase, thus reducing it in the manner required. It is apparent that this will reduce the gain and represents 10 dB of negative voltage feedback.

affects all other parts in a rather involved way. However, it is helpful to refer to the frequency response shown in Fig. 7.1(a) which is

typical of a simple amplifier. The output voltage for a constant input voltage is reduced at high and low frequencies by the shunt and series capacitive reactances, to which reference has already been made.

It would be possible to maintain the amplifier output voltage more nearly independent of frequency by subtracting a variable amount from the input signal. This would have the effect of reducing the true input to the amplifier and hence the output voltage. For example, at low frequencies the amount subtracted would be very small but as the frequency is increased a greater amount would be subtracted to keep the output voltage constant. Then again at high frequencies, smaller quantities would be subtracted in an effort to keep the output constant.

From this simple idea it can be seen that, if the output is to be kept more nearly constant as the frequency is varied, the voltage subtracted from the input signal must vary as a function of frequency in a manner similar to that shown in Fig. 7.1($b$). This leads directly to the idea of taking some fraction of the output voltage ($\beta$) and feeding it in series with the input signal so that it subtracts from this signal. Thus, when the amplifier gain is zero, there will be no voltage subtracted from the input and, when the gain is very large, a voltage almost equal to the input signal will be subtracted. In the first case we notice that the whole of the input signal is applied to the amplifier whereas, in the second, only a very tiny fraction of this signal reaches the



FIGURE 7.2

Amplifier $A$ with a proportion $\beta$ of its output fed back in series with the input.
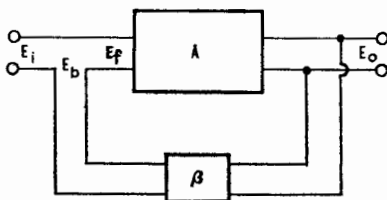
amplifier input. An approach to the desired conditions has therefore been made, for, when the amplifier gain is large, its true input is made small and conversely, when the gain is low, the input is made large.

**Basic Formula**

The simplified diagram of an amplifier with feedback is shown in Fig. 7.2. For convenience the amplifier has been shown as a block having an input voltage $E_f$, an output voltage $E_o$ and a gain $A$. In addition, it will be noticed that the input voltage is made up by combining the signal voltage, $E_i$, and the feedback voltage $E_b$, and

66

$E_b$ equals $\beta E_o$. Finally, it may be remembered that there is a definite relationship between the voltages $E_f$ and $E_o$, namely $E_o = E_f A$. So that

$$E_o = E_f A$$
$$\therefore E_o = (E_i + \beta E_o)A$$
$$= AE_i + A\beta E_o$$
$$\therefore E_o - A\beta E_o = AE_i$$
$$\text{or } E_o(1 - A\beta) = AE_i \tag{1}$$

The gain of the amplifier with feedback is given by $E_o/E_i$ which can be found from formula (1) to be

$$\frac{E_o}{E_i} = \frac{A}{1 - A\beta} \tag{2}$$

This is the fundamental equation which shows the gain of the amplifier with feedback in terms of the basic gain $A$ and the fraction $\beta$ of the output which is fed back. It should be noted that the gain without feedback is reduced by the factor $1 - A\beta$ which is referred to as the gain reduction factor $F$.

It will be noticed that, contrary to previous statements, equation (2) shows that the gain is increased since $1 - A\beta$ is less than unity. This apparent contradiction arises from the fact that the equation has been derived for the general case of $E_i + E_b$ and this is in fact positive feedback. The negative feedback case occurs when $E_b$ is negative with respect to $E_i$, i.e. if $A$ or $\beta$ (but not both) are negative. As the case of negative feedback is that most frequently employed in practice, the gain reduction factor $F$ is $1 - (-A\beta)$ which becomes $1 + A\beta$.

In terms of a simple sine wave input, $E_b$ can be made negative by shifting its phase by $180°$ from that of $E_i$. From the previous discussion of amplifier stages, it will be remembered that each valve stage does in fact shift the phase of the signal by $180°$. Thus, by selecting the number of stages over which feedback is applied, the appropriate sign of feedback voltage may be achieved. In other words, shifting the phase of a sine wave signal by $180°$ may just as conveniently be represented by multiplying the signal by $-1$.

Then, if $A$ is such as to introduce $180°$ phase shift, we may represent it by writing $-A$.

This then would give negative feedback and the equation for gain in this case becomes:

$$\frac{E_o}{E_i} = \frac{-A}{1 + A\beta}$$

The minus sign in this equation shows clearly that the output voltage will be 180° out of phase with respect to the input, and permits the effect of negative feedback to be calculated. As negative feedback is the usual case the minus sign is frequently ignored.

Assume that $A$ is large, say 200. Then, if the fraction feedback $\beta$ is zero, the amplifier gain becomes $\dfrac{E_o}{E_i} = \dfrac{200}{1}$, or 200. If $\beta = 1$, i.e. all of the output voltage is fed back to the input, the amplifier gain becomes $\dfrac{E_o}{E_i} = \dfrac{A}{1 + A\beta} = \dfrac{200}{201} = 0\cdot995$. Thus the gain in this latter case becomes very nearly equal to unity, and we notice that this is also the value of $\beta$.

It is possible to simplify the situation still further if we can assume that, in any given amplifier, the product $A\beta$ is very much larger than unity. Making this assumption, the equation for the gain of the amplifier with feedback becomes:

$$\frac{E_o}{E_i} \simeq \frac{A}{A\beta} = \frac{1}{\beta}$$

The validity of this assumption can be tested, and some interesting points illustrated, by taking a few numerical examples. Let $\beta = 0\cdot1$ in all examples, and in the first case let $A = 100$. Then by the approximate formula:

$$\frac{E_o}{E_i} \simeq \frac{1}{\beta} = 10$$

and precisely,

$$\frac{E_o}{E_i} = \frac{100}{1 + 100 \times 0\cdot1} = \frac{100}{11} = 9\cdot1$$

Thus in this case the approximate solution is 9% in error.
Now let $A = 1,000$.

Again the approximate formula will give $\dfrac{E_o}{E_i} = 10$, but by the precise method the result will be

$$\frac{E_o}{E_i} = \frac{1,000}{1 + 1,000 \times 0\cdot1} = \frac{1,000}{101} = 9\cdot9$$

Here the error is only 1%.
Finally, let $A_o = 10,000$. The precise calculation

$$\frac{E_o}{E_i} = \frac{10,000}{1 + 10,000} \times 0\cdot1 = \frac{10,000}{1,001} = 9\cdot99$$

68

Here the error is only $0 \cdot 1\%$ and in most practical cases would be quite negligible.

From this it can be seen that, providing $A\beta$ is greater than, say, 20, the errors involved in using the approximation are less than 5%, and for most audio frequency applications this approximation is satisfactory.

### Effect of Voltage Negative Feedback upon Gain

More interesting, however, is the fact that in the numerical examples given, although the gain without feedback, ($A$) changes from 100 to 10,000, i.e. by a factor of 100, the gain with feedback only changes from $9 \cdot 1$ to $9 \cdot 99$, i.e. by a factor of $1 \cdot 1$. This clearly illustrates one of the advantages of negative feedback, namely that the gain of an amplifier may be made sensibly independent of changes in the valve parameters, etc. It will also be clear that this advantage is greatest when $A\beta$ is very much greater than unity.

The fall off in the response of an amplifier at high and low frequencies is due to a reduction of $A$ at these frequencies. It is clear, therefore, that the application of negative feedback will tend to keep the response more nearly constant over a wide frequency range. However, it must be noted that the application of negative feedback to an amplifier may give rise to a sharp peak in its response at one or more frequencies. This is discussed more fully under the heading of stability.

### Effect of Negative Feedback upon Distortion

By arguments similar to those used in defining the effect of feedback on the overall gain, it can be shown that distortion generated in the amplifier is reduced in the same ratio as that of the gain. More precisely, if D is the harmonic distortion without feedback, then the distortion with feedback becomes:

$$\frac{D}{1 + A\beta}$$

There are a number of simplifying assumptions made in deriving this equation, but they do not invalidate its use in most practical problems.

### Input and Output Impedance

To calculate the change in the output impedance of a multi-stage amplifier due to the application of negative feedback, it is necessary to separate the parameters of the last stage from those preceding it.

69

Allowing the valve in the last stage to have an amplification factor $\mu$ and an anode resistance $r_a$, we may write the output impedance in the presence of feedback, $r_a'$ as:

$$r_a' = \frac{r_a}{1 + \beta\mu A_1}$$

where $A_1$ is the voltage gain of the stages before the last and $\beta$, as before, is the fraction of the output voltage which is fed back over the whole of the amplifier. To be strictly accurate, in calculating the true output impedance it is necessary to take into account the load resistance, $R_L$, of the output stage. Thus the true output impedance will be given by

$$\frac{R_L r_a'}{R_L + r_a'}$$

Finally, it is possible to calculate the input resistance of the complete amplifier with feedback and this can be shown to be $r_i' = r_i(1 + A\beta)$ where $r_i$ is the input resistance in the absence of feedback. From this it is clear that the application of feedback in the manner described will increase the input resistance.

It is important to remember that the results quoted above for an amplifier with feedback are limited to voltage negative feedback, in which the voltage feedback is placed in series with the input signal voltage. There are a number of other cases, such as current feedback, in which the results are quite different from those quoted here. However, negative voltage feedback is the form most commonly used, and the series connection is most favoured in power and other amplifiers where a level frequency response is required.

In frequency compensating stages, voltage feedback over one stage with parallel connection is frequently encountered. In this case the input resistance at the grid is reduced by feedback and becomes:

$$r_i' = \frac{r_i}{(1 + A\beta)}$$

**Phase Shift**

In this brief discussion of negative feedback it has been assumed that the only phase shifts introduced are those of 180° due to each stage of amplification. It will be obvious from the previous chapters that this assumption is not valid in practice, since each capacitance/resistance potential divider combination will not only effect the gain but will also introduce a frequency dependent phase shift. Thus at some mid-frequency, say 2 kc/s, the phase shift through the whole amplifier may be assumed to be 180° but, if the frequency is raised to

say 50 kc/s, the phase shift might easily be reduced to say 30°. Similarly, at 10 c/s the phase shift will be increased and might be 250°.

It is not possible to describe fully here the effects of this frequency dependent phase shift; but that of stability must be briefly dealt with. Taking the simplest situation, it is possible for the phase shift due to a number of capacitive reactances in the amplifier to be equal but opposite in sign from that introduced by the amplifier stages. Thus, at some frequency, the gain of the amplifier may become $+A$ instead of $-A$ as previously assumed. If this happens, the equation for gain with feedback becomes:

$$\frac{E_o}{E_i} = \frac{A}{1 - A\beta}$$

and we notice immediately that this is a condition of positive feedback and that the gain is increased.

In the limiting case, where $A\beta = 1$, we notice that the gain with feedback becomes:

$$\frac{E_o}{E_i} = \frac{A}{1 - 1} = \frac{A}{0} = \text{infinity}$$

(since any finite number divided by zero becomes infinite).

Put another way, this states that the amplifier, when in this condition, can produce an output for an indefinitely small input. An amplifier which reaches this condition will, in practice, oscillate at a frequency for which the phase shifts and gain round the whole amplifier are such as to make $A\beta = 1$.

Although the practical problems of feedback in complete amplifiers are discussed in Chapter 13, two points are worth emphasizing here. Firstly, that if $A\beta$ is positive, it will cause a rise or 'hump' in the frequency response curve (see Fig. 13.1). This indicates that the indiscriminate application of feedback does not necessarily improve the frequency response characteristic of an amplifier. Secondly, if $A\beta = 1$, the frequency of oscillation may be well above the audio frequency range and hence may not be detected except by its indirect effects on distortion.

Summarizing, it may be said that the performance of an amplifier may be greatly improved in all respects, except voltage gain, by the application of negative feedback. In applying feedback to an amplifier, however, extreme care must be taken if these advantages are to be fully realized. Some of the circuit arrangements commonly employed for producing negative feedback in high fidelity amplifiers will now be outlined.

**Applying Feedback in Series with Input**

The effect of current feedback was investigated by graphical analysis on page 42. The gain of the stage with current feedback is

$$\frac{\mu R_L}{R_L + r_a + R_k(1 + \mu)}$$

and the output impedance becomes $r_a' = r_a + R_k (1 + \mu)$.

The cathode resistor is therefore normally bypassed to avoid these effects, but in two cases it is necessary to omit the bypass capacitor.

1. The negative voltage feedback derived from the output of an amplifier is required to be placed in series with the input signal. The most convenient way of doing this is to feed back to the unbypassed cathode resistor of the first amplifier valve (see Fig. 7.3a).

In many cases the value of $R_k$, for the purpose of the feedback circuit, can be quite low, e.g. 100 $\Omega$, which is only a fraction of the normal value of a cathode resistor chosen on the basis of best biasing. Thus the cathode resistor can be split into two parts, $R_{k1}$ and $R_{k2}$, such that $R_{k2}$ is as low as the feedback circuit will permit, and $R_{k1} + R_{k2}$, the value required for correct biasing. $R_{k1}$ may then



(a)　　　　　　　　　　(b)
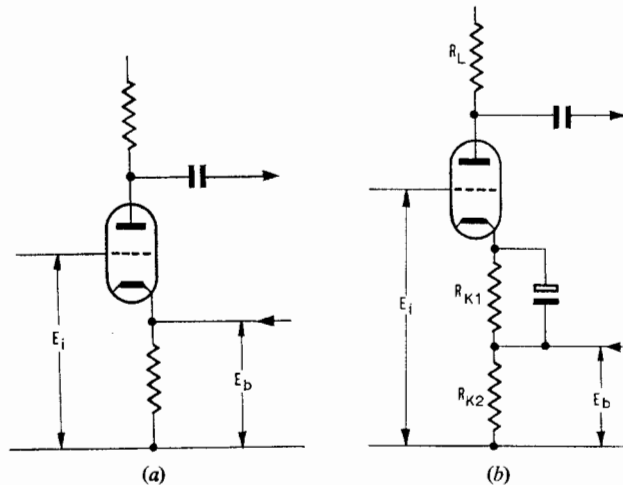
FIGURE 7.3

(a) Unbypassed cathode resistor permitting the application of negative voltage feedback in series with the input signal.

(b) Portion of cathode resistor unbypassed, $E_b$ is the negative feedback signal reducing current feedback.

be bypassed and the reduction in gain and increase in output impedance due to $R_{k2}$ will be small (see Fig. 7.3b).

2. The need for two equal inputs of opposite phase for a push-pull

72

output stage has been mentioned on page 63. One method of obtaining such inputs is to increase $R_{k2}$ so that it equals $R_L$ (see Fig. 7.4$a$).
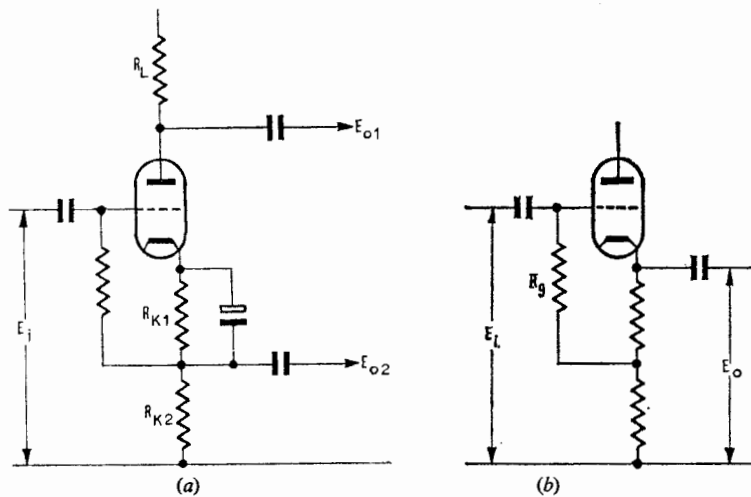


FIGURE 7.4

($a$) Increased cathode resistor giving equal outputs at anode and cathode.

($b$) Load resistor in cathode circuit only, to give cathode follower circuit.

The load of the stage is then $R_L + R_{k2}$, and the output will be divided equally between $R_L$ and $R_{k2}$. The output at the cathode will be in phase with the input and the output at the anode, 180° out of phase with the input. As one-half of the output operates as feedback, the factor $A\beta$ is high and distortion will be very low. On the other hand, the output impedance at the anode will be high. When $R_{k2}$ is high compared with $R_{k1}$, a cathode bypass capacitor will have little effect upon performance and may be omitted.

### Cathode Follower

If the load resistance is placed entirely in the cathode circuit, the whole of the output signal voltage appears at the cathode and is automatically in series with the input (see Fig. 7.4$b$).

This circuit is called a cathode follower, as the output is in phase with the input. The whole of the output is fed back so that $\beta = 1$ and, as the feedback is voltage feedback, the output impedance will be very low. When the stage gain without feedback exceeds 20, the output impedance is approximately $\frac{1}{g}$. The input impedance is high

73

and, when the circuit of Fig. 7.4$b$ is used, becomes $\dfrac{R_g}{1 - \dfrac{AR_L}{R_L + R_k}}$.

The cathode follower is often used to 'match' a high impedance to a low impedance, acting as an 'electronic transformer' of approximately unity gain.

FIGURE 7.5

Series negative voltage feedback applied to a single stage when the input is transformer coupled, i.e. the input circuit does not require to be earthed.
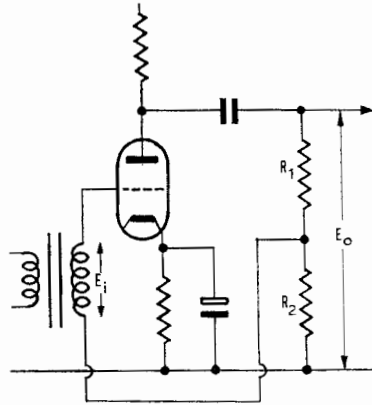
## Voltage Feedback over a Single Stage

A proportion of the output of a single stage can be fed back in series with its input when the stage has an input transformer (see Fig. 7.5). $\beta = \dfrac{R_2}{R_1 + R_2}$, the gain reduction factor is $F = 1 + A\,\dfrac{R_2}{R_1 + R_2}$ and the gain with feedback is $\dfrac{A}{F}$.

## Feeding Back over Two Stages

In the case of a two-stage amplifier, feedback can be applied at the cathode of the first valve as described (see Fig. 7.6).

To avoid the loss of gain due to reduction of the dynamic load impedance of $V_2$, $R_1$ should be high. To avoid reduction of the gain of $V_1$, the unbypassed resistance in its cathode circuit, $R_2$, must be low. If $R_1$ is high and $R_2$ low, the factor $\beta$ will be small, and may in practice be too small to be of significant effect. To obtain a useful value for $\beta$ in this case, it is necessary to reduce $R_1$ and increase $R_2$ to values which will just avoid serious loss of stage gain. The gain of $V_1$ can be increased by making its load resistance and its amplification factor high, and the use of a pentode valve will assist.

74
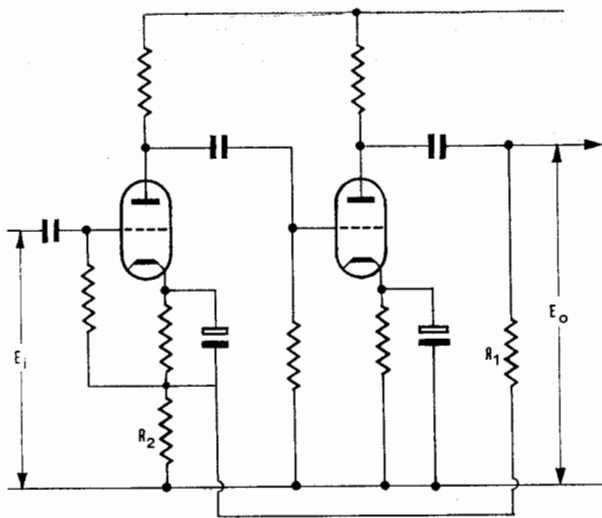
FIGURE 7.6

Series negative voltage feedback with earthed input, applied by
including an additional stage.

## Parallel Feedback

Another method of applying feedback to an amplifier is to feed
back in parallel with the grid circuit. This will have the effect of



(a)                                        (b)
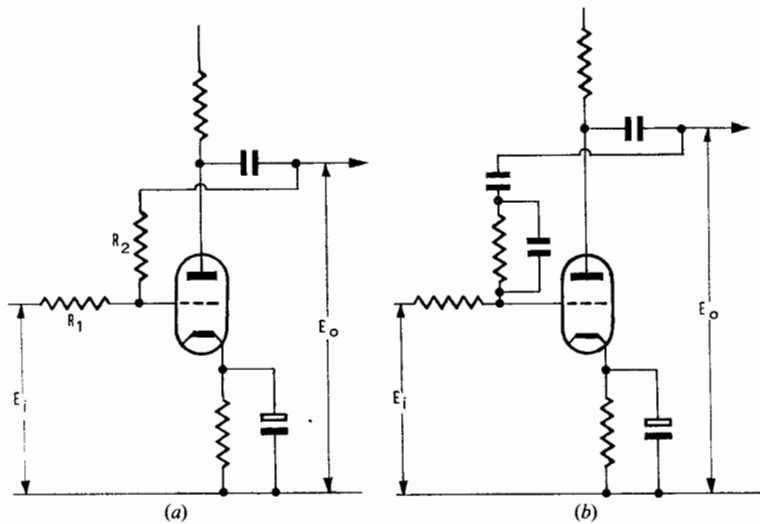
FIGURE 7.7

(a) Parallel voltage feedback applied to a single stage, reducing
input impedance.

(b) Parallel voltage feedback with frequency correction components.

75

making the input resistance low, and so this form of feedback is not often used except in the case of feedback over one amplifier stage. A suitable circuit is shown in Fig. 7.7$a$. When the stage gain is high, the gain with feedback is approximately $\frac{1}{\beta}$, (page 68) which is $\frac{R_2}{R_1}$.

**Frequency Correction by Feedback**

One application of this circuit is to vary the frequency response of the stage by means of capacitors placed in series or parallel with either $R_1$ or $R_2$. Thus for correction of the output of a pickup used to reproduce a l.p. record, it is necessary to increase low frequencies (bass boost) and reduce high frequencies (top cut).

Bass boost can be obtained by effectively increasing $R_2$ at low frequencies, thus reducing feedback and increasing gain at these frequencies. If a capacitor is connected in series with $R_2$, the impedance of the pair will increase as frequency is reduced, giving the required result. In a similar way, feedback can be increased by connecting a capacitor in parallel with $R_2$, so that the impedance of the pair reduces as frequency increases, i.e. giving top cut. The circuit then becomes that of Fig. 7.7$b$. Other uses of this circuit are discussed in Chapter 15.

76

# PHASE SPLITTING

PHASE splitting is the conversion of the single phase output of an amplifier stage to a two-phase output with a phase difference of 180°, for driving a pair of push-pull output valves.

The simplest method of achieving this is to use transformer coupling, and arrange for the transformer to have two secondaries feeding the output valves out of phase. Transformers are not favoured for high fidelity amplifiers, as it is not easy to maintain 180° phase shift at high and low frequencies. Also the transformer, to be effective, needs to be large and so is liable to pick up hum from the mains transformer.

## Cathode Follower

In Chapter 7, a circuit was discussed which provides two out-of-phase outputs from one valve, by dividing the load equally in the anode and cathode circuits. This is a popular and reliable circuit which has two disadvantages.

1. The output of the stage is split between the circuits and is therefore limited.

2. The output impedance at the anode is high, due to current feedback. It therefore gives little damping to the following grid and also makes it difficult to maintain an extended high frequency response when there is high shunt capacitance. Negative voltage feedback gives a low output impedance at the cathode and hence good damping on the following grid circuit. The circuit can be used when the output valves require input voltages not exceeding some 10 or 20 V r.m.s. and is thus suitable for small output pentodes which have a low input capacitance.

The circuit is that of Fig. 8.1 and the performance can be found by graphical analysis. If the valve is one section of an ECC 83 (12AX7), and the h.t. supply voltage is 300 V, $R_L$ and $R_{k2}$ can be 47 kΩ, so that the total load is 94 kΩ. A suitable working point is at $V_g = -2$,

when the anode current is $0 \cdot 82$ mA. The bias resistor is then $\dfrac{2}{\cdot 82}$ kΩ or $2 \cdot 4$ kΩ. From the load line the following table can be constructed:

TABLE 8.1   *Phase Splitter Analysis*

| Grid voltage | Grid Change (A) | Anode Voltage | Total Voltage Change | Anode & Cathode Change (B) | Input Change (A + B) |
|---|---|---|---|---|---|
| $-1 \cdot 0$ | $+1 \cdot 0$ | 166 | $-56$ | 28 | $+29$ |
| $-1 \cdot 5$ | $+0 \cdot 5$ | 195 | $-27$ | $13 \cdot 5$ | $+14$ |
| $-2 \cdot 0$ | 0 | 224 | 0 | 0 | 0 |
| $-2 \cdot 5$ | $-0 \cdot 5$ | 249 | $+27$ | $13 \cdot 5$ | $-14$ |
| $-3 \cdot 0$ | $-1 \cdot 0$ | 269 | $+47$ | $23 \cdot 5$ | $-24 \cdot 5$ |
| $-3 \cdot 5$ | $-1 \cdot 5$ | 288 | $+66$ | 33 | $-34 \cdot 5$ |

The input and output signal voltage can be plotted in the form of a dynamic characteristic, Fig. 8.2, which illustrates how the feedback has improved linearity. It will be seen that the gain from input to one output is less than unity, about $0 \cdot 96$, or a total gain of $1 \cdot 92$ for both outputs.

This performance illustrates the use of the basic feedback formula. The stage gain without feedback is about 50, and

$$\beta = \frac{R_L}{R_L + R_{k2}} = 0 \cdot 5,$$

so that $1 + A\beta = 26$. The total gain is therefore $\dfrac{50}{26} = 1\cdot94$. The second harmonic distortion for 1 volt peak input is

$$\frac{56 - 47}{2(56 + 47)} \times 100 = 4\cdot36\%$$

which is reduced to $\dfrac{4\cdot36}{26} = 0\cdot17\%$.

**Direct Coupling**

It will be remembered that in Chapter 5 it was stated that the coupling capacitor between stages was necessary to prevent the standing voltage at the anode of the first stage from upsetting the biasing arrangements of the second. Circuits are available which permit direct coupling between an anode and the following grid, and are necessary when d.c. changes, i.e. zero frequency, have to be amplified. In high fidelity amplifiers, the lowest frequency ever required to be amplified is 16 c/s, so that the complication of direct coupling can be avoided. The cathode follower phase splitter, however, is one case where direct coupling can be used without circuit complication, in fact the circuit is simplified (Fig. 8.3). In the example above, the ECC 83 valve is operated so that the current
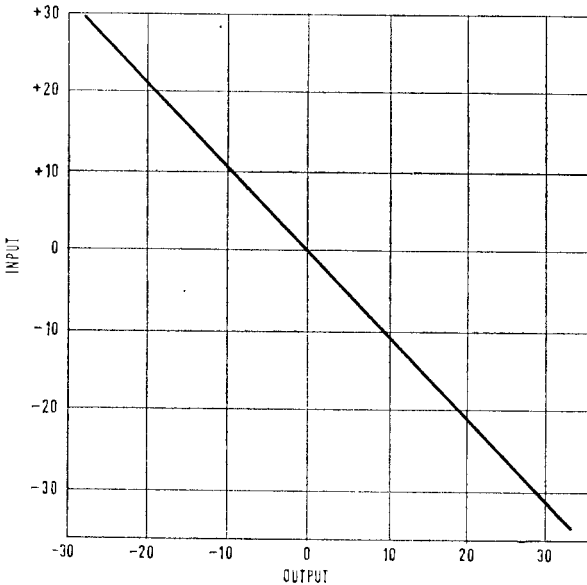


FIGURE 8.2

Dynamic characteristic of Fig. 8.1, showing how the feedback has improved linearity.

is $0 \cdot 82$ mA. The cathode load resistor is 47 K and the voltage across it is therefore $\dfrac{47,000}{0 \cdot 82 \times 10^{-3}} = 57 \cdot 4$ V.

If the cathode bias resistor is omitted and the anode of the preceding stage arranged to operate at $55 \cdot 4$ V and directly coupled to the grid, the grid bias will still be $-2$ V. It would seem at first sight that such a precise difference of voltage would be difficult to achieve and more difficult to maintain. This is not so, as the circuit
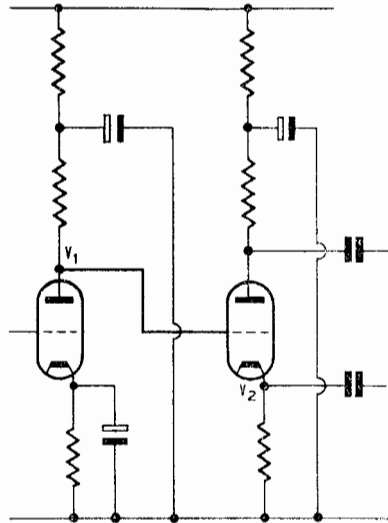


FIGURE 8.3

Version of cathode follower phase splitter circuit using direct coupling.

is largely self adjusting. For example, if the anode voltage applied to the grid rises, say to 60 V, the valve will tend to draw an increased current which will automatically increase the cathode voltage to compensate.

As this circuit operates down to zero frequency, it introduces no low frequency phase shift, a feature which can be of considerable value when feedback is to be applied over a number of stages.

## Anode Follower

It has been shown that, in a parallel feedback stage, the gain is approximately equal to the ratio of the values of the input and feedback resistors, and if these are equal, the gain is near unity. With equal resistors there is 100% feedback and the output impedance and distortion are low. If a normal amplifier stage feeds one output valve, it can feed the other output valve via a stage as described (see Fig. 8.4).

80

This arrangement is called the anode follower phase splitter and two valves are required; but each delivers the full output of which it is capable. The input impedance of the anode follower stage is approxi-

mately that of the input feedback resistor, and this must be taken into account when designing the input valve stage. As unity gain in the second stage is due to heavy feedback, the anode follower will give accurate phase reversal even when there is considerable variation in supply voltage and valve characteristics.
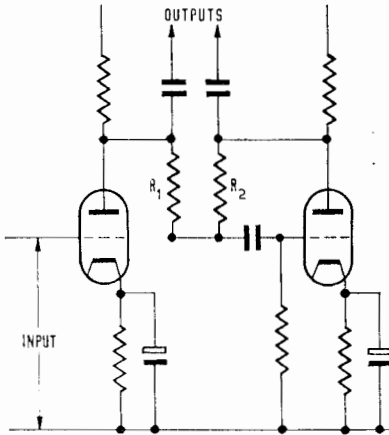
**Cathode Coupled Phase Splitter**

Another popular form of phase splitter uses two valves 'cathode coupled'. An amplifier stage feeds one output valve with 180° phase
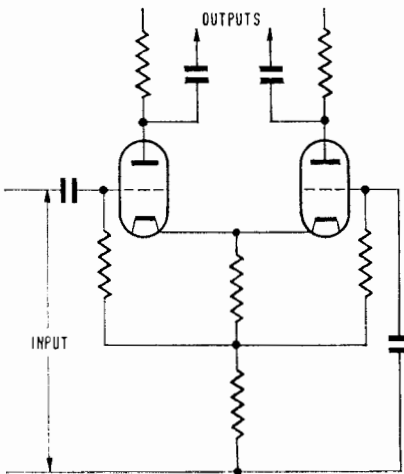
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shift, and its cathode resistor is unbypassed so that a small in-phase signal is obtained. This signal can be regarded as the input to the cathode of the second valve, the grid of which is earthed, so that the valve gives gain without phase reversal. The second valve then feeds the other output valve (see Fig. 8.5).

In practice, the valves have a common cathode resistor in which the signal currents due to the two valves are nearly equal and
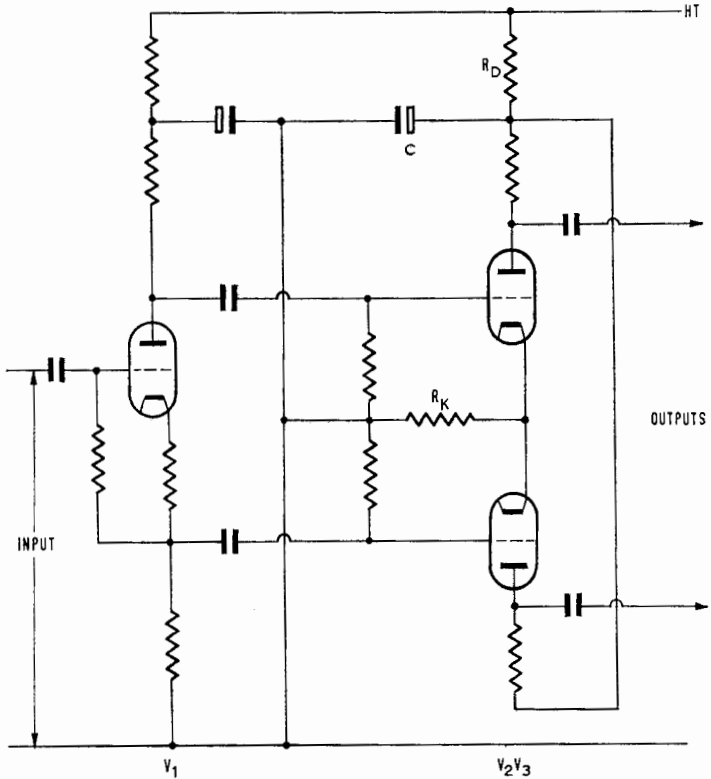


FIGURE 8.6

When push-pull outputs require a large input, drivers can be added to the phase splitter. $V_1$ is a cathode follower phase splitter and $V_2$ and $V_3$ are identical driver valves. $R_D$ is the decoupling resistor and $C_C$, the decoupling capacitor. As the signal currents in the cathode circuits of the drivers are equal and 180° out of phase, they cancel and a common cathode resistor $R_K$ may be used.

opposite, if the operating conditions of the two valves are identical. It is therefore necessary to unbalance the load resistors slightly and to make the common cathode resistor high in value. This circuit reduces stage gain to about one-half that of the basic input stage.

Biasing is arranged so that the common cathode resistor can be made as large as required, and balance can be determined experimentally, if a potentiometer is provided to unbalance the load resistors.

**Drivers**

When the output valves require a large input signal, it may not be possible to obtain this from any form of phase splitter without high distortion. In this event, 'driver' amplifier stages may be inserted between the phase splitter and the output valves.

A typical circuit is shown in Fig. 8.6, where $V_2$ and $V_3$ are 'drivers', and $V_1$ is a cathode follower phase splitter. As the signal currents in $V_2$ and $V_3$ are equal and of opposite phase, there is cancellation in $R_k$ which need not be bypassed. $R_D$ is a common decoupling resistor for the drivers, decoupled by the capacitor $C$.

# SUPPLY UNITS

THE voltage amplification and power output stages of a high fidelity amplifier require supply voltages for their operation. The heaters of the valves are normally operated at 6·3 volts. As the heater forms no part of the signal circuit, merely heating the cathode by radiation, the supply can be 6·3 volts a.c. which is made available by a suitable winding on the mains transformer.

## H.T. Supply

The h.t. supply required is d.c. and is usually between 250 and 550 volts. To obtain this supply from the a.c. mains, it is necessary to use a rectifier circuit, converting the a.c. supply to d.c. at the required voltage. The a.c. voltage applied to the rectifier will usually be different from the voltage of the mains supply and can be obtained by a further winding on the mains transformer. Some amplifiers are designed to work directly from either a.c. or d.c. mains, and a mains transformer cannot then be used. The omission of the mains transformer is not recommended, as the amplifier is then directly connected to one side of the mains supply and, unless the proper precautions are taken, there is always a danger of serious electrical shock. In any case, the h.t. supply voltage is limited to approximately that of the main supply.

The mains transformer therefore serves the double purpose of isolating the amplifier circuits from the mains and providing the exact h.t. supply voltage required.

## Rectification

Rectification is obtained by the use of a diode valve (or 'metal' rectifier) in which current flows only when the anode is positive with respect to the cathode. If a diode valve is connected in series with the load, and a large capacitor connected in parallel with the load and the circuit connected across an a.c. source, current will flow through the load and charge the capacitor during positive half

cycles of the a.c. supply. When the diode is not conducting, the capacitor will discharge through the load, so that the charge on the capacitor will drop. Whenever the charge is below the value of the a.c. voltage applied, the capacitor will be recharged and it will therefore receive a pulse of charging current in each cycle, corresponding to the positive peaks of the a.c. supply (see Fig. 9.1).
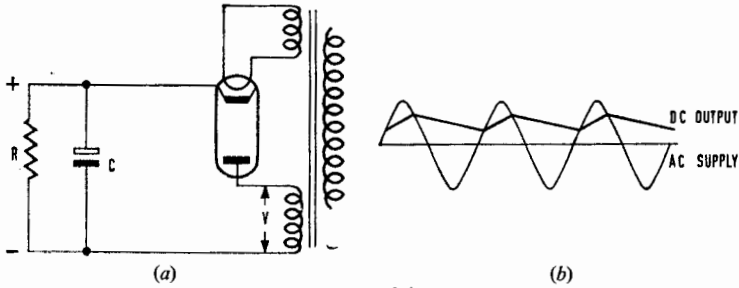


FIGURE 9.1

Indirectly heated diode, half wave rectifier circuit. The capacitor C charges on alternate half cycles and discharges through the load R. The ripple is 50 c/s.

If two diode circuits are used, and their supplies are connected out of phase, the outputs can be connected in parallel so that the
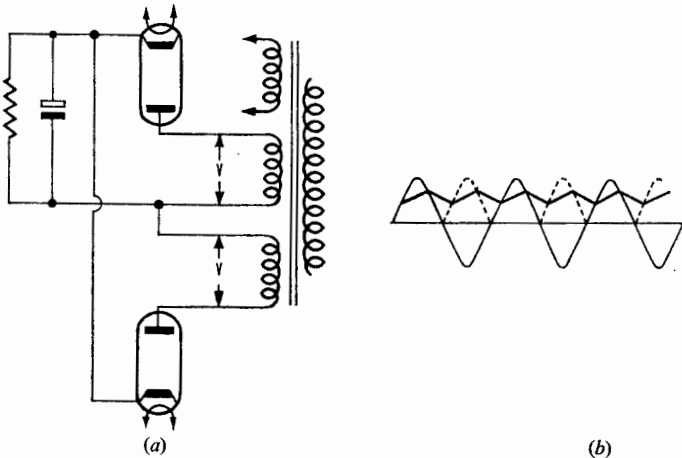


FIGURE 9.2

A second diode operating on alternate half cycles, giving full wave rectification. The ripple is 100 c/s.

capacitor will receive a charge for each half cycle of the alternating supply, i.e. there will be two charging pulses per cycle (see Fig. 9.2). This type of rectifier circuit, called a 'full wave' rectifier, is that most

commonly adopted, the two diodes being placed together in one valve and the cathodes internally connected (see Fig. 9.3). The two supplies are in the form of a single, centre-tapped winding on the mains transformer.

The rectifier valve is frequently made with a heater which requires 5 V a.c. This is obtained from a separate winding on the mains

FIGURE 9.3

The two diodes of Fig. 9.2 can be built into one envelope with connected cathodes, forming a full wave rectifier valve.
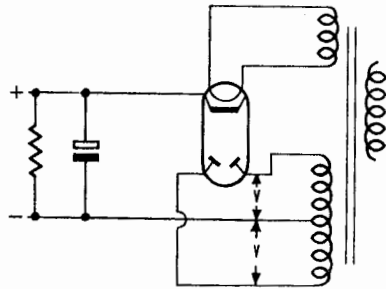
transformer. Some indirectly heated rectifiers require $6 \cdot 3$ V, in which case it is possible to use one heater winding for all the valves of the amplifier. As an indirectly heated cathode does not emit electrons until it has reached operating temperature, the h.t. supply will not be applied until all valve cathodes have been heated. Some rectifiers are made with a 'directly heated' filament, in which the electron-forming substance is deposited directly on to the heater. In this case the valve will commence to operate almost as soon as the supply to its heater is applied and, if the amplifier valves are of the indirectly heated type, they will initially draw no current, so that the h.t. supply will rise to the peak value of the rectifier a.c. supply. If this additional voltage will cause no damage to the amplifier circuits, then the effect can be ignored; otherwise some form of delay in applying the h.t. voltage is required.

In the bridge circuit, four rectifiers are required with an untapped h.t. supply. In the case of a voltage doubler, two diodes are used and the circuit arranged so that the reservoir capacitor receives a charging current from each diode in series, and the voltage output will be doubled, i.e. in the absence of a load, the voltage will be $2 \cdot 8 \times$ the a.c. supply voltage.

Diode valves in rectifier circuits can be replaced by rectifiers of the selenium or silicon type. These diodes require no heater supply and can be made small in size. This type of diode is equivalent to a valve without a heater, and its two connections are referred to as anode and cathode. They are frequently used in the form of a bridge circuit

and, due to their low impedance, form a very satisfactory h.t. supply unit (see Fig. 9.4).

Care must be taken not to exceed the rectifier maximum permitted peak anode current and peak inverse voltage. The maximum anode current is that which the cathode can supply without damage. This is normally taken to be about twice the d.c. current drawn from the h.t. supply.

The maximum peak inverse voltage (P.I.V.) is the maximum permitted negative voltage which can be applied to the anode and, in full-wave rectifier circuits, will be about three times the h.t. supply voltage. In bridge circuits, it will be about one and a half times this voltage.

### Smoothing

It has been mentioned that the reservoir receives a pulse of charging current during the period when the a.c. supply is at peak positive at the diode anode, the duration of the pulse depending upon the current drawn. In the case of a half-wave rectifier, this pulse will occur once in every cycle, i.e. 50 times a second when the a.c. supply is the 50 c/s house mains. This pulse will form a ripple on the d.c. supply which, if not removed, will be transferred to the signals appearing at the valve anodes. In a similar way a full-wave rectifier will receive two pulses a second, which in the case of 50 c/s mains, would give rise to a 100 c/s ripple. Ripple is removed by means of a decoupling circuit similar to that discussed for the voltage amplifier. In this case,
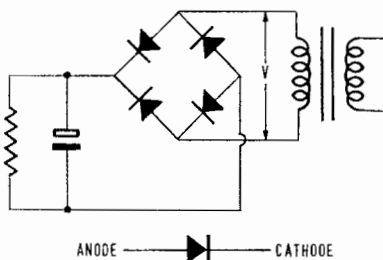


FIGURE 9.4

Full wave bridge circuit which requires four diodes, with a single transformer winding. Selenium or silicon diodes, which require no heater supply, are frequently used in this circuit.

ANODE ———▶︎|——— CATHODE

however, the current will be substantial and, if a resistor capacitor network were used, voltage drop due to the resistor would be high if the resistor was of an effective value.

This difficulty is avoided by using an inductor in place of the resistor. The d.c. resistance of the inductor can be quite low, whilst its de-coupling effect will depend upon its reactance at 50 or 100 c/s. The value of the decoupling capacitor can always be made as high

as conveniently possible, bearing in mind its working voltage and physical size (see Fig. 9.5).

The normal supply unit therefore comprises the mains transformer, rectifier valve and filter. The primary of the mains transformer may be tapped to correspond to various a.c. mains voltages, e.g. 200,

FIGURE 9.5

A practical full wave supply unit with capacitor input filter to remove ripple. The inductor ('choke') should be at least 5 H and of low d.c. resistance. The capacitors may be of the electrolytic type, 8 μF or more.
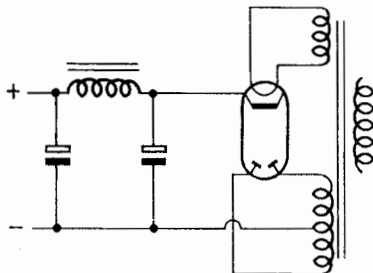


230, 240 and 250 V in Great Britain and 100–115 V in America. The secondary windings of this transformer will comprise a 6·3 V supply for the heaters of the voltage amplifier valves, a winding for the rectifier if a valve rectifier is used, and a high voltage winding suitable to the rectifier circuit employed. In the case of full-wave rectifiers, the high voltage winding is centre tapped and may be from 250–0–250 to 550–0–550.

The windings must be formed of wire of a gauge which will carry the required current without undue voltage drop and undue heating. The core of the transformer must also be composed of sufficient material to give efficient flux linkage and to avoid over-heating. A

FIGURE 9.6

Alternative choke input filter, which is lower in efficiency, but has improved voltage regulation. The choke must be a special 'swinging choke'.
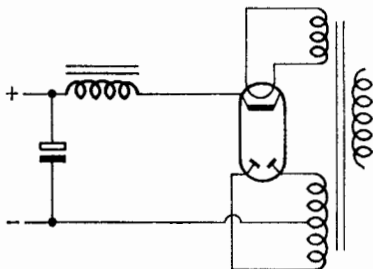


safe temperature for a transformer can be taken as 55° C above room temperature $(T_1)$ and can be calculated by measuring a winding resistance cold and after working on load for about five hours. If the cold resistance is $R_1$ and the hot resistor is $R_2$, then

$$\frac{R_2}{R_1} = \frac{T_2°C + 234·5}{T_1°C + 234·5}.$$

The diodes should be rated to supply the required current and also to withstand P.I.V. in the circuit employed. The reservoir capacitor must be rated to withstand the working voltage and also any higher voltage which may be present whilst valve heaters attain working temperature. The reservoir capacitor is also required to withstand a ripple current, which in the case of a half-wave rectifier is equal to twice the d.c. load current and, in full-wave circuits, the d.c. load current $\times$ 1·4. Smoothing capacitors should be of the largest capacitance economically possible and should be rated to withstand the working voltage. Reservoir and decoupling capacitors are normally
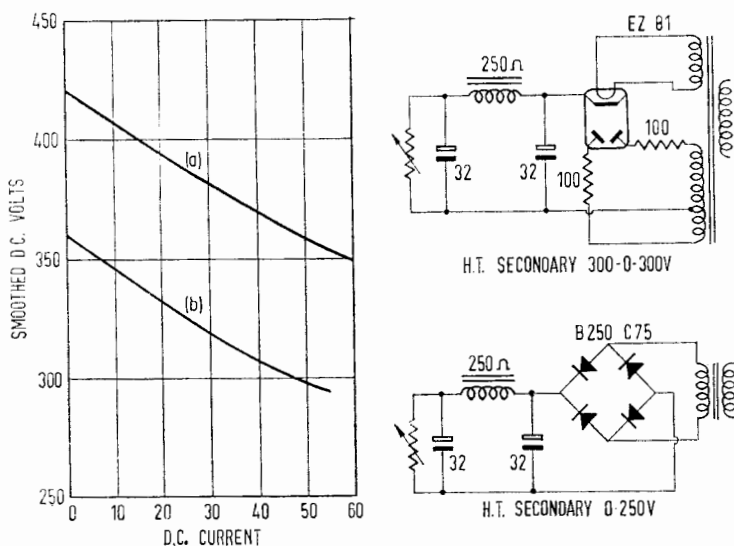


FIGURE 9.7

The regulation curves of two typical rectifier circuits. Circuit (a) uses a valve rectifier, and Circuit (b) uses a selenium bridge rectifier.

from 8 $\mu$F to 64 $\mu$F. The choke is required to have sufficient inductance at the d.c. working current to provide adequate smoothing and is normally in the range of 5 to 20 H.

A rectifier circuit has resistance and reactance which will cause a voltage drop which will increase as the current drawn increases. When no current is drawn, the reservoir capacitor will charge to approximately the peak value of the supply voltage, i.e. 1·4 $\times$ r.m.s. value. The way in which the output voltage drops with increasing current is called the regulation of the supply, so that units with low resistance will have better regulation than those with high resistance.

The output voltage/output current regulation of two typical power supply units is shown in Fig. 9.7, (a) employing a valve rectifier and (b) a selenium bridge rectifier. Fig. 9.8 is the case of a high voltage supply for a large amplifier using four separate silicon diode rectifiers with high P.I.V. rating.

**Choke Regulation**
The regulation can be improved by adopting the choke input filter circuit of Fig. 9.6. The choke is a special swinging choke, the inductance of which decreases as the current drawn increases. This has the effect of offsetting the drop in voltage, but the omission of the input capacitor considerably reduces the ratio of the output voltage to the supply voltage. In large amplifiers, where the current drawn by the output valves increases towards peak output, poor regulation
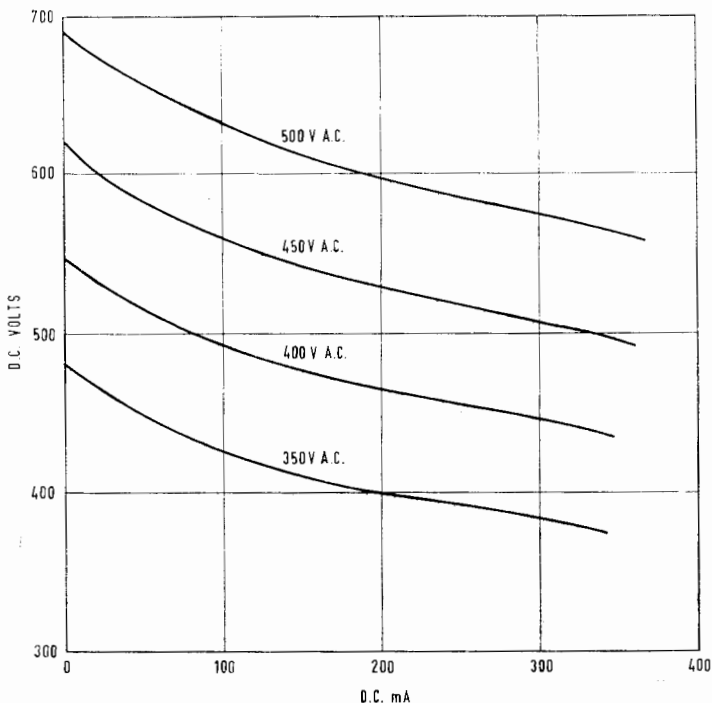


FIGURE 9.8

Regulation curves of silicon diode bridge rectifier circuit with 32 μF capacitor, for large amplifier.

of the h.t. supply would prevent maximum power being achieved. The use of the less efficient choke input filter circuit can be justified in such a case.

90

# DESIGN FOR A 3WATT AMPLIFIER

THE three basic sections of a high fidelity amplifier have been considered—the voltage amplifying stages, the output stage and the supply unit. These three sections are usually assembled together on one chassis to form a complete amplifier, although the voltage amplifying stages may be divided and the first stages built on to a separate chassis in the form of a control unit, sometimes called a pre-amplifier. The early stages often include frequency correction and tone and volume control circuits which are operated by means of selector switches and potentiometers, and it is an advantage to be able to place this unit close to the apparatus which will provide the input signals so that the controls are conveniently to hand.

The main amplifier is required to amplify its input signal and to deliver the power required to operate a loudspeaker system. These two matters are the first factors to be determined when drawing up the specification for an amplifier, other relevant factors being cost, size and standard of performance. The stages must be chosen and linked so as to produce an amplifier which will meet the given specification; cost and performance are obviously conflicting factors, the cost rising with standard of performance. An inexpensive amplifier may be small in size, and the size will increase with performance. The size may, however, be limited by the use of special components, but this will again affect the cost.

## Design Considerations

These factors result in a variety of amplifier designs as evidenced by the large number of commercially produced amplifiers available. The following general principles may be deduced from earlier chapters.

*Voltage Amplifiers.* Pentodes give the highest gain per stage and have the advantage of low input capacitance. The high valve impedance, shunted by the capacitance of the output circuit, will cause

a loss of high frequencies and will reduce the effectiveness of decoupling circuits. Pentodes introduce considerable distortion when required to provide a high output swing, i.e. more than about 20 V r.m.s.

Triodes will give only a moderate stage gain, but the valve impedance will assist in maintaining the high frequency response. There are a large number of triode valve types to choose from and the valve may be selected to suit a particular case. The higher the h.t. supply voltage available, the easier it will be to keep down distortion with high output, this applying equally to pentodes and triodes.

*Output Stages.* Pentodes and tetrodes are more efficient than triodes, i.e. will produce a greater power output for a given supply voltage and current drain. Thus if the h.t. supply is 250 V and 50 mA is drawn, i.e. $12 \cdot 5$ W, a single pentode output valve would produce about 6 W of audio power whereas a triode would produce only about 3 W. Pentodes and tetrodes will, however, introduce more distortion, which is mainly third harmonic.

The higher impedance of pentodes and tetrodes will provide little 'damping' of a loudspeaker (see Chapter 16).

Push-pull connection of two output valves will produce less distortion due to cancellation of even order harmonics, so that triodes will benefit most from this connection. Push-pull valves pass a standing current through the associated half of the primary of the output transformer, and one valve causes a magnetizing effect which is equal and opposite to, and hence cancels that of the other. The design of the output transformer is therefore eased. Push-pull, however, requires two out-of-phase, balanced inputs.

*Power Supply.* The larger the power output the greater will be the power required from the h.t. rectifier. The cost of the filter circuit components will increase as voltage and current requirements are increased. The power supply impedance should be as low as possible to give good regulation. Poor regulation will increase the possibility of coupling between stages and will prevent the output stage from developing full power.

When designing an amplifier, the general requirements are first set down and, using data from the application reports, a suitable circuit is developed. A detailed analysis of the performance of each stage can then be made and component values determined.

**Design Development**

The first consideration is the power required by the loudspeaker, which will be chosen according to the size of the room in which it is

used, and the sound level required. A loudspeaker system for a large auditorium may demand a power of 100 W or more from the amplifier. In a small room, where high sound level is not required, a high fidelity loudspeaker of high efficiency may not require more than 3 W from the amplifier. Output powers are rated as 'undistorted' although any amplifier will produce some distortion. Distortion is usually specified in terms of harmonics, and harmonics up to the 7th are the most significant. The highest total harmonic distortion which is acceptable is 5% and when distortion has been reduced to $0 \cdot 1\%$ it is negligible. Negative feedback can be employed to reduce distortion at the expense of gain, but excessive feedback may give rise to instability. Negative voltage feedback also reduces the apparent output impedance of the amplifier, a desirable feature when the load is a loudspeaker.

The next consideration is the input signal voltage which will be applied to the amplifier, and hence the voltage amplification required to enable the output stage to develop its maximum power output. A high quality microphone of the ribbon type may produce a signal voltage of the order of 100 $\mu$V, whilst a radio tuner might produce 1 V. When the amplifier is built in two sections, a control unit and a main amplifier, the control unit should accept inputs from microphone, gramophone pickup and radio tuner, and appropriately amplify each input signal, so that in each case about 1 V r.m.s. is delivered to the main amplifier.

As the simplest case, let us draw up a specification for an amplifier which will provide 3 W output power with total harmonic distortion not exceeding 5% from an input signal of 1 V r.m.s.

**Output Stage**

This is a low standard of performance, of the sort to be expected from a cheap amplifier. The output stage can therefore be a single pentode, for most efficient use of the power supply. As no available output pentode will produce 3 W from 1 V r.m.s. input, a voltage amplifier stage will be required.

Examination of the application reports suggest that the ECL86 would be suitable, producing 4 W of power at 10% distortion, with $3 \cdot 2$ V r.m.s. input and an h.t. supply of 250 V, the total current being about 42 mA. This valve includes a separate triode valve in the same envelope, which can be used as the voltage amplifier, with a gain of about 50. A gain of only 3 is required to load the output valve fully from 1 V, so that there is spare gain for the application of voltage feedback. A selenium bridge rectifier can be used, or

alternatively a rectifier valve such as the EZ80. The output transformer can be made to benefit from the feedback by including it in the feedback loop, and is required to present a load of 7 kΩ.

There will be a voltage drop at the anode due to the resistance of the output transformer primary, and the primary resistance should
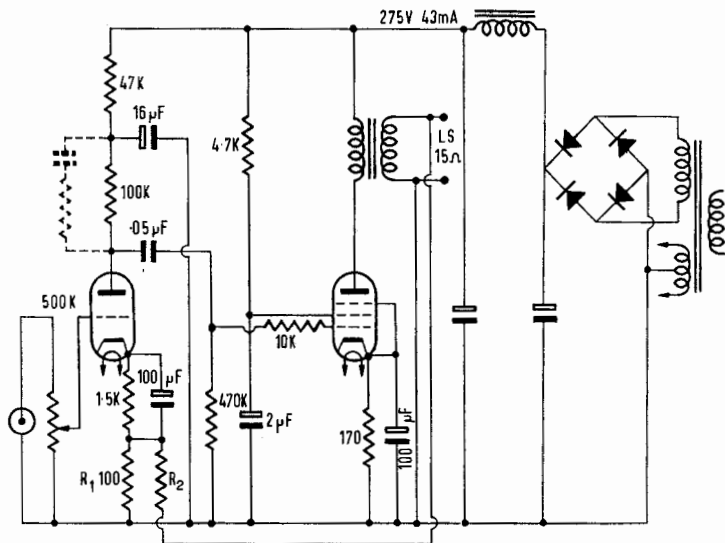
Complete circuit of ECL86 amplifier with selenium bridge h.t. rectifier. The h.t. supply voltage required is 275 V at 43 mA when the output transformer primary resistance is 500 ohms. Feedback is applied via $R_1R_2$ and when $R_2$ is 2·7 K ohms, the feedback factor is about 5 (14 dB). Four watts output requires an input of about 0·3 volts r.m.s. The series capacitor and resistor across the triode load resistor may be required to prevent high frequency instability due to phase shift.

therefore be as low as practicable. The voltage drop due to the primary must be allowed for when the h.t. supply voltage is being considered. A resistor in the screen circuit will drop the supply to 250 V, and should be decoupled. A typical output transformer may have a primary resistance of 500 Ω, so that the voltage drop is $\dfrac{500 \times 36}{1,000} =$ 18 V. Allowing 7 V for bias, the required h.t. supply is 275 V after the filter circuit. The screen draws 6 mA, so that the required value of dropping resistor is $\dfrac{25}{6} \times 1,000 = 4\cdot1$ kΩ, and a standard value of 4·7 kΩ would be suitable. The bypass capacitor should be not less than 2 μF. The bias resistor specified in the application report is

94

170 Ω. The maximum permitted grid to cathode resistor for the output valve is given in 'limiting valves' and is 1 MΩ.

## Voltage Amplifier

The grid resistance is effectively in parallel with the anode load of the voltage amplifier, reducing the gain and increasing the distortion of the stage. A high value will, however, increase distortion due to grid current when the output valve grid approaches zero volts, and a compromise of 470 kΩ has been taken. A suitable value for the coupling capacitor is $0 \cdot 05 \ \mu F$.

The output impedance of the triode valve and its load resistance will assist in 'damping' the effect of grid current of the output stage, so that a relatively low anode load for the triode is preferred. This will also help to maintain high frequency response, 100 kΩ being a reasonable value. Allowing 47 kΩ for the decoupling resistor, the d.c. load of the triode is 147 kΩ and the dynamic load about 80 kΩ. The performance of the triode stage can now be assessed, using the anode characteristic. (The triode is similar to one section of an ECC83 valve.)

As the expected stage gain is around 50, and the output stage requires an input of $3 \cdot 2$ V, the input to the triode will not exceed about $0 \cdot 06$ V r.m.s. The valve may therefore be operated close to the $-1$ V grid line.

Graphical analysis gives a working point of 165 V at the anode and a current of $0 \cdot 8$ mA, when the bias is $-1 \cdot 3$ V. The cathode resistor is $\dfrac{1 \cdot 3}{0 \cdot 8} \times 1{,}000 = 1 \cdot 6$ kΩ, which can conveniently be the standard values of $1 \cdot 5$ kΩ in series with 100 Ω, the latter resistor permitting series feedback to be applied from the output transformer secondary (see Fig. 10.1). The stage gain is then only slightly below 50.

## Frequency Response

The high frequency response of the triode stage depends upon the shunt capacitances which are:—

| | |
|---|---|
| $V1_{a-k}$ | $2 \cdot 5$ pF |
| $V2_{g-k, \ g2, \ g3}$ | $10 \cdot 0$ pF |
| $V2$ Miller effect | $21 \cdot 2$ pF |
| Strays | $10 \cdot 0$ pF |
| Total shunt | $43 \cdot 7$ pF |

The Miller effect for $V2$ is calculated from $C_{g-a}(1 + A)$. $C_{g-a}$ is $0\cdot4$ pF so that $C_{g-a}(1 + A)$ is $0\cdot4 \times 53 = 21\cdot2$ pF.

The response is 3 dB down when the reactance of the total shunt capacitance is numerically equal to $r_a$, $R_L$ and $R_G$ all in parallel (which is the same as $r_a$ in parallel with the dynamic load). At the chosen working point $r_a$ of $V1$ is about 70 kΩ and the dynamic load is 80 kΩ and $\dfrac{70 \text{ kΩ} \times 80 \text{ kΩ}}{70 \text{ kΩ} + 80 \text{ kΩ}} = 37$ kΩ. Assuming the shunt capacitance to be 50 pF, the table on page 164 shows that the reactance of 50 pF will be 37 kΩ between 70,000 and 100,000 c/s, which is satisfactory. The output valve produces 4 W with a load of 7 kΩ, and $E = \sqrt{WR} = 4 \times 7,000 = 168$ V r.m.s. The input for 4 W is $3\cdot2$ V r.m.s. so that $A \simeq 52$.

The frequency response of the output stage will depend upon the design of the output transformer, which is outside the scope of this work. It should be mentioned, however, that the output transformer is required to develop power in the secondary load (i.e. the loud-speaker) and in doing this it may introduce distortion which will be added to that introduced by the voltage amplifier and output stages. This can be kept sufficiently low in practice by using a large core constructed from high quality core material.

The response of the output valve and transformer can be assessed and the high frequency response will drop by $0\cdot707$ (i.e. $-3$ dB) when $Rp = r_a$.

The performance of the amplifier may be improved by the application of negative voltage feedback.

### Negative Feedback

It must be borne in mind that the feedback circuit is in parallel with the load and will thus draw power, but the loss will not be significant if the resistance of the feedback circuit is not less than 10 times the load resistance, i.e. 150 Ω when the load is 15 Ω. Taking the triode stage gain as being about 50, if the output valve requires an input of $3\cdot2$ V r.m.s. for 4 W output, the input to the amplifier will be $\dfrac{3\cdot2}{50}$ V r.m.s. or about $0\cdot06$ V. An output of 4 W will produce $\sqrt{4 \times 15}$ or $7\cdot75$ V r.m.s. when the load is 15 Ω, so that the over-all gain is $\dfrac{7\cdot75}{0\cdot004} \simeq 120$. If a moderate feedback factor of 5 is adopted, the proportion of the output required to be fed back can be calculated.

$$F = 1 + A\beta \text{ or } 5 = 1 + 120\,\beta, \text{ so that } \beta = \frac{1}{30}.$$

Now $\beta = \dfrac{R_1}{R_1 + R_2}$ and if $R_1$ is 100 $\Omega$, $R_2$ can be found to be 2·9 k$\Omega$, and 2·7 k$\Omega$ is the nearest 5% standard value. If the gain reduction factor is 5, the input for full output will be 0·32 V r.m.s. and the distortion will drop to approximately 2%. Adjustments to certain component values may be required when the amplifier is tested, but the specification for the amplifier has been theoretically exceeded and no difficulty should be experienced in meeting the specification in practice. An input volume control, in the form of a potentiometer, may precede the triode stage and should not exceed 500 k$\Omega$ to avoid high frequency loss due to the input capacitance of the valve. An input socket carries the signal to the volume control. The performance of an amplifier on test is described in Chapter 13.

CHAPTER 11

# DESIGN FOR A 10 WATT AMPLIFIER

THE single valve amplifier just described, whilst capable of producing acceptable quality of sound when used with a good quality, efficient loudspeaker, does not warrant the description high fidelity amplifier. For an adequate sound level in the home with a loudspeaker of average efficiency, an output power of about 10 W is required with low harmonic distortion content.

### Push-Pull Output Stage

A suitable output stage is formed by two EL84 valves in push-pull in ultralinear connection. When the screen tapping is at 43%, this stage will give an output of 11 W with a total harmonic distortion of 0·7% before feedback is applied. When moderate feedback is applied, i.e. a gain reduction factor of about 7, distortion can be reduced to 0·1%. The EL84 application report, page 203, shows that distortion increases gradually and smoothly up to maximum power.

It will also be observed that the current drawn by the stage is subject to little variation with power output. The performance as regards distortion and current variation can be compared with that illustrated on page 202 for a 20% screen tap connection. The h.t. supply line should be about 310 V at 80 mA and 40 mA of spare current can be allowed. The input required for full output is 16 V r.m.s. grid to grid, or 8 V r.m.s. per valve.

### Phase Splitter

For a good high frequency response, the preceding stage should have a low output impedance and, as the input voltage requirement of the output stage is relatively small, a valve with a high amplification factor can be used, the 12AT7 (ECC81) being suitable. Using the anode follower circuit, feedback can be applied from the output transformer secondary to the cathode of the input valve, the phase splitter valve having unity gain.

Using the anode characteristic for the ECC81, the load resistor can be chosen for a compromise between output and distortion, allowing 50 V drop in the decoupling resistor. As feedback is to be applied to the cathode resistor of the input valve, the resistor must be unbypassed. If the whole resistor were unbypassed, the output impedance of the stage would rise due to current feedback. A compromise can be struck by leaving a relatively small part of the resistor unbypassed. The value of the unbypassed section must be large enough to permit adequate feedback to be applied. The total resistance of the two feedback resistors is shunted across the output load, which will be 15 $\Omega$, and should be not less than 150 $\Omega$.

The 330 k$\Omega$ grid resistors, the maximum permitted for the EL84, are in shunt with the load resistors of the drivers, as are the two anode follower feedback resistors. The feedback resistors should therefore be of a high value, but the value should be limited to assist in the reduction of high frequency loss due to shunt capacitance. A good high frequency response will reduce phase shift and instability when feedback is applied.

The bass response depends upon the value of the coupling capacitors in relation to the grid resistors. A capacitor of $0 \cdot 05$ $\mu$F has a reactance of 318 k$\Omega$ at 10 c/s, and is a likely value to a first approximation.

The phase splitting stage is required to have unity gain, so that the feedback will be approximately 100%, and $R_4$ will be approximately equal to $R_5$. (see Fig. 11.2). As this stage has a high degree of feedback, its performance as regards frequency response and distortion will be considerably better than that of the input driver stage which feeds it. The output of the phase splitter valve will therefore closely correspond to that of the input stage, but with phase reversal. Again due to the high degree of feedback, the gain will be substantially independent of valve and supply voltage changes and hence will maintain balance over long periods. $C_1$ and $R_3$ must give a good low frequency response to the phase splitter.

A suitable load resistor for the 12AT7 stage is 47 k$\Omega$, which is shunted by the associated output valve grid resistor and the phase splitter feedback resistor. If these are both 330 k$\Omega$, the dynamic load to the stage is $\dfrac{47 \times 165}{47 + 165}$ k$\Omega$ = 37 k$\Omega$. The performance can be assessed by graphical analysis (see Fig. 11.1). After allowing 50 V for decoupling, the 47 k$\Omega$ load line can be drawn from 260 V to $5 \cdot 3$ mA. The maximum negative grid voltage to avoid grid current is $-1 \cdot 3$ V, and the required output is 8 V r.m.s. As the anticipated

gain is about 40, the input swing will be about $\frac{8}{40}$ V r.m.s. or $0\cdot28$ V peak, and the minimum bias is $-1\cdot6$ V. To allow a safety margin, $-2$ V is a reasonable choice at which the anode voltage will be 143 V and the anode current $2\cdot4$ mA. To provide $-2$ V bias at $2\cdot4$ mA,
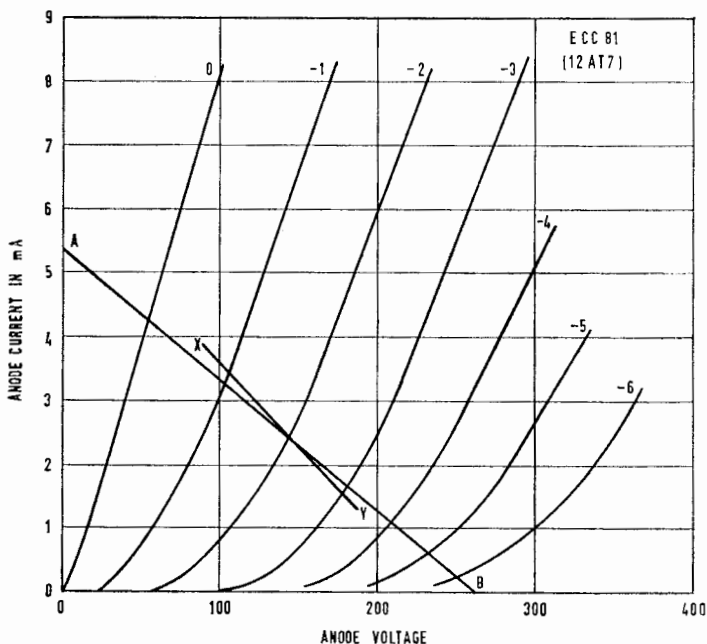


FIGURE 11.1

Anode characteristic of ECC83. Line A–B is for a load of 47 k$\Omega$, when supply is 260 V after decoupling. Dynamic load X–Y is 37 k$\Omega$, and the working point is $-2$ V at $2\cdot4$ mA.

the bias resistor will be $\frac{2}{2\cdot4}$ k$\Omega$ = 820 $\Omega$. The dynamic load shows that for an input of $+1$ V and $-1$ V the output will be $-37$ V and $+35$ V, a stage gain of 36 with about $1\cdot4\%$ distortion. The impedance of the valve at working point is $\frac{20}{1\cdot05}$ k$\Omega$ or 19 k$\Omega$, and the $\mu$ is 55. If 100 $\Omega$ is left unbypassed in the cathode circuit, the apparent output impedance will be $R_o = r_a + R_k (1 + \mu) = 19$ k$\Omega$ + 100 (56) or $24\cdot6$ k$\Omega$. The gain of the stage with current feedback is

$$A = \frac{\mu R_L}{r_a + R_L + R_K (1 + \mu)} = \frac{55 \times 37,000}{19,000 + 37,000 + 5,600} = 33.$$

100
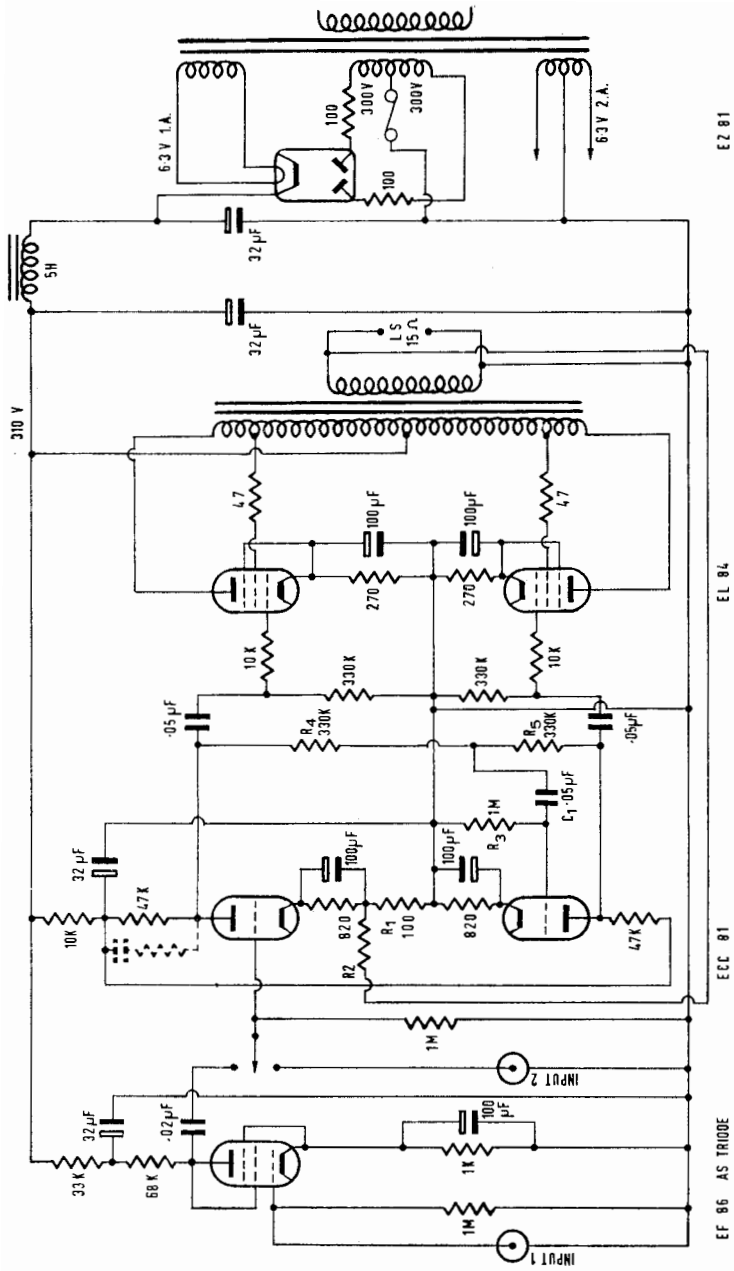
FIGURE 11.2

Complete circuit of high fidelity 10 watt amplifier with push-pull ultra-linear EL84 output stage. Feedback is applied via $R_1R_2$, and an output of 10 watts is obtained with a total harmonic distortion content of 0·1%. A series capacitor and resistor across the load of the input amplifier reduces phase shift and increases high frequency stability. A triode connected EF86 low noise input stage may be switched into circuit when required, to increase sensitivity.

101

The gain and distortion reduction factor is $F = 1 + A\beta$ or $1 + 36 \times \dfrac{100}{37,100} = 1 \cdot 1$. The gain with current feedback can be checked as $\dfrac{36}{1 \cdot 1} = 33$. The output stage delivers 11 W for 8 V r.m.s. per valve and 11 W output is therefore given by an input to the amplifier of $\dfrac{8}{33}$ or about $0 \cdot 25$ V r.m.s. The voltage corresponding to 11 W in 15 $\Omega$ is $\sqrt{WR}$, or $12 \cdot 85$ V r.m.s., so the overall gain is $\dfrac{12 \cdot 85}{0 \cdot 25} = 51 \cdot 5$.
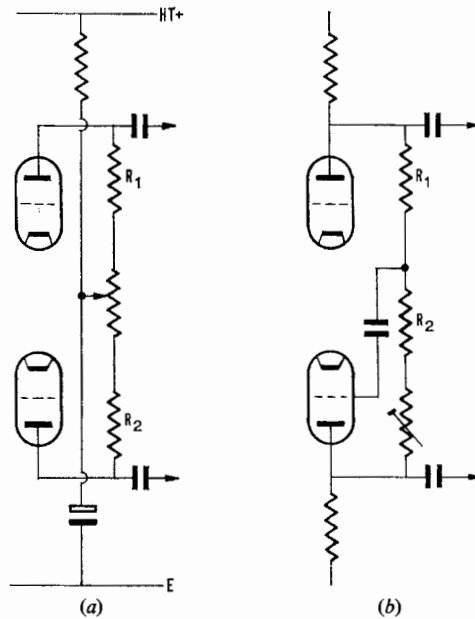


FIGURE 11.3

A push-pull amplifier may be balanced by load variation as in (a). This arrangement is applicable to the cathode coupled phase splitter circuit or to a pair of driver valves. In the case of the anode follower sphase splitter, a pre-set resistor enables the feedback to be reduced. thus increasing the gain of the anode follower stage.

## Negative Feedback

If feedback is applied so that the amplifier gain reduction factor is 7, $7 = 1 + A\beta$, i.e. $7 = 1 + 51 \cdot 5\beta$ and $\beta = \dfrac{6}{51 \cdot 5}$, about $\dfrac{1}{8 \cdot 5}$. If the gain reduction factor were made 10, $10 = 1 + 51 \cdot 5\beta$ and $\beta = \dfrac{9}{51 \cdot 5}$.

about $\frac{1}{5\cdot73}$. The unbypassed cathode resistor $R_1$ is 100 Ω, so that the feedback resistor $R_2$ can be determined, $\beta = \frac{100}{R_2 + 100}$ and when $\beta = \frac{1}{8\cdot5}$, $R_2 = 750$ Ω and when $\beta = \frac{1}{5\cdot73}$, $R_2 = 470$ Ω. In the latter case, $100 + R_2$ is 570 Ω which will draw little output power from the 15 Ω load. (See Fig. 11.2.)

The anode follower stage feedback resistors have been assumed to be 330 kΩ each and, as the stage gain is relatively high, the gain will be close to unity.
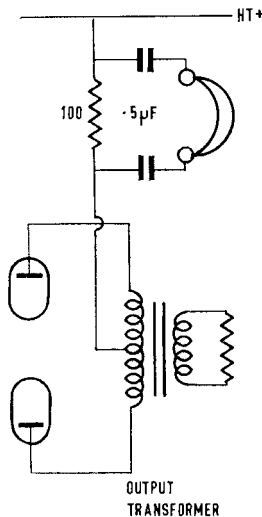
FIGURE 11.4

In both cases shown in Fig. 11.3, $R_1$ and $R_2$ are equal. Adjustment may be made for measured equality of the outputs; alternatively any unbalance in the output stage may be corrected by adjustment for minimum signal in the output transformer centre tap, using the temporary testing circuit shown here.

OUTPUT
TRANSFORMER

With a stage gain of 36 and 100% feedback, the gain with feedback is $\frac{36}{37}$ or 0·973. The resistor in the grid circuit of the anode follower stage will also affect the feedback factor and for these two reasons an adjustment to the feedback resistor is required for accurate balance, but calculation of the value is rather involved.

**Balancing**

In any event, for accurate balance close tolerance resistors would be required, so that determination of the value of the required resistor by measurement is preferred. Using two nominal 330 kΩ resistors, a 25 kΩ pre-set variable resistor is placed in series with the feedback resistor. The variable section can then be adjusted for balanced

output at the grids of the output valves, measured by means of a valve voltmeter, with a signal of 1,000 c/s applied at the input (see Fig. 11.3). It is possible to balance the whole of the push-pull amplifier including output valves and output transformer by this means, if a resistor of 100 $\Omega$ is temporarily connected in the h.t. supply to the output transformer centre tap, and a sensitive valve voltmeter or a headphone is connected across this resistor via two isolating capacitors. When a 1,000 c/s signal is applied to the amplifier, signal currents in the resistor should be equal and opposite so that the variable resistor is adjusted for minimum signal. If the amplifier is balanced in this way, adjustment may be required from time to time to compensate for change in output valve characteristics.

### Voltage Amplifier

When the gain reduction factor is 7, the input for maximum output is 1·75 V r.m.s., which is too high for some purposes. If required, a further stage of amplification can be added and can be a low noise stage such as a triode-connected EF86. The gain of this stage is about 25, so that the input for maximum output becomes about 0·07 V (70 mV) r.m.s. Sensitivity could be further increased by using the valve pentode connected, when the stage gain could be made to approach 200, giving a sensitivity of about 10 mV. In the demonstration amplifier, the first stage is triode connected and can be brought into circuit as required by means of a switch.

# DESIGN FOR A 30 WATT AMPLIFIER

FOR large rooms a maximum undistorted power of 25 W or more may be required and the output stage may then be push-pull EL34, KT66 or KT88 valves. These valves require an h.t. supply of 350 to 500 V and draw a current of 120 to 200 mA. A suitable power supply stage is required and rectifier valves such as GZ34 or U54 can be used. Under certain operating conditions it will be noticed that the current drawn by the output stage rises as the input increases. Unless the regulation of the power supply is good, the supply will drop with rise of current, so that the output stage will not develop maximum power. In such a case it is necessary to improve the power supply regulation and one method of doing this is to use a 'choke regulated' or 'choke input filter' circuit.

**Ultralinear Output Stage**
The application reports show that push-pull KT88 valves in 40% ultralinear connection with a supply voltage of 375 V, draw a current of 174 mA, which rises to 192 mA at maximum signal. This rise is not excessive, so that a full-wave rectifier with capacitor input filter is suitable .The stage requires an input of 71 V peak grid to grid, i.e. $35.5 \times 0.707$ or about 25 V r.m.s. per valve, for 30 W output at 1% total harmonic distortion.

The required input of $35.5$ V peak represents a relatively high output from the driver and this stage must be carefully designed to avoid distortion. The grid resistor of the output valve is in parallel with the load resistor of the driver, as regards a.c., and reduces the output. The grid resistor must therefore be as high as can be permitted, but it must be borne in mind that output valves of the type under consideration pass a higher grid current than valves of the EL84 type previously considered. When the total of the anode and screen power exceeds 35 W, as it does in the present case, the grid

resistor is limited by the valve manufacturers to 270 kΩ, and 220 kΩ is a suitable standard resistor.

### Driver Stage

A low output impedance for the driver valve will help to reduce the effects of grid current in the output stage and low impedance drivers can generally be made to give high output voltage without excessive distortion. Whilst the 12AT7 has a low output impedance, distortion is fairly high for large output voltages and, in any event, even if the driver stage gain were in the region of 40, the driver input would be nearly 1 V peak without the application of negative feedback to the amplifier.

The lower impedance valve 12AU7 is more suitable in this case, having an output impedance of about 10 kΩ under usual operating conditions and giving a stage gain of about 15. An additional stage will be required so that the amplifier input with feedback applied will be in the region of $\frac{1}{2}$ V r.m.s., and this stage can be brought within the feedback loop. In this case the lower output impedance of the 12AU7, and its relatively low stage gain, are an advantage as the high frequency response will be good and phase shift will be small at working frequencies. The 12AU7 driver stages can therefore be straight voltage amplifiers, and the phase splitter stage can feed the drivers, the driver input for full output with a stage gain of 15 being about $\frac{25}{15}$ or $1 \cdot 7$ V r.m.s.

### Phase Splitter

This is within the capabilities of any form of phase splitter, and the cathode follower phase splitter is frequently used in a circuit of this kind. This phase splitter operates with a cathode voltage in the region of 80–100 V, and has a gain of slightly less than unity from input to each output.

### Voltage Amplifier

A further amplifier stage will therefore be required to provide additional gain. If the operating anode voltage of this valve can be arranged to be equal to the cathode voltage of the phase splitter, the anode of the amplifier may be direct coupled to the grid of the phase splitter, which will be self-regulating as to its operating point. The direct coupling of these stages obviates one resistance-capacitance coupling and hence no low frequency phase shift is introduced here.

106
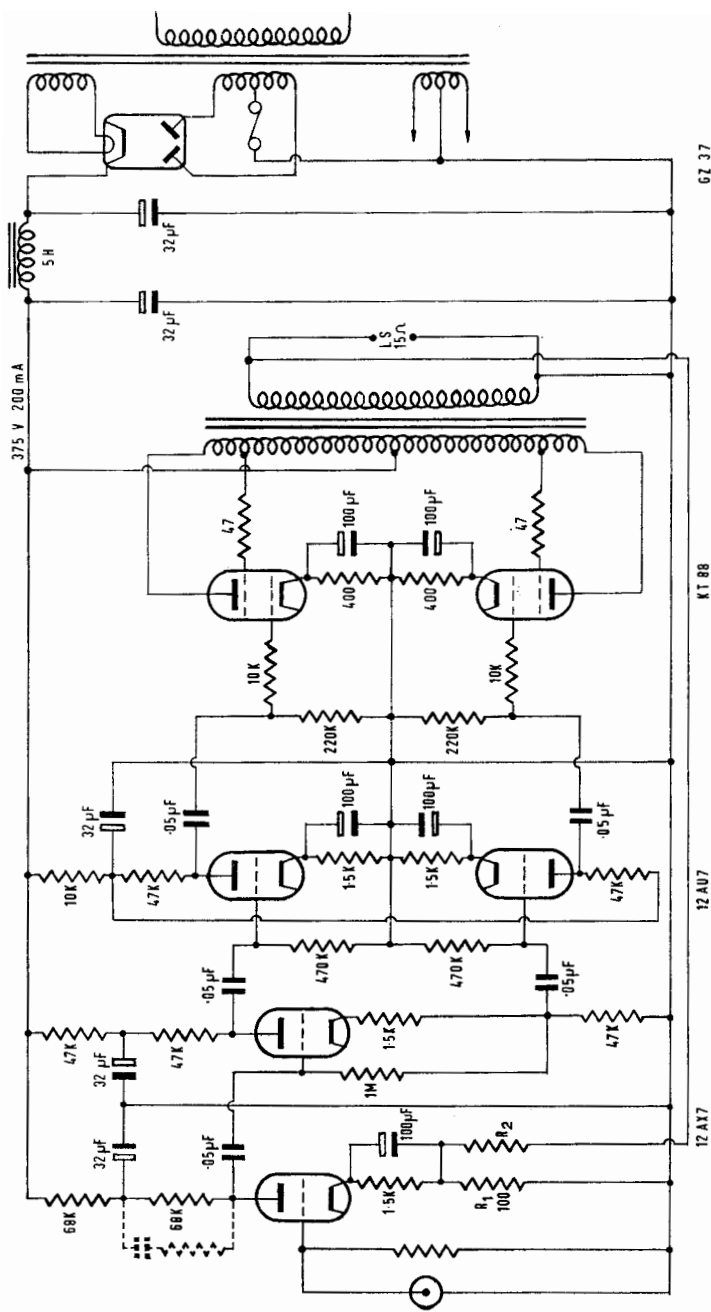
FIGURE 12.1

Complete circuit of high fidelity 30 watt amplifier with push-pull ultra-linear KT88 valves in the output stage. This stage requires 71 volts peak, grid-to-grid for 30 watts output, so that a pair of low impedance driver valves are inserted between the cathode follower phase splitter and the output stage. Negative feedback is applied via $R_1 R_2$. The output transformer should be of the highest quality to avoid introducing distortion and to reduce phase shift. The usual precautions against instability should be taken to ensure that the amplifier has a sufficient stability margin.

107

A second 12AU7 double-triode valve can be used in this position, and the resulting circuit is often referred to as the Williamson circuit.

With the best modern output transformers, the direct coupled circuit is not essential as a normal resistance-capacitance coupling can be made to operate with little phase shift down to a very low frequency, i.e. 1 c/s, due to the high input impedance of this phase splitter. Further, the phase splitter has a very low input capacitance so that a low output impedance for the first amplifier stage is not of prime importance. A valve such as the 12AX7 can therefore be used, with its high gain and low distortion, resulting in a more sensitive amplifier.

Further variations are possible using a 12AT7 anode follower phase splitter, similar to that described in the 10 W amplifier of Chapter 11, or alternatively a cathode coupled phase splitter.

### Negative Feedback

Negative voltage feedback can be applied from the output transformer secondary to the first amplifier valve cathode and, if the gain reduction factor is made to be 10, the amplifier will deliver 30 W with a total harmonic distortion of $0 \cdot 1 \%$, distortion at lower power outputs being less than this figure. This standard of performance can be regarded as virtually perfection. Stability and transient response should be thoroughly investigated and, if circuit changes can improve the performance in these respects, such improvements would have a more noticeable effect on sound output than would some increase in distortion over the figure quoted above.

The need to ensure that the output transformer is of high quality is again emphasized and, in an amplifier of this type, considerable care in design must be taken to avoid unwanted effects. The primary inductance must be high enough to give a good low frequency response and the core must be adequate to keep down distortion at low frequencies. High inductance and high power rating imply a large primary winding on a substantial core, so that the transformer will be large. This increases the problem of keeping down leakage inductance and stray capacitance, which affect the high frequency performance, so that special core material, such as C core, can be justified as this reduces core size and hence the amount of wire in the windings.

Various other devices are employed by transformer manufacturers to improve performance, and figures for primary inductance (high), primary resistance (low), self-capacitance (low), leakage inductance (low) and distortion at maximum rated output (low) should be

examined. A useful indication of the suitability of the transformer for a feedback amplifier is given by noting the frequency at which the self-capacitance combines with the leakage inductance to form the secondary 'resonance'. This frequency should be as high as possible, preferably 100 kc/s or more. (Ref. 1. Page 159).

Feedback can be applied to the unbypassed cathode resistor of the first valve, but this cathode resistor can be in two sections, as described for the case of the 10 W amplifier. The value of the resistors for the required feedback factor is calculated by using the formula $F = 1 + A\beta$. A 'step circuit' across the anode load of the first stage will almost certainly be required to reduce phase shift at high frequencies (see Chapter 13).

The output valves and the rectifier valves operate at a high bulb temperature and must be adequately spaced from each other and mounted on the chassis in a position where there is good ventilation.

# TESTING AMPLIFIERS

THE ultimate test of high fidelity equipment is whether it can reproduce sound so faithfully that the ear cannot distinguish between the original and the copy. In the home, the number of cases where this is possible is very small. The normal loudspeaker system radiates sound in a certain directional pattern, which will correspond with that of a very limited number of original sound sources. The sound might consist of two sources, such as piano and violin, and for accurate reproduction two amplifier channels and two suitable loudspeaker systems might be devised. As the number of sound sources increases, the complexity of such a system would increase and quickly become impractical. Whilst two or three performers might be present in an average room in the home, a small orchestra could not be accommodated, still less a full symphony orchestra. At the outset, therefore, it is clear that completely faithful reproduction of the original sound is not possible in the ordinary way, and all that can be achieved is a satisfactory compromise. The use of two channels in stereophony is an attempt to overcome some of these difficulties.

There is no standard by which a compromise can be judged, as this is largely a matter of personal preference; the aim of high fidelity reproducing equipment is to produce sound which gives a satisfactory impression of the original. So far as the high fidelity amplifier is concerned, any audible distortion it introduces will render the result less satisfactory. Theoretically, the amplifier should provide equal amplification for sounds of all frequencies from 20 to 20,000 c/s.

It has been stated that speech and music may be considered as being represented by a large number of sine waves added together. The tonal quality of different instruments sounding a note of the same fundamental pitch, say middle A, 440 c/s, is due to the presence of 'overtones' or 'harmonics' which are whole number multiples of the fundamental. Thus the twentieth harmonic of A (440) is 8,800 c/s.

If a very large number of harmonics are added in the correct proportion to the fundamental, the resultant wave shape becomes a square and in this case, theoretically, the rise time from zero to maximum is zero. A good amplifier should be able to reproduce such a waveform with a high degree of fidelity, the rise and fall occupying a very small period of time and the top of the square wave being flat and parallel with the base line. As the square wave contains very high order harmonics, which affect the 'squareness' of the corners, an extended high frequency response is required for faithful reproduction of this shape and much useful information can be obtained by square wave testing.

### Test Equipment

The essential equipment required to test an amplifier is a sine and square wave audio frequency generator, an oscilloscope for visual examination of the reproduced waveforms, and an a.c. voltmeter to measure power output and frequency response. The voltmeter should be accurate over the required frequency range, and a valve voltmeter is the most satisfactory type. A device for measurement of distortion is also very desirable. A high resistance D.C. voltmeter is used to check the h.t. supply voltage and the voltages at the valve anodes, screens and cathodes.

### Power Output Measurement

Taking the case of the simple amplifier described in Chapter 10, the first test is to ascertain whether the power output anticipated is obtainable. The amplifier, with its feedback line disconnected, is loaded by means of a resistor equal to the nominal loudspeaker impedance and rated to carry the power output. A 15 $\Omega$, 10 W resistor is suitable, across which is connected the voltmeter and oscilloscope. (Appendix 1 gives a table of power output related to r.m.s. voltage when the load is a 15 $\Omega$ resistor.)

The signal generator is set to 1,000 c/s and connected to the input. The calculated input for maximum output is 0·06 V so that the input can be set at 0·01 V and gradually increased, the power output and wave shapes being noted. If a distortion meter is available, the distortion can be measured at various power outputs. The distortion may also be observed by the departure of the waveform shape from a pure sine wave as the power output approaches maximum. (See Fig. 13.2.)

The power output at low and high frequencies, say 50 c/s and 10,000 c/s, should also be ascertained. The performance at these

111

frequencies should be equal to that at 1,000 c/s, and will be so in the case of the best high fidelity amplifiers.

### Frequency Response Test

Next the frequency response may be measured at a nominal output less than maximum. When the load resistor is 15 $\Omega$, 4 V, is a convenient reference, corresponding to about 1 W. The input at 1,000 c/s is adjusted until the voltmeter reads 4 V, and the input frequency is then raised until a frequency is reached at which the output voltage is reduced by $0 \cdot 707$, i.e. to about $2 \cdot 8$ V; the response is then $-3$ dB. When the output is halved, 2 V, the response is $-6$ dB, and it drops a further 6 dB each time the output reading is halved.

A similar test is then carried out at low frequencies and, from the results, a frequency curve may be drawn. The frequency response should then be checked at maximum rated output, when the result should closely approximate to that obtained at 1 W over the range 20 c/s to 20 kc/s.

### Feedback Measurements

The effect of feedback can now be measured. With the signal generator at 1,000 c/s and the output at 4 V, the feedback line is reconnected. The drop in output is noted, and the gain reduction factor is obtained by dividing 4 by this voltage. This factor may be expressed in decibels if required, and is $20 \log_{10} \left( \dfrac{4}{\text{new reading}} \right)$.

The power output can also be checked and a frequency response taken.

The application of feedback will introduce the possibility of the amplifier becoming unstable owing to phase shift. It should be ascertained that the amplifier is stable without feedback, as if it is not, faulty smoothing or decoupling is indicated. The amplifier may be stabilized at low frequencies by using an improved output transformer, reducing feedback, or altering the values of the coupling capacitors. In some cases it may be necessary to use a 'step circuit' consisting of a capacitor and resistor in parallel, in series with a grid coupling capacitor. In this case the capacitor can be about one-tenth the value of the coupling capacitor and the value of the resistor found by trial and error.

High frequency stability can be improved by means of a step circuit, consisting of a resistor in series with a capacitor connected across the load resistor of the first voltage amplifier within the feedback loop. The value of this resistor will usually be about one-tenth of that of the load resistor, and capacitors in steps of 50 pF from 50 pF to 350 pF can be tried to obtain best results.

112

## Square Wave Test

The square wave performance can be observed by injecting a 1,000 c/s square wave so that the a.c.-voltmeter reads about 4 V, deviation from a 'flat top' showing a loss or increase in low frequency response and rounding of the corners showing a loss of high frequency response. (The waveforms of Fig. 13.4 show the improvement in response brought about by negative feedback.) The square-wave test shows up overshoot and ringing, due to the output transformer, apparent as a 'pip' on the leading edge of the square wave, followed by a damped sine wave, which in some cases may persist along the whole of the top.
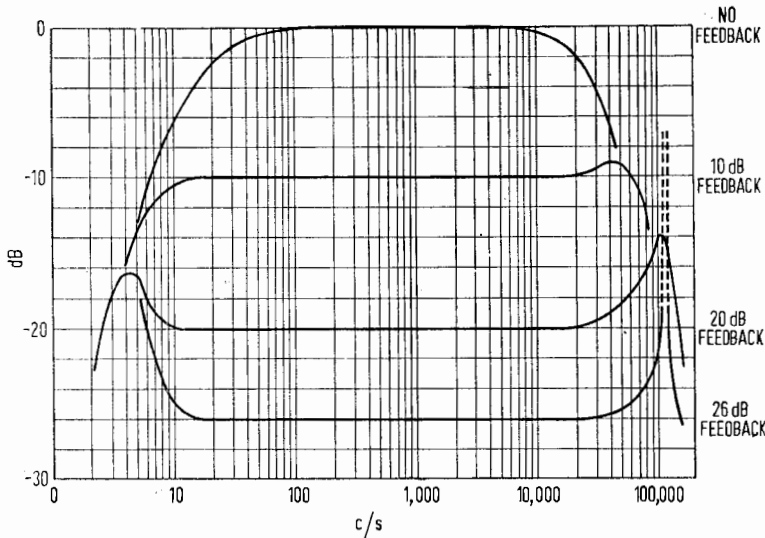


FIGURE 13.1

Frequency response of 10 watt amplifier with no negative feedback, and feedback of 10, 20 and 26 dB. The 10 dB feedback response can be compared with the theoretical requirement of Fig. 7.1. As feedback is increased, 'humps' appear at the extremes due to phase shift and instability may result, as in the case for 26 dB feedback where oscillation occurs at 135 Kc/s.

It is instructive to increase the feedback temporarily beyond that selected for normal operation, by reducing the feedback resistor until instability occurs (see Fig. 13.1). Any modifications required for stability can then be made and the feedback further increased. The increase in feedback before instability occurs expressed in decibels, is sometimes quoted as the 'stability margin' and a margin of 10 dB or greater is normally accepted as satisfactory.

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113

When any necessary changes have been made, the frequency response and power output should be checked to ensure that the performance is still satisfactory.

### Dummy Loudspeaker Load

The high fidelity amplifier is required to operate a loudspeaker system, whereas the tests have been carried out with a resistor load.

(a)



(b)



(c)
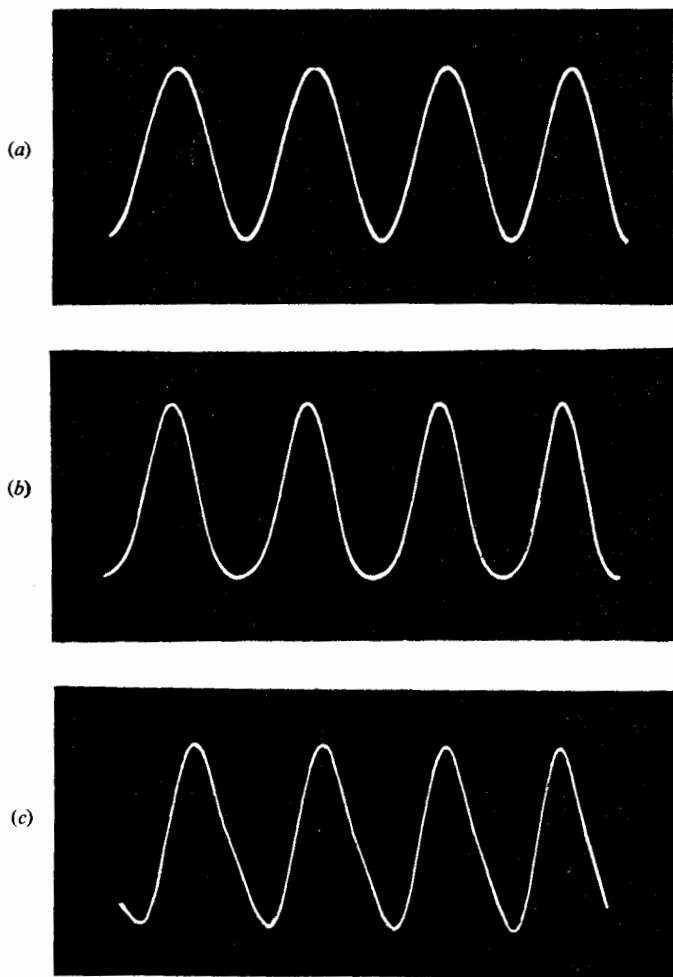


FIGURE 13.2

(a) 1,000 c/s sine wave test signal.
(b) Sine wave with the addition of 20 % second harmonic.
(c) Sine wave with 20 % second harmonic with 90° phase shift.

114

(a)



(b)
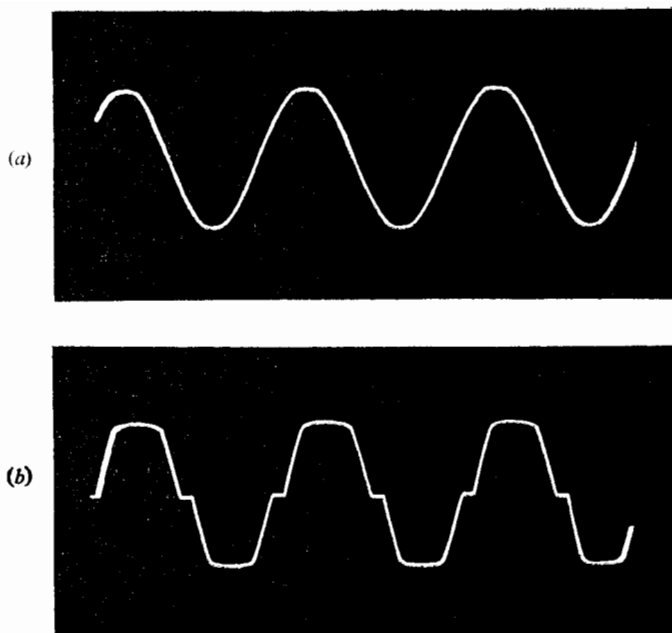
FIGURE 13.3

(a) Sine wave output of 10 watt amplifier at 10 watts showing onset of distortion.

(b) Sine wave output of 10 watt amplifier at 11 watts showing considerable distortion.

A 'dummy load' roughly equivalent to a loudspeaker should now be used to replace the resistor, and the 1,000 c/s square-wave test repeated (see Fig. 16.1). Some further adjustment may be found necessary to ensure satisfactory wave shape and stability with this load.

The amplifier with its input short-circuited or the input volume control set at minimum may now be connected to a loudspeaker. There should be no audible sound from the loudspeaker, when listening in a quiet room at a distance of 1 foot. Three types of sound may be heard:

*Hum at* 50 *and* 100 c/s, due to inadequate smoothing or decoupling, faulty heater to cathode insulation in a valve or hum pickup from the mains transformer or outside source.

*Hiss*, which may be due to resistor or valve noise.

*Intermittent crackling*, due to a faulty component or connection. (Chapter 14 may be referred to for notes on hum and noise.) If the amplifier produces no noise when the input control is at minimum,

115
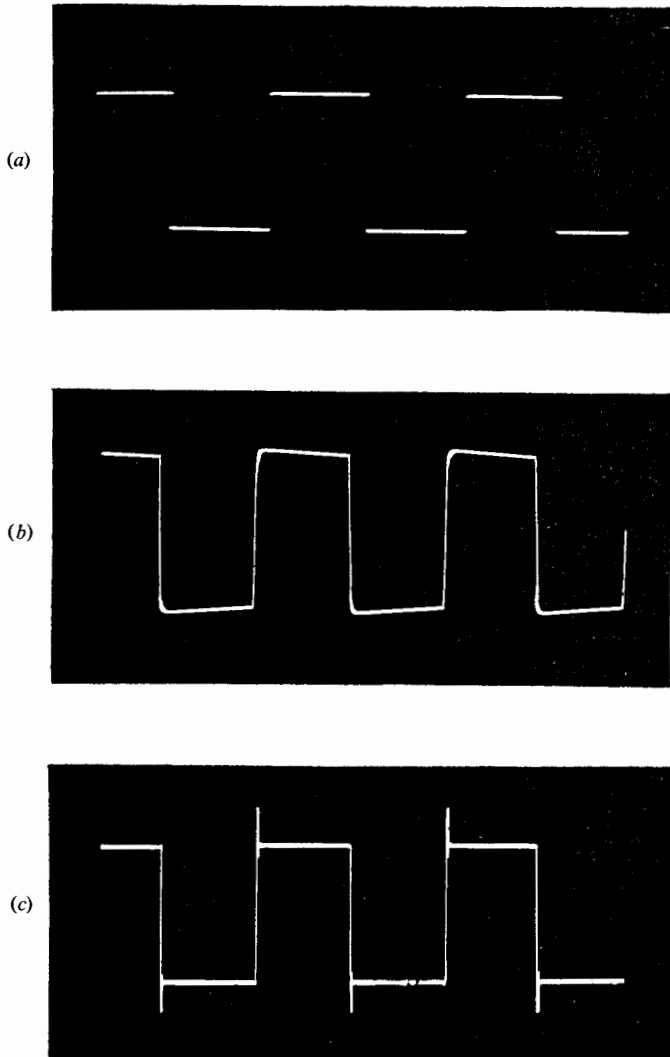
FIGURE 13.4

(a) 1,000 c/s square wave test signal.

(b) Square wave output of 10 watt amplifier without feedback. The slope of the horizontal lines indicates limitation of bass response and the rounding of the leading corners, limitation of treble response. The appearance of the vertical lines, hardly visible in (a), indicates that the amplifier does not immediately respond to sudden signal changes, i.e. transients.

(c) Square wave output of 10 watt amplifier with 15 dB feedback, showing improvement in bass and transient response. The high frequency response is improved but there is overshoot at the leading corner with slight ringing, due mainly to the output transformer.

116

the control should be increased to maximum without input connection, but with the input connector adequately screened. In sensitive amplifiers, some hiss due to the first voltage amplifier stage may be apparent, and, if the a.c. voltmeter is sufficiently sensitive, the noise output may be measured. An output of hum and hiss 80 dB below maximum output is virtually inaudible in a quiet room in the average home.

## Soak Test
The amplifier should be left in operation for five hours, and at the end of this time, switched off. All components, except valves, should be not more than comfortably warm to the touch. Resistors operating close to rating may become quite hot, but may continue to operate for a considerable period notwithstanding. If maximum reliability is required, components showing undue temperature rise should be exchanged for higher ratings.

## Oscilloscope Displays
Figs. 13.2–4 show typical oscilloscope waveforms encountered in amplifier testing.

# AMPLIFICATION FROM LOW VOLTAGE SOURCES

THE main amplifier circuits discussed have been considered on the basis that the input signal will be in the region of 1 V r.m.s. A number of sources will provide a signal which is less than 1 V, the lowest signal input met in practice being that from a ribbon microphone, which is normally matched to a low impedance line terminated by a step-up transformer. Here the output may not exceed 1 mV. To amplify this signal to 1 V requires a gain of 1,000. This is within the capability of two valve stages, which can readily be made to give individual stage gains of 33·3.

## Noise

When the signal is small it is difficult to keep the noise produced by the amplifier to a value which will give an acceptable signal-to-noise ratio. A satisfactory signal-to-noise ratio may be taken to be 60 dB (i.e. the noise signal is equal to one-thousandth of the signal voltage), and on this basis, if the signal is 1 mV, the maximum permitted noise figure is 1 $\mu$V.

Noise originates from a number of sources, which include thermal agitation in resistors and valve noise. The additional electrodes in pentodes cause 'partition' noise, so that a triode is usually preferred for the first stage of a low noise amplifier. If the gain of the first stage exceeds 20 times, the signal-to-noise ratio of the whole amplifier will be virtually that of this stage, so that this stage alone needs special consideration.

In this stage, all resistors associated with signal circuits should be special low noise types, generously rated when carrying current. The valve current and the anode load should be kept as low as possible, consistent with a satisfactory performance, and the grid resistor value selected to suit the input source. The noise produced by individual valves of the same type can vary and in some cases it may be necessary to select a low noise sample from a batch. Under normal working conditions the EF86 valve, pentode connected and used with an

anode load of 100 kΩ, will produce in its output a noise voltage equivalent to an input of 2 μV when the supply voltage is 250 V.

## A.C. Hum

When the valve heater is operated from an a.c. supply, hum will appear in the cathode. This may be reduced, however, by returning the heater supply to earth via the slider of a potentiometer connected across the heater supply and adjusted for minimum hum. (See Fig.



6·3 V WINDING ON
MAINS
TRANSFORMER

(a)

(b)
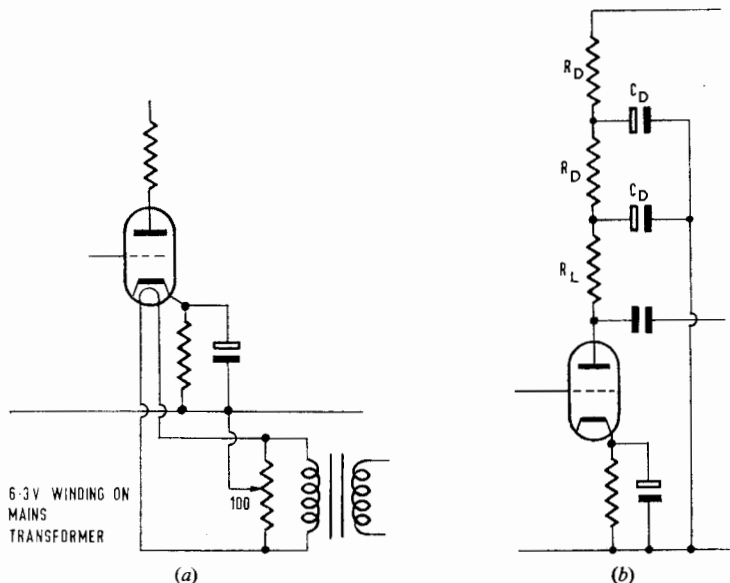
FIGURE 14.1

(a) A potentiometer connected across the heater a.c. supply, with slider connected to earth, can be used to balance the hum signal induced into the valve cathode.

(b) Improved filtering of ripple from the h.t. supply by double decoupling, i.e. the use of two $R_D C_D$ filter sections. When a very low noise level is required, the decoupling capacitor $C_D$ connected to the load resistor $R_L$ is a paper type rather than electrolytic.

14.1a.) A good quality valve-holder is required, as leakage and capacitive coupling between the valve pins will introduce considerable hum. Hum may be introduced into the stage in other ways and particularly through an inadequately decoupled h.t. supply. Considerable filtering may be needed to bring hum from this cause down to a level which permits the signal-to-noise ratio of 60 dB to be obtained, and double decoupling is often used. (See Fig. 14.1b)

119
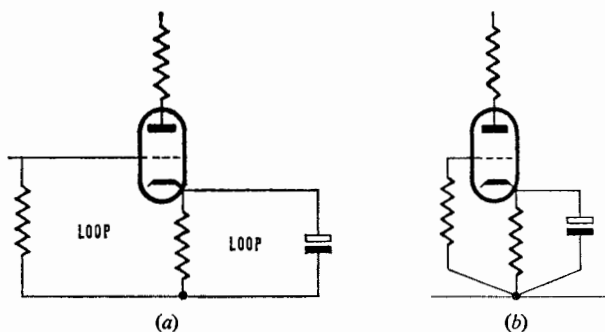
FIGURE 14.2

(a) Earth loops in the signal circuits of a sensitive amplifier stage into which hum can be induced by a mains transformer.

(b) Such loops can be avoided by using a single earthing point and placing components together to close the loop.

Hum may also arise if the valve is operated in the magnetic field of a mains transformer. The grid circuit of the valve is sensitive to electrostatic hum pickup which can be prevented by adequate metal screening.

It should be remembered that a microphone transformer connected to the grid circuit is particularly sensitive to hum pickup from the mains transformer due to magnetic linkage, and high quality magnetic screening (e.g. mu-metal) and orientation of the core may
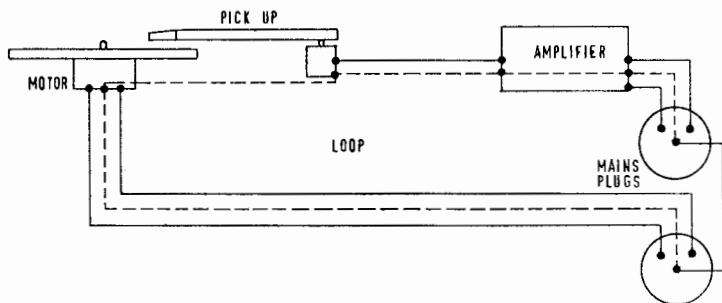


FIGURE 14.3

A loop caused by indiscriminate earthing may not always be immediately apparent. In the above example the loop may be broken by omitting the earth connection between the gramophone pickup and the motor.

be necessary. Whenever possible, a microphone transformer should be at least 3 feet from a mains transformer.

One other source of hum in sensitive stages which requires special consideration is that of earth loops formed by the wiring and circuit

120

components, into which the mains transformer can induce a 50 c/s signal. Examination of Fig. 14.2 shows that a number of loops formed by earth returns are possible. These can be minimized by earthing to a common point, preferably on the valve-holder, and by keeping components forming loops close together and parallel to each other. A loop can be formed if insufficient care is taken in the connection of screened leads, particularly in the grid circuit, and by the haphazard provision of external earth connections (see Fig. 14.3).

Where absolute freedom from hum is necessary, the first stage can be built into a separate, completely screened chassis well away from a.c. fields, and the valve-heater supplied from a d.c. scource. Double decoupling of the anode circuit will usually ensure that the h.t. supply is sufficiently free from ripple. A practical layout which avoids earth looping is shown in Fig. 18.2, Chapter 18.

# FREQUENCY CORRECTION AND TONE CONTROL

IN the case of a microphone, the frequency response is taken to be 'flat', although in some cases when the microphone is operated close to a sound source, i.e. less than 2 feet, the bass becomes accentuated. A modification of the frequency response is then required, and is almost invariably necessary when the input source is a tape head or magnetic gramophone pickup. Provided the departure from a level response is known, a network can be designed so that the overall response is 'flat'.

## Frequency Discrimination

In any amplifier it is possible to reduce the amplification at high or low frequencies. For example, the bass response may be reduced by
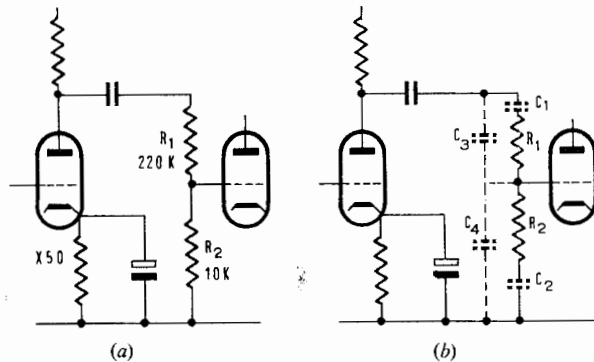


FIGURE 15.1

(a) Amplifier stage with divider $R_1 R_2$ in the output circuit. The overall gain is reduced by the factor $\dfrac{R_2}{R_1 + R_2}$. In the example, the gain is $50 \times \dfrac{10 \text{ k}}{10 \text{ k} + 220 \text{ k}} \simeq 2 \cdot 2$.

(b) Boost and cut at the maximum slope of 6 dB per octave can be obtained by introducing series or parallel capacitors into the divider network. $C_1$ bass cut; $C_2$ bass boost; $C_3$ treble boost; $C_4$ treble cut.

decreasing the capacitance of the coupling capacitors, and the treble response may be reduced by increasing the shunt capacitance at the output of a stage by connecting a capacitor from the anode to earth.

An increase in response can only be obtained (without resorting to resonant circuits) by reducing the gain at all frequencies other than that required to be 'boosted'. Naturally this entails a loss in amplification.

As several differing responses are usually required for various purposes, it is often necessary to add a valve stage, the stage providing the gain required for boosting. Various responses can be obtained by use of a variable resistor or by bringing into circuit various capacitors connected to a selector switch. The basic circuit is that of Fig. 15.1(a). If the valve gives a gain of 50, the divider $R_1$, $R_2$ will reduce the gain to $\dfrac{R_2}{R_1 + R_2} \times 50$.

The frequency response of the stage can be altered by placing capacitors in series or parallel with $R_1$ and $R_2$, as in Fig. 15.1(b). The maximum boost will be obtained when the impedance of the upper limb becomes insignificantly small compared with the lower limb. Thus, when a capacitor is connected in series with $R_2$, the reactance of the capacitor becomes high at low frequencies, e.g. a $0 \cdot 001$ $\mu$F
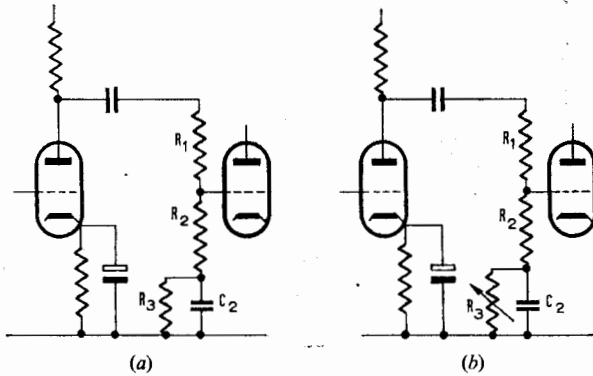


FIGURE 15.2

(a) A circuit providing boost of low frequencies. The maximum boost can be limited by $R_3$.

(b) The boost can be made variable if $R_3$ is a variable resistor.

capacitor has a reactance of $3 \cdot 18$ Megohms at 50 c/s giving bass boost. By choosing appropriate values for the resistors, the required degree of maximum boost can be obtained and, by choosing the

correct value capacitor, the frequency at which the boost commences can be set.

It must be remembered, however, that as the reactance of a capacitor doubles when the frequency is halved (i.e. drops one octave)
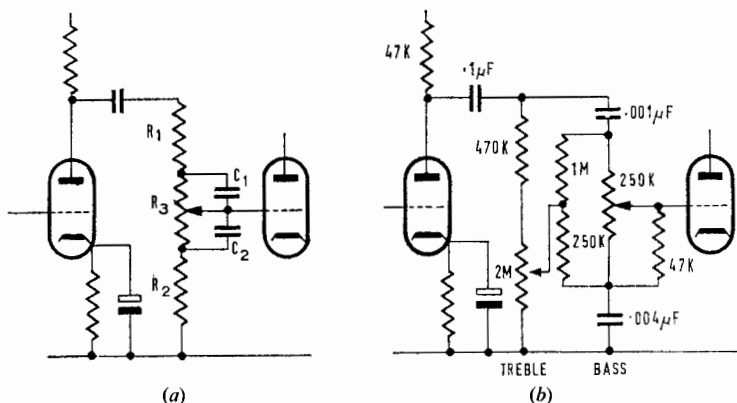


FIGURE 15.3

(a) Variable bass boost and cut with central 'flat' position, controlled by a single potentiometer.

(b) Complete tone control with variable bass and treble, boost and cut with flat central position, using two potentiometers (circuit due to P.G.A.H. Voigt).

the maximum boost which can be achieved is 6 dB per octave. It is of course possible to increase the boost by using a second divider network, when the rate of change becomes 12 dB per octave, but this entails a further loss of gain.

In cases where a slope greater than 6 dB per octave is required, a resonant circuit can be used, but steep slopes are seldom required in high fidelity reproduction, except for filtering (see page 128).

The boost is sometimes required to be limited, in which case the capacitor may be shunted by a resistor so that at a selected frequency further increase in reactance does not affect the response (see Fig. 15.2(a)).

In a similar way, a series capacitor can be introduced into the upper limb of the divider to give bass cut, whilst parallel capacitors connected across the upper and lower limbs of the divider will give treble boost and treble cut respectively. Again, the limit of boost and cut can be set by the introduction of limiting resistors.

**Frequency Correction**

A predetermined degree of boost or cut of treble or bass by selection of appropriate resistors and capacitors is called frequency correction.

124

A typical case for frequency correction is where a l.p. record is played by a magnetic pickup. Assuming a perfect pickup, the output will be in accordance with the international l.p. characteristic to which all modern microgroove records are made. The frequency correction required is the inverse of this characteristic, shown in Fig. 15.5(a), and is made up of (1) a treble cut circuit, (2) a bass boost circuit and (3) limitation of the bass boost at very low frequencies.

As frequency correction operates by reduction of gain, it is preferable to reduce gain by negative feedback so that distortion will be reduced and the basic frequency response improved. As feedback is applied via a resistor, the circuit can be made frequency discriminating by including capacitors in series or in parallel with the resistors. It must be borne in mind that boost is then obtained by the reduction of feedback giving increased gain, and at frequencies where feedback is virtually removed, reduction in distortion will not be obtained.



(a)                                   (b)
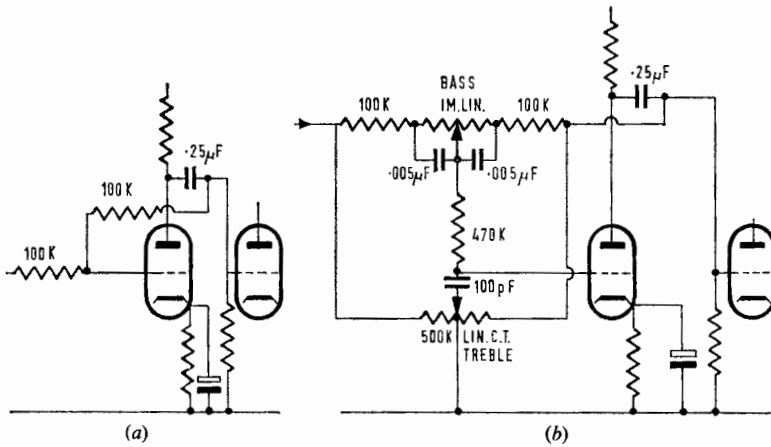
FIGURE 15.4

(a) Gain can be reduced by the parallel feedback circuit and is unity when the feedback resistors are equal. Feedback is varied by introducing capacitors and can be controlled by potentiometers. $R_1 = R_2 = 100 \text{ k}\Omega$

(b) A complete tone control circuit for bass and treble, boost and cut, with two linear potentiometers. (Circuit due to P. J. Baxandall.) This circuit provides approximately 20 dB boost and cut at 20 and 15,000 c/s. (Ref. 4. page 159).

A useful method of applying feedback to a single stage is the parallel feedback circuit of Fig. 15.4(a). Here the gain is nearly $\dfrac{R_2}{R_1}$ when the normal stage gain is high, and when $R_1 = R_2$, the stage gain is nearly unity. Frequency correction is obtained by placing

125

capacitors in series or parallel with $R_1$ or $R_2$ in the manner already discussed.

Thus the international l.p. characteristic is given by the circuit of Fig. 15.5($b$). It must be noted that the input impedance of this stage is virtually that of the input resistor, and this must be taken into account when the design of the preceding stage is considered.
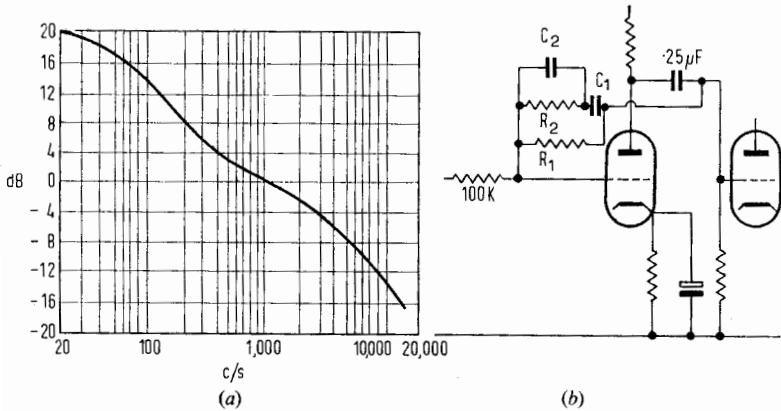


FIGURE 15.5

($a$) Reproducer frequency response required to give level output when a modern 33 or 45 r.p.m. gramophone record is played with a constant velocity pickup. This response can be expressed as a time constant in microseconds: top cut 75, bass boost 318, bass limitation 3180.

($b$) Parallel feedback circuit for international l.p. correction. When $R_2$ is 330 k$\Omega$, the gain with feedback and before correction is approximately $3\cdot3$. Due to interaction of the component parts, the values may be calculated as follows:

$$R_1 = 12\cdot4\,R_2 = 12\cdot4 \times 330\,k\Omega = 4\cdot1\,\text{M ohms}$$

$$C_1 \text{ in pF} = \frac{2940}{R_1(M\Omega)} = \frac{2940}{4\cdot1} = 718\,\text{pF}$$

$$C_2 \text{ in pF} = \frac{81\cdot2}{R_2(M\Omega)} = \frac{81\cdot2}{0\cdot33} = 246\,\text{pF}$$

Using standard 5% values, $R_1 = 3\cdot9$ M$\Omega$, $C_1 = 500$ pF $+ 220$ pF, $C_2 = 250$ pF.

Further, the frequency correction circuit will not operate correctly unless the impedance of the preceding circuit is low compared with the input resistor, i.e. not greater than one tenth.

A standard characteristic for $7\frac{1}{2}$ i.p.s. tape is the C.C.I.R. curve shown in Fig. 15.6($a$). The input from the tape head is low, usually about 5 mV, so that a low noise amplifier is required for a good signal-to-noise ratio. The triode has a low output impedance suitable for preceding a parallel feedback corrector circuit. The characteristic

126
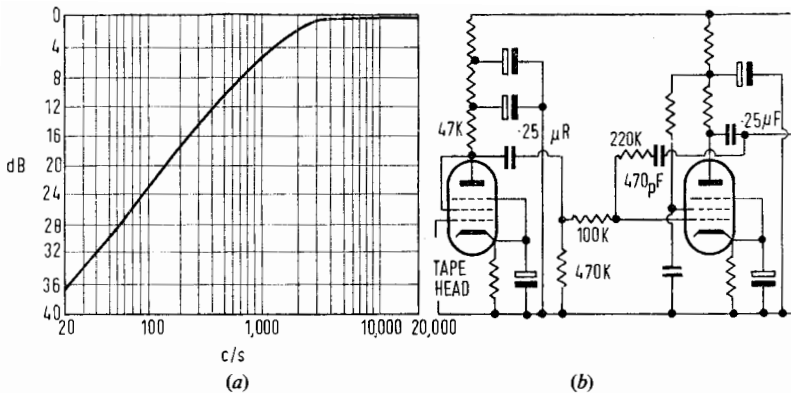
FIGURE 15.6

(a) C.C.I.R. $7\frac{1}{2}$ i.p.s. tape characteristic which can be expressed as a time constant of 100 microseconds. A bass boost circuit is required for correction.

(b) A low-noise triode amplifier gives amplification and provides a low output impedance preceding a parallel feedback corrector, giving bass boost. 220 k$\Omega$ × 470 pF = 100 $\mu$secs.

can be expressed as a time constant. The product $RC$, in megohms and picofarads, equals the time constant in microseconds. The $7\frac{1}{2}$ i.p.s. tape C.C.I.R. characteristic is 100 microseconds which is given by a $RC$ combination such as $0\cdot2$ M$\Omega$ and 500 pF (220 k$\Omega$ and 470 pF for standard values) as shown in Fig. 15.6(b).

The frequency correction stage can include a number of correction
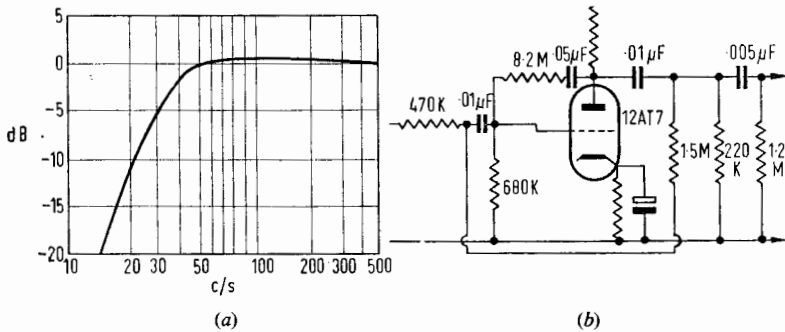


FIGURE 15.7

(a) Typical high pass filter frequency response to avoid low frequency noise and to improve decoupling at low frequencies.

(b) Circuit to produce response (a) (due to P. J. Baxandall).

circuits for various purposes brought into operation by means of a switch. The inputs can be brought into circuit by means of a second switch and the two switches coupled for single control operation.

127

**Tone Control**

It is desirable to be able to modify the low frequency and high frequency response of an amplifier, to adjust tonal balance. For this purpose a separate tone control stage is normally included in an amplifier and it is convenient to provide two controls, one for bass response and the other for treble response, each having a flat or unmodified response in its central position. The circuit is then arranged so that clockwise rotation of either control gives an increasing degree of boost and anticlockwise rotation an increasing degree of cut.

Either the divider circuit (called passive) or the feedback circuit can be used. The development of suitable circuits is traced in Fig. 15.3 for the passive case and Fig. 15.4 for the feedback case.

**Steep Cut Filters**

Whilst high fidelity reproduction normally requires a wide frequency range, ideally 20 c/s to 20 kc/s, in some instances a wide response will
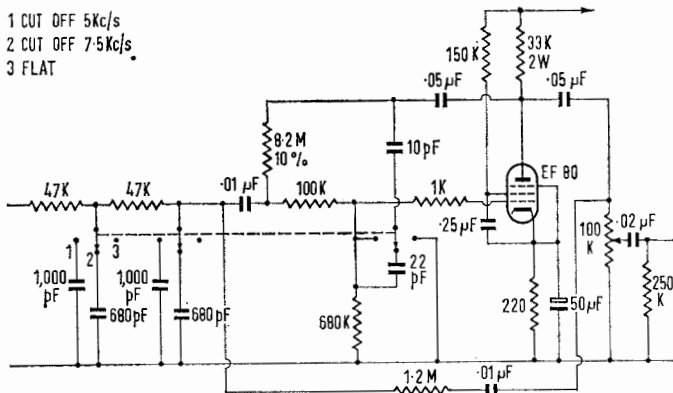


FIGURE 15.8

Circuit to produce a low pass filter with a cut-off rate of 18 dB per octave above a frequency determined by component values. The cut-off frequency can be changed by switching alternative capacitors into circuit, and the low pass filter can be combined in the same stage as the high pass filter. (Circuit due to P. J. Baxandall.)

pass unwanted low and high frequencies such as turntable rumble and gramophone record scratch. In these cases, the loss of quality due to restriction of the frequency range may be less important than the annoyance caused by the unwanted signals. Restriction of the high frequency range can best be achieved by inclusion of a low pass filter circuit, and restriction of low frequencies by a high pass filter. These filters enable the high frequency and low frequency loss to exceed 6 dB per octave.

128
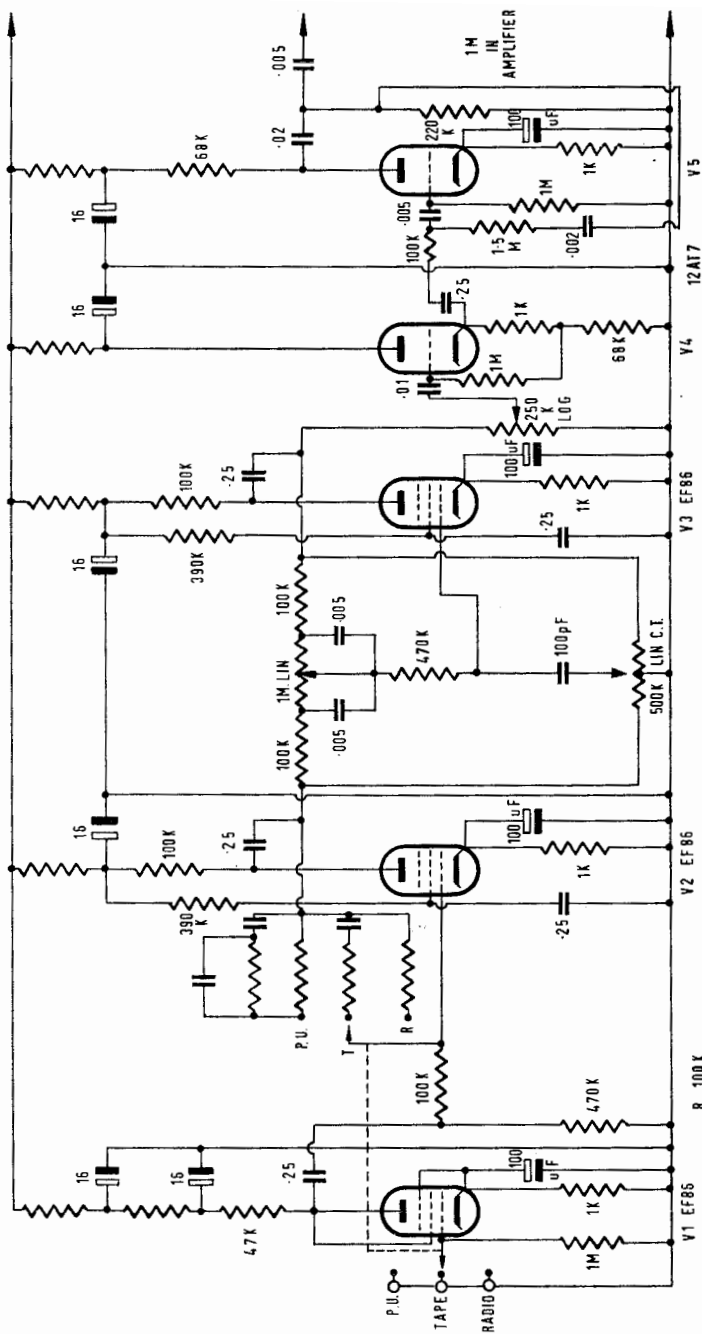
FIGURE 15.10

Complete control unit incorporating circuits discussed. $V_1$ low noise triode amplifier stage. $V_2$ pentode corrector for switched characteristics for international l.p., C.C.I.R. $7\frac{1}{2}$ i.p.s. tape and radio. The corrector switch is coupled to the input switch, connecting high quality low sensitivity pickup, tape head and radio tuner. $V_3$ tone control stage, unity gain. $V_4$ cathode follower coupling volume control. $V_5$ high pass filter. This control unit will give an output sufficient to drive the 10 watt amplifier at input 2.

129

Whilst the required response can be obtained by the use of resonant circuits, a feedback circuit requiring only resistors and capacitors is preferable, such as that illustrated in Figs. 15.7 and 15.8.

Such a circuit can be arranged to operate for several predetermined bandwidths by switching suitable capacitors into circuit. (Ref. to page 159.)

### Volume Control

A combination of the circuits discussed can be constructed as a control unit, used remotely from the main amplifier and provided with suitable controls. A volume control is introduced into the circuit in the form of a potentiometer. The 'placing' of this control in the amplifying chain must be carefully considered. The gain of the whole amplifier must be such that there is sufficient reserve to enable the smallest anticipated input signal to produce maximum power output.

FIGURE 15.9

Potentiometers are supplied in various gradings, and tone controls sometimes require a linear curve so that resistance is proportional to rotation. Volume controls are usually of the log type, the resistance curve of which is illustrated above.
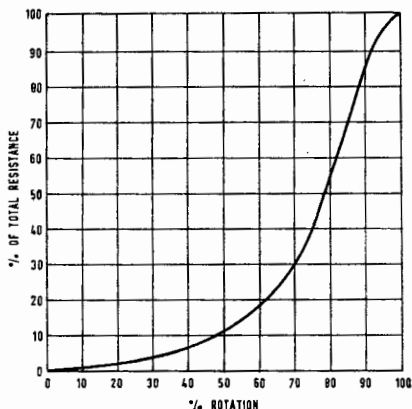
When the input signal is very small, the signal-to-noise ratio will be at its lowest, and background noise may be considerable. If the volume control is placed at the input terminals of the preamplifier, the background noise is not reduced even when the control is set at zero. As the noise is mainly that of the first stage, it can be reduced with gain if the control is placed after the first or a later stage. It must be remembered that if there is a reserve of gain, after one or two stages of amplification, there is the possibility of a valve being overdriven, thus causing unnecessary distortion. The volume control may therefore be placed after a stage where the output, with highest input signal and highest gain, does not exceed about 1 V r.m.s.

The volume control is usually graded logarithmically so that there is greater latitude of setting at low volume levels (see Fig. 15.9). The potentiometer is then referred to as a 'log' control.

130

# THE AMPLIFIER AND LOUDSPEAKER COMBINATION

THE performance of a high fidelity amplifier has so far been considered when operating into a resistive load. This type of amplifier will be required to operate a loudspeaker and the complicated nature of this as a load has been mentioned.

### Loudspeaker as a Load

Fig. 16.1 shows an equivalent circuit of a typical 12-inch loudspeaker unit and Fig. 16.2 shows its impedance over the audio frequency range. The equivalent circuit shows that the reactance is capacitive at some frequencies and inductive at others.

An amplifier should therefore be finally tested with a 'dummy load' built to the circuit of Fig. 16.1 and its performance checked particularly with regard to stability. The matter may be further complicated by the use of two or more loudspeakers, with a crossover filter which restricts the high frequency response of the bass unit and the low frequency response of the treble unit. The equivalent circuit of the loudspeaker might then appear as Fig. 16.3.

Whilst distortion has been considered in terms of the percentage of various harmonics of the fundamental present in the output,
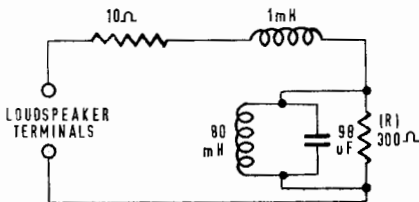


FIGURE 16.1

Approximate equivalent circuit of typical 12-inch moving coil loudspeaker. $10\Omega$ is the DC resistance of the coil, and 1 mH the inductance of the coil.

harmonics do not in themselves produce distortion and are component parts of natural musical sounds, pure sine waves being produced only by a signal generator, a tuning-fork and perhaps a
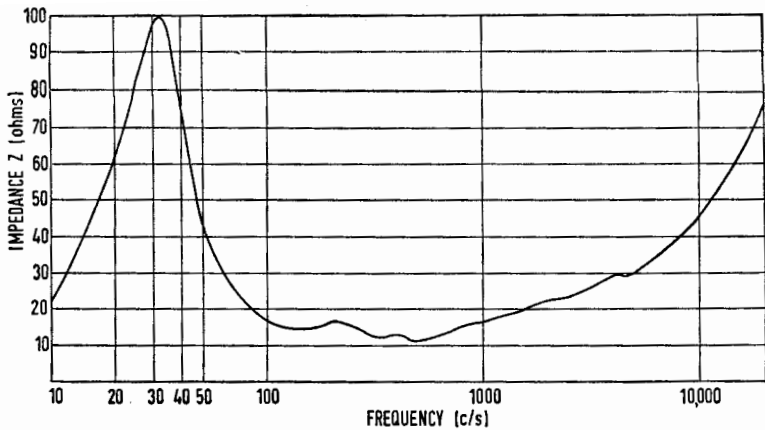
FIGURE 16.2

The impedance curve of a typical 12-inch loudspeaker quoted as
15 Ω at 400 c/s. Bass resonance at 32 c/s.

lightly blown flute. The measured harmonics are in fact an indication
of the non-linearity of the amplifier, the audible 'unpleasantness'
being due to intermodulation tones produced when two notes with
their harmonics are reproduced simultaneously. Intermodulation
tones thus produced are not whole number multiples of the funda-
mentals and cause unpleasant interference with the original sounds.
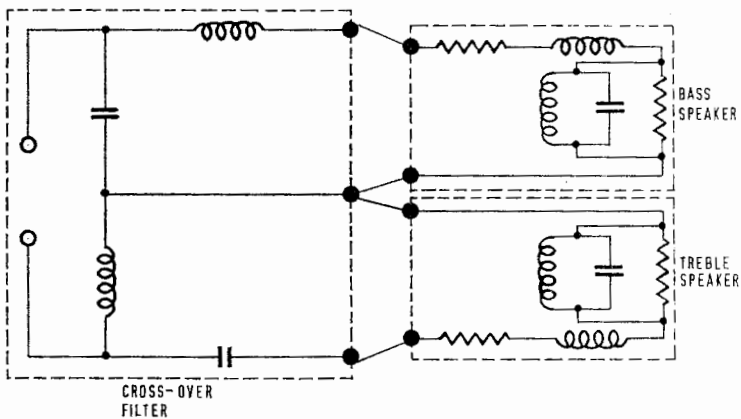


FIGURE 16.3

Approximate equivalent circuit of two loudspeakers with cross-over
filter sharing the signal between bass and treble units.

Thus a more realistic method of measurement of distortion is the
intermodulation method, and I.M. figures are given in some of the
application reports in the Appendix.

132

In practice, fortunately, most well-designed amplifiers which show up well on resistive load test with low harmonic distortion will give satisfactory results when operating with a loudspeaker.

### Effect of Loudspeaker upon Amplifier Performance

It has been shown that a valve power amplifier has an optimum load resistance which is a compromise between maximum output and minimum distortion. The way in which maximum power output for
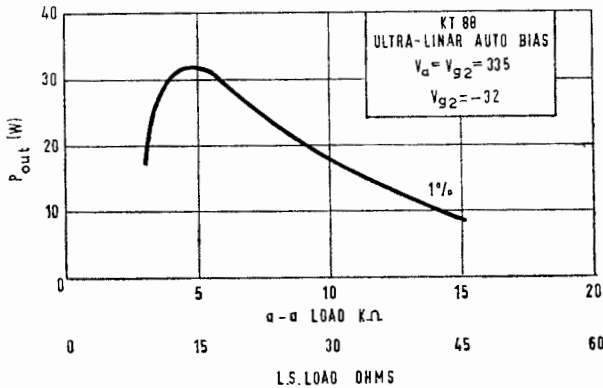


FIGURE 16.4
Variation of maximum power output for 1 % total harmonic distortion of ultralinear tetrode output stage, with change of load resistance.

a given percentage harmonic distortion varies with load resistance is shown in Fig. 16.4. This relates to a tetrode ultralinear push-pull stage, for which the change in performance with load is relatively small, a triode stage being rather better and a pentode stage worse.

If the minimum loudspeaker impedance, which occurs at about 400 c/s, is matched at optimum load impedance, then the value of the load impedance will rise at frequencies above and below 400 c/s so that the maximum power output capability of the output stage will fall. If the amplifier is rated for, say, 10 W on resistive load, this power will be available on loudspeaker load at 400 c/s. At higher frequencies the available power may decrease due to the output transformer and an increase in load impedance. Fig. 16.4 shows that if the impedance rose to 45 $\Omega$ at 10,000 c/s (not an unlikely figure) a 32 W amplifier would only deliver about 8 W maximum at this frequency. Average sounds, however, do not contain large high frequency components, so that an amplifier which is delivering 30 W

133

at 400 c/s will seldom be required to deliver as much as 8 W at 10,000 c/s.

At low frequencies there is an increase in loudspeaker impedance which rises to a peak at the resonant frequency of the parallel capacitor and inductor in the equivalent circuit. This peak is called the cone resonance of the loudspeaker and, for the loudspeaker in the example, is at 32 c/s. The rise in impedance is limited by the parallel resistance R in Fig. 16.1 which represents the damping provided by the cabinet in which the speaker is mounted, friction in the cone and air loading. The rise in impedance therefore varies from suspension speaker to speaker and with the nature of the cabinet in which it is enclosed.

It is the portion of $R$ which represents air loading in which acoustic power is developed. As this is in parallel with the resonant circuit and in series with the speech coil resistance and inductance, it is clear that the efficiency will be low. The increase in the impedance of the parallel circuit will cause the acoustic power output to rise at cone resonance. As a flat frequency response is required, this impedance rise must be controlled, and the cabinet in which the loudspeaker is mounted can be arranged to provide damping.

The reactance of the speech coil at the frequency of cone resonance is small compared with the speech coil resistance. The input terminals of the loudspeaker are coupled to the output valves via the output transformer, so that the anode impedance of the output valves appears across the secondary or output terminals of the amplifier, reduced by the square of the transformer turns ratio.

### Damping

If negative voltage feedback is applied, this impedance will be reduced and provides damping. The ratio of the amplifier output impedance to the load impedance is called the *damping factor*. Thus if the load is 15 $\Omega$ and the amplifier impedance 0·5 $\Omega$, the damping factor will be $\frac{15}{0·5} = 30$. It must be remembered that the amplifier output impedance is in series with the speech coil resistance so that, in the case of the 15 $\Omega$ loudspeaker, when the amplifier impedance drops below 1 $\Omega$ the increase in damping is negligible.

This can be illustrated if the amplifier damping is expressed as the ratio of load impedance to load impedance plus amplifier impedance, which cannot exceed unity. Thus in the example quoted above, the damping ratio is found to be 0·9. Any further reduction in amplifier impedance can only result in the damping factor more closely

approaching unity. The terms damping factor and damping ratio are due to F. Langford-Smith.

Fig. 16.5 shows the measured sound output over the audio frequency range of a twin cone loudspeaker mounted in a $4\frac{1}{2}$ cubic feet reflex cabinet, and Fig. 16.2 is the impedance characteristic of the same unit. The response was measured in an anechoic room on the
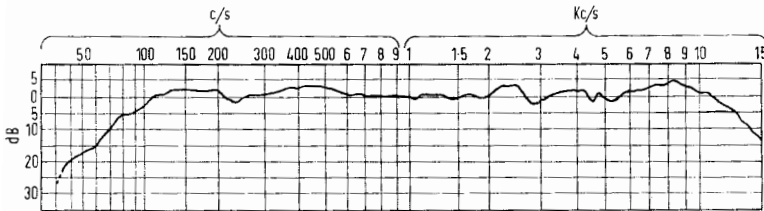


FIGURE 16.5

Free field frequency response curve of typical high fidelity 12-inch loudspeaker in a $4\frac{1}{2}$ cubic feet reflex cabinet, with constant voltage input. (Courtesy Goodmans Industries Ltd.)

axis of the speaker, the input being maintained at constant voltage, which is the equivalent of an amplifier with zero output impedance.

**Low Frequency Response**

It will be observed from Fig. 16.2 that the frequency of cone resonance is at 32 c/s and this is well damped in the frequency response. The low frequency response, however, shows some loss at frequencies below 100 c/s. This loss is exhibited by most loudspeaker arrangements and varies with the loudspeaker magnetic flux density, the amplifier output impedance and the cabinet design. In practice, the loss in bass response may be less than that indicated by Fig. 16.5 due to the influence of the room in which the unit is operated.

The loss at very low frequencies could be compensated by a special frequency correction circuit in the amplifier, but if this is done, care is required as increased power output at such frequencies will result in large cone excursions, so increasing loudspeaker distortion.

**High Frequency Response**

If the loudspeaker cone is considered to operate as a piston at frequencies above 100 c/s, the resistance due to air loading will be proportional to the square of the frequency. Hence the voltage across the speaker input should also be proportional to frequency, i.e.

135

constant current or high output impedance. This theoretical requirement does not seem to hold in practice and this may be due in part to complicated break-up of the cone, which varies its effective piston diameter. The response is also affected by reflections encountered in the normal listening room. It is significant that the response of Fig. 16.5 was taken with constant voltage input, apparently the condition under which the loudspeaker manufacturers intended the unit to be operated.

The complicated nature of the loudspeaker and cabinet combination and the effect of amplifier output impedance, with the possibility of some frequency compensation, would seem to indicate a strong case for designing amplifier, and speaker cabinet as a single unit for operation in an average listening room.

# TRANSISTORS

LIKE the thermionic valve, the transistor is a device in which the flow of charge carriers can be regulated in a manner which produces power gain. Thus, the input control signal when applied to a transistor in a suitable circuit will appear in amplified form at the output of the circuit. Apart from the fact that the valve and transistor both depend upon field effects for their action, they have little else in common. For example, the transistor is essentially a current operated device whereas the valve is voltage operated, i.e. for the transistor, the input signal *current* must be considered rather than an input signal *voltage*.

Again, the transistor does not depend upon thermionic emission, so no heater power is required and this results in a much greater efficiency. Another difference between the valve and the transistor can be seen in the power supply voltages that are required. It is common for valves to operate with supply voltages of, say, 100–400 V, whereas most transistors use supplies of 3–30 V. These and many other differences show that separate consideration must be given to transistors and their characteristics.

## Semiconductors

Unfortunately, the physical processes involved in transistor action are rather more complicated than those involved in a thermionic valve, and are such as to make a simplified description somewhat inaccurate. However, bearing this fact in mind, it will be remembered that the electrical conducting properties of solids were briefly discussed in Chapter 1. It was found that some materials are good conductors whilst others are very poor conductors. It is possible to

find materials which fall roughly midway between these extremes, and such materials are known as semiconductors. For the present discussion the most important semiconductor materials are germanium and silicon.

Now it is possible to alter the resistivity of an intrinsic semiconductor material by adding a very small quantity of an impurity material. Of particular interest are those impurities which combine with the intrinsic semiconductor to produce a material which has either an excess of electrons or a deficiency of electrons. The resultant materials are known as $n$-type and $p$-type respectively and the 'missing' electrons in the $p$-type are often referred to as 'holes'.

### Junctions

It is clear that, if a junction is formed between $n$- and $p$-type materials the result will be a redistribution of charge in the immediate vicinity



FIGURE 17.1
Distribution of charge in the region of a junction of semiconductor material.
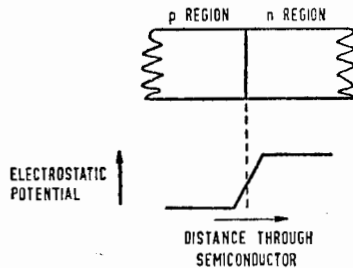
of the junction. This in turn generates a potential barrier as shown in Fig. 17.1 and, for the purposes of this simplified explanation, it is the barrier which is of primary interest. Once this barrier is established, it will reduce the facility with which electrons may flow from the $n$-type material to the $p$-type. Since we may regard the missing electrons or holes in the $p$-type as carriers of positive charge, this barrier will also inhibit the flow of these holes from the $p$- to $n$-type material.

Thus, in the absence of any external influence, the junction between the two types of material forms a kind of insulating barrier. The 'height' of this barrier can be influenced by the application of an external voltage source as shown in Figs. 17.2 and 17.3.

In the first case, assume that the polarity of the voltage is such as to make the $n$-type more positive with respect to the $p$-type. This will increase the potential barrier and hence even fewer charge carriers will have sufficient energy to pass over it. Hence the current in the external circuit will be very small. In the second case, assume that the

138

polarity of the external voltage is such as to make the *n*-type more negative with respect to the *p*-type. This will decrease the height of the potential barrier and will facilitate the movement of charge carriers. Hence the current in the external circuit will be relatively large.

It can be seen, then, that the *p-n* junction has the properties of a diode in that current may flow in one direction but not in the other.
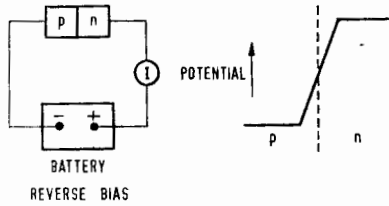


FIGURE 17.2

When the *n*-type is made positive there is an increase in potential barrier and hence electron flow is made more difficult.

It should perhaps be mentioned that the simplified description given here does not account for all of the properties of a semiconductor junction diode. For example, as the reverse bias voltage is increased, a point is reached at which the current will rise very rapidly with further increase in applied voltage. The point at which this occurs is known as the zener voltage and diodes, usually specially designed for the purpose, are sometimes used in this mode of operation as voltage stabilizing elements. Hence in normal rectifier circuits it is important that the maximum reverse voltage should not reach that of the zener breakdown.

### The Transistor
Having discussed how the diode action may be achieved with semiconductor elements, we may now proceed to consider the transistor, which enables signal amplification to be obtained. We can imagine a
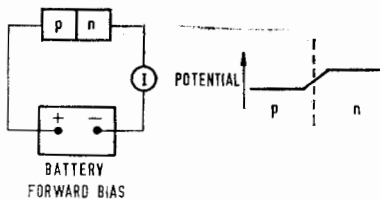


FIGURE 17.3

When the *n*-type is made negative there is a decrease in potential barrier and hence electron flow is made easier.

transistor as being made up of three layers of semiconducting material as shown in Fig. 17.4. We shall take the *p-n-p* type for discussion but *n-p-n* arrangements are now common. The principles involved apply to both types, the main practical difference being that of the polarity of the applied voltages.

139

We notice that the transistor resembles two diodes, placed back to back, formed in one piece of semiconducting material. In Fig. 17.4 the left- and right-hand *p*-type material will be referred to as the

FIGURE 17.4

Distribution of charge in a transistor with *p–n–p* junctions.
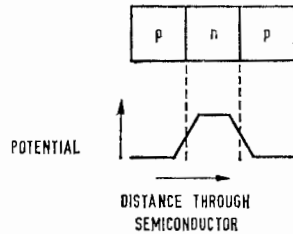
emitter and collector respectively, and the *n*-type as the base. Now if the collector/base junction is reverse biased, we should expect the current in this circuit to be very small because there are very few positive charges (holes) in the *n*-type and very few negative charges (electrons) in the *p*-type that are free to move in the potential gradient.

## Common Base Connection

However, it is possible to make some charge carriers available by applying a voltage to the emitter/base junction. If this voltage causes the base to become more negative relative to the emitter, holes will drift into the base region and some of these will reach the collector/base junction. Here they will be accelerated by the negative gradient and hence will constitute a current. The magnitude of this current will be dependent upon the number of charge carriers made available in the base region, and hence on the amount of forward bias placed

FIGURE 17.5

A *p–n–p* transistor with emitter biassed positive causing a large collector current to flow (top). A *p–n–p* transistor with emitter biassed negative when the collector current is small (below).
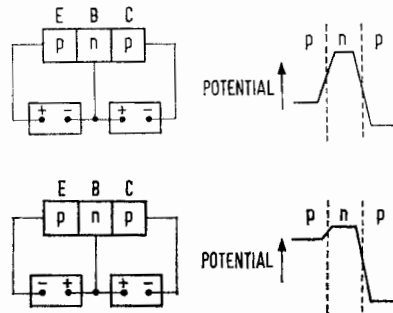
on the emitter/base junction. This mode of operation of a transistor is known as the common base configuration and is shown diagrammatically in Fig. 17.5.

140

In the common base configuration, it is clear that the current flowing in the collector circuit will be very nearly the same as that in the emitter circuit. Hence this method of connection does not provide current gain but does give power gain, since the resistance in the forward biased emitter is low and that of the reversed biased collector is high. (Power $= I^2R$, hence power gain will be $\dfrac{R_c}{R_e}$ approximately.)

## Common Emitter Connection

The very low input resistance (of the order of 100 $\Omega$) of the common base circuit makes it unsuitable for most audio-frequency applications, and hence an alternative mode of connection (common
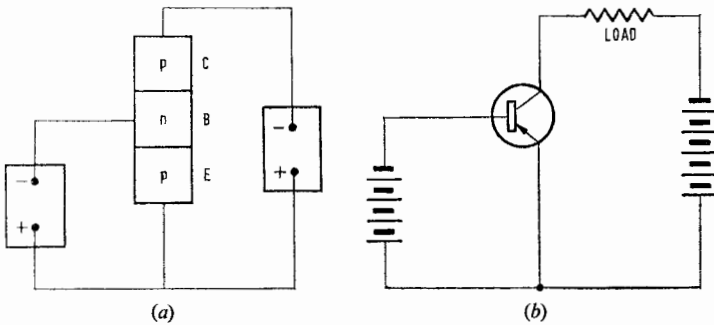


FIGURE 17.6

(a) Common emitter connection.

(b) Usual circuit symbol with load resistor included. The collector current is small when the base current is zero and large when the base current is negative.

emitter) is usually employed. Here the control signal is again applied to the emitter/base junction, but the emitter is earthed as shown in Fig. 17.6. The exact description of the action of the transistor in this configuration cannot be given here, but inspection of Fig. 17.6(a) shows that the output current also flows in the emitter/base junction. Therefore a given output current will be maintained with a much smaller input current than would be required in the common base connection. Thus, current gain can be achieved and might typically be around 50, whilst the input resistance is raised to, say, 5,000 $\Omega$.

## Common Collector Connection

The other mode of connection is used at times in audio-frequency applications and is known as the common collector or emitter follower configuration. As its name implies, it is analogous to the

141

valve cathode follower and, as with this circuit, its main function is that of an impedance convertor. With this configuration it is relatively easy to achieve input resistances of about 500 kΩ, but like the cathode follower the voltage gain must always be slightly less than unity.

From the preceding brief discussion of the principles of transistors it will be noted that they are essentially current operated devices and that there is always a finite resistance connecting the input and output circuits (unlike the thermionic valve where, at least in principle, there is no such connection between input and output). A further point to be borne in mind is the fact that the materials of which the emitter base and collector are made are semiconductors. Hence, the resistance of these elements in the transistor must be taken into account when calculating its performance in any circuit.

**Working Point**

In considering a transistor circuit suitable for providing linear amplification, the first requirement is that of selecting a satisfactory mean or quiescent working point. As with a valve stage of amplification, it is necessary to select the load resistance, bias values, etc., to ensure that linear amplification, or a close approach to it, will be achieved. In fact the process of selection is very similar to that given
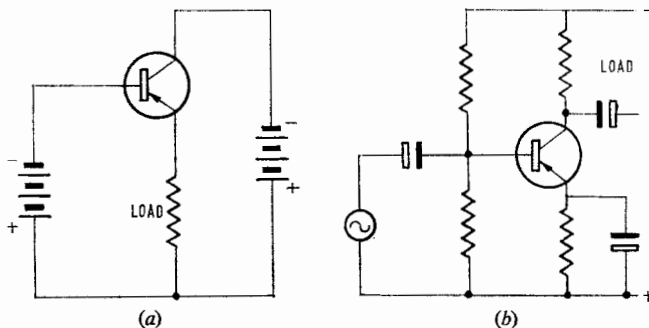


FIGURE 17.7

(a) Common collector or emitter follower connection.
(b) Practical circuit for transistor amplifier stage.

in Chapter 3 for valve amplifiers and hence this discussion will be limited to those aspects of the problem which make the transistor different from the valve.

It will be obvious from the preceding discussion that the transistor has to be biased in the forward direction; i.e. in the absence of a bias

142

currenc the transistor will be in the cut-off condition. This means that biasing cannot be achieved by voltage drop in resistance in the emitter circuit, for the emitter must be biased positively with respect to the base. One of the most common forms of bias arrangements is shown in Fig. 17.7(b) from which it will be seen that the base is biased in the negative direction by potential divider. It should be added that there are a number of other ways of obtaining bias current in a transistor circuit.

It will also be seen from Fig. 17.7(b) that a bypassed resistor is included in the emitter circuit. This resistor, which is usually about one-fifth of the value of the collector load resistance, is necessary to compensate for effects of temperature variation. Most of the transistors commonly available are the germanium type in which the collector current varies considerably with temperature. The inclusion of the resistance in the emitter circuit minimizes the change and hence keeps the quiescent operating point sensibly independent of normal variations in room temperature.

### Gain

The gain of a single transistor stage can be derived from the dynamic load line and a family of $I_c/V_c$ curves. However, the input quantity

TABLE 17.1  *Transistor Formulae*

| (a) Common Emitter Configuration | |
| --- | --- |
| Voltage Gain | $\dfrac{\alpha RL}{r_e + r_b(1 - \alpha)}$ |
| Input resistance | $\dfrac{r_b (1 - \alpha) + r_e}{1 - \alpha}$ |
| Output resistance | $\dfrac{R_g r_c(1 - \alpha) + r_e r_c}{R_g + r_b + r_e}$ |
| (b) Emitter Follower Configuration | |
| Voltage Gain | 1 approx |
| Input resistance | $\dfrac{R_L}{1 - \alpha}$ |
| Output resistance | $r_e + (1 - \alpha)(R_g + r_b)$ |

143

is in terms of base current and hence a direct statement of voltage gain cannot be obtained. For this and other reasons it is more usual to calculate the gain from a knowledge of the transistor and other
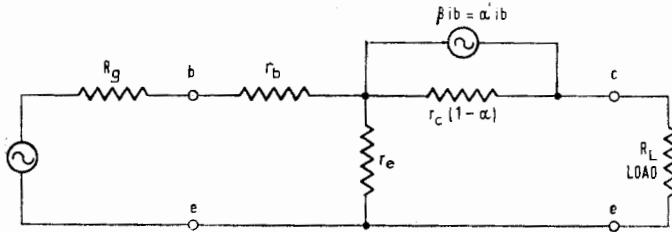


FIGURE 17.8
The T equivalent circuit for a.c. currents.

circuit parameters. In this method an equivalent circuit of the transistor is assumed and manufacturers quote the appropriate transistor parameters in their data sheets for each type of transistor.

Unfortunately there has been some lack of standardization in the drawing of the equivalent circuit of transistors and in particular the T and $\pi$ equivalents are frequently used. Since the T is easier to understand in the first instance, it is shown in Fig. 17.8 for the common emitter configuration. Solution of the circuit equations yields the results given in Table 17.1 from which the voltage gain, and the input and output resistances of the stage may be calculated. Another table is included to give similar formulae for the emitter follower configuration, this and the preceding configuration being those most commonly employed in audio-frequency work. It should be stressed that these are approximate formulae, but the errors involved in applying them to most practical circuits are small.

Some manufacturers quote only the hybrid or $h$-parameters, which are appropriate to the $\pi$ equivalent circuit. To assist in this direction, table 17.2 shows the relationship between the T and $\pi$ parameters, so that, knowing one set, the other can be calculated using the formulae in the table.

### Frequency Response
So far in this discussion no account has been taken of the frequency dependent effects in a transistor stage of amplification. As with a valve, there are a number of factors which cause the gain to change as the frequency is varied over a wide range.

Taking the low frequency range first, the capacitors which couple one stage to another and which bypass, say, the emitter resistance,

144

are the primary cause of gain variation. These effects are exactly the same as those which obtain in a valve amplifier circuit and so need not be discussed in great detail here. It is worth noting, however, that the input resistance of a transistor stage is very much lower than that of a valve stage (about 10 kΩ as opposed to $0·5$ MΩ). Consequently the coupling capacitor must be proportionately larger for a given low frequency response characteristic. Thus it is common to find coupling capacitors of about 10 $\mu$F in transistor circuits but, fortunately, because of the relatively large input current required to drive these circuits, the leakage current of an electrolytic capacitor can usually be accepted. The fall-off of low frequency response due to

TABLE 17.2 *Approximate relations between h parameters ($\pi$ equivalent) and T parameters*

| | |
|---|---|
| $h_{ib} = r_e + r_b(1 - \alpha)$ | $r_e = h_{ib} - \dfrac{h_{rb}}{h_{ob}}(1 + h_{fb})$ |
| $h_{rb} = \dfrac{r_b}{r_c}$ | $r_b = \dfrac{h_{rb}}{h_{ob}}$ |
| $h_{fb} = -\alpha$ | $r_c = \dfrac{1}{h_{ob}}$ |
| $h_{ob} = \dfrac{1}{r_c}$ | $\alpha = -h_{fb}$ |

the increasing reactance of the emitter bypass capacitor presents a problem similar to that of the cathode resistor bypass capacitor in the valve circuit.

The high frequency limitations of transistor circuits are more complicated to discuss but, for the purposes of audio-frequency work, perhaps the most important is the variation of $\alpha'$ (or $\beta$ as it is often designated) with frequency. Because of capacitative and other effects within the transistor, its current gain decreases with increasing frequency. Most manufacturers now quote $r'_{be}$ and $C'_{be}$ for the common emitter configuration, and a rough guide to the upper frequency limit can be gained by calculating the frequency at which the reactance of $c'_{be}$ is equal to $r'_{be}$. For example, for the OC45, $r'_{be} = 1·3$ kΩ and $c'_{be} = 1,000$ pF. Hence the frequency at which the capacitative reactance is equal to the resistance is approximately 125 kc/s and at this frequency the transistor gain will be reduced by

3 dB. It must be stated that the calculation of the true high frequency response is much more involved than this, but the above calculation is sufficient in most practical situations to indicate the suitability of a given transistor so far as its high frequency response is concerned.

The usual factors of circuit shunt capacitances must also be taken into account when calculating the high frequency response of an amplifier stage. With the much lower resistance values that are normally involved in transistor circuits, however, these factors are less troublesome than is the case in valve circuits.

### Distortion

A detailed discussion of non-linear effects in transistors cannot be pursued here, but as a rough guide in small signal amplifiers it may be assumed that distortion generated is about ten times greater than in an equivalent triode amplifier. Thus in general it is good practice to employ negative feedback circuits to reduce distortion from the figure of about $1\%$ in the absence of feedback to about $0\cdot1\%$. Of course with careful design it is possible to improve on this latter figure, but it does show that for high quality transistor voltage amplifiers the reduction of non-linear distortion needs rather more attention than in the equivalent valve amplifier.

### Noise

In the early stages of a voltage amplifier, noise generation is of course an important factor and, in the case of transistor stages, a rather complicated situation arises. However, a good general rule is to operate at least the first stage at a low collector current, since the noise generated in a transistor is related to the current. In addition, the noise factor is improved by selecting a transistor with low $I_{co}$, low $r_{bb}'$ and high $\alpha'$.

### Advantages and Disadvantages

The advantages accruing from the use of transistors in audio frequency amplifiers are (1) considerable space saving, (2) mechanically robust, (3) no heater supply required, (4) low impedance circuits reduce electrostatic hum pickup.

Disadvantages are (1) low impedance circuits are not suitable for use with many existing devices, e.g. pickups, tape heads, (2) low impedance circuits are more prone to earth loop hum, (3) the variation in transistor characteristics from unit to unit of the same type is large compared with that to be expected from valves, (4) transistor

146

characteristics vary with room temperature, (5) voltages and currents in excess of the maximum stated by the manufacturer can completely destroy the transistor's electrical properties, even though the excess be applied for a very short time. For example, earth leakage through a soldering iron can be sufficient to destroy a transistor, as can excessive heating of its leads due to poor soldering conditions.

The advantages mentioned above may be important where size and weight is the prime consideration, e.g. portable receivers. In the normal high fidelity equipment, the size and weight of the main amplifier is not of particular significance and the size of a control unit cannot be reduced, at least so far as the operating panel is concerned, due to the necessity of providing a number of controls of convenient size and disposition for manual operation.

In the first stages of a control unit there is some advantage over a valve as there is no heater supply which, if a.c. introduces hum into the stage. Further, low impedance circuits, provided loops are avoided, are less prone to hum pickup. Some transistors do not compare favourably with valves so far as hiss noise is concerned, but improvement here can be expected.

Transistors in the output stage of the main amplifier introduce considerable distortion, requiring heavy feedback for a reduction to the levels obtainable from valves. As transistors can develop high signal currents, there is the distinct advantage that the output transformer can be dispensed with. Transistors, however, have a finite power gain and if an output power of 10 W is required, it may be necessary to supply an input power to the stage of about 1 W. It is, therefore, generally necessary to consider at least the last two stages of a main amplifier from the power gain point of view.

Improvements in transistor construction and further experience with circuitry will undoubtedly result in a greater number of transistorized control units and main amplifiers being produced. Two circuit configurations by Mullard Ltd. are shown in Figs. 17.9 and 17.10.

The output stage of Fig. 17.9 uses a transformerless design and each channel of the amplifier delivers 10 W (sine wave) into a 15 Ω load. The output transistors are used in a new class AB mode of operation. This mode of operation, referred to as '$\pi$-mode class AB', differs from the commonly known class AB valve operation in that the total d.c. current remains constant with drive. For a power output of up to 40% of full power, the transistors operate in class A push-pull. As the drive is further increased, the operating conditions change so that the transistors operate in the class AB
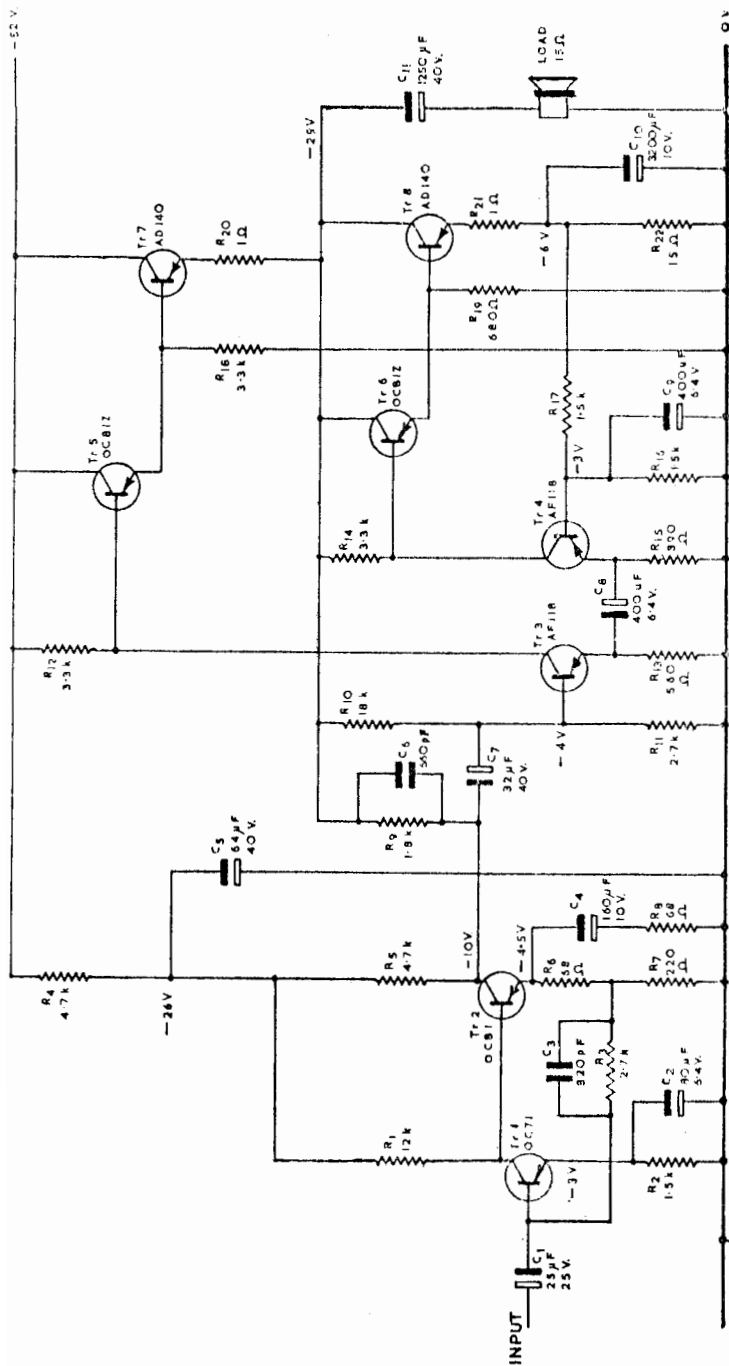
FIGURE 17.9

Transistorised 10 watt amplifier with class AB output stage. The circuit shown may form one channel of a 10 × 10 W stereo amplifier. (*Courtesy Mullard Ltd.*)
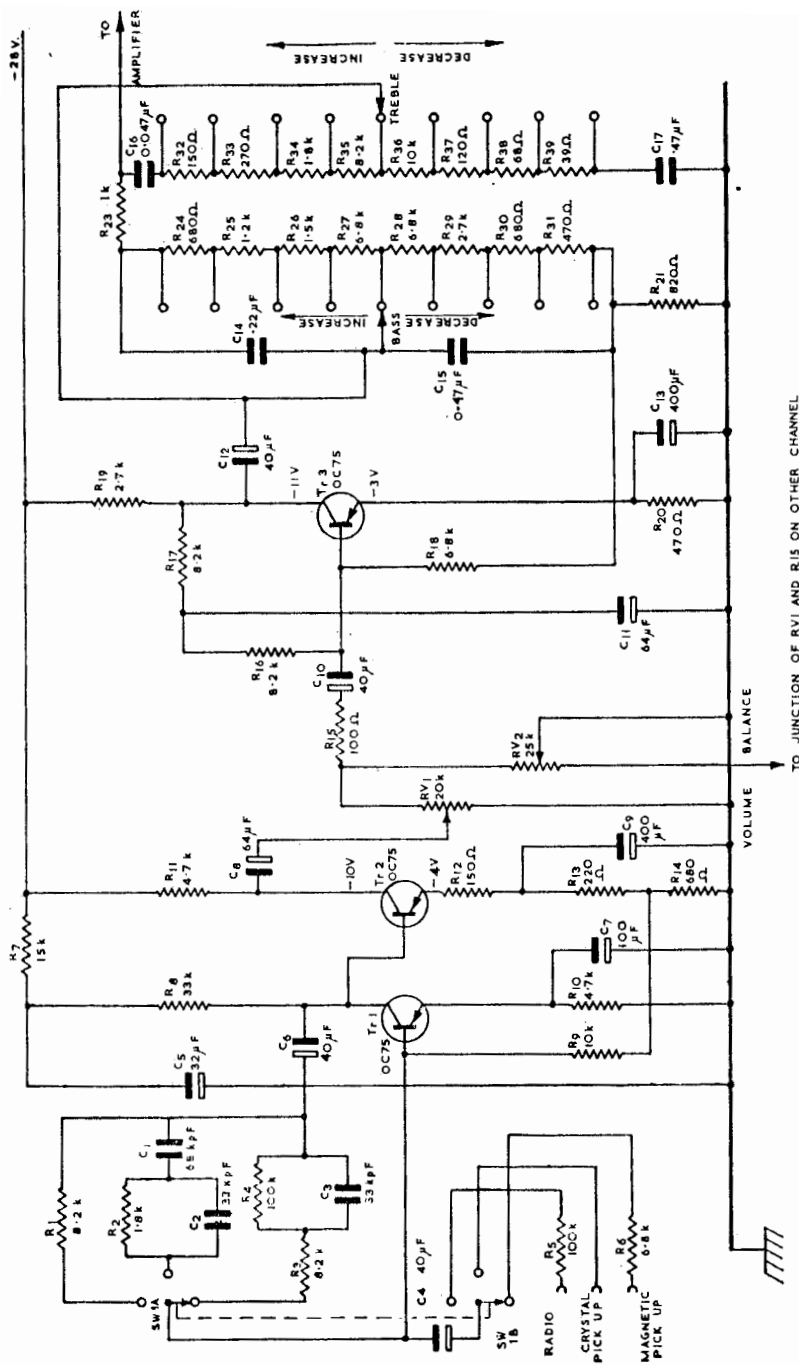
FIGURE 17.10

Three-transistor pre-amplifier with provision for crystal and magnetic pickups and a radio tuner. The circuit shown may form one channel of a stereo pre-amplifier. (*Courtesy Mullard Ltd.*)

149

mode until, at full power, class B operation is obtained. The output transistors (AD140), driver transistors (OC81Z) and phase splitter (AF118) are direct coupled. Overall negative feedback of 44dB is used over the three stages. This three-stage amplifier is preceded by a two-stage directly-coupled amplifier using an OC71 and an OC81.

*Performance*

| | |
|---|---|
| Frequency response (−3 dB): | 16 c/s to 50 kc/s at 500 mW |
| Power response (−3 dB): | 16 c/s to 18 kc/s at 10 W |
| Overall negative feedback: | 44 dB |
| Distortion at 1 kc/s: | 0·06% at 10 W |
| Sensitivity: | 140 µA for 10 W |
| Output impedance: | 0·2 Ω at 1 kc/s |
| Power supply requirements: | 0·9 A at 52 V |

The pre-amplifier of Fig. 17.10 consists of a two-transistor directly-coupled input stage with equalisation provided by overall negative feedback. The input stage is followed by the volume control, and a single stage incorporating the tone control circuits.

Provision is made for the following inputs:
1. Crystal pick-up giving an output of 500 mV.
2. Magnetic pick-up with an inductance of 500 mH and giving an output of 5 mV.
3. Radio tuner unit.

*Performance*

| | |
|---|---|
| Sensitivity (magnetic pick-up position): | 5 mV input for 140 µA output into <470 Ω |
| Distortion: | |
| 5 mV input | 0·05% |
| 50 mV input | 0·30% |
| (overload of 10 times) | |
| Tone control characteristics: | |
| Bass | +12 dB to −13 dB at 100 c/s |
| Treble | + 8 dB to −10 dB at 10 kc/s |
| Power supply requirement: | 24 mA at 28 V |

150

# CONSTRUCTION

WHEN an amplifier circuit has been developed, the amplifier is built up in prototype form and tested. After the complete circuit ha emerged, the final layout of the amplifier is considered. The form will depend upon many factors including cost and the number of amplifiers to be produced.

The basic requirement is a chassis to hold the components in convenient positions relative to one another; but the chassis may also be required to screen hum-sensitive circuits and to provide mechanical protection to delicate components. There is no reason why the form of the chassis, whilst performing these functional tasks, should not also be made attractive in appearance, and a good design will normally encourage the taking of the trouble required to present a well-finished product.

The whole of the amplifier from input connection to loudspeaker connection, including supply unit, may be constructed on one chassis provided care is taken in layout and the input stages do not pick up excessive hum signals from the mains transformer. This form is most commonly used for portable amplifiers contained in strong carrying cases.

The normal high fidelity amplifier is made in two sections, the control unit and the main amplifier. All controls and associated circuits are placed in the control unit, including a mains switch wired to the main amplifier. The control unit takes its heater and h.t. supplies from the main amplifier via cables and the input signal to the main amplifier is carried from the preamplifier via screened cable. As the screened cable has an outer earthed braiding, no other h.t. negative connection is required and indeed, if the chassis of the control unit and the main amplifier were connected via an additional 'earth', a loop would be formed. It should be noted that, if the supplies and signal cables are separate, there is the danger of receiving an electric shock if the equipment is switched on with the signal

151

cable disconnected. In this case, the chassis of the control unit is virtually at h.t. positive potential above earth.

### Chassis Forms

The most commonly used chassis form for the main amplifier is the 'inverted tray' made by 'cropping' the corners of a sheet of metal and bending to form sides. The large components and valves are mounted on the top surface of the tray and the small components are contained within, the sides being utilized for connecting sockets and controls.

In mass production, steel is a convenient chassis material being relatively cheap, strong and suitable for working with press tools and welding. For the amateur constructor, aluminium is a better material being more easily worked although, compared with steel, heavier gauges must be used to obtain the same strength. Aluminium has an advantage over steel in that it is non-magnetic and will not 'conduct' the magnetic field of the mains transformer to hum-sensitive parts of the amplifier.

A variation of the tray, suitable for small unit production, is the bending of the two opposite sides of a sheet of aluminium. The bending can be done easily in a large vice, and the bottom of the two sides tied by two square rods. This gives a rigid chassis form with the added advantage that the open sides allow free passage of air under the chassis, preventing overheating of the internal components. If the free edges of the chassis are given small 'sides', the same effect is achieved with added rigidity to the top, permitting perforated metal sides and a bottom to be fitted. A cover to protect valves and other components is sometimes provided, but this must be well ventilated. Carrying handles are also useful when the amplifier is large and heavy.

The chassis size will depend upon the layout of the component parts, which should be positioned so as to conform to as many as possible of the following requirements.

### Layout

The main transformer must be removed as far as possible from hum sensitive components (mainly the output transformer) and hum sensitive parts of the circuits (mainly grid circuits, and particularly the grid, cathode and anode circuits of the input valve).

The valves should be disposed so that associated components fall naturally into place and long connecting wires are avoided. This applies particularly to the anode-grid coupling components and wiring.

152

Valves and resistors which operate at high temperature should be well separated from each other and other components, and placed in positions where there is adequate ventilation.

Components associated with sockets and controls should be situated as close as possible to such sockets and controls.

A layout which fulfils all requirements is seldom possible and some compromise is necessary, particularly regarding the placing of the mains transformer. If it is acceptable to build the amplifier in two sections, the 'power unit', comprising mains transformer, rectifier,
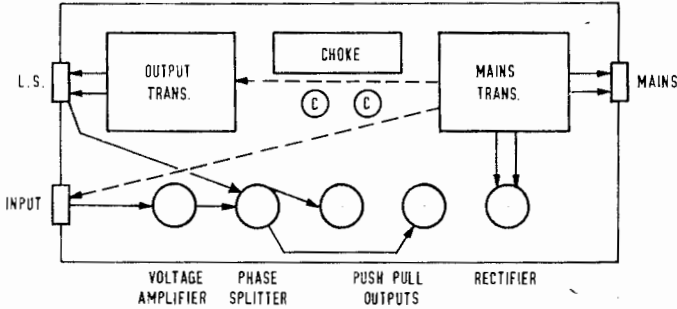


FIGURE 18.1
Amplifier layout. Full lines show short connections and dotted lines maximum spacing.

choke and smoothing capacitors, can be built on a separate chassis and coupled by means of a four-core cable with plugs and sockets.

As an example of a satisfactory layout, the 10 W design described in Chapter 11 should be studied. Here the mains transformer and output transformer are placed at opposite ends of the chassis, the input valve being in one corner as far removed from the mains transformer as possible (see Fig. 18.1). In this amplifier, negative feedback is applied from the output transformer secondary to the cathode of the second valve which is conveniently placed for this connection.

**Components**

Transformers are usually made in two forms: (1) complete units intended to be mounted above the chassis, or (2) with one cover cap so that the transformer core can be laid on the chassis and bolted to it. The latter method entails a rectangular cutout to accommodate windings. The first method is advisable when the chassis is steel, as this reduces circulating currents in the chassis arising from the field of the mains transformer. The second method is satisfactory when the

153

chassis is aluminium and has the advantage that the transformer connections are automatically brought under the chassis and can be efficiently and economically made by using the ends of the windings for direct connection to other components.

Smoothing and decoupling capacitors are invariably electrolytics in a metal or card tubular case, the smaller varieties having wire ends so that they are self-supporting.

Large electrolytics have tag connections and the case must be secured to the chassis by means of a circular metal 'clip'. Some care must be taken when this type of capacitor is used, as the negative connection is frequently not insulated from the metal casing and earth loops may result.

Resistors are normally provided with wire ends and are intended to be self-supporting. In all cases where small components have wire ends, it is preferable to wire direct to associated components. This reduces the number of connections and shortens the connecting wires, thus reducing stray capacitance and the possibility of electrostatic hum pickup.

When the weight of a component is sufficient to give rise to the possibility of strain if supported only by its wire ends, some form of mounting is advisable. In some cases it may be necessary only to provide connection tags so placed that the wire ends are reduced to a short length. In any layout it is preferable to avoid covering the underside of the valveholders with components, as this makes testing and replacement of faulty components difficult.

### Wiring

Components not provided with wire ends usually have tags or terminals to which a connection is made by means of tinned copper wire, over which is placed an appropriate length of Systoflex tubing. Alternatively, covered wire can be used and the covering stripped from the ends. In either case, connections are made by soldering except in the case of components fitted with screw terminals.

Soldering is preferred, as pressure contacts can deteriorate due to oxidization or the slackening of pressure due to vibration. Good soldering is essential, as poor joints can prevent correct operation, introducing resistance into the circuit or causing intermittent operation sometimes accompanied by noise. Whenever possible, a good mechanical joint should be made before soldering, as this holds the joint free from movement whilst the solder is cooling. All joints should be testing by pulling on the connecting wire to prove the soundness of the joint.

Valve heaters are normally connected in parallel to a 6·3 V winding on the mains transformer. The wire used for these connections should be of adequate gauge to prevent a voltage drop, bearing
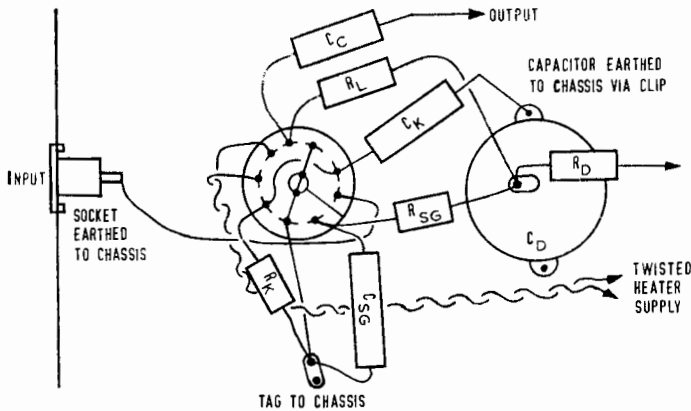
Practical layout of EF86 amplifier stage which is likely to be prone to hum due to unscreened grid circuit and earth loops in grid and cathode circuit.

in mind the total current at the transformer end of the pair may be 3 A or more. Electromagnetic hum from this supply is reduced by the cancelling effect obtained when they are twisted together. Where the signal voltage is low, i.e. less than 100 mV, it is desirable that the heater supply leads should be separated from the other wiring by taking them above the chassis to return at a point adjacent to the valveholder pins to be supplied.

**Earth Looping**

The earth connections of each stage require special consideration as hum due to earth looping can easily occur. This is particularly the case when earth connections are indiscriminately made to the chassis, a tempting method of reducing the amount of wiring. To avoid trouble from this effect, it is advisable to earth the chassis at a point adjacent to the rectifier capacitors only, with a good low resistance connection, and make all other earth connections to a 'bus-bar'. This bus-bar need not be physically apparent, although in some cases a sturdy length of bare copper wire is provided to emphasize its position and reduce resistance and inductive reactance.

The theory of earth looping was discussed in Chapter 14, and the stages where the signal is lowest are most prone to hum. Figs. 18.2

155

and 18.3 illustrate good and bad layout in the connections for a pentode voltage amplifier stage. Electrolytic capacitors mounted to the chassis by means of a clip should be considered as earthed at this point. This is satisfactory in the case of supply unit reservoir and smoothing capacitors, although a wire connection from the negative terminal to nut and bolt with a well-tightened shakeproof tag biting into the metal of the chassis is advisable. In all other cases, capacitors should be insulated from the chassis.
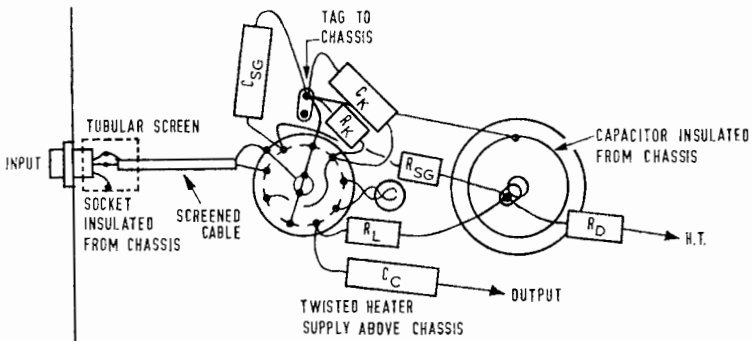


FIGURE 18.3

The same stage with improved layout. Note screening in grid circuit and disposition of wiring and components to avoid loops.

The signal input socket should be screened and insulated from the chassis. In low signal stages where the grid and anode connections exceed 1 inch in length, the connection is made by means of a screened cable which carries the earth bus-bar. If the bus-bar connection is not made via the cable, the screen braiding is connected to earth at one end only, to prevent looping.

**Finish**

Consistent with price, it should be arranged that adjustments to mains voltage tappings, load impedance changes, fuses and external connections may be made easily from outside the chassis without the need for tools, except that soldered connections to all plugs is preferred. Plugs should provide for cable clamping. Input sockets should be of the fully screened variety.

The chassis and components can be finished to present an attractive appearance in a number of ways, the actual finish depending on personal preference. A variety of cellulose finishes is available but steel or aluminium should be 'passivated', or otherwise protected from oxidization, before cellulose is applied. A well-finished, neatly-wired amplifier is likely to have been designed and constructed with more care than an untidy, cheaply-finished product.

# AMPLIFIERS FOR STEREOPHONY

STEREOPHONIC reproduction requires the simultaneous use of two identical channels. So far as the amplifier section of the system is concerned this means little more than the provision of two control units and two mains amplifiers.

## Chassis Form

For convenience, however, the two control units are usually built into one chassis with ganged controls and the two main amplifiers are similarly built on one chassis with a common supply unit.

Ganged potentiometers, i.e. two similar potentiometers operated by one spindle, should be accurately matched. Slight discrepancies in gain due to slight mismatch in the ganged potentiometer volume



(a)            (b)
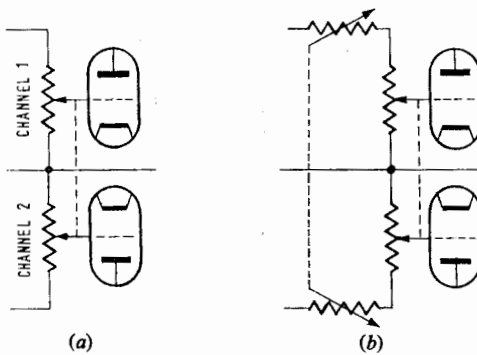
FIGURE 19.1

(a) Ganged volume control in stereo control unit.
(b) Ganged balance control added to volume control circuit.

control and dissimilar circuit gain, can be adjusted by means of a balance control. This control will enable other inequalities external to the amplifier to be corrected.

157

## Balancing

The simplest method of balancing is to add a special ganged variable resistor to the input side of the two volume controls as in Fig. 19.1. Unless these variable resistors are linear, one should be log graded and the other antilog. The two sections are arranged to be of equal



FIGURE 19.2

Balance control operating by variation of current feedback.
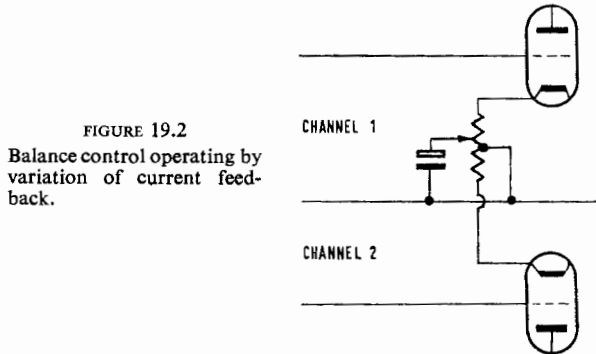
resistance at central position, one decreasing as the other increases upon operation of the control.

Balancing can be achieved by any method which will affect the gain of one channel relative to the other, and a variation of feedback will bring about this result. Whatever method is adopted, it is important to ensure that the balance control affects only the gain and has no unbalancing effect upon frequency response, input or output impedance, etc. A feedback balancing circuit is shown in Fig. 19.2, by which current feedback can be reduced in either channel, thus increasing gain. As this circuit operates by a variation of current



FIGURE 19.3

Two-pole, two-way switch used to reverse the connection to a loudspeaker for phase reversal.
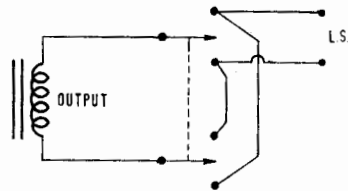
feedback the output impedance of the associated valves will vary and if such a circuit is adopted, it should be ascertained that there is no adverse effect upon performance.

Selector switches in the control unit can also be ganged and it is usual to provide a switch position whereby the inputs are coupled, so that both channels will operate from a single, i.e. monophonic,

158

source. Care should be taken that there is no 'breakthrough' from one channel to the other, particularly where parts of the two circuits are brought close together at ganged controls.

### Phase Reversal

A useful device is a phasing switch, which reverses the phase of one channel, enabling an audible check for correct phasing to be made. This can be done quite simply by arranging a switch to reverse the connections to one loudspeaker, as in Fig. 19.3. Other methods, such as the introduction of a phase reversing valve, are possible.

For the best stereophonic results, each channel should be designed to be a good monophonic reproducer. As there are two amplifiers in operation, the maximum power output per channel can be slightly reduced to save cost, but it must be remembered that there may be occasions when one channel is carrying full power.

# REFERENCE DATA

1. Voltage, Current and Power ratios expressed in decibels.
2. Linear and log. scales compared.
3. Voltage/Power for standard loads.
4. Current carrying capacity of standard resistors.
5. Reactance of standard capacitors.
6. Reactance of inductors.

1. *Voltage, Current and Power Ratios Expressed in Decibels*

| Voltage or Current Ratio | | dB | Power Ratio | |
|---|---|---|---|---|
| Up | Down | | Up | Down |
| 1·12 | ·89 | 1 | 1·26 | ·79 |
| 1·26 | ·79 | 2 | 1·58 | ·63 |
| 1·41 | ·71 | 3 | 1·99 | ·50 |
| 1·58 | ·63 | 4 | 2·51 | ·40 |
| 1·78 | ·56 | 5 | 3·16 | ·32 |
| 1·99 | ·50 | 6 | 3·98 | ·25 |
| 2·24 | ·45 | 7 | 5·01 | ·20 |
| 2·51 | ·40 | 8 | 6·31 | ·16 |
| 2·82 | ·35 | 9 | 7·94 | ·13 |
| 3·16 | ·32 | 10 | 10·00 | ·10 |
| 3·55 | ·28 | 11 | 12·59 | ·08 |
| 3·98 | ·25 | 12 | 15·85 | ·06 |
| 4·47 | ·22 | 13 | 19·95 | ·05 |
| 5·01 | ·20 | 14 | 25·12 | ·04 |
| 5·62 | ·18 | 15 | 31·62 | ·03 |
| 6·31 | ·16 | 16 | 39·81 | ·025 |
| 7·08 | ·14 | 17 | 50·12 | ·020 |
| 7·94 | ·13 | 18 | 63·10 | ·016 |
| 8·91 | ·11 | 19 | 79·43 | ·0126 |
| 10·00 | ·10 | 20 | 100·00 | ·0100 |
| 100 | ·01 | 40 | | |
| 1,000 | ·001 | 60 | | |
| 10,000 | ·0001 | 80 | | |
| 100,000 | ·00001 | 100 | | |

## 2. Linear and Logarithmic Scales Compared

Illustrating the log scale for frequencies ranging from 10 to 100 c/s.

| ·0 | | ·3 | | ·48 | ·6 | ·7 | ·78 | ·85 | ·9 | ·95 | 1·0 |
|----|---|----|---|-----|----|----|-----|-----|----|-----|-----|
| 10 | | 20 | | 30 | 40 | 50 | 60 | 70 | 80 | 90 | 100 |
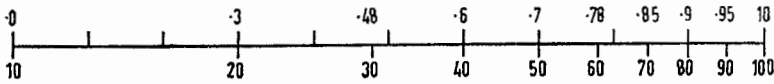
## 3. Voltage/Power for Standard Loads

| Measured Output (r.m.s. volts) | Power in load (Watts) | | |
|---|---|---|---|
| | $3\frac{3}{4}\,\Omega$ | $7\frac{1}{2}\,\Omega$ | $15\Omega$ |
| 1 | 0·266 | 0·133 | 0·067 |
| 2 | 1·07 | 0·52 | 0·27 |
| 3 | 2·4 | 1·2 | 0·60 |
| 4 | 4·3 | 2·1 | 1·07 |
| 5 | 6·7 | 3·3 | 1·57 |
| 6 | 9·6 | 4·8 | 2·4 |
| 7 | 13·1 | 6·5 | 3·3 |
| 8 | 17·1 | 8·5 | 4·3 |
| 9 | 21·6 | 10·8 | 5·4 |
| 10 | 26·6 | 13·3 | 6·6 |
| 11 | 32·2 | 16·1 | 8·1 |
| 12 | 38·4 | 19·2 | 9·6 |
| 13 | 45·0 | 22·5 | 11·3 |
| 14 | 52·3 | 26·2 | 13·1 |
| 15 | | 30·0 | 15·0 |
| 16 | | 34·1 | 17·0 |
| 17 | | 38·5 | 19·3 |
| 18 | | 43·2 | 21·6 |
| 19 | | 48·1 | 24·0 |
| 20 | | 53·5 | 26·6 |
| 21 | | | 29·4 |
| 22 | | | 32·2 |
| 23 | | | 35·3 |
| 24 | | | 38·4 |
| 25 | | | 41·7 |
| 26 | | | 45·1 |
| 27 | | | 48·6 |
| 28 | | | 52·2 |

L 4

## 4. *Current Carrying Capacity of Standard Resistors* (in mA)

| Ω | ½W | 1W | 2W | 3W | 5W | 10W |
|---|---|---|---|---|---|---|
| 10 | 224 | 316 | 447 | 548 | 707 | 1,000 |
| 15 | 185 | 258 | 366 | 447 | 578 | 817 |
| 22 | 151 | 213 | 310 | 380 | 477 | 675 |
| 33 | 123 | 174 | 246 | 302 | 389 | 550 |
| 47 | 108 | 146 | 206 | 253 | 326 | 461 |
| 68 | 86 | 121 | 173 | 210 | 272 | 384 |
| 100 | 71 | 100 | 141 | 176 | 224 | 316 |
| 150 | 58 | 82 | 115 | 143 | 182 | 258 |
| 220 | 48 | 67 | 95 | 117 | 151 | 213 |
| 330 | 39 | 55 | 78 | 95 | 123 | 174 |
| 470 | 33 | 46 | 65 | 80 | 103 | 146 |
| 680 | 27 | 38 | 54 | 66 | 86 | 121 |
| 1,000 | 22 | 32 | 45 | 55 | 71 | 100 |
| 1,500 | 18 | 26 | 36 | 45 | 58 | 82 |
| 2,200 | 15 | 21 | 31 | 38 | 48 | 67 |
| 3,300 | 12 | 17 | 25 | 30 | 39 | 55 |
| 4,700 | 10 | 14 | 21 | 25 | 33 | 46 |
| 6,800 | 8·6 | 12 | 17 | 21 | 27 | 38 |
| 10 k | 7·1 | 10 | 14 | 17 | 22 | 32 |
| 15 k | 5·8 | 8·2 | 11 | 14 | 18 | 26 |
| 22 k | 4·8 | 6·7 | 9·5 | 12 | 15 | 21 |
| 33 k | 3·9 | 5·5 | 7·8 | 9 | 12 | 17 |
| 47 k | 3·3 | 4·6 | 6·5 | 8 | 10 | 14 |
| 68 k | 2·7 | 3·8 | 5·4 | 7 | 8·6 | 12 |
| 100 k | 2·2 | 3·2 | 4·5 | 5·5 | 7·1 | 10 |
| 150 k | 1·8 | 2·6 | 3·6 | | | |
| 220 k | 1·5 | 2·1 | 3·1 | | | |
| 330 k | 1·2 | 1·7 | 2·5 | | | |
| 470 k | 1·0 | 1·4 | 2·1 | | | |
| 680 k | 0·86 | 1·2 | 1·7 | | | |
| 1 M | 0·71 | 1·0 | 1·4 | | | |
| 1·5 M | 0·58 | ·82 | 1·1 | | | |
| 2·2 M | 0·48 | ·67 | ·95 | | | |
| 3·3 M | 0·39 | ·55 | ·78 | | | |
| 4·7 M | ·33 | ·46 | ·65 | | | |
| 6·8 M | ·27 | ·38 | ·54 | | | |
| 10 M | ·22 | ·32 | ·45 | | | |

## 5. Reactance of Standard Capacitors (Ohms)

| c/s | 10 pF | 20 pF | 30 pF | 50 pF | 75 pF |
|---|---|---|---|---|---|
| 10 | | | | | |
| 20 | | | | | |
| 30 | | | | | 70·7 M |
| 50 | | | | 63·6 M | 42·4 M |
| 70 | | | 75·7 M | 45·4 M | 30·3 M |
| 100 | | 79·5 M | 53 M | 31·8 M | 21·2 M |
| 200 | 79·5 M | 39·8 M | 26·5 M | 15·9 M | 10·6 M |
| 300 | 53 M | 26·5 M | 17·7 M | 10·6 M | 7·07 M |
| 500 | 31·8 M | 15·9 M | 10·6 M | 6·36 M | 4·24 M |
| 700 | 22·9 M | 11·3 M | 7·57 M | 4·54 M | 3·03 M |
| 1,000 | 15·8 M | 7·95 M | 5·31 M | 3·18 M | 2·12 M |
| 2,000 | 7·95 M | 3·98 M | 2·65 M | 1·59 M | 1·06 M |
| 3,000 | 5·3 M | 2·65 M | 1·77 M | 1·06 M | 707 k |
| 5,000 | 3·18 M | 1·59 M | 1·06 M | 636 k | 424 k |
| 7,000 | 2·29 M | 1·13 M | 757 k | 454 k | 303 k |
| 10,000 | 1·59 M | 759 k | 532 k | 318 k | 212 k |
| 20,000 | 795 k | 398 k | 265 k | 159 k | 106 k |
| 30,000 | 530 k | 265 k | 177 k | 106 k | 70·7 k |
| 50,000 | 318 k | 159 k | 106 k | 63·6 k | 42·4 k |
| 70,000 | 229 k | 113 k | 75·7 k | 45·4 k | 30·3 k |
| 100,000 | 159 k | 79·5 k | 53 k | 31·8 k | 21·2 k |

| c/s | 100 pF | 120 pF | 150 pF | 180 pF | 220 pF |
|---|---|---|---|---|---|
| 10 | | | | 88·3 M | 72·3 M |
| 20 | 79·5 M | 66·3 M | 53 M | 44·1 M | 36·1 M |
| 30 | 53 M | 44·1 M | 35·3 M | 29·4 M | 24·1 M |
| 50 | 31·8 M | 26·5 M | 21·2 M | 17·7 M | 14·5 M |
| 70 | 22·9 M | 18·9 M | 15·1 M | 12·6 M | 10·3 M |
| 100 | 15·9 M | 13·2 M | 10·6 M | 8·83 M | 7·23 M |
| 200 | 7·95 M | 6·63 M | 5·3 M | 4·41 M | 3·61 M |
| 300 | 5·3 M | 4·41 M | 3·53 M | 2·94 M | 2·41 M |
| 500 | 3·18 M | 2·65 M | 2·12 M | 1·77 M | 1·45 M |
| 700 | 2·29 M | 1·89 M | 1·51 M | 1·26 M | 1·03 M |
| 1,000 | 1·59 M | 1·32 M | 1·06 M | 883 k | 723 k |
| 2,000 | 795 k | 663 k | 530 k | 441 k | 361 k |
| 3,000 | 530 k | 441 k | 353 k | 294 k | 241 k |
| 5,000 | 318 k | 265 k | 212 k | 177 k | 145 k |
| 7,000 | 229 k | 189 k | 151 k | 126 k | 103 k |
| 10,000 | 159 k | 132 k | 106 k | 88·3 k | 72·3 k |
| 20,000 | 79·5 k | 66·3 k | 53 k | 44·1 k | 36·1 k |
| 30,000 | 53 k | 44·1 k | 35·3 k | 29·4 k | 24·1 k |
| 50,000 | 31·8 k | 26·5 k | 21·2 k | 17·7 k | 14·5 k |
| 70,000 | 22·9 k | 18·9 k | 15·1 k | 12·6 k | 10·3 k |
| 100,000 | 15·9 k | 13·4 k | 10·6 k | 8·83 k | 7·23 k |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | 270 pF | 330 pF | 390 pF | 470 pF | 680 pF |
|---|---|---|---|---|---|
| 10 | 58·9  M | 48·2  M | 40·7  M | 33·8  M | 23·4  M |
| 20 | 29·4  M | 24·1  M | 20·4  M | 16·9  M | 17·7  M |
| 30 | 19·6  M | 16·1  M | 13·6  M | 11·3  M | 7·8  M |
| 50 | 11·8  M | 9·64 M | 8·15 M | 6·77 M | 4·67 M |
| 70 | 8·41 M | 6·85 M | 5·82 M | 4·83 M | 3·34 M |
| 100 | 5·89 M | 4·82 M | 4·07 M | 3·38 M | 2·34 M |
| 200 | 2·94 M | 2·41 M | 2·04 M | 1·69 M | 1·17 M |
| 300 | 1·96 M | 1·61 M | 1·36 M | 1·13 M | 780 k |
| 500 | 1·18 M | 964 k | 815 k | 677 k | 467 k |
| 700 | 841 k | 685 k | 582 k | 483 k | 334 k |
| 1,000 | 589 k | 482 k | 407 k | 338 k | 234 k |
| 2,000 | 294 k | 241 k | 204 k | 169 k | 117 k |
| 3,000 | 196 k | 161 k | 136 k | 113 k | 78 k |
| 5,000 | 118 k | 96·4 k | 81·5 k | 67·7 k | 46·7 k |
| 7,000 | 84·1 k | 68·5 k | 58·2 k | 48 3 k | 33·4 k |
| 10,000 | 58·9 k | 48·2 k | 40·7 k | 33·8 k | 23·4 k |
| 20,000 | 29·4 k | 24·1 k | 20·4 k | 16·9 k | 11·7 k |
| 30,000 | 19·6 k | 16·1 k | 13·6 k | 11·3 k | 7·80 k |
| 50,000 | 11·8 k | 9·64 k | 8·15 k | 6·77 k | 4·67 k |
| 70,000 | 8·41 k | 6·85 k | 5·82 k | 4·83 k | 3·34 k |
| 100,000 | 5·89 k | 4·82 k | 4·07 k | 3·38 k | 2·34 k |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | 800 pF | 1000 pF | ·002 $\mu$F | ·003 $\mu$F | ·005 $\mu$F |
|---|---|---|---|---|---|
| 10 | 19·2 M | 15·9 M | 7·95 M | 5·3 M | 3·18 M |
| 20 | 9·6 M | 7·95 M | 3·98 M | 2·65 M | 1·59 M |
| 30 | 6·4 M | 5·3 M | 2·65 M | 1·77 M | 1·06 M |
| 50 | 3·8 M | 3·18 M | 1·59 M | 1·06 M | 636 k |
| 70 | 2·73 M | 2·29 M | 1·13 M | 757 k | 454 k |
| 100 | 1·92 M | 1·59 M | 795 k | 530 k | 318 k |
| 200 | 960 k | 795 k | 398 k | 265 k | 159 k |
| 300 | 640 k | 530 k | 265 k | 177 k | 106 k |
| 500 | 380 k | 318 k | 159 k | 106 k | 63·6 k |
| 700 | 273 k | 229 k | 113 k | 75·7 k | 45·4 k |
| 1,000 | 192 k | 159 k | 79·5 k | 53 k | 31·8 k |
| 2,000 | 96 k | 79·5 k | 39·8 k | 26·5 k | 15·9 k |
| 3,000 | 64 k | 53 k | 26·5 k | 17·7 k | 10·6 k |
| 5,000 | 38 k | 31·8 k | 15·9 k | 10·6 k | 6·63 k |
| 7,000 | 27 k | 22·9 k | 11·3 k | 7·57 k | 4·54 k |
| 10,000 | 19 k | 15·9 k | 7·95 k | 5·3 k | 3·18 k |
| 20,000 | 9·6 k | 7·9 k | 3·98 k | 2·65 k | 1·59 k |
| 30,000 | 6·4 k | 5·3 k | 2·65 k | 1·77 k | 1·06 k |
| 50,000 | 3·8 k | 3·18 k | 1·59 k | 1·06 k | 636 |
| 70,000 | 2·7 k | 2·29 k | 1·13 k | 757 | 454 |
| 100,000 | 1·9 k | 1·59 k | 795 | 530 | 318 |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | ·01 $\mu$F | ·02 $\mu$F | ·05 $\mu$F | ·1 $\mu$F | ·25 $\mu$F |
|---|---|---|---|---|---|
| 10 | 1·59 M | 795 k | 318 k | 159 k | 63·6 k |
| 20 | 795 k | 398 k | 159 k | 79·5 k | 31·8 k |
| 30 | 530 k | 265 k | 106 k | 53 k | 21·2 k |
| 50 | 318 k | 159 k | 63·6 k | 31·8 k | 12·7 k |
| 70 | 229 k | 113 k | 45·4 k | 22·9 k | 9·08 k |
| 100 | 159 k | 79·5 k | 31·8 k | 15·9 k | 6·36 k |
| 200 | 795 k | 39·8 k | 15·9 k | 7·95 k | 3·18 k |
| 300 | 53 k | 26·5 k | 10·6 k | 5·3 k | 2·12 k |
| 500 | 31·8 k | 15·9 k | 6·36 k | 3·18 k | 1·27 k |
| 700 | 22·9 k | 11·3 k | 4·54 k | 2·29 k | 908 |
| 1,000 | 15·9 k | 7·95 k | 3·18 k | 1·59 k | 636 |
| 2,000 | 7·95 k | 3·98 k | 1·59 k | 795 | 318 |
| 3,000 | 5·3 k | 2·65 k | 1·06 k | 530 | 212 |
| 5,000 | 3·18 k | 1·59 k | 636 | 318 | 127 |
| 7,000 | 2·29 k | 1·13 k | 454 | 229 | 90·8 |
| 10,000 | 1·59 k | 795 | 318 | 159 | 63·6 |
| 20,000 | 790 | 398 | 159 | 79 | 31·8 |
| 30,000 | 530 | 265 | 106 | 53 | 21·2 |
| 50,000 | 318 | 159 | 63·6 | 31·8 | 12·7 |
| 70,000 | 229 | 113 | 45·4 | 22·9 | 9·8 |
| 100,000 | 159 | 79·5 | 31·8 | 15·9 | 6·36 |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | ·5 μF | 1 μF | 2 μF | 3 μF | 4 μF |
|---|---|---|---|---|---|
| 10 | 31·8 k | 15·9 k | 7·95 k | 5·30 k | 3·98 k |
| 20 | 15·9 k | 7·95 k | 3·98 k | 2·65 k | 1·97 k |
| 30 | 10·6 k | 5·3 k | 2·65 k | 1·77 k | 1·35 k |
| 50 | 6·36 k | 3·18 k | 1·59 k | 1·06 k | 795 |
| 70 | 5·54 k | 2·29 k | 1·13 k | 757 | 568 |
| 100 | 3·18 k | 1·59 k | 795 | 530 | 398 |
| 200 | 1·59 k | 795 | 398 | 265 | 197 |
| 300 | 1·06 k | 530 | 265 | 179 | 135 |
| 500 | 636 | 318 | 159 | 106 | 79·5 |
| 700 | 454 | 229 | 113 | 75 | 56·8 |
| 1,000 | 318 | 159 | 79·5 | 53 | 39·8 |
| 2,000 | 159 | 79·5 | 39·8 | 26·5 | 19·7 |
| 3,000 | 106 | 53 | 26·5 | 17·7 | 13·5 |
| 5,000 | 63·6 | 31·8 | 15·9 | 10·6 | 7·95 |
| 7,000 | 45·4 | 22·9 | 11·3 | 7·57 | 5·68 |
| 10,000 | 31·8 | 15·9 | 7·95 | 5·3 | 3·98 |
| 20,000 | 15·9 | 7·9 | | | |
| 30,000 | 10·6 | 5·3 | | | |
| 50,000 | 6·36 | 3·18 | | | |
| 70,000 | 4·54 | 2·29 | | | |
| 100,000 | 3·18 | 1·59 | | | |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | 5 $\mu$F | 8 $\mu$F | 10 $\mu$F | 16 $\mu$F | 25 $\mu$F |
|---|---|---|---|---|---|
| 10 | 3·18 k | 1·98 k | 1·59 k | 994 | 636 |
| 20 | 1·59 k | 994 | 795 | 497 | 318 |
| 30 | 1·06 k | 662 | 530 | 331 | 212 |
| 50 | 636 | 398 | 318 | 199 | 127 |
| 70 | 454 | 284 | 229 | 142 | 90·8 |
| 100 | 318 | 198 | 159 | 99·4 | 63·6 |
| 200 | 159 | 99·4 | 79·5 | 49·7 | 31·8 |
| 300 | 106 | 66·2 | 53 | 33·1 | 21·2 |
| 500 | 63·6 | 39·8 | 31·8 | 19·9 | 12·7 |
| 700 | 45·4 | 28·4 | 22·9 | 14·2 | 9·08 |
| 1,000 | 31·8 | 19·8 | 15·9 | 9·94 | 6·36 |
| 2,000 | 15·9 | 9·94 | 7·95 | 4·97 | 3·18 |
| 3,000 | 10·6 | 6·62 | 5·3 | 3·31 | 2·12 |
| 5,000 | 6·36 | 3·98 | 3·18 | 1·99 | 1·27 |
| 7,000 | 4·54 | 2·84 | 2·29 | 1·42 | |
| 10,000 | 3·18 | 1·98 | 1·59 | | |

*Reactance of Standard Capacitors (cont'd.)*

| c/s | 32 $\mu$F | 50 $\mu$F | 64 $\mu$F | 100 $\mu$F | 500 $\mu$F |
|---|---|---|---|---|---|
| 10 | 497 | 318 | 248 | 159 | 31·8 |
| 20 | 248 | 159 | 124 | 79·5 | 24·8 |
| 30 | 166 | 106 | 82·8 | 53 | 16·6 |
| 50 | 99·4 | 63·6 | 49·7 | 31·8 | 9·94 |
| 70 | 71 | 45·4 | 35·4 | 22·9 | 7·1 |
| 100 | 49·7 | 31·8 | 24·8 | 15·9 | 4·97 |
| 200 | 24·8 | 15·9 | 12·4 | 7·95 | 2·42 |
| 300 | 16·6 | 10·6 | 8·28 | 5·3 | 1·66 |
| 500 | 9·94 | 6·36 | 4·97 | 3·18 | |
| 700 | 7·1 | 4·54 | 3·54 | 2·29 | |
| 1,000 | 4·97 | 3·18 | 2·48 | 1·59 | |
| 2,000 | 2·48 | 1·59 | 1·24 | | |
| 3,000 | 1·66 | 1·06 | | | |
| 5,000 | | | | | |
| 7,000 | | | | | |
| 10,000 | | | | | |

## 6. Reactance of Inductors (Ohms)

| mH | 50 | 100 | 300 | 500 | 1,000 c/s |
|---|---|---|---|---|---|
| 1 | ·314 | ·628 | 1·88 | 3·14 | 6·28 |
| 2 | ·628 | 1·26 | 2·76 | 6·28 | 12·6 |
| 5 | 1·57 | 3·14 | 9·4 | 15·7 | 31·4 |
| 10 | 3·14 | 6·28 | 18·8 | 31·4 | 62·8 |
| 25 | 7·85 | 15·7 | 47 | 78·5 | 157 |
| 50 | 15·7 | 31·4 | 94 | 157 | 314 |
| 100 | 31·4 | 62·8 | 188 | 314 | 628 |

| mH | 5,000 | 10,000 | 50,000 | 100,000 c/s |
|---|---|---|---|---|
| 1 | 31·4 | 62·8 | 314 | 628 |
| 2 | 62·8 | 126 | 628 | 1,356 |
| 5 | 157 | 314 | 1,570 | 3,140 |
| 10 | 314 | 628 | 3,140 | 6,280 |
| 25 | 785 | 1,570 | 7,850 | 15,700 |
| 50 | 1,570 | 3,140 | 15,700 | 31,400 |
| 100 | 3,140 | 6,280 | 31,400 | 62,800 |

| H | 50 | 100 c/s |
|---|---|---|
| 1 | 314 | 628 |
| 2·5 | 785 | 1,570 |
| 5 | 1,570 | 3,140 |
| 10 | 3,140 | 6,280 |
| 20 | 6,280 | 12,600 |
| 50 | 15,700 | 31,400 |
| 100 | 31,400 | 62,800 |

# SELECTED VALVE APPLICATION REPORTS

## GENERAL OPERATIONAL CONDITIONS FOR VALVES

**Heater.** The heater voltage must be within $\pm 7\%$ of the rated value.

**Anode and screen.** The absolute-maximum ratings are limiting values which should not be exceeded under the worst probable conditions.

**Voltage between heater and cathode.** The maximum values of cathode-to-heater voltage are the maximum d.c. values quoted. Where a.c. and d.c. exist between heater and cathode (e.g. the cathode follower phase splitter) the d.c. component must not exceed the published value and in addition the maximum instantaneous value occurring must never exceed twice the published value or 300 V whichever is the lesser.

**External resistance between cathode and heater.** When cathode resistors of high value are used, the valve performance may be influenced by leakage between heater and cathode. A maximum value of 20 k$\Omega$ is therefore recommended although this may be increased to 1 M$\Omega$ if the d.c. component of the cathode-to-heater voltage is such that its instantaneous value never drops below three times the r.m.s. value of the heater voltage. The hum voltage produced across the resistance might assume a rather high value under these conditions.

**Rectifier cathodes.** Disintegration of the cathode coating may occur in both indirectly heated and directly heated rectifiers, if the total resistance in series with the anode is less than that specified in the Application Report. The value of this resistor depends upon the effective resistance, $Rt$, due to the mains transformer and $Rt = R_s + n^2 R_p$ where $R_s =$ the resistance of the transformer secondary in the anode circuit, $Rp =$ resistance of the transformer primary and $n =$ secondary to primary ratio, i.e.: secondary voltage/primary voltage. If the resistance $R_t$ is less than the minimum specified value, an additional series resistor must be included in the anode circuit. The wattage of this resistor should be at least three times that required for the d.c. current only.

**Control Grid.** It is good practice to keep the resistance of the circuit between control grid and cathode as low as possible. It should not exceed the maximum quoted in the Application Report, which is for fixed bias operation unless otherwise stated. The maximum value with cathode bias may be 2 or 3 times that quoted for fixed bias operation.

Although the cathode bias method is most frequently adopted, 'grid current biasing' may be employed. This relies upon the flow of current in the grid resistor developing a negative grid voltage, and in this case the grid resistor should be high (preferably greater than 10 M$\Omega$ but not

greater than 22 mΩ). The grid should not be allowed to become positive with respect to cathode.

Symbols Used in Application Report Data

$Va(b)$max.  The maximum voltage which can be applied to the anode when the valve is cold.

$Vg1(Ig1 = +0\cdot3\ \mu A)$ The control grid voltage at which the positive grid current i.s $0\cdot3\ \mu A$.

$\mu g1$–$g2$.  The 'inner—mu' is the amplification factor from control grid to screen grid $= \dfrac{\delta Vg2}{\delta Vg1}$ ($I_k$ constant).

$R_{eq}$.  The value of resistance which, if introduced into the grid circuit of a perfectly noiseless valve, would produce noise of the same level as that of the shot and partition noise occurring in the actual valve. It does not include flicker effect.

Noise factor.  The ratio of the signal-to-noise ratio at the input, to the signal-to-noise ratio at the output.

$c_{in}$.  Capacitance measured between the grid and other electrodes except the anode.

$c_{out}$.  Capacitance measured between the anode and other electrodes except the grid.

**Other symbols.** In this book, signal voltages have been indicated by $E$ and all other voltages by $V$. The anode resistance has been indicated by $R_L$. Other symbols are sometimes used but the reader should have no difficulty in recognizing $V_{out}$ and $V_{in}$ as output and input signal voltages or $R_a$ as the anode load resistance when these occur in the Application Reports and in other works.

**Valve Nomenclature.** One system of indicating valve types, consists of two or more letters followed by figures. The code is as follows:—

First letter gives the heater voltage. E = $6\cdot3$ V; G = 5 V.

Second letter shows the class of valve. C = triode; CC = double triode; F = voltage amplifying pentode; L = output pentode; Z = full wave rectifier; CL = triode and output pentode.

First figure gives the type of base. 3 = octal base; 8 = noval base (B9A).

Second and third figures are the serial numbers of design development; 0—1—2—3, etc.

The diagrams and data are reproduced by permission of the respective manufacturers and is gratefully acknowledged. The manufacturers concerned are:—

*Mullards Ltd.* EF86, ECC81, ECC82, ECC83, ECL86, EL84, EL34, GZ34, EZ80, EZ81.

*General Electric Co. Ltd. of England.* KT66, KT88, GZ37/U54.

## VOLTAGE AMPLIFYING PENTODE

*Low-noise pentode for use as r.c. coupled a.f. voltage amplifier, particularly suitable for the early stages of high-gain audio amplifiers, microphone pre-amplifiers and tape recorders.*

## Valve Data

| Heater (suitable for series or parallel operation, a.c. or d.c.) | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 200 | mA |

| Capacitances (measured without an external shield) | | |
|---|---|---|
| $c_{out}$ | 5·3 | pF |
| $c_{in}$ | 3·8 | pF |
| $c_{a-g1}$ | $<50$ | mpF |
| $c_{a1-h}$ | $<2·5$ | mpF |

| Characteristics | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_{g3}$ | 0 | V |
| $V_{g2}$ | 140 | V |
| $I_a$ | 3·0 | mA |
| $I_{g2}$ | 600 | $\mu$A |
| $V_{g1}$ | −2·0 | V |
| $g_m$ | 2·0 | mA/V |
| $r_a$ | 2·5 | M$\Omega$ |
| $\mu_{g1-g2}$ | 38 | |
| $V_{g1}$ max. ($I_{g1} = +0·3 \mu$A) | −1·3 | V |

## Operating Conditions as R.C. Coupled A.F. Amplifier

| Pentode connection | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| $V_b$ (V) | $R_a$ (k$\Omega$ ) | $R_{g2}$ (M$\Omega$) | $R_k$ (k$\Omega$) | $I_a$ (mA) | $I_{g2}$ ($\mu$A) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out}*$ ($V_{r.m.s.}$) | $R_{g1}†$ (k$\Omega$) |
| 400 | 100 | 0·39 | 1·0 | 2·6 | 550 | 137 | 95 | 330 |
| 350 | 100 | 0·39 | 1·0 | 2·3 | 480 | 130 | 81 | 330 |
| 300 | 100 | 0·39 | 1·0 | 1·95 | 400 | 127 | 68 | 330 |
| 250 | 100 | 0·39 | 1·0 | 1·6 | 350 | 122 | 53 | 330 |
| 200 | 100 | 0·39 | 1·0 | 1·35 | 280 | 115 | 38 | 330 |
| 150 | 100 | 0·47 | 1·5 | 0·8 | 180 | 104 | 26 | 330 |
| 400 | 220 | 1·0 | 2·2 | 1·2 | 250 | 213 | 80 | 680 |
| 350 | 220 | 1·0 | 2·2 | 1·1 | 220 | 208 | 70 | 680 |
| 300 | 220 | 1·0 | 2·2 | 0·9 | 190 | 202 | 59 | 680 |
| 250 | 220 | 1·0 | 2·2 | 0·8 | 160 | 196 | 48 | 680 |
| 200 | 220 | 1·0 | 2·2 | 0·7 | 120 | 188 | 37 | 680 |
| 150 | 220 | 1·0 | 2·7 | 0·5 | 100 | 165 | 26 | 680 |

*Output voltage at $D_{tot} = 5\%$.      †Grid resistor of following valve.

| | | Triode connection ($g_2$ connected to a, $g_3$ to k) | | | | | |
|---|---|---|---|---|---|---|---|
| $V_b$ (V) | $R_a$ (kΩ) | $I_a$ (mA) | $R_k$ (kΩ) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out}{}^*$ ($V_{r.m.s.}$) | $D_{tot}{}^*$ (%) | $R_{g1}$† (kΩ) |
| 400 | 47 | 3·5 | 1·2 | 27·3 | 68 | 5·2 | 150 |
| 350 | 47 | 3·0 | 1·2 | 27 | 57 | 5·0 | 150 |
| 300 | 47 | 2·6 | 1·2 | 26·7 | 47 | 4·6 | 150 |
| 250 | 47 | 2·2 | 1·2 | 26·5 | 36·5 | 4·3 | 150 |
| 200 | 47 | 1·8 | 1·2 | 25·8 | 26·5 | 3·9 | 150 |
| 400 | 100 | 1·9 | 2·2 | 31·5 | 77 | 4·8 | 330 |
| 350 | 100 | 1·65 | 2·2 | 31 | 65 | 4·7 | 330 |
| 300 | 100 | 1·45 | 2·2 | 30·5 | 54 | 4·6 | 330 |
| 250 | 100 | 1·2 | 2·2 | 30·5 | 42 | 4·2 | 330 |
| 200 | 100 | 1·0 | 2·2 | 30 | 31 | 4·0 | 330 |
| 400 | 220 | 1·0 | 3·9 | 33·2 | 78 | 4·3 | 680 |
| 350 | 220 | 0·9 | 3·9 | 33 | 66 | 4·3 | 680 |
| 300 | 220 | 0·8 | 3·9 | 32·8 | 54 | 4·2 | 680 |
| 250 | 220 | 0·68 | 3·9 | 32·5 | 43 | 3·9 | 680 |
| 200 | 220 | 0·55 | 3·9 | 32 | 31 | 3·7 | 680 |

*Output voltage and distortion at the start of positive grid current. At lower output voltages the distortion is approximately proportional to the voltage.
†Grid resistor of the following valve.

## Limiting Values

| | | |
|---|---|---|
| $V_{a(b)}$ max. | 550 | V |
| $V_a$ max. | 300 | V |
| $p_a$ max. | 1·0 | W |
| $V_{g2(b)}$ max. | 550 | V |
| $V_{g2}$ max. | 200 | V |
| $p_{g2}$ max. | 200 | mW |
| $I_k$ max. | 6·0 | mA |
| $R_{g1-k}$ max. ($p_a > 200$ mW) | 3·0 | MΩ |
| $R_{g1-k}$ max. ($p_a < 200$ mW) | 10 | MΩ |
| $V_{h-k}$ max. (cathode positive) | 100 | V |
| $V_{h-k}$ max. (cathode negative) | 50 | V |
| *$R_{h-k}$ max. | 20 | kΩ |

*When used as a phase inverter immediately preceding the output stage, $R_{h-k}$ max. may be 120 kΩ.

## OPERATING NOTES

### 1. Hum

When used as a normal voltage amplifier with a line voltage of 250 V, an anode load of 100 kΩ and a grid resistor of 470 kΩ the maximum hum level of the valve alone is 5 $\mu$V, the average value being about 3 $\mu$V when operated with one side of the heater earthed. This can be further reduced by centre-tapping the

heater to earth. Under these conditions the nominal hum level is 1 $\mu$V. The low level of hum attained with this valve can be completely masked by that due to an unsuitable valveholder, in which excessive leakage and capacitive coupling between pins will introduce considerable hum.

## 2. Noise

The low-frequency noise generated by a valve is most conveniently specified as an equivalent voltage on the control grid for a specific bandwidth. For the EF86 under normal conditions,



B9A Base
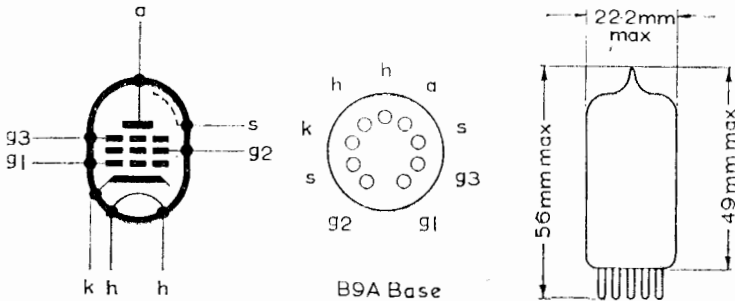
i.e. line voltage of 250 V and an anode load of 100 k$\Omega$, the equivalent noise voltage is approximately 2 $\mu$V for the frequency range of 25 to 10,000 c/s.

## 3. Microphony

Care in the design of the valve to ensure that the electrode structure and its mounting are as rigid as possible has reduced the microphony of the EF86 to a very low level. There are no appreciable internal resonances at frequencies below 1,000 c/s. At higher frequencies the effect of vibration is usually negligible on account of the damping provided by the chassis and the valveholder. In high-gain applications such as tape recording care should be taken in siting the valve, particularly when a loudspeaker is present in the same cabinet or when a motor is mounted on the same chassis. In such cases a flexible mounting for the valve holder or a separate weighted sub-chassis is advisable.
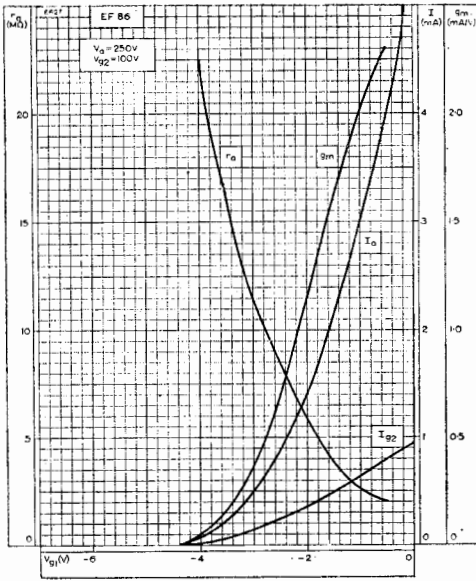
176

Anode and screen-grid currents, anode impedance and mutual
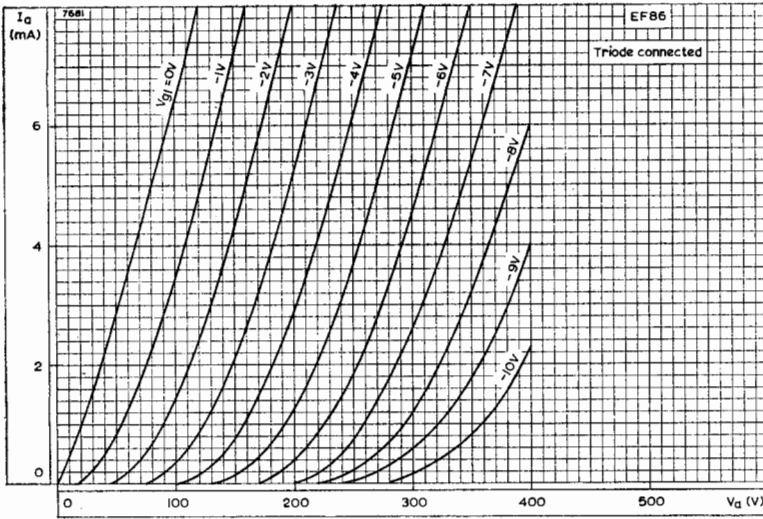conductance plotted against control-grid voltage.



Anode and screen-grid currents, anode impedance and mutual
conductance plotted against control-grid voltage.

M A
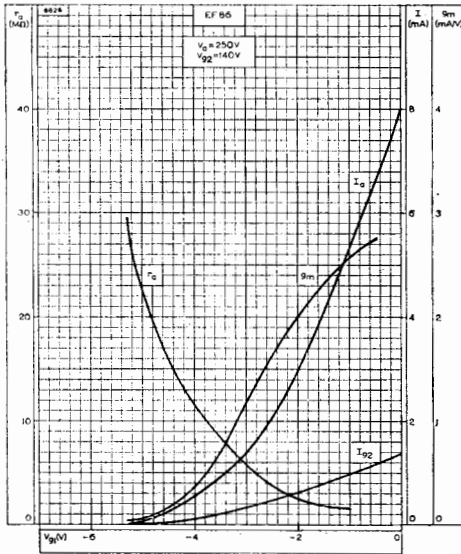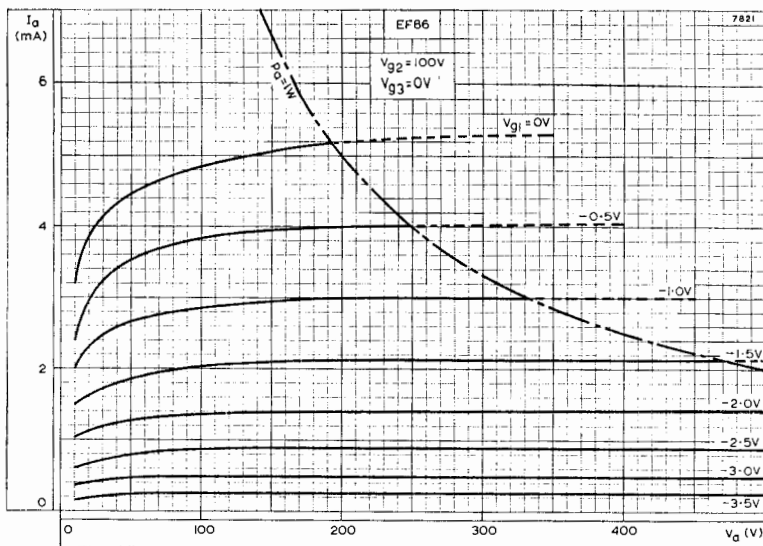
Anode current plotted against anode voltage with control-grid voltage as parameter. $V_{g2} = 100$ V.
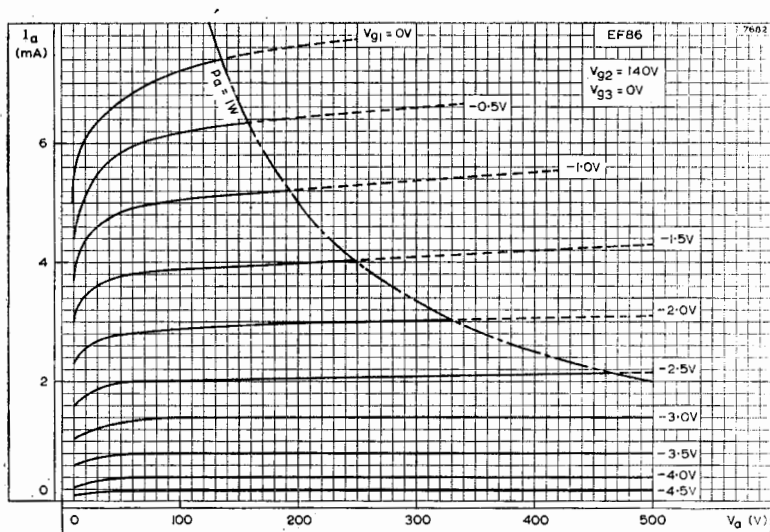


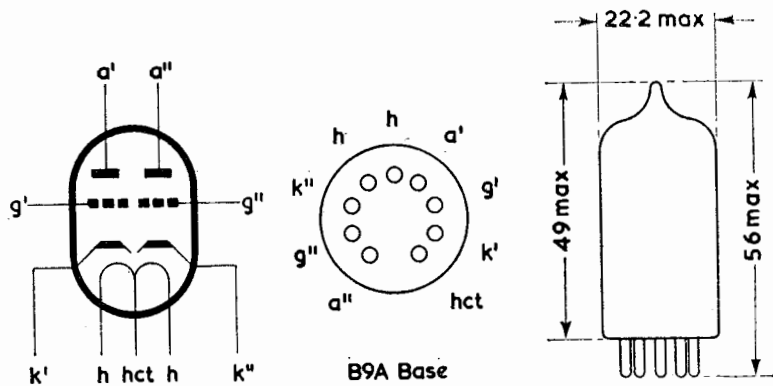Anode current plotted against anode voltage with control-grid voltage as parameter. $V_{g2} = 140$ V.

178

Anode current plotted against anode voltage with control-grid voltage as parameter when triode connected.

*Double triode primarily intended for use as a frequency changer or r.f. amplifier at frequencies up to 300 Mc/s.*

## Valve Data

Heater (suitable for series or parallel operation, a.c. or d.c. The heater is centre-tapped and the two sections may be operated in series or in parallel with one another).

|  | *Series* | $V_h$ applied between pins 4 and 5 | | |
|---|---|---|---|---|
|  | *Parallel* | $V_h$ applied between pin 9 and pins 4 and 5 connected together | | |
|  |  | *Series* | *Parallel* | |
| $V_h$ | | 12·6 | 6·3 | V |
| $I_h$ | | 150 | 300 | mA |



B9A Base

All dimensions in mm

| Capacitances | | |
|---|---|---|
| *$c_{a-g}$ | 1·6 | pF |
| *$c_{1n}$ | 2·3 | pF |
| $c_{a'-k'+h}$ | 0·45 | pF |
| $c_{a''-k''+h}$ | 0·35 | pF |
| *$c_{a-k}$ | 0·2 | pF |
| *$c_{h-k}$ | 2·5 | pF |
| *$c_{k-g+h}$ | 4·7 | pF |
| $c_{a'-g'+h}$ | 1·9 | pF |
| $c_{a''-g''+h}$ | 1·8 | pF |
| $c_{a'-a''}$ | $<$0·4 | pF |
| $c_{g-h}$ | $<$0·17 | pF |
| $c_{g'-g''}$ | $<$0·005 | pF |
| $c_{a'-g''}$ | $<$0·07 | pF |
| $c_{a''-g'}$ | $<$0·04 | pF |

*Each section.
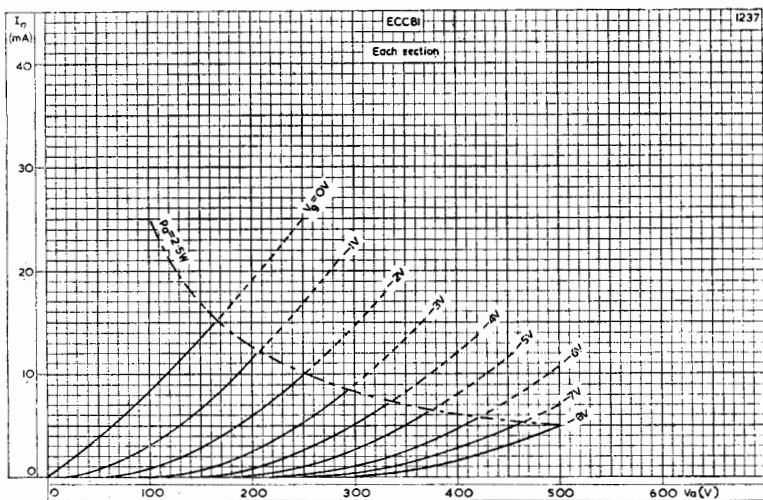
Anode current plotted against anode voltage, with grid voltage as parameter (each section).

## Characteristics (each section)

| $V_a$ | 100 | 170 | 200 | 250 | V |
|---|---|---|---|---|---|
| $I_a$ | 3·0 | 8·5 | 11·5 | 10 | mA |
| $V_g$ | −1·0 | −1·0 | −1·0 | −2·0 | V |
| $g_m$ | 3·75 | 5·9 | 6·7 | 5·5 | mA/V |
| $\mu$ | 62 | 66 | 70 | 60 | |
| $r_a$ | 16·5 | 11 | 10·5 | 11 | kΩ |
| *$r_{g-k}$ | 21 | 16 | 14 | 25 | kΩ |

*Measured at f = 50 Mc/s.

## Limiting Values (each section)

| | | |
|---|---|---|
| $V_{a(b)}$ max | 550 | V |
| $V_a$ max. | 300 | V |
| $p_a$ max. | 2·5 | W |
| $I_k$ max. | 15 | mA |
| −$V_g$ max. | 50 | V |
| $V_g$ ($I_g = +0·3\,\mu$A) | −1·3 | V |
| $R_{g-k}$ max. (self-bias) | 1·0 | MΩ |
| $V_{h-k}$ max. | 150 | V |
| $R_{h-k}$ max. | 20 | kΩ |

181

*Low μ double triode having separate cathodes, primarily intended for use as an amplifier or oscillator.*

## Valve Data

Heater (suitable for series or parallel operation, a.c. or d.c. The heater is centre-tapped and the two sections may be operated in series or in parallel with one another).

|  |  |  |  |
|---|---|---|---|
| *Series* | $V_h$ applied between pins 4 and 5 | | |
| *Parallel* | $V_h$ applied between pin 9 and pins 4 and 5 connected together | | |
|  | *Series* | *Parallel* | |
| $V_h$ | 12·6 | 6·3 | V |
| $I_h$ | 150 | 300 | mA |



B9A Base

Capacitances (measured without an external shield)

|  |  |  |
|---|---|---|
| *$c_{a-g}$ | 1·5 | pF |
| *$c_{in}$ | 1·8 | pF |
| $c_{out'}$ | 370 | mpF |
| $c_{out''}$ | 250 | mpF |
| *$c_{g-h}$ | $<$135 | mpF |
| $c_{a'-a''}$ | $<$1·1 | pF |
| $c_{a''-g'}$ | $<$60 | mpF |
| $c_{a'-g''}$ | $<$110 | mpF |
| $c_{g'-g''}$ | $<$10 | mpF |

*Each section.

## Characteristics (each section)

| | | | |
|---|---|---|---|
| $V_a$ | 100 | 250 | V |
| $I_a$ | 11·8 | 10·5 | mA |
| $V_g$ | 0 | −8·5 | V |
| $g_m$ | 3·1 | 2·2 | mA/V |
| $\mu$ | 19·5 | 17 | |
| $r_a$ | 6·25 | 7·7 | kΩ |
| $V_g$ max. ($I_g = +0·3\ \mu A$) | | −1·3 | V |

## Operating Conditions (each section)

| | | | As an a.f. amplifier | | | | |
|---|---|---|---|---|---|---|---|
| $V_b$ (V) | $R_a$ (kΩ) | $I_k$ (mA) | $R_k$ (kΩ) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out}^*$ ($V_{r.m.s.}$) | $D_{tot}^*$ (%) | $R_g$† (kΩ) |
| 400 | 47 | 5·0 | 1·2 | 13·5 | 59 | 6·7 | 150 |
| 350 | 47 | 4·3 | 1·2 | 13·5 | 51 | 6·6 | 150 |
| 300 | 47 | 3·7 | 1·2 | 13·5 | 43 | 6·5 | 150 |
| 250 | 47 | 3·0 | 1·2 | 13·5 | 34 | 6·4 | 150 |
| 200 | 47 | 2·4 | 1·2 | 13·5 | 26 | 6·3 | 150 |
| 150 | 47 | 1·8 | 1·2 | 13·5 | 18 | 6·1 | 150 |
| 100 | 47 | 1·2 | 1·2 | 13·5 | 11 | 5·6 | 150 |
| 400 | 100 | 2·6 | 2·2 | 14 | 57 | 6·2 | 330 |
| 350 | 100 | 2·3 | 2·2 | 14 | 49 | 6·1 | 330 |
| 300 | 100 | 2·0 | 2·2 | 14 | 41 | 6·0 | 330 |
| 250 | 100 | 1·6 | 2·2 | 14 | 32 | 5·9 | 330 |
| 200 | 100 | 1·3 | 2·2 | 14 | 25 | 5·8 | 330 |
| 150 | 100 | 1·0 | 2·2 | 14 | 17 | 5·6 | 330 |
| 100 | 100 | 0·7 | 2·2 | 14 | 10 | 4·8 | 330 |
| 400 | 220 | 1·3 | 3·9 | 14·5 | 50 | 5·1 | 680 |
| 350 | 220 | 1·2 | 3·9 | 14·5 | 43 | 5·0 | 680 |
| 300 | 220 | 1·0 | 3·9 | 14·5 | 36 | 4·9 | 680 |
| 250 | 220 | 0·8 | 3·9 | 14·5 | 28 | 4·8 | 680 |
| 200 | 220 | 0·7 | 3·9 | 14·5 | 22 | 4·7 | 680 |
| 150 | 220 | 0·5 | 3·9 | 14·5 | 15 | 4·4 | 680 |
| 100 | 220 | 0·3 | 3·9 | 14·5 | 8 | 4·0 | 680 |

*Output voltage and distortion at start of positive grid current. At lower output voltage, the distortion is approximately proportional to the output voltage.
†$R_g$ = grid resistor of following valve.

## Limiting Values (each section)

| | | |
|---|---|---|
| $V_{a(b)}$ max. | 550 | V |
| $V_a$ max. | 300 | V |
| $p_a$ max. | 2·75 | W |
| $I_k$ max. | 20 | mA |
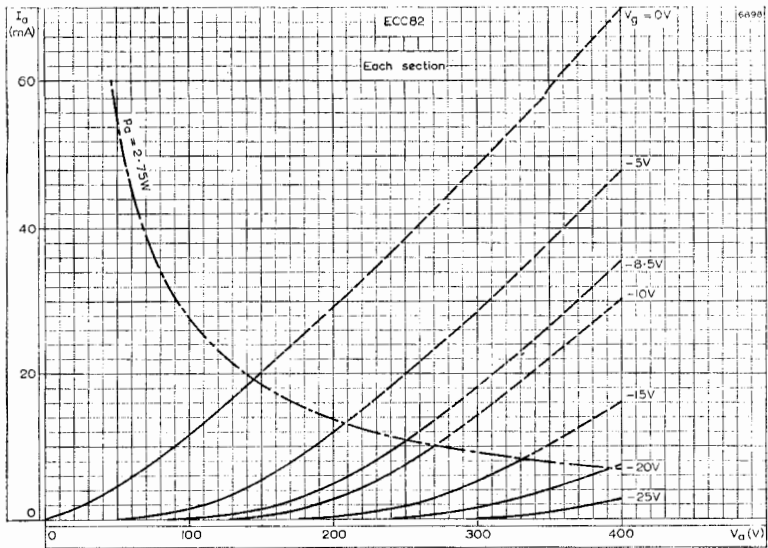
Anode current plotted against anode voltage with grid voltage as parameter (each section).

## Limiting Values (cont'd)

| | | |
|---|---|---|
| *$i_{k(pk)}$ max. | 150 | mA |
| $-V_g$ max. | 100 | V |
| $-v_{g(pk)}$ max. | 250 | V |
| $R_{g-k}$ max. (fixed bias) | 1·5 | MΩ |
| $V_{h-k}$ max. | 180 | V |
| †$R_{h-k}$ max. | 20 | kΩ |

*Maximum pulse duration=200 μs.
†When used as a phase inverter immediately preceding the output stage, $R_{h-k}$ max. may be 150 kΩ.

## OPERATING NOTES

This valve can be used without special precautions against microphony in equipment where the input voltage is not less than 10 mV for an output of 50 mW (or 100 mV for 5 W output).

With $V_h$ applied between pin 9 and pins 4 and 5 connected together, and with the centre tap of the heater transformer earthed the section connected to pins 6, 7 and 8 is the most favourable with regard to hum.

184

# DOUBLE TRIODE      ECC83

*High $\mu$ double triode, having separate cathodes, primarily intended for use as a resistance-coupled amplifier or phase inverter.*

## Valve Data

Heater (suitable for series or parallel operation, a.c. or d.c. The heater is centre-tapped and the two sections may be operated in series or in parallel with one another).

| | | | |
|---|---|---|---|
| *Series* | $V_h$ applied between pins 4 and 5 | | |
| *Parallel* | $V_h$ applied between pin 9 and pins 4 and 5 connected together | | |
| | *Series* | *Parallel* | |
| $V_h$ | 12·6 | 6·3 | V |
| $I_h$ | 150 | 300 | mA |



B9A Base

**All dimensions in mm**

| Capacitances | | |
|---|---|---|
| $c_{out'}$ | 330 | mpF |
| $c_{out''}$ | 230 | mpF |
| *$c_{in}$ | 1·6 | pF |
| *$c_{a-g}$ | 1·6 | pF |
| $c_{a'-a''}$ | <1·2 | pF |
| $c_{a''-g'}$ | <100 | mpF |
| $c_{a'-g''}$ | <110 | mpF |
| $c_{g'-g''}$ | <10 | mpF |
| *$c_{g-h}$ | <150 | mpF |

*Each section.

185

## Characteristics (each section)

| | | | |
|---|---|---|---|
| $V_a$ | 100 | 250 | V |
| $I_a$ | 0·5 | 1·2 | mA |
| $V_g$ | −1·0 | −2·0 | V |
| $g_m$ | 1·25 | 1·6 | mA/V |
| $\mu$ | 100 | 100 | |
| $r_a$ | 80 | 62·5 | kΩ |
| $V_g$ max. ($I_g = +0·3\ \mu A$) | | −0·9 | V |

## *Operating Conditions as Resistance Coupled A.F. Amplifier with grid current bias ($R_g = 10\ M\Omega$)*

| $V_b$ (V) | $R_a$ (kΩ) | $R_g$′** (kΩ) | $I_a$ (mA) | $Z_s = 0\ k\Omega$ $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$* (V) | $Z_s = 220\ k\Omega$ $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$† (V) |
|---|---|---|---|---|---|---|---|
| 400 | 47 | 150 | 3·5 | 48 | 43 | 33 | 49 |
| 350 | 47 | 150 | 2·8 | 46 | 35 | 33 | 41 |
| 300 | 47 | 150 | 2·2 | 45 | 28 | 32 | 33 |
| 250 | 47 | 150 | 1·7 | 43 | 21 | 31 | 25 |
| 200 | 47 | 150 | 1·2 | 40 | 14 | 30 | 17 |
| 400 | 100 | 330 | 2·1 | 63 | 58 | 42 | 68 |
| 350 | 100 | 330 | 1·7 | 62 | 48 | 42 | 57 |
| 300 | 100 | 330 | 1·4 | 61 | 38 | 41 | 46 |
| 250 | 100 | 330 | 1·1 | 59 | 29 | 41 | 36 |
| 200 | 100 | 330 | 0·8 | 56 | 20 | 39 | 25 |
| 400 | 220 | 680 | 1·2 | 75 | 69 | 49 | 82 |
| 350 | 220 | 680 | 1·0 | 74 | 57 | 48 | 70 |
| 300 | 220 | 680 | 0·8 | 72 | 45 | 47 | 57 |
| 250 | 220 | 680 | 0·6 | 70 | 34 | 46 | 44 |
| 200 | 220 | 680 | 0·4 | 66 | 24 | 44 | 32 |

*Output voltage measured at $D_{tot} = 5\%$.

$\dfrac{V_{out}}{V_{in}}$ measured with $V_{in\ r.m.s.} = 100$ mV.

**Grid resistor of following valve.

†When operating this valve with grid current bias and a high source impedance, the second harmonic distortion rises to a peak at quite low levels of output (about 10 $V_{r.m.s.}$) and then falls with increasing drive. The third harmonic then begins to rise, and $D_{tot}$ finally reaches 5% at a much higher output level than with zero source impedance. The maximum value of this distortion peak varies inversely with the anode load, being about 5·5% with $R_a = 47$ kΩ, 4·5% with $R_a = 100$ kΩ and 4% with $R_a = 220$ kΩ.

$I_a$ (mA)

ECC83   6234

Each section

$V_g = 0V$, $-0.5V$, $-1.0V$, $-1.5V$, $-2.0V$, $-2.5V$, $-3.0V$, $-3.5V$, $-4.0V$, $-4.5V$, $-5.0V$

$V_a$(V)
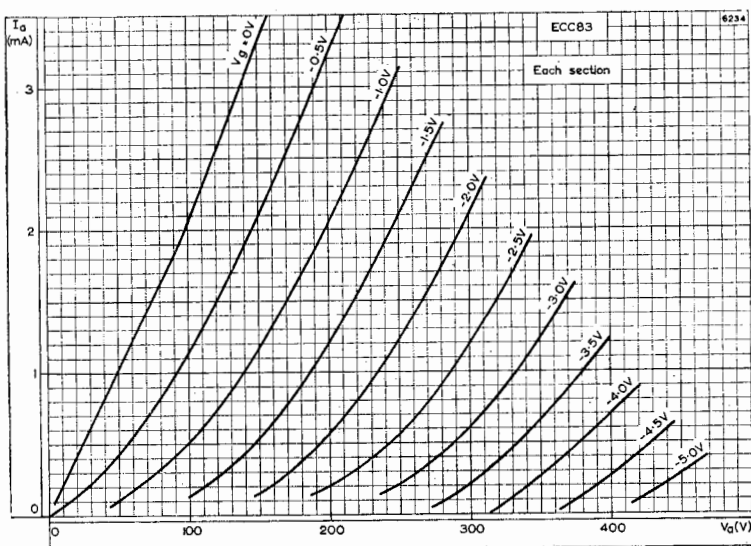
Anode current plotted against anode voltage with grid voltage as parameter (each section).

## Operating Conditions as Resistance Coupled A.F. Amplifier with Cathode Bias

| $V_b$ (V) | $R_a$ (kΩ) | $I_a$ (mA) | $R_k$ (kΩ) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$* (V) | $D_{tot}$* (%) | $R_g$† (kΩ) |
|---|---|---|---|---|---|---|---|
| 400 | 47 | 2·2 | 1·0 | 43 | 40·5 | 5·0 | 150 |
| 350 | 47 | 1·7 | 1·2 | 42 | 31 | 5·0 | 150 |
| 300 | 47 | 1·3 | 1·5 | 40 | 22 | 5·0 | 150 |
| 250 | 47 | 0·9 | 2·2 | 36 | 12·5 | 5·0 | 150 |
| 400 | 100 | 1·4 | 1·5 | 59 | 59 | 5·0 | 330 |
| 350 | 100 | 1·1 | 1·8 | 57 | 45 | 5·0 | 330 |
| 300 | 100 | 0·88 | 2·2 | 55 | 32·5 | 5·0 | 330 |
| 250 | 100 | 0·6 | 3·3 | 50 | 18·5 | 5·0 | 330 |
| 400 | 220 | 0·88 | 2·2 | 71 | 63 | 3·7 | 680 |
| 350 | 220 | 0·7 | 2·7 | 69 | 60 | 5·0 | 680 |
| 300 | 220 | 0·5 | 3·9 | 65 | 38·5 | 5·0 | 680 |
| 250 | 220 | 0·38 | 4·7 | 62 | 27 | 5·0 | 680 |

*Output voltage measured at $D_{tot} = 5\%$ or at start of positive grid current. At lower output voltages the distortion is approximately proportional to the output voltage.
† Grid resistor of following valve.
At lower values of $V_b$, grid current bias should be used.

187

*Limiting Values (each section)*

| | | |
|---|---:|---|
| $V_a$ max. | 300 | V |
| $p_a$ max. | $1 \cdot 0$ | W |
| $I_k$ max. | $8 \cdot 0$ | mA |
| $-V_g$ max. | 50 | V |
| $R_{g-k}$ max. (fixed bias) | $1 \cdot 0$ | $M\Omega$ |
| $V_{h-k}$ max. | 180 | V |
| *$R_{h-k}$ max. | 20 | $k\Omega$ |

*When used as a phase inverter immediately preceding the output stage, $R_{h-k}$ max. may be 150 kΩ.

## OPERATING NOTES

### 1. Microphony

This valve may be used without special precautions against microphony in equipment where the input voltage is not less than 5 mV for an output of 50 mW (or 50 mV for 5W output).

### 2. Hum

With $V_h$ applied between pin 9 and pins 4 and 5 connected together and the centre tap of the heater transformer earthed, the section connected to pins 6, 7 and 8 is the most favourable with regard to hum, and should be used for the input section when the two sections are used in cascade.

When used as a normal voltage amplifier with $V_b = 250$ V, $R_a = 100$ kΩ, $R_g = 330$ kΩ, $R_k = 1 \cdot 5$ kΩ (suitably decoupled), the maximum hum level of the input triode is 10 $\mu$V, the average value being 6 $\mu$V. If one side of the heater is earthed, rather than the centre tap, it is preferable to earth pins 4 and 5. The average value of hum under these conditions may be 50 $\mu$V.

*Combined high-μ triode and output pentode for use in audio amplifier circuits.*

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 700 | mA |
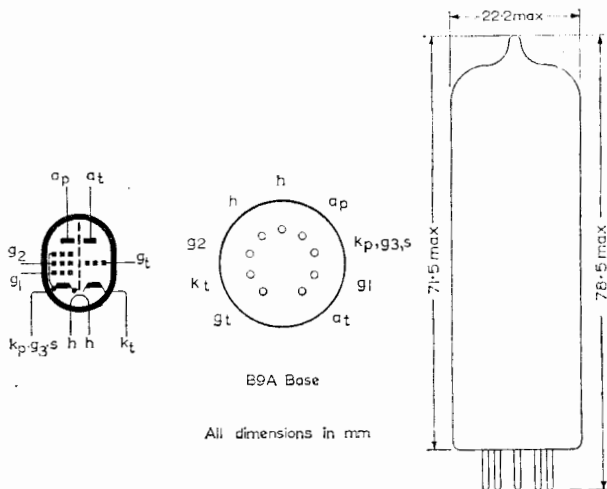


B9A Base

All dimensions in mm

| Capacitances | | |
|---|---|---|
| $C_{ap-gt}$ | $<6·0$ | mpF |
| $c_{at-g1}$ | $<200$ | mpF |
| $c_{gt-g1}$ | $<20$ | mpF |
| $c_{at-ap}$ | $<150$ | mpF |
| Pentode section | | |
| $c_{in}$ | 10 | pF |
| $c_{a-g1}$ | $<400$ | mpF |
| $c_{g1-h}$ | $<200$ | mpF |
| Triode section | | |
| $c_{in}$ | 2·3 | pF |
| $c_{out}$ | 2·5 | pF |
| $c_{a-g}$ | 1·4 | pF |
| $c_{g-h}$ | $<6·0$ | mpF |

## Characteristics

### Pentode section

| | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_{g2}$ | 250 | V |
| $V_{g1}$ | −7·0 | V |
| $I_a$ | 36 | mA |
| $I_{g2}$ | 6·0 | mA |
| $g_m$ | 10 | mA/V |
| $r_a$ | 48 | kΩ |
| $\mu_{g1-g2}$ | 21 | |

### Triode section (see ECC83)

| | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_g$ | −1·9 | V |
| $I_a$ | 1·2 | mA |
| $g_m$ | 1·6 | mA/V |
| $\mu$ | 100 | |
| $r_a$ | 62 | kΩ |

## Operating Conditions as Single Valve Amplifier

### Pentode section

| | | | | |
|---|---|---|---|---|
| $V_a$ | 250 | 250 | 250 | V |
| $V_{g2}$ | 210 | 250 | 250 | V |
| $R_k$ | 130 | 270 | 170 | Ω |
| $I_a$ | 36 | 26 | 36 | mA |
| $I_{g2}$ | 5·6 | 4·4 | 6·0 | mA |
| $R_a$ | 7·0 | 10 | 7·0 | kΩ |
| $P_{out}$ | 4·0 | 2·8 | 4·0 | W |
| $V_{in(r.m.s.)}$ | 3·1 | 2·7 | 3·2 | V |
| $D_{tot}$ | 10 | 10 | 10 | % |
| $V_{in(r.m.s.)}$ ($P_{out} = 50mW$) | 280 | 280 | 300 | mV |

## Operating Conditions for Two Valves in Push-Pull

### Cathode bias

| | | | |
|---|---|---|---|
| $V_{a(b)}$ | 250 | 300 | A |
| $V_{g2(b)}$ | 250 | 300 | V |
| $R_k$ (per valve) | 180 | 290 | V |
| $R_{a-a}$ | 8·0 | 9·0 | kΩ |
| $I_{a(o)}$ | $2 \times 35$ | $2 \times 31$ | mΩ |
| $I_{g2(o)}$ | $2 \times 5·6$ | $2 \times 5·0$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 10·2 | 17·4 | V |
| $P_{out}$ | 10 | 14·3 | W |
| $D_{tot}$ | 4·5 | 5·0 | % |
| $V_{in(r.m.s.)}$ ($P_{out} = 50$ mW) | 480 | 520 | mV |
| $I_{a}$ (max. sig.) | $2 \times 37·3$ | $2 \times 37$ | mA |
| $I_{g2}$ (max. sig.) | $2 \times 9·0$ | $2 \times 10·6$ | mA |

## Operating Conditions for Triode Section as Resistance Coupled A.F. Amplifier

| | | | Cathode bias | | | | |
|---|---|---|---|---|---|---|---|
| $V_b$ (V) | $R_a$ (kΩ) | $I_a$ (mA) | $R_k$ (kΩ) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$* (V) | $D_{tot}$* (%) | $R_g$† (kΩ) |
| 300 | 47 | 1·3 | 1·5 | 40 | 22 | 5·0 | 150 |
| 250 | 47 | 0·9 | 2·2 | 36 | 12·5 | 5·0 | 150 |
| 300 | 100 | 0·88 | 2·2 | 55 | 32·5 | 5·0 | 330 |
| 250 | 100 | 0·6 | 3·3 | 50 | 18·5 | 5·0 | 330 |
| 300 | 220 | 0·5 | 3·9 | 65 | 38·5 | 5·0 | 680 |
| 250 | 220 | 0·38 | 4·7 | 62 | 27 | 5·0 | 680 |

*Output voltage measured at $D_{tot} = 5\%$ or at start of positive grid current. At lower output voltages the distortion is approximately proportional to the output voltage.
†Grid resistor of following valve.
At lower values of $V_b$, grid current bias should be used.

### Grid current bias ($R_g = 10$ MΩ)

| | | | | $Z_s = 0$ kΩ | | $Z_s = 220$ kΩ | |
|---|---|---|---|---|---|---|---|
| $V_b$ (V) | $R_a$ (kΩ) | $R_g$† (kΩ) | $I_a$ (mA) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$* (V) | $\dfrac{V_{out}}{V_{in}}$ | $V_{out(r.m.s.)}$‡ (V) |
| 300 | 47 | 150 | 2·2 | 45 | 28 | 32 | 33 |
| 250 | 47 | 150 | 1·7 | 43 | 21 | 31 | 25 |
| 200 | 47 | 150 | 1·2 | 40 | 14 | 30 | 17 |
| 300 | 100 | 330 | 1·4 | 61 | 38 | 41 | 46 |
| 250 | 100 | 330 | 1·1 | 59 | 29 | 41 | 36 |
| 200 | 100 | 330 | 0·8 | 56 | 20 | 39 | 25 |
| 300 | 220 | 680 | 0·8 | 72 | 45 | 47 | 57 |
| 250 | 220 | 680 | 0·6 | 70 | 34 | 46 | 44 |
| 200 | 220 | 680 | 0·4 | 66 | 24 | 44 | 32 |

*Output voltage measured at $D_{tot} = 5\%$.
$\dfrac{V_{out}}{V_{in}}$ measured with $V_{in(r.m.s.)} = 100$ mV.
†Grid resistor of following valve.
‡When operating this valve with grid current bias and a h gh source impedance, the second harmonic distortion rises to a peak at quite low levels of output (about 10 $V_{r.m.s.}$) and then falls with increasing drive. The third harmonic then begins to rise, and $D_{tot}$ finally reaches 5% at a much higher output level than with zero source impedance. The maximum value of this distortion peak varies inversely with the anode load, being about 5·5% with $R_a = 47$ kΩ, 4·5% with $R_a = 100$ kΩ and 4% with $R_a = 220$ kΩ.

Anode and screen-grid currents plotted against anode voltage with
control-grid voltage as parameter. Pentode section. $V_{g2} = 250$ V.



Anode and screen-grid currents plotted against anode voltage with
control-grid voltage as parameter. Pentode section. $V_{g2} = 300$ V.
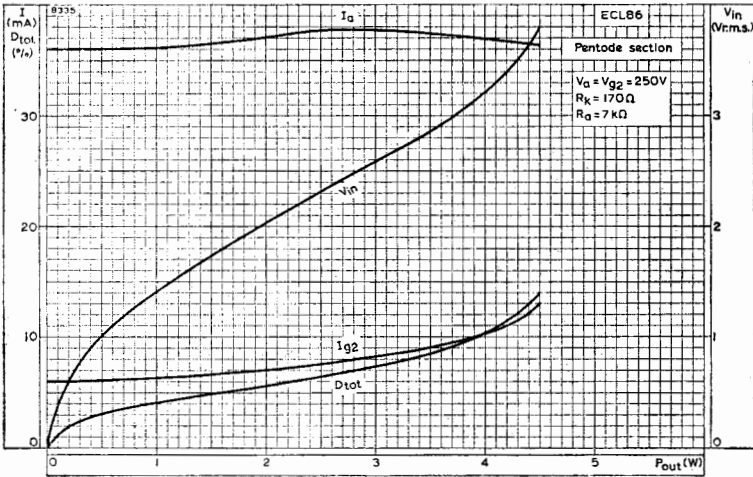
192

Performance of ECL86 as single valve amplifier. Pentode section.
$V_a = V_{g2} = 250$ V.
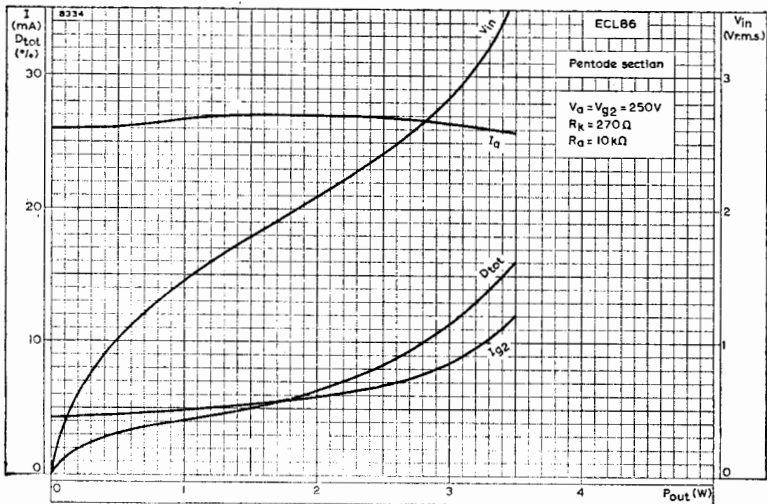


Performance of ECL86 as single valve amplifier. Pentode section.
$V_a = V_{g2} = 250$ V.
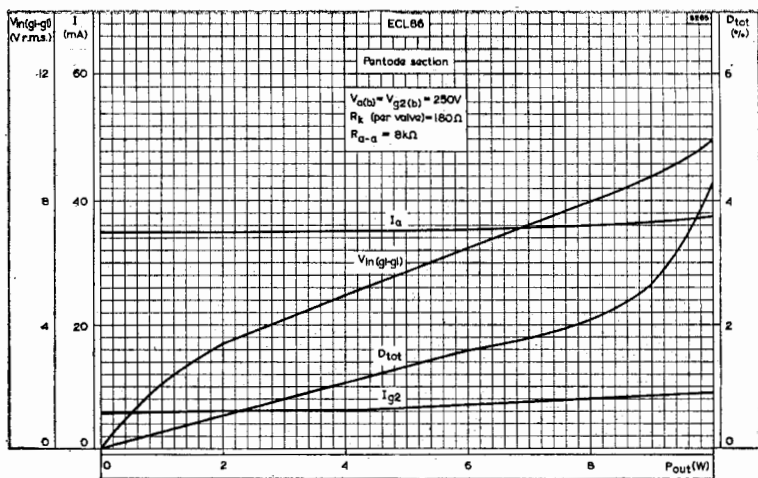
Performance of ECL86 in push-pull. Pentode section.
$V_{a(b)} = V_{g2(b)} = 250$ V.
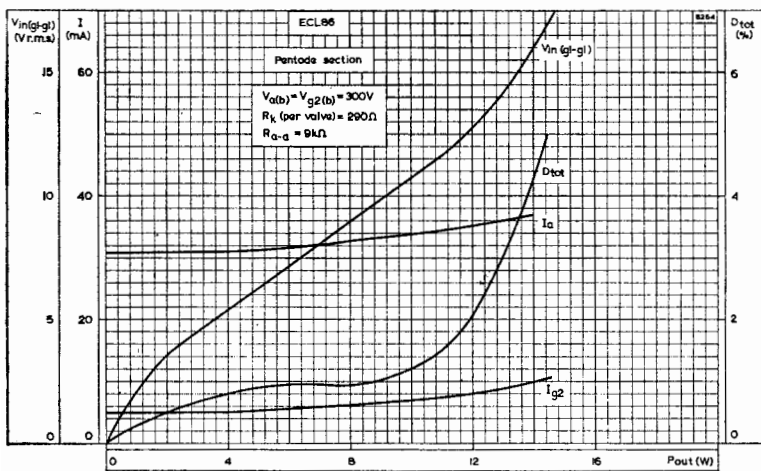


Performance of ECL86 in push-pull. Pentode section.
$V_{a(b)} = V_{g2(b)} = 300$ V.

194

| Pentode section | | |
|---|---:|---|
| $V_{a(b)}$ max. | 550 | V |
| $V_a$ max. | 300 | V |
| $p_a$ max | 9·0 | W |
| $V_{g2(b)}$ max. | 550 | V |
| $V_{g2}$ max. | 300 | V |
| $p_{g2}$ max. | 1·8 | W |
| $I_k$ max. | 55 | mA |
| $R_{g1-k}$ max. | 1·0 | MΩ |
| $V_{h-k}$ max. | 100 | V |
| $R_{h-k}$ max. | 20 | kΩ |

| Triode section | | |
|---|---:|---|
| $V_{a(b)}$ max. | 550 | V |
| $V_a$ max. | 300 | V |
| $p_a$ max. | 500 | mW |
| $I_k$ max. | 4·0 | mA |
| $R_{g-k}$ max. | 1·0 | MΩ |
| $V_{h-k}$ max. | 100 | V |
| *$R_{h-k}$ max. | 20 | kΩ |

*When used as a phase inverter immediately preceding the output stage, $R_{h-k}$ max. may be 120 kΩ.

## OPERATING NOTES

### 1. Microphony

This valve may be used without special precautions against microphony in equipment where the input voltage is not less than 4 mV for an output of 50 mW.

### 2. Hum

To obtain the minimum value of hum, pin 4 should be earthed.

## EL84 OUTPUT PENTODE

*Output pentode rated for 12 W anode dissipation, primarily intended for use in a.c. mains operated equipment.*

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 760 | mA |

| Capacitances | | |
|---|---|---|
| $c_{in}$ | 10·8 | pF |
| $c_{out}$ | 6·5 | pF |
| $c_{a-g1}$ | $<$500 | mpF |
| $c_{g1-h}$ | $<$250 | mpF |



B9A Base

22·2 mm max
71 mm max
78 mm max
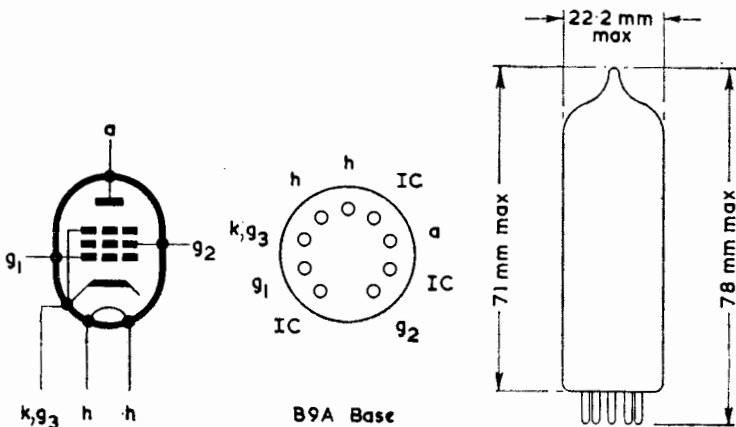
| Characteristics | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_{g2}$ | 250 | V |
| $I_a$ | 48 | mA |
| $I_{g2}$ | 5·5 | mA |
| $V_{g1}$ | −7·3 | V |
| $g_m$ | 11·3 | mA/V |
| $r_a$ | 38 | kΩ |
| $\mu_{g1-g2}$ | 19 | |

## Operating Conditions as Single Valve Amplifier

| Pentode connection | | | |
|---|---|---|---|
| $V_a$ | 250 | 250 | V |
| $V_{g2}$ | 250 | 250 | V |
| $R_a$ | | 5·2 | 4·5 | kΩ |
| $V_{g1}$ | -7·3 | -7·3 | V |
| $I_a$ | 48 | 48 | mA |
| $I_{g2}$ | 5·5 | 5·5 | mA |
| $V_{in(r.m.s.)}$ ($P_{out} = 50$ mW) | 300 | 300 | mV |
| $V_{in(r.m.s.)}$ ($D_{tot} = 10\%$) | 4·3 | 4·4 | V |
| $P_{out}$ ($D_{tot} = 10\%$) | 5·7 | 5·7 | W |
| $D_3$ | 9·5 | 8·0 | % |
| $D_2$ | 2·0 | 5·0 | % |

| Triode connection ($g_2$ connected to a) | | |
|---|---|---|
| $V_a$ | 250 | V |
| $R_a$ | 3·5 | kΩ |
| $V_{g1}$ | -9·0 | V |
| $I_{a(o)}$ | 34 | mA |
| $V_{in(r.m.s.)}$ ($P_{out} = 50$ mW) | 1·0 | V |
| $V_{in(r.m.s.)}$ | 6·0 | V |
| $P_{out}$ | 1·5 | W |
| $D_{tot}$ | 6·0 | % |
| $I_{a(max. sig.)}$ | 39 | mA |

## Operating Conditions for Two Valves in Push-Pull

| Pentode connection | | | |
|---|---|---|---|
| $V_a$ | 250 | 300 | V |
| $V_{g2}$ | 250 | 300 | V |
| $R_k$ (per valve) | 270 | 270 | Ω |
| $R_{a-a}$ | 8·0 | 8·0 | kΩ |
| $I_{a(o)}$ | 2 × 31 | 2 × 36 | mA |
| $I_{g2(o)}$ | 2 × 3·5 | 2 × 4·0 | mA |
| $V_{in(g1-g1)}$ r.m.s. | 16 | 20 | V |
| $P_{out}$ | 11 | 17 | W |
| $D_{tot}$ | 3·0 | 4·0 | % |
| $I_{a(max. sig.)}$ | 2 × 37·5 | 2 × 46 | mA |
| $I_{g2(max. sig.)}$ | 2 × 7·5 | 2 × 11 | mA |
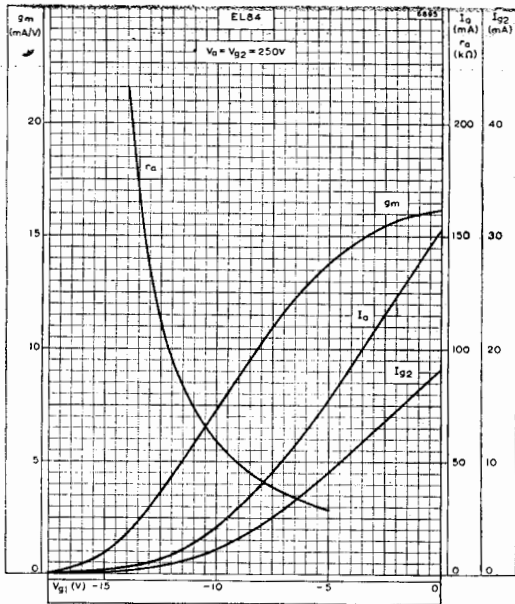
Anode and screen-grid currents, mutual conductance and anode impedance plotted against control-grid voltage.
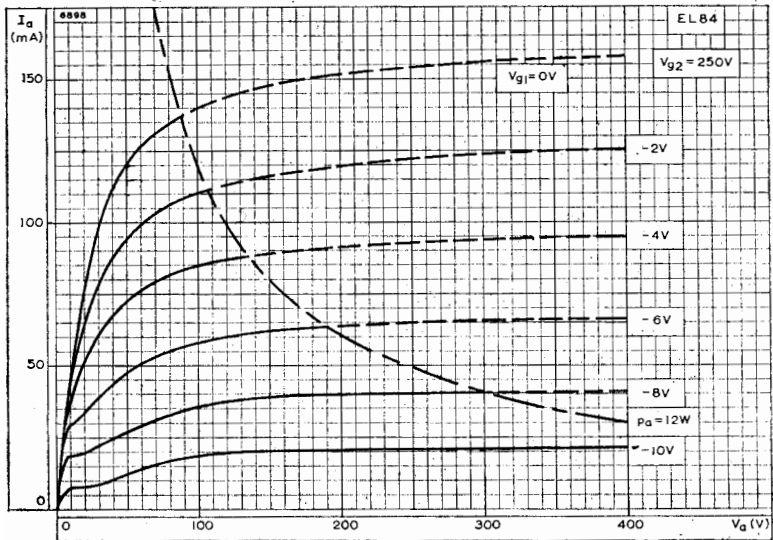


Anode current plotted against anode voltage with control-grid voltage as parameter. $V_{g2} = 250$ V.
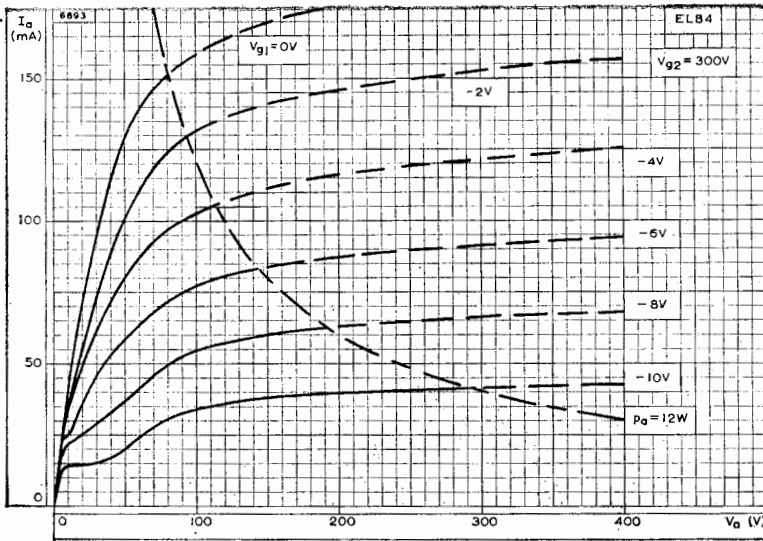
Anode current plotted against anode voltage with control-grid
voltage as parameter. $V_{g2} = 300$ V.
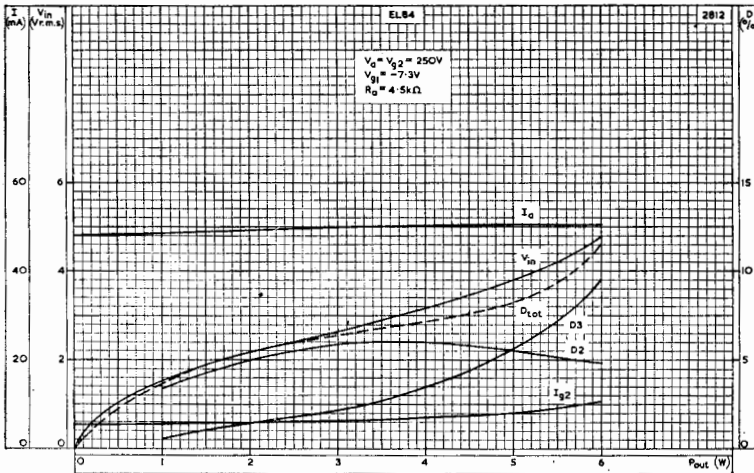


Performance of EL84 when used as a single valve amplifier. $V_a =$
$V_{g2} = 250$V.
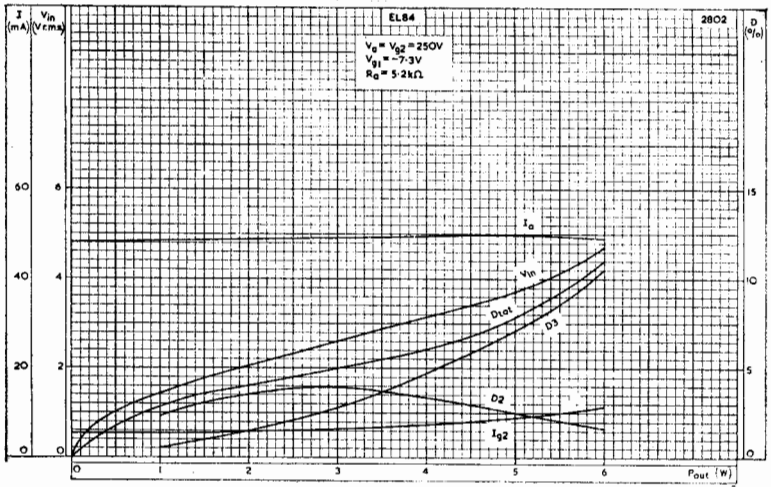
Performance of EL84 when used as a single valve amplifier. $V_a = V_{g2} = 250$ V.
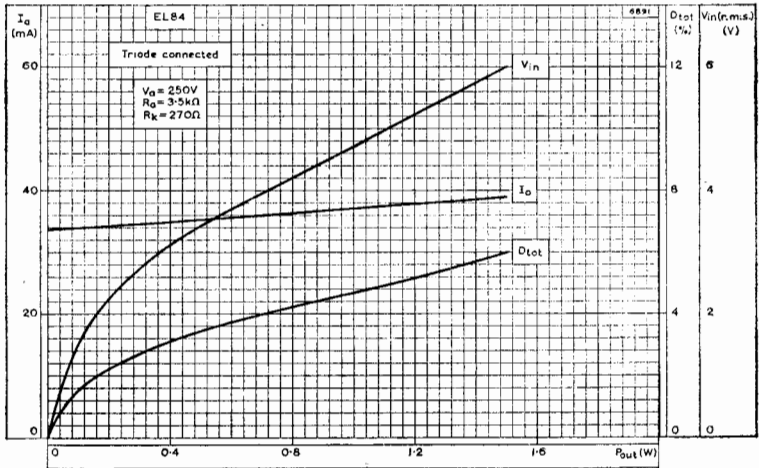


Performance of EL84 triode connected as a single valve amplifier. $V_a = 250$ V.
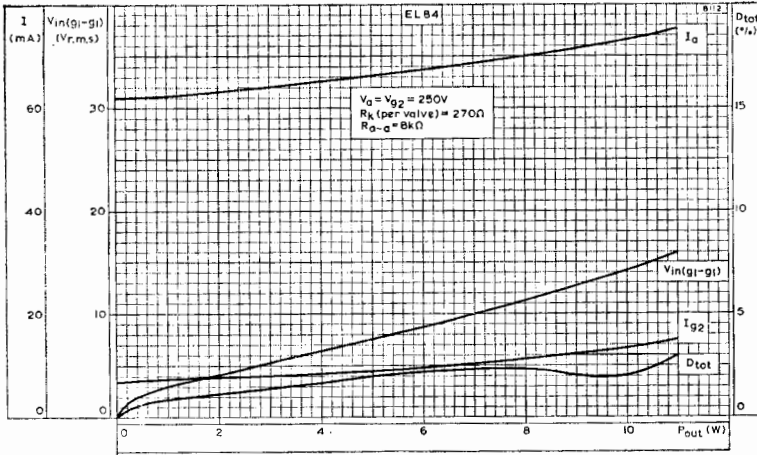
Performance of two EL84 in push-pull. $V_a = V_{g2} = 250$ V.
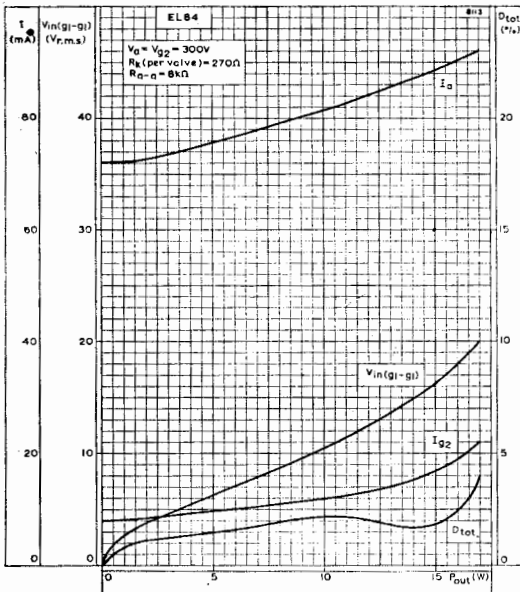


Performance of two EL84 in push-pull. $V_a = V_{g2} = 300$ V.
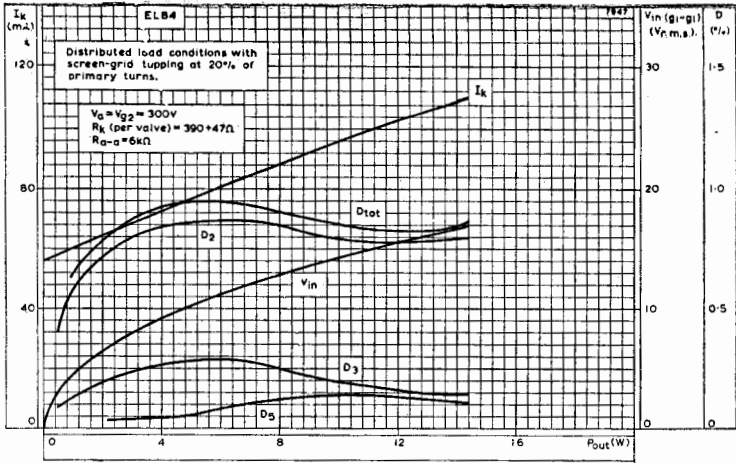
201
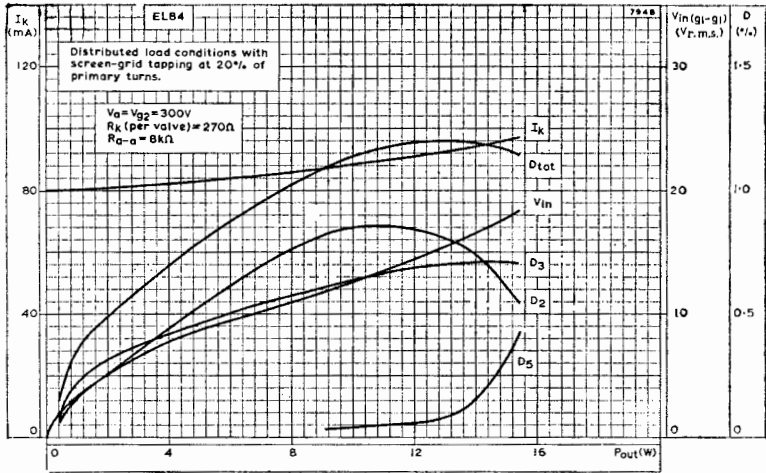
Performance of two EL84 in push-pull with distributed load conditions. Screen-grid tapping at 20% of primary turns. $V_a = V_{g2} = 300$ V.



Performance of two EL84 in push-pull with distributed load conditions. Screen-grid tapping at 20% of primary turns. $V_a = V_{g2} = 300$ V.

**Graph 1 (top):**

$I_k$ (mA)

EL84

Distributed load conditions with screen-grid tapping at 43% of primary turns.

$V_a = V_{g2} = 300V$
$R_k$ (per valve) = 390+47Ω
$R_{a-a} = 6kΩ$

$V_{in(g_1-g_1)}$ (V r.m.s.)   D (%)

$I_k$
$V_{in}$
$D_{tot}$
$D_2$
$D_3$
$D_5$

$P_{out}$ (W)

Performance of two EL84 in push-pull with distributed load conditions. Screen-grid tapping at 43% of primary turns. $V_a = V_{g2} = 300$ V.

**Graph 2 (bottom):**

$I_k$ (mA)

EL84

Distributed load conditions with screen-grid tapping at 43% of primary turns.

$V_a = V_{g2} = 300V$
$R_k$ (per valve) = 270Ω
$R_{a-a} = 8kΩ$

$V_{in(g_1-g_1)}$ (V r.m.s.)   D (%)

$V_{in}$
$I_k$
$D_{tot}$
$D_2$
$D_3$

$P_{out}$ (W)

Performance of two EL84 in push-pull with distributed load conditions. Screen-grid tapping at 43% of primary turns. $V_a = V_{g2} = 300$ V.

203

Performance of two EL84 triode connected in push-pull.
$V_{a+g2} = 250$ V.



Performance of two EL84 triode connected in push-pull.
$V_{a+g2} = 300$ V.

204

Distributed load conditions for maximum output (screen-grid tapping at 20% of primary turns)

| | | | |
|---|---|---|---|
| $V_a$ | 300 | 300 | V |
| $V_{g2}$ | 300 | 300 | V |
| $R_k$ (per valve) | 390 + 47 | 270 | Ω |
| $R_{a-a}$ | 6·0 | 8·0 | kΩ |
| $I_{k(o)}$ | 2 × 28 | 2 × 40 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 17 | 18·3 | V |
| $P_{out}$ | 14·4 | 15·4 | W |
| $D_{tot}$ | 0·85 | 1·17 | % |
| $I_{k(max. sig.)}$ | 2 × 55 | 2 × 48·5 | mA |

Distributed load conditions for minimum distortion (screen-grid tapping at 43% of primary turns)

| | | | |
|---|---|---|---|
| $V_a$ | 300 | 300 | V |
| $V_{g2}$ | 300 | 300 | V |
| $R_k$ (per valve) | 390 + 47 | 270 | Ω |
| $R_{a-a}$ | 6·0 | 8·0 | kΩ |
| $I_{k(o)}$ | 2 × 28 | 2 × 40 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 16·8 | 16 | V |
| $P_{out}$ | 10·1 | 11 | W |
| $D_{tot}$ | 0·72 | 0·7 | % |
| $I_{k(max. sig.)}$ | 2 × 47 | 2 × 45 | mA |

Triode connection ($g_2$ connected to a)

| | | | |
|---|---|---|---|
| $V_a$ | 250 | 300 | V |
| $R_k$ (per valve) | 560 | 560 | Ω |
| $R_{a-a}$ | 10 | 10 | kΩ |
| $I_{a(o)}$ | 2 × 20 | 2 × 24 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 16·5 | 20 | V |
| $P_{out}$ | 3·4 | 5·2 | W |
| $D_{tot}$ | 2·5 | 2·5 | % |
| $I_{a(max. sig.)}$ | 2 × 21·5 | 2 × 26 | mA |

## Operating Conditions with Continuous Sine Wave Drive

Single valve

| | | | |
|---|---|---|---|
| $V_a$ | 250 | 250 | V |
| $V_{g2(b)}$ | 250 | 250 | V |
| *$R_{g2}$ | 4·7(±10%) | 3·9(±10%) | kΩ |
| $R_k$ | 130 | 130 | Ω |
| $R_a$ | 5·25 | 4·5 | kΩ |

| | | | |
|---|---|---|---|
| $I_{a(o)}$ | 44 | 44 | mA |
| $I_{g2(o)}$ | 5·1 | 5·2 | mA |
| $V_{in(r.m.s.)}$ | 4·4 | 4·65 | V |
| $P_{out}$ | 5·4 | 5·6 | W |
| $D_{tot}$ | 12·5 | 13·9 | % |
| $I_{a(max. sig.)}$ | 40 | 42 | mA |
| $I_{g2(max. sig.)}$ | 8·6 | 8·4 | mA |
| $P_{g2}$ | 1·8 | 1·8 | W |

*Decoupled by 8 μF capacitor.

### Two valves in push-pull

| | | |
|---|---|---|
| $V_a$ | 300 | V |
| $V_{g2(b)}$ | 300 | V |
| *$R_{g2}$ | $1·8(\pm10\%)$ | kΩ |
| $R_k$ (per valve) | 270 | Ω |
| $R_{a-a}$ | 8·0 | kΩ |
| $I_{a(o)}$ | $2 \times 35$ | mA |
| $I_{g2(o)}$ | $2 \times 4·0$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 17·4 | V |
| $P_{out}$ | 15 | W |
| $D_{tot}$ | 3·4 | % |
| $I_{a(max. sig.)}$ | $2 \times 42$ | mA |
| $I_{g2(max. sig.)}$ | $2 \times 7·0$ | mA |
| $P_{g2}$ | 1·93 | W |

*Screen-grid resistor common to both valves.

## Limiting Values

| | | |
|---|---|---|
| $V_{a(b)}$ max. | 550 | V |
| $V_a$ max. | 330 | V |
| $p_a$ max. | 12 | W |
| $V_{g2(b)}$ max. | 550 | V |
| $V_{g2}$ max. | 300 | V |
| $p_{g2}$ max. | 2·0 | W |
| $I_k$ | 65 | mA |
| $-V_g$ max. | 100 | V |
| $R_{g1-k}$ max. | 300 | kΩ |
| $V_{h-k}$ max. | 100 | V |
| $R_{h-k}$ max. | 20 | kΩ |

*A beam tetrode with an absolute maximum anode dissipation rating of 30 W. It is designed for use in the output stage of an a.f. amplifier, or as a series valve in a stabilized power supply.*

*The KT66 is a commercial version of the CV1075.*

## Base Connections and Valve Dimensions
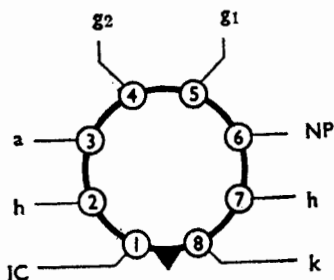
Base: B8-0 (Octal)
Bulb: Dome top tubular
Max. overall length:    135 mm
Max. seated length:    121 mm
Max. diameter:    53 mm

View from underside of base.

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 1·3 (approx) | A |

| Maximum ratings | | | |
|---|---|---|---|
| | *Design Max.* | *Absolute Max.* | |
| $V_a$ | 500 | 550 | V |
| $V_{g2}$ | 500 | 550 | V |
| $-V_{g1}$ | 200 | 200 | V |
| $I_k$ | 200 | 200 | mA |
| $p_a$ | 25 | 30 | W |
| $p_{g2}$ | 3·5 | 4·5 | W |
| *$p_{a+g2}$ | 27 | 32 | W |
| $V_{h-k}$ | 150 | 150 | V |
| $T_{bulb}$ | 250 | 250 | °C |
| $R_{g1-k}$ (cathode bias): | | | |
| $\quad p_{a+g2} \leqslant 27\,W$ | | 1·0 | $M\Omega$ |
| $\quad p_{a+g2} > 27\,W$ | 500 | | $k\Omega$ |
| $R_{g1-k}$ (fixed bias): | | | |
| $\quad p_{a+g2} \leqslant 27\,W$ | 250 | | $k\Omega$ |
| $\quad p_{a+g2} > 27\,W$ | 100 | | $k\Omega$ |

*Triode or ultralinear operation.

207

## Capacitances (measured on a cold unscreened valve)

$c_{g1-all,\,less\,a}$: $14 \cdot 5$ pF    $c_{a-all,\,less\,g1}$: $10 \cdot 0$ pF    $c_{a-g1}$: $1 \cdot 1$ pF

## Characteristics

### Tetrode connection

| | | |
|---|---:|---|
| $V_a$ | 250 | V |
| $V_{g2}$ | 250 | V |
| $-V_{g1}$ | 15 | V |
| $g_m$ | 7 | mA/V |
| $r_a$ | $22 \cdot 5$ | kΩ |

### Triode connection

| | | |
|---|---:|---|
| $V_a$ | 250 | V |
| $-V_{g1}$ | 15 | V |
| $g_m$ | $7 \cdot 3$ | mA/V |
| $r_a$ | $1 \cdot 3$ | kΩ |

## Typical Operation

### Triode connection. Class A. Single valve. Cathode bias.

| | | | |
|---|---:|---:|---|
| $V_{a,g2(b)}$ | 270 | 440 | V |
| $V_{a,g2}$ | 250 | 400 | V |
| $-V_{g1}$ (approx) | 20 | 38 | V |
| $v_{in(pk)}$ | 20 | 38 | V |
| $R_k$ | 330 | 600 | Ω |
| $I_{a+g2(o)}$ | 60 | 63 | mA |
| $p_{a+g2(o)}$ | 15 | 25 | W |
| $R_L$ | $2 \cdot 75$ | $4 \cdot 5$ | kΩ |
| $P_{out}$ | $2 \cdot 2$ | $5 \cdot 8$ | W |
| $D_{tot}$ | 6 | 7 | % |

### Triode connection. Class AB1. Push-pull. Cathode bias

| | | | |
|---|---:|---:|---|
| $V_{a,g2(b)}$ | 270 | 440 | V |
| $V_{a,g2}$ | 250 | 400 | V |
| $-V_{g1}$ (approx) | 19 | 38 | V |
| $v_{in(g1-g1)(pk)}$ | 38 | 76 | V |
| *$R_k$ | $2 \times 345$ | $2 \times 615$ | Ω |
| $I_{a+g2(o)}$ | $2 \times 55$ | $2 \times 62$ | mA |
| $p_{a+g2(o)}$ | $2 \times 14$ | $2 \times 25$ | W |
| $R_{L(a-a)}$ | $2 \cdot 5$ | $4 \cdot 0$ | kΩ |

| | | | |
|---|---|---|---|
| $P_{out}$ | 4·5 | 14·5 | W |
| $D_{tot}$ | 2·0 | 3·5 | % |
| †IM | 3·0 | 3·0 | % |
| $z_{out}$ | 3·5 | 3·5 | kΩ |

*It is essential to use two separate cathode bias resistors.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

## Tetrode connection. Class AB1. Push-pull. Cathode bias

| | | |
|---|---|---|
| $V_{a(b)(o)}$ | 450 | V |
| $V_{a(b)(max.\ sig.)}$ | 425 | V |
| $V_{a(o)}$ | 415 | V |
| $V_{a(max.\ sig.)}$ | 390 | V |
| $V_{g2(o)}$ | 300 | V |
| $V_{g2(max.\ sig.)}$ | 275 | V |
| $-V_{g1}$ (approx) | 27 | V |
| $I_{a(o)}$ | 2 × 52 | mA |
| $I_{a(max.\ sig.)}$ | 2 × 62 | mA |
| $I_{g2(o)}$ | 2 × 2·5 | mA |
| $I_{g2(max.\ sig.)}$ | 2 × 9 | mA |
| $P_{a(o)}$ | 2 × 21 | W |
| $P_{a(max.\ sig.)}$ | 2 × 9·0 | W |
| $P_{g2(o)}$ | 2 × 0·75 | W |
| $P_{g2(max.\ sig.)}$ | 2 × 2·5 | W |
| *$R_k$ | 2 × 500 | Ω |
| $R_{L(a-a)}$ | 8 | kΩ |
| $v_{in(g1-g1)(pk)}$ | 70 | V |
| $P_{out}$ | 30 | W |
| $D_{tot}$ | 6 | % |

*It is essential to use two separate cathode bias resistors.

## Ultralinear connection. Push-pull. 40% taps. Class AB1. Cathode bias.

| | | |
|---|---|---|
| $V_{a(b)}$ | 450 | V |
| $V_{a,g2(o)}$ | 425 | V |
| $V_{a,g2(max.\ sig.)}$ | 400 | V |
| $I_{a+g2(o)}$ | 2 × 62·5 | mA |
| $I_{a+g2(max.\ sig.)}$ | 2 × 72·5 | mA |
| $P_{a+g2(o)}$ | 2 × 26·5 | W |
| $P_{a+g2(max.\ sig.)}$ | 2 × 13·0 | W |
| *$R_k$ | 2 × 560 | Ω |
| $-V_g$ (approx) | 35 | V |
| $P_{out}$ | 32 | W |
| $R_{L(a-a)}$ | 7 | kΩ |
| $z_{out}$ | 9 | kΩ |
| $D_{tot}$ | 2 | % |
| †IM | 4 | % |

*It is essential to use two separate cathode bias resistors.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

O $^A$

Ultralinear connection. Class AB1. Push-pull. 40% taps. Fixed bias

| | | |
|---|---|---|
| $V_{a,g2(o)}$ | 525 | V |
| $V_{a,g2(max.\ sig.)}$ | 500 | V |
| $I_{a+g2(o)}$ | $2 \times 35$ | mA |
| $I_{a+g2(max.\ sig.)}$ | $2 \times 80$ | mA |
| $P_{a+g2(o)}$ | $2 \times 18$ | W |
| $P_{a+g2(max.\ sig.)}$ | $2 \times 15$ | W |
| * $-V_{g1}$ (approx) | 67 | V |
| $R_{L(a-a)}$ | 8 | k$\Omega$ |
| $v_{in(g1-g1)(pk)}$ | 127 | V |
| $P_{out}$ | 50 | W |
| $D_{tot}$ | 3 | % |
| †IM | 15 | % |
| $z_{out}$ | 10 | k$\Omega$ |

*A negative bias range of $\pm 25\%$ of this value should be available for each valve.
†Intermodulation distortion: measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

## Installation

The valve may be mounted in any position but when horizontal it should be orientated as shown in the diagram below. No retaining device or external screening is normally necessary.
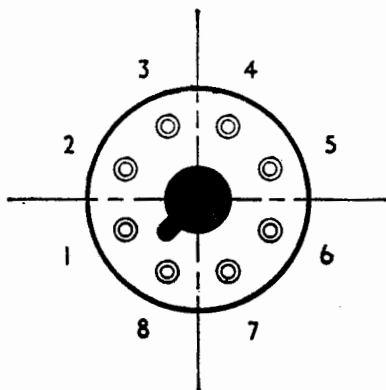


*Correct orientation of the valve socket for horizontally mounting the KT66.*

Adequate ventilation should be provided. A pair of valves working at maximum ratings should be mounted at not less than $3 \cdot 5$ in. (9 cm) between centres.

For the prevention of parasitic oscillation, a series resistor of 100–300 $\Omega$ should be connected close to the screen tag of the valve socket. When the valve is triode connected, this resistor should be connected between screen and anode. A control grid series resistor of 10–50 k$\Omega$ is also recommended. In push-pull applications having a large change in anode current between the quiescent and full output conditions, an
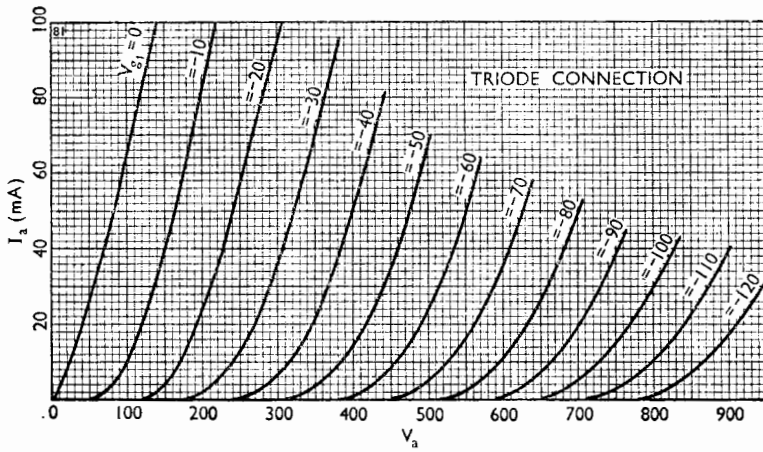
KT66. Anode current plotted against anode voltage with control-grid voltage as parameter when triode connected.
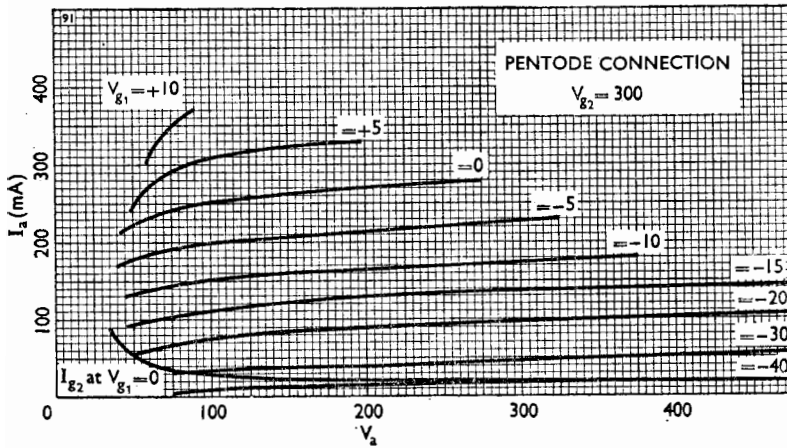


KT66. Anode and screen-grid currents plotted against anode voltage with control-grid voltage as parameter. $V_{g2} = 300$ V.

211
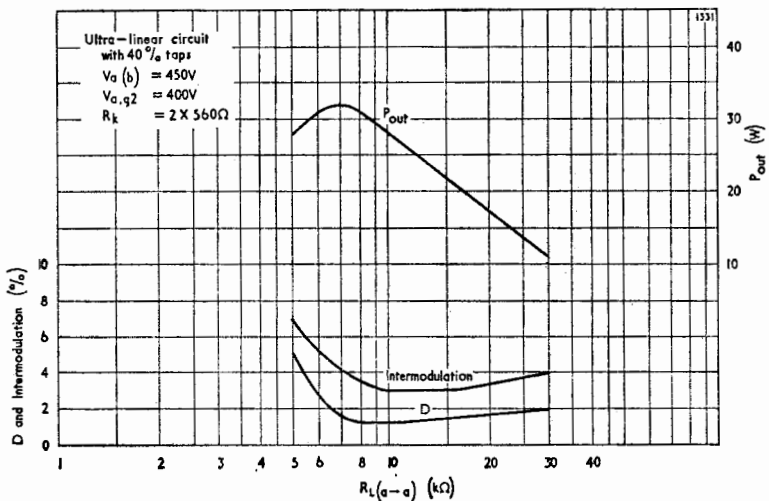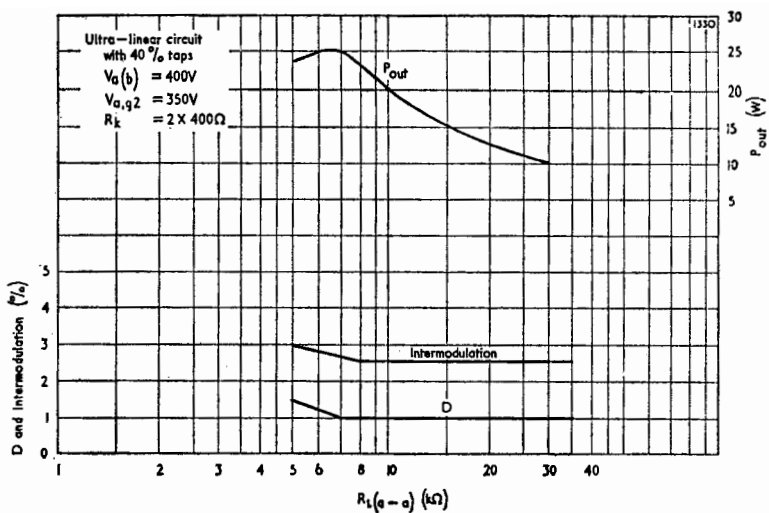
KT66. The above curves show power output, harmonic distortion and intermodulation plotted against load impedance. Ultra-linear operation at two values of $V_{a(b)}$.
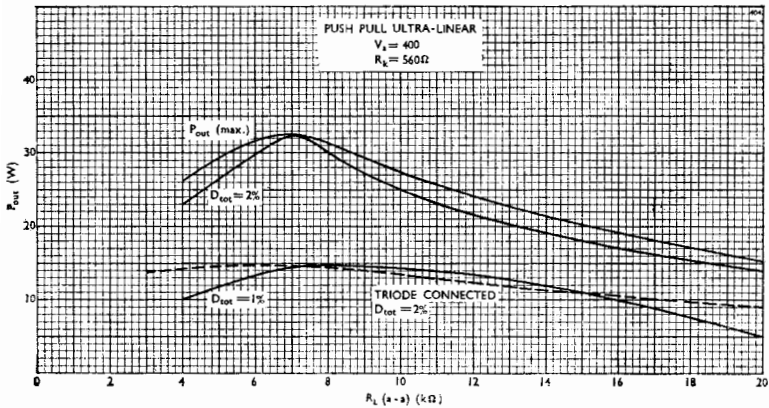
KT66. A comparison of ultra-linear with triode operation.
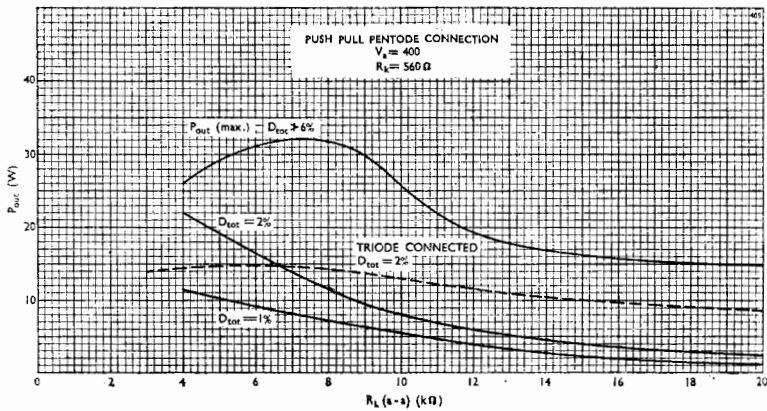


KT66. A comparison of pentode with triode operation.

inductance input filter circuit of good regulation should be used. A badly regulated supply will cause a fall in power output and/or excessive quiescent anode dissipation.

## INTERMODULATION IN ULTRALINEAR AMPLIFIERS

The curves below illustrate the various degrees of intermodulation with varying positions of the screen taps on the output transformer from 0% (tetrode operation) to 100% (triode operation). The measurements were made under SMPE* conditions using frequencies of 50 and 6000 c/s at a ratio of 4:1. It will be noticed that the lowest level of intermodulation is obtained with taps of about 40%. A slight increase in intermodulation is produced at 33% or 57% and a significant increase at 25% or 75%.

The triode performance (100%) is inferior to the ultralinear at all outputs except below 10W with 25% taps. The high intermodulation of the tetrode arrangement (0%) is to be expected.

The application of negative feedback will reduce intermodulation roughly in proportion to the reduction in gain. For example, a 3:1 reduction in intermodulation is obtained with 10db and a 5:1 reduction with 14 db feedback. It can be seen that it is not difficult to reduce intermodulation to below 1% at 50W output by the use of feedback.

*S.M.P.E. = Society of Motion Picture Engineers.
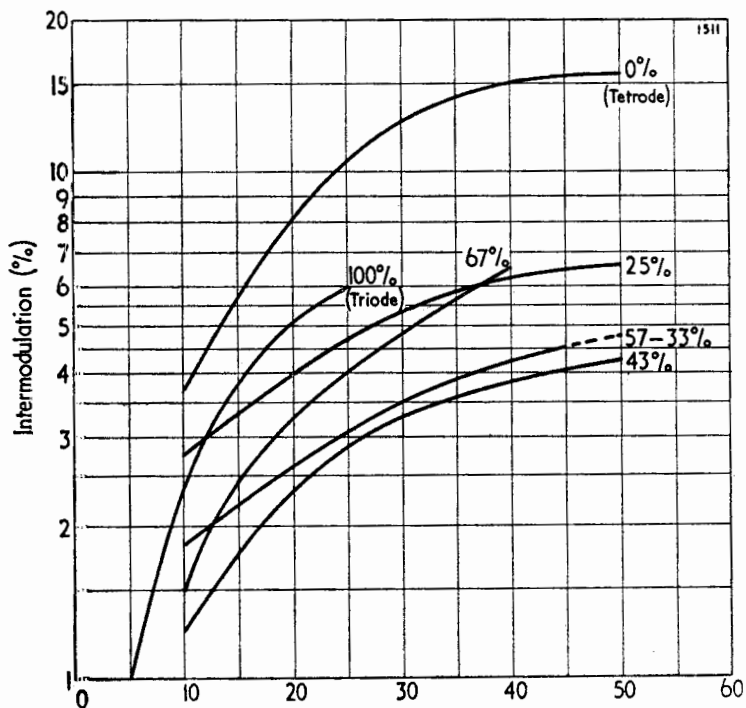


The various degrees of intermodulation with varying positions of the screen taps on the output transformer from 0° (tetrode operation) to 100% (triode operation).

214

## Comparison of Operating Conditions

| | Triode | Pentode* | Ultralinear | |
|---|---|---|---|---|
| $V_{a(b)}$ | 450 | 450 | 450 | V |
| $V_a$ | 400 | 400 | 400 | V |
| $V_{g2}$ | 400 | 400* | 400 | V |
| $I_{a+g2}$ | 125 | 125 | 125 | mA |
| $I_{g2(o)}$ | 5 | 5 | 5 | mA |
| $I_{a+g2(max.\ sig.)}$ | 135 | 135 | 145 | mA |
| $I_{g2(max.\ sig.)}$ | 8 | 24* | 15 | mA |
| $R_k$ (per valve) | 560 | 560 | 560 | Ω |
| $V_k$ (app) | 36 | 36 | 36 | V |
| $P_{in(o)}$ | 50 | 50 | 50 | ·W |
| $P_{in\ (max.\ sig.)}$ | 54 | 62 | 58 | W |
| $P_{out}$ | 15 | 32 | 32 | W |
| †$\eta$ | 28 | 52 | 55 | % |
| $D_{(max.\ sig.)}$ | 2 | ≯6 | 2 | % |
| $R_{L(a-a)}$ | 6 | 7 | 7 | kΩ |
| $z_{out}$ | 3·5 | 35 | 9 | kΩ |
| $V_{in(g1-g1)(r.m.s.)}$ | 52 | 42 | 56 | V |
| Relative $P_{out}$ for given $V_{in}$ $\Big\{$ | ×1 | ×5 | ×2·5 | |
| | 0 | +7 | +4 | db |

*Suitable for intermittent operation only, due to excessive screen dissipation at full output. When continuous full output is required, use the operating condition: $V_a = 400$, $V_{g2} = 300$.
†$\eta$ = % efficiency, i.e. $P_{out} \times 100 / P_{input}$.

215

*Output pentode rated for 25 W anode dissipation, intended for use in a.c. mains operated equipment*

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 1·5 | A |

| Capacitances | | |
|---|---|---|
| $c_{out}$ | 8·4 | pF |
| $c_{in}$ | 15·2 | pF |
| $c_{a-g1}$ | $<1·0$ | pF |
| $c_{g1-h}$ | $<1·0$ | pF |
| $c_{h-k}$ | 11 | pF |



All dimensions in mm
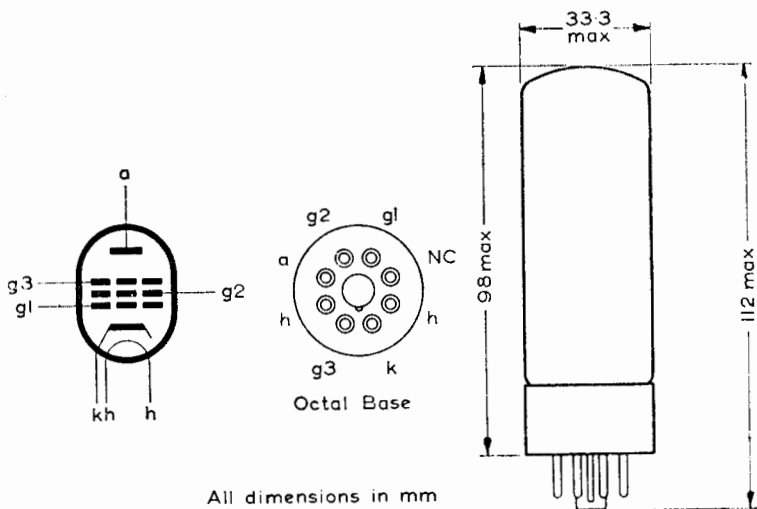
| Characteristics | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_{g2}$ | 250 | V |
| $V_{g3}$ | 0 | V |
| $I_a$ | 100 | mA |
| $I_{g2}$ | 15 | mA |
| $V_{g1}$ | −12·2 | V |
| $g_m$ | 11 | mA/V |
| $r_a$ | 15 | kΩ |
| $\mu_{g1-g2}$ | 11 | |
| $V_{g1}$ max. ($I_{g1} = +0·3\ \mu A$) | −1·3 | V |

## Operating Conditions as Single Valve Class 'A' Amplifier

| Pentode connection | | | |
|---|---|---|---|
| $V_a$ | 250 | 300 | V |
| $V_{g2}$ | 250 | 300 | V |
| $V_{g3}$ | 0 | 0 | V |
| $R_k$ | 106 | 190 | $\Omega$ |
| $R_a$ | 2·0 | 3·5 | k$\Omega$ |
| $I_a$ | 100 | 83 | mA |
| $I_{g2}$ | 15 | 13 | mA |
| $V_{in(r.m.s.)}$ ($P_{out} = 50$ mW) | 500 | 450 | mV |
| $V_{in(r.m.s.)}$ | 8·0 | 8·2 | V |
| *$P_{out}$ | 11 | 11 | W |
| *$D_{tot}$ | 10 | 10 | % |

*$P_{out}$ and $D_{tot}$ are measured at fixed bias and therefore represent the power output available during the reproduction of speech and music. When a sustained sine wave is applied to the control-grid the bias across the cathode resistor will readjust itself as a result of the increased anode and screen-grid currents. This will result in a reduction in power output of approximately 10%.

## Operating Conditions for Two Valves in Push-Pull

Distributed load conditions for maximum output (screen-grid tapping at 20% of primary turns)

| | | |
|---|---|---|
| $V_b$ | 450 | V |
| $R_{g2}$ (per valve) | 1·0 | k$\Omega$ |
| $R_k$ (per valve) | 500 | $\Omega$ |
| $R_{a-a}$ | 7·0 | k$\Omega$ |
| $I_{a(o)}$ | $2 \times 55$ | mA |
| $I_{g2(o)}$ | $2 \times 9·0$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 55·2 | V |
| $P_{out}$ | 40 | W |
| $D_{tot}$ | 4·5 | % |
| $I_{a(max. sig.)}$ | $2 \times 74$ | mA |
| $I_{g2(max. sig.)}$ | $2 \times 9·0$ | mA |

Distributed load conditions for minimum distortion (with screen-grid tapping at 43% of primary turns)

| | | | |
|---|---|---|---|
| $V_b$ | 430 | 430 | V |
| $R_{g2}$ (per valve) | 1·0 | 1·0 | k$\Omega$ |
| $R_k$ (per valve) | 470 | 470 | $\Omega$ |
| $R_{a-a}$ | 6·0 | 6·0 | k$\Omega$ |
| $I_{a(o)}$ | $2 \times 62·5$ | $2 \times 62·5$ | mA |
| $I_{g2(o)}$ | $2 \times 10$ | $2 \times 10$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 35 | 50 | V |
| $P_{out}$ | 20 | 34 | W |
| $D_{tot}$ | 0·35 | 2·5 | % |
| $I_{a (max. sig.)}$ | $2 \times 65$ | $2 \times 70$ | mA |
| $I_{g2(max. sig.)}$ | $2 \times 10·2$ | $2 \times 14$ | mA |
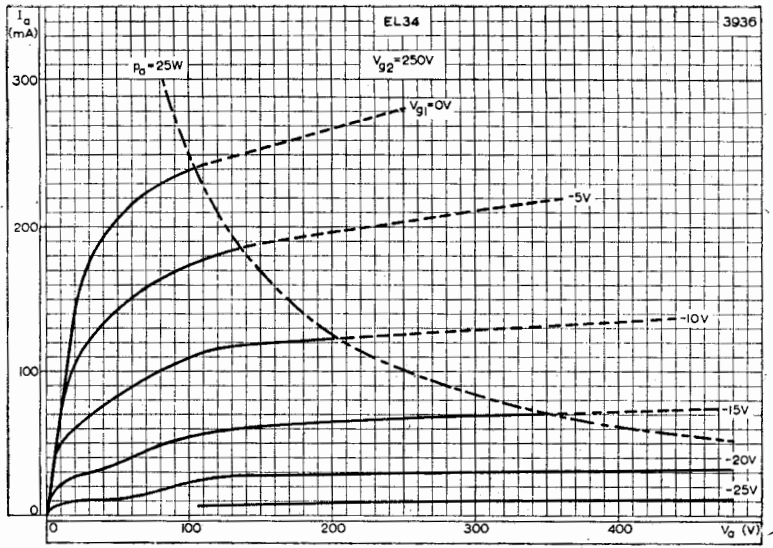
Anode current plotted against anode voltage with control-grid voltage as parameter.
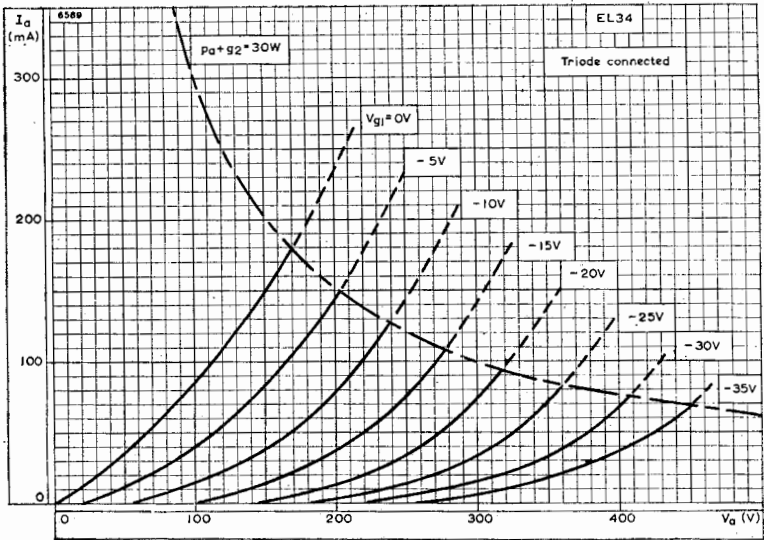


Anode current plotted against anode voltage with control-grid voltage as parameter when triode connected.

218
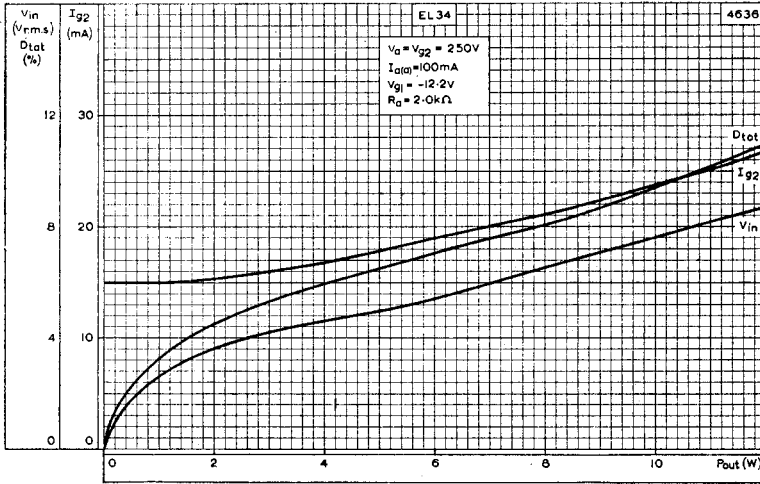
Performance of EL34 when used as a single valve Class 'A' amplifier.
$$V_a = V_{g2} = 250 \text{ V}.$$
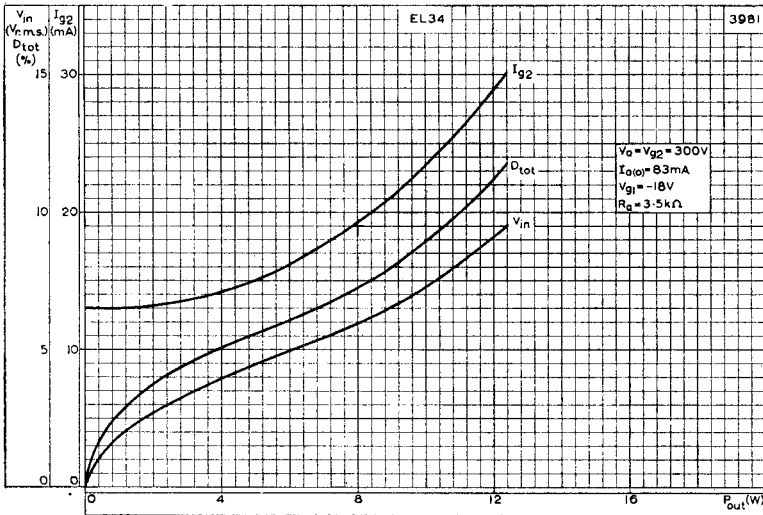


Performance of EL34 when used as a single valve Class 'A' amplifier.
$$V_a = V_{g2} = 300 \text{ V}.$$
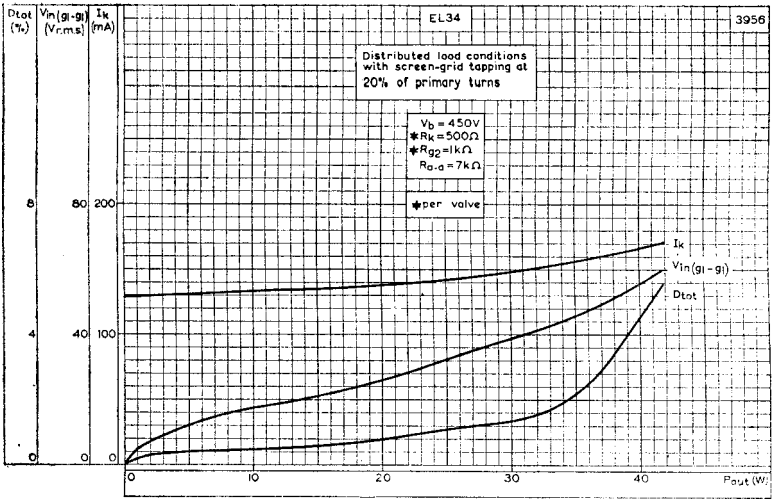
219
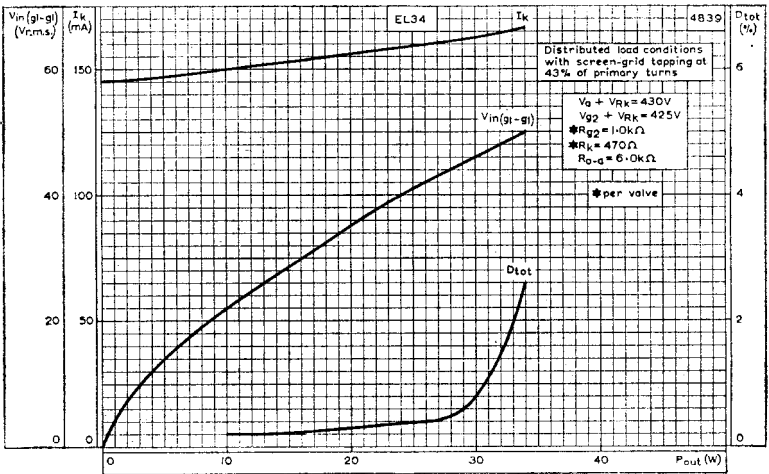
Performance of two EL34 in push-pull with distributed load conditions. Screen-grid tapping at 20 % of primary turns. $V_b = 430$ V.



Performance of two EL34 in push-pull with distributed load conditions. Screen-grid tapping at 43 % of primary turns. $V_b = 430$ V.

220
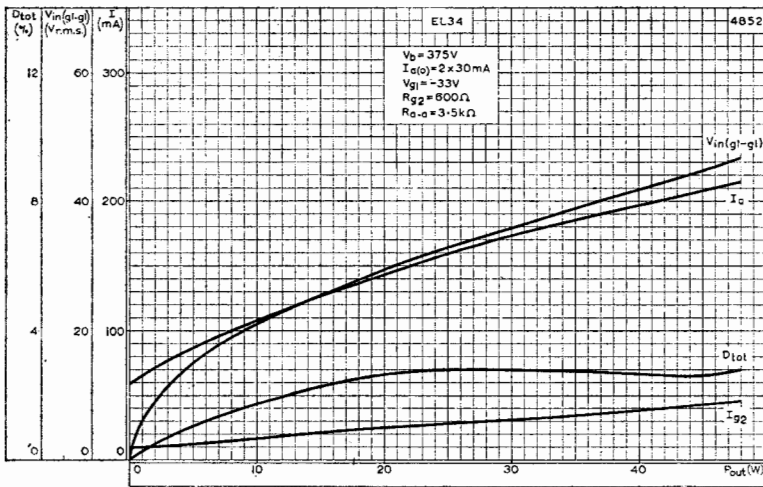
Performance of two EL34 in push-pull with fixed bias. $V_b = 375$ V.
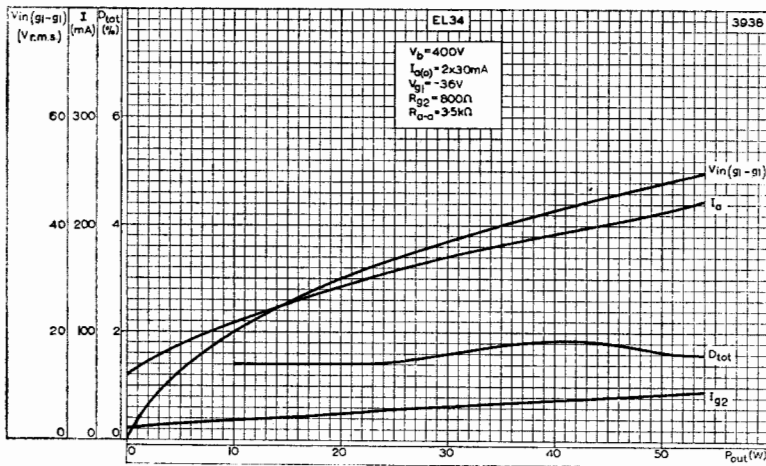


Performance of two EL34 in push-pull with fixed bias. $V_b = 400$ V.

221
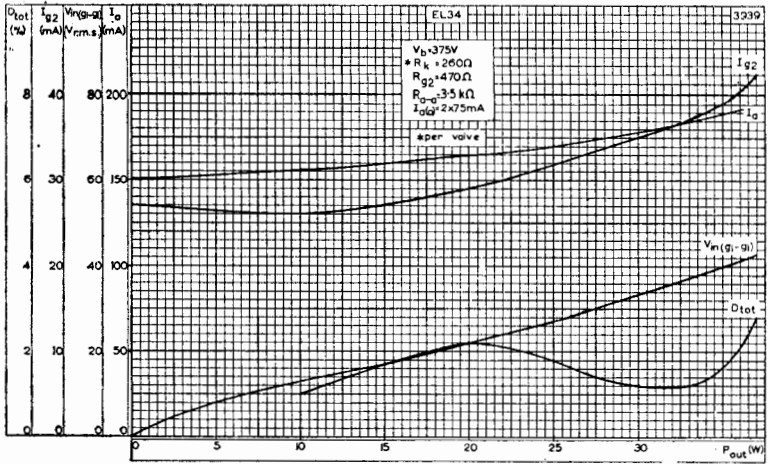
Performance of two EL34 in push-pull with cathode bias.
$V_b = 375$ V.
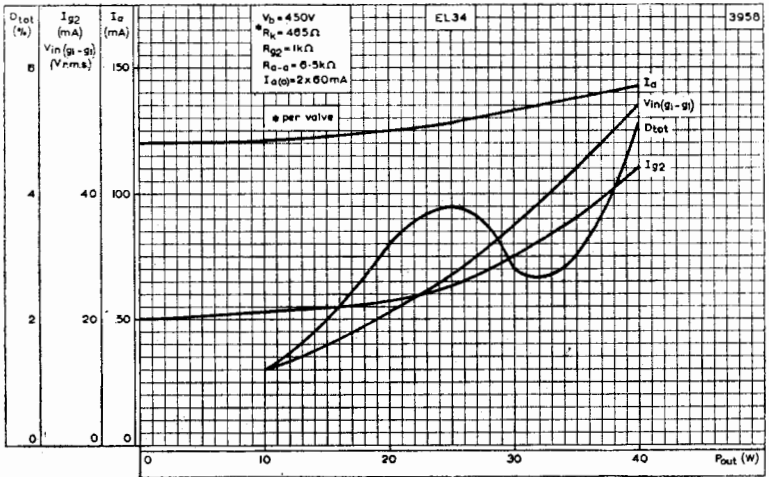


Performance of two EL34 in push-pull with cathode bias.
$V_b = 450$ V.

222
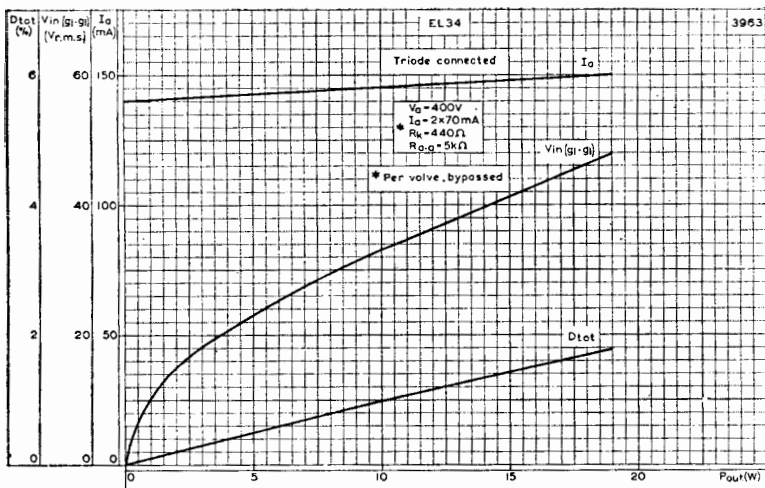
Performance of two EL34 in push-pull when triode connected and the cathode bypassed. $V_a = 400$ V.
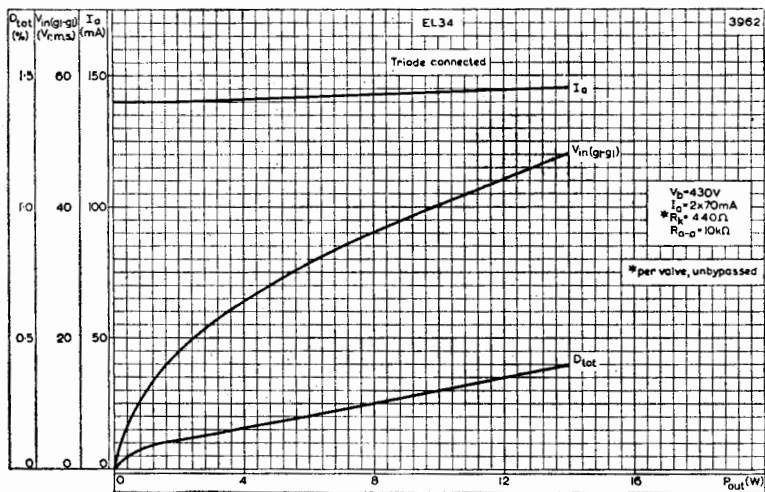


Performance of two EL84 in push-pull. when triode connected and the cathode unbypassed. $V_b = 430$ V.

223

## Operating Conditions for Two Valves in Push-Pull

| Fixed bias | | | |
|---|---|---|---|
| $V_b$ | 375 | 400 | V |
| $V_{g3}$ | 0 | 0 | V |
| $*R_{g2}$ | 600 | 800 | $\Omega$ |
| $V_{g1}$ | −33 | −36 | V |
| $R_{a-a}$ | 3·5 | 3·5 | k$\Omega$ |
| $I_{a(o)}$ | 2 × 30 | 2 × 30 | mA |
| $I_{g2(o)}$ | 2 × 4·7 | 2 × 4·5 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 46·7 | 50 | V |
| $P_{out}$ | 48 | 54 | W |
| $D_{tot}$ | 2·8 | 1·6 | % |
| $I_{a(max.\ sig.)}$ | 2 × 107·5 | 2 × 110·5 | mA |
| $I_{g2(max.\ sig.)}$ | 2 × 23·5 | 2 × 23 | mA |

*Screen-grid resistor common to both valves.

| Cathode bias | | | |
|---|---|---|---|
| $V_b$ | 375 | 450 | V |
| $V_{g3}$ | 0 | 0 | V |
| $*R_{g2}$ | 0·47 | 1·0 | k$\Omega$ |
| $R_k$ (per valve) | 260 | 465 | $\Omega$ |
| $R_{a-a}$ | 3·5 | 6·5 | k$\Omega$ |
| $I_{a(o)}$ | 2 × 75 | 2 × 60 | mA |
| $I_{g2(o)}$ | 2 × 12·5 | 2 × 10 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 40 | 54 | V |
| $P_{out}$ | 34 | 40 | W |
| $D_{tot}$ | 1·7 | 5·1 | % |
| $I_{a(max.\ sig.)}$ | 2 × 94 | 2 × 71·5 | mA |
| $I_{g2(max.\ sig.)}$ | 2 × 19·5 | 2 × 22 | mA |

*Screen-grid resistor common to both valves.

## Operating Conditions for Two Valves in Push-Pull

Triode connection ($g_2$ connected to a, $g_3$ to k) with separate cathode bias resistors.

| With $R_k$ bypassed | | |
|---|---|---|
| $V_b$ | 430 | V |
| $V_a$ | 400 | V |
| $V_{g3}$ | 0 | V |
| $R_k$ (per valve) | 440 | $\Omega$ |
| $R_{a-a}$ | 5·0 | k$\Omega$ |
| $I_{a(o)}$ | 2 × 70 | mA |
| $V_{in(g1-g1)r.m.s.}$ | 48 | V |
| $P_{out}$ | 19 | W |
| $D_{tot}$ | 1·8 | % |
| $I_{a(max.\ sig.)}$ | 2 × 75 | mA |

| | | |
|---|---|---|
| $V_b$ | 430 | V |
| $V_a$ | 400 | V |
| $V_{g3}$ | 0 | V |
| $R_k$ (per valve) | 440 | $\Omega$ |
| $R_{a-a}$ | 10 | k$\Omega$ |
| $I_{a(o)}$ | $2 \times 70$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 48 | V |
| $P_{out}$ | 14 | W |
| $D_{tot}$ | 0·4 | % |
| $I_{a(max. sig.)}$ | $2 \times 73$ | mA |

## Operating Conditions for Two Valves in Push-Pull with Continuous Sine Wave Drive

### Fixed bias

| | | | |
|---|---|---|---|
| $V_b$ | 375 | 400 | V |
| $V_{g3}$ | 0 | 0 | V |
| $R_{g2}$ | 1·0 | 1·5 | k$\Omega$ |
| $V_{g1}$ | −32 | −35·5 | V |
| $R_{a-a}$ | 3·5 | 3·5 | k$\Omega$ |
| $I_{a(o)}$ | $2 \times 30$ | $2 \times 30$ | mA |
| $I_{g2(o)}$ | $2 \times 4·4$ | $2 \times 4·4$ | mA |
| $V_{in(g1-g1)r.m.s.}$ | 45 | 50 | V |
| $P_{out}$ | 42 | 51 | W |
| $D_{tot}$ | 3·0 | 1·8 | % |
| $I_{a(max. sig.)}$ | $2 \times 98$ | $2 \times 106$ | mA |
| $I_{g2(max. sig.)}$ | $2 \times 19$ | $2 \times 21$ | mA |

### Cathode bias

Any of the cathode bias conditions published in this data sheet are suitable for continuous sine wave drive.

## Design Centre Ratings

| | | |
|---|---|---|
| $V_{a(b)}$ max. | 2·0 | kV |
| $V_a$ max. | 800 | V |
| $p_a$ max. | 25 | W |
| $V_{g2(b)}$ max. | 800 | V |
| $V_{g2}$ max. | 500 | V |
| $p_{g2}$ max. | 8·0 | W |
| $I_k$ max. | 150 | mA |
| $R_{g1-k}$ max. | 500 | k$\Omega$ |
| $V_{h-k}$ max. | 100 | V |
| $R_{h-k}$ max. | 20 | k$\Omega$ |

### Triode connected

| | | |
|---|---|---|
| $V_a$ max. | 600 | V |
| $p_{a+g2}$ max. ($V_a = 500$ V) | 30 | W |
| $p_{a+g2}$ max. ($V_a = 600$ V) | 15 | W |

# BEAM TETRODE

*The KT88 has an absolute maximum anode dissipation rating of 42W and is designed for use in the output stage of an a.f. amplifier. Two valves in Class AB1 give a continuous output of up to 100 W. The KT88 is also suitable for use as a series valve in a stabilised power supply.*

*The KT88 is a commercial version of the CV5220 and is similar to the 6550.*

## Base connections and valve dimensions

Base: Metal shell wafer octal
Bulb: Tubular

Max. overall length: 125 mm
Max. seated length: 110 mm
Max. diameter: 52 mm

$g_2$ — 4
$g_1$ — 5
a — 3
6 — NP
h — 2
7 — h
s — 1
8 — k, bp

View from underside of base.
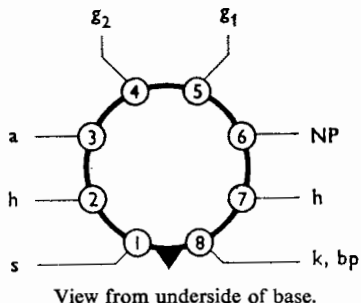
## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 1·6 (approx) | A |

| Maximum ratings | | | |
|---|---|---|---|
| | **Absolute** | **Design Maximum** | |
| $V_a$ | 800 | 800 | V |
| $V_{g2}$ | 600 | 600 | V |
| $V_{a,g2}$ | 600 | 600 | V |
| $-V_{g1}$ | 200 | 200 | V |
| $p_a$ | 42 | 35 | W |
| $p_{g2}$ | 8 | 6 | W |
| $p_{a+g2}$ | 46 | 40 | W |
| $I_k$ | 230 | 230 | mA |
| $V_{h-k}$ | 250 | 200 | V |
| $T_{bulb}$ | 250 | 250 | °C |
| $R_{g1-k}$ (cathode bias) | | | |
| $\quad p_{a+g2} \leqslant 35W$ | | 470 | kΩ |
| $\quad p_{a+g2} > 35W$ | | 270 | kΩ |
| $R_{g1-k}$ (fixed bias) | | | |
| $\quad p_{a+g2} \leqslant 35$ W | | 220 | kΩ |
| $\quad p_{a+g2} > 35$ W | | 100 | kΩ |
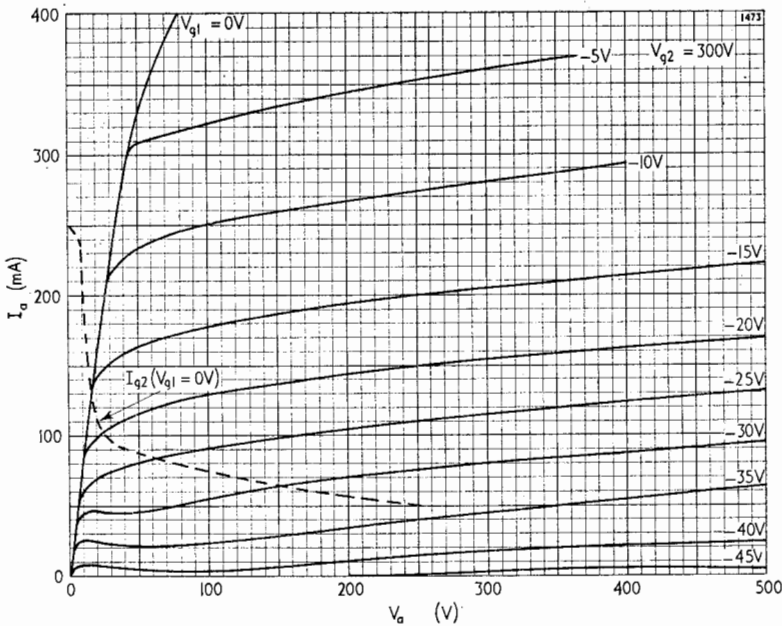
KT88. Anode and screen-grid current plotted against anode voltage with control-grid voltage as parameter. $V_{g2} = 300$ V.

## Capacitances (measured on a cold unscreened valve)
### Triode connection

$c_{g1-a,g2}$: $7 \cdot 9$pF    $c_{g1-\text{all less a,g2}}$: $9 \cdot 3$pF    $c_{a,g2-\text{all less g1}}$: $17$pF

### Tetrode connection

$c_{g1-a}$: $1 \cdot 2$pF    $c_{g1-\text{all less a}}$: $16$pF    $c_{a-\text{all less g1}}$: $12$pF

## Characteristics
### Tetrode connected

| | | |
|---|---|---|
| $V_a$ | 250 | V |
| $V_{g2}$ | 250 | V |
| $I_a$ | 140 | mA |
| $I_{g2}$ | 3 (approx) | mA |
| $-V_{g1}$ | 15 (approx) | V |
| $g_m$ | $11 \cdot 5$ | mA/V |
| $r_a$ | 12 | k$\Omega$ |
| $\mu_{g1-g2}$ | 8 | — |

### Triode connected

| | | |
|---|---|---|
| $V_{a,g2}$ | 250 | V |
| $I_{a+g2}$ | 143 | mA |
| $-V_{g1}$ | 15 (approx) | V |

227

| | | |
|---|---|---|
| $g_m$ | 12 | mA/V |
| $r_a$ | 670 | $\Omega$ |
| $\mu$ | 8 | — |

## Typical Operation

### Push-pull. Class AB1. Cathode bias. Tetrode connection

| | | |
|---|---|---|
| $V_{a(b)}$ | 560 | V |
| $V_{a(o)}$ | 521 | V |
| $V_{g2}$ | 300 | V |
| $I_{a(o)}$ | $2 \times 64$ | mA |
| $I_{a(max.\ sig.)}$ | $2 \times 73$ | mA |
| $I_{g2(o)}$ | $2 \times 1\cdot7$ | mA |
| $I_{g2(max.\ sig.)}$ | $2 \times 9$ | mA |
| $R_{L(a-a)}$ | 9 | k$\Omega$ |
| *$R_k$ | $2 \times 460$ | $\Omega$ |
| $-V_{g1}$ | 30 (approx) | V |
| $P_{out}$ | 50 | W |
| $D_{tot}$ | 3 | % |
| †I.M. | 11 | % |
| $p_{a(o)}$ | $2 \times 33$ | W |
| $p_{a(max.\ sig.)}$ | $2 \times 12$ | W |
| $p_{g2(o)}$ | $2 \times 0\cdot5$ | W |
| $p_{g2(max.\ sig.)}$ | $2 \times 2\cdot7$ | W |
| $V_{in(g1-g1)(pk)}$ | 60 | V |

*It is essential to use two separate cathode bias resistors.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

### Push-pull. Class AB1. Fixed bias. Tetrode connection

| | | |
|---|---|---|
| $V_{a(b)}$ | 560 | V |
| $V_{a(o)}$ | 552 | V |
| $V_{g2}$ | 300 | V |
| $I_{a(o)}$ | $2 \times 60$ | mA |
| $I_{a(max.\ sig.)}$ | $2 \times 145$ | mA |
| $I_{g2(o)}$ | $2 \times 1\cdot7$ | mA |
| $I_{g2(max.\ sig.)}$ | $2 \times 15$ | mA |
| $R_{L(a-a)}$ | $4\cdot5$ | k$\Omega$ |
| *$-V_{g1}$ | 34 (approx) | V |
| $P_{out}$ | 100 | W |
| $D_{tot}$ | $2\cdot5$ | % |
| †I.M. | 10 | % |
| $p_{a(o)}$ | $2 \times 33$ | W |
| $p_{a(max.\ sig.)}$ | $2 \times 28$ | W |
| $p_{g2(o)}$ | $2 \times 0\cdot5$ | W |
| $p_{g2(max.\ sig.)}$ | $2 \times 4\cdot5$ | W |
| $V_{in(g1-g1)(pk)}$ | 67 | V |

*It is essential to provide two separately adjustable bias voltage sources, having a voltage adjustment range of $\pm25\%$.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).
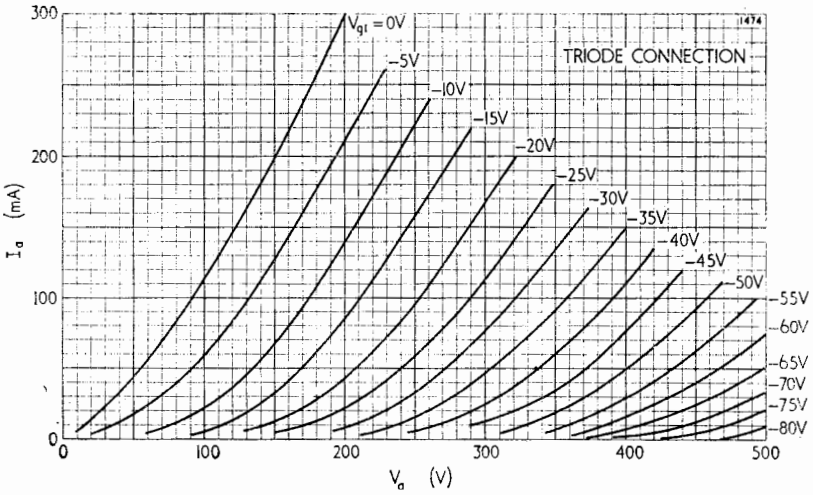
KT88. Anode current plotted against anode voltage with control-grid as parameter when triode connected.
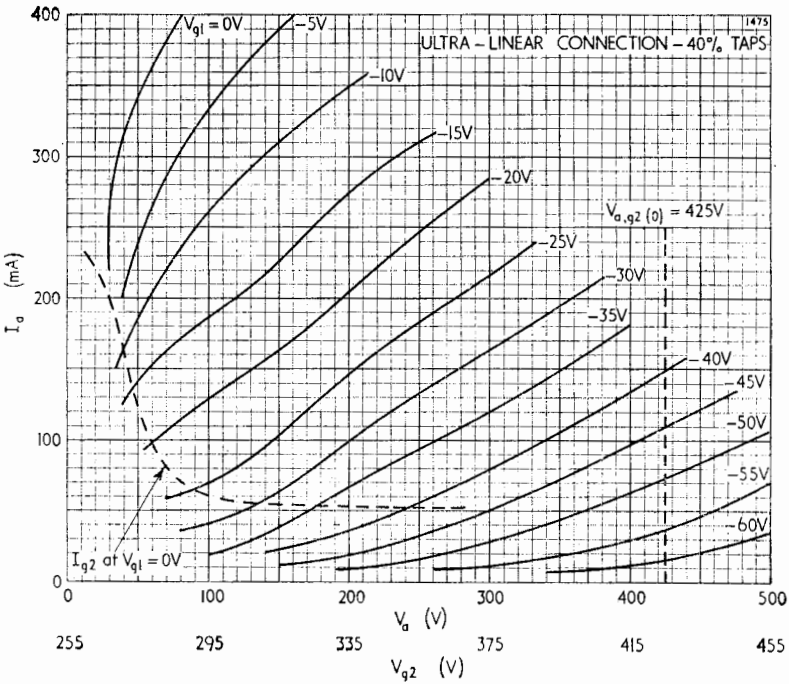


KT88. Anode and screen-grid current plotted against voltage with control-grid as parameter when 40% ultralinear connected.

229

Push-pull. Class AB1. Cathode bias. Ultralinear connection (40% Tapping points)

| | | | |
|---|---|---|---|
| $V_{a,g2(b)}$ | 500 | 375 | V |
| $V_{a,g2(o)}$ | 436 | 328 | V |
| $I_{a+g2(o)}$ | $2 \times 87$ | $2 \times 87$ | mA |
| $I_{a+g2(max. sig.)}$ | $2 \times 99$ | $2 \times 96$ | mA |
| $R_{L(a-a)}$ | 6 | 5 | k$\Omega$ |
| *$R_k$ | $2 \times 600$ | $2 \times 400$ | $\Omega$ |
| $-V_{g1}$ | 52 (approx) | 35 (approx) | V |
| $P_{out}$ | 50 | 30 | W |
| $D_{tot}$ | 1·5 | 1 | % |
| †I.M. | 4 | 3 | % |
| $p_{a+g2(o)}$ | $2 \times 38$ | $2 \times 28·5$ | W |
| $p_{a+g2(max. sig.)}$ | $2 \times 17$ | $2 \times 16$ | W |
| $v_{in(g1-g1)(pk)}$ | 104 | 71 | V |
| $Z_{out}$ | 4·8 | 4·5 | k$\Omega$ |

*It is essential to use two separate cathode bias resistors.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

Push-pull. Class AB1. Fixed bias. Ultralinear connection (40% Tapping points)

| | | | |
|---|---|---|---|
| $V_{a,g2(b)}$ | 560 | 460 | V |
| $V_{a,g2(o)}$ | 553 | 453 | V |
| $I_{a+g2(o)}$ | $2 \times 50$ | $2 \times 50$ | mA |
| $I_{a+g2(max. sig.)}$ | $2 \times 157$ | $2 \times 140$ | mA |
| $R_{L(a-a)}$ | 4·5 | 4 | k$\Omega$ |
| *$-V_{g1}$ | 75 (approx) | 59 (approx) | V |
| $P_{out}$ | 100 | 70 | W |
| $D_{tot}$ | 2 | 2 | % |
| †I.M. | 11 | 10 | % |
| $p_{a+g2(o)}$ | $2 \times 27·5$ | $2 \times 22·5$ | W |
| $p_{a+g2(max. sig.)}$ | $2 \times 33$ | $2 \times 27$ | W |
| $v_{in(g1-g1)(pk)}$ | 140 | 114 | V |
| $Z_{out}$ | 7 | 6·5 | k$\Omega$ |

*It is essential to provide two separately adjustable bias voltage sources, having a voltage adjustment range of $\pm 25\%$.
†Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

Push-pull. Class AB1. Cathode bias. Triode connection

| | | | |
|---|---|---|---|
| $V_{a,g2(b)}$ | 400 | 485 | V |
| $V_{a,g2(o)}$ | 349 | 422 | V |
| $I_{a+g2(o)}$ | $2 \times 76$ | $2 \times 94$ | mA |
| $I_{a+g2(max. sig.)}$ | $2 \times 80$ | $2 \times 101$ | mA |
| $R_{L(a-a)}$ | 4 | 4 | k$\Omega$ |

| | 40 (approx) | 50 (approx) | V |
|---|---|---|---|
| $-V_{g1}$ | 40 (approx) | 50 (approx) | V |
| $P_{out}$ | 17 | 31 | W |
| $D_{tot}$ | 1·5 | 1·5 | % |
| *I.M. | 5·6 | 5·6 | % |
| $P_{a+g2(o)}$ | $2 \times 26·5$ | $2 \times 40$ | W |
| $P_{a+g2(max. sig.)}$ | $2 \times 19$ | $2 \times 27$ | W |
| $R_k$ | $2 \times 525$ | $2 \times 525$ | $\Omega$ |
| $V_{in(g1-g1)(pk)}$ | 78 | 114 | V |
| $Z_{out}$ | 2 | 1·9 | k$\Omega$ |

*Intermodulation distortion; measured using two input signals at 50 and 6000 c/s (ratio of amplitudes 4:1).

## INSTALLATION

The valve may be mounted either vertically or horizontally.

When a pair of valves is mounted vertically it is recommended that the centres of the valve sockets are not less than 4 in. apart and that pins 4 and 8 of each valve are in line.

When a pair of valves is mounted horizontally it is recommended that the centres of the valve sockets are not less than 4 in. apart and that pins 4 and 8 of each valve are in the same vertical line.

Free air circulation around the valve is desirable.

The KT88 may be used in pairs in either triode, tetrode, or ultralinear push-pull circuits for outputs of 30 to 400 watts in various conditions of operation. The ultralinear circuit is recommended for outputs up to 100 W when the maximum output coupled with the lowest distortion is required and it may also be used with multiple pairs, four pairs producing 400 W.

The curves opposite show the output and approximate distortion obtained from push-pull KT88 valves with various positions of the screen grid taps on the output transformer. At the left-hand side (0 % taps) the valves are, of course, tetrode-connected and on the right-hand side (100% taps) they are triode-connected. Generally speaking, the taps should be spaced at 20% to 50% of the turns on each half-primary from the centre, with 30% to 40% as the optimum. The dotted curve shows the reduction in output impedance obtained.

Either cathode bias or fixed bias may be used. The former has the advantage of simplicity whereas the latter provides the normal maximum output of 100 W from a pair of valves and higher efficiency.

The output power of the ultralinear circuit is not less than that of the tetrode at any given supply voltage and has the advantage that a low impedance screen supply is not required. The ultralinear circuit does, in fact, show a rather higher efficiency in that a lower current is required from the power supply. For example, with fixed bias and a supply voltage of 460 an output of 70W is obtained with an anode current of 280mA in both the tetrode and the ultralinear arrangements but the tetrode requires, in addition, a screen current of 35 mA. Furthermore, it is desirable that the tetrode screen supply be stabilised. The output impedance and distortion are both more favourable in the ultralinear circuit, the former being 6·5 k$\Omega$ in this example, compared with 50 k$\Omega$, and the distortion is almost entirely independent of load impedance above the rated value.

231

Screen tapping points). (% turns)
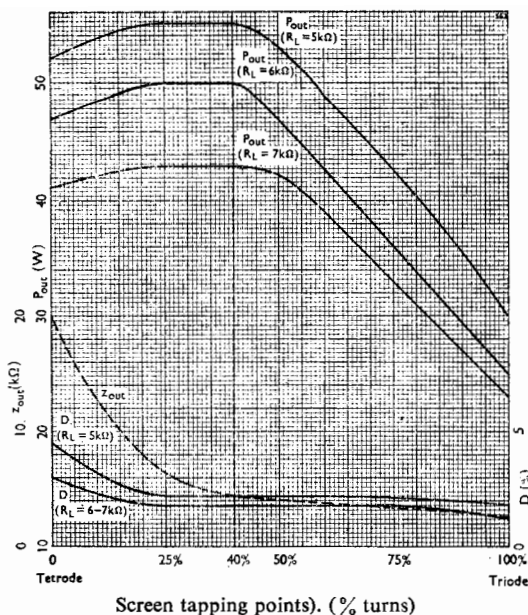
The triode connection is sometimes preferred when a moderate power output with low distortion is required. An output of up to 31 W is obtainable with an h.t. supply of 485 V and cathode bias and, even at lower voltages, the output is adequate for domestic amplifiers. The distortion will depend on the degree of matching between pairs but it is normal to obtain a distortion below 2% without negative feedback by the selection of two out of any three valves since this procedure, at worst, halves the normal variation in characteristics.

Due to the high mutual conductance of the KT88 some precautions have to be taken against parasitic oscillation whatever the circuit arrangement. Grid and screen series resistors of about 10 kΩ and 270 Ω respectively, are recommended. In the ultralinear circuit, resistance-capacitance networks may be found necessary between the anode and screen taps on the output transformer. With some transformers they will not be needed and they become less necessary as the taps include a greater part of the primary due to the consequent reduction in leakage inductance.

When an output exceeding 100 W is needed, the KT88 valve may be used in multiple pairs in parallel push-pull instead of a single pair of larger valves. One of the advantages of this method is the low cost of the power supply which is required to give an h.t. of only 550 V. Another advantage is that a valve failure in the output stage merely reduces the available output power with a probable increase in distortion.

The bulb temperature rating of 250°C must not be exceeded and any cabinet used must provide adequate ventilation. Under free air conditions this temperature is not reached at maximum ratings, but where a valve is enclosed it is recommended to check that adequate air circulation exists.

232

Although the KT88 may be mounted in any desired position it is recommended that, when valves are mounted horizontally, pins 4 and 8 should be in a vertical line as in fig. 2.
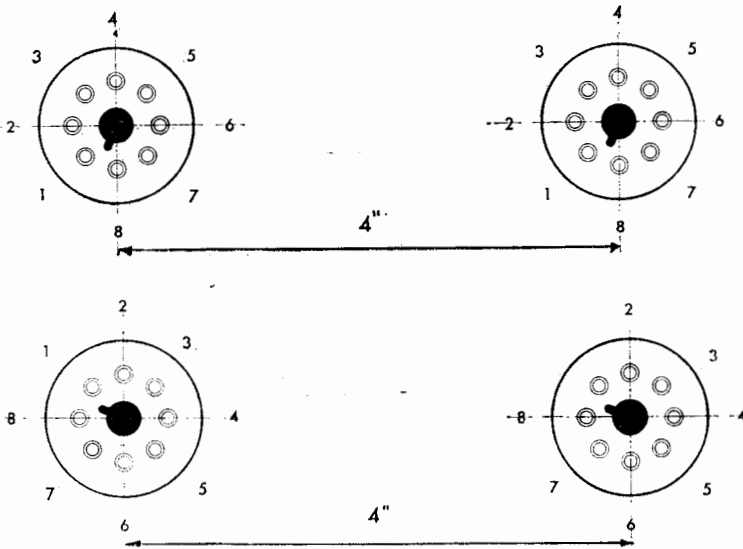


Correct orientation and spacing of the valve sockets when the valves are mounted horizonally (top) and vertically (below).

When two valves are mounted vertically, in the more usual manner, on a horizontal chassis, there is a slight advantage obtained if pins 4 and 8 of both valves are mounted in line, as in fig. 3. The coolest part of the bulbs then face each other whilst the hottest parts (nearest the anodes) are at right-angles.

### Power supply

The type of power supply required will depend on the operating conditions but a capacitance input filter circuit is satisfactory for ultralinear cathode bias amplifiers.

With fixed bias, the large change in anode current requires a low impedance power supply and an inductance input filter is essential. It is desirable for the smoothing capacitor to be of high value to prevent an instantaneous fall in voltage upon the occurrence of a transient. Satisfactory performance will be obtained with a single inductor and a capacitance of $50$–$150\ \mu\text{F}$. Two $160\ \mu\text{F}$, $450$ V electrolytic capacitors in series have been found very successful with the ultralinear higher voltage working conditions.

The choice of a rectifier will depend on the power output required. The U52 or U54 are suitable for powers up to 50 W, but for amplifiers designed to cover the range 50–100 W, two U19 or two GXU50 rectifiers should be used

*Indirectly heated full-wave rectifier primarily intended for use in a.c. mains operated equipment.*

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 5·0 | V |
| $I_h$ | 1·9 | A |

## Limiting Values

| | | |
|---|---|---|
| P.I.V. max. | 1·5 | kV |
| $i_{a(pk)}$ max. | 750 | mA |
| C max. | 60 | $\mu$F |
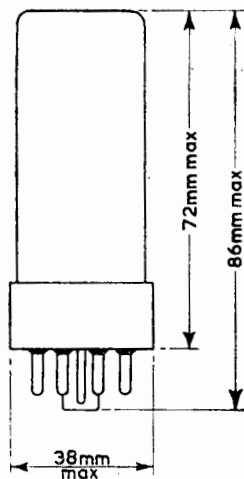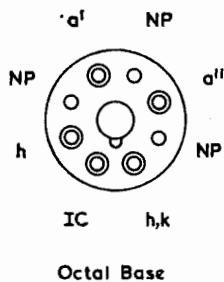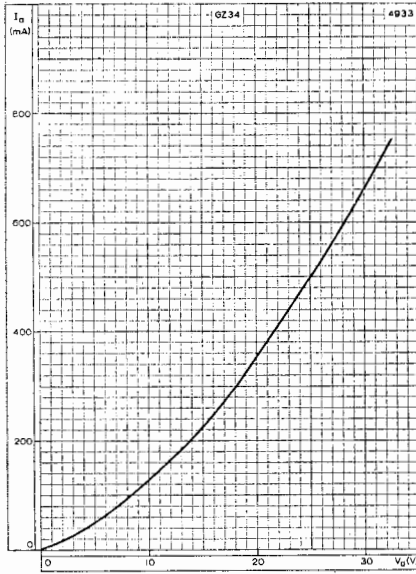


Octal Base

Anode current plotted against anode voltage.



Boundary of operation with capacitor or choke input filter.

235
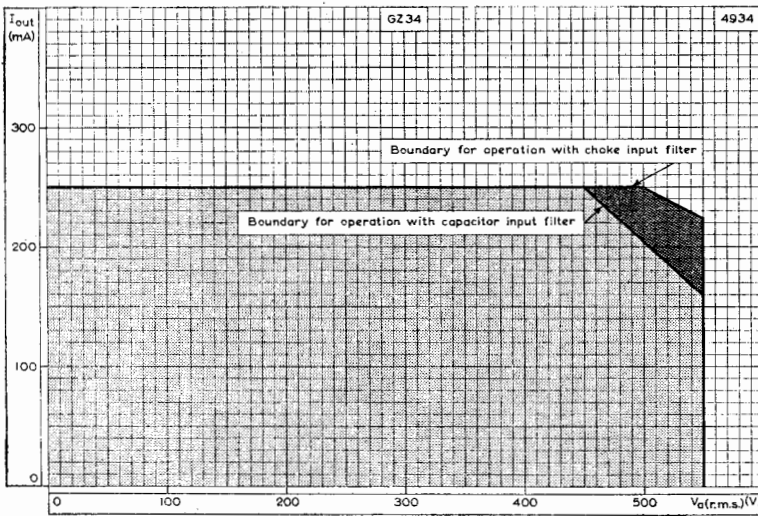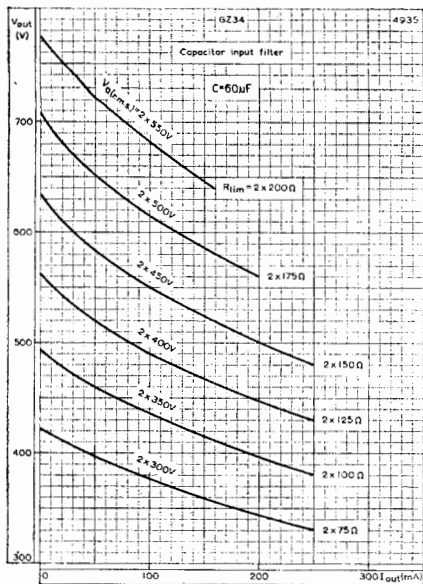
Capacitor input filter regulation curves.
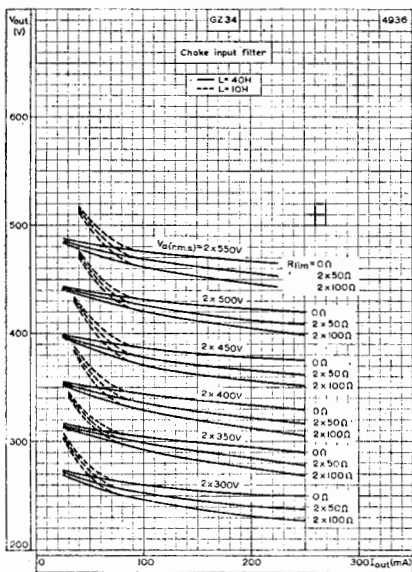


Choke input filter regulation curves.

236

| $V_{a(r.m.s.)}$ | $2 \times 300$ | $2 \times 350$ | $2 \times 400$ | $2 \times 450$ | $2 \times 500$ | $2 \times 550$ | |
|---|---|---|---|---|---|---|---|
| | | | Capacitor input | | | | |
| $I_{out}$ max. | 250 | 250 | 250 | 250 | 200 | 160 | mA |
| $R_{lim}$ min. (per anode) | 50 | 75 | 100 | 125 | 150 | 175 | Ω |
| | | | Choke input | | | | |
| $I_{out}$ max. | 250 | 250 | 250 | 250 | 250 | 225 | mA |
| $R_{lim}$ min. (per anode) | 0 | 0 | 0 | 0 | 0 | 0 | Ω |

## Typical Operating Conditions

| | | Capacitor input | | |
|---|---|---|---|---|
| $V_{a(r.m.s.)}$ | $I_{out}$ | C | $R_{lim}$ (per anode) | $V_{out}$ |
| (V) | (mA) | ($\mu$F) | (Ω) | (V) |
| $2 \times 300$ | 250 | 60 | 75 | 330 |
| $2 \times 350$ | 250 | 60 | 100 | 380 |
| $2 \times 400$ | 250 | 60 | 125 | 430 |
| $2 \times 450$ | 250 | 60 | 150 | 480 |
| $2 \times 500$ | 200 | 60 | 175 | 560 |
| $2 \times 550$ | 160 | 60 | 200 | 640 |

| | | Choke input | | |
|---|---|---|---|---|
| $V_{a(r.m.s.)}$ | $I_{out}$ | L | $R_{lim}$ (per anode) | $V_{out}$ |
| (V) | (mA) | (H) | (Ω) | (V) |
| $2 \times 300$ | 250 | 10 | 0 | 250 |
| $2 \times 350$ | 250 | 10 | 0 | 290 |
| $2 \times 400$ | 250 | 10 | 0 | 330 |
| $2 \times 450$ | 250 | 10 | 0 | 375 |
| $2 \times 500$ | 250 | 10 | 0 | 420 |
| $2 \times 550$ | 225 | 10 | 0 | 465 |

## EZ80 *(•x·)*   FULL-WAVE RECTIFIER

*Indirectly-heated full-wave rectifier, primarily intended for use in a.c. mains-operated equipment*

### Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 6·3 | V |
| $I_h$ | 600 | mA |

### Limiting Values

| | | |
|---|---|---|
| $V_{a(r.m.s.)}$ max. | $2 \times 350$ | V |
| $I_{out}$ max. | 90 | mA |
| C max. | 50 | $\mu$F |
| $i_{a(pk)}$ max. (per anode) | 270 | mA |
| $v_{h-k(pk)}$ max. | 500 | V |

### Operating Conditions

| | | | | | |
|---|---|---|---|---|---|
| $V_{in(r.m.s.)}$ | $2 \times 250$ | $2 \times 275$ | $2 \times 300$ | $2 \times 350$ | V |
| $R_{lim}$ min. (per anode) | 125 | 175 | 215 | 300 | $\Omega$ |
| C | 50 | 50 | 50 | 50 | $\mu$F |
| $I_{out}$ | 90 | 90 | 90 | 90 | mA |
| $V_{out}$ | 265 | 285 | 310 | 360 | V |



B9A Base
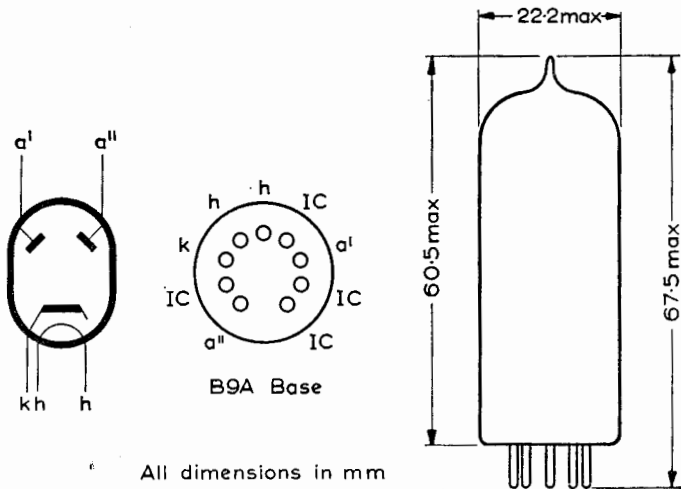
All dimensions in mm
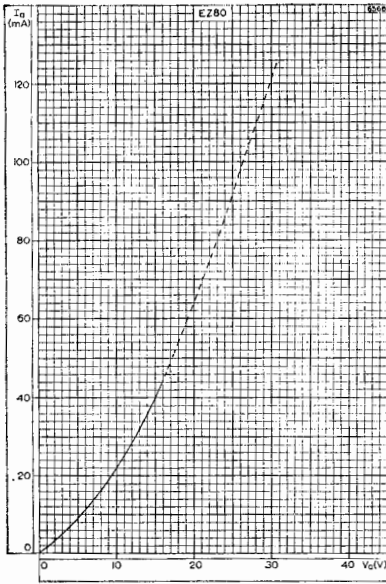
Anode current plotted against anode voltage.
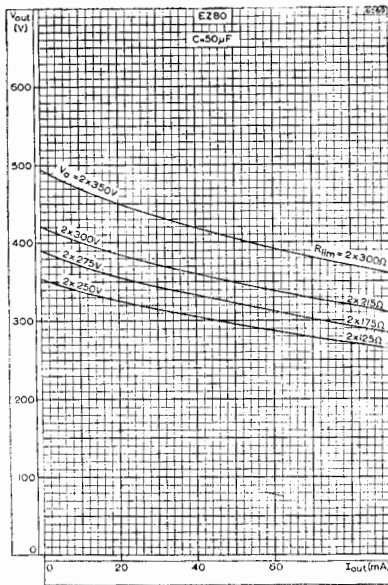


Regulation curves.

# EZ81      FULL-WAVE RECTIFIER

*Indirectly heated full-wave rectifier with $6 \cdot 3$ V heater.*

## Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | $6 \cdot 3$ | V |
| $I_h$ | $1 \cdot 0$ | A |

## Limiting Values

| | | |
|---|---|---|
| P.I.V. max. | $1 \cdot 3$ | kV |
| $i_{a(pk)}$ max. | 500 | mA |
| $i_{a(surge)}$ max. | $1 \cdot 8$ | A |
| $V_{h-k}$ max. (cathode positive) | 500 | V |

| Capacitor input | | |
|---|---|---|
| $I_{out}$ max.   $\Big\}$ See rating chart 1 | | |
| $V_{in(r.m.s.)}$ max. | | |
| $R_{lim}$ min. | See rating charts 2 and 3 and capacitor input regulation curves. | |

| Choke input | | |
|---|---|---|
| $I_{out}$ max.   $\Big\}$ See rating chart 1 | | |
| $V_{in(r.m.s.)}$ max. | | |
| $L_{min}$ (at 50 c/s) | See choke regulation curves | |



a′   a″

h   h   IC
k   a′
IC   IC
a″   IC

B9A
Noval Base

k h   h

22·2mm max

71·5mm max

78·5mm max
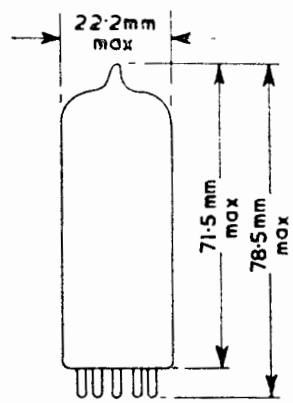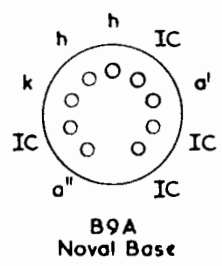
| Characteristic | | |
|---|---|---|
| Anode voltage drop ($I_{out} = 150$ mA) | $19 \cdot 8$ | V |
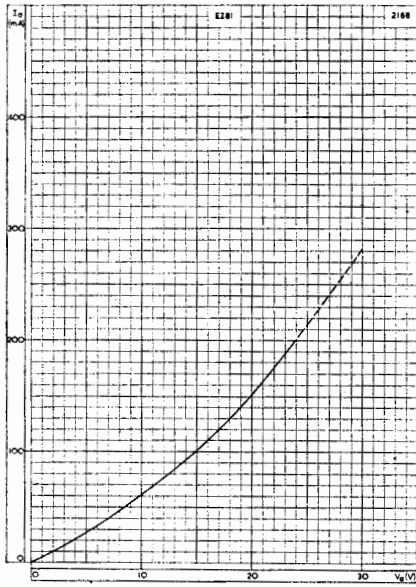
Anode current plotted against anode voltage.
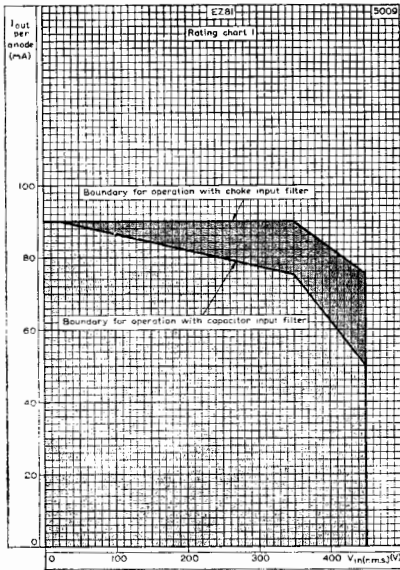


Boundary of operation with capacitor or choke input filter.

*Operating Conditions*

| Capacitor input | | | | |
|---|---|---|---|---|
| $V_{iu(r.m.s.)}$ | 2 × 250 | 2 × 350 | 2 × 450 | V |
| $R_{lim}$ (per anode) | 150 | 230 | 310 | Ω |
| C | 50 | 50 | 50 | $\mu$F |
| $I_{out}$ | 160 | 150 | 100 | mA |
| $V_{out}$ | 245 | 352 | 497 | V |

| Choke input | | | | |
|---|---|---|---|---|
| $V_{in(r.m.s.)}$ | 2 × 250 | 2 × 350 | 2 × 450 | V |
| L | 10 | 10 | 10 | H |
| $I_{out}$ | 180 | 180 | 150 | mA |
| $V_{out}$ | 199 | 288 | 378 | V |

## OPERATING NOTES

The design of a power circuit starts with a knowledge of the output conditions and from this information the transformer and secondary or input voltage can be chosen. Reference to the rating charts will indicate whether a rectifier is suitable for a particular application.

### Rating chart 1

This shows all the combinations of input voltage and output current considered safe for both capacitor and choke input filters.

### Rating chart 2

This chart shows the minimum series resistor per anode necessary to restrict the maximum switching surge in a capacitor input filter, to its limiting value over the range of supply voltages.

### Rating chart 3

This shows the relationship between the maximum rectification efficiency and output current.

### Capacitor input filter circuits

Reference should be made to rating charts 2 and 3 and the regulation curves. The circuit is set up and the input and output voltage and output current are measured. If the operating conditions lie within the boundary lines of the regulation curves, an improvement in the rectification efficiency may be effected by reducing the value of the limiting resistance. Rating chart 2 gives the minimum value of the limiting resistance against open circuit secondary,

242

voltage; this resistance will guard against excessive switching currents.

Comparison of the calculated rectification efficiency $\dfrac{V_{out}}{\sqrt{2} \times V_{in(r.m.s.)}}$ with rating chart 3 will show whether the limiting resistance must be increased to lower the rectification efficiency to the area of safe operation. Operation within this area indicates that the limiting value $i_{a(pk)}$ has not been exceeded.

### Choke input circuit

Reference should be made to rating chart 1. A suitable value of choke can be obtained from the choke regulation curves.

## U54/GZ37    BI-PHASE HALF-WAVE RECTIFIER

*An indirectly heated rectifier capable of supplying up to 250 mA for a 500 V r.m.s. input to each anode in capacitance input filter circuits and up to 350 mA for a similar input in inductance input filter circuits.*

### Base connections and valve dimensions

Base: B8-0 (Octal)
Bulb: Dome topped pear.
Max. overall length:    142 mm
Max. seated length:    128 mm
Max. diameter:    57·5 mm

View from underside of base.

a′    NP

NP ——③        ⑥ —— a″

h ——②        ⑦ —— NP

IC ——        h, k
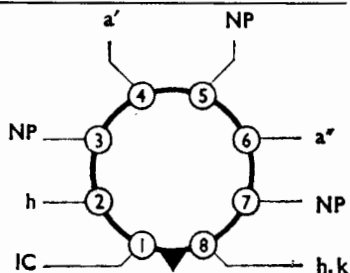
### Valve Data

| Heater | | |
|---|---|---|
| $V_h$ | 5·0 | V |
| $I_h$ | 2·8 (approx) | A |

| Maximum ratings (design centre) | | | |
|---|---|---|---|
| | Capacitance input filter | Inductance input filter | |
| PIV | 1·6 | 1·85 | kV |
| $i_{a(pk)}$ (each anode) | 0·75 | — | A |
| $I_{out}$ | 250 | 350 | mA |
| $R_{source}$ (min) (each anode) | 75 | — | $\Omega$ |
| $T_{bulb}$ | 225 | 225 | °C |

### Typical Operation

| Bi-Phase Half-Wave Rectifier. Capacitance Input Filter (Fig. 1) | | | | |
|---|---|---|---|---|
| $V_{a(r.m.s.)}$ (each anode) | 300 | 400 | 500 | V |
| $C_{reservoir}$ | 4·0 | 4·0 | 4·0 | $\mu$F |
| $R_{source}$ (each anode) | 75 | 75 | 75 | $\Omega$ |
| $I_{out}$ (full load) | 250 | 250 | 250 | mA |
| $V_{out}$ (full load) | 238 | 358 | 486 | V |

| Bi-Phase Half-Wave Rectifier. Inductance Input Filter (Fig. 2) | | | | |
|---|---|---|---|---|
| (Inductor resistance 100 $\Omega$) | | | | |
| $V_{a(r.m.s.)}$ (each anode) | 300 | 400 | 500 | V |
| L | 10 | 10 | 10 | H |
| $I_{out}$ (full load) | 350 | 350 | 350 | mA |
| $V_{out}$ (full load) | 207 | 298 | 381 | V |

244
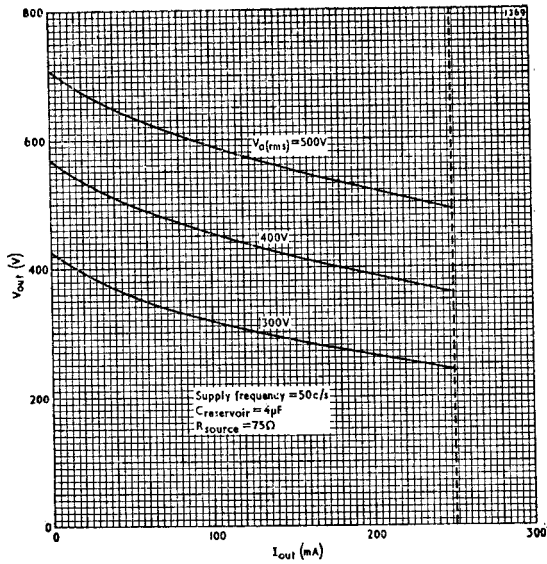
Capacitance input filter operation.
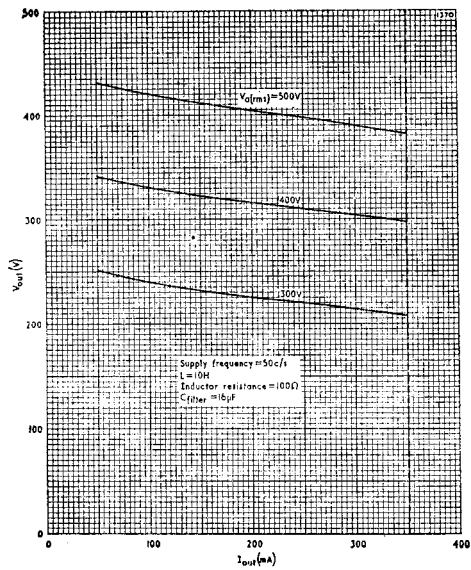


Inductance input filter operation.

# REFERENCES

1. UL OUTPUT TRANSFORMERS. D. M. Leakey and R. B. Gibson. *Wireless World*, January 1956.

2. HOW GOOD IS YOUR AMPLIFIER, SERIES. J. Somerset Murray. *Hi-Fi News*. IMPULSE TESTING, February 1959. LOW FREQUENCY POWER OUTPUT LIMIT, April 1959. HIGH FREQUENCY POWER LOSSES IN THE TRANSFORMER, December 1959. LEAKAGE RESONANCE INTERMODULATION DISTORTION, January 1960.

3. STABILIZING FEEDBACK AMPLIFIERS. T. Roddam. *Wireless World*. March 1951.

4. NEGATIVE FEEDBACK TONE CONTROL. P. J. Baxandall. *Wireless World*. October 1952.

5. GRAMOPHONE AND MICROPHONE PREAMPLIFIER. P. J. Baxandall. *Wireless World*. January and February 1955.

# GLOSSARY OF TERMS

A.C./D.C. – Equipment capable of operating from a.c. or d.c. mains supplies without circuit modification. A transformer cannot be employed so that valve heaters are connected in series.

AMPERE – One ampere is the current which, passing through a solution of nitrate of silver in water, deposits silver at the rate of $0 \cdot 00111800$ gram per second. This represents a flow of approximately $6 \times 10^{18}$ electrons.

AMPLIFIER – Apparatus which produces an electric output that is greater than, but in all other respects identical to, the signal applied to the input.

AMPLITUDE – The maximum displacement of a moving particle from the mean position. Synonymous with 'peak' (page 17).

ANGULAR VELOCITY – If $f$ revolutions occur in one second and each revolution represents $360°$ (or $2\pi$ radians), then $2\pi f$ radians are swept out in one second and this is called the angular velocity. Symbol $\omega$ (see Sine wave).

ATTENUATOR – A device for reducing voltage, current or power. The simplest forms are the resistor potential divider (fixed attenuator) and the potentiometer or volume control (variable attenuator). The output resistance varies from zero to that of the control and thus depends upon the setting. By using multiple resistors the input and/or the output resistance can be made constant and popular arrangements are the T section and the $\pi$ section.

AUTOMATIC BIAS – A means of producing the correct d.c. operating conditions in a valve or transistor circuit without the use of bias batteries.

AUDIO FREQUENCIES – Sound waves of a pitch audible to the human ear. The lowest frequency which the ear will recognise as a continuous tone is about 16 c/s whilst the upper frequency limit is in the region of 10,000 to 16,000 c/s. The upper limit varies from person to person usually dropping with increasing age.

AUDIO POWER – The energy delivered into the air by a loudspeaker, measured in watts. Loudspeaker efficiency varies from unit to unit over a wide range; from about 3% to 25%. The power required from the amplifier driving a loudspeaker, to produce a given sound level, therefore, depends upon loudspeaker efficiency. To produce an audio power of 1 watt will require a driving power of 33 watts when the loudspeaker is 3% efficient, and 4 watts when the loudspeaker is 25% efficient. It should be remembered that an increase of power of 3 dB produces the minimum change in sound level which the ear can detect and this represents a power increase of $\times 2$. To give a worthwhile increase on audio power, it is therefore necessary at least to double the power output of the amplifier.

CASCODE AMPLIFIER – An amplifier stage which employs two valves in series, so that one valve becomes a part of the anode load of the other. The effect is equivalent to a single valve with a high $\mu$, capable of a high stage gain. Cascode triode amplifiers can provide a high signal-to-noise ratio, but the circuit is complex and must be used with care.

DECIBEL – Unit which expresses the ratio of two different levels of sound intensity or electrical power.

DECOUPLING – A *CR* or *LR* circuit used to prevent a portion of the signal developing in a part of the circuit where it is not required.

DE-EMPHASIS – *See* Pre-emphasis.

DISTORTION – Any difference between the input and output signals of an amplifier excepting amplication. Non-linearity distortion is mainly due to curvature of valve characteristics and can be measured in terms of harmonics. If a sine wave is passed through an amplifier with non-linearity distortion, harmonics are produced which are whole number multiples of the sine wave frequency. The harmonics present in the output can be measured by a 'distortion meter' or calculated by a Fourier analysis of the output waveform. When distortion is small, total harmonic distortion can be expressed as a percentage of the fundamental by $\sqrt{H_2{}^2 + H_3{}^2 + H_4{}^2}$ etc., where $H_2$, $H_3$, $H_4$ etc. are the percentage distortions of the individual harmonics.

Harmonics are present in nearly all sounds and account for the difference in tone of sounds of the same fundamental frequency. Their presence does not therefore explain the unpleasant nature of distortion which is due to inter-modulation distortion. Non-linearity causes intermodulation of simultaneously amplified signals, producing difference and summation frequencies which are not whole number multiples of the fundamental frequencies. Intermodulation distortion measurement is therefore a more realistic approach to an assessment of amplifier performance.

EMITTER FOLLOWER – A transistor version of the cathode follower with very similar properties.

EQUALISATION – The process of restoring a level frequency response, usually after de-emphasis or pre-emphasis (q.v.)

FADER – A device for reducing volume, i.e. a volume control or attenuator (q.v.). A centre-tapped volume control enables a fade to be made from one circuit to another.

FREQUENCY – The rate of vibration of an electrical or mechanical oscillation. Measured by the number of complete cycles executed in one second.

FREQUENCY RESPONSE – A graph of output voltage against frequency for an amplifier or attenuator. The horizontal frequency scale is usually logarithmic.

FUSE – A device for protecting apparatus in the event of the failure of components or faults in wiring. It is usual to provide a fuse in series with the mains transformer primary and in the a.c. supply to the h.t. rectifier. In the case of a full wave rectifier (Fig. 9.5, page 88), a fuse in the connection to the centre tap of the h.t. secondary winding will protect against a failure of the capacitors or the rectifier valve. A fuse connected in this way is shown in the amplifier circuit on page 101.

GAIN – The increase in signal strength from input to output terminals of an amplifier. Loosely synonymous with amplification.

HIGH FIDELITY – Reproduction of a signal without significant distortion by the apparatus employed – either from disc or tape or broadcast.

IMPEDANCE – The ohmic value of combined resistance and reactance. See vector Diagram.

LOAD LINE – A straight line drawn on $I_A V_A$ (or $I_C V_C$) characteristics representing the variation of anode (or collector) volts with valve (or transistor) current for a resistive load. Used to determine working point and signal handling capacity of the device.

MIXER – A circuit which enables the signals from two or more circuits to be combined, each signal being

248

continuously variable in the output. A common form of circuit employs two potentiometers, with resistors connected in series with the variable contacts, the ends of the resistors being joined to couple the circuits. Constant resistance attentuators can be arranged for mixing.

MONOPHONIC SOUND – Sound emerging from a single speaker and giving no impression of spatial distribution of the several sources of sound (musical instruments, voices, etc.) which comprise it.

MUTUAL CONDUCTANCE ($g_m$) – The change of anode current produced by a change of 1 volt on the grid of a valve. Unit mA/volt.

NYQUIST DIAGRAM – It has been shown (page 71) that in a feedback amplifier, the amplifier will oscillate at a frequency for which the phase shift and gain round the whole amplifier are such as to make $A\beta = 1$. The mid-frequency gain can be measured, and at this frequency the phase shift will normally be zero. At low and high frequencies gain will decrease and phase shift occur, and the value of $A\beta$ can be determined and the phase angle measured for a number of frequencies over a wide range. These results may be plotted with polar co-ordinates in the form of a Nyquist diagram. When $A\beta$ closely approaches 1 at any frequency, the amplifier is stable but is in a condition where any change in operating conditions might bring $A\beta$ to unity causing instability, e.g. during warming up or when driven beyond maximum output. The amplifier is then said to be 'conditionally stable'. An amplifier should be designed to be unconditionally stable. On the polar diagram, if the point 1,0. is enclosed by the curve, the amplifier is unstable.

OHM – The resistance of a column of mercury at 0° C, 14·4521 grammes in mass, of constant cross-sectional area of 1 sq. mm. and 106·300 cms. in length.

PARAPHASE – A circuit for providing two signals with a 180° phase difference to feed a push-pull stage. This term was originally applied to a circuit in which an amplifier valve feeds one output valve direct, and also a second amplifier valve. The second amplifier introduces 180° phase shift and a potential divider is arranged to exactly offset this amplifier stage gain, its output being fed to the second output valve. This term is now almost synonymous with phase inversion, and circuits employing negative feedback are preferred as distortion is reduced and a degree of self-balance can be obtained.

PENTODE – A five electrode valve. $k$, $g_1$, $g_2$, $g_3$, anode, developed from the triode to reduce input to output capacitance. This is achieved by the screen grid ($g_2$) which is earthed to a.c.

POTENTIOMETER – A network, the function of which is to provide an output at some level lower than that of the input. Composed of resistors, the attenuation is independent of frequency.

POWER AMPLIFIER – An amplifier of sufficient power to drive a loudspeaker.

PRE-AMPLIFIER – An amplifier to increase the small signals from a microphone, tape head, etc., sufficiently to be able to apply them to a power amplifier.

PRE-EMPHASIS – The boosting of a band of frequencies in accordance with a specified characteristic. It is convenient to arrange for this by means of a $CR$ network, and the pre-emphasis can then be specified as a time constant. To achieve a level response an inverse characteristic is employed in another part of the circuit, called de-emphasis. This arrangement is normally used to increase the signal-to-noise ratio of a system, where the noise mainly occurs over a part of the frequency spectrum. An example is the International L.P. recording characteristic, in which frequencies above 1,000 c/s are emphasised during

recording to improve the signal to 'record scratch' ratio. A de-emphasis circuit is then used in the reproducer.

REACTANCE – The 'resistance' of a coil or capacitor to alternating current. Given by $2\pi fL$ for a coil, $1/2\pi fC$ for a capacitor. The unit is the ohm.

RECTIFICATION – The process by which a d.c. voltage is derived from an a.c. supply. Achieved by use of a rectifier (diode valve or junction) and usually followed by a smoothing circuit.

REGULATED POWER SUPPLY – The rectified voltage of a power supply will fall as the current drawn is increased. The regulation can be improved by using a 'choke input filter' (see page 90) or a 'series' or 'shunt' valve regulator. A well designed valve regulated supply can be made to give a voltage which is virtually independent of current and this implies a power supply of very low internal impedance.

RESONANCE – The condition where current and applied voltage are in phase in a circuit including capacitance and inductance. Frequency at which this occurs given by $f = 1/2\ \pi\sqrt{LC}$. Series resonance gives minimum impedance which is equal to $R$ of circuit. Parallel resonance gives maximum impedance which is equal to $L/CR$.

SIGNAL-TO-NOISE RATIO – The ratio between the amplitude of the required signal and that of all other signals. The unwanted signals include hiss, hum, flicker and partition noise in pentodes. For comparison it is usual to state the ratio of the signal to the noise in decibels, and with a bandwidth of 50 to 10,000 c/s, a ratio of 40 dB is acceptable, 50 dB good, 60 dB excellent and 70 dB near perfection. As the noise will increase as the lower and upper frequency limits are extended, the bandwidth in c/s should be stated when a signal-to-noise ratio is specified.

SINE WAVE – The 'graph' of the voltage or current change in a circuit connected to an alternator. The instantaenous voltage is peak voltage × sin $\omega t$, where $\omega$ is the angular velocity, and $t$ the time.

SPECIFIC RESISTANCE $(\rho)$ – The resistance at opposite faces of a one centimetre cube of a substance. The resistance of a length $l$ of a conductor of constant cross-sectional area $A$ is then $R = \rho l/A$. At 20° C copper has a specific resistance of $1 \cdot 7 \times 10^{-12}$ ohms.

STEREOPHONIC SOUND (stereo) – Sound which appears to emanate from sources (violin, piano, etc.) which are spatially distributed. Achieved by the use of spaced speakers supplied from two or more separate microphones at the studio, or from multi-track tape or disk.

SQUARE WAVE – See Transient.

THERMIONIC EMISSION – The emission of electrons (thermions) from a solid surface by heat. In a thermionic valve this emission takes place from the heated cathode.

TRANSIENT – A transient is a complex waveform which does not repeat periodically as does a sine wave. It is characterised by sudden increases and decreases of amplitude, so that the performance of an amplifier on square wave test is of considerable interest. The rise and fall time of a square wave is theoretically zero and, in practice, square wave generators can be made with extremely short rise and fall times.

VECTOR DIAGRAM – When resistance (in ohms) is combined with reactance (in ohms) the resultant can be found by a diagram in the form of a right-angled triangle. The length of the sides adjacent to the right angle, represent the ohmic value of the resistance $(R)$ and reactance $(X)$. The resultant, called the impedance $(Z)$, is represented by the hypotenuse, and can be found from Pythagoras, so that $Z = \sqrt{R^2 + X^2}$. The resistance and reactance are 90° out of phase, and the relative phase angle of the resultant is shown by the diagram.

VOLT – The electrical pressure which will cause a current of one ampere to flow through a circuit resistance of one ohm.

# INDEX