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AMPLIFIER MANUAL No. 3

by

J. S. KENDALL
ASSOC. BRIT. I.R.E.

practical circuits for the

Amateur Constructor

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CONTENTS

Preface

Of the many branches of radio open to the amateur, audio amplification and the construction of low frequency amplifiers is perhaps the most popular. As recording technique has improved, and now that long playing records have become firmly established, the audio amplifier is receiving even closer attention than hitherto, with the result that much chaff has been separated from the wheat. Just as the old " General Purpose " valve has disappeared, so audio amplifiers have become classified; and to-day there are suitable designs for every purpose. For many years Bernards (Publishers) Ltd. have endeavoured to present the amateur with the best designs in this branch of the radio field, and in presenting this manual by J. S. Kendall they are confident that the exacting standards of design demanded by the modern constructor have been maintained.

In general the subject has been broadly treated to provide the home constructor with sufficient data to design his own equipment; in part, or as a whole. Complete designs are also given, suitable for P.A. work, gramophone reproduction in the home and high-quality reproduction for the serious gramophile.

Of particular interest is the 5 valve 10 watt amplifier by Mullard Ltd. This is a new design developed by this famous Company and released for the benefit of the home-constructor. Full details of this remarkable design together with others, each with an outstanding performance are contained in this manual.

Walter J. May,

Technical Editor.

CHAPTER I

THE POWER SUPPLY

No matter what type of amplifier is being constructed; the first consideration is the type of power supplies, is it to be A.C., A.C./D.C., or run off batteries.

An A.C. supply mains has to be converted to the correct voltage, rectified, and finally all the mains hum and noise must be removed. The main advantages of A.C. mains are that the amplifier chassis does not have to be directly joined to the mains as with D.C. or A.C./D.C. techniques. Again the voltages and power for the heaters can be obtained as required without any trouble. It is this type of supply that will be first considered.

Battery power units are of two main types, one is the rotary converter, which is the better of the two, and at the time of writing these units are readily available on the surplus market. The other is the vibrator pack, this however is a little difficult to design as the unit if not correctly screened and filtered causes considerable noise. Chokes have to be placed in all the heater leads to stop the R.F. generated by the sparking of the vibrator from getting into the amplifier. It is better to purchase these units ready made rather than try to construct them.

A.C. Supplies

The A.C. mains power pack is built around the mains transformer, this is usually purchased ready made although they can be constructed. The requirements of the transformer are, the supply of the H.T. voltage, and two or more heater voltages. For example an amplifier requires 250 volts at 60 m.A. for H.T., with heaters for the amplifier valves requiring 6.3 volts at 2 amps. and the rectifier perhaps 5 volts at 2 amps. On looking at a catalogue, one finds perhaps a transformer listed as providing 250-0-250v. at 80 m.A., 6.3 volts at 3 amps and 5 volts at 2 amps., this would be quite suitable for the example quoted.

Many constructors are under the impression that the transformer will put out the current at which it is rated irrespective of the circuit requirements; this is not true, the current drawn is dependent on the load imposed by the amplifier on the transformer. Under no circumstances may a transformer be used that has a lower rating than the requirements of the circuit with which it is to be used. When choosing a transformer, a small safety margin must be allowed in case the smoothing condensers deteriorate and pass a small amount of current.

The choice of the rectifier valve requires careful consideration. The writer has on occasion seen low power circuits used with unnecessarily large rectifiers. For example, a receiver with a H.T. of 250 volts at a current of 80 m.A. used a GZ32 rectifier, this valve is designed for an output current of 250 m.A., over three times the current required for the receiver. The valve that should have been chosen here was the GZ30, this provides a maximum current of 125 m.A. or just over half as much again as the radio required. Heater construction should be considered; should one use directly or indirectly heated valves? The former consume less power at the heater than the latter, but the latter are of lower impedance and require the use of surge limiting resistors in the anode circuits. Again directly heated valves require only a few seconds to warm up whilst the indirectly heated types take perhaps thirty seconds. If the other valves in the equipment are indirectly heated, an indirectly heated rectifier should be used, this prevents the voltage at the cathode of the rectifier building up to a dangerous level. If, however, the output valves are of the PX4, ACO44 types, which are directly heated, a directly heated rectifier will be quite in order. Remember the voltage at the cathode of the rectifier valve with no load being drawn will rise almost to half as much again as the rated voltage of the transformer H.T. secondary.

There is very little in the design of the power unit that is otherwise than conventional. The circuit shown in Fig. 1 is typical. The choice of valve and transformer have been discussed so that only the capacitors and choke remain. L.F. chokes should be of ample current rating and as the commercial chokes are made in fewer ranges than capacitors, it is best to calculate the minimum value of capacitance required. This minimum value is only a rough guide, and can often be increased with advantage. With 50 cycle mains and full wave rectification, the capacity of the two capacitors added together and then multiplied by the choke inductance in Henrys should be at least 200. For example, a 10 Henry choke is available so that a factor of 20 is the condenser capacity required. This can be raised to 24 mfd. divided into two parts, the reservoir capacitor at the cathode of the rectifier being 8 mfd. and the smoothing capacitor 16 mfd. This use of condensers of differing capacities is quite common and for some applications a reservoir capacitor is not used at all. If the capacity of the reservoir capacitor is too high, surge limiting resistors must be included in the anode circuits of the rectifier. In many cases the D.C. resistance of the H.T. winding of the mains transformer is quite high enough to provide this, and extra resistors are not required.

Often it is found that the H.T. voltage is too high even though the transformer of the correct rating has been used. This is because the output voltage at the cathode of the rectifier is almost 1.4 times the R.M.S. value of the transformer H.T. secondary. There is of course the drop across the valve to be considered also the voltage drop across the choke. A transformer rated at 250 volts will, after there has been a voltage drop of 50 volts across the valve and perhaps another of 10 volts across the choke, still deliver about 280 volts rectified (1.4 x 250 is 350, allow the 60 volts lost in the circuit and 290 volts remain). It must be remembered that a transformer secondary rated at 250 volts actually develops 350v. This is because the rating quotes the R.M.S. (root mean square) voltage whereas under working conditions the peak voltage (R.M.S. x 1.4) must be considered. Assuming the amplifier requires only 250 volts, how then are we to get rid of the 40 volts excess? The remedy is to insert a suitable resistor in the circuit having calculated the value of resistance that will, at the current drawn by the amplifier, drop 40 volts. Assume the amplifier requires 80 m.A., by a simple application of Ohms Law, the value of resistance can be derived. This law states that the current drawn by a circuit is directly proportional to the applied voltage and inversely proportional to the resistance of the circuit.

i.e. $$E = I \times R$$
$$I = \frac{E}{R}$$
$$R = \frac{E}{I}$$

where E=voltage I=current in amps and R=resistance in ohms.

This formula should be remembered as it is the most used in electrical and electronic work.

In this case R equals E/I, substituting here E is 40, I is .08 amps (notice here the current has been converted from m.A. to the fundamental unit of the ampere) R= 40/.08 or 500 ohms.

Not any 500 ohm resistor can be used here as considerable power will be dissipated. The power rating can easily be found by multiplying the current in amps by the voltage drop. As the voltage is 40 and the current .08 amperes the wattage will be 3.2 Watts the use of a 5 Watt resistor here would be the best choice.

A typical circuit is shown in Fig. 2. Note that an extra capacitor has been included in the circuit in order that common impedance coupling be kept to a minimum.

A.C./D.C. Supplies

A.C./D.C. supplies sometimes misnamed "Universal," employ a different technique, one side of the supply is connected to the chassis of the apparatus, half wave rectification is usual, and valve heaters are always wired in series instead of in parallel as with A.C. equipment. Where the heaters are wired in parallel, the voltage is important, and each valve must have the same voltage rating, the current consumption is of secondary importance. With series wiring, the current consumed is what matters and the voltage rating is of secondary importance. For instance, a domestic receiver designed for A.C. operation may use the following valves ECH35 (6.3v 0.3A), EF39 (6.3v 0.2A), EBC33 (6.3v 0.2A), EL33 (6.3v 0.9A), and EZ35 (0.6A). The total requirements for the heaters is 6.3v 2.2A. A similar design for use on A.C./D.C. mains would use CCH35 (7.0v 0.2A), EF 30 (6.3v. 0.2A), EBC33 (6.3v. 0.2A), CL33 (33.0v 0.2A) and CY31 (20.0v 0.2A). Power required for this heater chain is 72.6v 0.2A. It will be observed that the A.C. equipment consumes nearly 14 watts of power (6.3 × 2.2A) whereas the A.C./D.C. system uses just over 14.5 watts (72.6 x 0.2). In practice the A.C./D.C. chain is by far the more costly to run because the difference between the supply voltage, and that required by the valve heater chain, is dropped through a large resistor and dissipated in the form of heat. Thus for the A.C./D.C. chain mentioned on a 230v supply, the power used will be 230 x 0.2=46 watts.

Fig. 3 illustrates the wiring of a series heater chain complete with dropper. Note that two resistors are shown on the circuit and that one has a black dot on it. This indicates a special resistor known as a Thermistor. When the valve heaters are cold their resistance is practically nil, so that a heavy current surge occurs, this is injurious to the valves and reduces their useful life considerably. The characteristic of the thermistor is the reverse of a valve heater, when cold, it has a very high resistance which falls as it heats up.

When calculating the value of a dropper resistor for use with a series heater chain, the resistance of this component may be ignored. It will be appreciated that by including a thermistor in the heater circuit, the injurious surge is avoided and the life of the valves prolonged. Thermistors are manufactured by several companies, one of the most popular is that made by Standard Telephones & Cables and marketed under the name BRIMISTOR.

There are several different A.C./D.C. valve series available at the present time, designed for the following current ratings, 0.3A, 0.2A, 0.15A and 0.1A. American manufacturers are mainly concerned with 0.3A and 0.15A types but British and European manufacturers concentrate on 0.2A and 0.1A valves. Thermistors are made in a number of types to suit any of the popular valve series.

The value of the dropper resistor is found by applying Ohms Law. Subtract the combined heater voltages of the valves used from the mains supply voltage. Divide the resultant by the heater current in amps of the valve series chosen. The dividend is the required value in ohms. Assume the valve sequence referred to earlier is being used (CCH35, EF39, EBC33, CL33, and CY31) on 240v. mains. Heater voltages total 72.6, say 72, therefore the voltage to be dropped equals 168 and the value of the dropper resistor equals

$$\frac{168}{0.2} = 840 \text{ ohms.}$$

A very useful alternative is to calculate the voltage to be dropped in the usual way, and then for

0.3 Amp. valve series $\times 10$ and $\div 3$
0.2 Amp. " " $\times 5$
0.15 Amp. " " $\times 20$ and $\div 3$
0.1 Amp. " " $\times 10$

Dropper resistors when purchased, usually have a higher resistance than will be required and are provided with an adjustable clip so that they can be set to suit all conditions. A 0.2A type as would be required in the case of the example

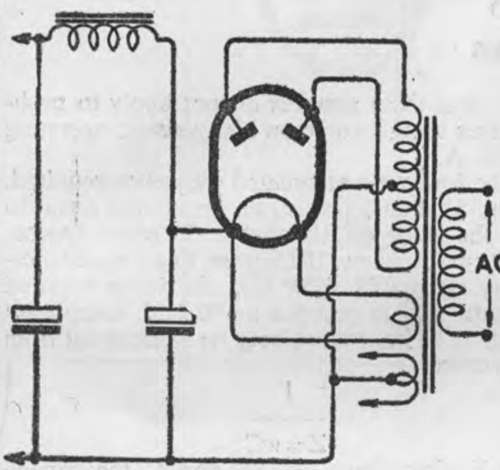


FIG. 1.

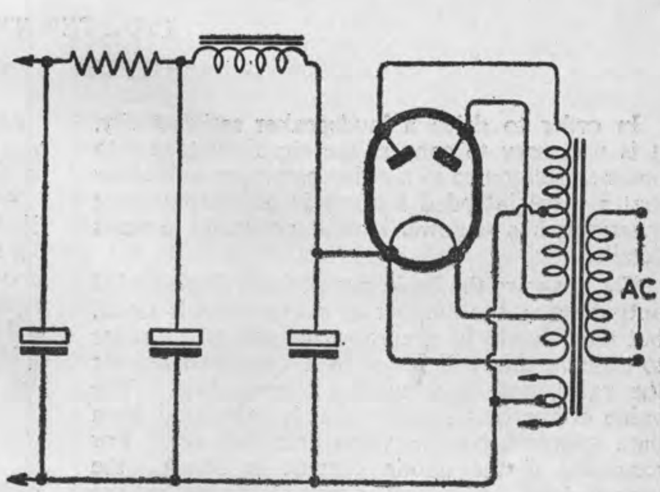


FIG. 2.

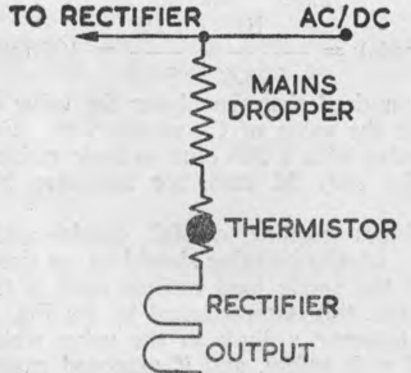


FIG. 3.

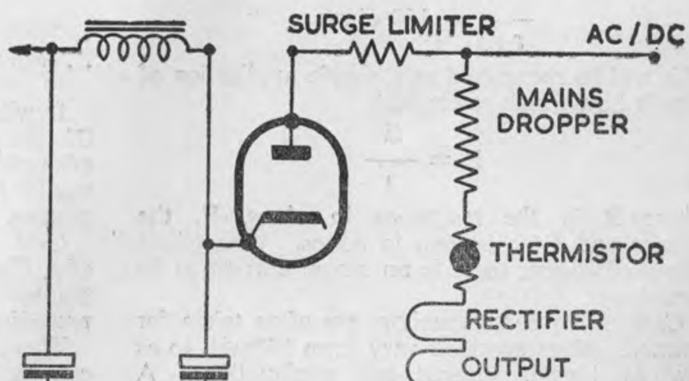


FIG. 4.

quoted has a nominal value of 1,000 ohms. The dropper should be set by means of an ohm-meter and then under working conditions can be finally adjusted with an ammeter inserted in series with the heaters.

Having dealt with the heater supplies, there is still the question of H.T. This is obtained directly from the mains, as shown in Fig. 4. It will be seen that there is a resistor between the mains and the rectifier valve anode. This is what is known as a surge limiter and has a value of some 50 to 250 ohms, the value depends on the size of the reservoir capacitor and the valve used and is quoted in the valve maker's tables. The value in micro-farads of the smoothing capacitor should be double that used with full-wave circuits. When the values of the reservoir and smoothing capacitors are added together and multiplied by the choke inductance in henrys, the result should be at least 400. Voltage drop is very important because the output voltage is always less than that of the mains input and every volt counts. For this reason the modern trend is to use low resistance and low inductance smoothing chokes, and to use very high value capacitors. A few years ago 16 mfd. for both reservoir and smoothing was normal practice but with modern amplifiers, capacitors of 64 mfd. or even 100 mfd. are quite common. An advantage is, that large capacitors are much cheaper than high inductance chokes.

CHAPTER TWO

OUTPUT STAGES

In order to drive a loudspeaker satisfactorily, it is necessary to convert the signal voltage into power. Reference to a valve catalogue will show that a valve intended for use in an output stage invariably has its power handling capacity in watts detailed.

Fig. 5 shows the basic circuit for a single-ended output stage, the number of components is small, but they should be accurately calculated. Failure to observe this will result in a shortened life for the valve, and poor quality reproduction. The value of the bias resistor (Rk) is calculated from data supplied by the valve manufacturer. For example, if the anode current is 36mA., the screen 4mA., the total current flowing in the cathode circuit will be 40mA. If the grid bias required is –6 volts, the value of the resistor will be

$$\frac{6}{0.04} = 150 \text{ ohms}$$

This will be recognised as a simple application of Ohm's Law which states that

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

where R is the resistance in ohms, E. the voltage and I the current in Amps. With triode valves of course, there is no screen current to be considered.

Cathode by-pass capacitors are often taken for granted, values specified vary from 100mfd. to as low as 10mfd. without any explanation. A capacitor is required in this position because signal voltage is present in the cathode circuit just as it is in the anode and if permitted to develop across Rk, current negative feedback occurs which causes loss of gain and adds to the overall impedance of the output circuit, this in itself is highly undesirable.

Note that these remarks do not apply to push-pull stages using a common bias resistor, operating in Class A.

At the lowest un-attenuated frequency required, Ck (Fig. 5) should present an impedance equal to 0.2 of the value of Rk, that is 30 ohms approx. since Rk=150 ohms. Reference to an impedance-capacity chart will show that the value required is 106mfd. and in practice a 100 mfd. component is ideal. The impedance may be worked out from the formula

$$\frac{1}{Z = wC}$$

where Z=impedance $w = 2\pi$ and C the capacitance in farads.

$$\text{i.e. } Z = \frac{1}{wc} = \frac{1}{2\pi fc} = \frac{1}{6.28 \text{ fc}}$$

$$\therefore \text{(Microfarads)} = \frac{10^6}{6.28Zf} = \frac{10^6}{9420} = 106 \text{ mfd.}$$

It will be noticed that the higher the value of Rk the lower the value of Ck needs to be. For example a valve with a 500 ohm cathode resistor would require only 32 mfd. for adequate by-passing.

Grid resistors require careful consideration (Rg, Fig. 5). Ideally its value should be ten times the value of the anode load resistor used in the preceding stage, this is represented by Ra Fig. 6.

There is however a limit to the valve which can be used with safety, and if exceeded causes grid current to flow, this increases to anode current beyond its rating and quickly ruins the valve. According to Prof. F. E. Termon the American authority, grid resistance when cathode bias is employed, should not exceed $\frac{10}{gm^2}$ MΩ where

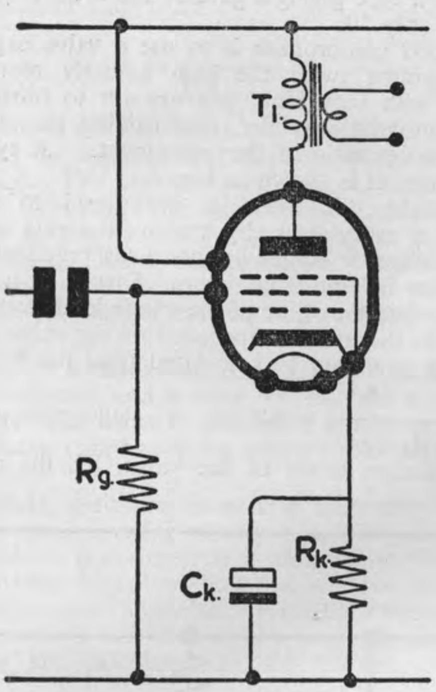


FIG. 5.

gm= the mutual conductance of the valve expressed in milli-amps, per volt. Consider the EL 41, this valve has a gm or "slope" of 10 mA/v.

Applying the formulae $Rg = \frac{10}{100}$ MΩ = 100kΩ

With modern British Valves this limit can be, and often is exceeded by as much as 100% but. the author prefers to regard Prof. Termon's recommendation as the safety limit.

Valves with a lower slope permit the use of a higher grid resistance, type 6F6 is typical, it has a gb. of 2.5 m.A/v therefore $Rg = \frac{10}{2.5^2}$ MΩ = $\frac{10}{6.25}$ MΩ = 1.6 MΩ

There now remains only the calculation of the output transformer T.1 (Fig. 5) to complete the output stage. This component transforms the signal voltages appearing at the anode into power, and suitably matches them to the load presented by the speaker voice coil. This latter item is an impedance which for calculation purposes may be regarded as a resistance. Now the power at the anode is the voltage (Ea) squared, divided by the anode load resistance (Ra). Power developed by the speaker voice coil, equals applied voltage (Es) divided the impedance of the voice coil (Rs). We are therefore concerned with the ratio of the squares of two voltages i.e. $\frac{Ea^2}{Ra} = \frac{Es^2}{Rs}$ which when transposed gives $\frac{Rs}{Ra} = \frac{Ea^2}{Es^2}$ so that the ratio of T.1 is equal to the square root of

$$\frac{Ea}{Es} \text{ or } \sqrt{\frac{Ea}{Es}}$$

Example: for an anode load of 4,000 ohms and a voice coil impedance of 10 ohms, T.1 ratio $= \sqrt{\frac{4000}{10}} = \sqrt{400} = 20:1$

CHAPTER THREE

VOLTAGE AMPLIFIERS

Whilst the output stage provides power to feed the loud-speaker this stage in turn must be provided with a fairly high input voltage to enable it to develop its maximum power. Certain modern output valves require only 4 or 5 volts at the grid for full output, but even this is far in excess of the output of modern gramophone pick-ups. To quote an example, the Philips high-fidelity head provides an output of 0.7 volts and there are others whose output is considerably less. From this it will be understood that it is invariably necessary to provide an extra stage or even two stages of amplification between the pick-up and output stage. Such a stage is a voltage amplifier and is often referred to as a driver stage.

Triodes are very suitable for use as voltage amplifiers, there are many types on the market, and one should be chosen with a suitable amplification factor. Obviously it is useless to try and obtain a gain of 40 from a valve designed to provide a gain of 20, equally so it is wasteful to employ a type giving a gain of 100 if the required gain is only 20.

A good compromise is to use a valve capable of providing twice the gain actually required. Stages with very high gain are apt to introduce noise into the amplifier, thus ruining the overall signal noise ratio of the equipment. A typical triode circuit is shown in Fig. 6.

Inevitably, distortion is introduced to some extent in every stage, by keeping the gain of the driver stage or stages as low as is practical, the distortion is kept at minimum. Distortion will be lowest when the value of the anode load resistor is equal to the anode impedance of the valve used.

Stage gain may be calculated from the formula

$$\propto = \frac{\mu R_L}{R_L + r_a}$$

where $\propto$ = stage gain μ the amplification factor of the valve, R_L the anode

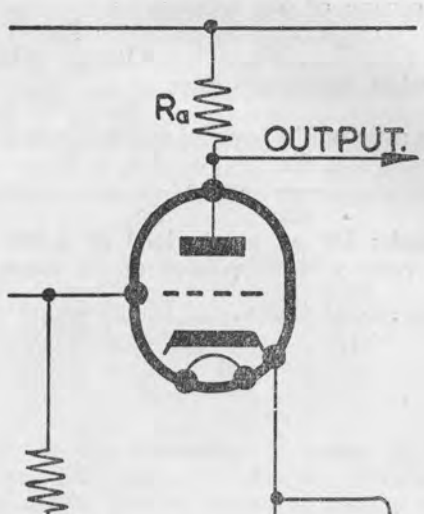


FIG. 6.

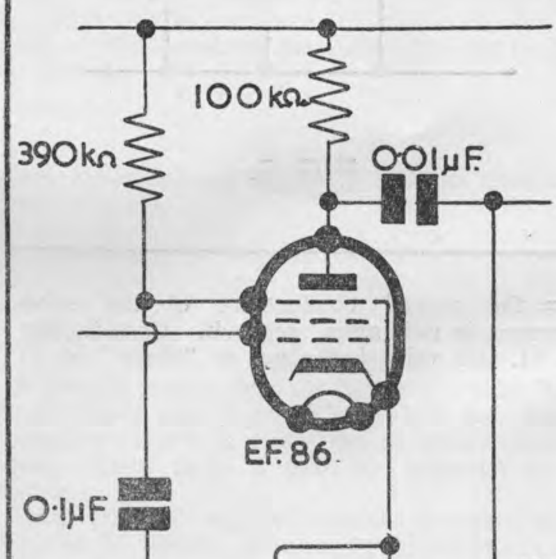


FIG. 7.

load resistor in ohms and ra the anode impedance of the valve. This formula is useful for calculating the gain for a given value of R_L, but in practice it is better to use the following since it permits the value of R_L to be calculated for any particular value of stage gain

$$R_L = ra \left(\frac{\alpha}{\mu - \alpha} \right)$$

Coupling is usually effected by means of a capacitor and the impedance of this component must be watched. For efficient performance, this should be less than 0.2 of the associated grid resistor, this at the lowest frequency for which the amplifier is designed. From this it will be observed that the **larger** the permissable value of the grid resistor, the **smaller** the required coupling capacitance.

It is not a matter of using a high capacity for good measure and hoping for the best however, such practice leads to instability due to common impedance coupling in the anode circuit and other evils.

Modern equipment is making increasing use of R.F. pentodes, as L.F. voltage amplifiers, although the circuits are simple it is unusual to find any explanation given on how the various component values are calculated. With the exception of the screen resistor and de-coupling capacitor, the values are calculated as for triodes. Screen circuit values are calculated as follows: Calculate the anode resistor, note the anode and screen current required by the valve and observe the ratio, i.e. a valve with a consumption of 10 mA's anode and 2 mA's screen has a ratio of 1 : 5. Now the value of the screen resistor = Anode resistor × ratio × 2. For an anode load of 100 kΩ the value of the screen resistor = 100,000 × 5 × 2 = 500,000 × 2 = 1MΩ

There are special circuits where the screen operates at a higher potential than the anode, but for general application the screen is operated at about 0.5 of the anode potential. A little thought will show that the above formulæ provides for this. Working under these conditions the valve impedance and amplification factor remains constant and distortion is avoided.

As a point of interest, output pentodes usually operate with a slightly higher screen potential than that of the anode, this of course is a matter of design.

Heptodes are sometimes used in audio circuits, some of these have a higher screen consumption than that demanded by the anode, however, calculations of component values is carried out as with R.F. pentodes.

A typical voltage amplifier circuit using an R.F. pentode type EF86 is shown in Fig. 7, close tolerance resistors, preferably with a high stability characteristic should be used to minimise risk of the resistance value changing with age.

CHAPTER FOUR
FEEDBACK

Low frequency amplifiers are apt to appear deceptively simple when layed out in theoretical form.

Basically the L.F. amplifier is simple, but it is choosing the correct component values that taxes the ability of the designer. Insufficient decoupling or poor layout will cause instability. An unstable amplifier will give low output, distortion and be generally unsatisfactory in every way.

Examine Fig. 8. This is a simple circuit consisting of an RF pentode voltage amplifier resistance-capacity coupled to an output pentode. Apparently it has only one input, and to all appearances it should be stable. This is not so: g2 is fed to the H.T. rail by a resistor and decoupled by a 0.1 μF capacitor. At low frequencies the impedance of this capacitor is quite high and will not have any effect on the circuit, with the result that small variations of H.T. rail voltage caused by the operation of the output valve, are reflected to g2 of the voltage amplifier, causing this valve to draw increased H.T. current. This increased current causes a voltage drop through the anode load and arrives at the control grid of the output valve, by means of the coupling capacitor, as a negative charge.

Current drawn by the output valve falls, allowing the H.T. rail voltage to rise, thus the original fluctuation is returned to g2 of the voltage amplifier in aggravated form. Eventually the negative charge arriving at the output stage will be of sufficient magnitude to cause the valve to "cut off" that is, current will cease to flow. At this moment the circuit momentarily stablises. Once the negative charge at the grid commences to decay, current again flows, the H.T. rail voltage falls, causing reduced current to be drawn by the voltage amplifier. This deposits a positive charge at the grid of the output valve, which builds up until grid current is drawn, after which the whole cycle of events starts again. The frequency at which this series of events or oscillation occurs may be too low to be clearly audible, but the quality of the reproduction and the erratic performance of the equipment will leave no doubt that something is seriously wrong. Sometimes high-frequency oscillation is encountered, this gives excessive distortion or even a continuous roar. Oscillation is caused by positive feedback and is best overcome or avoided by the judicious use of de-coupling. High capacity smoothing capacitors are a great help in avoiding this form

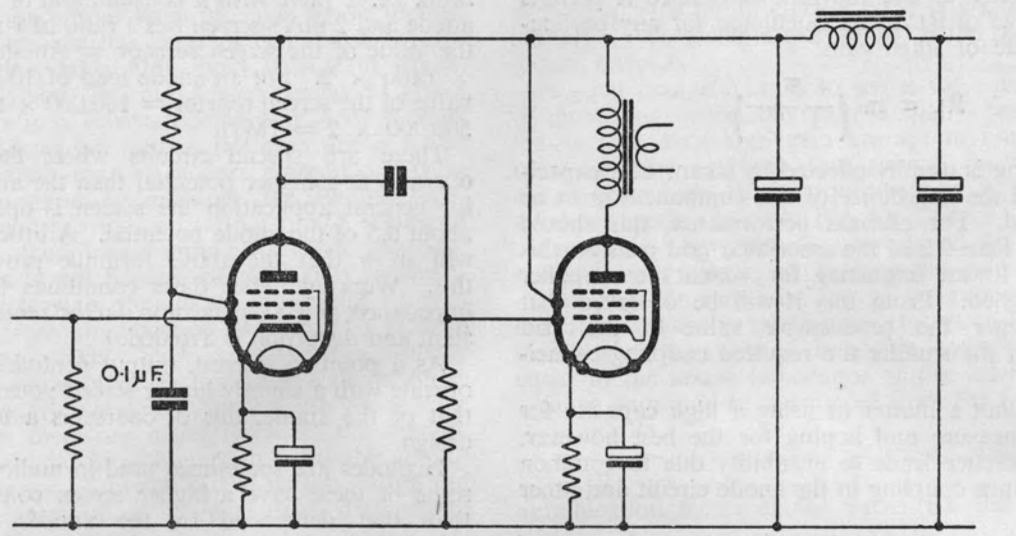


FIG. 8.

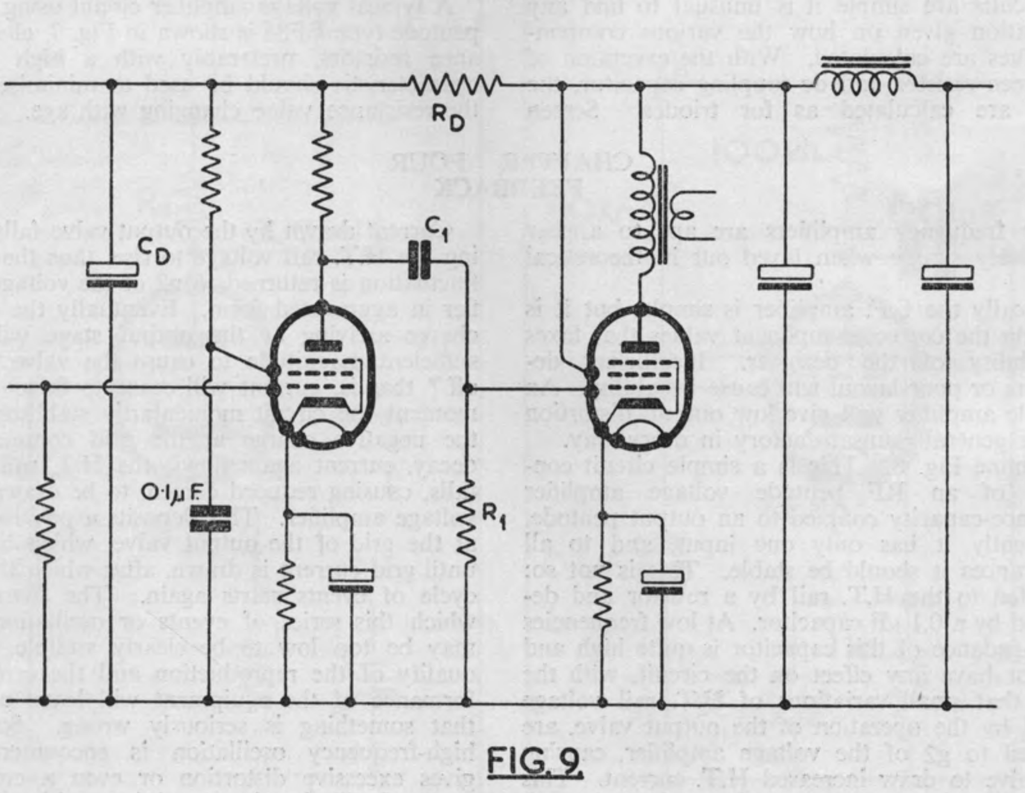


FIG. 9

of trouble but the use of a resistance-capacitor network is by far the most satisfactory solution. Compare Fig. 9 with that of Fig. 8. With the exception of Rd and Cd they are identical, but whereas the circuit of Fig. 8 is liable to oscillate continuously that of Fig. 9 is perfectly stable.

Values for Rd and Cd depend on the amount of feedback present, an average value is 22 kΩ for Rd and 2 μF for Cd.

The efficiency of a given de-coupling network can be roughly calculated by comparing the time constant of the de-coupling network with that of the coupling components. An R.C. network constant is found by multiplying the capacitor value in micro-farads by the resistor value in megohms. Refer to Fig. 9:

if C1 = 0.01 μF and R1 = 470 kΩ, constant = 0.0047, again if Cd = 8, μF and Rd=22 kΩ, constant=0.176.

therefore ratio=4.7 : 176 or 1: 37.

Note that Rd is a resistance, a choke should not be used because at very low frequencies its impedance would probably not exceed the D.C. resistance value, that is 200 to 500Ω.

So far, the feedback discussed has been of a positive nature, there is, however, another form which can be highly beneficial to L.F. amplifiers. Negative-feedback as its name implies produces the reverse effect to positive feedback. By applying it intelligently, stability is increased and overall distortion reduced.

Degeneration or negative-feedback is applied by feeding a small amount of the output back to the input in such a manner that the gain is reduced (just as regeneration or positive feedback is applied to T.R.F. receivers to increase the gain).

At first view this might appear a pointless procedure, after all, gain can be reduced by using lower gain valves. It is the reduction in overall distortion that makes the presence of negative-feedback so worthwhile.

Every amplifier generates a certain amount of distortion: when negative feed-back is applied, the distortion present in the output is fed back in such a polarity as to produce amplified distortion voltages which tend to cancel out the distortion generated within the amplifier.

A.C. Cossor Ltd. were one of the first Companies in England to appreciate its advantages.

A modern version of the circuit used by this Company in the early 1930's is shown in Fig. 10. Observe that the suppressor grid of the driver valve is connected to the cathode of the output pentode from which the usual by-pass capacitor is omitted.

A very useful amplifier for general application can be constructed from this circuit. the EF36 is specially designed for L.F. work, it uses a special electrode assembly technique ensuring that hum and microphony are negligible. As a voltage amplifier this valve is capable of very high gain, and when used with an EL41, a remarkably high slope output valve, adequate power is available from very small inputs such as provided by modern gram-pickups. Though an excellent design this circuit has one drawback, no account is taken of any distortion due to the output transformer since the feedback loop does not include this component.

Distortion can be introduced in one of two ways by this component, iron saturation and amplitude distortion caused by falling impedance at low frequencies.

Iron saturation is due either to bad design or because the transformer cannot handle the current demanded by the output valve, it has the effect of distorting the wave-forms by clipping the peaks and introducing harmonics. When purchasing a transformer, ensure that the output valve current does not exceed the maximum **Standing** current at which the component is rated. Be careful to differentiate between Max. D.C. and Max. Standing D.C. The former rating is commonly used on cheap components, it does not refer to the standing current as drawn by the output valve, but to peak conditions, thus a transformer rated at Max. D.C. 50mA cannot with safety be used with a valve drawing more than 25mA because under working conditions output valves momentarily draw double their rated standing current during each cycle at full drive. With reputable components rated at a certain max. standing current, constructors can rest assured that the peak conditions have been allowed for in the design.

Falling impedance at low frequencies, which gives rise to amplitude distortion is entirely a matter of design. For purposes of low frequency response calculation, the transformer can be regarded as an inductance shunted by the load impedance of the output valve. When the inductive impedance of the transformer is twice the load impedance there is a 2 : 1 (3db) power loss, and when the impedances are equal there is a 4 : 1 (6db) power loss.

In the latter case it might be considered that with the impedances equal, half the power is developed in each and that the power loss can only be 3db. Remember, however, that they are in parallel and that the voltage drop across the two will be halved and the current shared.

Medium priced transformers usually have a falling response at the rate of 3db per octave below 1,000 cycles. High grade components maintain their response down to 150 cycles or even lower and then fall off at 3db per octave. Falling bass response can be compensated for, by including the secondary of the output transformer in the feed-back network. As the gain tends to fall due to a falling off in the transformer response, the amount of feed-back is reduced which automatically increases the gain. At the same time any unwanted resonant peaks which

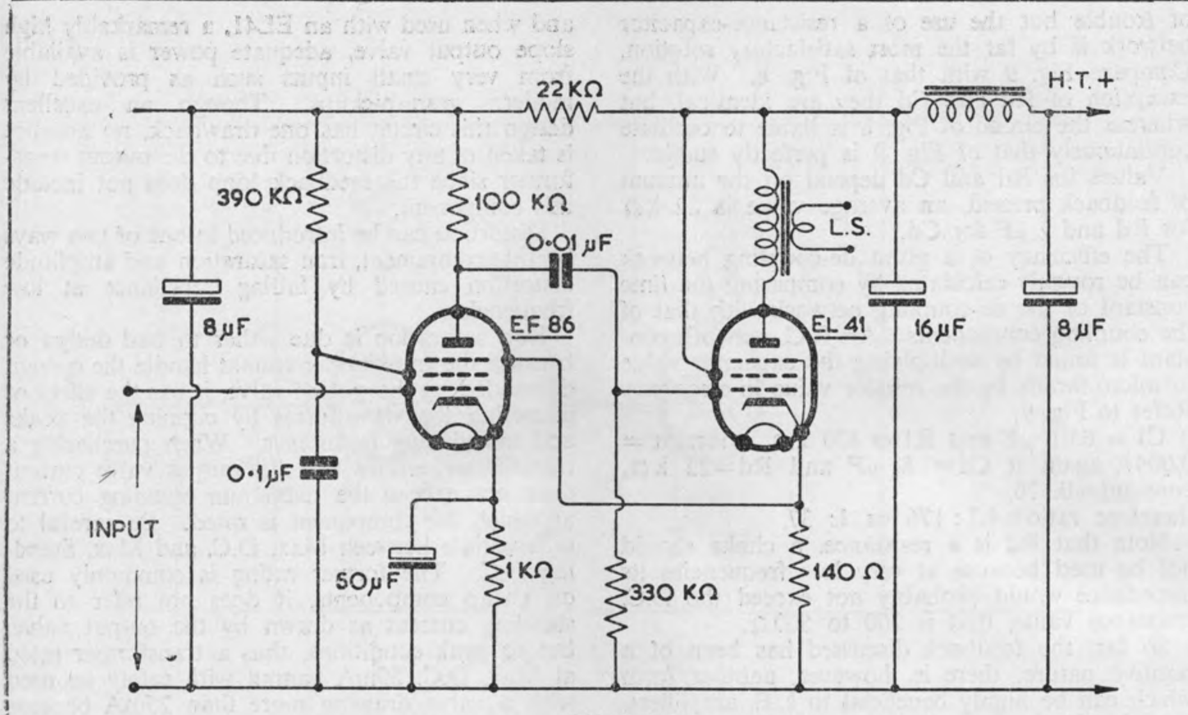


FIG. 10.

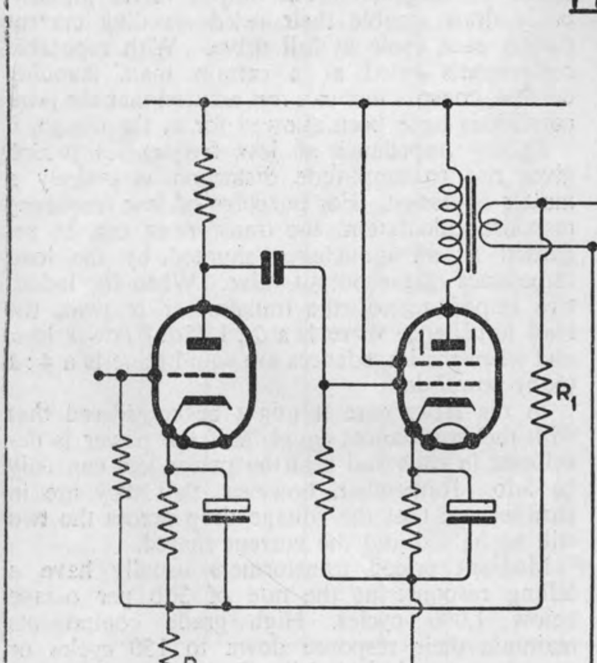


FIG. 11.

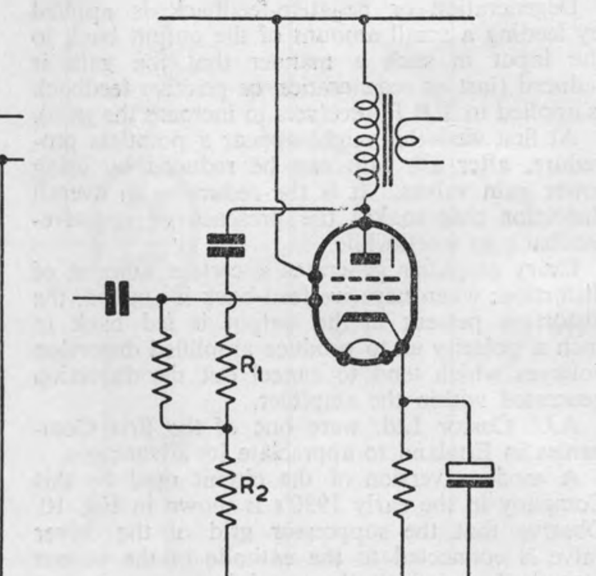


FIG. 12.

may be present are smoothed out.

It should not be thought that negative feed-back is a cure for badly designed transformers, quite the reverse, with such components phase-shift is almost certain to occur with the result that positive feed-back is introduced and the amplifier will oscillate uncontrollably. Fig. 11 shows a basic circuit of a two-stage amplifier with feed-back. R1 and R2 are the feed-back resistors, it will be noticed that they are in series across the output of the amplifier thus forming a voltage divider. The degree of feed-back is given by multiplying the ratio of R1 to R2, by the ratio of the output transformer.

It is unwise to introduce the feed-back over more than two stages because the small amount of phase-shift between stages is sufficient to cause positive feed-back and of course, oscillation.

A third method of introducing feed-back is shown in Fig. 12, it consists of an R.C. network from anode to grid. The frequency at which the feed-back becomes operative is dependant on the amount of capacitance included, it is a useful circuit for providing bass-boost to compensate for the recording characteristic of gramophone records. Ideally the combined values of R1 and R2 should be some ten times the optimum load of output valve, but care must be taken to ensure that the combined values of the grid resistor and R2 do not exceed the maximum value permitted for the output valve used. One final point, when applying feed-back it is necessary to experiment with the connections to the secondary of the output transformer, since in one direction the feed-back is positive and the amplifier will oscillate, to cure this the feed-back connections are reversed on the output transformer secondary.

CHAPTER FIVE

PHASE-SPLITTING

So far, all the circuits discussed have been designed for single-ended output. These are certainly very useful in many applications, but they have one serious drawback from a quality point of view.

Irrespective of how carefully the equipment has been designed, and regardless of precautions taken, some distortion is bound to be present. Now single-ended stages generate all orders of harmonics, that is the 2nd, 3rd, etc., of which the 2nd harmonic is the dominant. Fortunately push-pull amplifiers cancel the even harmonics (2nd, 4th, 6th, etc.), thereby reducing the overall distortion content. This low harmonic distortion content is one reason why push-pull amplifiers sound superior to the average person. Again, by eliminating the even harmonic distortion content, a higher output per valve is possible than with single-ended stages.

To operate two valves in push-pull, it is necessary to supply a signal to the control grid of each valve. Not only must the signals be of equal intensity but they must be in opposite phase; that is, one is 180° out of phase with the other.

Until the mid 1930's transformers were the only method used to accomplish this.

Fig. 13 illustrates the method. Note that the transformer secondary is centre-tapped. It will be observed that the signal at each control grid is equal and opposite since by a natural law, voltages developed in a winding must be of opposite polarity at each end.

With the circuit shown the voltage amplifier provides a gain of about 16, the transformer is parallel fed because by isolating the primary from the D.C., risk of distortion due to iron saturation is reduced. Another reason for using parallel feed is that like output transformers, inter-valve transformers suffer from a falling frequency response at low frequencies. The combined effect of the coupling capacitor and transformer primary inductance provides a bass boost at or near the resonant frequency of this network, though below this frequency the response falls rapidly.

Split secondary transformers are a later development, they enable negative feed-back to be provided over the output stage. Examine Fig. 14 this amplifier is capable of delivering some 15 watts with a low distortion factor. With the circuit constants specified feed-back amounts to 10% of the output. Transformers are hardly ever used in modern equipment, because the cost of a high grade component is considerable and at best they are not as good as a properly designed electronic phase-splitter. One of the first, and probably the most popular electronic phase-splitter is the "Concertina" circuit. This consists essentially of a triode valve with equal anode and cathode loads. From Fig. 15 it will be seen that bias is derived in the normal way from a cathode resistor suitably by-passed. The grid resistor is terminated at the junction of the cathode bias resistor and the cathode load resistor. Because of the heavy negative feed-back due to the degenerative effect of the cathode load resistance, distortion is negligible.

Load resistors L1 and L2 are equal in value and may be up to ten times the anode impedance of the valve used. R.F. pentodes can be used in this circuit if strapped as triodes. Both coupling capacitors, and grid load resistors following the phase-splitter, are of equal value. An ingenious

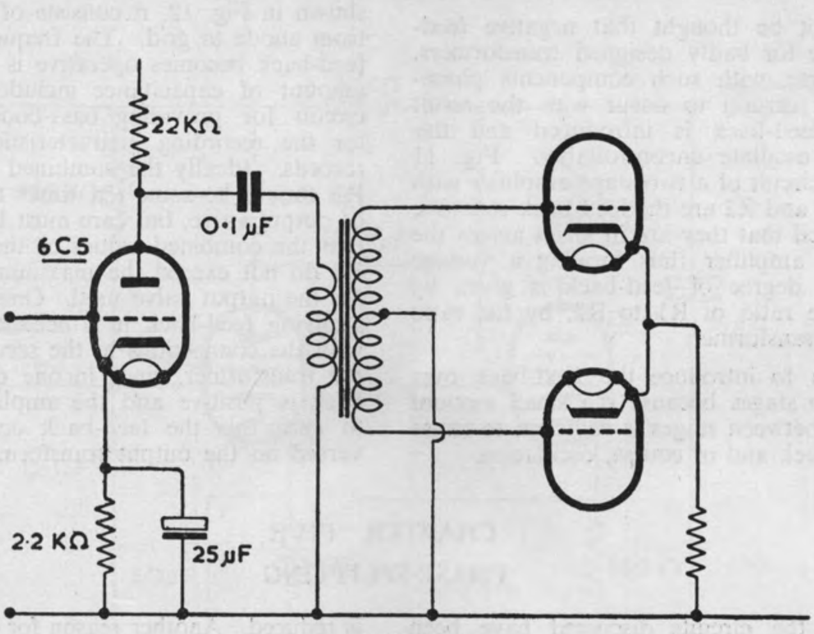

22KΩ
0·1 μF
6C5
2·2 KΩ
25μF

FIG. 13.

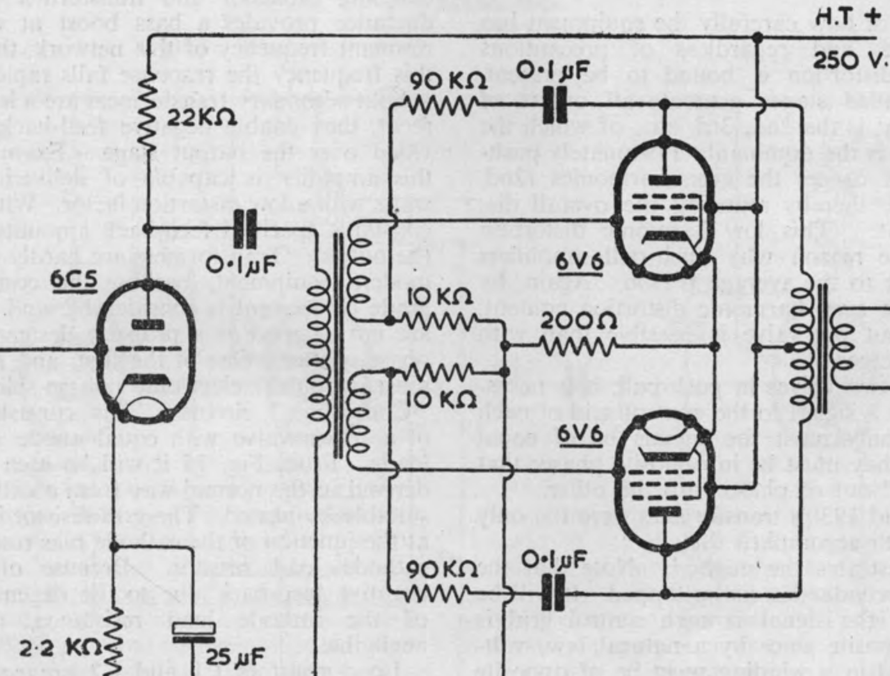

H.T +
250 V.
90 KΩ
0·1 μF
22KΩ
0·1 μF
6C5
10 KΩ
200Ω
10 KΩ
L.S.
6V6
6V6
90 KΩ
0·1 μF
2·2 KΩ
25μF

FIG. 14.

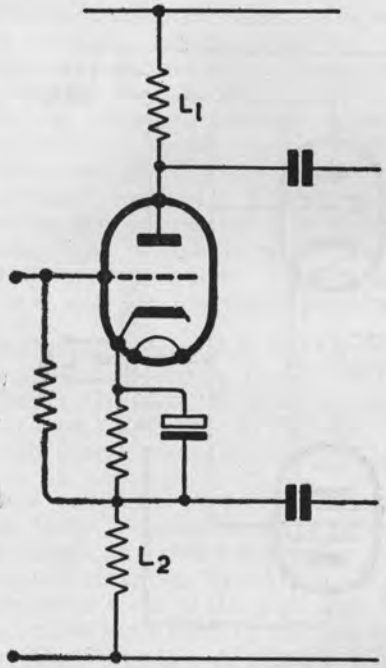


FIG. 15.

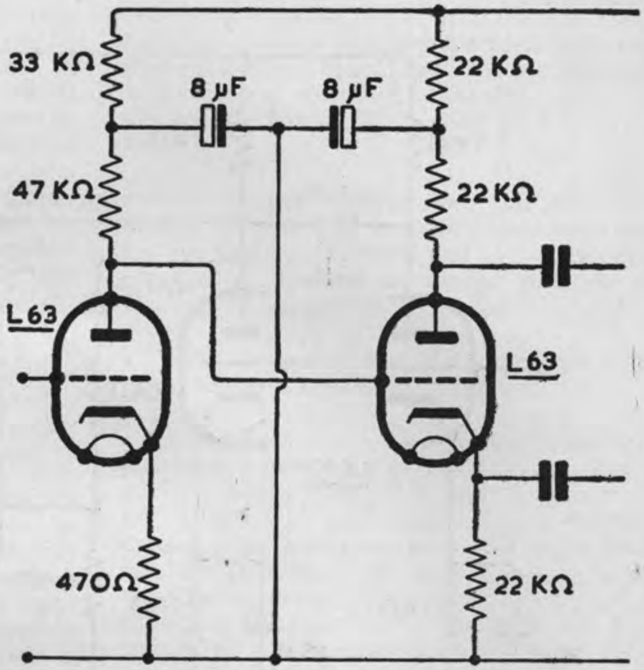


FIG. 16.

circuit was introduced by the G.E.C. a few years ago, the voltage amplifier is direct coupled to the phase-splitter. This circuit is reproduced in Fig. 16. Phase shift is lessened by the elimination of the more usual coupling capacitor. In operation the anode of the voltage amplifier is at a potential of +100v., which of course also applies to the grid of the phase-splitter. Due to the high value cathode resistor, a voltage drop of 105v. is obtained, this providing a grid bias of –5v. for this stage. Fig. 17 shows another form of splitter or invertor circuit, at one time quite popular but little used now, a double triode is best for this type though of course two separate triodes could be used. Its operation is as follows: the first section is designed to provide a gain of 18. This stage feeds one of the output valves, and also the second section of the phase-splitter. Note that in this case the anode load is much lower and the gain is 10, thus from input to the output of the second section there is a gain of 180. Since the input of the second section is fed from the output of the first it follows that the phase will be reversed, as required. Now one output valve is fed from the first section of the phase-splitter with a gain of 18, if the second output valve was fed from the output of the second section with its gain of 180 the circuit would be unbalanced. In practice the difficulty is overcome by tapping the second output valve's grid down the grid resistor. Two resistors are used and the values so arranged that the valve receives only one tenth of the available output, that is, there is a gain of 18 from input to each output valve grid.

In the opinion of the Author, the Schmidt phase-splitter is the best of them all, this appears in Fig. 18. It is simple and self-balancing, almost any double triode can be made to operate satisfactorily. With an ECC35 a gain of 30 to each output valve grid is possible.

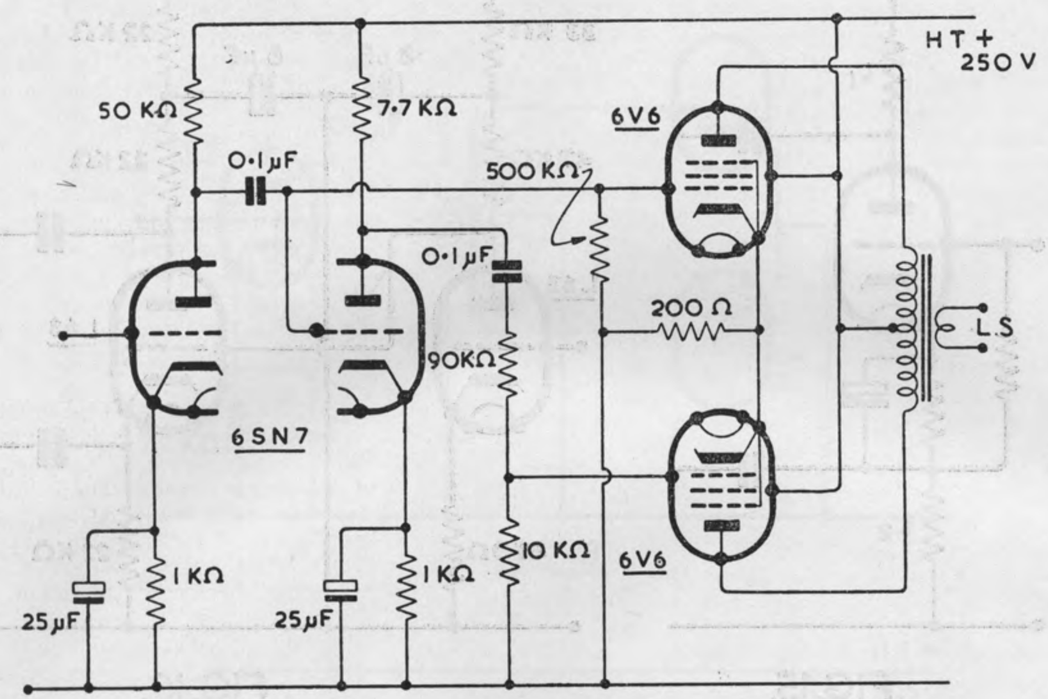

HT +
250 V
50 KΩ
7·7 KΩ
0·1 μF
500 KΩ
6V6
0·1 μF
200 Ω
L.S
90KΩ
6SN7
10 KΩ
6V6
1 KΩ
25 μF
1 KΩ
25 μF

FIG. 17.

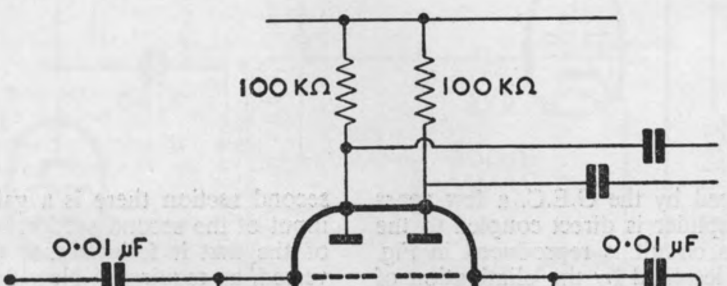

100 KΩ
100 KΩ
0·01 μF
0·01 μF
1 MΩ
1 MΩ
1 KΩ
47 KΩ

FIG.18.

CHAPTER SIX

MIXER CIRCUITS AND TONE CONTROL

Often the constructor wishes to fit several inputs to an amplifier. Existing amplifiers can easily be converted by the addition of an extra unit. A small power pack to supply the extra unit is simple to construct, though there is usually sufficient reserve in an existing amplifier to feed the extra circuit. Fig. 19 shows a conventional arrangement using an ECC40 double triode. Available gain is low, but it is of use where the intended input is high as in the case of a radio-tuner or crystal pick-up. Voltage gain for this unit is 4, and two inputs are provided.

Another useful circuit is that shown in Fig. 20. Two valves are shown though in practice any number can be used. Each valve should be of a similar type, 6SN7, B65, ECC83, etc. Types with differing characteristics should not be mixed.

Rk is calculated according to the number of valves used. Assume three ECC83's are to be used: Ia per triode=0.5 mA and a bias of 1 volt is required therefore Rk=330Ω. Rc should be ten times the value of Rk multiplied by the number of triode units used, in this case six, therefore Rc= 20kΩ approx. A suitable value of anode load resistance is some four times the anode impedance of one triode which in this case amounts to 320kΩ. Gain will be 40 for each unit. Such a gain will usually avoid the use of extra pre-amplification for microphones.

Though satisfactory for most applications this circuit has its troubles if the voltage drop across Rc is excessive. With six inputs using type ECC83 the drop is 60 volts, had 6SL7's been used, the drop would have been 120v. For efficient operation at least 150 volts apart from the drop across Rc is required. Where a sufficiently high voltage is not readily available Rc may be replaced by a valve as in Fig. 21.

It is not generally realised that the impedance of a valve is considerably higher than its D.C. resistance. An EL33 has a D.C. resistance of 7,000 ohms, but the impedance is 50,000 ohms (Vgl,-6v.). At zero voltage on gl the impedance remains constant but the D.C. resistance falls to 3,000 ohms. It is quite in order to operate the valve under zero bias conditions provided the max. power ratings are not exceeded.

Tone Control

Tone correction is a difficult subject. It would seem that by providing an amplifier with a level response, everyone should be satisfied. Unfortunately the human ear plays a big part and no two people are satisfied with the same overall response. Again gramophone recordings do not provide a linear output, and the modern pick-up requires some correction. Even the loud-speaker has certain characteristics which must be provided for.

With modern designs, two variable controls usually give all the flexibility necessary, that is, top lift and cut, bass lift and cut. Consider the effect of a capacitor in parallel with the anode load of an amplifying valve.

As the frequency rises, the impedance will fall, thus providing a measure of top cut.

Fig. 22 illustrates a practical application of this technique. Here we have a triode amplifier with a 33kΩ anode load a 0.01 mfd. coupling capacitor, a 470 kΩ grid resistor and a 0.005 mfd. capacitor for tone correction. At audio frequencies the impedance of the coupling capacitor is small compared to the anode load and that of the 8 mfd. de-coupling electrolytic capacitor may be considered a short circuit as far as audio frequencies are concerned. In effect, therefore, the anode load, grid resistor and tone corrector capacitor are all in parallel.

Now consider the effect on the gain, if the anode load is equal to four times the valve's impedance, gain will be 0.8 of the amplification factor quoted for the valve. As frequency rises the impedance of the tone corrector capacitor falls until a point is reached where the effective value of the anode load resistor is equal to the valve's impedance, the gain then falls to 0.5 of the amplification factor, in other words the high frequency response has been cut or attenuated. By including resistance in series with the tone corrector capacitor the effect can be limited so that above a selected frequency the gain remains constant. A very useful flexible arrangement of this principle is shown in Fig. 23. It can be used either across the voltage amplifier anode load or across the output transformer primary. The former is the better arrangement, because with the latter there is risk of creating a resonant condition with the output transformer primary, resulting in unintentional bass lift giving an unpleasant booming effect. It is also possible to provide top lift by use of capacitors as in Fig. 24. The original grid resistor is replaced by two of equal value, capacitors of various values are switched in circuit across to top section. Values for these capacitors are best found by trial, the higher the value of the resistance network the smaller the required capaci-

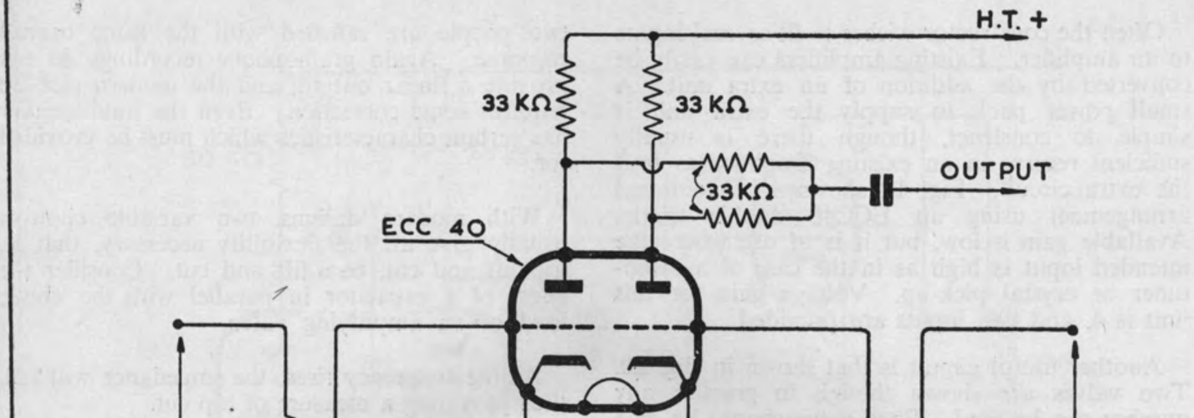

H.T. +
33 KΩ
33 KΩ
33 KΩ
OUTPUT
ECC 40
INPUT 1
·25 MΩ
25μF
600 Ω
INPUT 2
·25 MΩ

FIG. 19.

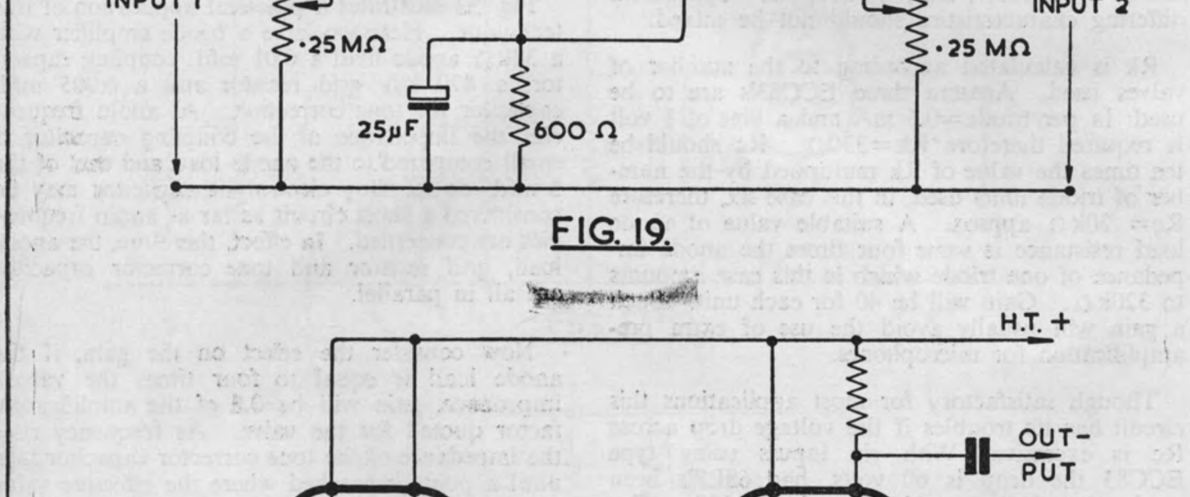

H.T. +
OUT-PUT
R
K
R
C
INPUT 1
INPUT 2
INPUT 3
INPUT 4

FIG. 20.

tance to provide a given amount of top lift.

Bass cut circuits are very simple. One of the most effective is to reduce the value of the coupling capacitance between voltage amplifier and output valve. This can be arranged with a rotary switch as for the top cut and top lift circuits. Three values will usually provide sufficient range, 0.1 mfd., 0.01 mfd. and 0.001 mfd.

Bass lift can be obtained in a number of ways, one is to include a tuned circuit in the grid wiring of one stage, for practical reasons it is usually better to include it in the anode circuit as shown in Fig. 25. With this circuit the resistance of the anode load resistor is only a fraction of the impedance of the valve.

The choke and capacitor should resonate at some 70 cycles, 50 or 100 cycles must be avoided otherwise residual mains hum and pick-up hum will be amplified out of all proportion. This method can be used over two or more stages so as to give boost at different frequencies.

Resonant circuits using chokes are rarely used in modern equipment and a much better system is shown in Fig. 26. An extra valve is used for bass amplification. It will be obvious that the amount of boost possible is greatly increased.

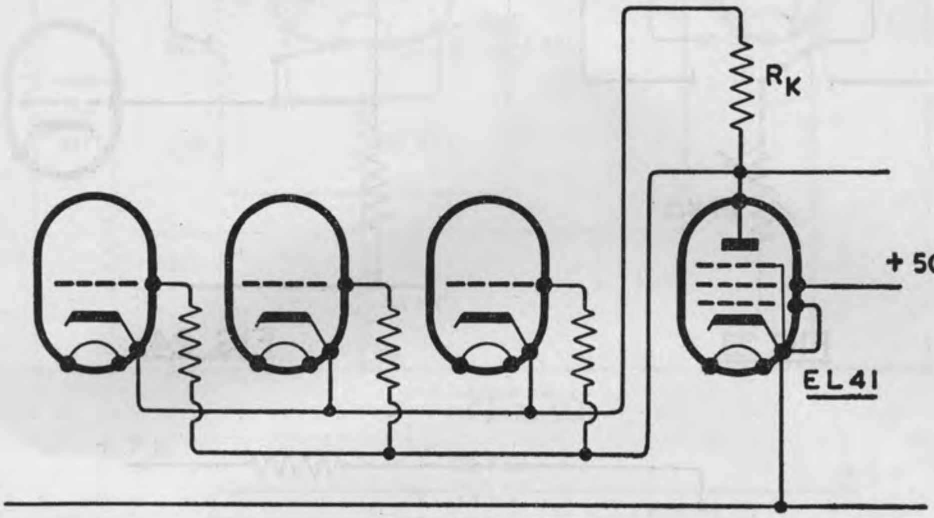


FIG. 21.

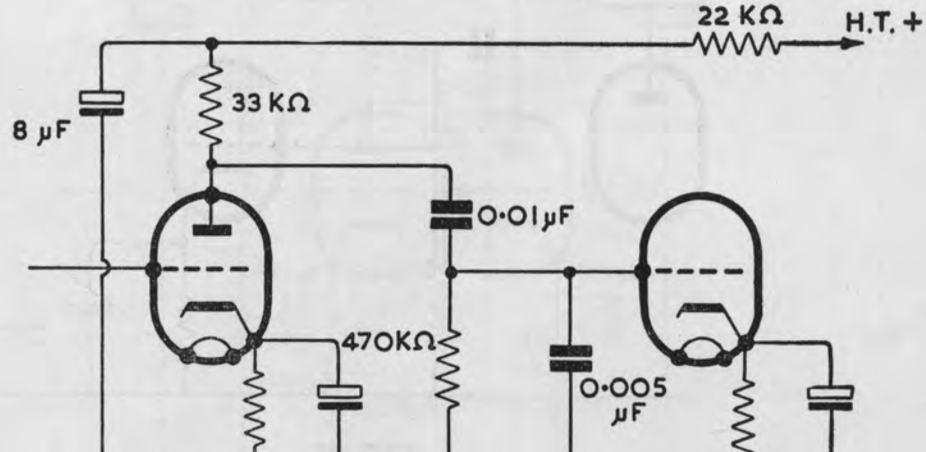


FIG. 22.

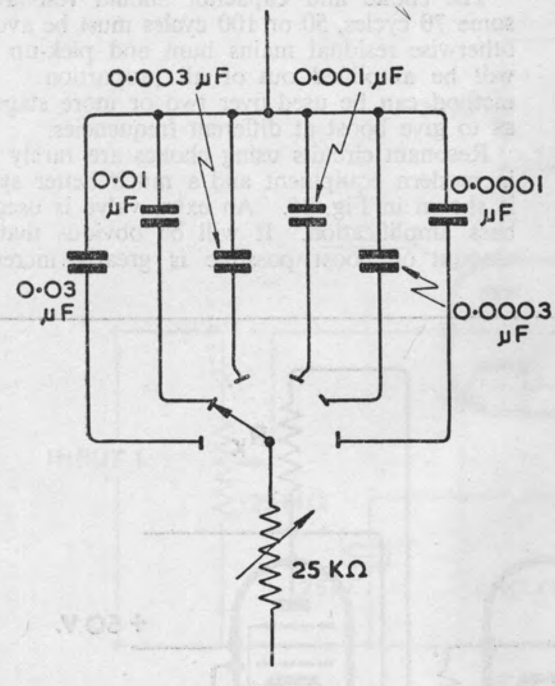


FIG. 23

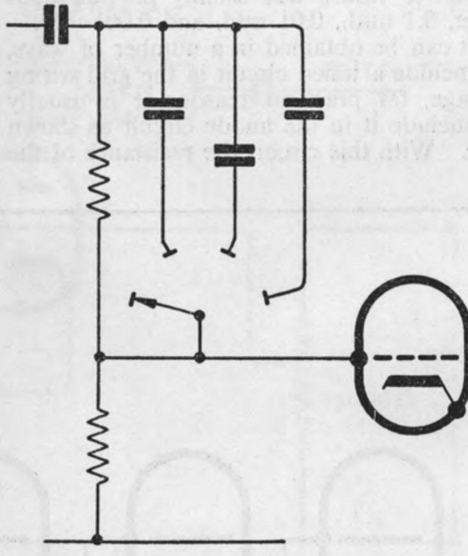


FIG. 24

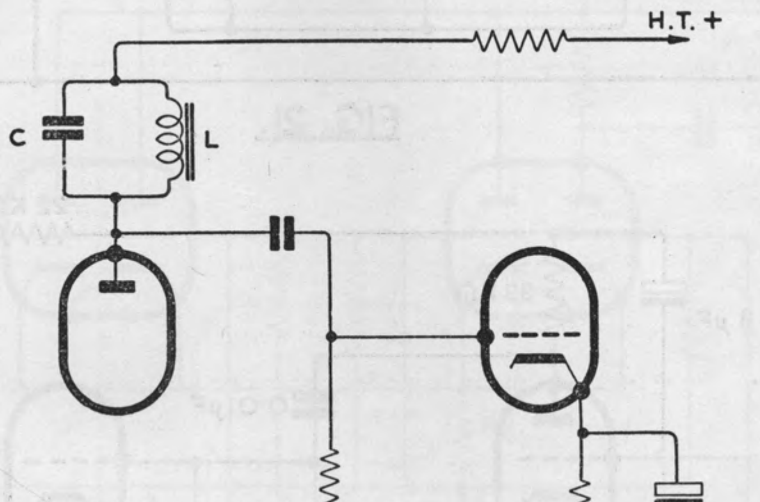


FIG.25

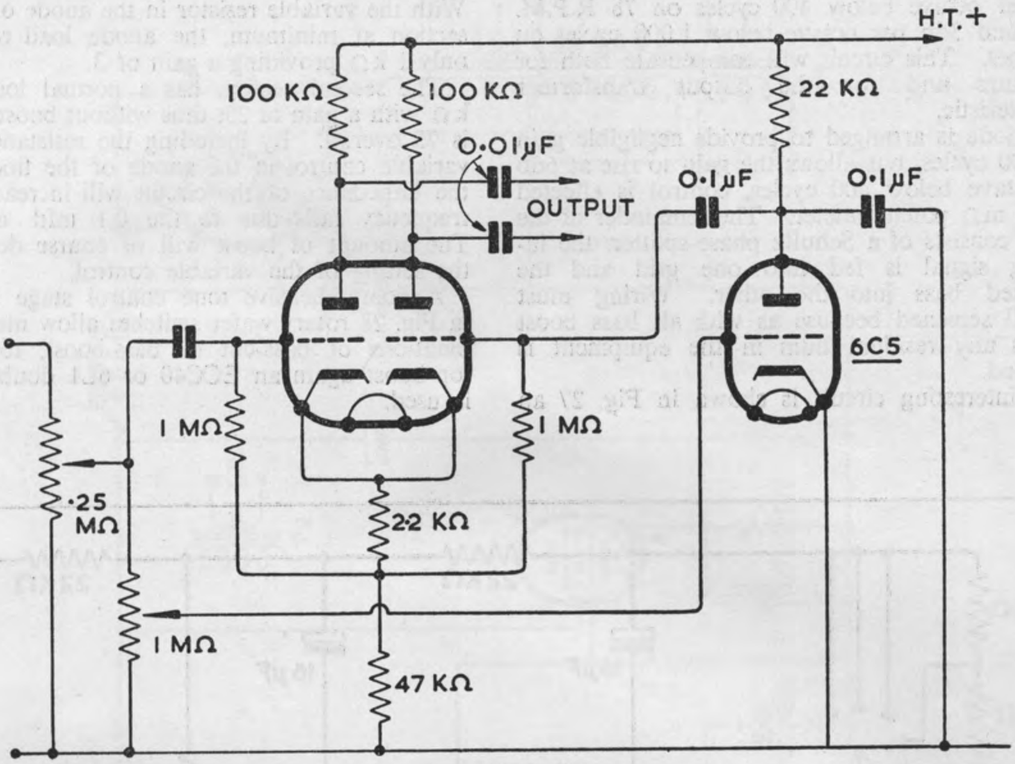

H.T. +
100 KΩ
100 KΩ
22 KΩ
0·01μF
OUTPUT
0·1μF
0·1μF
6C5
1 MΩ
1 MΩ
·25 MΩ
2·2 KΩ
1 MΩ
47 KΩ

FIG. 26

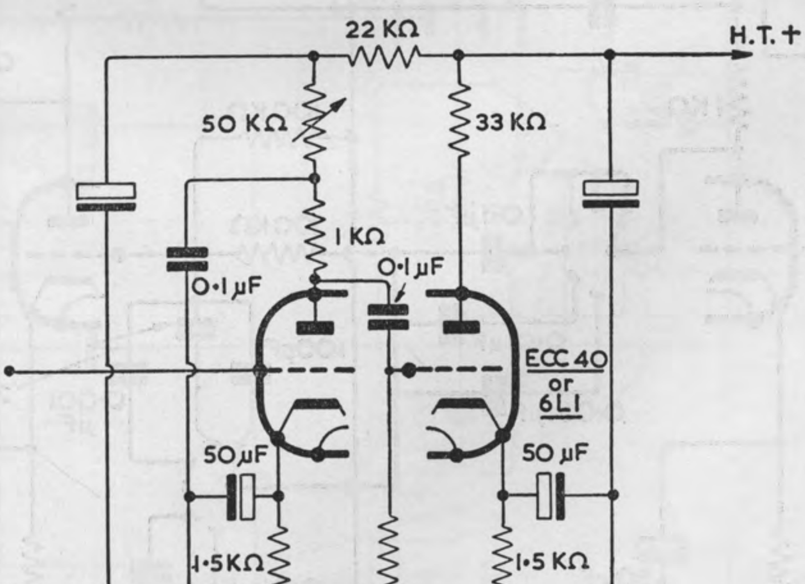

22 KΩ
H.T. +
50 KΩ
33 KΩ
1 KΩ
0·1μF
0·1μF
ECC 40 or 6L1
50μF
50 μF
1·5KΩ
1·5 KΩ

FIG. 27.

Gramophone records have a reduced response of 3db per octave below 400 cycles on 78 R.P.M. types and 3db per octave below 1,000 cycles on LP types. This circuit will compensate both for recordings and for the output transformer characteristic.

A triode is arranged to provide negligible gain at 1,000 cycles, but allows the gain to rise at 6db per octave below 400 cycles, control is effected by a 1 mΩ potentiometer. The remainder of the circuit consists of a Schultz phase-splitter, the incoming signal is fed into one grid and the amplified bass into the other. Wiring must be well screened because as with all bass boost circuits any residual hum in the equipment is amplified.

An interesting circuit is shown in Fig. 27 an ECC40 or 6L1 is best for this arrangement. With the variable resistor in the anode of the first section at minimum, the anode load resistor is only 1 kΩ providing a gain of 3.

The second section has a normal load of 33 kΩ with a gain of 25, thus without boost the gain is 75 overall. By including the resistance of the variable control in the anode of the first section the impedance of the circuit will increase as the frequency falls due to the 0.1 mfd. capacitor. The amount of boost will of course depend on the setting of the variable control.

A comprehensive tone control stage is shown in Fig. 28 rotary wafer switches allow many combinations of bass-cut or bass-boost, top-cut or top-boost again an ECC40 or 6L1 double triode is used.

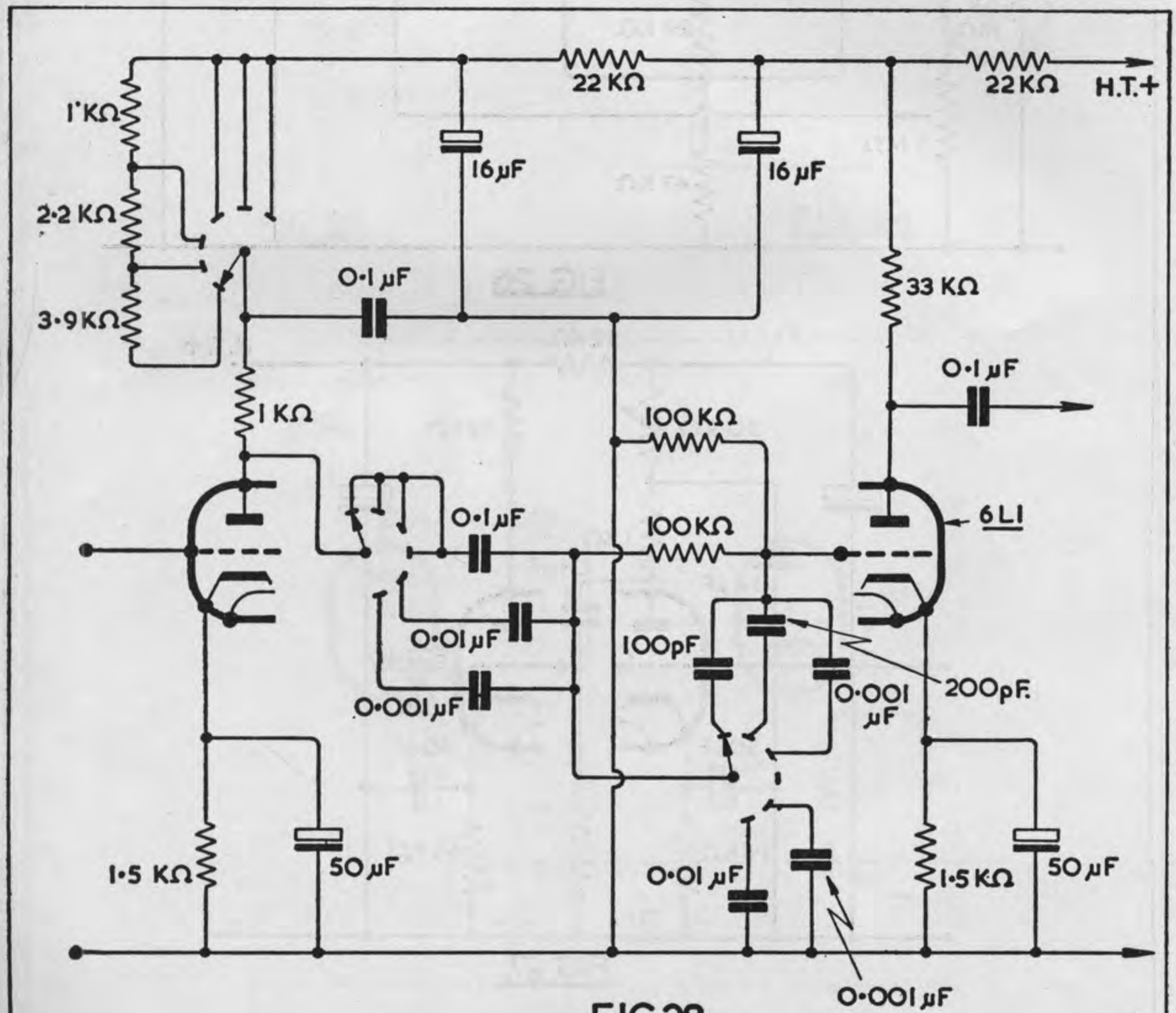


FIG.28.

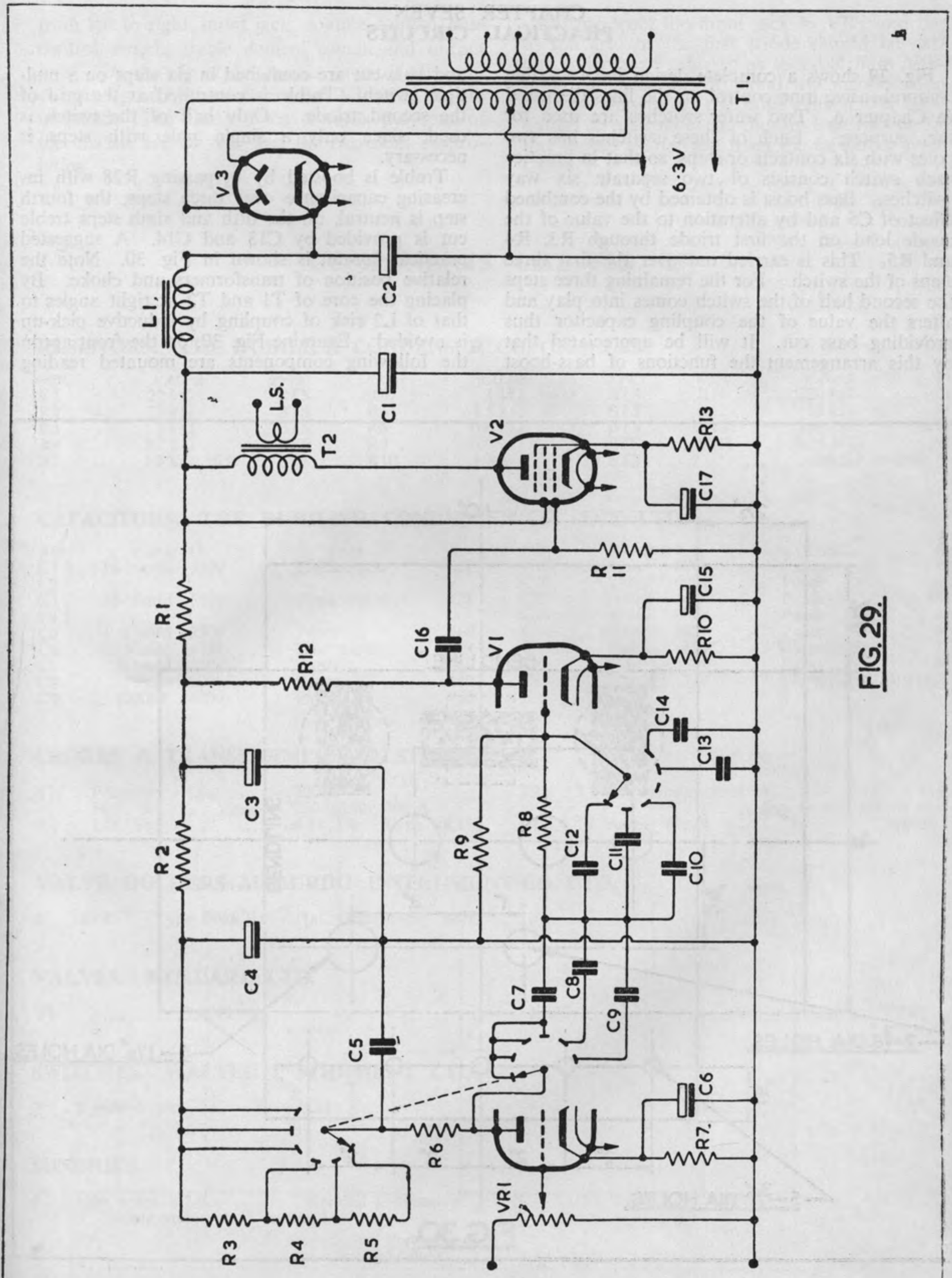

T1
6·3V
V3
C2
L1
C1
LS.
T2
V2
R13
C17
R
11
C15
R1
C16
V1
R10
R12
C14
C13
C3
R8
C12
C11
R9
R2
C10
C4
C8
C7
C9
C5
C6
R6
R7.
VR1
R3
R4
R5

FIG. 29.

CHAPTER SEVEN
PRACTICAL CIRCUITS

Fig. 29 shows a complete design incorporating comprehensive tone control on the lines discussed in Chapter 6. Two wafer switches are used for this purpose. Each of these switches has two poles with six contacts or steps, so that in practice each switch consists of two separate six way switches. Bass boost is obtained by the combined effect of C5 and by alteration to the value of the anode load on the first triode through R3, R4 and R5. This is carried out over the first three steps of the switch. For the remaining three steps the second half of the switch comes into play and alters the value of the coupling capacitor thus providing bass cut. It will be appreciated that by this arrangement the functions of bass-boost and bass-cut are combined in six steps on a multiple switch. Treble is controlled at the grid of the second triode. Only half of the switch is used, since only a single pole with steps is necessary.

Treble is boosted by by-passing R28 with increasing capacitance over three steps, the fourth step is neutral, on the fifth and sixth steps treble cut is provided by C13 and C14. A suggested practical lay-out is shown in Fig. 30. Note the relative position of transformers and choke. By placing the core of T1 and T2 at right angles to that of L2 risk of coupling by inductive pick-up is avoided. Examine Fig. 30, on the front apron the following components are mounted reading

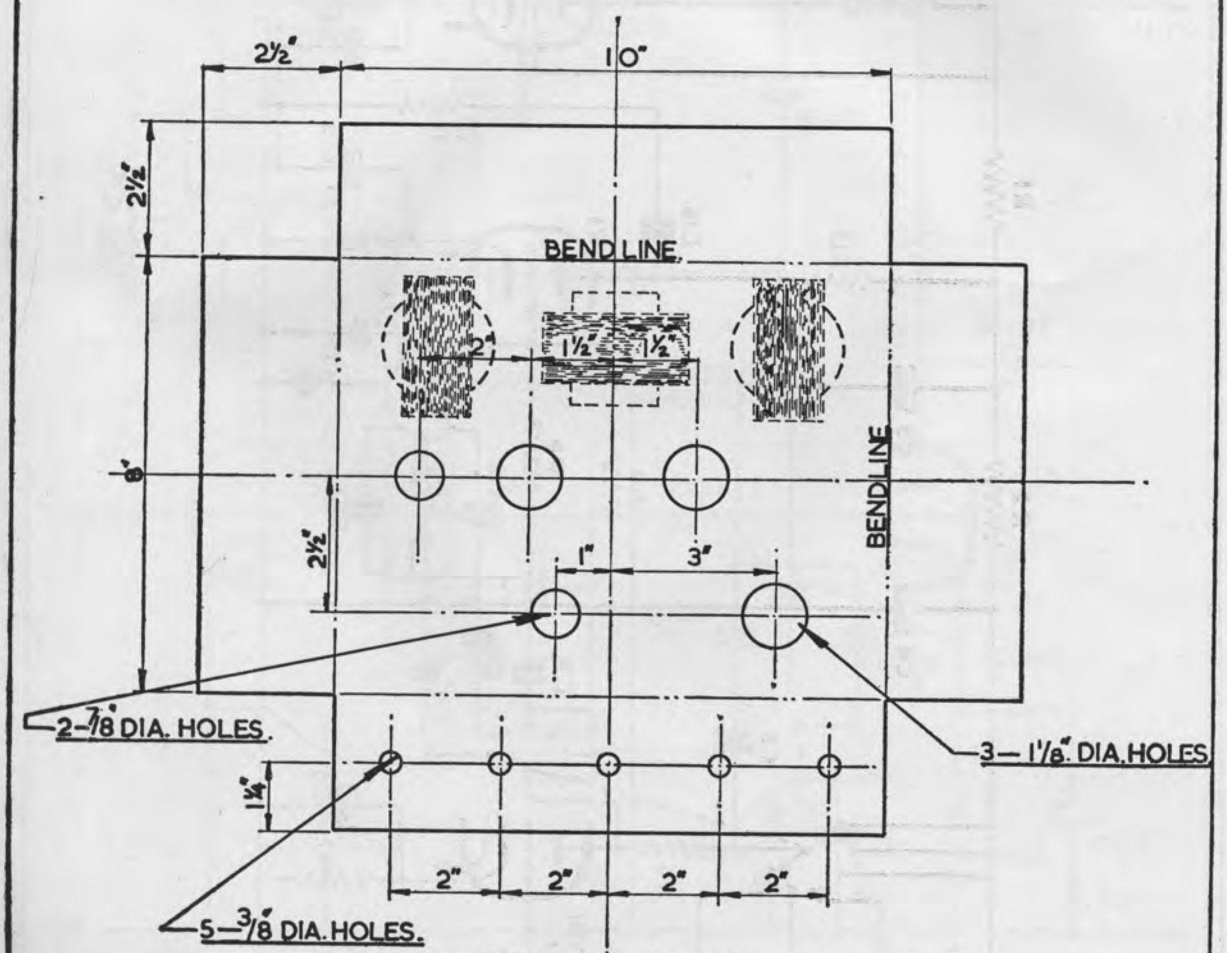


FIG. 30.

from left to right, input jack, volume control, bass control switch, treble control switch and output jack. On the top of the chassis, towards the front chassis apron, valve-holders for V1 and V2 are mounted. The three holes towards the rear of the chassis are for V3, and the two dual electrolytics.

Wiring from the input jack to VR1 and thence to the grid of the first triode should be carried out in screened cable so as to avoid hum pick-up. It is also necessary to bond the metal casing of VR1 to chassis for the same reason. If it is required to calculate the output transformer ratio, this should be derived from the formulae given in Chapter 3.

COMPONENTS LIST FIG. 29

RESISTORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	*VALUE*	*TYPE*	*REF.*	*VALUE*	*TYPE*	*REF.*	*VALUE*	*TYPE*
R1	22 kΩ	BTS	R6	1 kΩ	BTS	R11	220 kΩ	BTS
R2	22 kΩ	BTS	R7	1.5 kΩ	BTS	R12	33 kΩ	BTS
R3	1 kΩ	BTS	R8	100 kΩ	BTS	R13	150 Ω	BWF2
R4	2.2 kΩ	BTS	R9	100 kΩ	BTS	VR1	0.25 MΩ with S.P. switch	CBS
R5	3.9 kΩ 10%	BTS	R10	1.5 kΩ	BTS			

CAPACITORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	*CAPACITY*	*TYPE*	*No.*	*REF.*	*CAPACITY*	*TYPE*	*No.*
C1 } C2 }	16-16mfd 350v	Electrolytic	CT	C10	100pF 600v	Paper	400
				C11	200pF 600v	Paper	400
C3 } C4 }	16-16mfd 350v	Electrolytic	CT	C12	1000pF 500v	Paper	400
				C13	100pF 600v	Paper	400
C5	0.1mfd 350v	Paper	460	C14	1000pF 500v	Paper	400
C6	50mfd 12v	Electrolytic	BR	C15	50mfd 12v	Electrolvtic	BR
C7	0.1mfd 350v	Paper	460	C16	0.1mfd 350v	Paper	460
C8	0.01mfd 1000v	Paper	460	C17	50mfd 12v	Electrolytic	BR
C9	1000pF 500v	Paper	400				

CHOKES & TRANSFORMERS. ELSTONE.

T1 Primary 200/250v
H.T. Sec. 250-0-250v 80mA
L.T. Sec. ... ... 6.3v 3A Type SR250

T2 7,000Ω Primary load Type MR/7
Sec. to suit speech coil.
L1 10 Henry 80mA ... Type SC/80

VALVE HOLDERS McMURDO INSTRUMENT CO. LTD.

2 B8A Type BM8/E1 Int. Octal Type B8/U

VALVES. MULLARD LTD.

V1 ... ECC40 V2 ... EL33 V3 ... EZ40

SWITCHES. WALTER INSTRUMENT LTD.

2 2 pole 6 way ... Type CH

SUNDRIES.

1 10in. x 8in. x 2½in. ... Kendall & Mousley chassis and case if required.

2 Jacks, Type J.6 ... Bulgin
2 Plugs, Type P.38 ... Bulgin

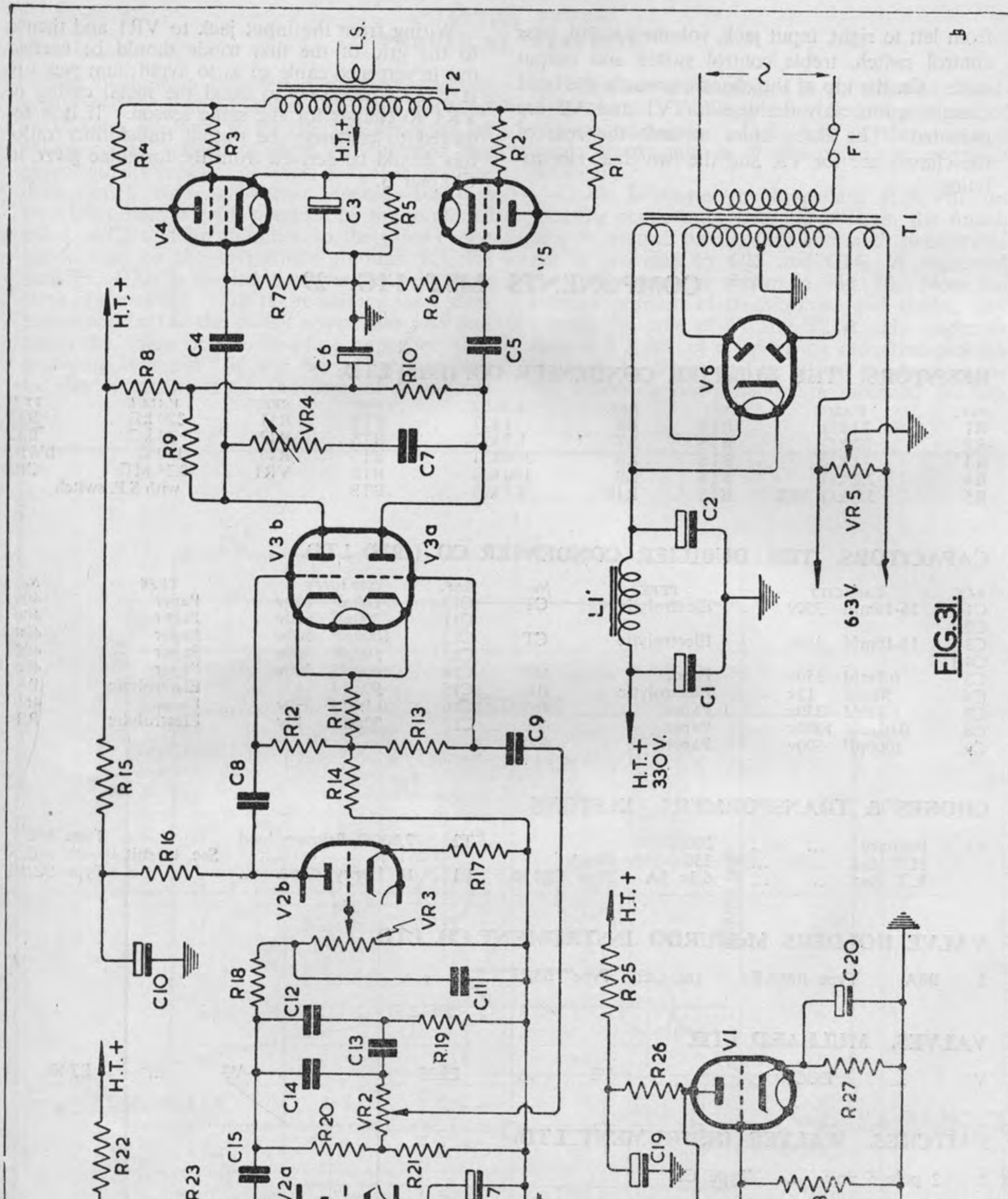

L.S.
T2
H.T.+
R1
R2
R3
R4
R5
R6
R7
R8
R9
R10
R11
R12
R13
R14
R15
R16
R17
R18
R19
R20
R21
R22
R23
R24
R25
R26
R27
R28
C1
C2
C3
C4
C5
C6
C7
C8
C9
C10
C11
C12
C13
C14
C15
C16
C17
C18
C19
C20
V1
V2a
V2b
V3a
V3b
V4
V5
V6
VR1
VR2
VR3
VR4
VR5
F1
T1
L1
6·3V
H.T.+ 330V.
S1
R
G
RADIO INPUT
GRAM INPUT

FIG. 31.

HARMONIC DISTORTION AT 1,000 CYCLES

(tone controls at flat frequency response —)

OUTPUT (watts)	% 2nd HARMONIC	% 3rd HARMONIC	% 4th HARMONIC
2	0·68	0·34	<0·02
3	0·56	0·64	0·20
4	0·75	0·46	0·07
5	0·18	5·00	0·30

FREQUENCY RESPONSE

(—gain control at maximum)

FREQUENCY (cycles per second)	MIN. TOP MIN. BASS (flat response, 40 mV input.) dB	MAX. TOP MIN. BASS (40mV input) dB	MIN TOP MAX. BASS (5mV input) dB	MAX TOP MAX. BASS (5mV input) dB	TOP CUT MAX TOP BOOST MIN BASS BOOST MIN dB
50	−0·5	−0·4	+20 (3·5 watts)	+20 (1·4 watts)	−0
100	0	0	+14·5	+16	−0
200	0	+0·4	+9·5	+10·5	0
400	0	+0·4	+5·2	+5·5	−0·4
1,000	0 (0·5 watt)	0 (0·33 watt)	0 (0·035 watt)	0 (0·014 watt)	−2
2,500	−0·2	+1·4	−2·5	−3·5	−7
5,000	−0·2	+7	−3·5	+2·5	−13
10,000	+0·2	+10	−3·5	+7	−19
20,000	+0·2	+11 (4·1 watts)	−3·5	+7·5 (0·78 watts)	−30

FIG. 32.

The author is indebted to D. N. Corfield of Standard Telephones & Cables Ltd. for permission to use the three following designs (Fig. 31, 34 and 36). The first of these is designed for A.C. operation. It includes a number of interesting features all of which contribute to its fine performance as a high-fidelity amplifier.

A separate voltage amplifier is used as a pre-amp for gramophone reproduction. Gain is high, and an input of 150 mV will fully load the output stages. Miniature moving iron pick-ups provide a very small output and usually require a separate pre-amplifier, this design is however quite suitable, without any additional amplification. Crystal heads are reasonably sensitive and it may be an advantage to reduce R26 if one of these is used. R28 is specified as 1M Ω but this value may be changed to suit the requirements of the particular pick-up used.

Adequate tone-control is obtained through V2, a double triode arranged as a voltage amplifier, and bass amplifier VR2 controls the top response and VR3 the bass. Top cut is provided by VR4.

Signal voltage is fed from the output of V2a to V3a, while V2b feeds V3b, any unbalance is corrected by R14. It is essential that the grid and anode resistors are within the specified tolerances, otherwise the quality cliamed will not be achieved. With the specified values, the push-pull voltages applied to the 6V6G grids are balanced within 1db between 50 cycles and 12Kc/s. Available output power is between 2 and 5 watts at an extremely low distortion factor. This can be seen from Fig. 32 which also shows the frequency response that can be expected at various settings of the controls. Both L.P. and standard recordings can be brilliantly reproduced through this design. A suggested layout showing the position of the main components appears in Fig. 33. To avoid hum both V1 and V2 should be enclosed in screening cans.

For constructors with facilities for winding their own output transformer the specification is as follows:—

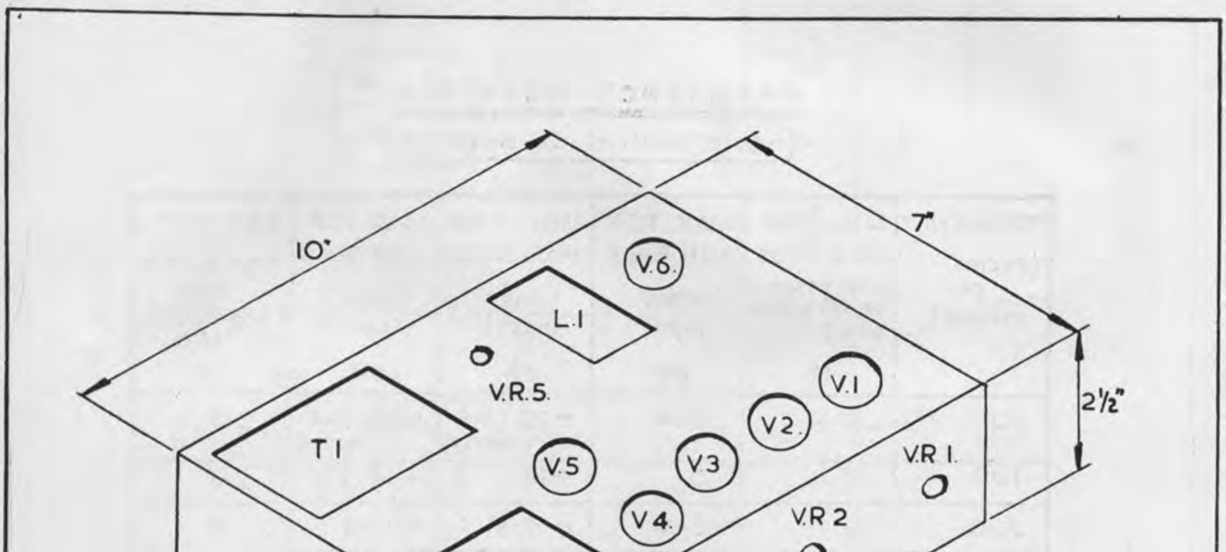


FIG.33.

Laminations:	Sankey 60A, ¾" stack, butt joint.
Primary:	2 sections each 1,800 turns 36 s.w.g. S.S. Enam.
Secondary:	3 sections each 70 turns 24 s.w.g. Enam.

All sections are interleaved to reduce leakage inductance.

Total primary resistance 280 Ω, secondary 2.2 Ω. First wind on the first section of the secondary. Each layer must be wound evenly, and a layer of paper is inserted between layers. After the first secondary section is complete, insulate with several paper layers and wind on the first primary section. Continue this process until all sections are complete. This specification will match the amplifier to a loud speaker with a speech coil impedance of 15 Ω.

At first sight the output power may seem low considering that two beam tetrodes are used in the output stage, but it must be appreciated that in the interests of high quality, they are triode connected. It is interesting to note that the high quality claimed for this design is obtained without recourse to voltage negative feed-back.

When carrying out initial tests, VR5 is adjusted for minimum hum, the rest of the controls require no further comment.

COMPONENTS LIST FIG. 31

RESISTORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	VALUE	TYPE	REF.	VALUE	TYPE	REF.	VALUE	TPPE
R1	100 Ω	BWF2	R12	1 MΩ	BTS	R23	47 kΩ	BTS
R2	100 Ω	BWF2	R13	1 MΩ	BTS	R24	1 kΩ	BTS
R3	100 Ω	BWF2	R14	47 kΩ	BTS	R25	47 kΩ	BTS
R4	100 Ω	BWF2	R15	47 kΩ	BTS	R26	100 kΩ	BTS
R5	240 Ω 5w	A1/1	R16	47 kΩ	BTS	R27	3.3 kΩ	BTS
R6	100 kΩ 10%	BTS	R17	1 kΩ	BTS	R28	1 MΩ	BTS
R7	100 kΩ 10%	BTS	R18	220 kΩ	BTS	VR1	500 kΩ	C
R8	10 kΩ	BTS	R19	220 kΩ	BTS	VR2	500 kΩ	C
R9	47 kΩ 5%	R425	R20	220 kΩ	BTS	VR3	1 MΩ	C
R10	47 kΩ 5%	R425	R21	22 kΩ	BTS	VR4	250 kΩ	C
R11	470 Ω	BTS	R22	10 kΩ	BTS	VR5	100 Ω	

All resistors ½w. 20% unless otherwise stated. Colvern CLR 1206/95

CAPACITORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	CAPACITY		TYPE	No.	REF.	CAPACITY			TYPE	No.
C1 } C2 }	16mfd	500v	Dual Electrolytic	CT	C11	0.1mfd	350v		Paper	460
					C12	60pF	350v	10%	Mica	S635
C3	50mfd	25v	Electrolytic	BR	C13	40pF	350v	10%	Mica	S635
C4	0.1mfd	350v	Paper	460	C14	10 pF	350v	10%	Mica	S635
C5	0.1mfd	350v	Paper	460	C15	0.1mfd	350v		Paper	460
C6	8mfd	350v	Electrolytic	BR	C16	8mfd	350v		Electrolytic	BR
C7	5000pF	1000v	Paper	460	C17	25mfd	12v		Electrolytic	MBR
C8	0.1mfd	350v	Paper	460	C18	0.1mfd	350v		Paper	460
C9	0.1mfd	350v	Paper	460	C19	8mfd	350v		Electrolytic	BR
C10	8mfd	350v	Electrolytic	BR	C20	25mfd	12v		Electrolytic	MBR

VALVES. BRIMAR

V1 ...	6J5GT	V3 ...	6SN7GT	V5 ...	6V6G
V2 ...	6SN7GT	V4 ...	6V6G	V6 ...	5Z4G

SWITCHES & FUSES. A. F. BULGIN

1	Gram/Radio	S.P. Changeover Type S273	F.1	Fuseholder	... ... Type F27/1
1	ON/OFF	S P ... ... Type S263	1	2 Amp Fuse	... ... Type F108

CHOKES & TRANSFORMERS.

T1	Primary	To suit requirements	T2	Primary Load . .	4,500Ω
	H.T. Sec. ...	320v-0-320v 120 mA			Sec. to suit loudspeaker
	L.T. Sec. ...	6.3v 4A 5v 2A	L1	7.5-10 Hy ...	120mA.

SUNDRIES.

2 Concentric plugs and Sockets ... ... Belling & Lee

1 10in. x 7in. x 2½in. Chassis and Case if required Kendall & Mousley

2 Octal valve screening cans

VALVE HOLDERS, McMURDO INSTRUMENT CO. LTD.

6 Int. Octal. Type B8/U

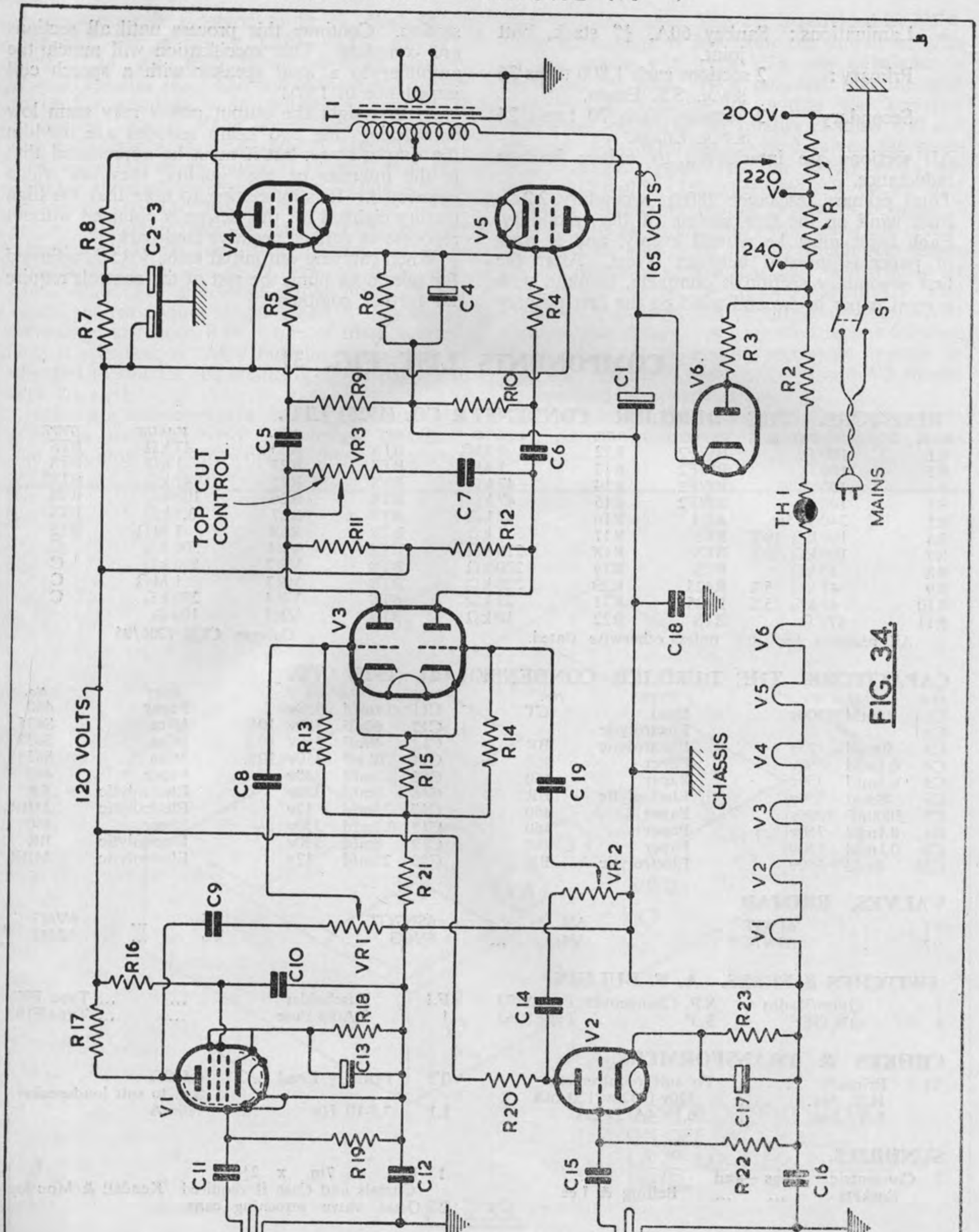

T1
R8
C2-3
V4
R7
R5
R6
C4
V5
R4
165 VOLTS
200V
220 V
240 V
R 1
S1
R 3
V6
C1
R2
R9
R10
C5
C6
TOP CUT CONTROL
VR3
C7
R11
R12
TH 1
MAINS
C18
V3
120 VOLTS
R13
R14
R15
C8
C19
CHASSIS
V1
V2
V3
V4
V5
V6
R21
VR2
C9
R16
C10
VR1
R17
R18
C14
R23
C13
V1
R20
V2
C17
C11
R19
C12
C15
R22
C16
MIC
P.U.

FIG. 34

A rather different design is that shown in Fig. 34.

It is suitable for either A.C. or D.C. operation, consequently no mains transformer is used. As is usual with such designs a series heater chain is employed, the valves are all designed for this purpose and consumption is 0.15 Amp. Note the thermistor TH1, Fig. 34. This prevents the injurious surge usually associated with this type of equipment. When first switched on TH1 has a high resistance whereas that of the valve heaters is practically nil, as the circuit warms up the resistance of the thermistor falls until the correct operating conditions are arrived at. It is important that the heaters are wired up in the order shown, otherwise hum may be picked up in the early stages. Overall frequency response is shown in Fig. 35. An output of 4 watts is obtained with an input of 60mV at the gram. input and 20 mV at the microphone input.

It is essential that under any conditions the 100 Ω surge limiting resistor (R3) in the anode circuit of V6 should not be omitted; equally so the anode loads R11 and R12, and the following grid resistors R4 and R5 must be within the tolerances specified. This latter precaution ensures that the push-pull voltages at the grids of V4 and V5 is balanced to within 1dB between 50 cycles and 12 Kc/s. If desired the output transformer can be home constructed, the specification is as follows:—

Laminations:	$\frac{3}{4}''$ stack, Stalloy stype 60A, 0.014″ thick.
Primary:	2 sections, each 1,800 turns. 38 s.w.g. S.S. Enam.
Secondary:	Single section 195 turns. 26 s.w.g. Enam.

The secondary winding is wound between each half of the primary to prevent excessive leakage inductance. Both sides of the secondary are electrostatically screened by 0.007″ soft copper sheet suitably insulated, this screen must be earthed. Great care must be taken over the insulation since any contact between two layers of copper will result in a short circuit turn and the transformer will not operate.

Layout is not critical and can be arranged to suit individual requirements, the usual precautions against long grid and anode leads should be observed and it is as well to screen the input grid wiring to V1 and V2. Valves V1 and V2 should be enclosed in screening cans.

An earth is necessary with this amplifier, note that the input jacks are isolated from the chassis, and have one side directly connected to earth. This avoids any possibility of a shock from the metal screening of the microphone or pick-up connecting leads since they are automatically earthed. Under certain conditions, the chassis may be live during operation, it is therefore essential to house it in a wooden or similar container and to ensure that no metal parts directly connected to the chassis are exposed to accidental contact. Input jacks and the toggle switch are mounted off the chassis on an insulated panel, the variable controls are on the chassis and care must be taken to ensure the grub screws on the control knobs do not protude. The tone control is simple but effective, this can be clearly seen from Fig. 35.

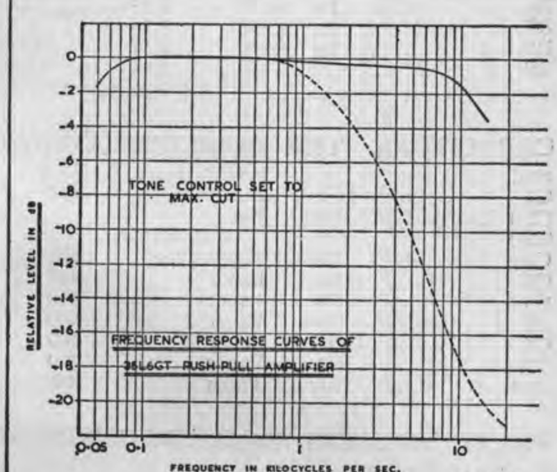


FIG. 35.

COMPONENTS LIST FIG. 34

RESISTORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	VALUE	TYPE	REF.	VALUE	TYPE	REF.	VALUE	TYPE
R1	80Ω 10w	A2/1A	R10	220kΩ	BTS	R19	470kΩ	BTS
R2	220Ω 25w	D2/1	R11	100kΩ	BTS	R20	100kΩ	BTS
R3	100Ω 5w	A1/1	R12	100kΩ	BTS	R21	47kΩ	BTS
R4	100kΩ	BTS	R13	1MΩ	BTS	R22	470kΩ	BTS
R5	100kΩ	BTS	R14	1MΩ	BTS	R23	3.9kΩ	BTS
R6	75Ω 1w	R627	R15	47kΩ	BTS	VR1	1MΩ	C
R7	2.2kΩ 2w	R850	R16	2.7MΩ	BTS	VR2	1MΩ	C
R8	2.7kΩ 2w	R850	R17	470kΩ	BTS	VR3	500kΩ	C
R9	220kΩ	BTS	R18	5.6kΩ	BTS			

All Resistors 10% unless otherwise specified.

CAPACITORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	CAPACITY	TYPE	No.	REF.	CAPACITY	TYPE	No.
C1	16μF 500v	Electrolytic	CT	C11	0.02μF 350v	Paper	Met. Min.
C2 } C3 }	32μF+32μF 350v	Dual Electrolytic	CT	C12	0.02μF 350v	Paper	Met. Min.
				C13	25μF 12v	Electrolytic	MBR
C4	50μF 12v	Electrolytic	BR	C14	0.1μF 350v	Paper	460
C5	0.1μF 350v	Paper	460	C15	0.02μF 350v	Paper	Met. Min.
C6	0.1μF 350v	Paper	460	C16	0.02μF 350v	Paper	Met. Min
C7	0.002μF 1000v	Paper	460	C17	25μF 12v	Electrolytic	MBR
C8	0.01μF 1000v	Paper	460	C18	0.01μF 1000v	Paper	460
C9	0.1μF 350v	Paper	460	C19	0.01μF 1000v	Paper	460
C10	0.1μF 350v	Paper	460				

VALVE HOLDERS McMURDO INSTRUMENT CO. LTD.

6 Int. Octal Type B8/U

VALVES. BRIMAR: STANDARD TELEPHONES & CABLES LTD.

V1 ...	12J7GT	V3 ...	25SN7GT	V5 ...	35L6GT
V2 ...	12J5GT	V4 ...	35L6GT	V6 ...	35Z4GT

TRANSFORMERS.

T1 Pri. load = 5500Ω Sec. to suit speech coil.

SUNDRIES.

2	Input Jacks	Bulgin J.6
	Jack-Plugs	Bulgin P38
1	Thermistor (TH1) ..	Brimar CZ2
1	DPST toggle switch ...	Bulgin S267
	Chassis	10in. x 8in. x 2½in.
2	Valve Screening Cons.	(int. octal size)
1	Grid Clip	Bulgin P96

The Amplifier shewn in Fig. 36-37 is similar in form to the previous one. In this case, however, a large output has been provided for, 12-16 watts. This is achieved by a parallel push/pull output, and a heavy-duty power pack. Every care should be taken to ensure the heaters are wired up in the order shown in Fig. 37 otherwise pick-up hum will result. Both V1 and V2 should be enclosed in screening cans which also reduces any tendency to pick up unwanted hum. Sensitivity is excellent, measured at 400 cycles, an input of 40mV at the P/U jack or 7.5mV at the microphone jack will provide the full output of 15 watts. As will be readily appreciated, this amplifier is very suitable for P.A. work. Reference to the components list will show that a number of 5% and 10% resistors are specified, these values and tolerances should be strictly adhered to otherwise the performance will suffer. Except where otherwise specified, resistors are all 20% tol.

R1 requires some explanation, it is made up from a standard 0.3 Amp. dropper such as the Bulgin MR64 or similar type with a number of adjustable clips. Set one clip in the centre of the resistor this represents 240v input. Sections A and B are set with the aid of further clips, these sections regulate the heater supply at 220v and 200 v, each section should be adjusted to a resistance of 30Ω.

The remaining sections are for surge limiting, and C and D are set at 67Ω each. It can be seen from Fig. 37, that provided the input is to remain constant at 200v, R1 can be omitted. A simple top-cut filter is incorporated; its action is illustrated in Fig. 38.

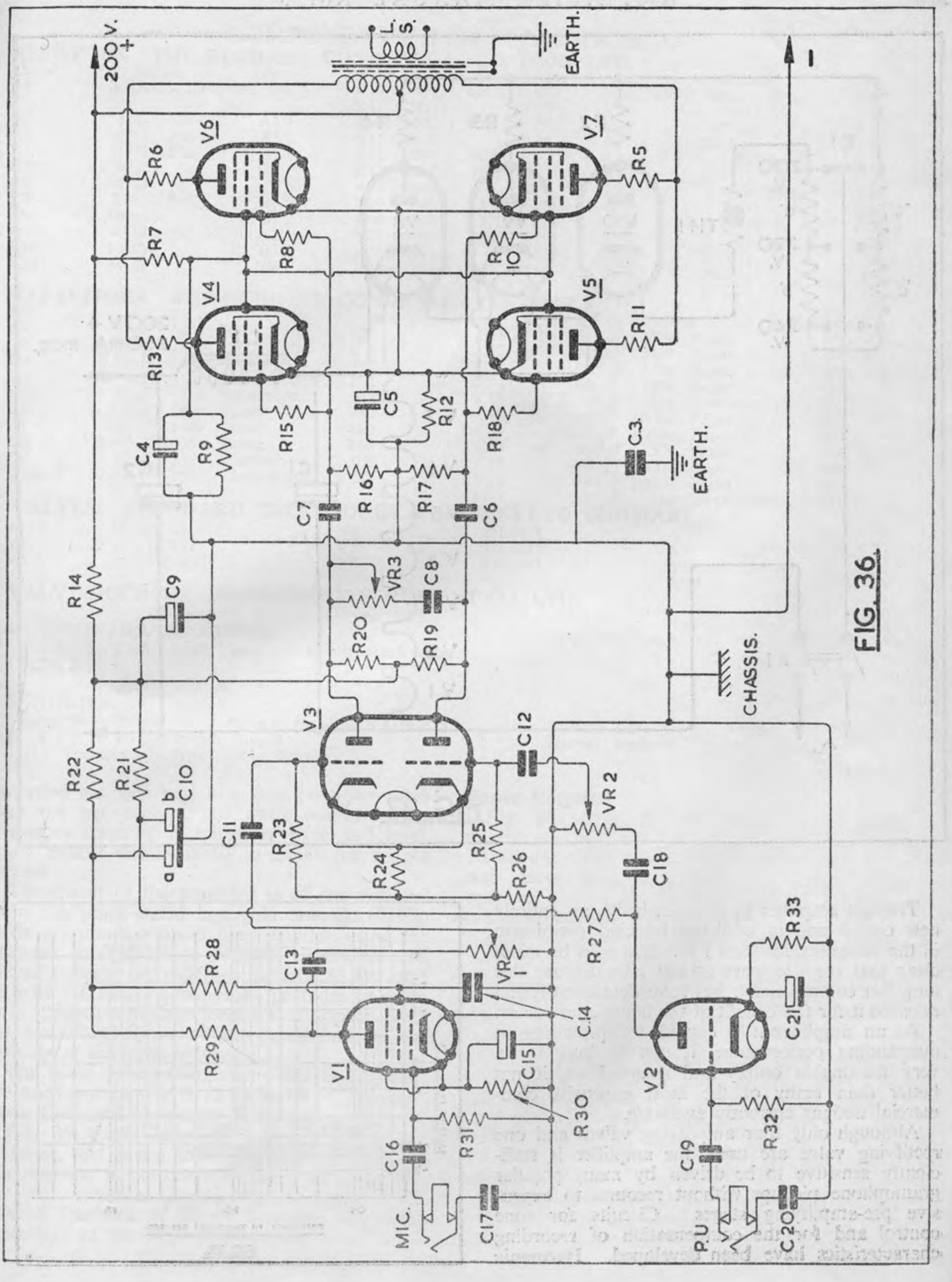

200 V. +
L.S.
EARTH.
V6
V7
R6
R5
R7
R8
R 10
V4
V5
R13
R11
C4
R9
R15
C5
R12
R18
C3
EARTH.
C7
R16
R17
C6
R14
C9
VR3
C8
R20
R19
CHASSIS.
V3
C12
R22
R21
b
C10
a
C11
R23
R24
R25
R26
VR 2
C18
R27
R33
R28
C13
VR 1
C14
R29
V1
C15
R30
C21
V2
R31
R32
C16
C19
C17
MIC.
P.U.
C20

FIG. 36.

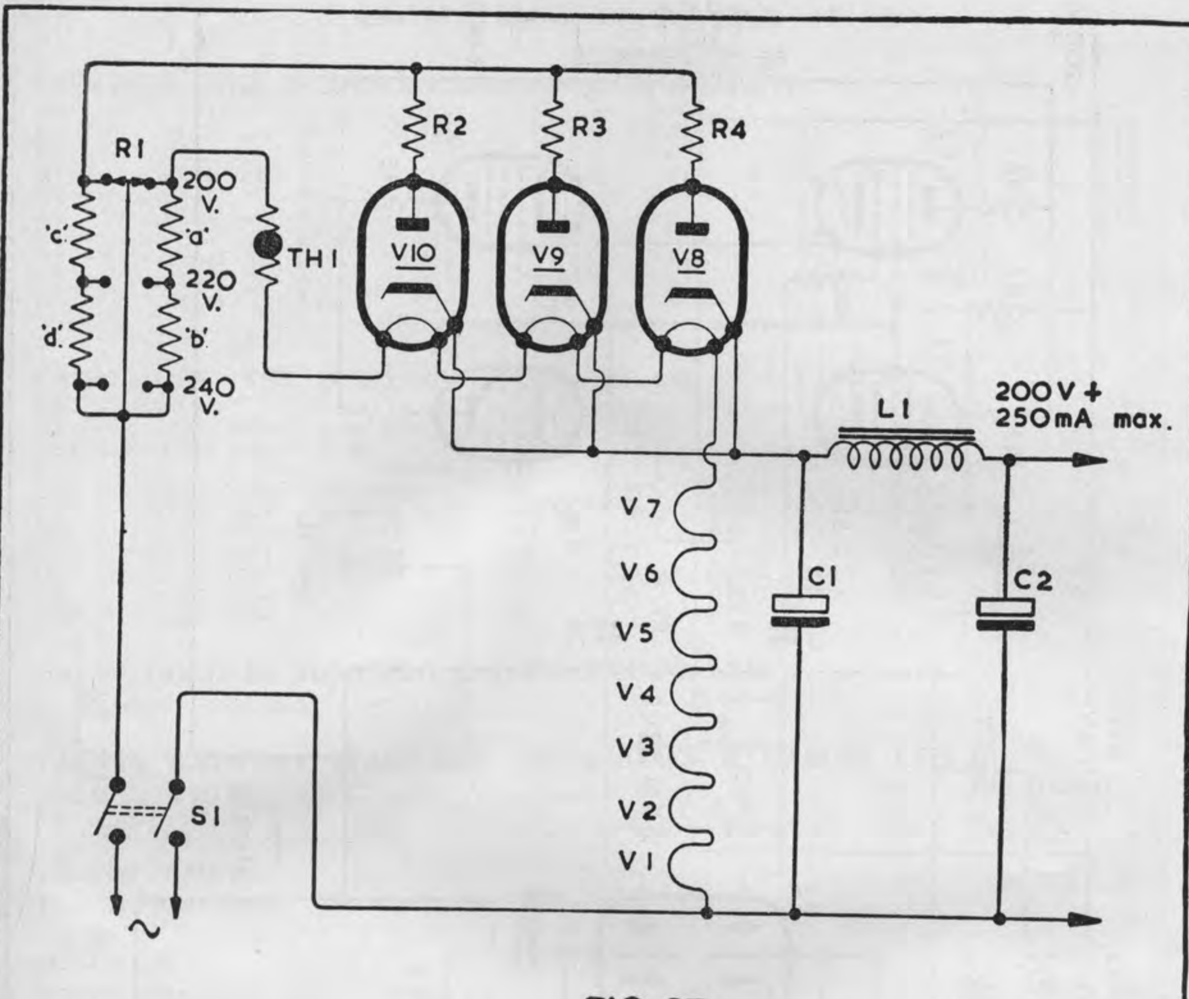


FIG. 37.

The last amplifier to be described is an entirely new circuit and is published by kind permission of the designers, Mullard Ltd. It should be made clear that the designers do not manufacture this amplifier commercially, but have developed it and released it for the benefit of the home constructor.

As an amplifier it is capable of providing an outstanding performance, it can be built for a very reasonable outlay and is equal to, if not better than many of the most expensive commercial designs currently available.

Although only four amplifying valves and one rectifying valve are used, the amplifier is sufficiently sensitive to be driven by many popular gramophone pickups without recourse to expensive pre-amplifying stages. Circuits for tone control and for the compensation of recording characteristics have been developed. Harmonic

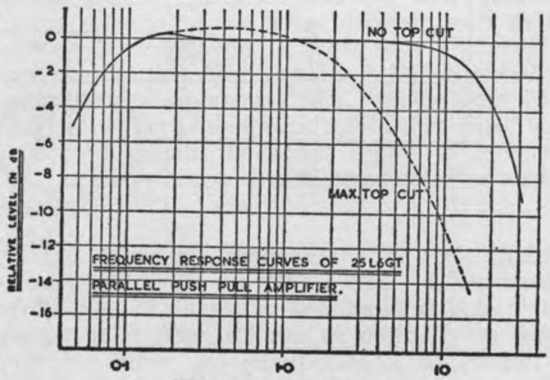


FIG. 38.

COMPONENTS LIST FIGURE 36-37

RESISTORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	VALUE	TYPE	REF.	VALUE	TYPE	REF.	VALUE	TYPE
R1	0.3 Amp. Dropper, see text		R12	47Ω 5w	A1/1	R24	680Ω 10%	BTS
			R13	47Ω	BWF	R25	1MΩ	BTS
R2	47Ω 5w	A1/1	R14	22kΩ 1w	BTB	R26	47kΩ	BTS
R3	47Ω 5w	A1/1	R15	100kΩ	BTS	R27	220kΩ	BTS
R4	47Ω 5w	A1/1	R16	220kΩ 10%	BTS	R28	1MΩ	BTS
R5	47Ω	BWF	R17	220kΩ 10%	BTS	R29	220kΩ	BTS
R6	47Ω	BWF	R18	100kΩ	BTS	R30	2.7kΩ 10%	BTS
R7	2.7kΩ 5w	A1/1	R19	100kΩ 5%	R425	R31	1MΩ	BTS
R8	100kΩ	BTS	R20	100kΩ 5%	R425	R32	1MΩ	BTS
R9	4.7kΩ 5w	A1/1	R21	68kΩ 10%	BTS	R33	2.7kΩ 10%	BTS
R10	100kΩ	BTS	R22	47kΩ	BTS	VR1	500kΩ	C
R11	47Ω	BWF	R23	1MΩ	BTS	VR2	500kΩ	C
						VR3	500kΩ	C

CAPACITORS. THE DUBILIER CONDENSER CO. (1925) LTD.

REF.	VALUE	TYPE	No.	REF.	VALUE	TYPE	No.
C1	50μF 280v	Electrolytic	CT	C11	0.01μF 350v	Paper	Met. Min.
C2	100μF 280v	Electrolytic	CT	C12	0.01μF 350v	Paper	Met. Min.
C3	0.01μF 1000v	Paper	460	C13	0.05μF 350v	Paper	Met. Min.
C4	8μF 150v	Electrolytic	BR	C14	0.5μF 150v	Paper	410
C5	50μF 12v	Electrolytic	MBR	C15	25μF 12v	Electrolytic	MBR
C6	0.05μF 350v	Paper	460	C16	0.01μF 1000v	Paper	460
C7	0.05μF 350v	Paper	460	C17	0.01μF 1000v	Paper	460
C8	0.005μF 1000v	Paper	460	C18	0.05μF 350v	Paper	Met. Min
C9	32μF 350v	Electrolytic	BR	C19	0.01μF 1000v	Paper	460
C10a-b	32-32μF 250v	Electrolytic	BR	C20	0.01μF 1000v	Paper	460
				C21	25μF 12v	Electrolytic	MBR

VALVES. STANDARD TELEPHONES & CABLES LTD. (BRIMAR).

V1	...	6J7GT	V4	...	25L6GT	V7	...	25L6GT
V2	...	6J5GT	V5	...	25L6GT	V8	...	25Z4G
V3	...	6SL7GT	V6	...	25L6GT	V9	...	25Z4G
						V10	...	25Z4G

VALVE-HOLDERS. McMURDO INSTRUMENT CO. LTD.

10 ... ... Int. Octal ... Type B8/U

OUTPUT TRANSFORMER.

Primary ... 2750Ω load. Sec. ... to suit speech coil.

CHOKE

L1 ... 5 Henry 250mA

SUNDRIES.

2 input jacks Bulgin J6 Plugs as required Bulgin P38
S1 D.P.S.T. Toggle switch, Bulgin.
TH1 Thermistor, type CZ1 Brimar.

Chassis 14in. x 10in. x 3in.
Output sockets, tag-strips, etc.

distortion has been kept to a very low figure—less than 0.4 per cent. at 10 watts output. The frequency response is extremely wide and level, being almost flat from 10 to 20,000 cycles per second.

The circuit of the amplifier is of conventional form. A single-ended high-gain pentode (EF86) feeds a cathode-coupled phase-splitter using the high-mu double triode ECC83. The balanced output voltages derived from the ECC83 are used to drive the grids of two EL84 pentodes in push-pull. Negative voltage feedback is applied from the secondary of the output transformer to the cathode of the input valve.

The rated output of the amplifier is 10 watts, the maximum power is 12-13 watts.

Loop Gain and Frequency Response

Fig. 39 shows loop gain, overall frequency response, and phase shift. Relative to 1,000 c/s the response is not down by more than ½ db at the two extremes of 10 c/s and 20,000 c/s. Overall feedback of 26 db is taken from the secondary of the output transformer.

Power Response

Fig. 40 shows the maximum output power relative to 10W for frequencies between 15 and 30,000 c/s. The measurements were made for a sine wave drive and with the output stage operated under its normal loading conditions (anode-to anode load of 8,000).

From 40–10,000 c/s, max. output is 1db relative to 10W.
„ 20–16,000 c/s max. output is 0db (10W.)
„ 16–30,000 c/s, max. output is −2db relative to 10W.

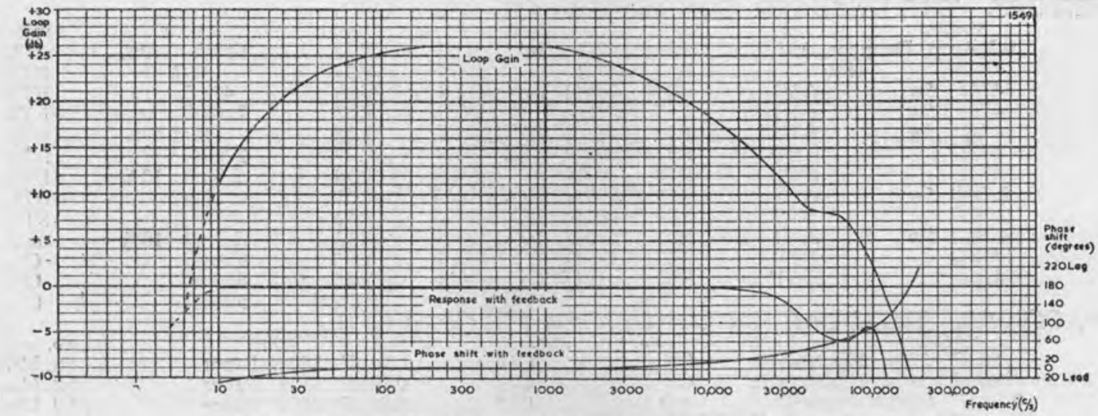
Distortion

Total harmonic distortion has been measured at 40 c/s, 400 c/s, and 2,000 c/s (Fig. 42a). For the rated output of 10W the total distortion is:

Less than 0.4% at 40 c/s
„ „ 0.2% „ 400 c/s
„ „ 0.3% „ 2,000 c/s

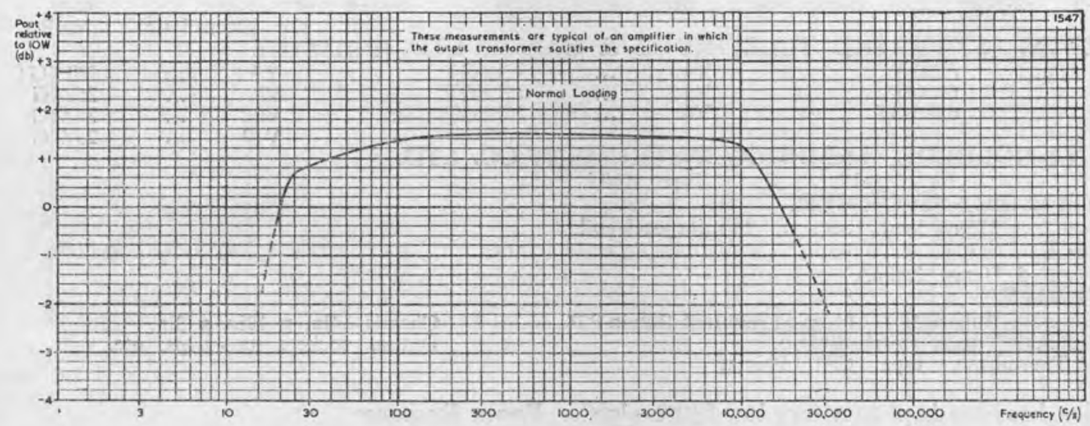
Hum and Noise

With the ear close to the loudspeaker no hum can be detected and residual noise is only a slight rustle. Under normal listening conditions hum



TYPICAL PHASE SHIFT AND RESPONSE CHARACTERISTICS OF COMPLETE AMPLIFIER.

FIG. 39.



MAXIMUM OUTPUT POWER OF AMPLIFIER PLOTTED AGAINST FREQUENCY.

FIG. 40.

and noise are completely inaudible, being some 73 db below 10W, or 74 db below the maximum rated output of 12.5W.

Output Resistance

The output resistance is 0.9 Ω on a 15 Ω output and is sufficiently small in practice to ensure adequate electrical damping of the speaker coil.

Tone Control

The tone control unit provides 10 db boost in treble and bass, 10 db attenuation in treble, and 5 db attenuation in bass (Fig. 41).

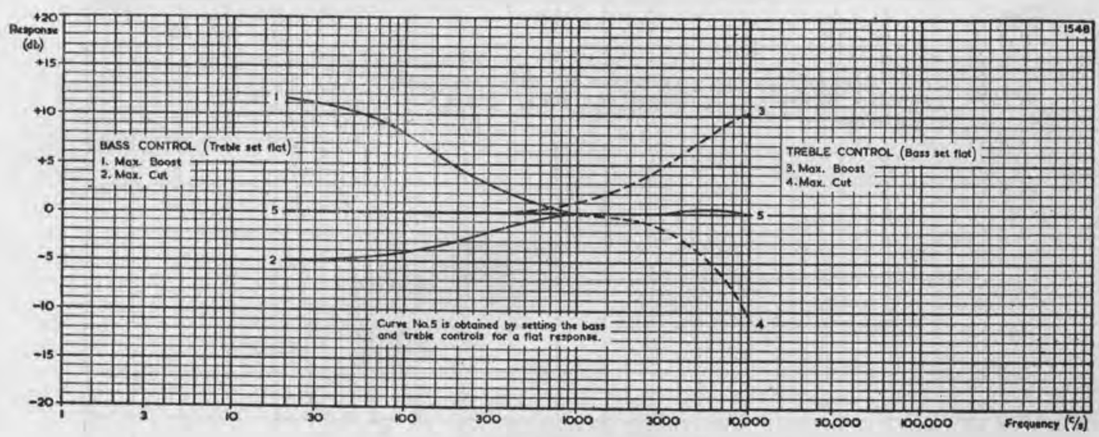
Measurements for curves No. 1 and 2 have been made with the treble control set flat and show the effect of varying the bass control. The reverse procedure was adopted for curves No. 3 and 4. In curve No. 5 a flat response has been obtained by a suitable setting of the treble (RV23) and bass (RV24) controls.

The range of control should be sufficient to compensate for widely differing listening conditions.

This tone control produces an attenuation of about 12 times. Because of the high sensitivity of the amplifier—50 mV at Y–Y with feed-back—the tone control unit is suitable for use with a crystal pick-up having a relatively large output, without the need arising for a separate valve pre-amplifier. The input voltage at X–X must be approximately 600 mV to load the amplifier fully.

Pick-ups and Equalising Networks

The Collaro 'O' and 'P' "Studio" pick-up heads and the Acos Hi-g microgroove and standard pick-ups are particularly suitable for use with



FREQUENCY RESPONSE OF TONE CONTROL CIRCUIT

FIG. 41.

the amplifier. Details of some equalising networks which have been found very satisfactory are given in Fig. 43. They are designed to match into the input impedance of the tone control circuit, and were derived using the Decca K1804A (78 r.p.m.) and Decca LXT 2695 (33 1/3 r.p.m.) recordings.

First Amplifying Stage

The first stage of the amplification is provided by the EF86 in a circuit having a gain of approximately 150 times. The negative feedback voltage from the secondary of the output transformer is introduced across the 100 Ω resistor, R5, in the cathode circuit. In a feedback amplifier with a wide frequency response, stability can be achieved only if the required difference in phase is maintained between the input signal and the feedback voltage. The EF86 has accordingly been coupled directly to the following stage in order to reduce the phase shift at low frequencies. The C-R network (C13, R22) shunting the anode load produces an advance in phase, which increases the stability of the amplifier at high frequencies.

Phase Splitter

The output stage is fed by an ECC83 double triode operated as a cathode coupled phase splitter. The two grids are coupled together by R8, the second being capacitively earthed by C7. The cathodes of the ECC83 are biased to have the same voltage as the anode of the EF86 (= 70V). Anode resistors R9 and R10 (=100 kΩ) should be matched within 5%, R10 being given the larger value.

The use of the cathode coupled circuit provides for low distortion and facilitates direct coupling to the first stage. The gain obtained with the cathode coupled circuit is about half that obtained from each valve section operated as a normal voltage amplifier. Nevertheless it is sufficient as the ECC83 has an amplification factor of 100.

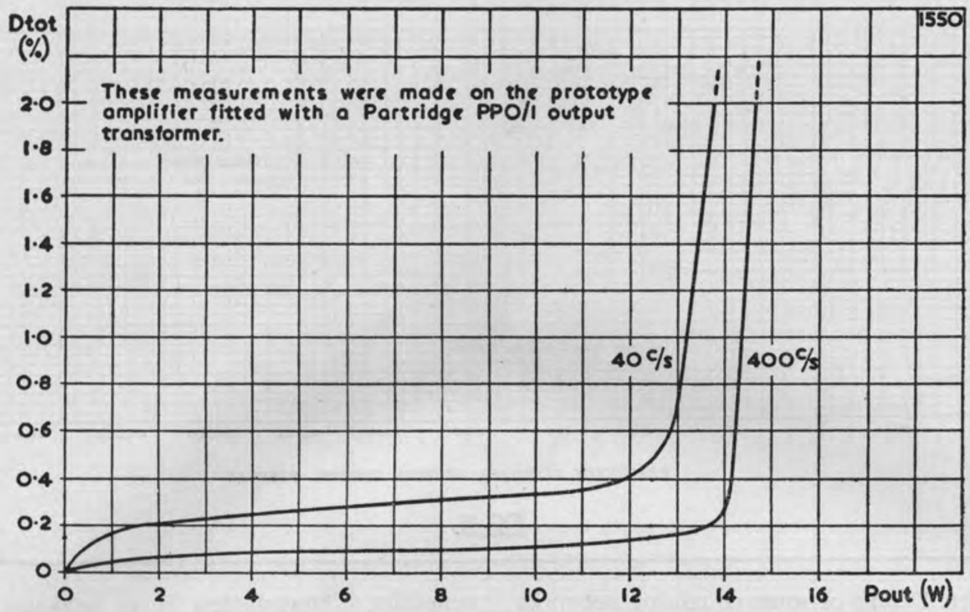
Output Stage

The output stage is equipped with two EL84 output pentodes operated in a self-biased push-pull circuit. The anodes are fed from the reservoir capacitor C1, the screen grids and the rest of the amplifier being supplied via R18 and C2. Separate bias resistors R16, R17 are used. Stopper resistors (R14, R15, R19, R20) are included in the control- and screen-grid leads.

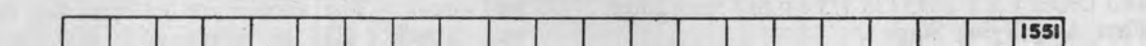
A resistor with a value of about 1k Ω may be placed across the output terminals to prevent instability from occurring with a disconnected loudspeaker.

Operating Conditions

Alternative modes of operating two EL84's in a push-pull output stage can be compared by referring to the curves showing total harmonic distortion plotted against output power (Fig. 42b). Curve No. 2 is plotted for cathode bias and refers to a sine wave input. It is obtained under the conditions given for Class AB operation in the valve data; these conditions are necessary for testing the amplifier with a sine wave input up to full power. They can be referred to as the 'normal loading' conditions, the anode-to-anode impedance being 8000 Ω and the quiescent anode current 2×36 mA. With speech and music inputs, however, the output stage operates with approximately fixed bias. As a result, when the normal loading conditions are used for speech and music (Curve No. 1), the distortion above 10W is considerably greater than might be expected from the data. It can be seen that above 12.5W the distortion for speech and music inputs would be nearly twice that obtained for a sine wave input.

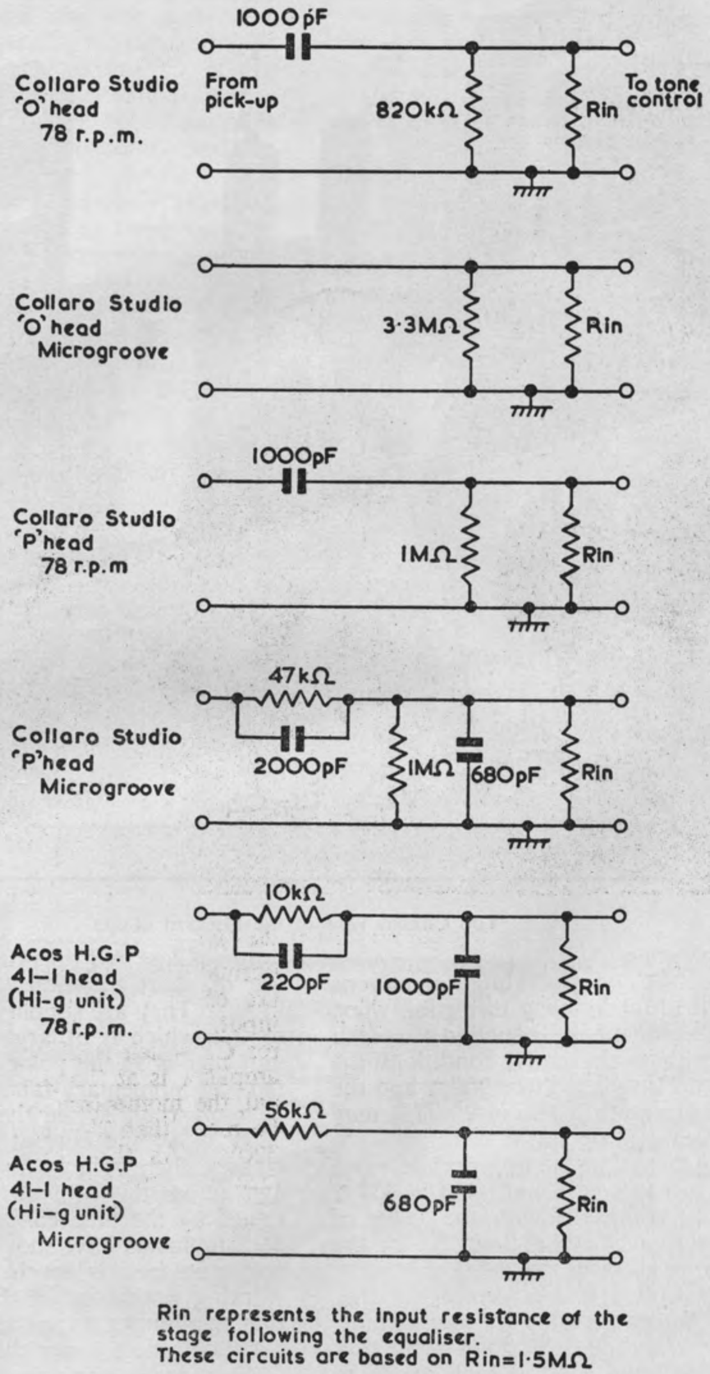


a. TOTAL HARMONIC DISTORTION PLOTTED AGAINST OUTPUT POWER FOR THE COMPLETE AMPLIFIER.



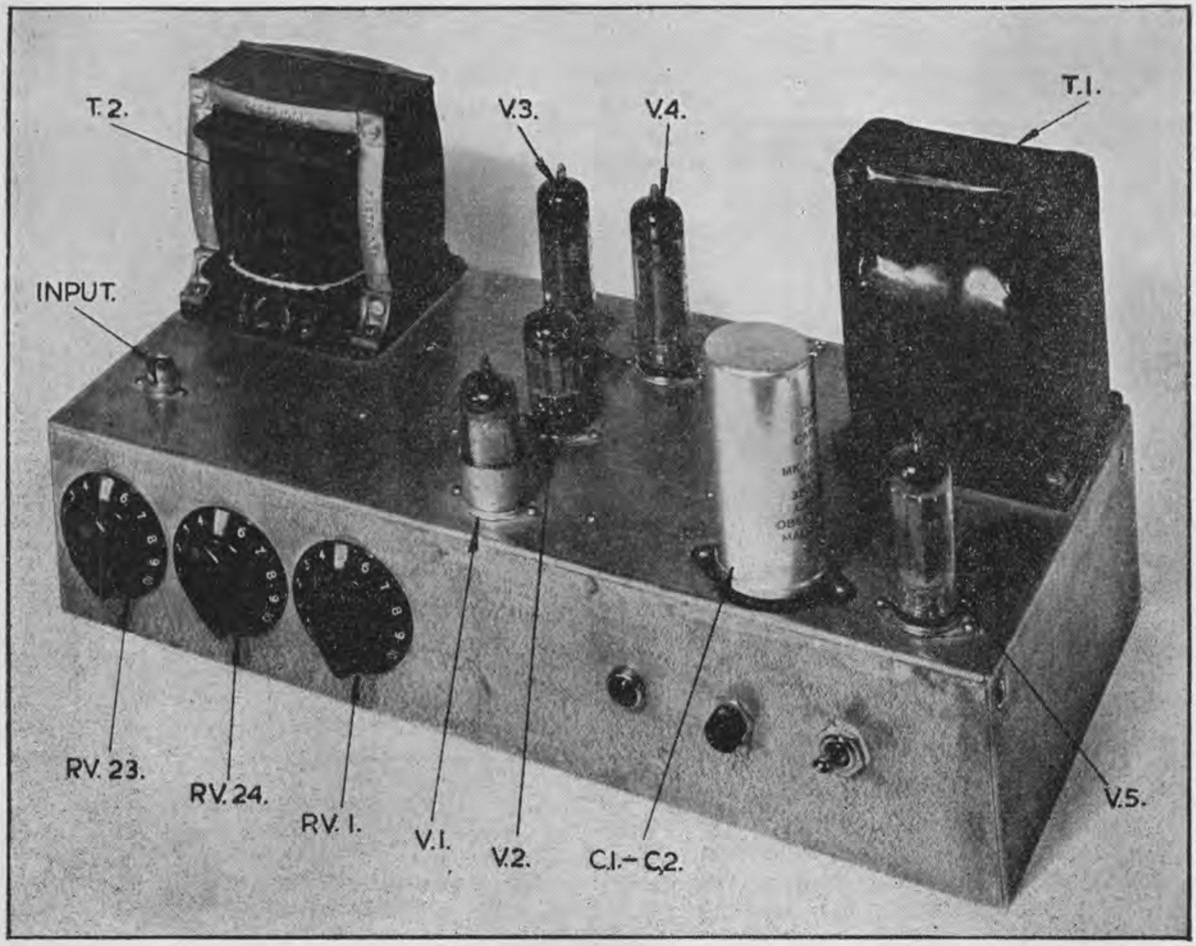
b. TOTAL HARMONIC DISTORTION PLOTTED AGAINST OUTPUT POWER FOR TWO EL84's OPERATED IN PUSH-PULL, WITH ANODE AND SCREEN GRID VOLTAGES OF 300 V.

FIG. 42.



EQUALISING NETWORKS SUITABLE FOR USE WITH TONE CONTROL UNIT.

FIG. 43.



Top Chassis view of the Mullard design.

Low Loading Operation

An alternative set of operating conditions (Curve No. 3) will result in lower distortion when the amplifier is used for the reproduction of speech and music. Under these alternative conditions, the anode-to-anode load is reduced to 6,000 Ω and the quiescent anode current to 2 × 24 mA. This may be termed 'low loading' operation.

For low loading operating the appropriate value of both cathode resistors R16 and R17 is 437 Ω (=390 Ω + 47 Ω, as compared with the value of 270 Ω each for normal loading (that is, for the Class AB conditions given in the data).

The H.T. consumption is considerably smaller when the output stage is adjusted for low loading. In consequence the standing dissipation in the output stage is reduced from 11W at each anode to 7.5W at each anode, the output valves then being run well below their maximum permissible anode dissipation of 12W. There will also be less ripple on the H.T. line.

Peak Handling Capacity

Larger peak currents are produced in the output stage under low loading conditions than with normal Class AB operation. These peak currents are of short duration with a speech and music input. They are supplied by the reservoir capacitor C1 which is of large value (50μF). When the amplifier is at the point of overload on peak signal, the momentary fall in line voltage should not be more than 2V on the nominal line voltage of 320V. As the current in the output stage increases, there follows an increase in the bias voltage across the cathode resistors at a rate determined by the time constant of the bias networks. Measurements have shown that in practice this increase in bias is not likely to exceed 1V. The working conditions of the output stage are such that the output valves are then driven back into a region where lower distortion is obtained.

As a result, however, of any change in the bias of the output stage, a variation in gain will occur; but the distortion which is introduced in this way is held to a low level by the large amount of negative feedback.

Output Transformer

The output transformer is the most important component in a feed-back amplifier, and it is essential that it shall give adequate performance. It is therefore advisable to obtain the output transformer from a manufacturer who has undertaken to build this component specially for this amplifier. For this reason, a detailed specification of the transformer is provided. It is essential that a component meeting the minimum specification be used, otherwise there will be instability and deterioration in performance.

Of the output transformers currently available, the Partridge PPO was selected as being suitable. The distortion curves (Fig. 42) were obtained with the prototype amplifier fitted with this output transformer.

Messrs. Partridge have now issued Type P3650 which conforms to the designers specification and should be used in preference to Type P.P.O.

Rectifier

The GZ30 full-wave rectifier can supply a current drain of 125 mA and is completely suitable for all applications since it has sufficient current reserve to supply an F.M., or other radio unit in conjunction with the amplifier.

Construction

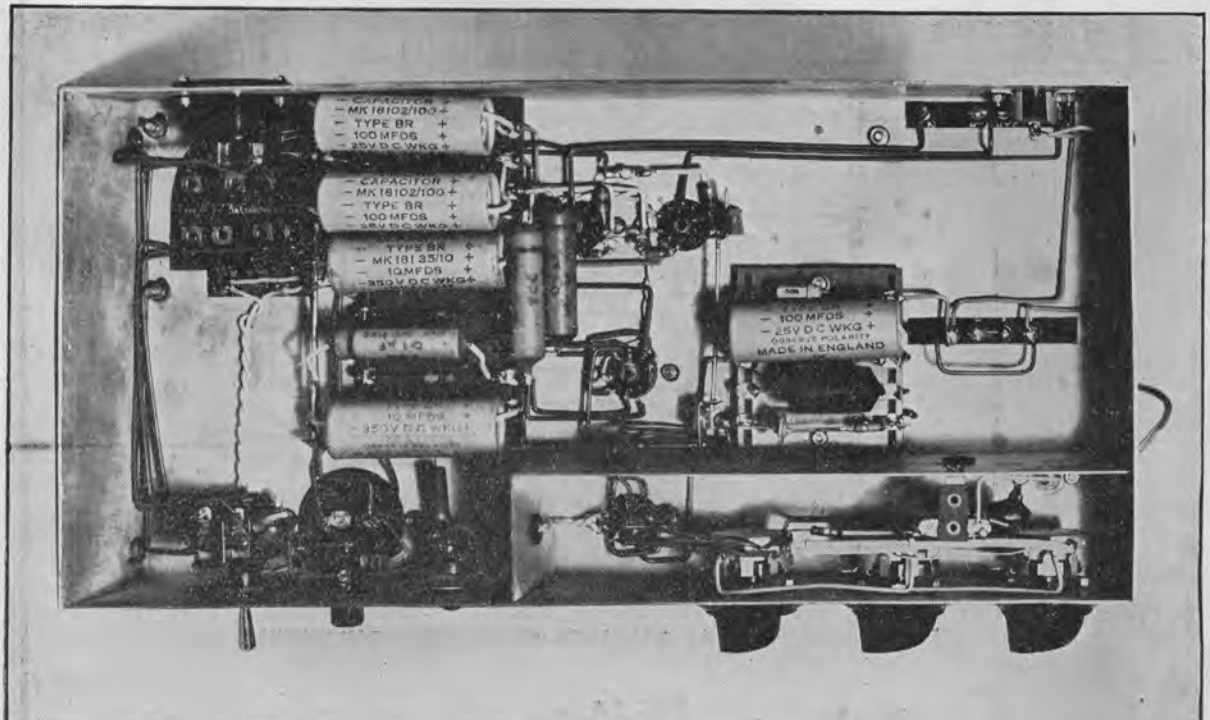
The layout used on the original design can be easily followed from the two photographs. Note that V.1 (EF86) together with the volume and tone control networks, input socket, etc., is screened off from the remainder of the circuit. This screen can be bent into shape from a strip of aluminium 3″ wide. V.1 anode resistors R2 and R22 together with C13 can be seen positioned immediately behind the screen, on the tag board.

V.1 uses a screened valve holder, the can has been removel for photographic purposes. The reservoir and smoothing condensers are contained in a dual electrolytic capacitor, "ear" mounting can type by Dubilier, the remaining electrolytics are all Dubilier B.R. series. Input from the pick-up is by way of a Belling-Lee concentric plug and socket, which is both neat and effective. Output and mains sockets, fuse holders and other sundries are a matter of personal preference. Chassis size is 14″ × 7″ × 3″.

CIRCUIT VOLTAGES

TESTING Point	*VOLTAGE (V)* D.C.	A.C.	*METER RANGE*
C1	320	Normal loading 4 Low loading 2.5	1000 V
C2	310		1000 V
Cathodes V3, V4	12		100 V
Anodes V3, V4	310		1000 V
Screen grids V3, V4	312		1000 V
C3	255		1000 V
Anodes V2	210		1000 V
Cathodes V2	70		1000 V
C4	182		1000 V
Anode V1	70		1000 V
Screen grid V1	65		1000 V
Cathode V1	1.5		25 V

These voltages were measured with Model 8 Avometer (20,000 Ω/V) with zero input signal.



Under Chassis, showing wiring and layout.

SPECIFICATION FOR OUTPUT TRANSFORMER (TR2)

Matching

8000 Ω or 6000 Ω primary to 15 Ω or 3.75 Ω secondary.

Note: In winding table (A) refers to 8000 Ω primary, (B) to 6000 Ω primary.

Winding Data

Core: Square stack of 1 1/8-inch centre limb, E and I laminations, "NO WASTE" series.

Pattern No. Sankey No. 158, proposed British Standards No. 29.

Core Material: 4% Si steel, 0.014″ laminations.

Core Size (External): 3.375 × 2.8 × 1.125 inches. Bobbin to suit.

Insulation: Between layers of primary sections: 0.001″ paper. Between layers of secondary sections: 0.003″ presspahn. Between various sections: 1 layer of 0.003″—0.004″ presspahn, and 1 layer of 0.002″ paper.

If P1 and P2 are wound clockwise then all other windings are wound anticlockwise.

The following windings are connected in parallel: P1 and P4 to form one half of primary; P2 and P3 to form other half of primary.

S1 and S2 are brought out separately for series (15 Ω) or parallel (3.75 Ω) connection.

Electrical Characteristics

Primary shunt inductance (8000 Ω) 40H at 10V 50 c/s.

Primary leakage inductance 25mH

Primary d.c. resistance 240 Ω per half primary (8000 Ω)

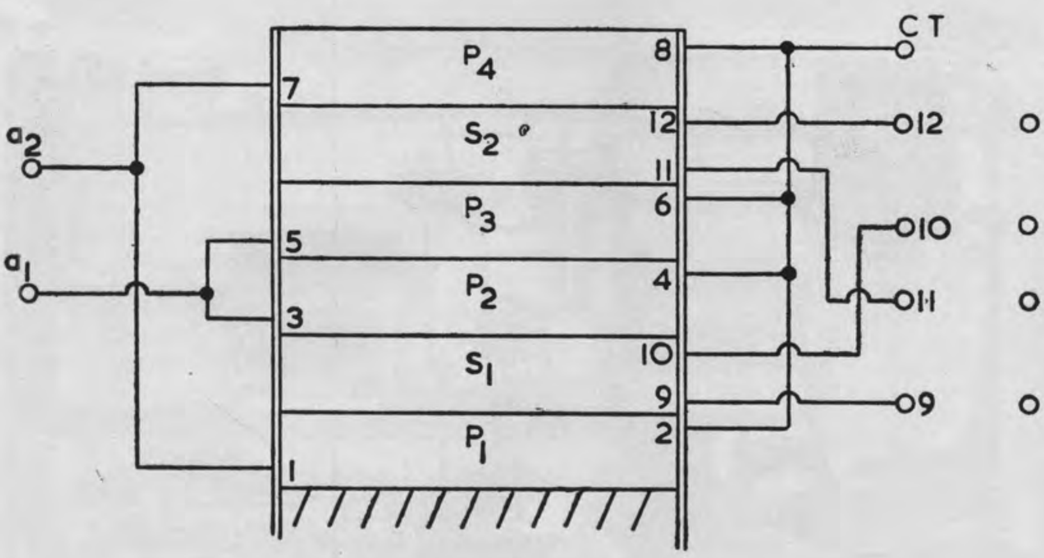
Secondary resistance S1 approx. 0.5 Ω

S2 approx. 0.5 Ω

WINDING TABLE

Winding	No. of Turns	No. of Layers	Turns per layer	Wire SWG
P1	1650 (A)	7	236 (A)	40
	1450 (B)	7	207 (B)	40
S1	76	2	38	22
P2	1650 (A)	7	236 (A)	40
	1450 (B)	7	207 (B)	40
P3	1650 (A)	7	236 (A)	40
	1450 (B)	7	207 (B)	40
S2	76	2	38	22
P4	1650 (A)	7	236 (A)	40
	1450 (B)	7	207 (B)	40

Wire: Enamelled Copper. Width of Windings: 1.35 inches.



INTERCONNECTION OF WINDINGS ON OUTPUT TRANSFORMER.

FIG.44.

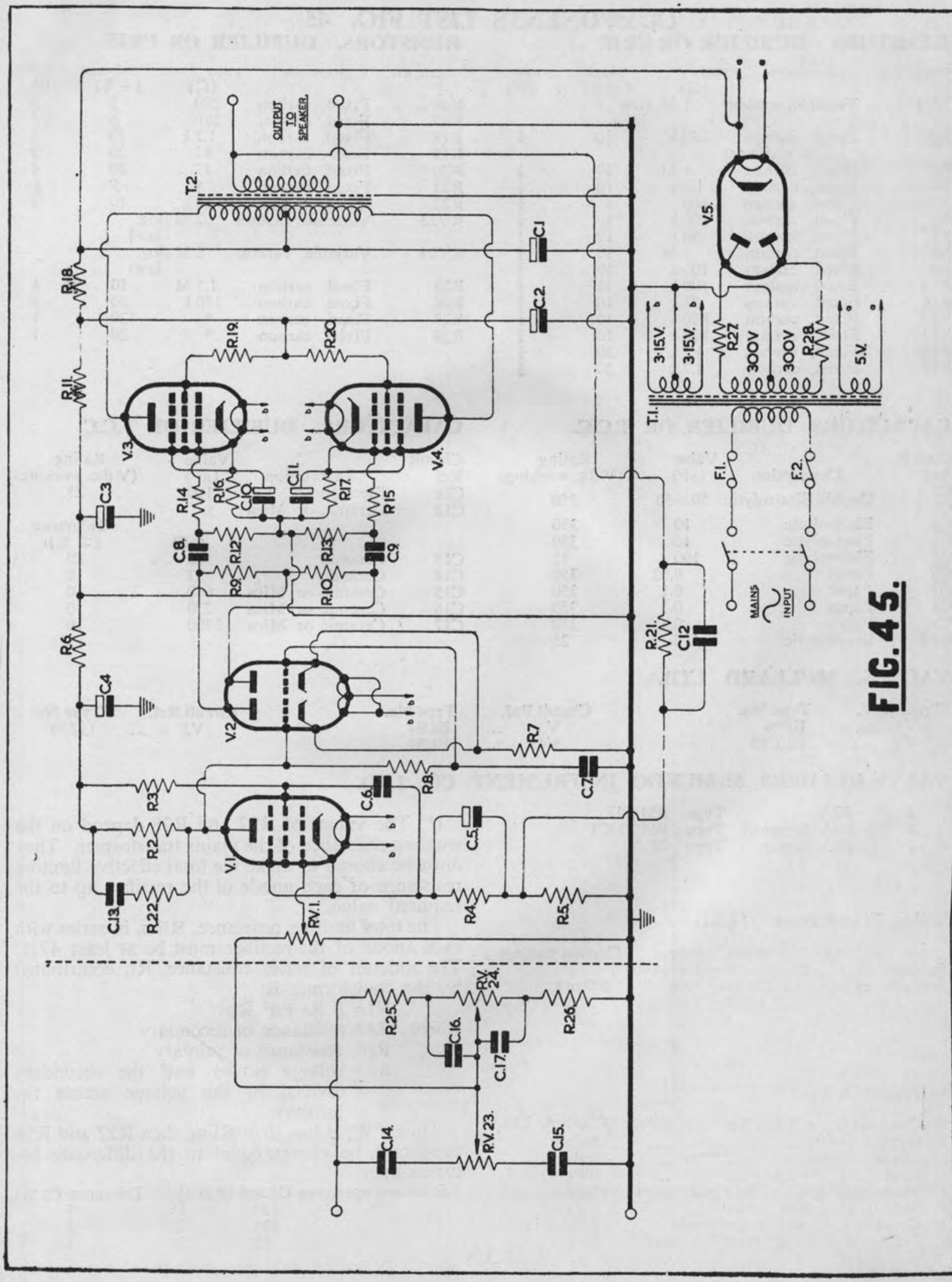

OUTPUT TO SPEAKER
T.2
T.1
V.1
V.2
V.3
V.4
V.5
C.1
C.2
C.3
C.4
C.5
C.6
C.7
C.8
C.9
C.10
C.11
C.12
C.13
C.14
C.15
C.16
C.17
R.2
R.3
R.4
R.5
R.6
R.7
R.8
R.9
R.10
R.11
R.12
R.13
R.14
R.15
R.16
R.17
R.18
R.19
R.20
R.21
R.22
R.25
R.26
R.27
R.28
RV.1
RV.23
RV.24
3·15V
3·15V
300V
300V
5V
F.1
F.2
MAINS INPUT

FIG. 4.5.

COMPONENTS LIST FIG. 45

RESISTORS DUBILIER OR ERIE

Circuit Ref.	Description	Value (Ω)	Tolerance (+%)	Rating (W)
RV1	Variable, carbon	1 M (log law)		
R2	Fixed, carbon (high stability)	180 k	10	1
R3	Fixed, carbon	1 M	10	¼
R4	Fixed, carbon	1.8 k	10	¼
R5	Fixed, carbon	100	5	¼
R6	Fixed, carbon	100 k	10	¼
R7	Fixed, carbon	68 k	10	1
R8	Fixed, carbon	1 M	10	¼
R9	Fixed, carbon	100 k	10	½
R10	Fixed, carbon	100 k	10	½
R11	Fixed, carbon	33 k	10	¼
R12	Fixed, carbon	820 k	10	¼
R13	Fixed, carbon	820 k	10	¼
R14	Fixed, carbon	4.7 k	20	¼
R15	Fixed, carbon	4.7 k	20	¼

RESISTORS. DUBILIER OR ERIE

Circuit Ref.	Description	Value (Ω)	Tolerance (+%)	Rating (W)
R16	Fixed, carbon	270	5	3
R17	Fixed, carbon	270	5	3
R18	Fixed, carbon	1.2 k	10	1
R19	Fixed, carbon	47	20	¼
R20	Fixed, carbon	47	20	¼
R21	Fixed, carbon	†	5	¼
R22	Fixed, carbon	18 k	10	¼
RV23	Variable, carbon	2 M (log. law)		
RV24	Variable, carbon	2 M (log. law)		
R25	Fixed, carbon	1.5 M	10	¼
R26	Fixed, carbon	150 k	10	¼
R27	Fixed, carbon	*	20	1
R28	Fixed, carbon	*	20	1

CAPACITORS DUBILIER OR T.C.C.

Circuit Ref	Description	Value (μF)	Rating (V d.c. working)
C1 } C2 }	Double Electrolytic	50+50	350
C3	Electrolytic	10	350
C4	Electrolytic	10	350
C5	Electrolytic	100	12
C6	Paper	0.02	350
C7	Paper	0.1	350
C8	Paper	0.1	350
C9	Paper	0.1	350
C10	Electrolytic	100	25

CAPACITORS. DUBILIER OR T.C.C.

Circuit Ref	Description	Value (μF)	Rating (V d.c. working)
C11	Electrolytic	100	25
C12	Ceramic or Mica	†	
		(pF)	Tolerance (± %)
C13	Ceramic	100	20
C14	Ceramic	33	10
C15	Ceramic or Mica	680	10
C16	Ceramic or Mica	270	10
C17	Ceramic or Mica	3300	10

VALVES. MULLARD LTD.

Circuit Ref.	Type No.	Circuit Ref.	Type No.	Circuit Ref.	Type No.
V1 ...	EF86	V3 ...	EL84	V5 ...	GZ30
V2 ...	ECC83	V4 ...	EL84		

VALVE-HOLDERS McMURDO INSTRUMENT CO. LTD.

4	B9A	Type	BM9/U
1	B9A Screened	Type	BM9/UC1
1	B9A Screen	Type	75

Mains Transformer. (TR1)

	Voltage Tappings	Current Ratings
Primary :	10—0—200—220—240	
Secondaries :	300—0—300v.	100 mA
	3.15—0—3.15v.	2 A
	0—5v.	2 A

SUNDRIES

1	Chassis 14in. x 7in. x 3in.	Denco (Clacton) Ltd.
1	D.P.S.T. switch	Bulgin
2	Single fuse-holders	Bulgin
1	Pilot lamp assembly	Bulgin
1	Mains plug and socket	Bulgin
1	Output plug and socket ...	Bulgin
1	Co-axial plug and socket (input)	Belling-Lee
	Tag strips and tag mounting boards	Bulgin

(*) The values of R27 and R28 depend on the winding resistance of the mains transformer. They must be chosen to make the total effective limiting resistance of each anode of the rectifier up to the required value.

The **total** limiting resistance, Rlim, in series with each anode of the rectifier must be at least 47 Ω. The amount of series resistance, Rt, contributed by the transformer is:

$$Rt = \tfrac{1}{2}\ Rs + n^2\ Rp$$

where Rs = resistance of secondary
Rp = resistance of primary
n = voltage across **half** the secondary divided by the voltage across the primary.

Thus if R2 is less than Rlim, then R27 and R28 must both be chosen equal to the difference between them.

† Speaker Impedance	C12 (pF)	R21 (kΩ)	Tolerance (±%)
3.75	180	15	5
7	120	22	5
15	82	33	5

BERNARDS RADIO BOOKS

No.		
35.	Dictionary of Mathematical Data	2/-
61.	Amateur Transmitter's Construction Manual	2/6
64.	Sound Equipment Manual	2/6
65.	Radio Designs Manual	2/6
66.	Communications Receivers' Manual	2/6
68.	Frequency Modulation Receivers' Manual	2/6
69.	Radio Inductance Manual	2/6
70.	Loudspeaker Manual	2/6
73.	Radio Test Equipment Manual	2/6
78.	Radio and Television Laboratory Manual	2/6
81.	Using Ex-Service Apparatus	2/6
83.	Radio Instruments and their Construction	2/6
86.	Midget Radio Construction	3/6
96.	Crystal Set Construction	1/-
97.	Practical Radio for Beginners, Book 1	3/-
99.	One Valve Receivers	1/6
100.	A Comprehensive Radio Valve Guide, Book 1	5/-
101.	Two Valve Receivers	1/6
102.	40 Circuits Using Germanium Diodes	3/-
103.	"Radiofolder" A. The Master Colour Code Index for Radio and Television	1/6
104.	Three Valve Receivers	1/6
106.	Radio Circuits Handbook No. 4	2/6
107.	Four Valve Circuits	1/6
108.	Five Valve Circuits	2/6
110.	The Emperor: A Radio-gram for Home Constructor	3/6
112.	"Radiochart" Electronic Multimeter Construction	2/6
115.	Constructors' Handbook of Germanium Circuits	2/6
120.	Radio and Television Pocket Book	2/6
121.	A Comprehensive Radio Valve Guide, Book 2	5/-
122.	Wide Angle Conversion: For Home Constructed Televisors; Constructional Envelope	3/6
123.	"Radiofolder" F. The Beginners' Push-Pull Amplifier	1/6
124.	"At a Glance" Valve and Valve Equivalents for Radio and Television	5/-
125.	Listeners' Guide to Radio and Television Stations of the World	2/6
126.	The Boys' Book of Crystal Sets	2/6
127.	Wireless Amplifier Manual No. 3	3/6
128.	Practical Transistors and Transistor Circuits	3/6
129.	Universal Gram-Motor Speed Indicator	1/-
130.	Practical F.M. Circuits for the Home Constructor	5/-
131.	"Radiochart" Guide to Modern Valve Bases	2/6
132.	Reactance—Frequency Chart for Designers and Constructors	1/6
133.	Radio Controlled Models for Amateurs	5/-
134	F.M. Tuner Construction	2/6
135.	All Dry Battery Portable Construction	2/6
137.	An A.M.-F.M. De-Luxe Receiver	3/6
138.	How to make T.V. and F.M. Aerials Bands 1, 2 and 3	2/6
139.	Practical Radio for Beginners, Book 2	3/6
140.	Television Servicing for Beginners, Book 1	4/6
141.	Radio Servicing for Amateurs	3/6
142.	Modern Television Circuits and Fault Finding Guide	4/6
143.	A Comprehensive Valve Guide, Book 3	5/-
144.	The New "At a Glance" Valve and Television Tube Equivalents	5/-
	Servicing the Modern Radio Receiver	1/6
	A Magnetic Tape Recorder	3/6
	The Electronic Photographic Speedlamp: How to Make it and how to Use it	3/6