AN APPROACH TO

AUDIO FREQUENCY AMPLIFIER

DESIGN

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A Publication of

The G.E.C. Valve and Electronics Department

AN APPROACH TO AUDIO FREQUENCY AMPLIFIER DESIGN

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PREFACE

This book has been prepared under the auspices of the G.E.C. Valve and Electronics Department with the aim of presenting, in a readily accessible form, the essential details and circuit diagrams of a comprehensive range of audio-frequency amplifiers. All the designs reflect various aspects of modern technique and have been built and tested. In this field, the experience and accumulated knowledge of the G.E.C., dating back to the earliest days of radio communication and large power valve design, are unique.

Altogether seventeen amplifiers are described, ranging from a "junior" 5-watt ultra-linear circuit to an 1100-watt Class AB2 amplifier. In addition, various pre-amplifier units are discussed so that the information in the following chapters deals with all the equipment necessary between the signal source and the loudspeaker in most domestic, public address and industrial sound installations.

Since the book is essentially a work of reference it contains no constructional information—neither does it claim to be a text book, although as an exposition of modern design practice, it will be of considerable value in this respect.

An Approach to Audio Frequency Amplifier Design is the culmination of considerable design and development work on a.f. amplifiers carried out by W. I. Heath, B.Sc. (Eng.) and D. M. Leakey, B.Sc. (Eng.) in the Research Laboratories of the General Electric Co. Ltd. and by G. R. Woodville in the Applications Laboratory of the M-O. Valve Co. Ltd., a subsidiary of the G.E.C.

LONDON,
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C. E. KNIGHT-CLARKE,
Editor

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CHAPTER 1

THE DESIGN OF AN AUDIO FREQUENCY AMPLIFIER

The design of an audio frequency amplifier may conveniently be divided into four parts, which are to some extent inter-related. The output stage ; the input stage ; the intermediate stages, and the power supply.

The purpose for which the equipment is required naturally determines the type of output stage required. Public address and industrial amplifiers are seldom required to provide a power output of less than 25 watts and may require to supply as much as 1000 watts, whereas amplifiers for domestic use do not often exceed 25 watts and are more usually of 10-15 watts output. The signal source will determine the type of input stage required ; obviously, a tape recorder will require a higher sensitivity than a radio receiver. The design of the intermediate stages, including any tone controls will, of course, depend on the other two sections. Finally, there is the power supply, which may consist of one or more units.

THE OUTPUT STAGE

1-1. Major Design Features.—The usé to which the amplifier is to be put will decide the four major design features of the output stage, which are :

1. Maximum power output.

2. Frequency range.

3. Maximum distortion.

4. Output impedance.

Amplifiers for domestic use normally have a relatively low output of from 5W to 25W and hence the efficiency, that is, the power output for a given anode dissipation, is not important and often a pair of triodes in Class A push-pull is used. The frequency range should be as wide as possible, say 30 to 25,000c/s or at least equal to that of the associated loudspeaker, record player, etc. The maximum distortion should be low, less than 0·5% for the best equipment rising to perhaps 2% for cheaper amplifiers. The output impedance should be low in order to offset one of the defects inherent in nearly all loudspeakers.

The four points take on a different emphasis when an industrial or public address amplifier is being designed. The power output required is much higher; up to about 1000W, and hence it is usual to design for maximum efficiency, since an efficient amplifier will be cheaper, smaller and will run cooler than an inefficient one. Class A amplifiers are not used and, for the higher powers, the choice is between Class AB and Class B triodes. The acceptable frequency range is usually narrower; a lower limit of 100c/s for music and 300c/s for speech with an upper limit of 7-10kc/s. The maximum distortion is not very important and usually 5% or 6% is not objectionable. The output impedance should be low to prevent an undue rise in output voltage with a change of load impedance.

1-2. The Output Stage of a Domestic Amplifier.—Current design practice for amplifiers used in the home comprises a push-pull output stage either triode-connected or arranged in an " ultra-linear " (or " distributed load ") circuit. The ultra-linear circuit will normally provide twice the output of push-pull triode-connected valves for any given type and h.t. voltage with little, if any, addition to the circuit. There is little doubt that the ultra-linear amplifier is the most satisfactory for providing from 5W to 100W output power when low distortion and good frequency response are the major considerations. Compared with push-pull pentodes, the same high efficiency is obtained but with the low distortion and uncritical load impedance of triodes.

The curves in fig. 1-1 show the output and approximate distortion obtained from push-pull KT88 valves with various positions of the screen grid taps on the output transformer. At the left-hand side (0% taps) the valves are, of course, pentode-connected and on the right-hand side (100% taps) they are triode-connected. Generally speaking, the taps should be spaced at 20% to 50% of the turns on each half-primary from the centre, with 30% to 40% as the optimum. The dotted curve shows the reduction in output impedance obtained.

This curve is representative of the behaviour of most types of output pentodes although the actual distortion and power output will vary with different samples and types. Badly matched valves will give a distortion up to twice that shown in fig. 1-1. Information on valve matching is given on page 7.

Low distortion is obtained not only by correct operation of the output valves but also by the application of negative feedback, usually from the output transformer secondary to one of the early stages. It is essential to use a high quality output transformer; a cheap or improperly designed component will result not only in higher distortion or lower output, but almost certainly in oscillation which will be often at an ultrasonic frequency and, consequently, not easy to detect without suitable test gear. In such cases it is usually possible to obtain stable operation with a sacrifice of quality by restricting the frequency range within the feedback path. Information on improving the stability of feedback amplifiers is given in Appendix B (page 119).
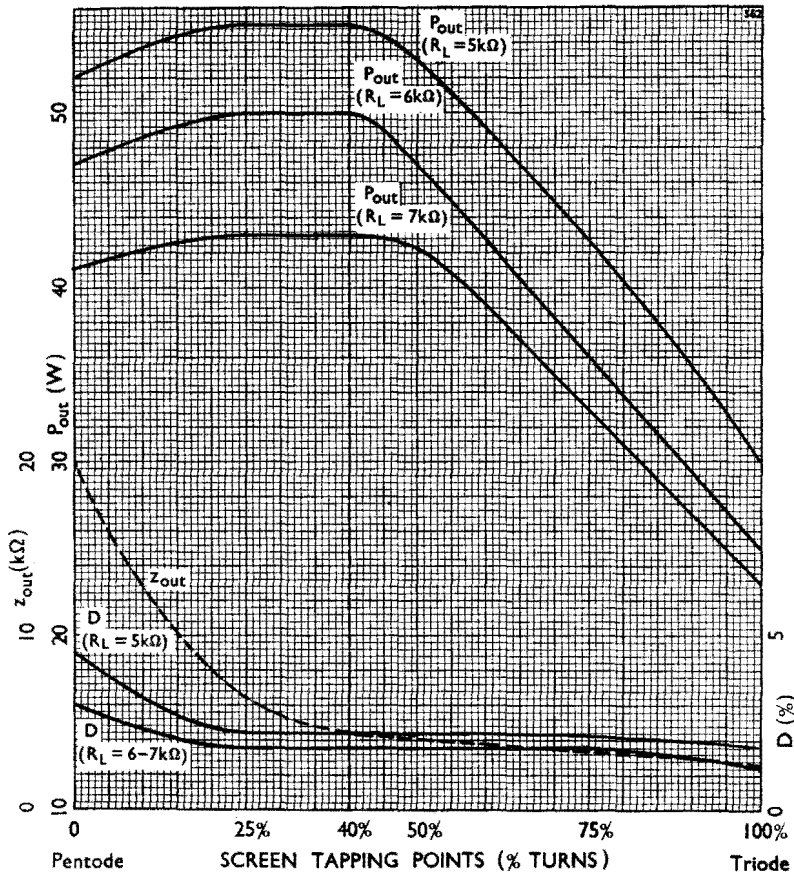
Fig. 1-1. Output power, distortion and output impedance of a KT88 ultra-linear stage at various positions of the screen taps from 0% to 100% of each half-primary from the centre tap.

This class of amplifier always employs cathode bias (auto bias) and each output valve should be provided with its own cathode resistor unless, as in the " Williamson " amplifier, provision is made for balancing the anode currents (see also Section 1-5). However, separate cathode resistors are preferable for domestic equipment as they lessen the need for additional controls capable of being wrongly set. If a common cathode resistor is used it will be necessary to operate the valves conservatively as, otherwise, with " limit " valves, one of the pair will be over-run. With either common or individual cathode biasing, the bypass capacitor(s) should not be omitted as it will be found in practice that, with bypassing, even Class A triodes will give a lower distortion because of minor differences in characteristics.

1-3.   The Output Stage of a Public Address Amplifier.—A p.a. amplifier may be conveniently classified as an equipment giving a power output from 25W to about 1000W.   The smaller designs do not differ greatly from the domestic amplifier, in that they normally use a pair of cathode biased output valves either as ultra-linear or as normally connected pentodes.   The power output of the ultra-linear circuit is not less than that of the pentode connection at any given supply voltage and the ultra-linear circuit has the advantage that a separate screen supply is not required.   Taking the KT88 as an example (Chapter 5), the ultra-linear circuit shows, in fact, a higher efficiency in that a lower total current is required from the h.t. supply.   With fixed bias and an h.t. potential of 460V, an output of 65W is obtained with an anode current of 240mA in both the ultra-linear and pentode arrangements but the pentode circuit requires, in addition, a screen current of 35mA.   The output impedance and distortion are both more favourable, the former being 6·5kΩ in this example, compared with 50kΩ, and the distortion is almost entirely independent of load impedance in the ultra-linear circuit.

From 50W to 100W the same type of valve may be used with fixed bias or in multiple pairs with cathode bias.   A given pair of valves operating with fixed bias will usually provide about twice the output obtainable with cathode bias.   With fixed bias the working point is usually chosen so that the valves operate at a reduced anode dissipation in the quiescent condition, permitting a smaller power supply transformer and a lower temperature rise.   The bias voltage that gives one half of the maximum dissipation is usually correct.

Above 100W a pair of triodes is nearly always used, the operating conditions being either Class AB1, AB2 or Class B.*   The same type of valve may be used for either Class AB1 or AB2 ; for example, the DA100, which has a low amplification factor and low impedance and requires a fixed negative grid bias of 100-250V (Chapter 7).   The principal difference between Class AB1 and AB2 is that the Class AB1 circuit operates without grid current and consequently presents a relatively high impedance to the previous stage.   In Class AB2 operation the grids are driven positive and draw considerable current.   This imposes a low and variable load on the previous stage and on the grid bias supply and both must be of low impedance.

Valves designed for Class B operation have a high amplification factor (about 70), high impedance and operate at or near zero grid bias.   Their high impedance restricts the anode current to a low value in the quiescent condition, even though the anode voltage may be a kilovolt or more.   The DA42 is an example of this type of valve.   Because of the grid current in a Class B circuit a low impedance driver stage is essential and it could be stated that the success or failure of a Class B amplifier is dependent upon the design of this stage.   The

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*An alternative and in some cases more attractive method is to use a number of smaller valves in parallel push-pull.   An example of this type of circuit is described in Appendix A (page 112).

absence of a separate bias supply renders the circuit self-protective and no safety device is necessary to prevent application of the anode voltage prior to the grid bias, as is essential with a Class AB amplifier. Descriptions of amplifiers using the DA42 in Class B are given in Chapter 6.

When the required power output exceeds 200W (the maximum obtainable from a pair of DA42 valves), either a single pair of suitable valves in Class AB1 or AB2 may be used, or the load may be divided into groups of 200W, each supplied by a pair of DA42 valves. Two DA100 in Class AB2 will provide 270W with an anode supply of 1,250V and the appropriate operating data is given in Chapter 7. A pair of V1505 will give up to 400W in Class AB1 or up to 1100W in AB2, at anode voltages of 2kV and 2·5kV respectively (Chapter 8).

The design of the driver stage for a Class B or Class AB2 amplifier will follow the same pattern since both arrangements require a similar driving power. For example, at 1000V anode potential, the outputs of a pair of DA42 and DA100 valves are 175W and 200W respectively. They need driving powers of 5W and 7W respectively but, as the Class B DA42 circuit requires a grid to grid input of 100+100V against the 240+240V required by the Class AB2 DA100, its driver anode voltage may be lower. It should be remembered that the driver stage is itself a power amplifier, voltage gain being of secondary importance.

Valves larger than the DA42 have been seldom used in Class B zero bias amplifiers but the advantages of this type of circuit, when it is used in conjunction with the improved driver stage described below, will undoubtedly encourage the manufacture of suitable Class B valves for use in the 500-1000W range.

Class B Driver Stages.—Since the driver stage is so intimately associated with the Class B output stage its features will be discussed here instead of in the section describing intermediate stages (page 12).

The most satisfactory driver stage is undoubtedly that using a cathode-coupled, bridged transformer.* (See fig. 1-2). The transformer contains four sections, all of which must have an identical number of turns, but the d.c. resistance of the primary is often made higher than the secondary (which should be of low resistance) in order to provide the correct cathode bias potential for the driver valves. Two capacitors of 2-16μF, which may be of the electrolytic type, join the ends of the primary and secondary and hence short-circuit the transformer leakage inductance, which is largely responsible for the bad reputation of Class B amplifiers for distortion. The KT66 is frequently used in cathode-coupled stages of this type, in which it operates at about 50mA cathode current with a cathode bias of 30-40V. To provide this bias, each half of the transformer primary should have a resistance of 750Ω, but should it be less, an additional resistor—common to both valves—may be connected between the

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*This circuit was originally developed in the Research Laboratories of the G.E.C. (Patent No. 487953).
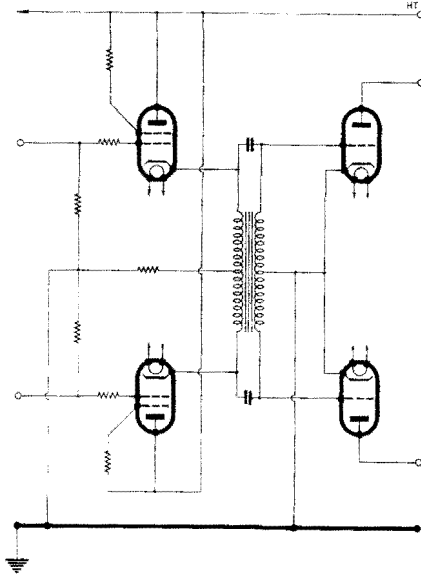
Fig. 1-2. The cathode-coupled bridged transformer stage. The circuit of a Class B amplifier using this type of stage will be found on page 67. Details of a typical transformer are given on page 75.
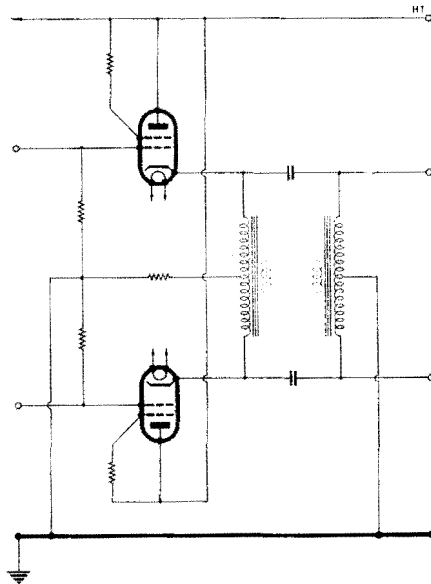

Fig. 1-3. A variation of fig. 1-2, using two centre-tapped inductors or two push-pull output transformers.

centre tap and earth. Should the correct type of transformer not be available, satisfactory operation can be obtained with two centre-tapped inductors, for example two push-pull output transformers (which need not be identical), with their secondaries left unused as shown in fig. 1-3. The KT66 cathodes feed the output valve grids through large capacitors as before. Their working voltage must allow for the KT66 bias and any d.c. grid voltage applied to the output valves.

If a conventional transformer without capacitance bridging is used, transient oscillations are often set up by the incidence of grid current and the superiority of the bridged transformer in this respect is clearly evident from an oscilloscope display of the waveform.

The cathode-coupled stage has low output and high input impedances. The output impedance is given, as a rough approximation, by $1/g_m$, this value being doubled for a push-pull pair. A typical stage using two KT66 valves has, therefore, an output impedance of 350Ω and a pair of KT55 valves ($g_m = 20mA/V$) would have an output impedance of 100Ω. The only disadvantage, a minor one, is that, like all circuits of the cathode-follower type, there is a small overall loss in signal amplitude. The loss in this case is below 10% and since the input impedance is high, this may be overcome in the previous stage. Examples of the cathode-coupled driver stage may be found in the circuits of Chapters 6, 7 and 8. The preceding voltage amplifying stage is discussed in Section 1-8, page 12.

Class AB2 Driver Stages.—The recommended circuit for driving a Class AB2 output stage is essentially the same as that for Class B (fig. 1-2) but, as stated previously, the Class AB2 amplifiers described in this book are of over 200W output and a higher output voltage is required from the driver stage. This necessitates a slightly different arrangement for the preceding voltage amplifier and circuit details are given in Section 1-8, page 12.

1-4. Output Stages for d.c./a.c. Amplifiers.—When only a d.c. mains supply is available, the anode voltage is limited to 220V or so. The G.E.C. markets two low impedance output beam pentodes designed for d.c./a.c. operation : the KT33C and KT55. Suitable amplifier designs and the measures taken to obtain maximum output are given in Chapters 3 and 4. It is important to use a smoothing inductor of low d.c. resistance (100Ω or less) and an efficient output transformer, also of low resistance, in order not to reduce still further the low available anode voltage.

A pair of KT33C or KT55 valves will give 14W or 25W output respectively but multiple pairs may also be used. Chapters 3 and 4 give circuits, of 14W and 25W output respectively, for operation from d.c. or a.c. mains and Chapter 4 includes information on the KT55 for designers of marine equipment.

1-5. Matching of Valves in Triode and Ultra-Linear Circuits.—The minimum distortion at any given output is obtained when the two output valves have similar characteristics and are operated under similar conditions.

Valves, like any other product produced in quantity, have manufacturing tolerances, so that two valves of the same type are not necessarily identical in electrical characteristics. However, it is possible to ensure that the minimum distortion is introduced either by valve selection or suitable circuit design. Selection of valves is normally undesirable although, if an additional valve is available, it is possible to minimise differences by selecting from the three valves two which have similar quiescent anode currents. This assumes identical values of cathode bias resistance and wire-wound resistors of not more than 5% tolerance are recommended. Suitable cathode shunt capacitors are essential and it may be noted that the conventional 50μF type is scarcely large enough for the KT55, for which 200-250μF should be used for satisfactory low frequency performance. An adequate working voltage should be allowed ; for example, a KT88 amplifier running at the maximum rated conditions would need 100V capacitors.

When fixed bias is used, such as in the larger amplifiers described in later chapters, the quiescent currents will be equalised by the grid bias supply potentiometers. Minimum distortion at full output may then be obtained if desired by equalising the anode currents with an additional potentiometer as, for example, R5 in fig. 5-5 page 59. This potentiometer is adjusted at 90% of full output to give identical cathode currents in the output valves. This arrangement is suitable for the KT55, KT66, KT88 and similar valves in fixed bias circuits.

This method of compensating for different characteristics and differences in signal voltages cannot be used in cathode-biased output stages. However, the following somewhat similar circuit arrangement coupled with a different measuring technique has been found useful :

Fig. 1-4 illustrates the slight modification (comprising the addition of R and RV) of the double triode penultimate stage in cathode-biased amplifiers of the KT66 and KT88 class such as that shown in fig. 4 on page 42. R is a half-watt resistor of 1·5kΩ to 5kΩ and RV is a 10kΩ wire-wound potentiometer. Resistor R introduces feedback which equalises the two triode systems and renders RV more effective.

The balancing procedure is as follows :

1. Connect an oscilloscope and the correct load impedance to the output transformer secondary.

2. Apply to the amplifier a sine wave signal of 200-2000c/s of sufficient amplitude to give nearly full output and increase it until either the upper or lower peaks of the waveform on the oscilloscope begin to flatten. If both peaks begin to flatten simultaneously, the output stage will be already in perfect balance.

3. If necessary, adjust RV to give a symmetrical waveform. With a little practice, it is possible to obtain a balance nearly as good as that obtained with distortion factor equipment.

Fig. 1-4. A method of signal balancing for amplifiers with cathode biased output stages.
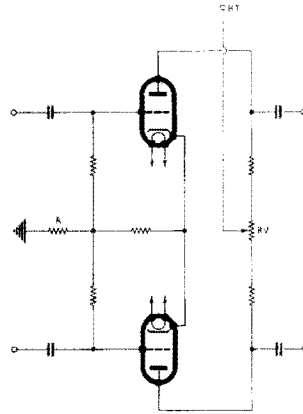
After this procedure, the distortion should be minimum at all output levels and will not greatly exceed 1% at full output (before applying negative feedback) with the ultra-linear or triode amplifiers of up to 50W output described in the following chapters. In pentode amplifiers, it is not considered that the improvement justifies the additional components.

1-6. Recommended Precautions in Output Stage Design.

1. Adequate ventilation must be allowed. If in doubt, temperature sensitive paint* may be used to determine the temperatures of valve bulbs. For long-term reliability the bulb temperature of output valves should not exceed 200°C.

2. Unnecessarily high values of grid resistance should be avoided. For valves up to the KT66 class with cathode bias, the grid-cathode resistance should not exceed 560kΩ. This permits the use of a standard 470kΩ resistor of 20% tolerance. With larger valves the maximum value is lower and with fixed bias it should not exceed one third to one half that permitted with cathode bias. If the valves are run at dissipations less than the maximum, the grid circuit resistance may be increased in inverse proportion.

3. With fixed bias some form of interlock is recommended to prevent the application of anode voltage before the bias, or to isolate the anode supply if the bias should fail. Typical safety circuits which allow an amplifier to operate at reduced output are given in fig. 5-9, page 64 and fig. 8-7, page 96.

4. Large Class B, AB1 or AB2 amplifiers require a protective device across the output transformer to absorb surges. A pair of resistors and capacitors has been found satisfactory. The capacitors may have a reactance equal to the load impedance at, say, 10-15kc/s and the resistors may equal the load impedance and have a rating of about one tenth of the full output power of the amplifier.

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*A suitable paint, called " Tempilac " is obtainable from J. M. Steel & Co. Ltd., 36 Kingsway, London, W.C.2.
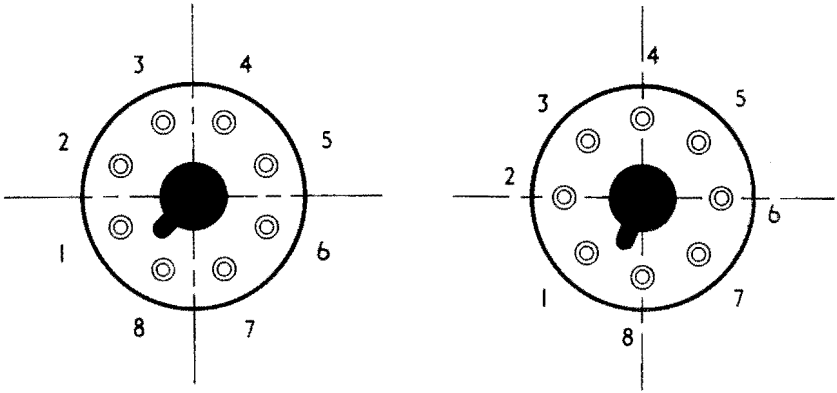
Fig. 1-5.   Correct orientation of the valve sockets for horizontally mounting the KT66 (left) and the KT88 (right).

5.   When valves must be mounted horizontally, it is preferable to arrange for the major axis to be vertical.   The proper orientation of the valve socket depends upon the valve type but the correct positions for the KT66 and KT88 are given in fig. 1-5 above.   By arranging the axis of the valve in this way, variation in characteristic due to sagging of the grid wires after a long period of life is avoided.   A similar precaution is desirable with directly heated power rectifiers.

THE INPUT STAGE

1-7. Design Requirements.—The input or first stage of an amplifier will be designed according to the service for which the equipment is required. Normally, amplifiers within the scope of this book must be capable of operating from a radio receiver or record reproducer. Industrial or public address amplifiers are required to work from a microphone but few domestic amplifiers are used with a microphone except in conjunction with a tape recorder.

When a microphone or tape recorder is the source, the input stage must be designed with a low noise level, including a low hum level, and a high anti-microphony standard. If the valve around which this stage is designed is correctly chosen, no difficulty should be experienced in attaining the required performance.

The G.E.C. Z729 is unequalled for this service and it may be connected either as a triode or pentode. It may be used satisfactorily with an a.c. heater supply at low-level inputs since the hum level referred to the grid does not exceed $1\cdot5\mu$V and a high standard of anti-microphony is maintained. The following few precautions will be of assistance in obtaining trouble-free service from the Z729 :

1. The valve socket must be of good quality and of non-hygroscopic high-insulation material. The type with a large metal centre boss is particularly suitable in that the boss screens the grid pins from the heater pins. An additional screen across the socket is sometimes advantageous.

2. The heater supply wires should consist of a twisted pair kept well clear of the grid connection at pin 9.

3. Cathode bias should be used, the resistor being bypassed by a large electrolytic capacitor. The bias should be from 1V to 1·5V. It is not recommended that some or all of the cathode resistor of the input stage be left unshunted because of the possibility of hum being induced from the heater supply. The leads of the cathode bypass capacitor should be arranged closely adjacent to the lead to the signal grid pin and the capacitor earthed at the signal input socket.   By this means, the loop formed by the cathode-grid circuit is kept as small as possible to minimise electro-magnetic hum pick-up.

4. A centre-tapped heater supply should be provided but there is seldom justification for a variable balance control.

5. When the Z729 is required to operate from a low-level source the anode and control grid resistors should be of the high stability type to reduce noise to a minimum. Similar resistors may be used in the screen and cathode circuits, but this is not essential as both are shunted by large capacitors. The screen should be bypassed to earth through a paper capacitor of at least 0.5μF. Sometimes, for example in electro-medical equipment, this may with advantage be increased to 2μF.

6. The ratio of the screen resistor to the anode load resistor should be at least 5:1. In low-level circuits this may be increased to 10:1 to obtain the optimum signal-to-noise ratio.

Some consideration must be given to the relative levels of the various signal sources, as they are likely to differ considerably. A satisfactory method of catering for several inputs is used in the two alternative input units designed for the G.E.C. " 912-PLUS " (figs. 3-4 and 3-5, pages 31 and 32). The pre-amplifier circuit of fig. 3-5 provides four types of equalisation for magnetic pickups of low output (10mV), a microphone input of 1mV, and a radio input of 50mV.

Tape recorders normally contain their own pre-amplifier equalisation and the Z729 is unsurpassed in this application. As tape equalisation must boost the lower frequencies by 20-30db, the precautions listed above should be followed carefully. In lightweight portable equipment an anti-microphonic valve socket may remove any objectionable noise resulting from operation of the controls. To obtain the maximum suppression the valve socket may require to be mounted on a sub-chassis having an appreciable mass, this being attached to the main chassis through a resilient mounting. This is, however, seldom necessary. Flexible connections must be used with any type of anti-microphonic socket.

Chapter 9 discusses the design of pre-amplifier tone control units for use with most of the complete amplifiers described in this book which do not contain built-in tone control facilities.

THE INTERMEDIATE STAGES

1-8. Some Typical Intermediate Stage Circuits.—After the signal has passed through the first stage or low level amplifier it will have attained an amplitude of some 50-100mV. The intermediate stages amplify the signal to the level required by the output stage for full power. At the input to the intermediate stages, the signal has sufficient amplitude to ensure that the signal-to-noise ratio does not show any apparent change with operation of the gain control.

The ideal valve for the first stage of this part of the amplifier is the Z729, triode connected, which will give a gain of about 25-30db.   This may be followed by the tone control network, if this is to be incorporated in the main amplifier, which will have a similar order of insertion loss, the level then being raised to 1-2V by a further triode ampifier stage.

In the smaller equipments one section of a double triode is often used for the stage following the tone controls, the other section being used as a phase splitter for the output stage.   This arrangement is used in the " 912-PLUS " (Chapter 3), but it is essential to select a valve capable of giving the necessary performance.   The G.E.C. B719 can be recommended and will be found relatively free from microphony.   Originally developed for f.m. service, it is a low-impedance valve capable of a stage gain of about 35db (50 times).   It may replace the B309 in the " 912-PLUS " with only a minor change of the heater wiring, the other connections remaining the same.*   The B719 will give a slightly higher overall gain than the B309 but it has a similar impedance of 10kΩ.

The complete amplifiers described in the later chapters of this book all require an input of between 0·5V and 2V and have either two or three amplifier stages before the output stage.   Some of the circuits will be examined to explain modern practice.

In the KT66 or KT88 30W ultra-linear amplifier of fig. 4-4, page 42, a pair of B65 double triodes are used which, with the addition of 14db negative feedback, ensure full output from the complete amplifier at less than 0·5% distortion.   One half of the first B65 acts as the input stage, which is direct-coupled to the other half, a conventional phase splitter.   The second B65 is a push-pull voltage amplifier and together they provide a well-balanced signal for the output stage.   Any slight unbalance is neutralised by negative feedback.

A somewhat different intermediate stage is used in the 50W amplifier employing KT88 valves.   (Fig. 5-2, page 55).   In this circuit, the first double triode is arranged as a floating paraphase phase inverter.   The penultimate stage is provided with a control so that the input signal to the KT88 valves can be adjusted to give a balanced output from the output stage with consequent minimum distortion.   This control is adjusted for equal cathode currents in the output stage at about 90% of full output on a steady tone signal.   Its use is optional and only necessary when the lowest possible distortion is required.

The intermediate circuit recommended for driving Class B and AB2 output stages and designed around the cathode-coupled bridged transformer has already been discussed in Sect. 1-3, page 4.   The amplifier which has been designed to precede this type of driver circuit is shown in fig. 1-6 overleaf.

*The B719 heater is not centre-tapped and the connections are brought out to pins 4 and 5.   Pin 9 (the heater centre-tap in the B309) is connected to an internal screen in the B719 and should be earthed.
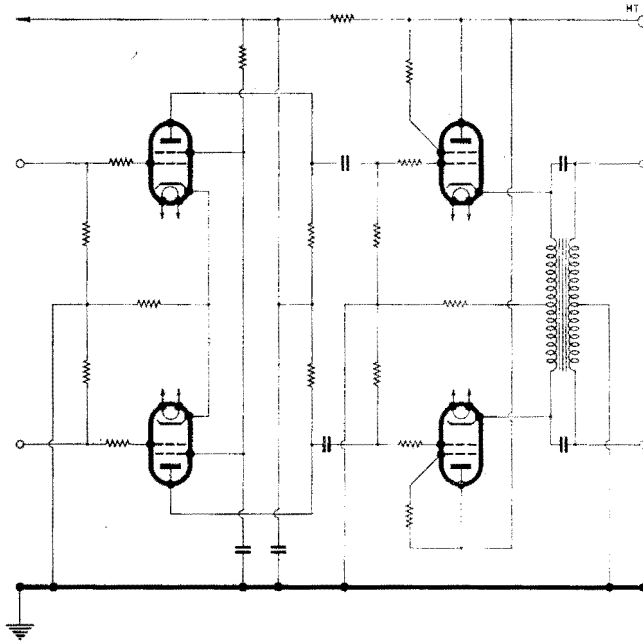
Fig. 1-6. A cathode-coupled driver stage (on the right of the diagram) driven by a pair of beam pentodes arranged as a normal push-pull voltage amplifier with resistance load. This type of intermediate stage is used in the Class B amplifiers on pages 67 and 78.

The high input impedance of the cathode-coupled stage permits advantage to be taken of the high voltage gain obtainable from a pair of small power pentodes, in this case the N709. Normally used as an output valve, the N709 is used here with the relatively high anode load of $47k\Omega$ and gives a gain of about 40db (100 times).

The circuit of fig. 1-6 is capable of an output of $125+125V$ peak at not more than 2% distortion and this is adequate for use in the Class B circuits given in Chapter 6. However, the Class AB2 amplifiers described in Chapters 7 and 8 require a driving signal of some $200+200V$ and, as there is a slight loss of gain in the cathode-coupled driver stage, the preceding stage must supply a signal somewhat greater than this. The output of the circuit of fig. 1-6 is limited by the relatively low anode and screen potentials inherent in R-C coupled amplifiers with normal power supply circuits and some alternative is necessary.

Fig. 1-7 shows the circuit for Class AB2 amplifiers and it is, in fact, a simple modification of fig. 1-6. The anode resistors are shunted by a centre-tapped inductor with the result that the anode potential is practically that of the h.t. line and the maximum output from the stage is increased to over $200+200V$
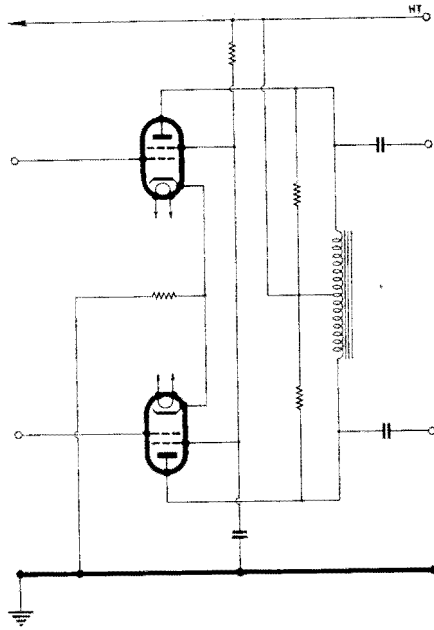
Fig. 1-7. If the voltage amplifier stage of fig. 1-6 is modified by adding a centre-tapped inductor, a higher output voltage is obtainable and the circuit is suitable for large Class AB2 amplifiers such as that shown on page 85.

peak. This circuit retains the excellent frequency and distortion characteristics of that shown in fig. 1-6 and would seem to be an improvement on the more conventional transformer-coupled triode amplifier. The inductance value is determined by the lowest frequency required and will be of the order of 100-200H for satisfactory operation at 75c/s. An input of 2-3V grid-to-grid will give an output of 200+200V peak at the secondary of the cathode-coupled transformer in the following driver stage.

When a voltage amplifier feeds an output stage of unusually low input impedance, a cathode follower buffer stage with its high input and low output impedances will maintain the output of the voltage amplifier. This type of intermediate stage is used in the multiple-pair push-pull amplifier described in Appendix A, in which the parallel-connected output valves must have a common low resistance grid circuit.

THE POWER SUPPLY

1-9. Power Supply Categories.—Power supplies for audio amplifiers may be grouped broadly into two types : 1: Those in which the rectifier feeds directly into an inductor, usually 5 to 20H, without a reservoir capacitor (inductance-input), and 2 : those provided with a capacitor as the first element of the smoothing circuit (capacitance-input).

Each type has its advantages and the selection of one or the other will depend on the design of the amplifier connected to it.

The inductance-input type provides a more constant output voltage with varying load current.   It is said, therefore, to have " better regulation " or " lower impedance."   This type of power supply will be found in all the larger amplifiers, say above 25-50W output, where fixed bias is used to provide a low anode current demand in the quiescent periods.

The capacitance-input type provides a higher output voltage, other things being equal, but the change in voltage with current is greater, i.e. the regulation is poorer.   It is nearly always used to supply voltages below 450 to amplifiers with cathode bias in which little or no change in anode current occurs.   The first capacitor after the rectifier (the reservoir capacitor) is subjected to a peak voltage considerably higher than the d.c. output voltage and which is approximately equal to one and a half times the r.m.s. voltage supplied by the mains transformer.   Consequently, for d.c. output voltages above 350-400, it may be necessary to connect two or more electrolytic capacitors in series to obtain the required value of capacitance with the necessary working voltage.

Correctly designed, the inductor of an inductance-input filter prevents the high peak voltage reaching the smoothing capacitor and single electrolytic capacitors may be used in higher voltage power supplies, but the mains transformer is usually required to provide a higher voltage for the same d.c. output potential.   The resistor and capacitor connected across the first inductor of the inductance-input circuits in this book (for example, C and R in fig. 1-9, opposite) suppress parasitic high voltages which would otherwise tend to break down the insulation of the components.   Suitable values are 10kΩ and 0.01μF. They should not be omitted.

An inductance-input filter may be followed by a further inductor and capacitor when a high degree of smoothing is required, although in the larger amplifiers, where a small amount of hum is not objectionable, a single section filter is often used.

Television receiver design techniques have resulted in the economical production of large-value electrolytic capacitors of relatively high working voltage.   A capacitance of $100\text{-}200\mu$F is obtainable in a small can and two or more of these may be series-connected to operate at any required voltage. Three capacitors each 150μF, 450V rating have been used with a DA42 Class

B amplifier to give an effective $50\mu F$ at 1350V.   A high capacitance of this order is desirable with all amplifiers in which a large change in anode current is produced with changes in power output, but even domestic amplifiers of the 10-15W type are improved by the resulting low impedance power supply.

The rectifying circuits used to provide the d.c. anode supplies in most of the amplifiers in this book are either the bi-phase half-wave (fig. 1-8), or the full-wave (" bridge ") circuit (fig. 1-9).   Each has its advantages.   The former is
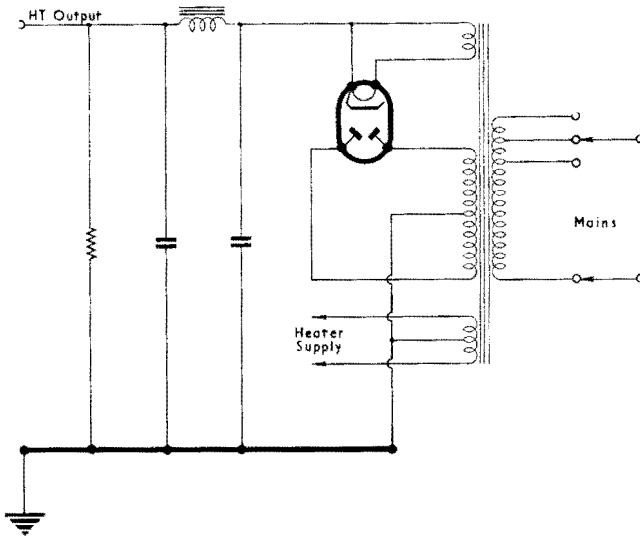
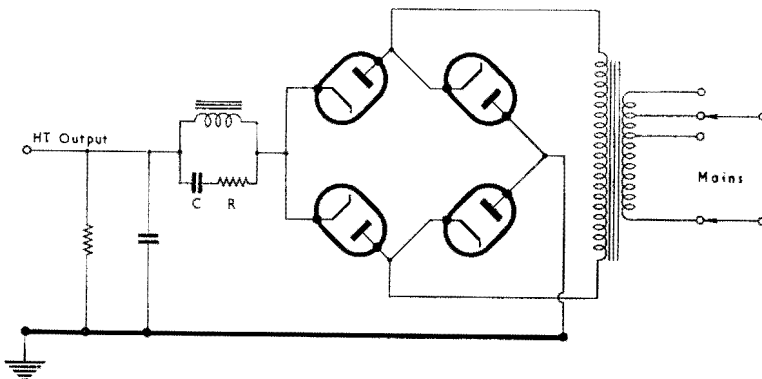**Fig. 1-8.**   The bi-phase half-wave rectifier circuit.

Fig. 1-9.   The full-wave or " bridge " rectifier circuit.   C and R prevent parasitic high voltages across the inductor as explained opposite.

nearly always used for h.t. voltages below 750 because of the simplicity of a single valve, but the bridge circuit, although requiring four valve systems, calls for a transformer giving only half the total secondary voltage. In high voltage equipment this effects a considerable saving in insulation.

The bridge circuit, an example of which is given in fig. 6-4 page 71, is capable of providing two d.c. output voltages simultaneously, one of which is half the other. Fig. 1-10 is a simplified diagram in which the low voltage part of the circuit is in full line and the high voltage part in broken line. A variation of this circuit, using three rectifiers, is given in fig. 1-11 and it will be seen that electrically the arrangement is precisely the same. The input inductor of the low voltage supply in either circuit must not be omitted.

High-powered Class AB1 and AB2 amplifiers require, in addition to the anode supply, a source of negative grid bias. No particular precautions are necessary in a Class AB1 amplifier and a bias supply impedance of 20-30kΩ is satisfactory, thus the bias source need not supply more than 5-10mA. In a Class AB2 amplifier, considerable grid current flows at full output and as this may amount to 30-40mA it is important that the bias remains fairly constant. The voltage will tend to increase if the bias supply is of high resistance and
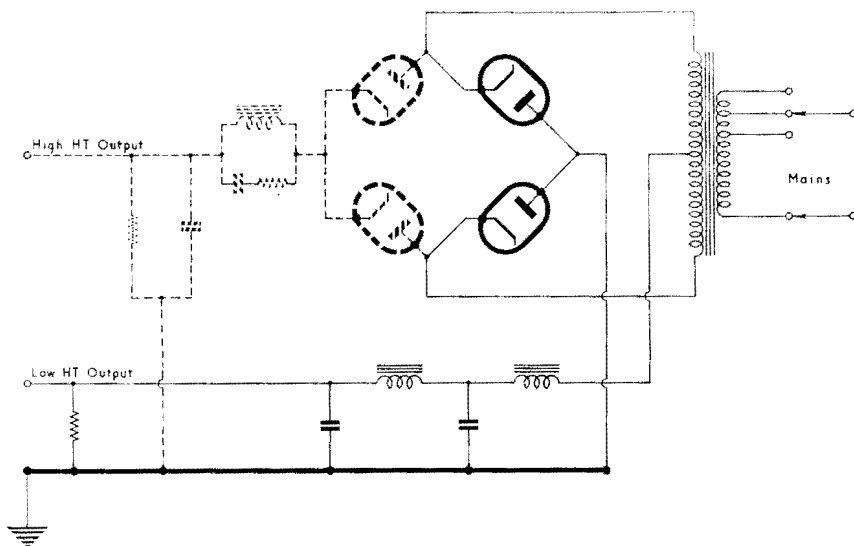
Fig. 1-10. The bridge circuit arranged to provide two output voltages, one of which is half the other.

serious crossover distortion and limitation of output results. Either a stabilised supply or one supplying 50-100mA may be used. The latter is simpler and only slight crossover distortion will be apparent at 95% of the full output of the amplifier.

The bias supply for a Class AB1 amplifier may be of the capacitance-input filter type but inductance-input should always be used for Class AB2. With the low-impedance valves used in these amplifiers it is essential that the anode voltage is not applied without grid bias and an interlock should be provided to hold-off or disconnect the anode supply should the bias fail. An alternative and probably better arrangement is to arrange for the automatic insertion of a cathode bias resistor which will permit the amplifier to function at reduced output until serviced. See fig. 5-9, page 64 and fig. 8-7, page 96.

Fig. 1-11. The dual power supply of fig. 1-10 obtained with three valves.

CHAPTER 2

A 5-WATT JUNIOR AMPLIFIER

This circuit has been designed to meet the need for an inexpensive amplifier of modest output power and simple tone control. It has an output of 5W with less than 2% distortion and is very suitable for use with lower priced loudspeakers and record players.*

Four valves are used : a DH77 triode and two LN309 triode-pentodes give five stages of amplification with only three valves ; and the fourth, a U709, is the rectifier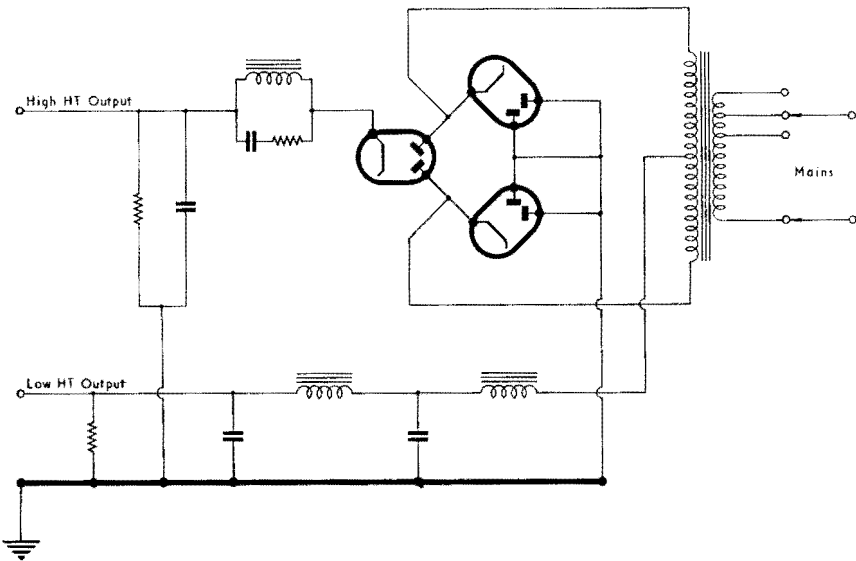. The circuit, which contains an ultra-linear output stage, is given in fig. 2-1. The tone controls, although simple and requiring fewer components than those in the G.E.C. " 912-PLUS," for example, are capable of giving satisfactory results from both LP and 78 r.p.m. records. The full range of tone control is also available when the amplifier is used with a radio receiver. An input of half a volt r.m.s. will give full output.

Fig. 2-2 shows that it is not critical as to load impedance ; fig. 2-3 gives the distortion of the amplifier at various power outputs and fig. 2-4 shows the frequency response ; the uncritical load impedance is one of the many desirable features of the ultra-linear circuit which enables the G.E.C. Junior Amplifier to out-perform many other low-priced amplifiers.

A pair of LN309 valves are used in the ultra-linear output stage, with the screen grids taken to taps in each half of the output transformer primary at 20% to 40% of the turns from the centre-tap. The triode-pentodes are V2 and V3, the triode V2A being an amplifier and V3A the phase splitter. The splitter load resistors R13 and R14 should be a matched pair, but, if this is impracticable, satisfactory results will be obtained with 10% resistors. The value of the negative feedback resistor R23 is determined, strictly speaking, by the impedance of the loudspeaker. However, a compromise value of 220Ω has been used and gives good results with either a 3·5Ω or 15Ω loudspeaker.

The amplifier requires an h.t. supply of 250V at 65mA. The two 6·3V windings on the mains transformer are connected in series to provide a centre-

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*A complete description of the " Junior " amplifier was given in *The Radio Constructor*, November and December 1956, published by Data Publications Ltd.
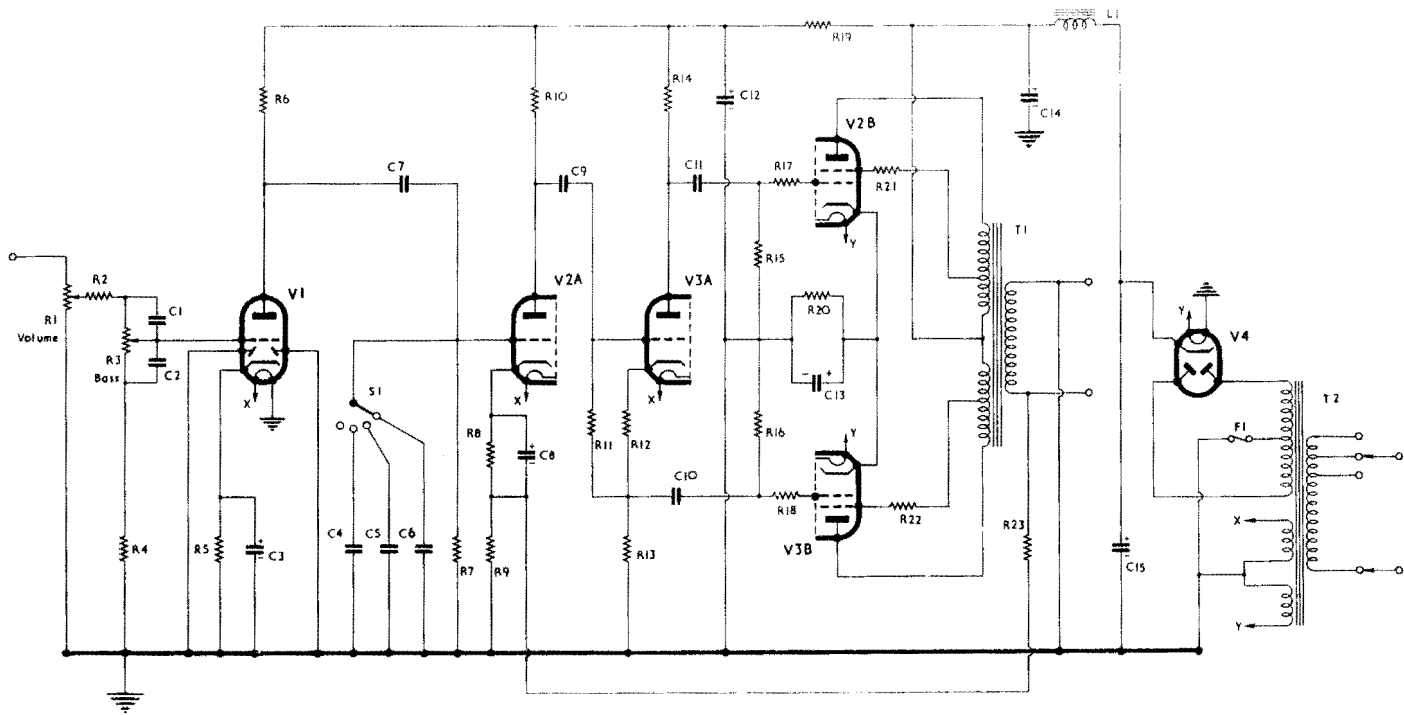
Fig. 2-1. Circuit of the 5W Junior Amplifier. Component values are given on page 23.

tapped 12·6V supply, the LN309 heaters being connected across the total supply and the DH77 and U709, which require 6·3V, being each connected across one half-winding. When connecting the 6·3V windings they must be series-aiding (in phase), otherwise the LN309 will not operate although no damage will be done.

The bass control R3 should be wired up so that C2 is connected between the " tapered " end of the potentiometer and the slider. The tapered end of a normal " volume control " is that at which the slider is fully anticlockwise.

The output transformer is provided with two secondary windings which may be connected in series or parallel for either 15 or 3·5Ω loudspeakers. The respective ratios are 24:1 and 48:1.

TABLE 2-I

OPERATING CONDITIONS OF THE LN309 OUTPUT STAGE OF FIG. 2-1

$V_{a\ (b)}$	250	V
$V_{a,\ g2}$	230	V
$I_{a+g2\ (o)}$	2×28	mA
$p_{a+g2\ (o)}$	$2 \times 6·5$	W
R_k	2×330	Ω
V_g	$-18·5$	V
P_{out}	5·5	W
$R_{L\ (a-a)}$	8	kΩ
z_{out}	4	kΩ
D	1·7	%
$V_{in\ (rms)}$ (to first stage)	0·5	V
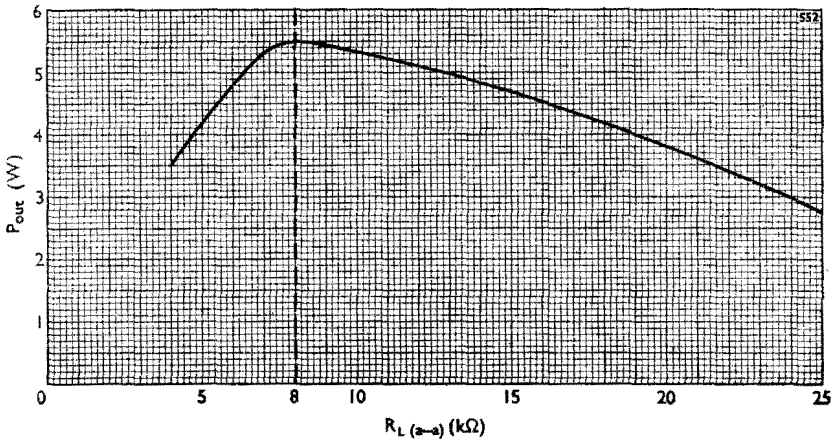
Fig. 2-2. 5W Junior Amplifier : power output against load impedance.

COMPONENT VALUES FOR FIG. 2-1

5W JUNIOR AMPLIFIER

G.E.C. VALVES
V1 DH77/6AT6
V2 LN309
V3 LN309
V4 U709

RESISTORS

(20%, 0·25W unless otherwise shown)
R1 1MΩ Log
R2 470kΩ
R3 2MΩ Log.
R4 47kΩ
R5 2·2kΩ
R6 150kΩ
R7 1MΩ
R8 2·2kΩ
R9 47Ω
R10 68kΩ
R11 470kΩ
R12 470Ω
R13 15kΩ ⎫
R14 16kΩ ⎬ Matched to 5%
R15 470kΩ ⎭
R16 470kΩ
R17 22kΩ
R18 22kΩ
R19 15kΩ 0·5W
R20 330Ω 10% 1W
R21 220Ω
R22 220Ω
R23 220Ω

CAPACITORS
C1 1000pF
C2 0·01μF
C3 50μF 12V
C4 500pF
C5 0·001μF
C6 0·002μF
C7 0·01μF
C8 50μF 12V
C9 0·01μF
C10 0·01μF
C11 0·01μF
C12 8μF 250V
C13 25μF 25V
C14 ⎫
C15 ⎬ 32+32μF 450V

MISCELLANEOUS
L1 5H 65mA 100Ω
T1 5W Ultra-linear transformer
      8kΩ anode-anode
      Dual ratio : 24:1 and 48:1
      Primary inductance : ≮20H
      Leakage inductances :
      Prim.—sec. : ≯15mH
      ½ prim.—UL tap : ≯10mH
T2 Mains transformer
      Secondaries :
      250-0-250V 65mA
      6·3V 1A
      6·3V 1A
S1 1-pole, 4-way

Fig. 2-3. 5W Junior Amplifier : distortion characteristics.

c

Fig. 2-4. 5W Junior Amplifier : effects of the bass and treble controls.

CHAPTER 3

AMPLIFIERS OF 12 TO 14 WATTS

3-1. A 14W Ultra-Linear Amplifier.—This class of amplifier is probably the most popular for use in domestic equipment. As it is frequently required to work from a diversity of signal sources—radio tuners, record players, tape recorders, microphones etc., the domestic amplifier is generally used in conjunc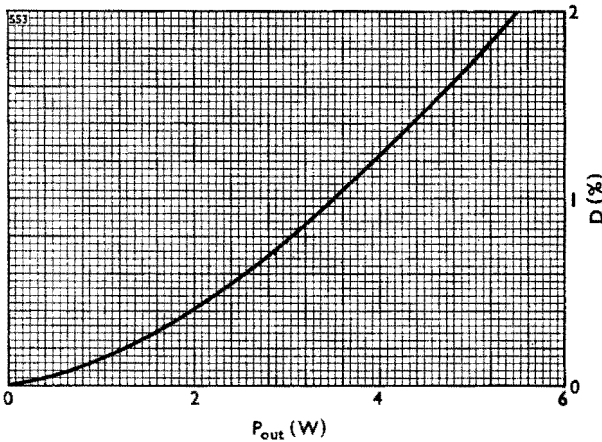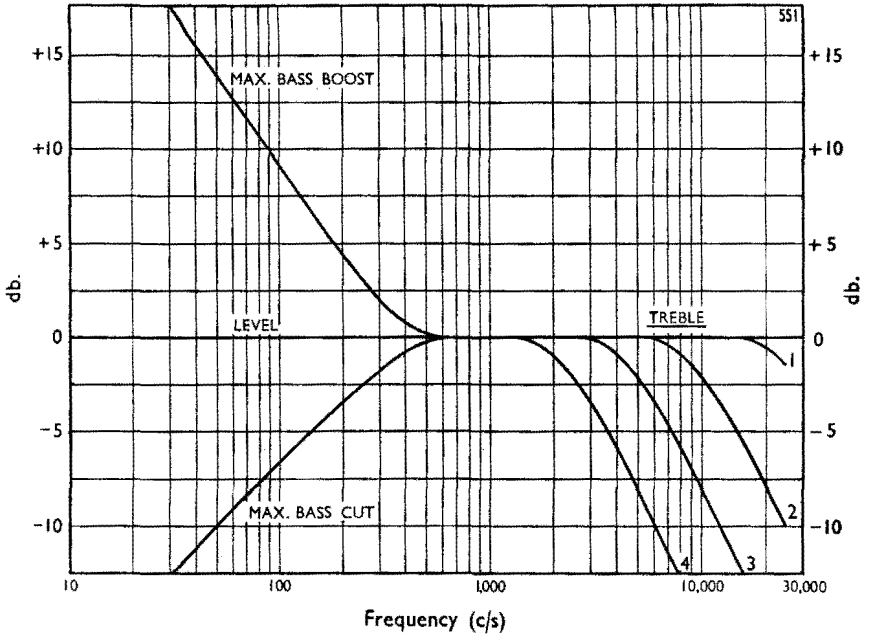tion with auxiliary pre-amplifying and equalisation equipment. Consequently, a slight departure is made here by describing a complete amplifier with tone controls and, in addition, alternative input selection and equalisation units. Higher power amplifiers suitable for domestic use are described in Chapter 4 and separate pre-amplifier and tone control circuits are discussed in Chapter 9.

The main amplifier circuit is given in fig. 3-3 and the two types of input unit appear in figs. 3-4 and 3-5. This design is based on the very popular " G.E.C. 912-PLUS " amplifier.* The full output of 12-14W (the precise value depending upon the quality of the output transformer) is obtained with less than 1% total harmonic distortion over the frequency range of 30c/s to 20kc/s. An input of 50mV will give 12W output.

Comprehensive tone controls are built into the main amplifier and both the alternative input selection units provide switching for radio input, four different record characteristics and microphone. The pre-amplifiers described in Chapter 9 are very suitable for combining with the G.E.C. 912-PLUS when the flexibility of a separate pre-amplifier and tone control unit is desired. Details are given at the end of this chapter.

One of the units contains passive networks (i.e. there is no amplification) for equalising the four record characteristics and is intended for use with crystal pickups and microphones. The alternative unit is a single-valve pre-amplifier for which details are given not only for equalising the four record characteristics

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*Full constructional information, with point-to-point wiring instructions, is given in *The G.E.C. 912-PLUS*, published by the G.E.C. Valve and Electronics Dept.

from different types of pickup but also for matching radio and microphone outputs to the main amplifier.

The passive input unit (fig. 3-4) has a sensitivity of 150mV on the " record " positions of the selector switch.   The " radio " and " microphone " inputs are switched straight through to the main amplifier and consequently the original sensitivity of 50mV is retained.

The pre-amplifier (fig. 3-5) can be made specifically to suit the requirements of the particular pick-up, radio and microphone used, but maximum sensitivities are : 10mV for the " record " positions of the selector switch, 1mV for " microphone," and, nominally, 50mV for " radio."   The required input impedance and sensitivity at any particular position of the input selector switch will depend upon the pickup or microphone used.   Table 3-II gives appropriate alternative values for the resistors which determine these characteristics.

The circuit of the main amplifier is given on page 28.   The first stage consists of a Z729 low-noise pentode followed by the tone control networks, described later.   Next, the first section of a B309 double triode* is a typical voltage
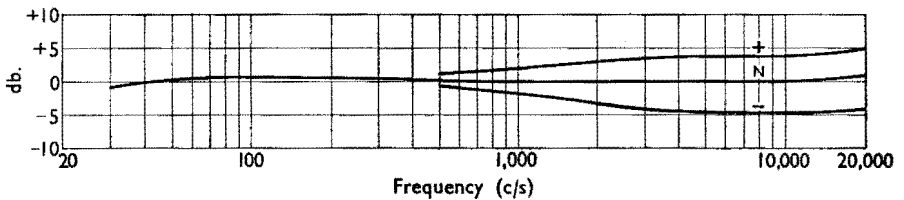
*A B719 may be used ; see page 13.

Fig. 3-1.   12W ultra-linear amplifier :   effect of the " presence " control, with bass and treble controls at the level positions (mid-travel).

amplifier with negative feedback applied to the cathode circuit from the output transformer secondary. The feedback network contains an additional tone control (" presence ") as described later. The second half of the B309 is a phase-splitter which feeds the N709 ultra-linear output stage, the characteristics of which are given in Table 3-I. The output transformer is tapped at 20% to 40% of the turns of each half-primary from the centre tap. The power supply is of conventional design with a U709 feeding a capacitance-input filter.

Bass Loudness.—This control is continuously variable, giving linear response at the centre, bass attenuation anticlockwise, and bass accentuation clockwise.

Treble.—This is a switched control providing a linear position, two degrees of attenuation and one degree of boost.

Treble Slope.—A continuously variable control which alters the slope of the high frequency cut selected by the treble switch.

Presence.—This switch, the effect of which is shown in fig. 3-1, is designed to vary the listener's apparent position when listening to music. It introduces a frequency-selective network into the negative feedback loop.*

*The full significance of this control is explained in *The G.E.C. 912-PLUS*, page 18. See also Sect. 9-3, page 108.
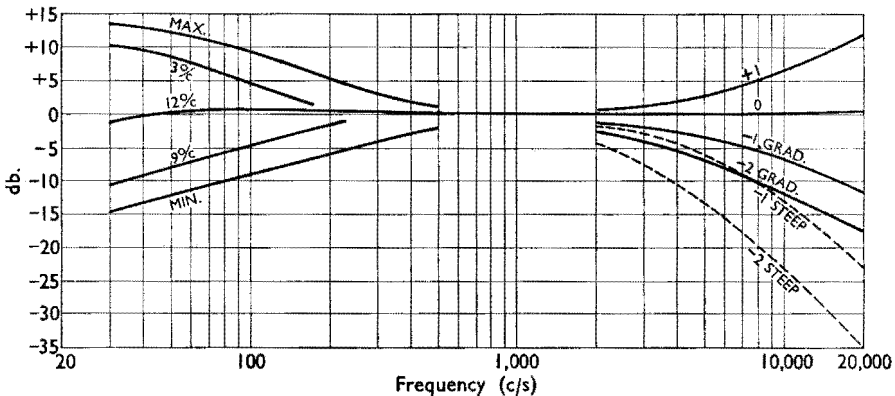
Fig. 3-2. 12W ultra-linear amplifier : the effect of the bass, treble and treble slope controls. The right-hand solid curves illustrate the four positions of the treble switch. The dotted curves show how the two treble cut curves are modified by fully rotating the treble slope control. The position of each dotted curve will alter as treble slope is decreased until, when it is at minimum, the treble response is that shown by the corresponding solid curve.
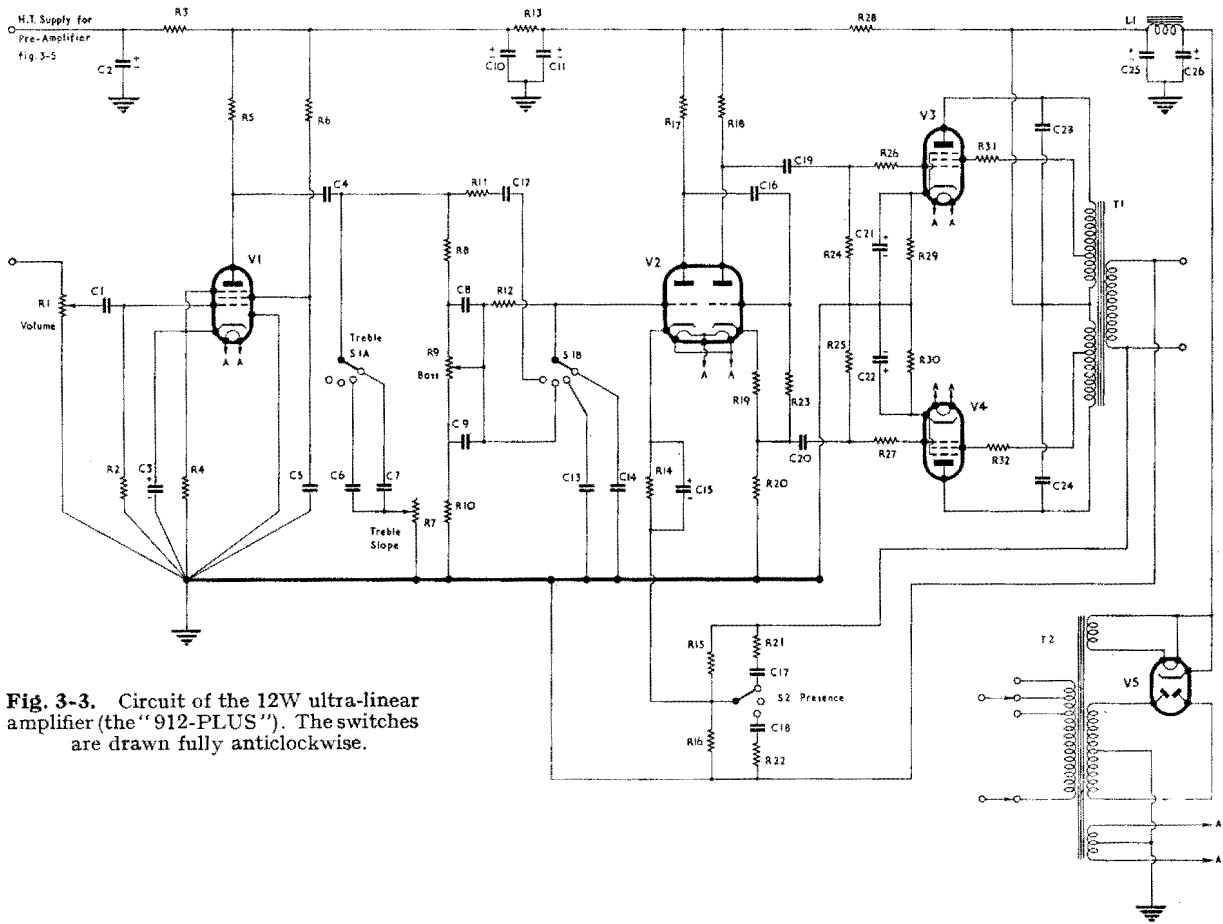
Fig. 3-3. Circuit of the 12W ultra-linear amplifier (the "912-PLUS"). The switches are drawn fully anticlockwise.

COMPONENT VALUES FOR FIG. 3-3

14W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 B309/12AT7
V3 N709
V4 N709
V5 U709

RESISTORS

(20%, 0·5W *unless otherwise shown*)
R1 1MΩ Log.
R2 2·2MΩ
R3 47kΩ 0·25W
R4 1·2kΩ
R5 100kΩ
R6 330kΩ
R7 1MΩ Log.
R8 220kΩ 5%
R9 1MΩ Log.
R10 22kΩ 5%
R11 150kΩ
R12 100kΩ 5%
R13 22kΩ
R14 680Ω
R15 1kΩ
R16 68Ω
R17 47kΩ
R18 22kΩ 5% 1W
R19 680Ω
R20 22kΩ 5% 1W
R21 1kΩ
R22 47Ω
R23 1MΩ
R24 220kΩ
R25 220kΩ
R26 10kΩ
R27 10kΩ
R28 10kΩ 1W
R29 270Ω 10% 1W
R30 270Ω 10% 1W
R31 47Ω
R32 47Ω

CAPACITORS

C1 0·005μF
C2 8μF 450V
C3 25μF 25V
C4 0·05μF
C5 0·1μF
C6 470pF 5%
C7 1000pF 5%
C8 2000pF 5%
C9 0·02μF
C10 16μF 450V
C11 8μF 450V
C12 22pF 5%
C13 220pF 5%
C14 470pF 5%
C15 25μF 25V
C16 0·1μF
C17 0·1μF
C18 2μF (paper)
C19 0·05μF
C20 0·05μF
C21 50μF 25V
C22 50μF 25V
C23 0·001μF
C24 0·001μF
C25 16μF 450V
C26 8μF 450V

MISCELLANEOUS

L1 10H 110mA
T1 14W Ultra-linear transformer
         7kΩ anode-anode
         Primary inductance : ≮50H
         Leakage inductances :
         Prim.—sec. : ≯20 mH
         ½ prim.—UL tap : ≯ 10mH
T2 Mains transformer
    Secondaries :
         300-0-300V 110mA
         6·3V 2·5A CT
         6·3V 1A
S1 2-pole 4-way m.b.b.
S2 1-pole 3-way m.b.b.

TABLE 3-I

OPERATING CONDITIONS OF THE N709 OUTPUT STAGE OF FIG. 3-3

$V_{a\ (b)}$	300	V
$V_{a,\ g2}$	295	V
$I_{a+g2\ (o)}$	2×44	mA
$I_{a+g2\ (max\ sig)}$	2×50	mA
R_k	2×270	Ω
V_g	-11	V
P_{out}	12	W
$R_{L\ (a-a)}$	7	kΩ
z_{out} (approx.)	2	kΩ
D	<1	%
$V_{in\ (rms)}$ (to first stage)	50	mV

TABLE 3-II

ALTERNATIVE COMPONENT VALUES FOR FIG. 3-5

Gramophone Input (all figures for 50mV input to main amplifier)

Pickup Output (at 4cm/s as a guide)	Required Load Impedance	R2	R3	R5	R19
5mV	10kΩ 22kΩ 47kΩ	15kΩ 47kΩ Infinity	47kΩ	100kΩ	47kΩ
10mV	22kΩ 47kΩ 100kΩ	33kΩ 100kΩ Infinity	100kΩ	100kΩ	47kΩ
20mV	22kΩ 47kΩ 100kΩ	33kΩ 100kΩ Infinity	100kΩ	0	220kΩ
50mV	47kΩ 100kΩ 220kΩ	68kΩ 220kΩ Infinity	220kΩ	0	220kΩ
100mV	100kΩ 220kΩ 470kΩ	150kΩ 470kΩ Infinity	470kΩ	0	220kΩ
200mV	100kΩ	100kΩ	1MΩ	0	220kΩ
500mV	100kΩ	100kΩ	2·2MΩ	0	220kΩ

Microphone Input (1mV for full output)

Approximate Input Impedance	R1	R7
2·2MΩ	Infinity	2·2MΩ
1MΩ	2·2MΩ	2·2MΩ
470kΩ	680kΩ	2·2MΩ
220kΩ	220kΩ	2·2MΩ

Radio Input (Input impedance = R4)

Required input for full output $= \dfrac{50\times R4}{220}$ mV (R4 in kΩ).

Standard value for R4 = 220kΩ (input = 50mV).
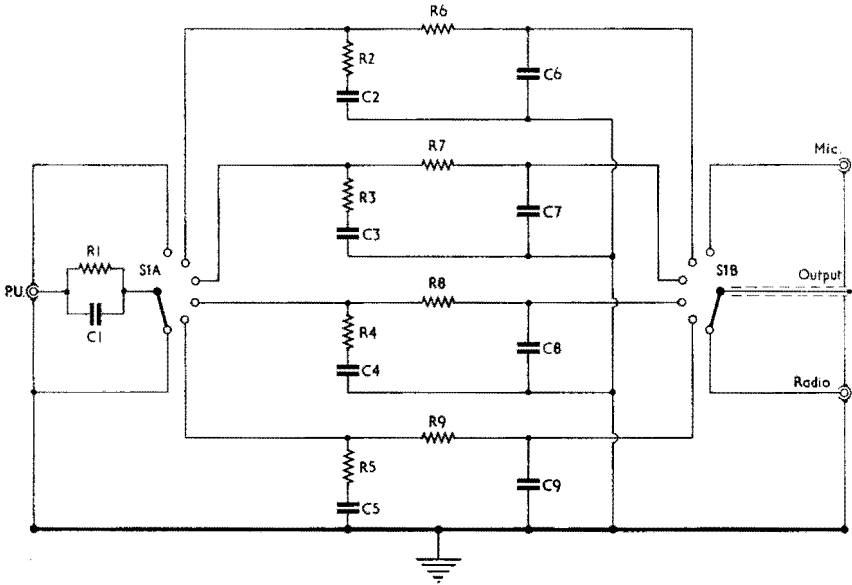
Fig. 3-4. Circuit of the passive input selector unit.   Both switches are shown fully anticlockwise.   The G.E.C. publication *The G.E.C. 912 PLUS* shows how all components may be mounted on a specially designed wafer switch.   Component values are given below.

RESISTORS

(10%, 0·25W)

R1	
R2	470kΩ
R3	
R4	
R5	150kΩ
R6	
R7	
R8	470kΩ
R9	

CAPACITORS

C1	470pF	10%
C2	0·002μF	25%
C3	0·005μF	25%
C4	0·005μF	25%
C5	0·01μF	25%
C6	220pF	10%
C7	100pF	10%
C8	100pF	10%
C9	33pF	10%

SWITCH

S1 2-pole 6-way m.b.b.
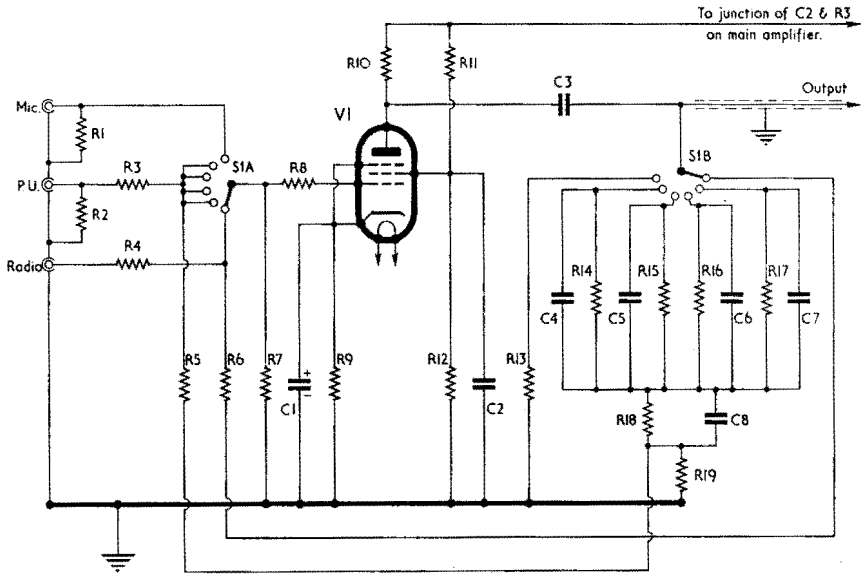
Fig. 3-5. Circuit of the pre-amplifier input selector unit. The equalising components are incorporated in a negative feedback loop (see page 98). The switches are shown fully anticlockwise. Component values are given below.

G.E.C. VALVE

V1 Z729

RESISTORS

(20%, 0·25W unless otherwise shown)

*R1 ⎫
R2 ⎪
R3 ⎬ See Table 3-II
R4 ⎪
R5 ⎭
R6   220kΩ
R7   2·2MΩ
R8   2·2kΩ
*R9   3·3kΩ   5%
*R10 220kΩ   5%   0·5W
*R11 470kΩ   5%
*R12 330kΩ   5%
R13 220kΩ
R14 220kΩ   5%

High Stability

R15 330kΩ   5%
R16 330kΩ   5%
R17 470kΩ   5%
R18 3·3MΩ
R19 See Table 3-II

CAPACITORS

C1 25μF   25V
C2 0·1μF
C3 0·05μF
C4 330pF   10%
C5 150pF   10%
C6 68pF   10%
C7 33pF   10%
C8 1000pF   10%

SWITCH

S1 2-pole 6-way m.b.b.

Using a Separate Pre-Amplifier.—The 912-PLUS may be converted for use with the pre-amplifier described in Section 9-3 (fig. 9-3) in the following way:

Referring to fig. 3-3, the connection to the grid of the left-hand section of V2 is removed so that the whole of the circuit to the left-hand side of V2 is discarded. The " presence " components are removed from the feedback loop, leaving only the series feed resistor R15 and the lower part of the V2 cathode resistance R16. This results in a normal application of negative feedback from the output transformer secondary to the cathode circuit of the first valve of what is now the main amplifier.

The h.t. line is terminated at C10 and C11, R13 is removed and C10, C11 are strapped to give a total capacitance of 24μF. The h.t. series resistor R28 is reduced to 3·3kΩ (1W) and the h.t. connection to the pre-amplifier is taken from the junction of C11 and R28. No additional decoupling is required in the pre-amplifier h.t. line. The heater supply for the pre-amplifier should consist of a separate pair of twisted wires running direct to the mains transformer.

The pre-amplifier connections should be made through an octal socket mounted on the main amplifier chassis near V2 and, to prevent damage to V2 by disconnecting the pre-amplifier when the power is switched on, a grid leak of 1MΩ should be connected between the input grid of V2 and earth.

This combined equipment will have somewhat less overall distortion than the normal 912-PLUS, the second harmonic distortion being, in fact, as low as 0·3%.

The pre-amplifier of fig. 9-2 is also suitable for operation with the 912-PLUS (see page 102).

3-2. A 14W d.c./a.c. Amplifier.—Although the d.c./a.c. amplifier is hardly to be recommended when high quality is the main consideration, fig. 3-6 shows a circuit for use on d.c. or a.c. mains of 190-250V. Two KT33C valves are used in an ultra-linear output stage which provides from 7W to 14W depending upon the mains voltage. About 20db of negative feedback is applied from the output transformer secondary to the cathode circuit of the first valve, a Z729 pentode.

Two U31 half-wave rectifiers are used, the final h.t. voltage being from 180 to 240, depending upon the mains voltage. Resistors R23 and R25, in series with the rectifier cathodes, limit the h.t. voltage on a.c. mains to about the same as that obtained from d.c. mains of the same value. In addition these resistors limit the rectifier peak current and equalise the flow through the two valves. The power supply circuit in fig. 3-6 on 190-250V a.c. or d.c. mains but, for continuous operation on 240-260V a.c., R23 and R25 may be increased to 180Ω each in order to limit the anode and screen dissipation of the output valves.

For d.c. operation only, the rectifiers, resistors R23, R25 and the reservoir capacitor C14 may be omitted. Without rectifiers there is no protection for the electrolytic capacitors against reversed polarity of the mains and paper types should be substituted. A lower capacitance of 4μF will be adequate for C8 and C11 on most d.c. mains.

The current in the heater chain is controlled by a 303 barretter for mains voltages of 190-250 but for continuous operation above 240V a Type 304 should be substituted.

On low mains voltages, the resistances of L1 and T1 are important and should be as low as possible.

Table 3-III (page 36) gives the operating data for the output stage of this amplifier.

COMPONENT VALUES FOR FIG. 3-6

KT33C 14W DC/AC ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 L63/6J5
V3 KT33C
V4 KT33C
V5 U31
V6 U31
V7 Barretter 303

RESISTORS

(20%, 0·5W unless otherwise shown)
R1 500kΩ
R2 250 √speech coil impedance
R3 2·2kΩ
R4 47Ω
R5 220kΩ
R6 1MΩ
R7 15kΩ
R8 2·2MΩ
R9 15kΩ   1W ⎫
R10 15kΩ   1W ⎬ matched to 5%
                ⎭
R11 1·5kΩ
R12 220kΩ
R13 220kΩ
R14 15kΩ
R15 15kΩ
R16 220Ω   5%   1W
R17 220Ω   5%   1W
R18 15kΩ   1W
R19 100Ω
R20 100Ω

R21 470—1500Ω
R22 470—1500Ω
R23 100Ω   5W
R24 62Ω   5W
R25 100Ω   5W

CAPACITORS

C1 0·02μF
C2 25μF   25V
C3 0·1μF
C4 4μF   350V
C5 0·05μF
C6 0·1μF
C7 0·1μF
C8 32μF   350V
C9 25μF   25V
C10 25μF   25V
C11 32μF   350V
C12 1000pF
C13 1000pF
C14 32μF   450V

MISCELLANEOUS

L1 5—10H   200mA
T1 14W Ultra-linear transformer
        3·2kΩ anode-anode
        Primary inductance : ≮25H
    Leakage inductances :
        Prim.—sec. : ≯5mH
        ½ prim—UL tap : ≯5mH
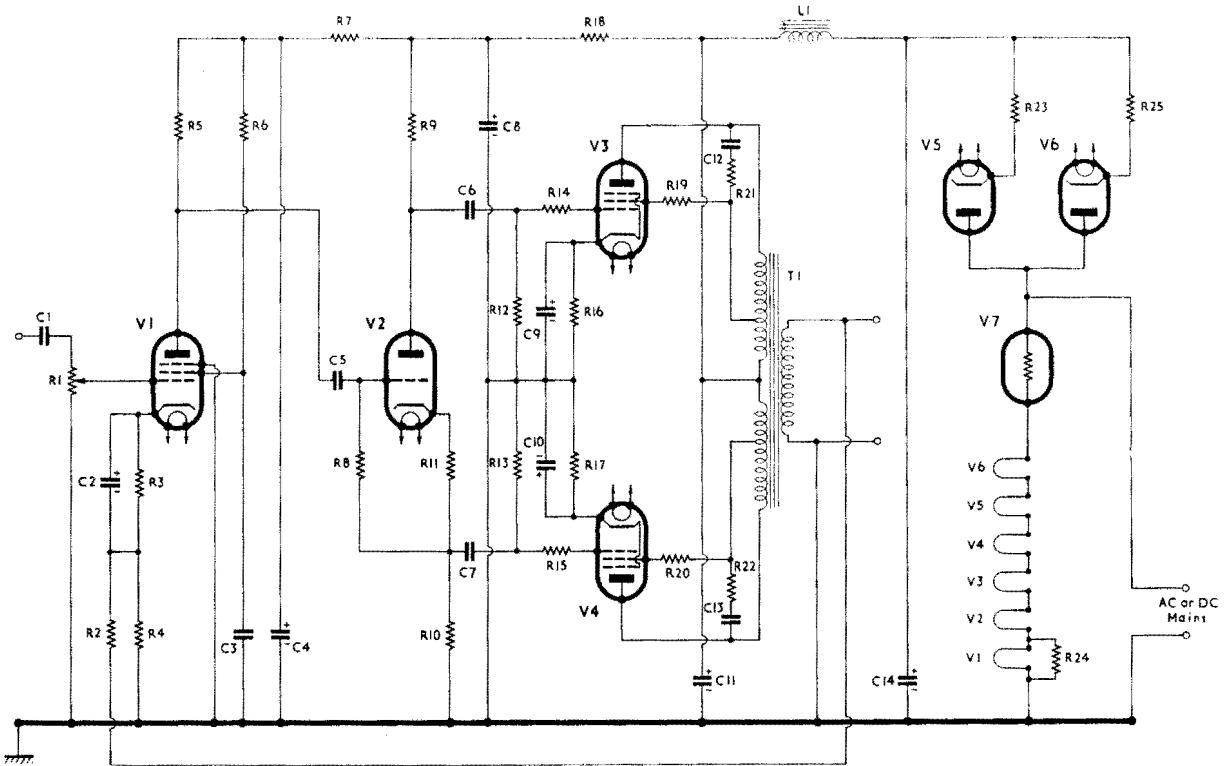
Fig. 3-6.   Circuit of the d.c./a.c. 14W amplifier.

TABLE 3-III

Operating Conditions of the KT33C Output Stage of Fig. 3-6

$V_{a(b)}$ (approx)	235	V
$V_{a, g2}$ (approx)	220	V
$I_{a+g2(o)}$	2×70	mA
$I_{a+g2\,(max\,sig)}$	2×75	mA
$p_{a+g2(o)}$	2×15	W
$p_{a+g2\,(max\,sig)}$ (approx)	2×10	W
R_k	2×220	Ω
V_g	-15	V
P_{out}	14	W
$R_{L\,(a-a)}$	3·2	$k\Omega$
z_{out} (approx)	3	$k\Omega$
D	<0.5	%
$V_{in\,(rms)}$ (to first stage)	0·5	V

CHAPTER 4

AMPLIFIERS OF 15 TO 30 WATTS

Amplifiers giving from 15W to 50W output are adequately catered for by the three beam pentodes, KT55, KT66 and KT88. This chapter gives details of a 14-15W KT66 triode amplifier, two versions of a 30W ultra-linear amplifier (KT66 and KT88), and a 25W KT55 ultra-linear amplifier for d.c. or a.c. mains. Chapter 5 describes amplifiers of 50W to 100W using the KT55 and the KT88.

The KT66 has achieved a world-wide reputation in setting a standard of performance in its class. Introduced in 1937, it was used in very large quantities by the military services in World War II because of its efficiency and reliability and it has since become the basis of many amplifiers with output powers up to 50W. It is widely used in domestic and industrial amplifiers connected in push-pull triode or ultra-linear circuits for outputs of high quality from 15W to about 30W. A well-known example of a push-pull triode design is the " Williamson " amplifier introduced by the M-O. Valve Co. Ltd. in 1947.

The KT88 might be described as a larger version of the KT66 and this chapter includes details of a high quality KT88 30W amplifier suitable for domestic use. This valve was primarily designed for amplifiers of up to 100W output and Chapter 5 gives circuit information for amplifiers of that class and compares the characteristics of the KT88 and the KT66. Appendix A describes the use of the KT88 in multiple pairs for obtaining outputs of several hundred watts.

The KT55 has a 52V, 0·3A heater and was designed specifically to give a high-power output from the low h.t. voltage inherent in the d.c./a.c. amplifier. Details are given of a d.c./a.c. amplifier of 25W output at 1·5% distortion with notes on operation with low mains voltages.

4-1. KT66 Triode Amplifier.—The circuit of the original Williamson amplifier is shown in fig. 4-1*. An output of 14-15W is given at a distortion of 0·1%. Table 4-I gives the operating conditions of the output stage.

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*A complete description of the " Williamson " amplifier and associated tone control circuits is published by Iliffe & Sons Ltd. for *Wireless World*.
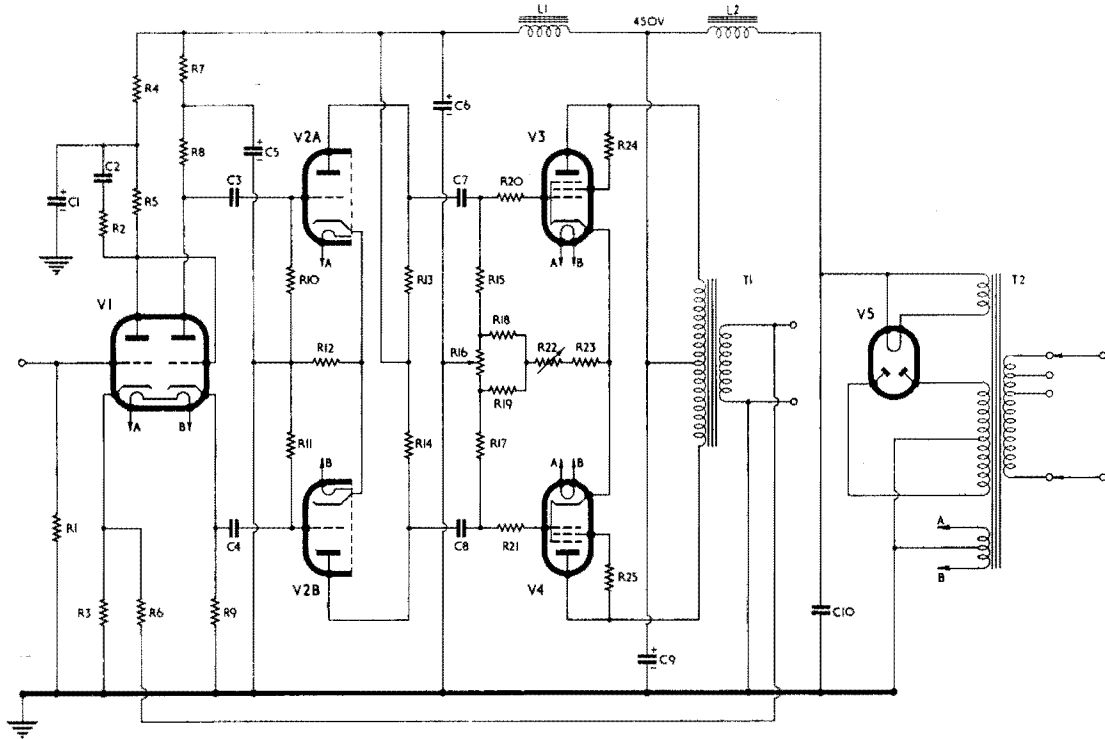
Fig. 4-1. Circuit of the " Williamson " amplifier. R22 is adjusted for a total cathode current of 125mA for the two output valves. The output stage is balanced by connecting a d.c. voltmeter across the whole of the output transformer primary and adjusting R16 for zero reading. C2 and R2 increase the margin of stability at high frequencies.

COMPONENT VALUES FOR FIG. 4-1

" WILLIAMSON " AMPLIFIER

G.E.C. VALVES
V1 B65/6SN7 or 2 × L63/6J5
V2 B65/6SN7 or 2 × L63/6J5
V3 KT66
V4 KT66
V5 U52/5U4 or U54

RESISTORS
(20%, 0·25W unless otherwise shown)
R1 1MΩ
R2 4·7kΩ
R3 470Ω   10%
R4 33kΩ   1W
R5 47kΩ   1W
R6 1200√ speech coil impedance
R7 22kΩ   1W
R8 22kΩ   1W ⎫ Matched to 5%
R9 22kΩ   1W ⎭
R10 470kΩ
R11 470kΩ
R12 390Ω   10%
R13 47kΩ   2W ⎫ Matched to 5%
R14 47kΩ   2W ⎭
R15 100kΩ   10%
R16 100Ω   2W   w.w.
R17 100kΩ   10%
R18 100Ω   1W
R19 100Ω   1W
R20 1kΩ
R21 1kΩ

R22 100Ω   2W   w.w.
R23 150Ω   5%   3W   w.w.
R24 100Ω   0·5W
R25 100Ω   0·5W

CAPACITORS
C1 8μF 500V
C2 200pF
C3 0·05μF
C4 0·05μF
C5 8μF 500V
C6 8μF 500V
C7 0·25μF
C8 0·25μF
C9 8μF 500V
C10 8μF 600V

MISCELLANEOUS
L1 30H 20mA
L2 10H 150mA
T1 14W Output transformer
    10kΩ anode-anode
    Primary inductance : ≮ 100H
    Leakage inductance : ≯ 30mH
T2 Mains transformer
    Secondaries :
    425-0-425V 150mA
    6·3V 4A CT
    5V 3A

TABLE 4-I

OPERATING CONDITIONS OF THE KT66 OUTPUT STAGE OF FIG. 4-1

$V_{a\,(b)}$	450	V
$V_{a,\,g2}$ (approx)	410	V
$I_{a+g2\,(o)}$	$2 \times 62 \cdot 5$	mA
$I_{a+g2\,(max\,sig)}$	$2 \times 72 \cdot 5$	mA
$p_{a+g2\,(o)}$	2×25	W
$p_{a+g2\,(max\,sig)}$	2×21	W
R_k	See fig. 4-1	
V_g (approx)	-37	V
P_{out}	15	W
$R_{L(a\text{-}a)}$	10	kΩ
z_{out}	300	Ω
D	$<0 \cdot 1$	%
$v_{in\,(pk)}$ (approx) (to first stage)	1·9	V

D

4-2. Two 30W Ultra-Linear Amplifiers.—The basic circuit of fig. 4-4 may be used with either KT66 or KT88 valves and the components list on page 43 gives suitable values for each type of valve.

With negative feedback, the KT66 amplifier will give 32W output with about 0·5% distortion at an anode potential of 400V and the KT88 will give 32W with 0·25% distortion at an anode potential of 335V.* The input signal to the first stage of the amplifier for full output in the KT66 version is 600mV whereas the KT88 version requires 500mV. With no negative feedback these figures become 120mV and 100mV respectively. It will be noted that the KT88 version has less distortion, higher sensitivity and requires a lower h.t. voltage than the KT66 for the same maximum output. If feedback is omitted the KT88 amplifier has only 1% distortion at 32W output as against the 2% of the KT66 version. As the KT88 valves are conservatively run in this circuit they will have a long life.

The output stage is preceded by a conventional double triode voltage amplifier which is fed by a triode phase-splitter comprising one half of a further double triode. The other half of this valve is the input stage voltage amplifier, which is directly coupled to the phase-splitter. As in other ultra-linear amplifiers in this book, instability is guarded against by the capacitors and resistors across part of each output transformer half-primary and by grid and screen " stopper " resistors in the output stage. The curves of figs. 4-2 and 4-3 illustrate, respectively, the performance of the KT66 and KT88 output stages without feedback.

Negative Feedback.—14db of negative feedback is used and this is adequate for all normal purposes. This value will reduce the output impedance, distortion and sensitivity of the basic amplifiers by a factor of 5.

Feedback from the output transformer secondary is introduced into the cathode circuit of the first stage via R2. Since the sensitivity of either amplifier without feedback is approximately 100mV, a feedback voltage of about 500mV is required for 14db feedback. As the voltage across the output transformer secondary for 30-32W is about 21·5 for a 15Ω load and about 11V for a load of 4Ω, the resistors R2 and R4 are chosen so that 500mV will exist at their junction at full output. Assuming R4 to be 22Ω, R2 is given by $225\sqrt{Z_o}$ (where Z_o = the loudspeaker impedance) and the nearest standard value may be used. If $Z_o = 15\Omega$, R2 should be 1kΩ and if $Z_o = 4\Omega$, R2 should be 470Ω.

The operating conditions for the output stage of the amplifier of fig. 4-4 are given in Table 4-II.

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*The KT88 output stage is, in fact, operated here almost in Class A, hence the reduced distortion compared with the KT66 version.
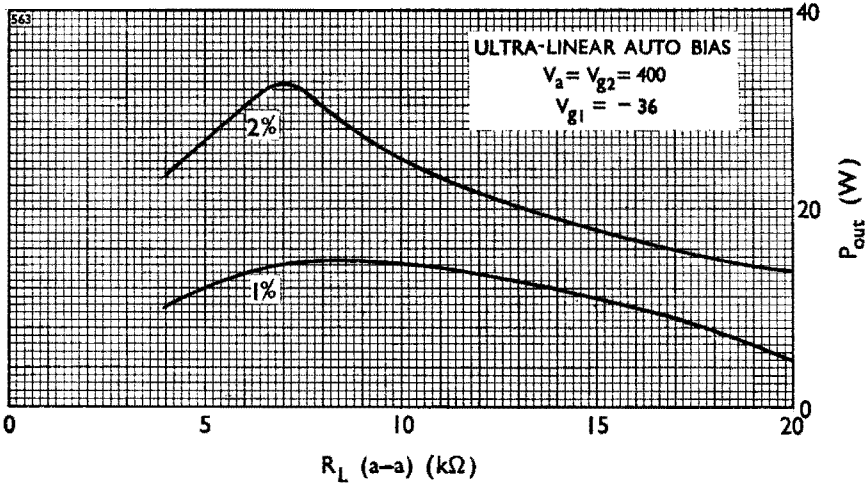
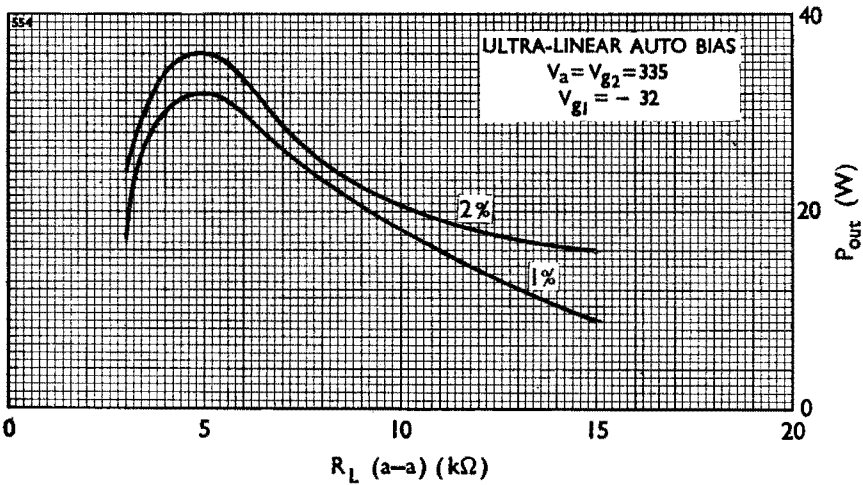**Fig. 4-2.** Performance of the KT66 version of the amplifier in fig. 4-4.

Fig. 4-3. Performance of the KT88 version of the amplifier in fig. 4-4.
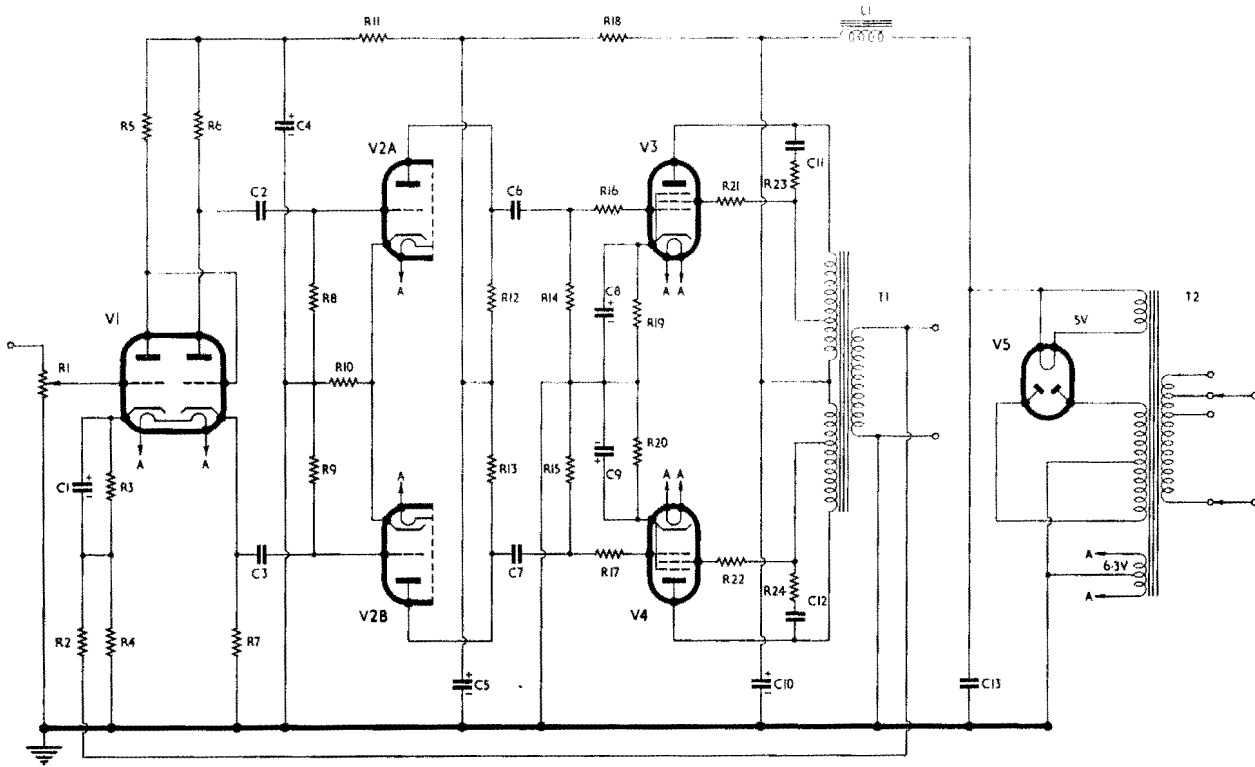
Fig. 4-4. Circuit of a 30W ultra-linear amplifier using KT66 or KT88 valves in the output stage. The component values and circuit conditions differ : see the components list opposite.

COMPONENT VALUES FOR FIG. 4-4
KT88 30W ULTRA-LINEAR AMPLIFIER
Values required for KT66 valves are indicated where necessary

G.E.C. VALVES
V1 B65/6SN7
V2 B65/6SN7
V3 KT88 (KT66)
V4 KT88 (KT66)
V5 U54

RESISTORS
(20%, 0·25W unless otherwise shown)
R1 1MΩ Log.
*R2 225√ speech coil impedance
R3 1kΩ
R4 47Ω
R5 100kΩ
R6 15kΩ 0·5W ⎫
R7 15kΩ 0·5W ⎬ Matched to 5%
R8 470kΩ 10%
R9 470kΩ 10%
R10 1kΩ
R11 22kΩ 1W
R12 33kΩ 10% 1W
R13 33kΩ 10% 1W
R14 220kΩ
R15 220kΩ
R16 10kΩ
R17 10kΩ
R18 4·7kΩ 1W
R19 400Ω (KT66 : 560Ω) 5% 5W
R20 400Ω (KT66 : 560Ω) 5% 5W
R21 270Ω 0·5W
R22 270Ω 0·5W
R23 470—1500Ω 0·5W
R24 470—1500Ω 0·5W
*For 14db feedback.

CAPACITORS
C1 50μF 12V
C2 0·05μF
C3 0·05μF
C4 8μF 350V
C5 8μF 450V
C6 0·05μF
C7 0·05μF
C8 50μF 50V
C9 50μF 50V
C10 8μF 500V
C11 1000pF
C12 1000pF
C13 8μF 500V (KT66 : 600V†)

MISCELLANEOUS
L1 10H 200mA
T1 35W Ultra-linear transformer
    6kΩ anode-anode (KT66 : 7kΩ)
    ‡Primary Inductance : ≮ 50H
    Leakage inductances :
    ‡Prim.—sec. : ≯ 10mH
    ‡‡ prim.—UL tap : ≯ 10mH
T2 Mains transformer
    Secondaries :
    375-0-375V 200mA
    (KT66 : 450-0-450V 150mA)
    6·3V 5A CT (KT66 : 4A)
    5V 3A

†Or two 16μF 350V in series.

‡With these values, R23, R24 and C11, C12 may be omitted.

TABLE 4-II
OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 4-4

	KT66 Valves	KT88 Valves	
$V_{a\ (b)}$	450	375	V
$V_{a,\ g2}$	400	335	V
$I_{a+g2\ (o)}$	$2 \times 62·5$	2×80	mA
$I_{a+g2\ (max\ sig)}$	$2 \times 72·5$	2×85	mA
$P_{a+g2\ (o)}$	2×25	2×27	W
$P_{a+g2\ (max\ sig)}$	2×15	2×12	W
R_k	2×560	2×400	Ω
V_g (approx)	—36	—32	V
P_{out}	32	30	W
$R_{L\ (a-a)}$	7	5	kΩ
z_{out}	1·8	1	kΩ
D	0·5	0·25	%
$V_{in\ (rms)\ (approx)}$ (to first stage)	600	500	mV

If negative feedback is omitted, the last three values are as follows :

z_{out}	9	4·5	kΩ
D	2	1	%
$V_{in\ (rms)\ (approx)}$ (to first stage)	120	100	mV

4-3. A 25W d.c./a.c. Amplifier.—The main problem in the design of an amplifier for d.c./a.c. operation lies in providing adequate output power with a limited h.t. voltage, and the KT55 beam pentode has been specifically designed for this purpose.

Two KT55 valves in push-pull will provide 25W output with a mains supply of 225V.

Circuit Description.—The recommended circuit is shown in fig. 4-5 and it is designed round one Z729, one L63 and two KT55 valves, the h.t. supply being provided by a metal rectifier when the amplifier is operated on a.c. mains.   The use of a thermionic rectifier is impracticable due to the high current required, which is about 275mA.

The input signal is applied to a Z729 voltage amplifier, followed by a conventional triode phase splitter which feeds the KT55 ultra-linear output stage.

The sensitivity of the amplifier without negative feedback is high, full output being obtained for an input of 55mV.   The sensitivity is reduced to 300mV by the application of negative feedback as indicated in fig. 4-5.   Before feedback is applied, the hum and noise level with the volume control at maximum is 55db below full output.

COMPONENT VALUES FOR FIG. 4-5
KT55 25W DC/AC AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 L63/6J5
V3 KT55
V4 KT55
V5 Barretter 303

RESISTORS

(20%, 0·25W unless otherwise shown)
R1 1MΩ Log.
*R2 300 $\sqrt{\text{speech coil impedance}}$
R3 2·2kΩ
R4 22Ω
R5 220kΩ 10% 0·5W
R6 1MΩ 10%
R7 10kΩ
R8 1MΩ
R9 22kΩ 0·5W $\Big\}$ matched to 5%
R10 22kΩ 0·5W
R11 1·5kΩ
R12 220kΩ
R13 220kΩ
R14 10kΩ
R15 10kΩ
R16 185Ω 5% 5W w.w.
R17 185Ω 5% 5W w.w.
R18 10kΩ 10% 1W
R19 47Ω
R20 47Ω
R21 15+15Ω 10% 10W w.w.

For 14 db feedback

R22 Thermistor CZ1 or TH1
R23 63Ω 5% 5W
R24 470-1500Ω 0·5W
R25 470-1500Ω 0·5W

CAPACITORS

C1 0·01μF
C2 50μF 12V
C3 0·1μF
C4 50 pF
C5 0·01μF
C6 0·1μF
C7 0·1μF
C8 16μF 350V
†C9 50μF 25V
†C10 50μF 25V
C11 200μF 275V
C12 1000pF
C13 1000pF
C14 100μF 275V

MISCELLANEOUS

L1 1·5-2H 300mA 75Ω
T1 25W Ultra-linear transformer
    2kΩ anode-anode
    Primary Inductance : ≮ 10H
  Leakage inductances :
    Prim.—sec. : ≯ 20mH
    ½ prim.—UL tap : ≯10mH
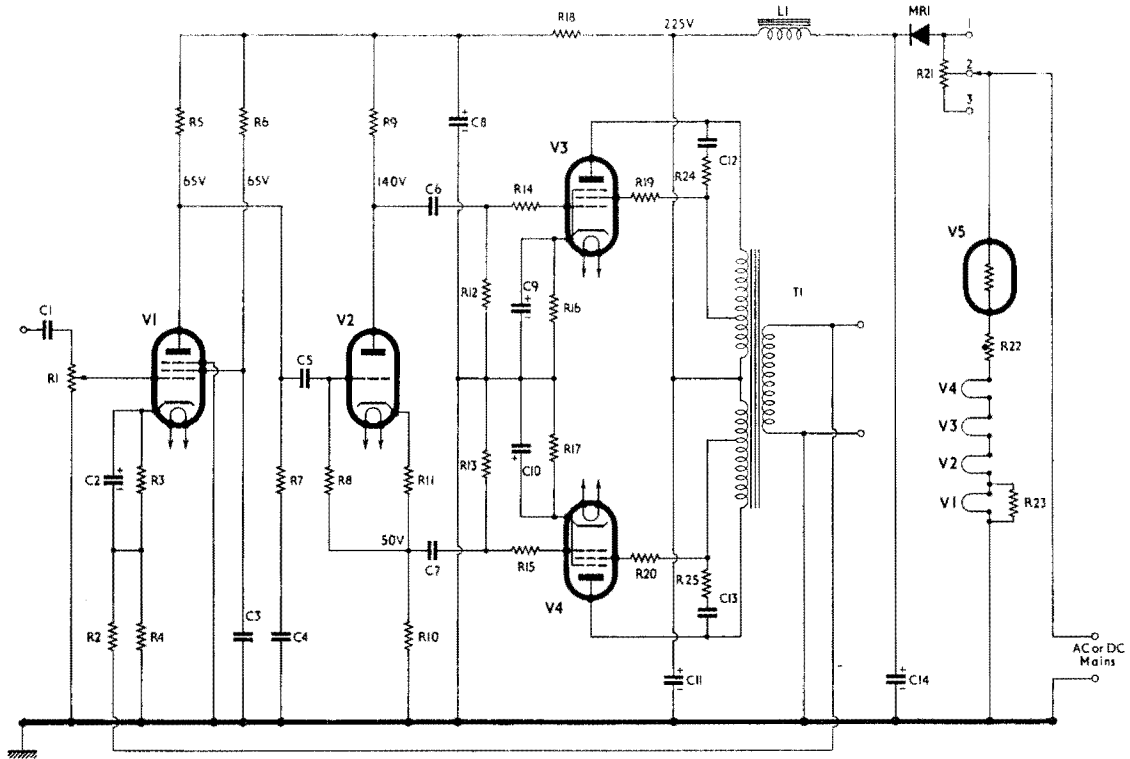MR1 250V 275mA (G.E.C. 13H16XG)

†*See page 47.*

Fig. 4-5. Circuit of the 25W d.c./a.c. amplifier. R7 and C4 increase the margin of stability at high frequencies (see Appendix B, page 119).
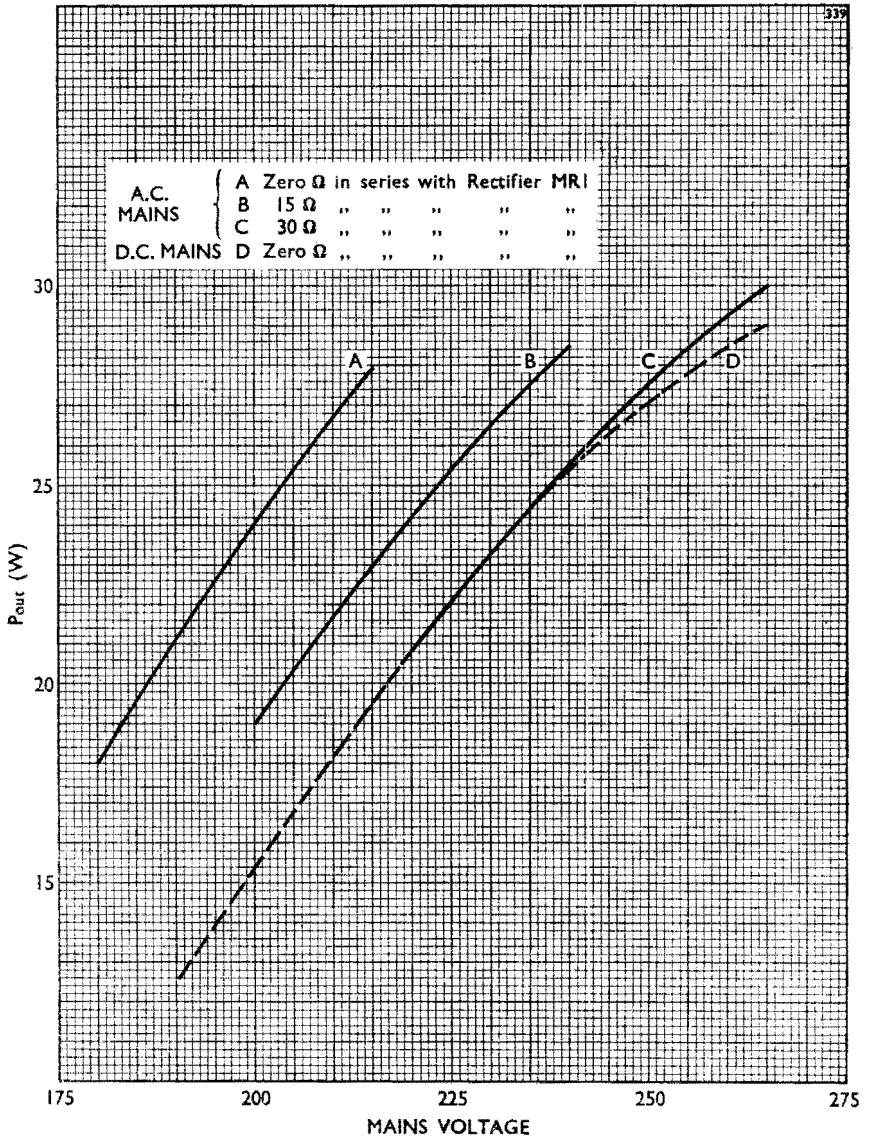
Fig. 4-6. The output power obtained from the amplifier of fig. 4-5 at mains voltages from 180 to 265. The appropriate value of rectifier series resistance is also given.

Separate bias resistors are essential in the output stage in view of the high mutual conductance. The $50\mu F$ cathode bypass capacitors give a loss of 6db at 50 c/s. They may be increased to $250\mu F$ when this loss is objectionable.

The output transformer is tapped at 40% of the turns on each half-primary from the centre tap. The small capacitors, C12, C13 and resistors R24, R25 are required with some output transformers to prevent the possibility of spurious oscillation.

Negative feedback is applied over three stages, about 14db giving a satisfactory reduction in distortion and output impedance without introducing the danger of instability. It is probable that more could be applied with high-quality output transformers but with this value of feedback an output of 25W at 0·25% distortion is obtained. R7 and C4, in the grid circuit of the phase splitter, assist in ensuring stability by reducing the loop gain at ultrasonic frequencies—they result in a loss of 6db at 20kc/s. (See Appendix B, page 119).

A low impedance power supply is obtained by the use of components primarily designed for television receivers. L1, which should have an inductance of about 2H and a d.c. resistance of about 75Ω, is used with two large-value electrolytic capacitors C11 and C14. On d.c. the mains adjusting resistor is not in circuit and the metal rectifier protects the capacitors against reversed polarity. The heater current of 0·3A is controlled by a barretter and thermistor.

The circuit of fig. 4-5 may be changed to pentode operation by simply connecting R19 and R20 to the output transformer centre tap. This will, of course, result in somewhat higher distortion. Compared with pentode operation the ultra-linear circuit reduces the output impedance from $9k\Omega$ to $2·35k\Omega$, giving unity ratio with the anode-to-anode load. The distortion is 1·5% compared with the 2% of pentode operation. The advantages of ultra-linear operation are gained at the expense of a slight fall in the overall sensitivity of the amplifier, that is, it will require a slightly larger input signal to give the same output.

Performance.—The performance of the output stage in fig. 4-5 may be judged from the curves of fig. 4-6 and the data in Table 4-III. The curves, which cover four different groups of mains voltages, were prepared with two considerations in mind : (a) the maximum dissipation of the KT55, and (b) the maximum rating of the metal rectifier. For d.c. mains voltages between 190 and 260 and for a.c. voltages between 190 and 215, no limiting resistance is required in series with the rectifier to meet either of the two conditions (a) or (b). With a.c. mains exceeding 215V, a series resistor (R21) is necessary and this is tapped to cover operation up to 260V.

The curves of fig. 4-6 overlap at the low voltage end in order to convey a complete picture of the performance ; it is suggested that the three mains taps be marked as follows :

1.   190-215V a.c. and 190-260V d.c.
2.   215-235V a.c.
3.   235-260V a.c.

The heater current is maintained within the limits 285mA to 315mA by the barretter Type 303. A satisfactory performance will be obtained over a rather wider range than the usual 200-250V without adjustment. The thermistor prevents surges during switching-on periods when the valves are cold. No thermistor shunt is required.

TABLE 4-III

OPERATING CONDITIONS OF THE KT55 OUTPUT STAGE OF FIG. 4-5

$V_{a\ (b)}$	225	V
$V_{a,\ g2}$	200	V
$I_{a+g2\ (o)}$	2×120	mA
$I_{a+g2\ (max\ sig)}$	2×127	mA
$P_{a+g2\ (o)}$	2×23	W
$P_{a+g2\ (max\ sig)}$	$2 \times 11 \cdot 5$	W
R_k	2×185	Ω
V_g	-22	V
P_{out}	25	W
$R_{L\ (a-a)}$	2	kΩ
Z_{out}	325	Ω
D	$0 \cdot 25\%$	$\%$
$V_{in\ (rms)}$ (to first stage)	$1 \cdot 5$V	mV

If negative feedback is omitted, the last three values become $2 \cdot 35$kΩ, $1 \cdot 5\%$ and 300mV.

Operation from Sub-Normal Mains Voltages.

Low d.c. Mains.—The KT55 may be used successfully on low-voltage d.c. mains supplies, as in some marine installations. The curve in fig. 4-7 shows the expected output on mains voltages of 100-200.

The recommended circuit in this case is similar to fig. 4-5 but, to obtain maximum h.t., the rectifier should be omitted and the output transformer connected to the mains side of the smoothing inductor. C14 also should be omitted and C8 and C11 made 4μF paper capacitors. The output stage bias resistors R16 and R17 should be reduced to 150Ω and the optimum load to $1 \cdot 5$kΩ.

The heater connections depend upon the value of the mains voltage. From 150 to 200V the existing series arrangement may be used with a 305 barretter, which is replaced by a suitable resistor for 125-150V. Below 125V two chains will be required, the KT55 valves being connected in series in one chain with a small resistor, if necessary, and a 303, 304 or 305 barretter in series with the remaining valves in the second chain. A suitable tapped resistor may be used, if preferred, instead of the barretter.
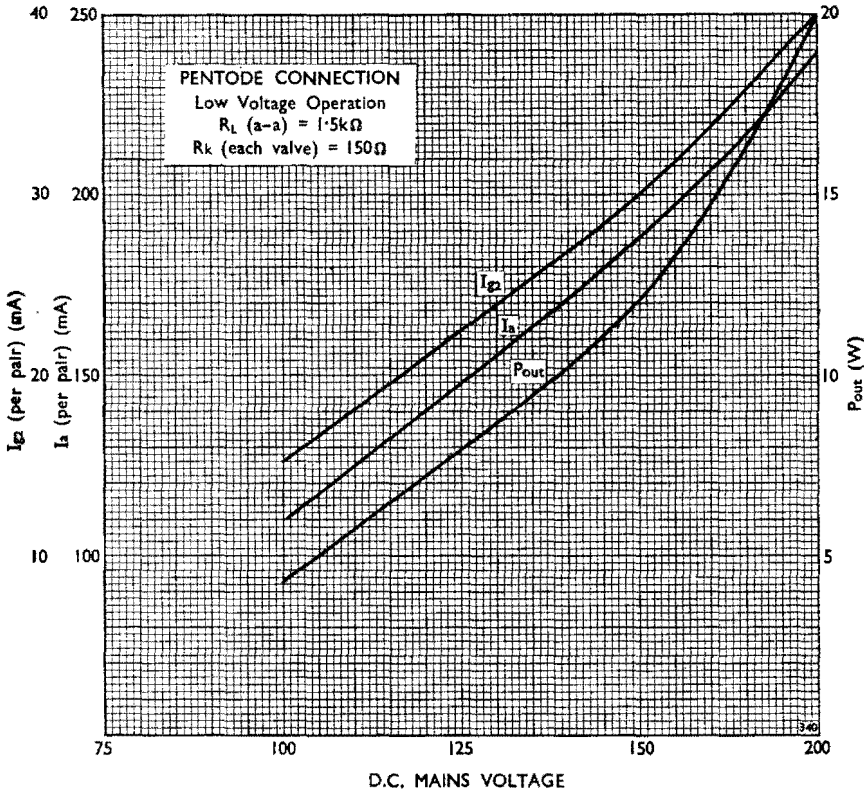
Fig. 4-7. Characteristics of the amplifier of fig. 4-5 on low d.c. mains. The circuit is modified as described opposite. Although drawn for pentode operation, these curves apply also to the ultra-linear circuit but the screen current will be about half that shown.

Low a.c. Mains.—For low a.c. mains voltages an auto-transformer may be used for the h.t. supply, with the heaters connected as for d.c. mains. An alternative method of obtaining the h.t. supply from low a.c. mains is shown in fig. 4-8. A voltage doubler circuit is used to give an h.t. of 220V from 110V a.c.

The smoothing inductor used for the circuit of fig. 4-8 has a lower resistance (20Ω) and the specified rectifier is specially designed for voltage doubling.

The h.t. voltage and power output at various mains voltages is as follows :

AC Mains Voltage	HT Voltage	Output Power
110	200	19W
110	220	23W
120	235	27W

Heater Circuit for Four Valves.—The heaters of four KT55 valves may be connected (via a shunted thermistor) directly to a mains supply between 200 and 215V. For supplies between 215 and 250V, additional series resistance is necessary. See fig. 4-9 for details of the recommended arrangement.
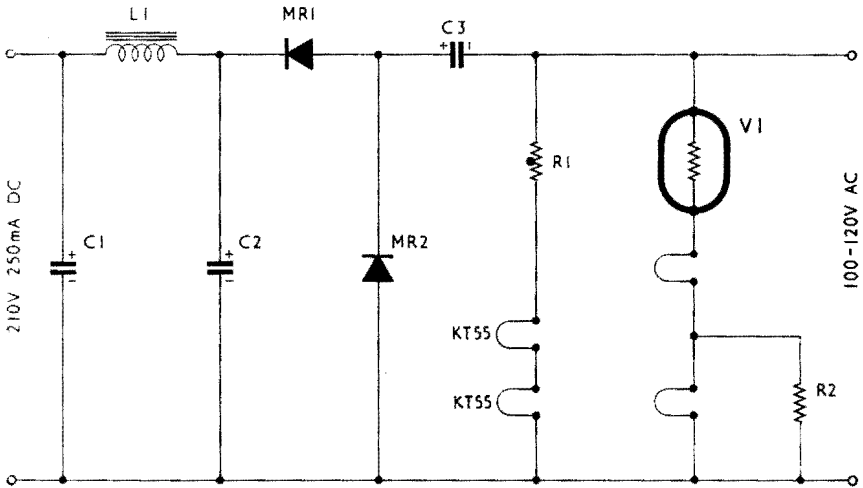
Fig. 4-8. Voltage doubler circuit for the KT55 amplifier. Component values :
V1 : G.E.C. Barretter 303, 304 or 305 (depending upon voltage across heater chain) ;
R1 : Thermistor CZ1 or TH1 ; R2 : 60Ω 5W ; C1 and C3 : 200μF 275V ; C2 : 100μF
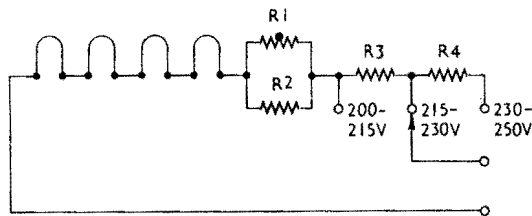275V ; L1 : 1H 20Ω ; MR1 and MR2 : 250V 300mA (G.E.C. 13D8XG).

Fig. 4-9. Heater circuit for four KT55 valves. Component values : R1 : Thermistor
CZ1 or TH1 ; R2 : 750Ω 10% 2W ; R3 and R4 : 50Ω 5% 10W.

CHAPTER 5

AMPLIFIERS OF 50 TO 100 WATTS

Although the amplifiers described earlier in this book, giving outputs up to about 50W, are adequate for domestic purposes and small public address equipment, the demand often arises for a 50W amplifier with a conservative rating and with the possibility of increasing its output to 100W by modifying the circuit and operating conditions. For such amplifiers the KT88 beam pentode is particularly suitable.

This chapter discusses the use of the KT88 in a design for a 50W ultra-linear amplifier with cathode bias and in a 100W fixed bias ultra-linear amplifier.

Details are also given of a relatively simple and inexpensive a.c. amplifier, with KT55 valves in the output stage, which gives 50W output at 5% distortion.

5-1. Comparison of KT66 and KT88.—Compared with the KT66, the KT88 has an increased anode dissipation of 35W, a higher mutual conductance and a cathode of larger emissive area. The physical characteristics are also different in that it is mounted on the more modern wafer octal base which eliminates the glass pinch. The higher anode potential permitted by this type of construction results in a power output from a push-pull pair up to twice that obtainable from the KT66.

Table 5-I overleaf compares the principal characteristics of the two valves and, as they have the same base connections, initial trials of the KT88 in existing KT66 equipment are facilitated.

The difference between the two valves may be further judged from fig. 5-1 in which the two sets of characteristic curves are superimposed.

It is usually possible to substitute the KT88 for the KT66 in most push-pull triode or ultra-linear amplifiers without circuit modification, since the cathode bias resistor is required to be about the same (500-600Ω) for both valves. No great increase in output will be obtained, however, unless the circuit conditions are suitably modified.

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TABLE 5-I

COMPARISON OF KT66 AND KT88

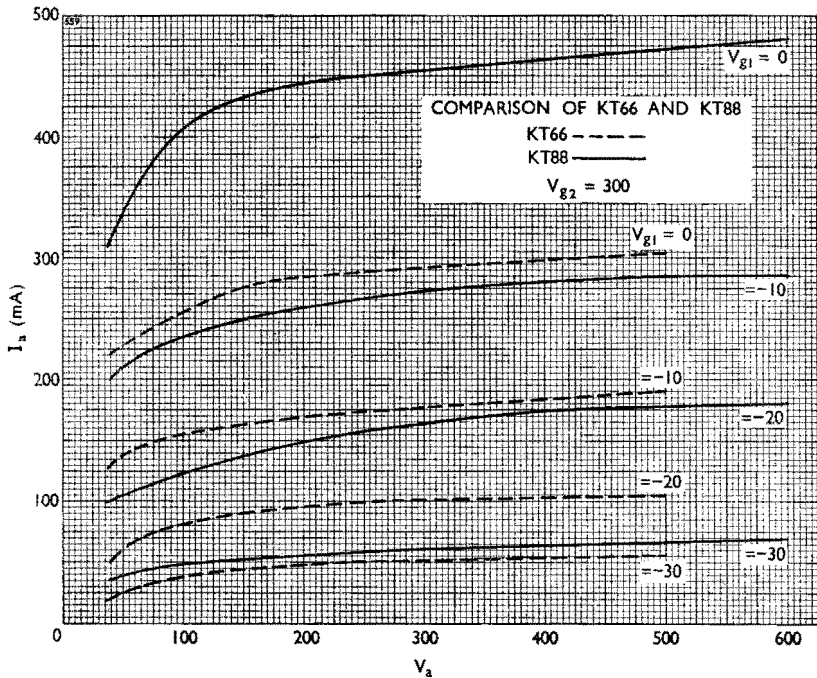|	*KT66*	*KT88*	
V_h	6·3	6·3	V
I_h	1·27	1·8	A
$V_{a\ (max)}$	500	600	V
$V_{g2\ (max)}$	400	600	V
$p_{a\ (max)}$	25	35	W
$p_{g2\ (max)}$	3·5	6	W
g_m	6·3	11	mA/V
*P_{out} (cathode bias, ultra-linear)	32	50	W
*P_{out} (fixed bias, ultra-linear)	50	100	W
P_{out} (cathode bias, triode-connected)	14	27	W
Overall length	135	120	mm
Seated length	120	105	mm
Diameter	52	52	mm

*Class AB1 Push-Pull.

Fig. 5-1. Comparison of the characteristics of the KT66 and KT88.

5-2. KT88 50W Ultra-Linear Amplifier.—The circuit of a complete amplifier with a KT88 ultra-linear output stage giving 50W output at 0.2% distortion is shown in fig. 5-2*. The design follows the practice for other amplifiers in this book but also incorporates the capacitors C8 and C9 shunted by R12 and R13 to ensure stability at frequencies below the cut-off frequency of the output transformer. The desirability of incorporating these components is discussed more fully in Appendix B (page 119). Also shown in fig. 5-2 are networks for reducing the amount of overshoot and consequent " ringing " in the output transformer. These are C6, R10 and C7, R11 between the first two stages. The method of introducing these networks is a variation of the recommendations given in Appendix B for eliminating high-frequency instability in feedback amplifier circuits.

In other respects, the amplifier follows conventional practice. The first double triode, V1, is arranged as a self-balancing floating paraphase phase inverter which feeds V2, the following push-pull voltage amplifier. Potentiometer R22 allows the signal input to the output stage to be adjusted for output stage dynamic balance, as explained on page 8.

The power supply incorporates a thermistor in the h.t. output line in order to reduce the surge from the directly heated rectifier while the remaining valves are warming up. The performance of this amplifier is illustrated in fig. 5-3 and the output stage characteristics are given in Table 5-II below :

TABLE 5-II

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-2

$V_{a\ (b)}$	500	V
$V_{a,\ g2}$	425	V
$I_{a+g2\ (o)}$	2×87	mA
$I_{a+g2\ (max\ sig)}$	2×100	mA
$p_{a+g2\ (o)}$	2×40	W
$p_{a+g2\ (max\ sig)}$	2×18	W
R_k	2×525	Ω
V_g (approx)	-50	V
$\dagger P_{out}$	50	W
$R_{L\ (a-a)}$	5	$k\Omega$
$\dagger D$	0.2%	$\%$
$V_{in\ (rms)}$ (to first stage)	500	mV

*A detailed article describing this amplifier : *Design for a 50-Watt Amplifier* by W. I. Heath and G. R. Woodville was published in the April 1957 issue of *Wireless World*.

†These figures refer to an average pair.

COMPONENT VALUES FOR FIG. 5-2

KT88 50W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES
V1 B339/12AX7
V2 B329/12AU7
V3 KT88
V4 KT88
V5 U52/5U4

RESISTORS
(20%, 0·5W *unless otherwise shown*)
R1 1MΩ
R2 3·3kΩ
R3 100Ω
R4 3·3kΩ
R5 1MΩ
R6 220kΩ 10%
R7 220kΩ 10%
R8 1MΩ
R9 1MΩ
R10 10kΩ
R11 10kΩ
R12 4·7MΩ
R13 4·7MΩ
R14 100kΩ
R15 470kΩ
R16 470kΩ
R17 10kΩ
R18 680Ω
R19 10kΩ
R20 4·7kΩ
R21 33kΩ 2W
*R22 25kΩ 4W
R23 33kΩ 2W
R24 220kΩ
R25 220kΩ
R26 10kΩ
R27 10kΩ
R28 525Ω 5% 6W
R29 525Ω 5% 6W
R30 100Ω
R31 100Ω

Optional potentiometer for dynamically balancing the output stage. If omitted, R21 and R23 should be 47kΩ, 2%, 2W.

R32 100kΩ 1W
R33 100kΩ 1W
R34 Thermistor CZ6
†R35 6·8kΩ
†R36 6·8kΩ

CAPACITORS
C1 8μF 500V
C2 50μF 12V
C3 50μF 12V
C4 0·25μF
C5 0·25μF
C6 470pF
C7 470pF
C8 0·005μF
C9 0·005μF
C10 0·5μF
C11 0·5μF
C12 50μF 100V
C13 50μF 100V
C14 24μF 350V
C15 24μF 350V
C16 4μF 750V
†C17 8μF 500V

MISCELLANEOUS
L1 5H 250mA
T1 50W Ultra-linear transformer
   5kΩ anode-anode
   Primary inductance : ≮30H
   Leakage inductances :
   Prim.—sec. : ≯ 10mH
   ½ prim.—UL tap : ≯ 10mH
T2 Mains transformer
   Secondaries:
   500-0-500V 250mA
   6·3V 5A CT
   6·3V 1-2A CT (for pre-amp.)
   5V 3A
F1 1A delayed fuse

†*R35, R36 and C17, shown dotted in fig. 5-2, provide the decoupled h.t. feed for the pre-amplifier of fig. 9-3 which was designed to work with this amplifier.*
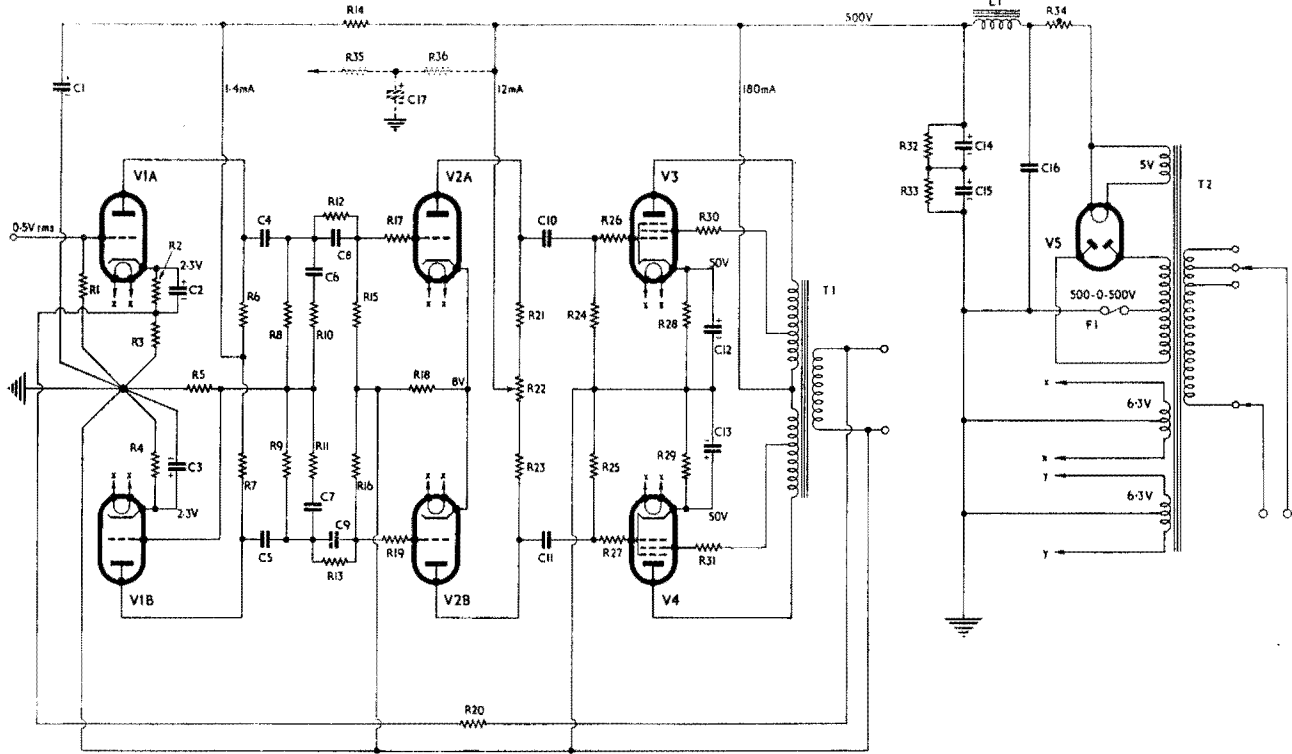
Fig. 5-2. Circuit of the KT88 50W amplifier (the " 88-50 "). The components shown in broken line provide a decoupled h.t. feed for the pre-amplifier of fig. 9-3 (page 105).
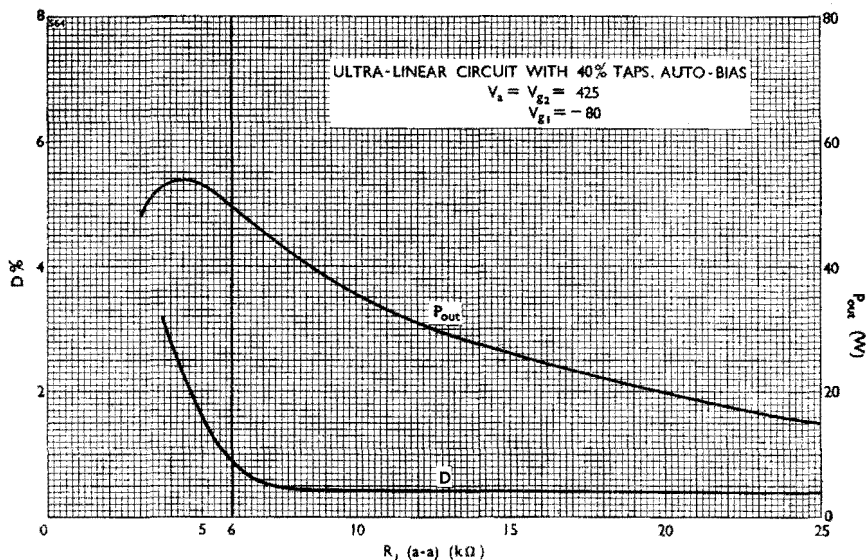
Fig. 5-3. Performance of the KT88 amplifier of fig. 5-2.

5-3   KT55 50W Fixed Bias Ultra-Linear Amplifier.—Although the KT55 was designed primarily for d.c./a.c. amplifiers, its characteristics lend themselves well to the design of a 50W a.c. mains amplifier. This circuit, illustrated in fig. 5-5, provides an economical method of obtaining 50W output. The use of fixed bias in the ultra-linear arrangement enables a pair of KT55 valves to give 50W output at about 5% distortion with an h.t. potential of only 300V.

There is a large change in anode current between the quiescent and full output conditions and a power supply of very low impedance is necessary. The circuit diagram includes details of a single-phase half-wave rectifier power supply having an impedance of 100Ω over the working range. The grid bias voltage is rectified by a small rectifier from the 52V heater supply for the KT55 valves.

The h.t. supply is derived from an auto-transformer which provides 255V r.m.s. for the metal rectifier MR1. Two other windings of 52V and 6·3V supply the heaters of the KT55 valves and earlier valves, respectively, and a small metal rectifier MR2 provides the grid bias potential for the output stage. The range of negative bias required is from 50V to 60V.

The auto-transformer does not need to be of large size since it has only to cater for the voltage difference between the mains and the rectifier supply of 255V.   It is this feature of the auto-transformer that is responsible for the excellent regulation of the power supply circuit.   Satisfactory results are obtained from a transformer similar in size to those used in domestic radio receivers.

The capacitance-input filter contains two large capacitors of $160\mu F$ and a low-resistance 1H smoothing inductor and provides a hum-free d.c. supply.

The correct quiescent current of 70-75mA is set for each valve on the meter in the cathode circuit by adjusting potentiometers R8 and R9.   Resistors R17 and R18 are meter shunts, the value of which will depend upon the meter characteristics and the type of meter circuit adopted.   The potentiometer R5 in the anode circuit of the first stage is adjusted at 90% of maximum output to give equal cathode currents in the output stage for obtaining dynamic balance. Resistor R19, in the bias line, reduces the potential at C8 to 60V from the 70V it would otherwise be.

The curves of fig. 5-4 illustrate the performance of this amplifier and Table 5-III gives the characteristics of the output stage.
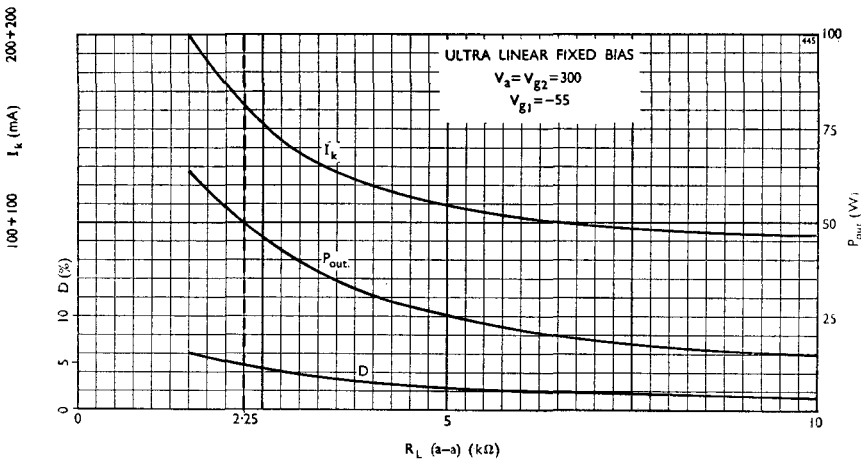

Fig. 5-4.   Performance of the KT55 amplifier of fig. 5-5.

TABLE 5-III

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-5

$V_{a, g2 (o)}$	325	V
$V_{a, g2 (max sig)}$	300	V
$I_{a+g2 (o)}$	2×75	mA
$I_{a+g2 (max sig)}$	2×165	mA
$p_{a+g2 (o)}$	2×25	W
$p_{a+g2 (max sig)}$	2×12.5	W
V_{g1}	-50 to -55	V
P_{out}	50	W
$R_{L (a-a)}$	2.25	$k\Omega$
z_{out}	2.7	$k\Omega$
*D	5	%
$V_{in (rms)}$ (to first stage)	2×2.5	V

*Assuming the cathode currents are balanced at 90% of maximum output by R5

COMPONENT VALUES FOR FIG. 5-5

KT55 50W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 B65/6SN7
V2 KT55
V3 KT55

RESISTORS

(20%, 0·25W *unless otherwise shown*)
R1 1MΩ
R2 1MΩ
R3 1kΩ
R4 33kΩ   1W
R5 20kΩ   w.w. preset
R6 33kΩ   1W
R7 100kΩ   10%
R8 10kΩ   w.w. preset
R9 10kΩ   w.w. preset
R10 100kΩ   10%
R11 10kΩ   10%   1W
R12 10kΩ
R13 10kΩ
R14 4·7kΩ   10%   1W
R15 270Ω
R16 270Ω
R17 ⎫
R18 ⎭ Meter shunts

R19 150Ω
R20 470-1500Ω   0·5W
R21 470-1500Ω   0·5W

CAPACITORS

C1 16μF   350V
C2 0·1μF
C3 0·1μF
C4 160μF   450V
C5 1000pF
C6 1000pF
C7 160μF   450V
C8 32μF   100V

MISCELLANEOUS

L1 1H   350mA   20Ω
T1 50W Ultra-linear transformer
    2·25kΩ anode-anode
    Primary inductance : ≯ 12H
    Leakage inductances :
    Prim.—sec. : ≯ 10mH
    ½ prim.—UL tap : ≯ 10mH
T2 Mains auto-transformer (see text)
MR1 275V   350mA   (G.E.C. Z13H17XG)
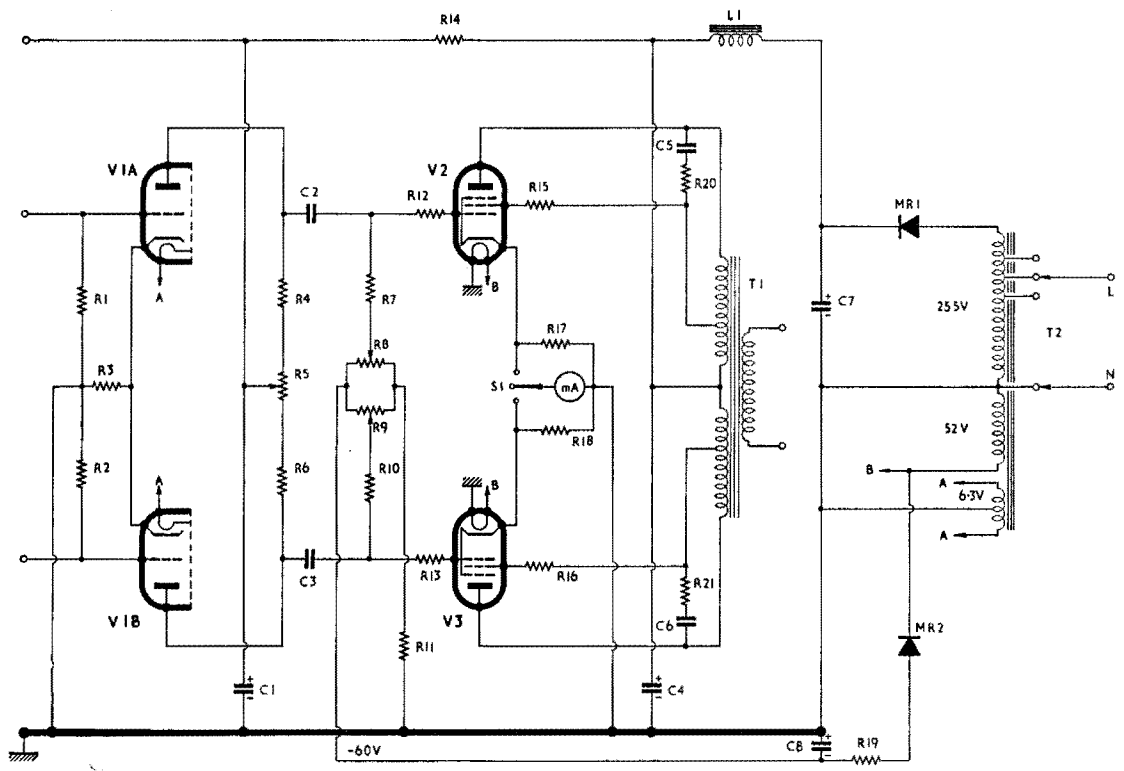MR2 60V   10mA   (G.E.C. Z11H4X)
S1 1-pole 3-way

Fig. 5-5. Circuit of the KT55 50W amplifier. The earth line or chassis is connected to one side of the mains supply. If this amplifier is preceded by a triode phase-splitting stage such as that in fig. 5-8, the required signal input will be 2·5V.

5-4.   KT88 100W Fixed Bias Ultra-linear Amplifier.—The circuit of this amplifier is given in fig. 5-8 and it follows, in general, the main details of the KT55 50W amplifier described in the previous section.   Points of difference include the addition of a phase-splitter at the input and, of course, the h.t. power and grid bias supplies.

This amplifier provides 100W output at 5% distortion with an h.t. potential of 560V and its performance is shown graphically in fig. 5-6.   Table 5-IV gives the characteristics of the output stage.

With fixed bias, the large change in anode current necessitates a low impedance power supply and, with normal rectifier circuits, an inductance-input smoothing filter is essential.   The smoothing capacitor should be of high value to prevent an instantaneous fall in h.t. potential upon the occurrence of a transient signal.   Satisfactory performance will be obtained with a single inductor and a capacitance of 50-150μF.   The circuit diagram shows two 160μF 450V electrolytic capacitors in series as an economical method of obtaining the required capacitance.   Reducing the h.t. potential to 460V, the load impedance
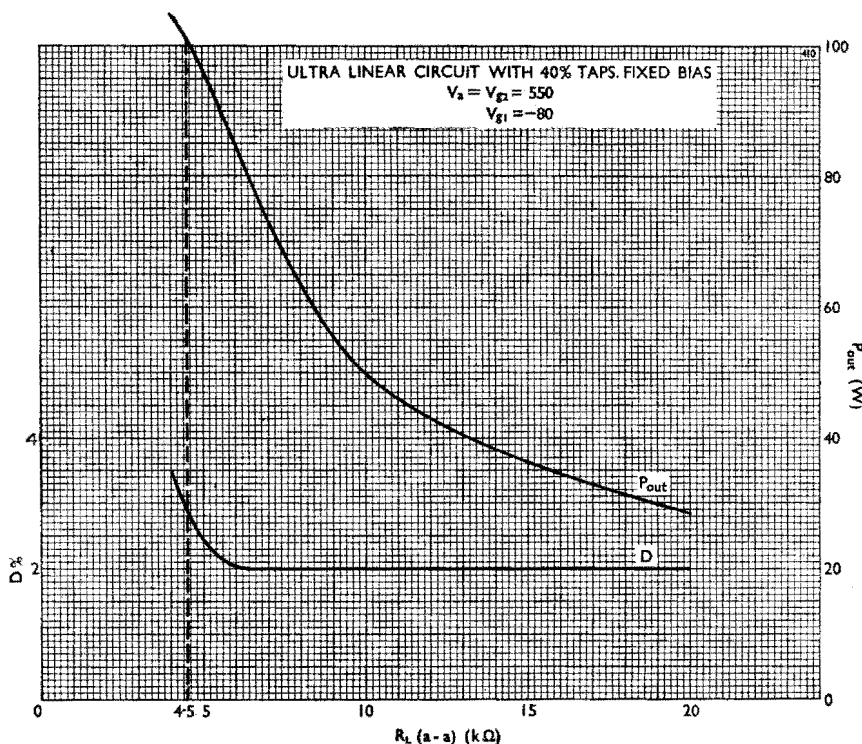

ULTRA LINEAR CIRCUIT WITH 40% TAPS. FIXED BIAS
$V_a = V_{g_2} = 550$
$V_{g_1} = -80$

Fig. 5-6.   Performance of the KT88 100W amplifier of fig. 5-8.

to 4kΩ and the grid bias to −65V results in an amplifier giving 65W output. The performance of this version is shown in fig. 5-7, while the output stage characteristics are included in Table 5-IV.
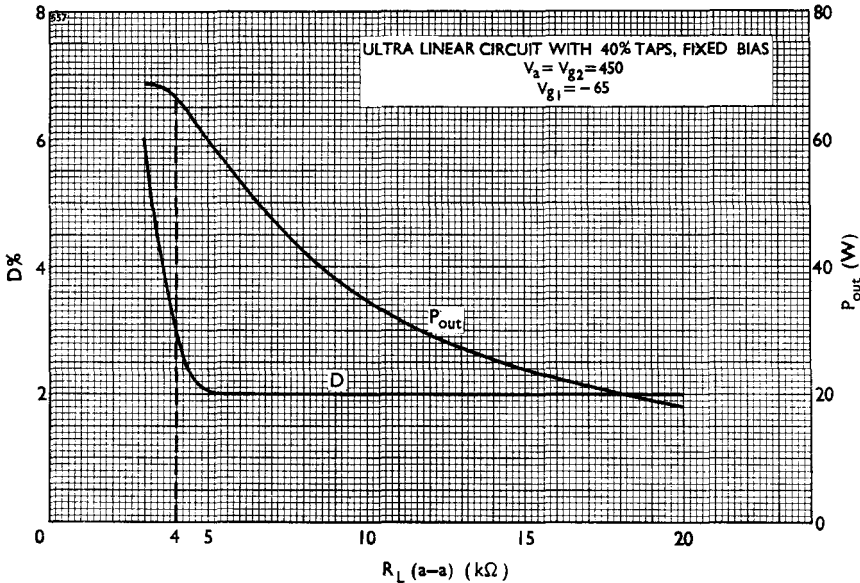

Fig. 5-7. The modified characteristics of the amplifier in fig. 5-8 when it is arranged to give 65W maximum output (see Table 5-IV).

TABLE 5-IV

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-8

$V_{a\ (b)}$	460	560	V
$V_{a,\ g2}$	450	550	V
$I_{a+g2\ (o)}$	2×50	2×50	mA
$I_{a+g2\ (max\ sig)}$	2×120	2×150	mA
$p_{a+g2\ (o)}$	2×25	2×30	W
$p_{a+g2\ (max\ sig)}$	2×20	2×33	W
*V_{g1}	−65	−80	V
P_{out}	65	100	W
$R_{L\ (a-a)}$	4	4·5	kΩ
z_{out}	6·5	6·5	kΩ
†D	3-6	3-6	%
$V_{in\ (rms)}$ (to first stage)	1 to 1·5	1 to 1·5	mV

*A bias voltage range of at least $\pm25\%$ is recommended.

†The distortion will vary according to the degree of matching by R11, as explained on page 8.

Protection against Bias Failure.—Should the bias supply fail, the KT88 anode currents would increase excessively and it is recommended that some device be incorporated for protecting the output valves in the event of bias failure. The arrangement illustrated in fig. 5-9 inserts a suitable resistor into the output stage cathode circuit which will enable the amplifier to function temporarily at half maximum output.

A triode, which could be one half of a double triode used also in the first stage of the amplifier, is connected in series with a relay across the main h.t. supply. The relay contacts are normally closed and short-circuit the emergency cathode resistor R1. The triode is held at cut-off by the connection of its grid to the bias supply at a point about 50V negative to earth. Should the bias fail, the grid of the triode will rise to earth potential and current will flow through the triode. This energises the relay, the contacts of which will open and bring into circuit the cathode bias resistor.

COMPONENT VALUES FOR FIG. 5-8
KT88 100W FIXED BIAS AMPLIFIER

G.E.C. VALVES
V1 L63/6J5
V2 B65/6SN7
V3 KT88
V4 KT88
V5 ⎱
V6 ⎰ U19 (or GXU50 with delay)

RESISTORS
(20%, 0·25W *unless otherwise shown*)
R1 1MΩ
R2 1·5kΩ
R3 33kΩ 1W ⎱ matched to 5%
R4 33kΩ 1W ⎰
R5 470kΩ 10%
R6 470kΩ 10%
R7 33kΩ 1W
R8 1kΩ
R9 33kΩ 10% 1W
R10 10kΩ
R11 33kΩ 10% 1W
R12 100kΩ 10% 0·5W
R13 20kΩ w.w. preset
R14 20kΩ w.w. preset
R15 100kΩ 10% 0·5W
R16 10kΩ 10% 1W
R17 4·7kΩ 1W
R18 5·6kΩ
R19 5·6kΩ
R20 ⎱ Meter shunts
R21 ⎰
R22 270Ω 0·5W
R23 270Ω 0·5W
R24 470-1500Ω 0·5W
R25 470-1500Ω 0·5W
R26 100kΩ 10% 1W
R27 100kΩ 10% 1W
R28 10kΩ 0·5W

CAPACITORS
C1 0·01μF
C2 0·05μF
C3 0·05μF
C4 8μF 350V
C5 8μF 450V
C6 0·1μF
C7 0·1μF
C8 1000pF
C9 1000pF
C10 8μF 250V
C11 160μF 450V ⎱ each 100+60 dual
C12 160μF 450V ⎰
C13 8μF 250V
C14 0·01μF 750V

MISCELLANEOUS
L1 5H 325mA
L2 20H 10mA
T1 100W Ultra-linear transformer
    4·5kΩ anode-anode (100W)
    4kΩ anode-anode (65W)
    Primary inductance : ≮ 40H
    Leakage inductances :
    Prim.—sec. : ≯ 6mH
    ½ prim.—UL tap : ≯ 6mH
T2 Mains transformer
    Secondaries:
    700-0-700V 325mA
    6·3V 5A CT
    5V 7A
    75V 10mA (bias)
S1 1-pole 3-way
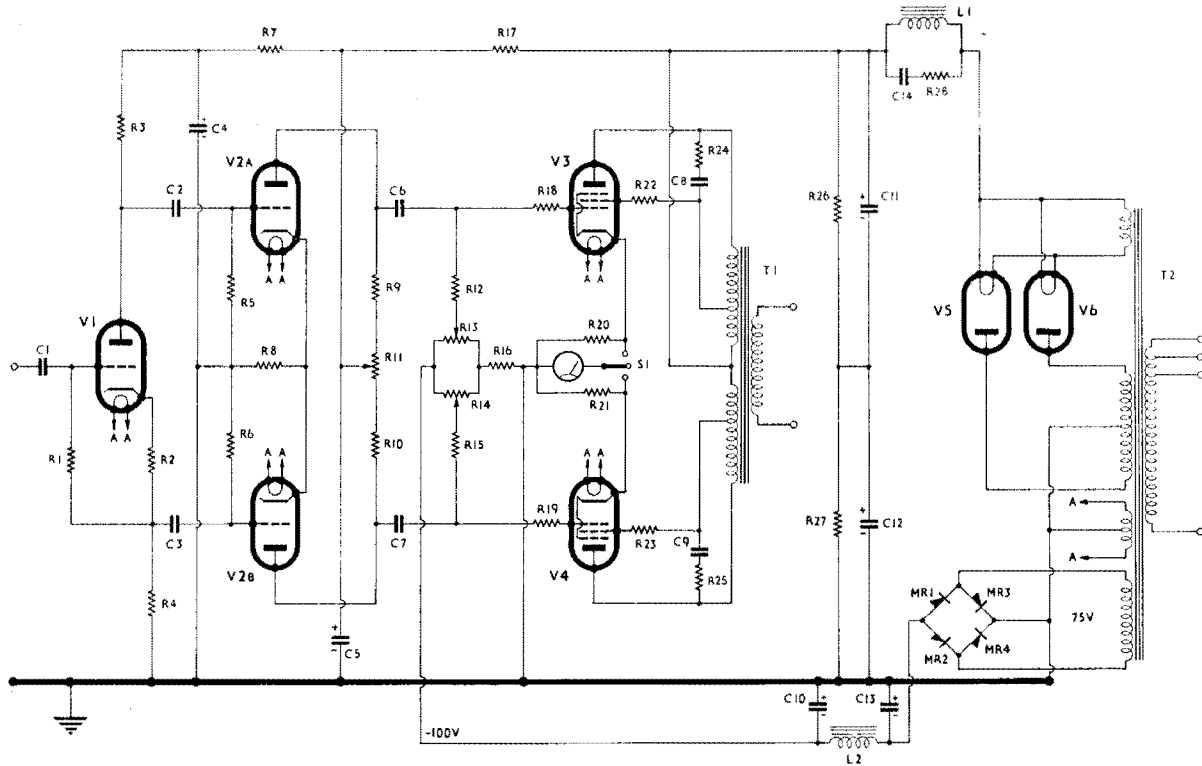MR1 ⎫
MR2 ⎬ 75V 10mA
MR3 ⎪
MR4 ⎭

Fig. 5-8. Circuit of the KT88 fixed bias amplifier. Table 5-IV gives the operating conditions for obtaining either 65W or 100W maximum output from this circuit. R26 and R27 equalise the voltages across C11 and C12, the series-connected smoothing capacitors. C14 and R28 prevent the build-up of high voltage transients across L1.
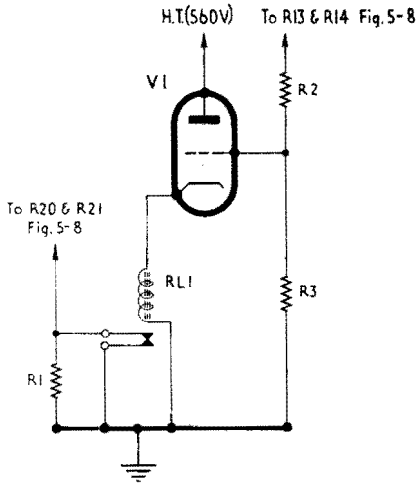
Fig. 5-9. Bias failure protection for the KT88 fixed bias amplifier. R2 is connected to the positive (earthy) junction of R13 and R14 in fig. 5-8. R2 and R3 take the place of R16 in the amplifier. The earth connection to the meter in fig. 5-8 is broken and the meter (and R20, R21) taken to R1. Component values : V1 : G.E.C. L63/6J5GT ; R1 : 330Ω 10% 5W ; R2 : 6·8kΩ 10% 0·5W ; R3 : 15kΩ 10% 0·5W ; RL1 : 20kΩ.

CHAPTER 6

CLASS B AMPLIFIERS OF 175 TO 200 WATTS

The Class B amplifier is recognised as an economical method of obtaining a high audio output because of its high efficiency and the simplicity of the circuit arrangement. The absence of a separate grid bias supply renders the valve self-protective and eliminates the need for an interlock for delaying the application of h.t. until bias is established. It is necessary to qualify this statement in view of the small degree of bias used in the output stage of the amplifier of fig. 6-5. In this case, however, the bias serves only to limit the quiescent dissipation and, unlike normal fixed bias arrangements, no harm results from running the output stage for short periods of a few minutes without this bias being fully established.

Much of the distortion associated with the Class B amplifier in the past was due to the leakage inductance of the driver stage coupling transformer but this has now been largely overcome by using the cathode-coupled bridged transformer. This arrangement was discussed in more detail in Section 1-3.

This chapter describes 175W and 200W amplifier circuits containing the DA42, an indirectly heated triode designed specifically for Class B operation.

6-1. DA42 175W Amplifier.—The complete amplifier circuit is shown in fig. 6-1 and comprises—apart from the output stage—a KT66 driver stage, which supplies the necessary 75+75V for full output, and a push-pull intermediate stage consisting of two N709 pentodes. The resistance-coupled N709 valves are required to deliver 80+80V, and this output is obtained at low distortion because of the high h.t. voltage available, provided the screen and cathode resistors are of optimum value. The input stage may utilise any small triode such as the DH77 shown, which draws an anode current of 2mA.

The combined anode and screen current of each N709 is 6mA, the anode and screen potentials being 150 and 60, respectively. An input not exceeding 1+1V to this stage gives an output of 80+80V at less than 2% distortion.

With 500V h.t. supply for the KT66 valves a bias of 45V is required, giving an anode current of 70-80mA for the pair. A common cathode resistance of 600Ω is necessary to give this bias and this is partly supplied by the resistance of

65

the half-primaries of the coupling transformer. An additional resistor (R19), common to both valves, is usually necessary. In the transformer used for the prototype, the resistance of each half-primary was about 300Ω and R19 was found to require a value of 440Ω.

The coupling transformer bridging capacitors are not critical in value within the limits of $2\mu F$ to $16\mu F$.

The output transformer should have low leakage inductance between the two half-primaries and between the primary and secondary. For a 100V output line (60Ω in this case), the primary/secondary ratio is 13:1.

The amplifier of fig. 6-1 gives 175W output at 6% distortion for an input of less than 1V r.m.s. and the performance of the output stage is displayed by the curves of fig. 6-3. Table 6-I gives the operating conditions.

COMPONENT VALUES FOR FIG. 6-1

DA42 175W AMPLIFIER

G.E.C. VALVES
V1 DH77/6AT6
V2 N709
V3 N709
V4 KT66
V5 KT66
V6 DA42
V7 DA42

RESISTORS
(20%, 0·25W *unless otherwise shown*)
R1 470kΩ
R2 1·5kΩ
R3 22kΩ 0·5W ⎱ matched to 5%
R4 22kΩ 0·5W ⎰
R5 470kΩ 10%
R6 470kΩ 10%
R7 10kΩ
R8 10kΩ
R9 47kΩ 0·5W
R10 220Ω 0·5W
R11 470kΩ 10% 0·5W
R12 47kΩ 5% 5W w.w.
R13 47kΩ 5% 5W w.w.
R14 4·7kΩ 10% 1W
R15 470kΩ
R16 470kΩ

R17 10kΩ
R18 10kΩ
R19 (See text above)
R20 100Ω
R21 100Ω
R22 5kΩ 10% 10W
R23 5kΩ 10% 10W

CAPACITORS
C1 0·01μF
C2 0·01μF
C3 0·01μF
C4 4μF 450V
C5 0·5μF
C6 16μF 450V
C7 0·05μF
C8 0·05μF
C9 16μF 100V
C10 16μF 100V
C11 2000pF 1kV
C12 2000pF 1kV
C13 150 pF

TRANSFORMERS
T1 Bridged transformer (see fig. 6-7)
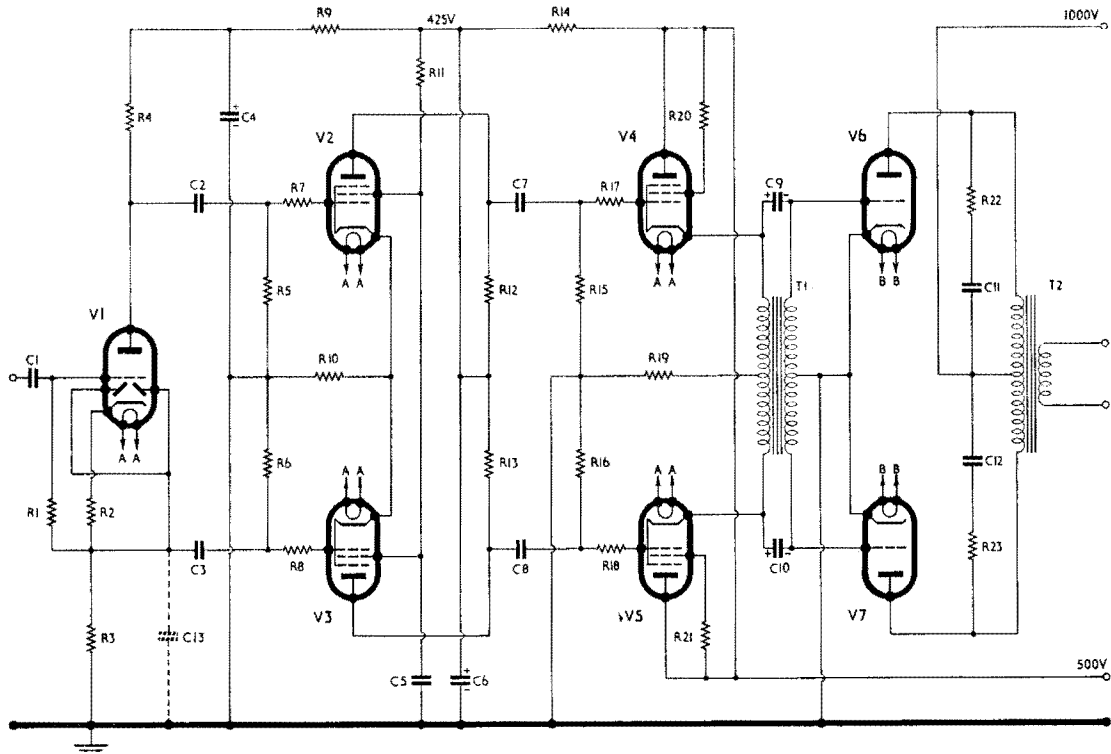T2 Output transformer (see fig. 6-8)

Fig. 6-1. Circuit of the DA42 175W Class B amplifier. The slight unbalance in signal in the output of V1, caused by phase shift, may be eliminated by shunting R3 with a capacitor C13, shown dotted.

TABLE 6-I

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 6-1

$V_{a (b)}$	1 (approx)	kV
V_a	1	kV
$I_{a (o)}$	2×25	mA
$I_{a (max sig)}$	2×140	mA
$p_{a (o)}$	2×25	W
$p_{a (max sig)}$	2×50	W
V_g	0	V
$I_{g (max sig)}$	2×25	mA
P_{out}	175	W
$R_{L (a-a)}$	10	kΩ
z_{out}	15	kΩ
D	6	%
p_{dr}	5	W
$V_{in (g-g) (pk)}$	175	V
$V_{in (rms)}$ (to first stage)	$\not> 1$	V

These conditions should be used only for normal speech and music. A continuous tone at maximum output will exceed the permitted dissipation and for such applications $R_{L (a-a)}$ must be increased by at least 20%.

The addition of negative feedback, as described below, results in the following conditions :

P_{out}	175	W
z_{out}	3 to 4	kΩ
D	$\not> 2$	%
$V_{in (rms)}$ (to first stage)	$\not> 5$	V

Negative Feedback.—An improved performance is obtained at the expense of a reduction in sensitivity by adding negative feedback. The circuit modifications are given in fig. 6-2 and this arrangement applies 12-14 db feedback, which provides a damping factor of about 4·5 and reduces the distortion to 2% at 175W.

The coupling transformer T1 introduces negligible phase shift so that feedback may be taken from the output transformer secondary to the N709 cathode circuit. The single cathode bias resistor is replaced by R1 and R2, and resistors R3 and R4 provide the feedback paths from each end of the transformer secondary.

The values of R3 and R4 are determined by the load impedance but, with a 100V line (60Ω), they are each 3·3kΩ.

Fig. 6-2. Adding negative feedback to the DA42 175W amplifier. Component values : R1 and R2 : 330Ω 10% 0·5W ; R3 and R4 : 3·3kΩ 10% 1W (for 100V line output). Other designations are as shown on fig. 6-1.
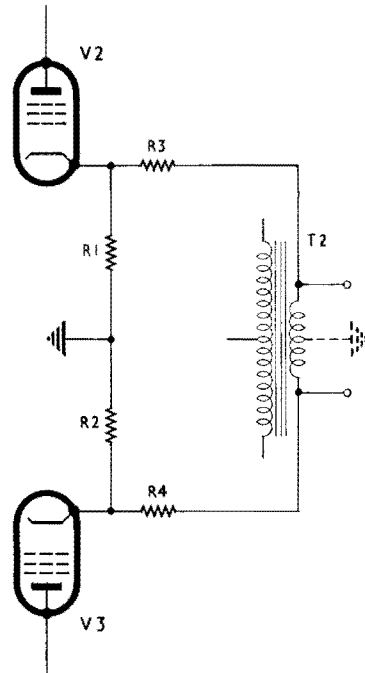
Power Supply.—The power supply circuit of fig. 6-4 is unconventional in that one set of rectifiers and one transformer provide two h.t. voltages, one of which is twice the value of the other. Compared with conventional circuits, this arrangement, which is described more fully in Section 1-9, would appear to offer a saving in both initial cost and space.

The rectifiers are not equally loaded, two of them having to supply the low voltage current of 100mA in addition to the 275mA at 1000V required by the DA42 stage at maximum output. However, this is well within the capability of the four U19 rectifiers specified since their rated maximum output in this type of circuit is 500mA at 4kV. The GXU50 xenon rectifier may be used as an alternative to the U19 ; a 10-second switching delay being then required.

As the power supply impedance causes some interaction between the low and high voltage supplies, a sudden demand for maximum current would cause a reaction in the 500V supply if a conventional 4-8μF smoothing capacitor is used in the high voltage supply. Connecting three 160μF, 450V electrolytic capacitors in series achieves an effective capacitance of 50-60μF, with a working voltage of 1350. Resistors R1, R2 and R3 equalise the voltage across each capacitor. The same principle is applied to the capacitors in the low voltage supply.
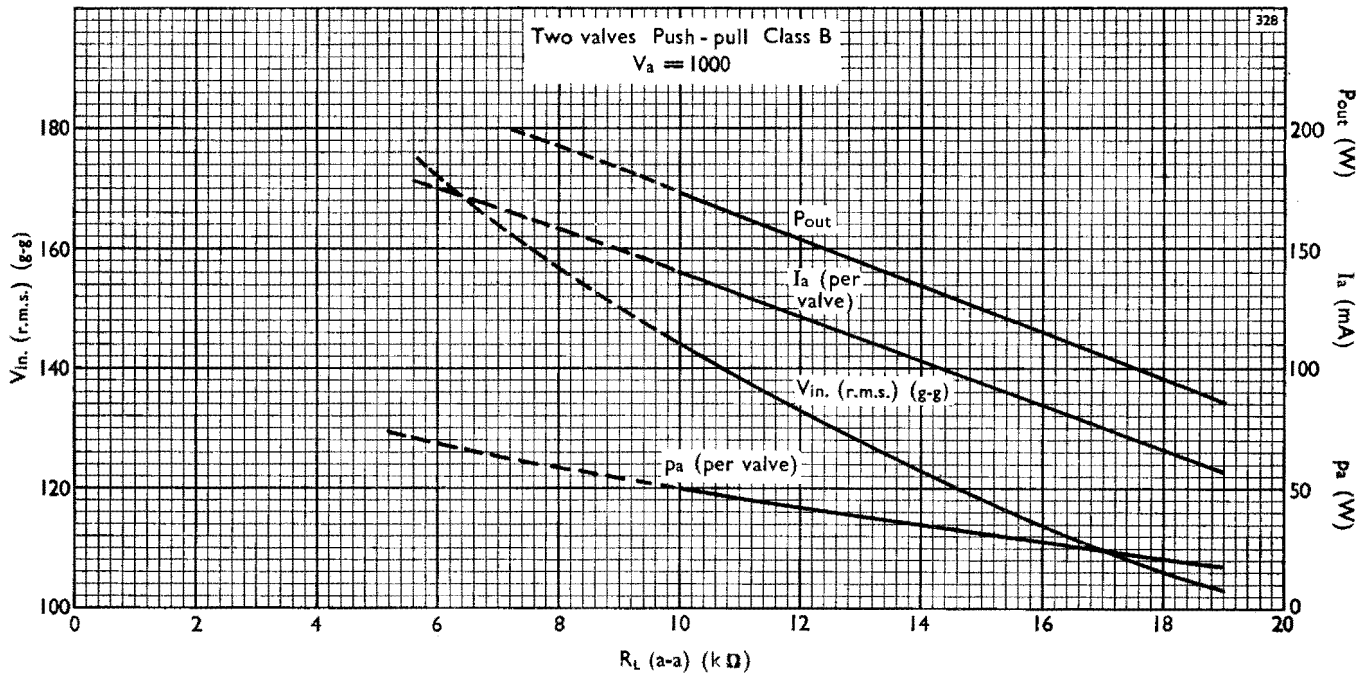
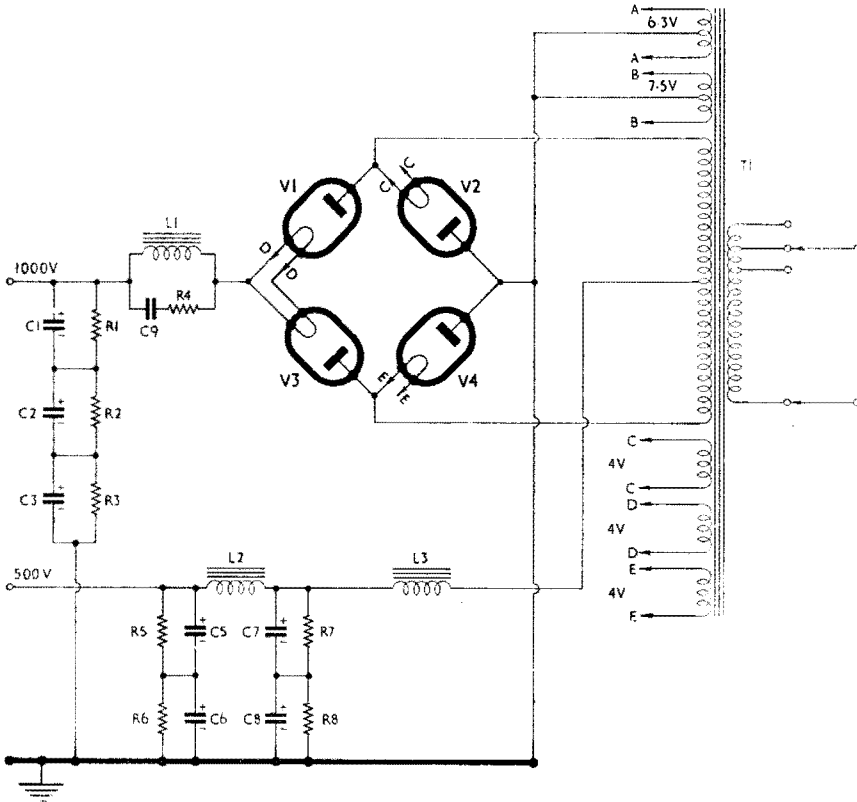**Fig. 6-3.**   Characteristics of the 175W amplifier of fig. 6-1.

Fig. 6-4.   Power supply circuit for the 175W and 200W amplifiers of figs. 6-1 and 6-5. Although three secondary windings can be used for the rectifier filaments, as shown, transformer winding is simplified if four identical windings are used.   Component values are given below.

G.E.C. VALVES

V1 ⎫
V2 ⎬ U19 or GXU50
V3 ⎪
V4 ⎭

MISCELLANEOUS

L1 10H   275mA
L2 10H   100mA
L3 10H   100mA
T1 Mains transformer
    Secondaries
        1200V CT 275mA
        6·3V   5A   CT
        7·5V   2·5A CT
        4V   3·5A
        4V   3·5A
        4V   7·0A (or two 3·5A)

RESISTORS

(10%, 1W unless otherwise shown)
R1 ⎫
R2 ⎬ 100kΩ
R3 ⎭
R4 10kΩ   2W
R5 ⎫
R6 ⎬ 100kΩ
R7 ⎪
R8 ⎭

CAPACITORS

C1 ⎫
C2 ⎪
C3 ⎪
C5 ⎬ 160µF   450V (each 100 + 60 dual)
C6 ⎪
C7 ⎪
C8 ⎭
C9 0·01µF   1·5kV

F

6-2. DA42 200W Amplifier.—The amplifier of fig. 6-1 may be modified to give 200W output by increasing the output stage anode potential to 1250V and by applying about 4V negative grid bias in order to limit the quiescent dissipation.

The modifications are shown in fig. 6-5.   The bias potential is obtained by inserting a resistor (R25) in the negative line of the h.t. supply for the driver stage.   The value of R25 will depend upon the total h.t. current, but it will be about 47Ω in the arrangement shown.   The bias is applied to the output stage grids by connecting it to the centre tap of the coupling transformer secondary.

The bias is set to 4V negative in the quiescent condition and at full output it will increase to about 5V as a result of the DA42 grid current.   This has very little effect upon the performance and is in any case partially suppressed by C15.

The performance of the output stage is shown by the curves of fig. 6-6 and Table 6-II gives the operating conditions.

COMPONENT VALUES FOR FIG. 6-5
DA42 200W AMPLIFIER

G.E.C. VALVES

V1 DH77/6AT6
V2 N709
V3 N709
V4 KT66
V5 KT66
V6 DA42
V7 DA42
V8 U54

RESISTORS

(20%, 0·25W *unless otherwise shown*)
R1 470kΩ
R2 1·5kΩ
R3 22kΩ   0·5W ⎫ matched to 5%
R4 22kΩ   0·5W ⎭
R5 470kΩ   10%
R6 470kΩ   10%
R7 10kΩ
R8 10kΩ
R9 47kΩ   0·5W
R10 220Ω   0·5W
R11 470kΩ   10%   0·5W
R12 47kΩ   5%   5W
R13 47kΩ   5%   5W
R14 4·7kΩ   10%   1W
R15 470kΩ
R16 470kΩ
R17 10kΩ
R18 10kΩ
R19 (See page 66)
R20 100Ω
R21 100Ω

R22 7·5kΩ   10W
R23 7·5kΩ   10W
R24 47kΩ   5W
R25 47Ω   1W

CAPACITORS

C1 0·01μF
C2 0·01μF
C3 0·01μF
C4 4μF   450V
C5 0·5μF
C6 16μF   450V
C7 0·05μF
C8 0·05μF
C9 16μF   100V
C10 16μF   100V
C11 2000pF   1kV
C12 2000pF   1kV
C13 8μF   500V
C14 8μF   600V
C15 100μF   12V
C16 150pF

MISCELLANEOUS

L1 10H   120mA
T1 Bridged transformer (see fig. 6-7)
T2 Output transformer (see fig. 6-8)
T3 Mains transformer
      Secondaries
          425-0-425V   120mA
          6·3V   5A   CT
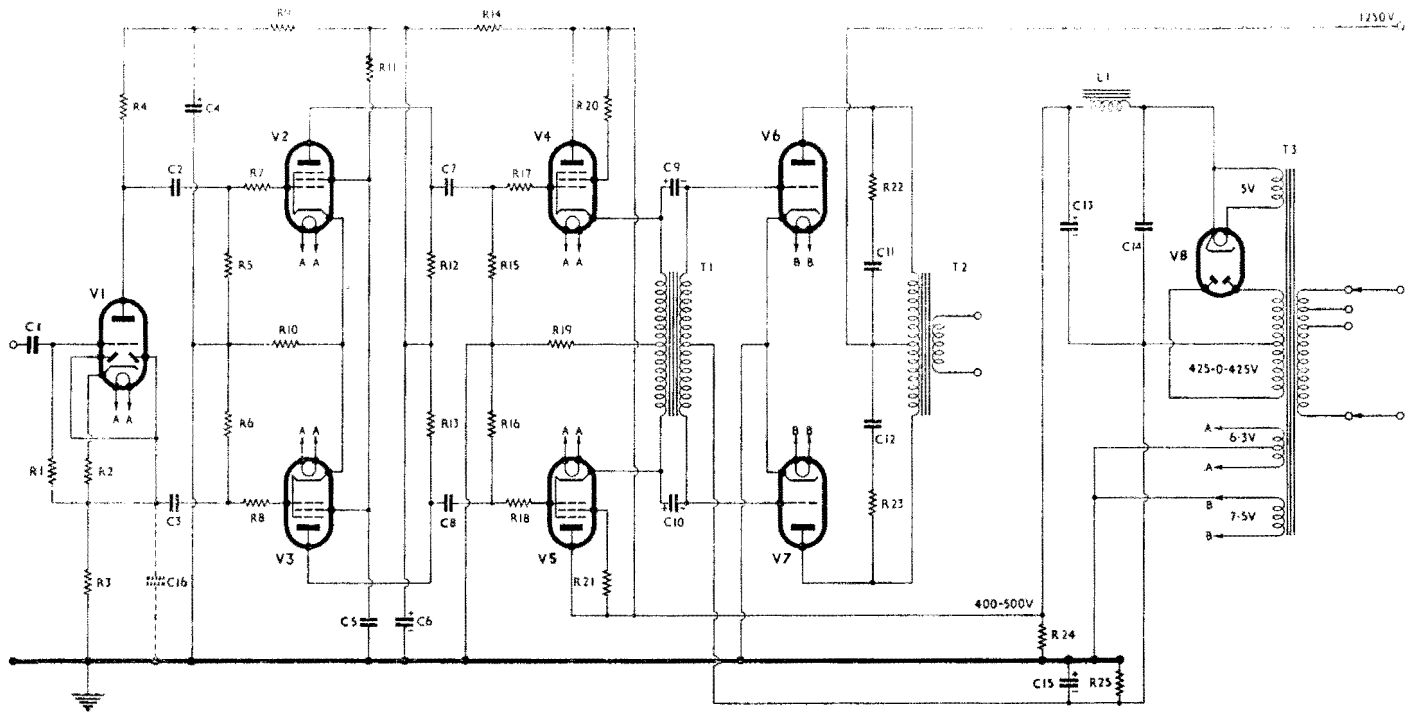   7·5V   2·5A   CT
          5V   3A

Fig. 6-5. Circuit of the DA42 200W Class B amplifier. Apart from the low voltage power supply and provision for grid bias in the output stage, this circuit is identical to that in fig. 6-1. See page 67 for the reason for C16, shown dotted.

TABLE 6-II

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 6-5

$V_{a\ (b)}$	1·25 (approx)	kV
V_a	1·25	kV
$I_{a\ (o)}$	2×20	mA
$I_{a\ (max\ sig)}$	2×120	mA
$p_{a\ (o)}$	$2 \times 12·5$	W
$p_{a\ (max\ sig)}$	2×25	W
$V_{g\ (o)}$	-4	V
$I_{g\ (max\ sig)}$	2×20	mA
P_{out}	200	W
$R_{L\ (a-a)}$	13	kΩ
z_{out}	15	kΩ
D	6	%
p_{dr}	4·5	W
$V_{in\ (g-g)\ (pk)}$	200	V
$V_{in\ (rms)}$ (to first stage)	$\not> 1$	V

These conditions should be used only for normal speech and music. A continuous tone at maximum output will exceed the permitted dissipation and for such applications $R_{L\ (a-a)}$ must be increased by at least 20%.
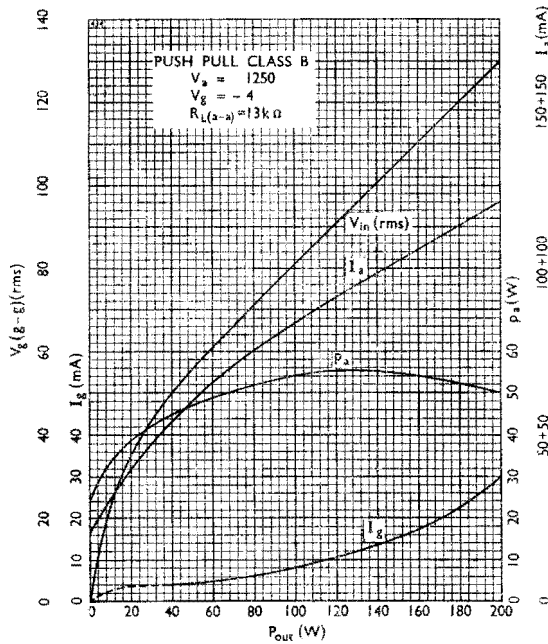

Fig. 6-6. Performance of the DA42 200W amplifier of fig. 6-5.

Power Supply.—The combined power supply in fig. 6-4 cannot be used because of the common negative connection.   Two separate supplies are required : one inductance-input supply giving 240mA at 1250V for the output stage and one, which may be capacitance-input, of 400-500V for the earlier stages.   A suggested arrangement for the latter is incorporated in fig. 6-5.
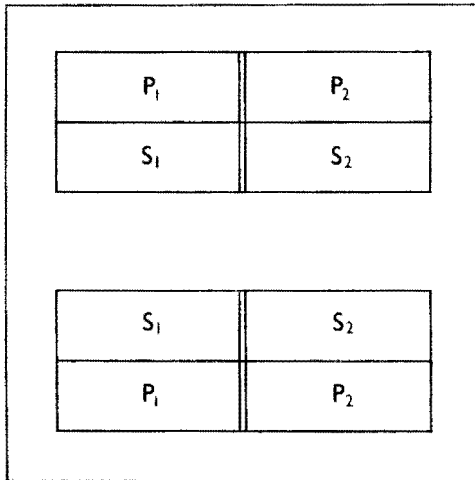

Fig. 6-7.   The coupling transformer T1 in figs. 6-1 and 6-5.   Winding details are as follows :

> Core :   Stalloy No. 4 ; square section ; no gap.
> Primary :   2000+2000 turns, 28 s.w.g.
> Secondary : 2000+2000 turns, 30 s.w.g.

The primary and secondary half-windings P1 and S1 are wound in the opposite direction to P2 and S2 so that, when the inner ends of each half-winding are joined, they are series-aiding.   The number of turns in each of the four sections must be identical.
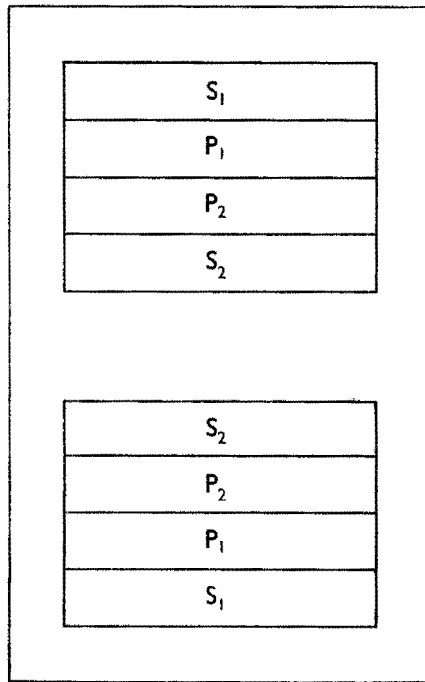
Fig. 6-8.   The 100V line (60Ω) output transformer T2 in figs. 6-1 and 6-5.   Winding details are as follows :

Core :	Stalloy No. 66 ; square section ; no gap.
Primary :	1250+1250 turns, 28 s.w.g.
Secondary :	100+100 turns, 15 s.w.g.

All windings are wound in the same direction.   The secondary is connected in series-aiding.

CHAPTER 7

CLASS AB AMPLIFIERS OF 100 TO 300 WATTS

The DA100 triode is designed for use in Class AB circuits and this chapter gives design information for Class AB1 and AB2 amplifiers which provide output powers of 175W and 270W respectively. The data includes the conditions for giving reduced outputs of 115W and 200W respectively thereby providing a choice of four amplifiers covering the range 115W to 270W.

7-1. DA100 115-175W Class AB1 Amplifier.—The circuit diagram of fig. 7-1 illustrates the output and driver stages of this amplifier and Table 7-I gives the operating data. The performance is shown by the curves of figs. 7-2, 7-3 and 7-4.

The DA100 grids are not driven positive and the preceding stage is of orthodox design and consists of a pair of triode-connected N709 or KT61 valves in a transformer-coupled push-pull Class A amplifier circuit. The output stage requires an input of 150+150V r.m.s. and this is obtained from a coupling transformer having a 1:2 ratio.

A diode is shunted across each DA100 grid circuit in order to prevent " triggering." If these diodes are absent, the output valves may be seriously damaged by excessive anode current in the event of " trigger effect " should their grids be driven positive.

Meters in the DA100 anode lines enable the grid bias to be adjusted to give the correct anode currents. These meters or other anode current measuring facilities as well as the anode circuit fuse (fig. 7-5) should always be incorporated.

A suitable power supply for the output stage is given in fig. 7-5 in which two GXU50 xenon rectifiers or U19 vacuum rectifiers are used in a bi-phase half-wave circuit having a source impedance of 400Ω. The grid bias is provided by the separate circuit also shown in fig. 7-5. The power supply for the driver and earlier stages may be of conventional design.

Table 7-II gives the data necessary for operating this amplifier at a reduced anode potential to give 115W output.
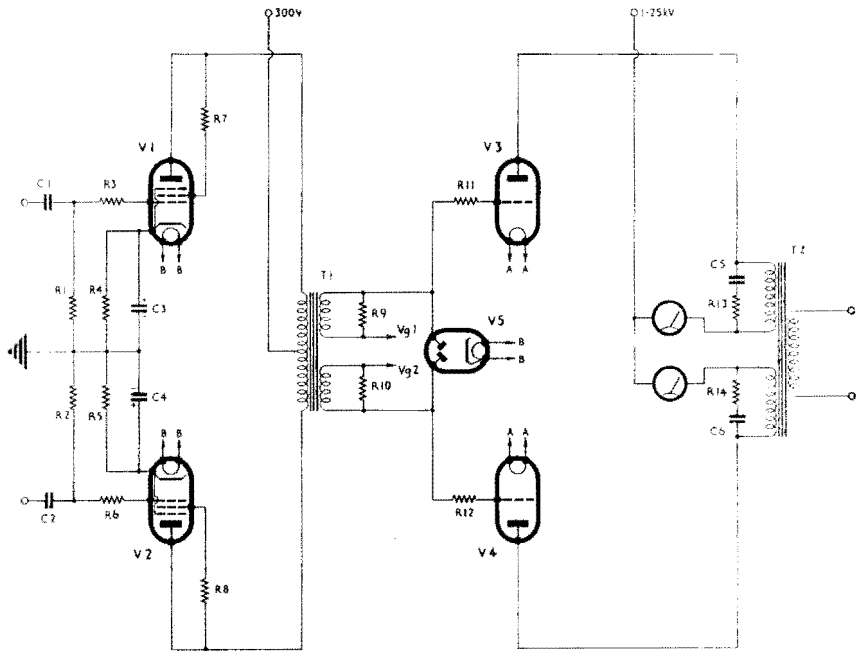
Fig. 7-1. Circuit of the driver and output stages of the DA100 Class AB1 115-175W amplifier. Component values are given below :

G.E.C. VALVES

V1 N709
V2 N709
V3 DA100
V4 DA100
V5 U709 or U78/6X4

RESISTORS

(20%, 0·25W unless otherwise shown)
R1 270kΩ
R2 270kΩ
R3 10kΩ
R4 330Ω   10%   0·5W
R5 330Ω   10%   0·5W
R6 10kΩ
R7 100Ω
R8 100Ω
R9 5kΩ
R10 5kΩ

R11 1kΩ
R12 1kΩ
R13 10kΩ   10%   20W
R14 10kΩ   10%   20W

CAPACITORS

C1 0·02μF
C2 0·02μF
C3 50μF   50V
C4 50μF   50V
C5 2000pF   1·5kV
C6 2000pF   1·5kV

TRANSFORMERS

T1 Intervalve transformer, 1 : 2
T2 Output transformer
       7kΩ anode-anode (175W)
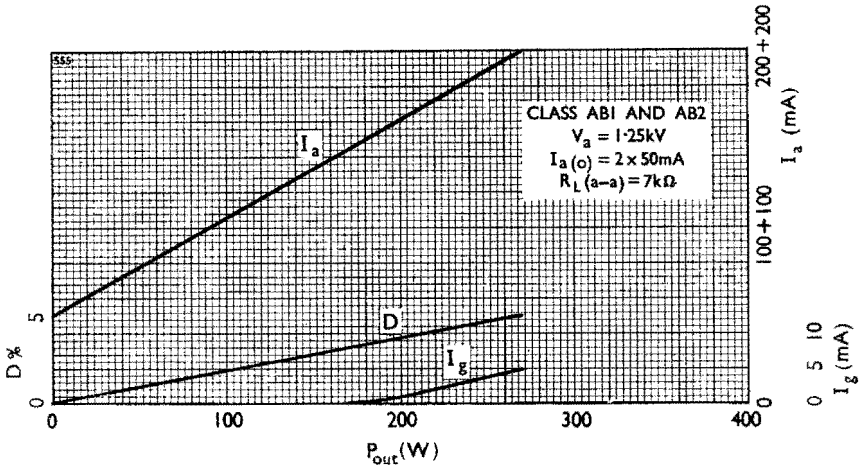   6kΩ anode-anode (115W)

Fig. 7-2. Characteristics of the amplifiers of fig. 7-1 (Class AB1 175W) and fig. 7-6 (Class AB2 270W).
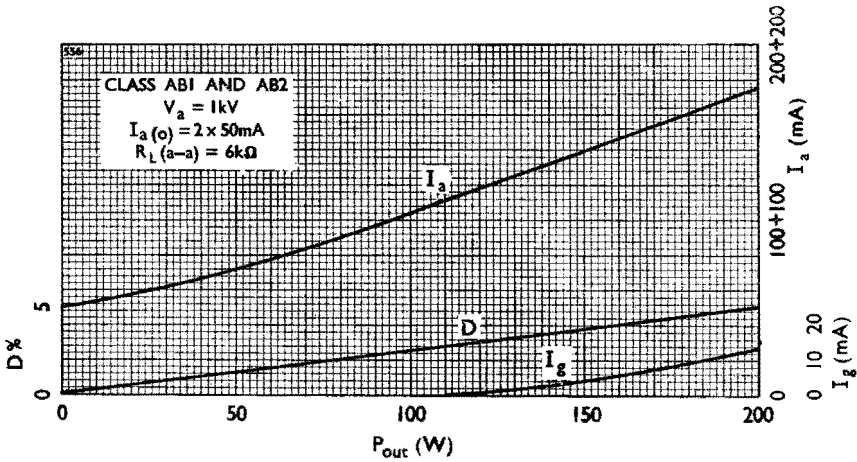

Fig. 7-3. Characteristics of both AB1 and AB2 amplifiers at the reduced ratings of 115W and 200W respectively.
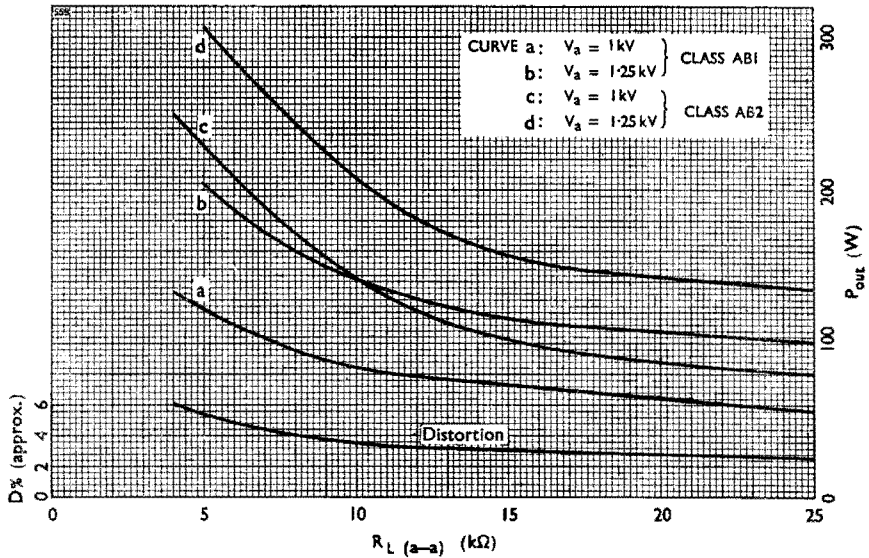
Fig. 7-4. Distortion and power output against anode load for the various AB1 and AB2 conditions given in Tables 7-I to 7-IV.

COMPONENT VALUES FOR FIG. 7-5

POWER SUPPLY FOR DA100 AB1 AND AB2 AMPLIFIERS

G.E.C. VALVES

V1 U19 or GXU50
V2 U19 or GXU50
V3 U52/5U4G

RESISTORS

R1 10kΩ 20% 0·5W
R2 2kΩ 5W w.w.
R3 2kΩ 5W w.w.
R4 2kΩ 10% 20W w.w.

CAPACITORS

C1 4μF 1·5kV
C2 4μF 1·5kV
C3 0·01μF 1·5kV
C4 16μF 350V

MISCELLANEOUS

*L1 5H 300mA
*L2 5-25H 300mA swinging choke
L3 20H 150mA
T1 4V 7A
T2 As required
T3 Bias transformer
   Secondaries
    250-0-250V 100mA (1kV opera-
     tion)
    325-0-325V 100mA (1·25kV
     operation)
    6V 4A CT
    6·3V CT as required
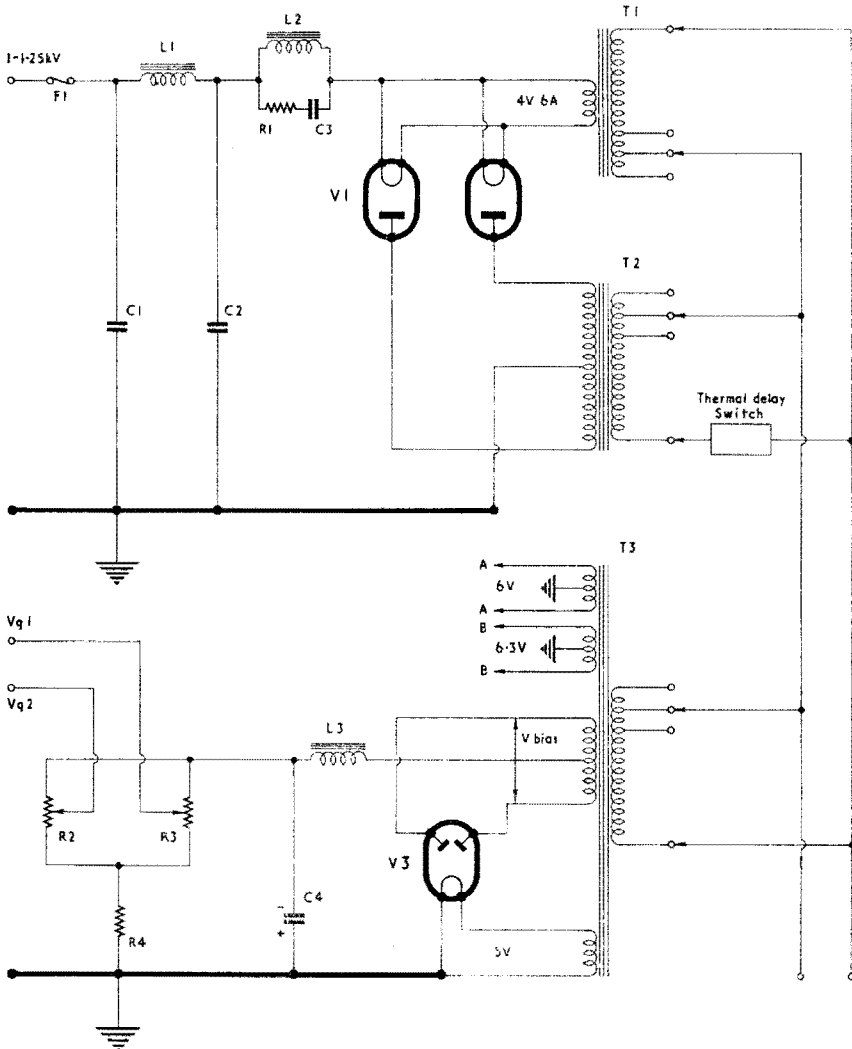   5V 3A
F1 500mA

*≯50Ω for AB2

Fig. 7-5. Power supply circuit for the AB1 and AB2 amplifiers of figs. **7-1** and 7-6. The thermal delay switch at T2 is not necessary with U19 vacuum rectifiers.

TABLE 7-I

OPERATING CONDITIONS OF THE CLASS AB1 175W AMPLIFIER OF FIG. 7-1

$V_{a (b)}$	1·25 (approx)	kV
V_a	1·25	kV
*$I_{a (o)}$	2×50	mA
$I_{a (max sig)}$	2×150	mA
$P_{a (o)}$	2×70	W
$P_{a (max sig)}$	2×100	W
$V_{g (o)}$	-225	V
P_{out}	175	W
$R_{L (a-a)}$	7	kΩ
D	3	%
$V_{in (pk) (g-g)}$	2×225	V
$V_{in (rms)}$ (to driver stage)	2×10	V

*Precise value unimportant and may be between 40 and 60mA.

TABLE 7-II

OPERATING CONDITIONS OF THE CLASS AB1 AMPLIFIER OF FIG. 7-1
FOR REDUCED OUTPUT OF 115W

$V_{a (b)}$	1 (approx)	kV
V_a	1	kV
*$I_{a (o)}$	2×50	mA
$I_{a (max sig)}$	2×135	mA
$P_{a (o)}$	2×55	W
$P_{a (max sig)}$	2×90	W
$V_{g (o)}$	-175	V
P_{out}	115	W
$R_{L (a-a)}$	6	kΩ
D	3	%
$V_{in (pk) (g-g)}$	2×165	V
$V_{in (rms)}$ (to driver stage)	2×8	V

*Precise value unimportant and may be between 40 and 60mA.

7-2. DA100 200-270W Class AB2 Amplifier.—This amplifier, shown in fig. 7-6, requires a driver stage capable of providing the 3W driving power for the DA100 grid circuit and two KT66 in a cathode-coupled bridged transformer arrangement are recommended.   This type of driver is discussed in Chapter 1, page 7.

The amplifier preceding the KT66 stage is a push-pull resistance-capacitance coupled arrangement using two N709 valves as described in Chapter 1, page 14. This circuit will provide the 200+200V required at the cathodes of the KT66 driver stage.

Meters are placed in the DA100 anode lines to enable the grid bias to be adjusted for correct anode current. It is inadvisable to omit these meters or the fuse in the anode circuit (fig. 7-5). There is no danger of " trigger effect " with the Class AB2 circuit because of the low d.c. grid circuit resistance, consequently, diodes are not required in this output stage.

The power supply of fig. 7-5 is also suitable for this amplifier and the grid bias network should provide a range of 175-250V negative to cater for differences in valve characteristics. The d.c. resistance of the inductors in the high voltage smoothing filter should not exceed 50Ω. If quality of reproduction is not important and some hum is tolerated the second inductor (L1) and capacitor (C1) may be omitted. In this event, the signal will be modulated at 100c/s ripple frequency at outputs exceeding 75% of the maximum.

The power supply for the driver and earlier stages may be of conventional design.

Table 7-III gives the output stage operating data for this amplifier and Table 7-IV the output stage characteristics for a reduced output of 200W at 1kV h.t.

TABLE 7-III

OPERATING CONDITIONS OF THE CLASS AB2 270W AMPLIFIER OF FIG. 7-6

$V_{a\ (b)}$	1·25 (approx)	kV
V_a	1·25	kV
$I_{a\ (o)}$	2×50	mA
$I_{a\ (max\ sig)}$	2×200	mA
$p_{a\ (o)}$	2×70	W
$p_{a\ (max\ sig)}$	2×90	W
$V_{g\ (o)}$	-225	V
$I_{g\ (max\ sig)}$	5	mA
$i_{g\ (pk)\ (max\ sig)}$	20	mA
P_{out}	270	W
$R_{L\ (a-a)}$	7	kΩ
z_{in}	25	kΩ
D	5	%
p_{dr}	3	W
$V_{in\ (pk)\ (g-g)}$	2×300 (approx)	V
$V_{in\ (rms)}$ (to driver stage)	2×15	V

TABLE 7-IV

Operating Conditions of the Class AB2 Amplifier of Fig. 7-6
for Reduced Output of 200W

$V_{a\ (b)}$	1 (approx)	kV
V_a	1	kV
$I_{a\ (o)}$	2×50	mA
$I_a\ (max\ sig)$	2×150	mA
$p_{a\ (o)}$	2×55	W
$p_a\ (max\ sig)$	2×85	W
$V_{g\ (o)}$	-175	V
$I_{g\ (max\ sig)}$	15	mA
$i_{g\ (pk)\ (max\ sig)}$	60	mA
P_{out}	200	W
$R_{L\ (a-a)}$	5	kΩ
z_{in}	7	kΩ
D	5	%
p_{d_1}	7	W
$v_{in\ (pk)\ (g-g)}$	2×250	V
$V_{in\ (rms)}$ (to driver stage)	2×10	V

COMPONENT VALUES FOR FIG. 7-6

DA100 200-270W AB2 AMPLIFIER

G.E.C. VALVES
V1 N709
V2 N709
V3 KT66
V4 KT66
V5 DA100
V6 DA100

RESISTORS
(20%, 0·25W *unless otherwise shown*)
R1 220kΩ
R2 220kΩ
R3 220Ω   10%   1W
R4 220kΩ   10%   1W
R5 47kΩ   3W ⎫
R6 47kΩ   3W ⎬ matched to 5%
R7 4·7kΩ   10%   1W
R8 470kΩ
R9 470kΩ
R10 10kΩ
R11 10kΩ
R12 100Ω

R13 100Ω
R14 (See Sect. 1-3)
R15 5kΩ   10%   50W
R16 5kΩ   10%   50W

CAPACITORS
C1 0·01µF
C2 0·01µF
C3 0·5µF
C4 0·05µF
C5 0·05µF
C6 4µF   350V
C7 4µF   350V
C8 2000pF   1·5kV
C9 2000pF   1·5kV

MISCELLANEOUS
L1 Centre-tapped inductor 100-200H
T1 Bridged transformer (See fig. 6-7)
T2 Output transformer
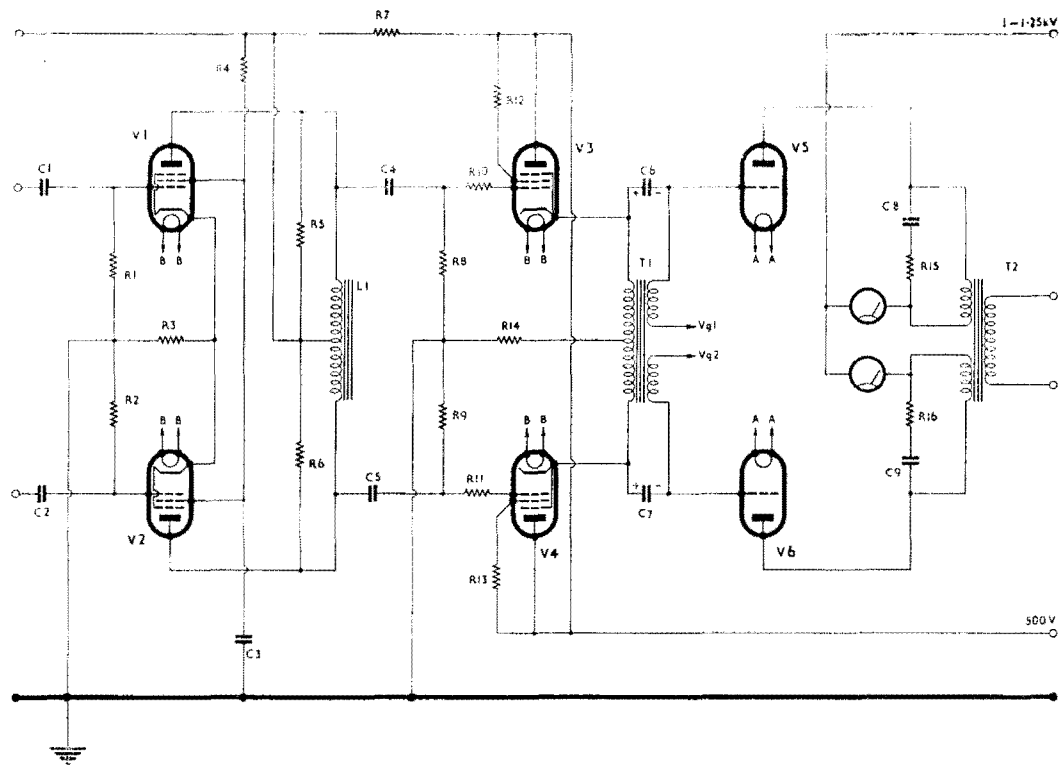   7kΩ anode-anode (270W)
    5kΩ anode-anode (200W)

Fig. 7-6. Circuit of the DA100 Class AB2 200-270W amplifier. The cathode-coupled driver stage is preceded by two N709 beam pentodes in a push-pull voltage amplifier stage as described on page 14.

CHAPTER 8

CLASS AB AMPLIFIERS OF 300 TO 1100 WATTS

The largest amplifiers dealt with in this book are designed round the V1505, a Class AB triode of 275W dissipation, and this chapter describes the design of both Class AB1 and AB2 circuits. The AB1 amplifier will provide up to 450W output and the AB2 amplifier up to 1100W. An output exceeding 900W requires considerable driving power and the data should be closely followed in order to obtain minimum distortion.

8-1. V1505 300-450W Class AB1 Amplifier.—Fig. 8-1 is the circuit diagram of the output and driver stages of this amplifier. Little explanation is required ; the driver stage contains a pair of triode-connected KT66 valves which are coupled to the output stage by a transformer of 1:2 ratio. The spark gap across the output transformer primary is optional but, if it is used, the series resistor (R13) must not be omitted. A load impedance of 10kΩ is specified and this should not be reduced as a lower load will result in excessive anode dissipation at maximum output.

Fig. 8-5 contains the power supply circuit recommended for providing h.t. for the output stage. The bias voltage is provided by a normal full-wave rectifier circuit which is also shown in fig. 8-5. The KT66 driver stage power supply may be a conventional capacitance-input circuit since there is no variation in h.t. current. The KT66 valves require about 350V at 100mA.

The performance of the amplifier is illustrated in figs. 8-3 and 8-4 and the output stage data are given in Table 8-I.

Table 8-II gives the data necessary for operating this amplifier at a reduced anode potential to give 300W output and fig. 8-2 illustrates the corresponding performance. At this reduced level, a load impedance lower than the 10kΩ specified is permissible and this would provide a somewhat higher output as is shown by the appropriate curve in fig. 8-2.
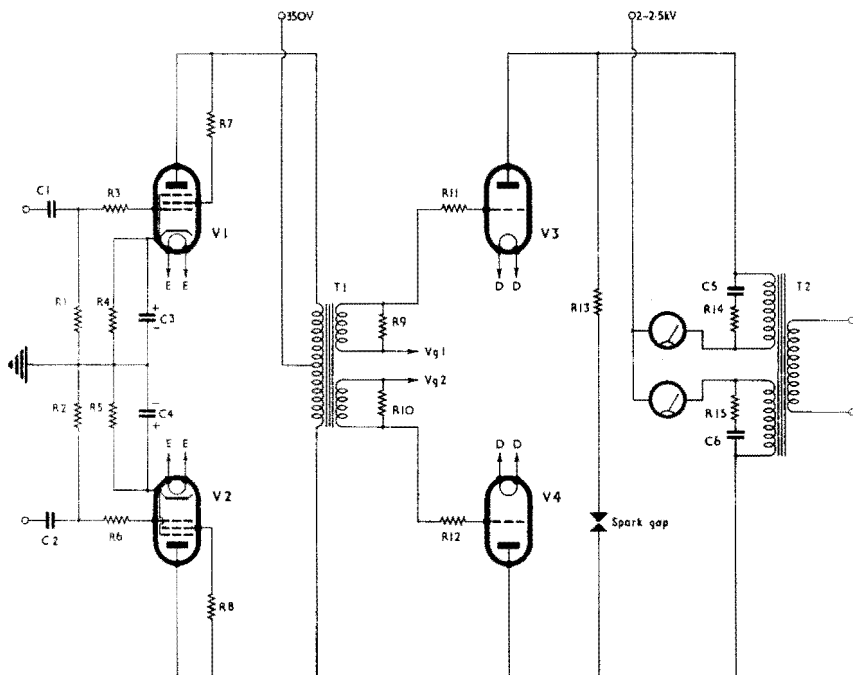
Fig. 8-1. Circuit of the output and driver stages of the V1505 Class AB1 300-450W amplifier. Component values are given below.

G.E.C. VALVES

V1 KT66
V2 KT66
V3 V1505
V4 V1505

RESISTORS

(20%, 0·25W unless otherwise shown)
R1 470kΩ
R2 470kΩ
R3 10kΩ
R4 600Ω 5% 3W
R5 600Ω 5% 3W
R6 10kΩ
R7 100Ω 0·5W
R8 100Ω 0·5W
R9 33kΩ 10% 0·5W
R10 33kΩ 10% 0·5W

R11 1kΩ
R12 1kΩ
R13 10kΩ 10% 100W (see p. 86)
R14 10kΩ 10% 50W
R15 10kΩ 10% 50W

CAPACITORS

C1 0·02μF
C2 0·02μF
C3 25μF 50V
C4 25μF 50V
C5 2000pF 1·5kV
C6 2000pF 1·5kV

TRANSFORMERS

T1 Intervalve transformer 1:2
T2 Output transformer
    10kΩ anode-anode

Solid lines: Driver supply 475 V
Broken lines: Driver supply 575 V
        see text for explanation

PUSH PULL OPERATION
        $V_a = 2.5$ kV
    $I_{a(o)} = 50$ mA per valve
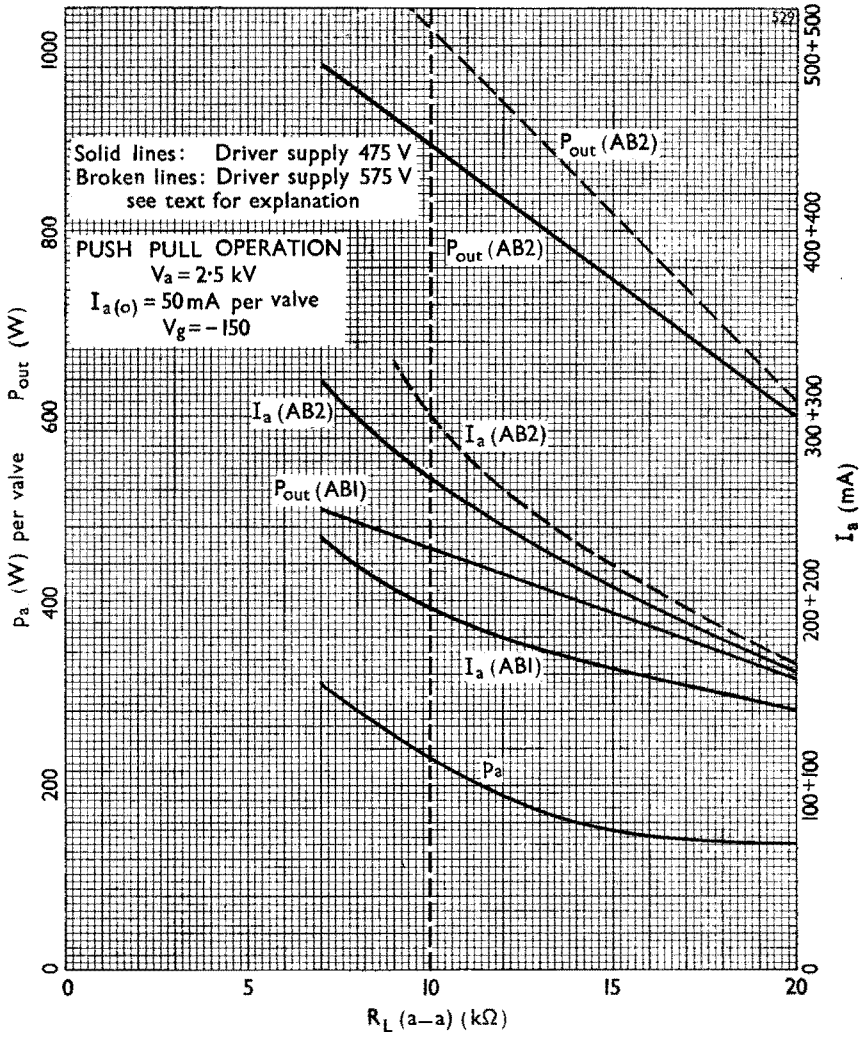 $V_g = -150$

Fig. 8-2. Characteristics of the amplifiers of fig. 8-1 (Class AB1 450W) and fig. 8-6 (Class AB2 900-1100W).
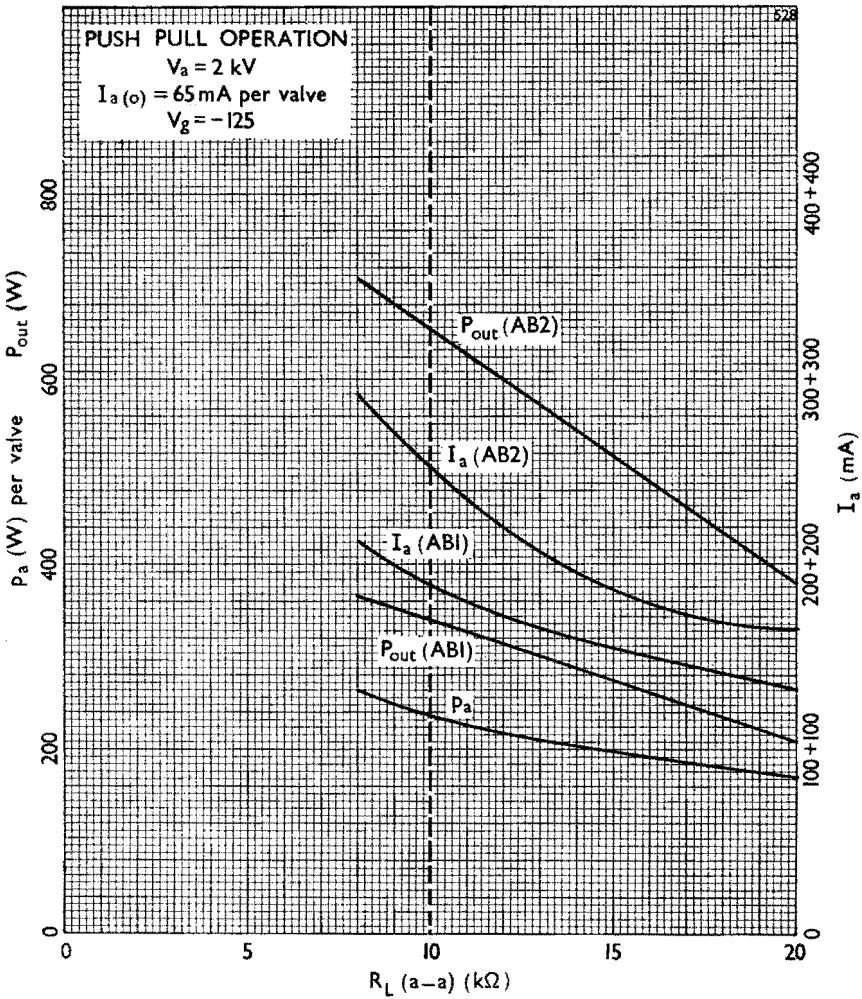
Fig. 8-3.   Characteristics of both AB1 and AB2 amplifiers at the reduced ratings of 300W and 450W respectively.
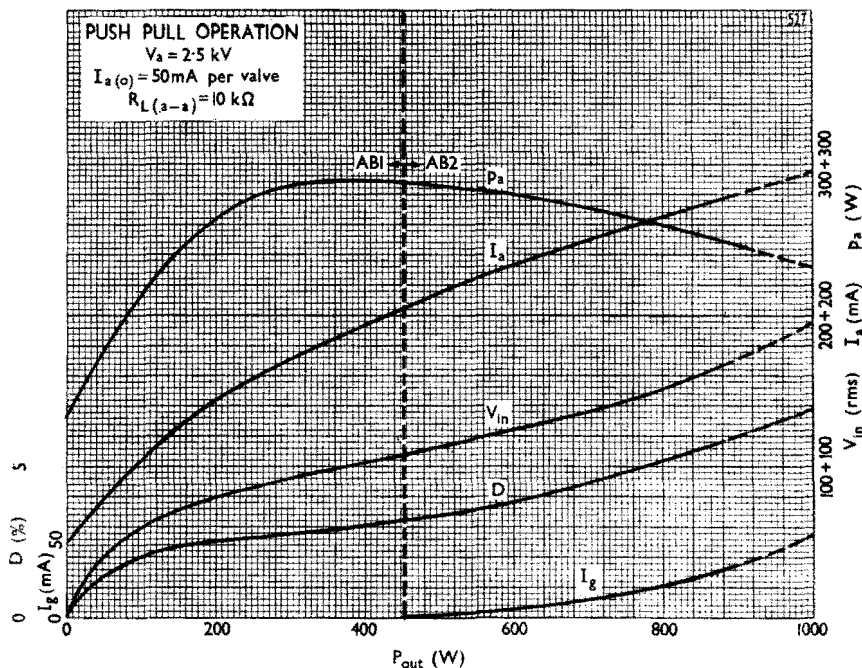
Fig. 8-4. Valve operating conditions and distortion against output power for the V1505 in both AB1 and AB2.

COMPONENT VALUES FOR FIGS. 8-5 AND 8-7

POWER SUPPLY FOR V1505 AB1 AND AB2 AMPLIFIERS

G.E.C. VALVES

V1 ⎱
V2 ⎰ U19 or GXU1
V3 ⎱
V4 ⎰
V5 U52/5U4

RESISTORS

R1 10kΩ 20% 0·5W
*R2 2kΩ 5W w.w.
*R3 2kΩ 5W w.w.
*R4 2kΩ 10% 10W
R5 700Ω 10% 25W (fig. 8-7)
*In Class AB1, may be increased to 5kΩ to reduce current consumption.

CAPACITORS

C1 4μF 3kV
C2 4μF 3kV

C3 0·01μF 3kV
C4 16μF 350V
C5 4μF 350V (fig. 8-7)

MISCELLANEOUS

F1 HV type
L1 5H 600mA
L2 5-20H 600mA swinging choke
L3 20H 100mA
T1 As required
T2 2·5V 5A
    2·5V 5A
    2·5V 10A (or two 5A)
T3 Bias transformer
    Secondaries :
      220-0-220V 85mA
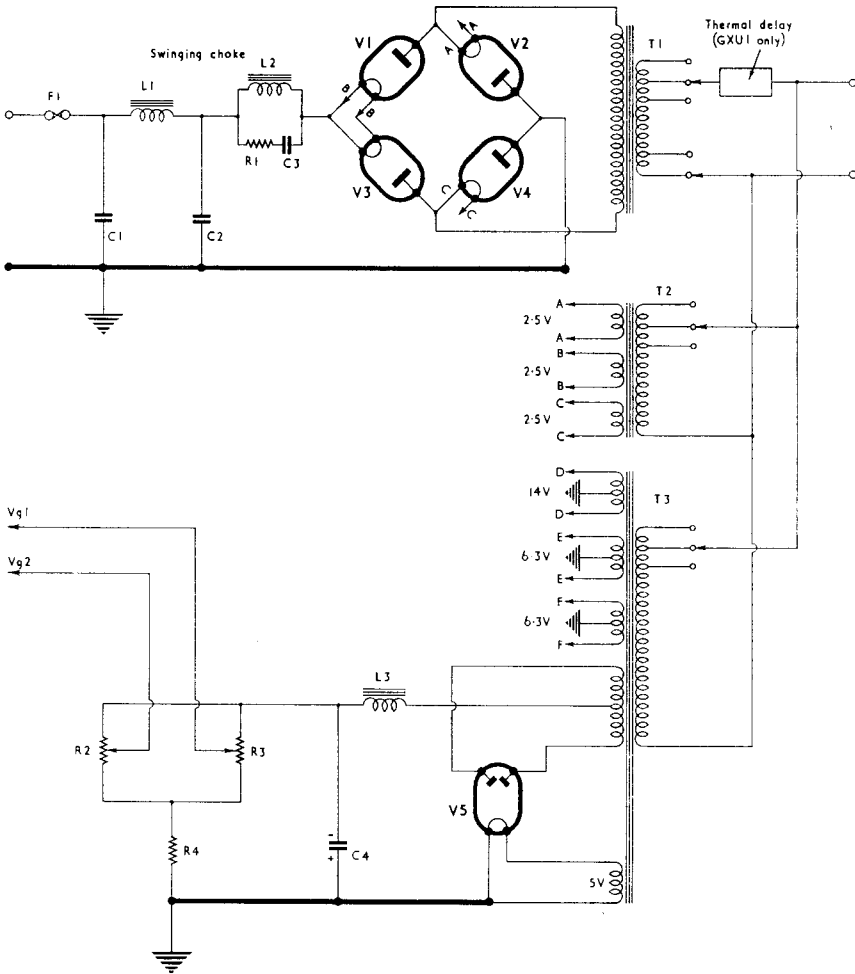   14V 13A
      6·3V as required
      5V 3A

Fig. 8-5. Power supply circuit for the V1505 Class AB1 and AB2 amplifiers in figs. 8-1 and 8-6.

TABLE 8-I

OPERATING CONDITIONS OF THE CLASS AB1 450W AMPLIFIER OF FIG. 8-1

$V_{a (b)}$	2·5 (approx)	kV
V_a	2·5	kV
*$I_{a (o)}$	2×50	mA
$I_{a (max sig)}$	2×200	mA
$p_{a (o)}$	2×130	W
$p_{a (max sig)}$	2×275	W
$V_{g (o)}$	-155	V
P_{out}	450	W
$R_{L (a-a)}$	10	kΩ
D	3	%
$v_{in (pk) (g-g)}$	2×150	V
$V_{in (rms)}$ (to driver stage)	2×10	V

*See section 8-3, page 96.

TABLE 8-II

OPERATING CONDITIONS OF THE CLASS AB1 AMPLIFIER OF FIG. 8-1
FOR REDUCED OUTPUT OF 300W

$V_{a (b)}$	2 (approx)	kV
V_a	2	kV
*$I_{a (o)}$	2×65	mA
$I_{a (max sig)}$	2×180	mA
$p_{a (o)}$	2×135	W
$p_{a (max sig)}$	2×200	W
$V_{g (o)}$	-125	V
P_{out}	300	W
$R_{L (a-a)}$	10	kΩ
D	3	%
$v_{in (pk) (g-g)}$	2×120	V
$V_{in (rms)}$ (to driver stage)	2×8	V

*See section 8-3, page 96.

8-2. V1505 600-1100W Class AB2 Amplifier.—The recommended circuit for Class AB2 operation of the V1505 is shown in fig. 8-6 which also contains the KT66 driver stage with its preceding amplifier.   Bridged transformer coupling is used, as described in Chapter 1, page 5, between the KT66 and V1505 stages.   Up to 900W will be obtained with a driver anode supply of 475V at a total current of 130mA.   The bias required for the KT66 valves for this operating point is 40V and this is provided by an effective cathode resistance of 600Ω per valve.   Part of this resistance is provided in each cathode line by the half-primaries of the coupling transformer, the remainder (R13) being inserted in the common earth line to the transformer centre-tap.

To obtain 1100W output, the V1505 valves will require a driving power approximately double that needed for 900W and a considerable signal voltage is necessary. The KT66 valves (the KT88 is also suitable) must operate at an anode supply of 575V and they will each consume 55mA. The required bias of about 55V is produced by an effective cathode resistance of 1000Ω per valve made up as above.

A load impedance of 10kΩ is specified and this should not be reduced as a lower load will result in excessive anode dissipation at the half-power point—where maximum dissipation occurs in Class AB2.

The power supply circuit is the same as that recommended for the Class AB1 arrangement and is shown in fig. 8-5.

The curves in figs. 8-3 and 8-4 show the performance in detail and Table 8-III gives the output stage operating data for either 900W or 1100W output.

For an output of 600W the V1505 anode supply is reduced to 2kV and the resulting performance is illustrated in fig. 8-2. The output stage data are given in Table 8-IV. At this reduced level, a load impedance lower than the 10kΩ specified is permissible and this would provide a somewhat higher output.

TABLE 8-III

OPERATING CONDITIONS OF THE CLASS AB2 900-1100W AMPLIFIER OF FIG. 8-6

$V_{a\ (b)}$	2·5 (approx)	2·5 (approx)	kV
V_a	2·5	2·5	kV
*$I_{a\ (o)}$	2×50	2×50	mA
$I_{a\ (max\ sig)}$	2×275	2×300	mA
$P_{a\ (o)}$	2×130	2×130	W
†$P_{a\ (max)}$	2×275	2×275	W
$V_{g\ (o)}$	—155	—155	V
$I_{g\ (max\ sig)}$	30	60	mA
$i_{g\ (pk)\ (max\ sig)}$	75	140	mA
P_{out}	900	1100	W
$R_{L\ (a—a)}$	10	10	kΩ
z_{in}	8-12	8-12	kΩ
D	6	7	%
P_{dr}	10‡	20§	W
$V_{in\ (pk)\ (g—g)}$	2×230	2×280	V
$V_{in\ (rms)}$ (to first stage)	2×5	2×6	V

*See section 8-3, page 96.

†$P_{a\ (max)}$ occurs at half output. (See fig. 8-4).

‡Driver stage operating at 475V, 130mA.

§Driver stage operating at 575V, 110mA.

TABLE 8-IV

OPERATING CONDITIONS OF THE CLASS AB2 AMPLIFIER OF FIG. 8-6
FOR REDUCED OUTPUT OF 600W

$V_{a (b)}$	2 (approx)	kV
V_a	2	kV
*$I_{a (o)}$	2×60	mA
$I_{a (max sig)}$	2×230	mA
$p_{a (o)}$	2×120	W
$p_{a (max sig)}$	2×170	W
$V_{g (o)}$	-125	V
$I_{g (max sig)}$	40	mA
$i_{g (pk) (max sig)}$	95	mA
P_{out}	600	W
$R_{L (a-a)}$	10	$k\Omega$
z_{in}	8-12	$k\Omega$
D	6	%
p_{dr}	10†	W
$V_{in (pk) (g-g)}$	2×200	V
$V_{in (rms)}$ (to first stage)	2×4	V

*See section 8-3, page 96.

†Driver stage operating at 475V, 130mA.

COMPONENT VALUES FOR FIG. 8-6

V1505 600-1100W AB2 AMPLIFIER

G.E.C. VALVES
V1 N709
V2 N709
V3 KT66
V4 KT66
V5 V1505
V6 V1505

RESISTORS
(20%, 0·25W *unless otherwise shown*)
R1 470kΩ
R2 470kΩ
R3 220Ω
R4 100Ω
R5 100Ω
R6 1·5-2·5kΩ   5W
R7 100kΩ   10%
R8 100kΩ   10%
R9 10kΩ
R10 10kΩ

R11 100Ω
R12 100Ω
R13 See Sect. 8-2, page 92
R14 10kΩ   10%   50W
R15 10kΩ   10%   50W

CAPACITORS
C1 0·02μF
C2 0·02μF
C3 4μF   600V
C4 4-16μF   350V
C5 4-16μF   350V
C6 2000pF   1·5kV
C7 2000pF   1·5kV

TRANSFORMERS
T1 Intervalve transformer 1:2
T2 Bridged transformer (see fig. 6-7)
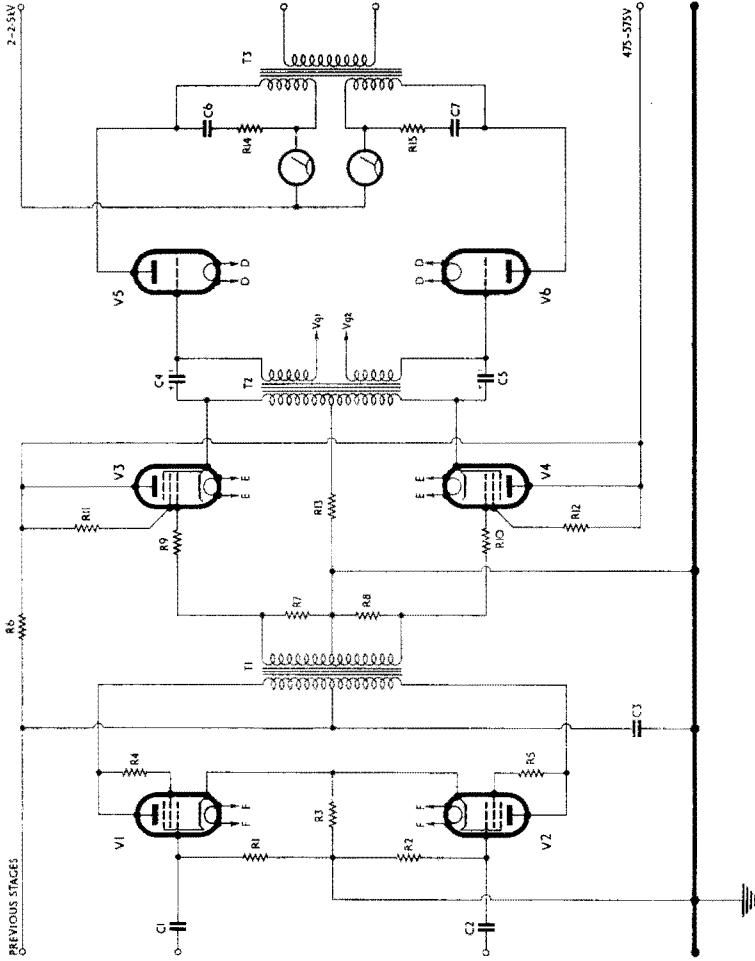T3 Output transformer
        10kΩ anode-anode

Fig. 8-6. Circuit of the V1505 Class AB2 600-1100W amplifier.

8-3. Quiescent Current in AB1 and AB2.—The quiescent currents selected (65mA and 50mA for anode potentials of 2kV and 2·5kV respectively) are not critical and any current of this order may be used. In both Class AB1 and AB2, the quiescent dissipation is low, at 150W, which is a desirable feature. However, too low a current will result in increased distortion whereas currents in excess of those recommended will not reduce distortion. These two values of current are chosen to give similar orders of the crossover distortion which occurs at full output as a result of the regulation of the grid bias supply circuit. Although a stabilised bias supply would remove the effect at the cost of some circuit elaboration it should be noted that the effect is virtually absent up to 90% of full output.

8-4. Protection against Bias Failure.—In the simple arrangement shown in fig. 8-7 a relay, with its operating coil connected to the bias supply, inserts a resistor in the V1505 filament circuit when its normally-open contacts are released by bias failure. The value of the resistor is chosen so that the V1505 valves operate with auto-bias at 100mA anode current per valve and give a reduced output of 200-300W.
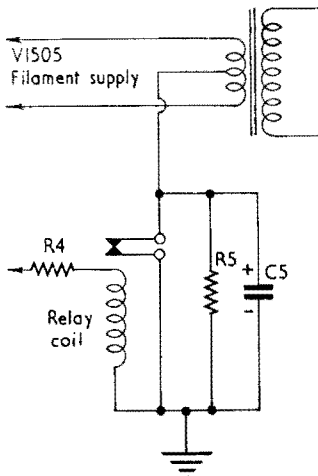

Fig. 8-7. Bias failure protection for the V1505 amplifiers. The relay coil is connected in series with R4 (fig. 8-5) and the normally-open contacts are held closed by the bias supply current. Bias failure opens the contacts and the V1505 filaments are then auto-biased by R5. See page 90 for component values.

CHAPTER 9

PRE-AMPLIFIERS AND TONE CONTROL

This chapter discusses some of the requirements of pre-amplifiers for use with the power amplifiers described in earlier chapters and includes details of two complete pre-amplifier circuits containing record and tape equalisation and tone controls.

9-1. The Purpose of a Pre-Amplifier.—This can be summed up, within the context of this book, in three requirements : (*a*) the amplification of small signals, of the order of millivolts, to a level suitable for operating a power amplifier without adding hum and other noise ; (*b*) frequency " equalisation " of the output from gramophone records, tape, etc. ; (*c*) the provision of variable tone controls, which, to satisfy present-day demands requires at least bass and treble cut and bass lift.

These three requirements are dealt with in more detail in the following paragraphs.

Amplification.—A comprehensive pre-amplifier of the type described later in this chapter must be suitable for use with microphones, tape playback heads and low-sensitivity pickups.

The output voltage of a typical crystal microphone of the diaphragm type connected directly to the grid of the first valve is about 2mV for a sound pressure of 1 dyne/cm². This is roughly equivalent to loud speech a foot or so from the microphone and normal conversation will give about 0·5mV. Signal magnitudes of this order will also be obtained at the input valve grid from a moving-coil or ribbon microphone connected via an appropriate step-up transformer. With such low level input voltages the hum and noise level at the grid must not exceed about 1·5µV for a signal-to-noise ratio approaching 50db (300:1), which is an acceptable standard for high-fidelity equipment.

The output from tape playback heads varies from about 1·5mV, at 1000c/s, for the cheapest kinds up to about 10mV for the more expensive heads. Obviously, the greater the output, the better will be the signal-to-noise ratio obtained with any given tape. Magnetic tape is recorded at constant peak flux density so that the head output falls with decreasing frequency at the rate of 6 db per octave. Consequently, the output at 50c/s will be about 24db less than that at

1000c/s, that is, about 0·1mV for a low-output head.   It follows that, with this type of head, a signal-to-noise ratio of 40db (100:1) will necessitate a noise level at the grid no higher than about $1\mu V$.

An input stage which meets the requirements of high signal-to-noise ratio with adequate gain can be built by using the Z729, a valve especially designed for this type of application.   The pre-amplifier circuits of figs. 9-2 and 9-3 both contain a tape playback amplifier of this kind, the valve being V1 in each case. Certain precautions must be observed, particularly with the wiring of grid, heater and cathode circuits and in the use of suitable components in order to take full advantage of the performance of the valve itself.

The grid and cathode wiring must be compact and direct in order to mini-mise the electrostatic pick-up of hum and the magnetic loop formed by the input terminal, grid, cathode and earth must be as small as possible.   A single earth connection is essential and this should be close to the input socket, which should be screened.   The part of the chassis containing the input stage may, with advantage, be totally enclosed with metal screens.   The heater wiring should consist of a twisted pair and should be kept well clear of the grid circuit.   The heaters should be wired up from the supply end starting with the output valve and finishing with the input stage.   In this way, the heater supply wires running to the first stage will carry only the current drawn by the input valve and the electro-magnetic field about these wires will be at a minimum.

Care in the choice of components used in the input stage is necessary and all capacitors should be of high quality if they are not to introduce noise due to variable insulation resistance etc.   Resistors should be of the high-stability type in order to minimise noise produced by thermal agitation (Johnson noise) and other effects.   The valve socket should be of unloaded plastics or ceramic material and should preferably contain a central screening boss, this being earthed together with contacts 2 and 7 which connect to the internal screening system of the Z729 valve.

Equalisation.   As explained above, the frequency characteristic of the output from a tape playback head or gramophone pickup is not level and a suitable equalising network must be incorporated in the pre-amplifier in order to re-produce these signals with a flat frequency characteristic.

Such a network introduces a loss in amplification which must be made good and it is convenient to combine it with a valve amplifier, the amplification of which makes up for the insertion loss of the correcting circuit.   By arranging the equalising network in the form of a negative feedback loop in this amplifier stage, valve distortion is reduced to a minimum.

The equalisation basically comprises bass boost of 10-15db at 50c/s for record reproduction and 24db or more at 50c/s for tape playback.   This is achieved by suitably attenuating the higher frequencies, full gain being obtained only at the lowest frequencies.   Fig. 9-1 is a simplified diagram of this circuit,
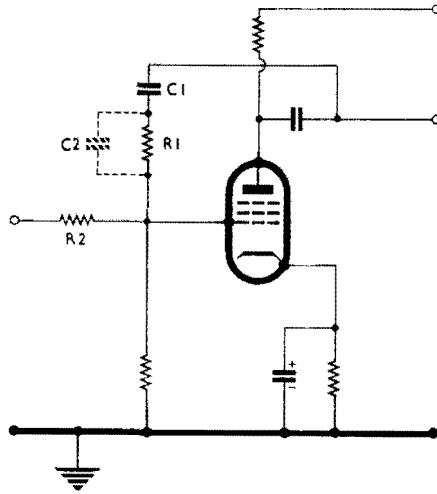
Fig. 9-1. Simplified circuit showing
how the response of an amplifier may
be modified by placing a suitable net
work in the feedback loop.

the corresponding valves in the full circuits of figs. 9-2 and 9-3 being V2 in each case.

The feedback loop can be made to modify the response of this amplifier stage in any desired way by introducing into it a suitable frequency selective network. To obtain bass boost, this network consists basically of C1 and R1 in fig. 9-1 and the values of these components are chosen so that the impedance of C1 becomes greater than the resistance of R1 at a particular low frequency. The amount of feedback thus becomes smaller with decreasing frequency and the amplification increases, giving bass boost. For record reproduction, high frequency de-emphasis is introduced by the addition of C2.

In the complete pre-amplifiers, a suitable equalising network is automatically switched into circuit for each type of input source. For radio reproduction the gain of this stage is reduced uniformly at all frequencies, by introducing pure resistance into the feedback loop, to suit the larger input signal from a radio tuner.

As stated above, when the signal is from a tape playback head considerable attenuation of all but the lower frequencies is necessary for true equalisation and an additional stage must precede the equalising amplifier. The requirements for this input stage were discussed under " Amplification " on page 97.

An additional amplifier is not necessary for record equalisation because the required bass boost is considerably less and, normally, the input signal is larger.

The gain control follows the equalising stage as the signal level is such that any noise generated by a potentiometer may be neglected.

Tone Control.—Having achieved a signal of uniform frequency characteristics from any of the input sources by the action of the equalising stage, tone controls can be introduced for adjusting the balance and frequency range to suit listening conditions, programme quality and personal taste.

Both pre-amplifier designs incorporate tone controls, those in the circuit of fig. 9-3 being somewhat more comprehensive. The details of the tone control circuits are given in the descriptions of the pre-amplifiers below.

9-2. Pre-Amplifier Circuit No. 1.—The circuit diagram of this pre-amplifier is given in fig. 9-2. This design, the simpler of the two described in this chapter, utilises three Z729 valves and is suitable for feeding an amplifier requiring about 0·5V input. V1 is the tape playback amplifier and the sensitivity at the input socket of this stage is 1mV. The output of this stage and the input signals from the pickup and radio sockets are connected to the grid of V2 via the input selector switch S1A.

COMPONENT VALUES FOR FIG. 9-2
PRE-AMPLIFIER CIRCUIT NO. 1

G.E.C. VALVES

V1 ⎫
V2 ⎬ Z729
V3 ⎭

RESISTORS

(10%, 0·25W unless otherwise shown)
*R1 2·2MΩ 5%
*R2 5% ⎱
*R3 5% ⎰ See page 102
*R4 3·3kΩ 5%
*R5 47kΩ 5%
*R6 220kΩ 5%
*R7 470kΩ 5%
*R8 220kΩ 5%
*R9 1MΩ 5%
*R10 470kΩ 5%
R11 220kΩ
R12 220kΩ
R13 47kΩ
R14 470kΩ
R15 1MΩ
R16 10MΩ
R17 1MΩ
R18 10MΩ
R19 220kΩ
R20 22MΩ
*R21 3·3kΩ 5%
*R22 220kΩ 5%
*R23 1MΩ 5%
R24 500kΩ Log.
R25 2·2MΩ
R26 47kΩ
R27 1·5kΩ 5%

High Stability

R28 100kΩ
R29 470kΩ
R30 220kΩ 5%
R31 1MΩ Log.
R32 22kΩ 5%
R33 22kΩ

CAPACITORS

C1 16μF 350V
C2 25μF 25V
C3 0·1μF
C4 0·05μF
C5 33pF 10%
C6 330pF 10%
C7 68pF 10%
C8 220pF 10%
C9 470pF 10%
C10 25μF 25V
C11 0·1μF
C12 0·05μF
C13 8μF 350V
C14 0·005μF
C15 25μF 25V
C16 0·1μF
C17 0·05μF
C18 2000pF 10%
C19 0·02μF
C20 16μF 450V
C21 1000pF 10%
C22 2000pF 10%
C23 0·05μF

SWITCHES

S1 2-pole 4-way m.b.b.
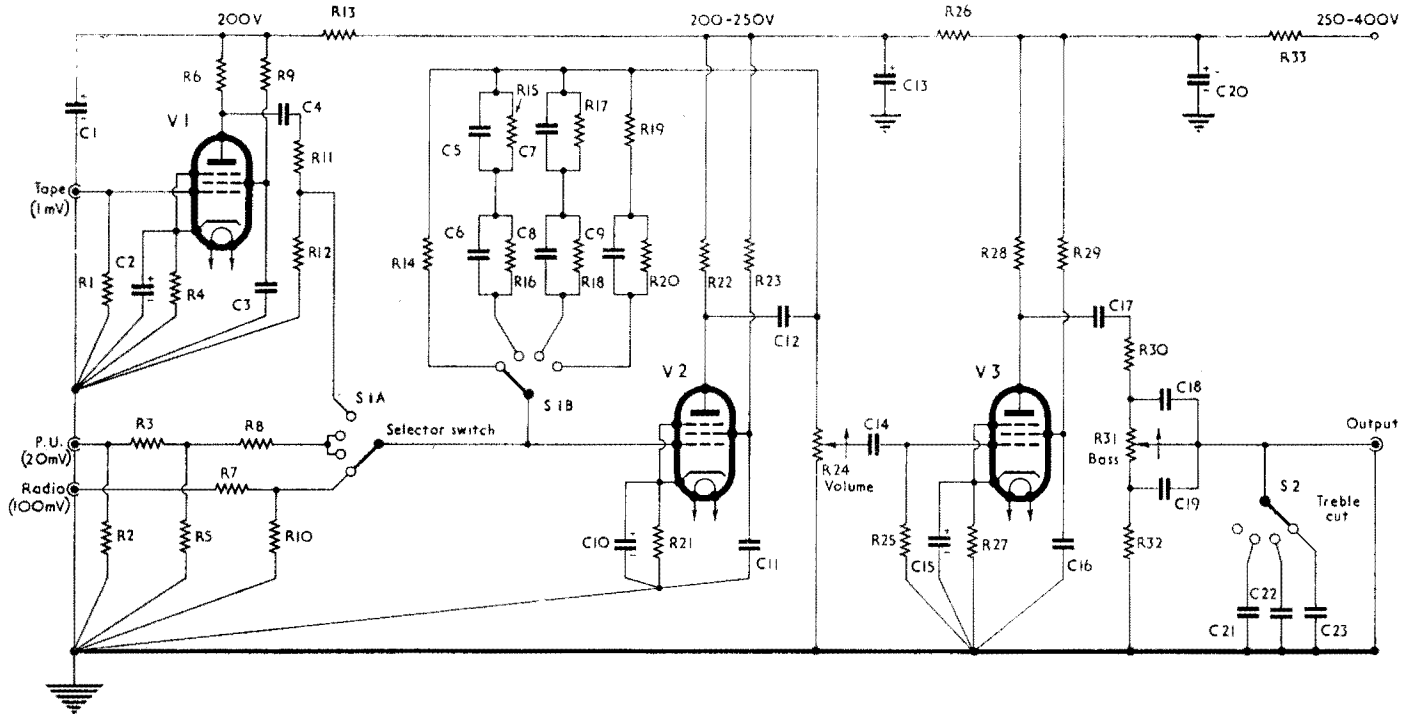S2 1-pole 4-way m.b.b.

Fig. 9-2. Circuit of Pre-Amplifier No. 1. All switches are shown fully anticlockwise and the arrows at the potentiometers indicate clockwise rotation.

Switch S1 has four positions and in clockwise order these are : radio (100mV), pickup 78 r.p.m. (20mV), pickup 33 r.p.m. (20mV) and tape playback (1mV). S1A selects the required input signal and S1B automatically inserts into the feedback loop of V2 the appropriate equalising network.

The values of resistors R2 and R3, in the pickup input circuit, will depend upon the type of pickup used. Variation of the value of R2 will alter the load impedance presented to the pickup and variation of R3 will alter the input sensitivity to suit the output of the pickup. With many moving iron (variable reluctance) pickups R2 should be open-circuited (R2 = $\infty\Omega$) and R3 short-circuited (R3 = OΩ) but if this results in full output with the volume control at less than half-travel, a suitable value of R3 will attenuate the pickup output and so prevent distortion in V2. A value of 47kΩ for R2 will provide a suitable load impedance for most magnetic pickups. The output of a crystal pickup can be made similar to a magnetic type by making R2 100kΩ and reducing its output by making R3 1MΩ (see fig. 9-8).

The tone control stage consists of V3 and its associated bass and treble controls R31 and S2. Both controls utilise conventional circuits, the bass control being a potentiometer which boosts the bass as the slider travels away from the centre in one direction and cuts as it is moved in the opposite direction, with the centre position giving a level response. The treble control consists of a four-position switch giving three degrees of cut and one level position.

The input sockets and the first valve should be grouped together with a single earth connection to the chassis. The earth connection from the equalising stage (V2) should be returned to the same point. The two selector switches may be of the bakelite wafer type, with make-before-break contacts.

The grid coupling capacitors are chosen so that their combined effect produces attenuation below 30c/s in order to reduce rumble.

As the output impedance of the pre-amplifier is about 22kΩ, up to four feet of ordinary screened wire or up to twenty feet of coaxial cable may be connected without high frequency loss.

This pre-amplifier may be used as a substitute for the built-in tone controls of the 912-PLUS amplifier when the flexibility of a separate control unit is preferred. The h.t. feed for the pre-amplifier is taken from the junction of C11 and R28 in fig. 3-3 (page 28). The output of the pre-amplifier, fed into the first grid of V2 in fig. 3-3, provides the correct input signal for the 912-PLUS.

9-3. Pre-Amplifier Circuit No. 2.—This pre-amplifier was designed for use with the KT88 50W amplifier described in Chapter 5* but it may be used to feed any power amplifier requiring an input signal of 0·5V. It is very suitable for use with the G.E.C. 912-PLUS amplifier as a substitute for the built-in tone control stage. Details of its use with the 912-PLUS are given in Chapter 3.

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*A detailed article giving constructional information was published in the July 1957 issue of *Wireless World* under the title *The ' 88-50 ' Pre-Amplifier* by W. I. Heath and D. M. Leakey.

The pre-amplifier circuit is illustrated in fig. 9-3 and it is designed to operate from a gramophone pickup, a tape playback head (for which a separate input stage, V1, is used), a high-impedance microphone and a radio tuner. Choice of any of these inputs is provided by a selector switch which also automatically modifies the pre-amplifier sensitivity and equalisation to suit the selected input source.

The six positions of the input selector switch are given in clockwise order by the table below which also shows the input sensitivity, input impedance and an indication of the equalisation obtained at each position.

Switch Position	Sensi- tivity*	Input Imped- ance	Frequency Correction	
			Bass Boost at	Treble Cut at
Radio	0·2V	470kΩ	—	—
78 r.p.m. Records	14mV	50kΩ	400c/s	6kc/s
British Microgroove Records ...	10mV	50kΩ	500c/s	3kc/s
American Standard Records ...	12mV	50kΩ	600c/s	2kc/s
Tape (7½in/s)	4mV	220kΩ	1·5kc/s	—
Microphone (high impedance) ...	1mV	1·1MΩ	—	—

*Input for 0·5V r.m.s. output from pre-amplifier.

Equalisation Networks.—The equalising stage is of the negative feedback type as described in Sect. 9-1 above and it utilises another G.E.C. Z729 low noise pentode. The various equalisation characteristics are illustrated in figs. 9-4 to 9-7. The tape playback equalisation is based on the recommendation that tape recorded at 7½in/s should be played back with a time constant of 100μs and that treble loss in recording should be offset by pre-emphasis.† This is equivalent to a bass boost of 6db per octave from 1500c/s and is obtained by C10 and R19 in the feedback loop. At other tape speeds, these components should be as follows :

Speed	*Time Constant*	*C10*	*R19*
3¾in/s	200μs	1000pF	220kΩ
15in/s	35μs	150-220pF	220kΩ

†Amendment No. 1, July 1954, B.S.1568:1953.

H

COMPONENT VALUES FOR FIG. 9-3

PRE-AMPLIFIER CIRCUIT NO. 2

G.E.C. VALVES
V1 Z729
V2 Z729
V3 B329/12AU7

RESISTORS
(10%, 0·25W unless otherwise shown)
R1 470kΩ
*R2 220kΩ 5%
R3 470kΩ
R4 10MΩ
R5 820kΩ 5%
R6 470kΩ
*R7 3·3kΩ 5%
*R8 10kΩ 5%
R9 22MΩ
R10 1·5MΩ 5%
R11 2·2MΩ
R12 2·2MΩ
R13 1MΩ
R14 1MΩ 5%
R15 10kΩ
R16 100kΩ (see page 107)
R17 22MΩ
R18 220kΩ
R19 220kΩ
*R20 100kΩ 5%
R21 22kΩ
*R22 3·3kΩ 5%
*R23 220kΩ 5%
R24 1MΩ
R25 500kΩ Log.
R26 22kΩ
R27 1kΩ
R28 47kΩ
R29 100kΩ
R30 1MΩ Log.
R31 10kΩ
R32 47kΩ
R33 500kΩ Log.
R34 3·3kΩ
R35 2MΩ Log.
R36 100kΩ
R37 100kΩ
R38 680kΩ
R39 1·5MΩ
R40 1kΩ
R41 47kΩ
R42 10kΩ
R43 150kΩ

High stability

R44 100kΩ
R45 100kΩ
R46 47kΩ
R47 100kΩ
R48 10kΩ
R49 100kΩ
R50 100kΩ
R51 2·2kΩ
R52 100kΩ
R53 100kΩ
R54 2·2kΩ

CAPACITORS
C1 25μF 12V
C2 470pF 5%
C3 33pF 5%
C4 220pF 5%
C5 33pF 5%
C6 0·02μF
C7 82pF 5%
C8 100pF 5% (see page 107)
C9 0·05μF
C10 470pF 5%
C11 8μF 450V
C12 8μF 450V
C13 25μF 12V
C14 0·05μF
C15 0·1μF
C16 25μF 12V
C17 0·1μF
C18 5000pF 5%
C19 0·04μF
C20 1000pF 5%
C21 0·02μF
C22 680pF 5%
C23 330pF 5%
C24 100pF 5%
C25 47pF 5%
C26 470pF 5%
C27 220pF 5%
C28 0·002μF
C29 25μF 12V
C30 0·02μF
C31 8μF 500V
C32 220pF 5%

SWITCHES
S1 2-pole 6-way m.b.b.
S2 3-pole 3-way m.b.b.
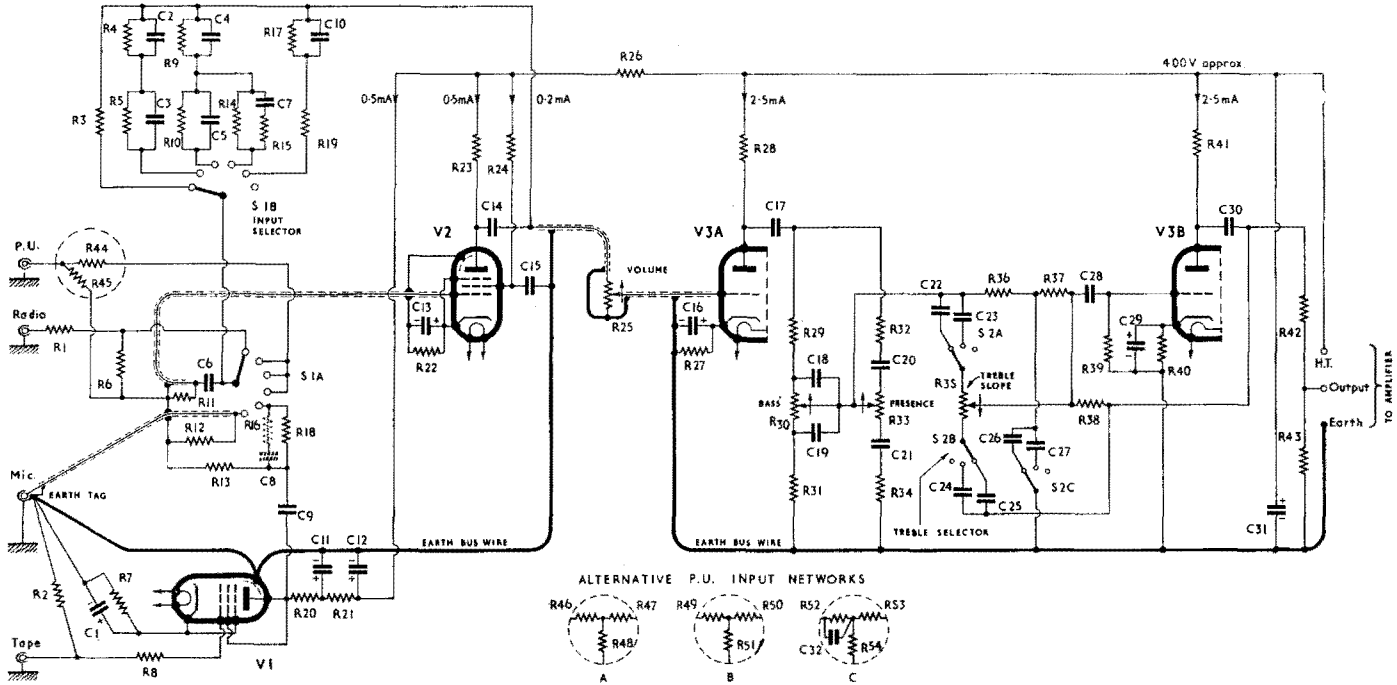
Fig. 9-3. Circuit of Pre-Amplifier No. 2. All switches are shown fully anticlockwise and the arrows at the potentiometers indicate clockwise rotation. The circuit is arranged to show the method of connecting all earth returns to a bus which is earthed to the chassis only at the microphone input socket. Details of the alternative pickup input networks are given on page 108.
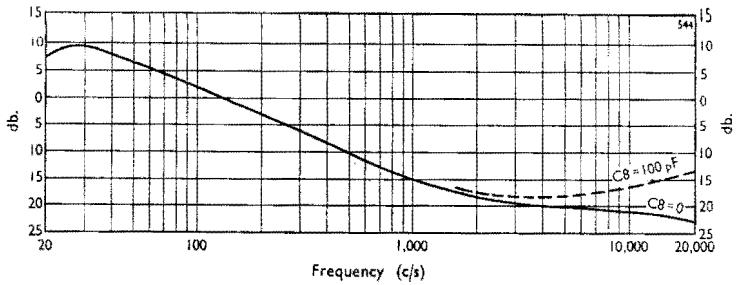
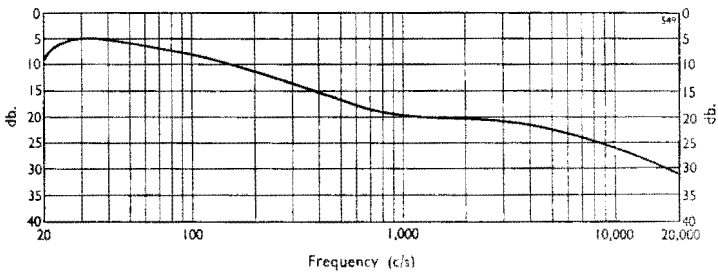**Fig. 9-4.**   Tape equalisation characteristic.   See page 107 for C8.

Fig. 9-5.   British 78 r.p.m. equalisation characteristic.
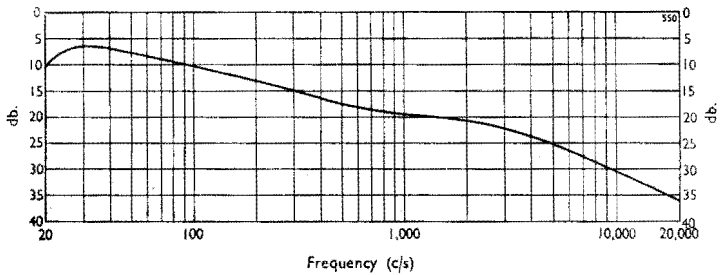
Fig. 9-6.   British Microgroove equalisation characteristic.
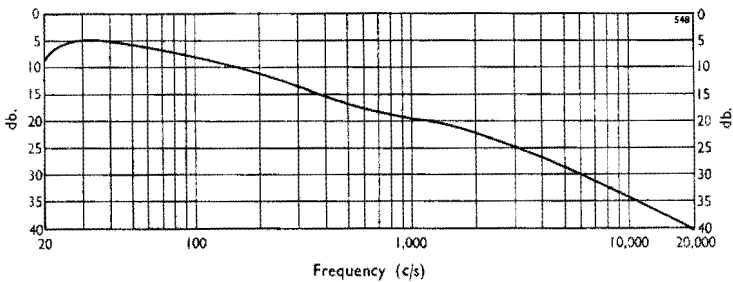
Fig. 9-7.   American Standard equalisation characteristic.

The selector switch may be provided with more positions if these additional equalisation characteristics are required to be built into the pre-amplifier. It should be noted that R17 is connected across C10 merely to eliminate switch clicks and does not form part of the equalisation circuit.

In general, it is undesirable to compensate for tape losses by treble boosting on playback as this will accentuate tape hiss. However, a degree of treble boost on playback is an advantage when the playback head itself is responsible for high frequency loss. A suitable amount of treble boost may be introduced by shunting R18, in the output circuit of V1, by C8 as shown in broken line in fig. 9-3. C8 should not exceed 100pF for 7½in/s and the effect of this capacitor is shown in fig. 9-4. The limiting resistor R16, in series with C8, should be 100kΩ.

Gramophone record equalisation is obtained in the same way as for tape, e.g. for American Standard records bass boost is provided by C4 and R14 with treble roll-off or de-emphasis by C7. Resistor R15, in series with C7, is inserted to limit feedback at very high frequencies which would cause instability. It was not found necessary to include this resistor in the other feedback paths. Resistors R4, R9 and R17 eliminate switch clicks and, in addition, limit the bass boost.

The record equalisation networks are designed for use with moving coil or moving iron (variable reluctance) pickups which give an output voltage proportional to recorded velocity. The pre-amplifier input impedance and sensitivity are suitable for use with most pickups of this type, which give a maximum output of 10mV to 50mV. When using a pickup of different characteristics, such as a crystal type, the input network must be appropriately modified. Normally, with the optimum load impedance of about 1MΩ, a good quality crystal pickup gives an approximately flat response from a record, in other words it provides its own bass boost to compensate the falling bass characteristic of the record. However, it has been found that a very smooth frequency response, extending up to the high frequency peak, can be obtained if a crystal pickup is connected to a
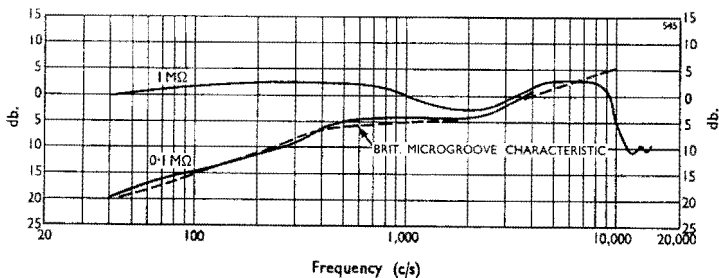

Frequency (c/s)

Fig. 9-8. Response of a typical crystal pickup connected across the normal load impedance of 1MΩ and compared with that obtained with a load of 100kΩ. The British Microgroove recording characteristic is also shown for comparison with the 100kΩ curve.

load lower than the optimum.   The resulting response is very similar to that of a velocity pickup and it is then corrected by the equalising network as for a magnetic type.   Fig. 9-8 shows the response of a Collaro " Studio P " crystal pickup to a British microgroove test record plotted with a normal load of 1MΩ and with a load of 100kΩ.   The fall in bass in the latter curve is caused by the internal capacitance of the pickup and the result closely resembles the British Microgroove recording characteristic, also shown in the graph.

As the output of a crystal pickup is considerably greater than that of a magnetic type, the input network of the pre-amplifier must be modified to attenuate the signal to the level required for correct operation of V2.   Inset in fig. 9-3 are three alternative pickup input networks.   Inset " A " is for use with magnetic pickups having a maximum output greater than 50mV.   Maximum output is here defined as that given by a recorded velocity of 7cm/s and corresponds to high modulation on an average record.   Inset " B " is for use with good-quality popular crystal pickups, such as the Collaro " Studio P." Inset " C " is for best quality crystal pickups having a less pronounced treble peak, such as the Collaro " Transcription."

Tone Control.—The gain control is inserted after the equalising stage and the various tone controls are built into a following double triode amplifier stage. V3A feeds two frequency selective potential divider circuits which provide variable control of bass and " presence."   Logarithmic controls are used for R30 and R33 in order to give a level response when their sliders are at the mid position.   The effect of these controls is shown in fig. 9-9.

The presence control alters the level of all frequencies above about 1500c/s. It therefore alters the balance between high and low frequencies so that an increase in presence, i.e. turning the control clockwise from mid-travel thereby lifting the high frequencies, gives a more " forward " quality.   This is as though the listener were to move nearer the orchestra or speaker.   Conversely, anticlockwise rotation from the mid position simulates the effect of an orchestra receding and approximates to listening from the back of the concert hall.

The signal passes from these controls to V3B, to which is applied negative feedback, the loop consisting of R38 in conjunction with isolating resistors R36 and R37.   Associated with these resistors are the treble boost capacitors C22 and C23, and the treble cut capacitors C24 to C27.   Appropriate combinations of these capacitors are switched into circuit by the treble selector S2 and the response curve so obtained may be modified by the treble slope potentiometer R35.   This control is inoperative when S2 is at the level position but an overall cut or *boost* is obtained by rotating the potentiometer when S2 is at either of the *cut* positions.   The variation in treble response with rotation of the treble slope control at both " cut " positions of S2 is shown in figs. 9-10 and 9-11.

If a normal logarithmic potentiometer is used for treble slope, it will attenuate the treble as it is rotated clockwise.   An inverse log potentiometer must be used to obtain increasing treble with clockwise rotation if this is preferred.
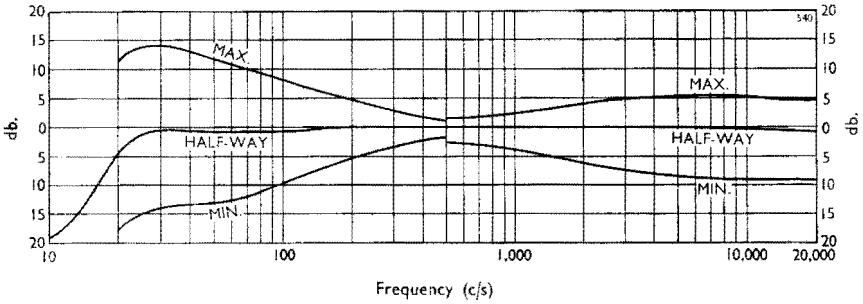
Frequency (c/s)

Fig. 9-9.   Effect of the bass and presence controls. The pronounced attenuation
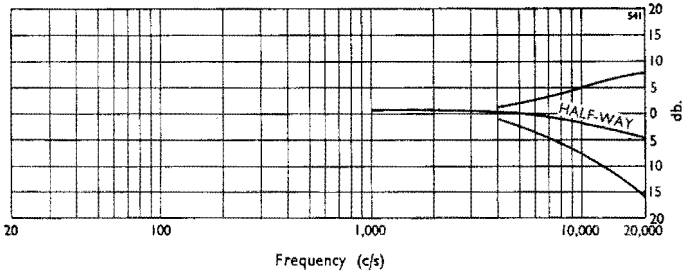below 30c/s is the result of the rumble filter (see overleaf).

Frequency (c/s)

Fig. 9-10.   The treble slope control gives this range of response at the first " cut "
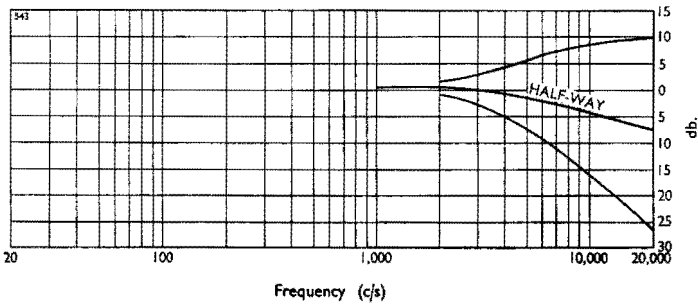position of the treble switch.

Frequency (c/s)

Fig. 9-11.   Range of response obtained with the treble slope control at the second
" cut " position of the treble switch.

Rumble Filter.—Within the feedback loop of V3B are two coupling networks, C28 R39 and C30 R42 R43, each having a time constant of 3ms.   As a result, low frequencies are attenuated and positive feedback is introduced below 50c/s, the combined effect giving a flat response down to 30c/s below which the response falls steeply down to about 15c/s, when the coupling capacitors elsewhere in the pre-amplifier provide additional attenuation.

The effect of the rumble filter is shown in fig. 9-9.   If it is desired to attenuate below 40c/s instead of 30c/s, R39 and R43 should be made 1MΩ and 100kΩ respectively.

Output Impedance.—The negative feedback on V3B maintains a low output impedance over the greater part of the frequency range of the pre-amplifier. The actual value is approximately equal to R42, i.e. about 10kΩ and up to five feet of ordinary screened wire may be connected to the output terminal with no deterioration in high frequency performance.

Construction.—The circuit diagram, fig. 9-3, is drawn in a way that illustrates the method of earthing all the components to a continuous earth wire which is connected to the chassis only at a tag mounted on the microphone input socket.

The two switches are bakelite wafer types, S1 having each pole on a separate wafer whilst S2 is a single wafer 3-pole 3-way switch.

The pre-amplifier earth bus wire should be connected to the main amplifier chassis only by a single wire which may conveniently be the screening braid on the signal output cable.   This cable itself may be one of the elements of a multi-core cable between pre-amplifier and main amplifier.

The precautions to be taken in constructing a pre-amplifier of this type are described on page 11 and if these are borne in mind and the circuit diagram intelligently interpreted, the construction should be quite straightforward.   The original article, mentioned on page 102, in which constructional information and illustrations were given, should be consulted by those with little experience in building this type of equipment.

Power Supplies.—The pre-amplifier was designed to draw its power supplies from the KT88 50W amplifier (Chapter 5) and the h.t. line smoothing is the minimum necessary for adequate ripple attenuation in order to maintain the h.t. as high as possible.   When using the KT88 amplifier, no added h.t. decoupling is necessary because the stabilising circuits in that amplifier and its good push-pull balance contribute greatly to its freedom from instability at low frequencies. It is possible that further smoothing and/or decoupling may be required when using the pre-amplifier with a different type of power amplifier.   This point is covered in Chapter 3 when using the pre-amplifier as part of the G.E.C. 912-PLUS (see page 33).

Performance.—The KT88 50W amplifier requires 0·5V r.m.s. input for maximum output and the pre-amplifier will supply this signal at a harmonic

distortion comparable with that of the KT88 amplifier.   The distortion does not increase with variation of the controls but to avoid the possibility of distortion in the equalising stage (V2) all input levels must be adjusted so that the full output of 50W from the KT88 amplifier (corresponding to 0·5V pre-amplifier output) is not obtained until the pre-amplifier gain control is beyond the centre of its travel.   Given this condition, an increase in input signal level of nearly 20db can be accepted with no increase in distortion.

The signal-to-noise ratio of the complete equipment comprising the pre-amplifier and the KT88 amplifier is —76db relative to 50W with the gain control at minimum and the tone controls set for level response.   With the gain control at maximum and the input sockets short-circuited to earth, the following signal-to-noise ratios were obtained at each position of the input selector on a prototype :   Radio :   —69db ;   A.S. Records :   —63db ;   Tape :   —52db ; Microphone :   —55db.

APPENDIX A

MULTIPLE-PAIR PUSH-PULL AMPLIFIERS

When an output exceeding 100W is needed, the KT88 valve may be used in multiple pairs in parallel push-pull instead of a single pair of larger valves.* One of the advantages of this method is the low cost of the power supply which is required to give an h.t. of only 550V. Another advantage is that a valve failure in the output stage merely reduces the available output power with a probable increase in distortion.

The circuit diagram of fig. A-1 illustrates the use of ten valves in a fixed bias ultra-linear output stage and this arrangement gives 400W output. More or less than five pairs of valves can be used, depending upon the power required.

A single control is used for grid bias adjustment and this simplifies the amplifier at the expense of somewhat higher distortion and lower output. For this to be successful the power supply must have good regulation and a suitable circuit is included in fig. A-3.

It is not essential to use accurately matched valves but the total current in each half of the push-pull stage should be equalised as closely as possible. This becomes easier with an increasing number of pairs and is facilitated by the cathode current meter built into the amplifier. The individual cathode currents will vary from about 35mA to about 60mA and each valve should be measured in turn and the valves sorted into two groups of approximately similar total current.

When the valves have been sorted and the two groups plugged into their respective halves of the amplifier, the optimum operating condition is obtained by adjusting the grid bias with R5 (fig. A-3) so that the current drawn by any single valve does not exceed 60mA. In this way an output of 400W will be obtained at a distortion of about 5%.

When using six or more valves in this type of amplifier, the value of the grid return resistors R14 and R15 is of importance and a low value is desirable. In order to facilitate the production of the necessary distortion-free input signal of 55+55V r.m.s., the output stage is driven by a pair of cathode followers V2B and V3B. With four output valves the grid resistors could be increased to

*A detailed article describing this type of amplifier: *400-watt Audio Amplifier* by G. R. Woodville, was published in the November 1957 issue of *Wireless World*.

100kΩ and the cathode followers dispensed with.   However, as they form parts of double triodes, the saving in cost is insignificant and, on the whole, it is preferable to retain them.

The method of measuring the cathode current of each KT88 valve is shown in fig. A-1.   A resistor of 10Ω is inserted in the cathode lead of each valve (i.e. R38, R39, etc.) and a meter M1 is connected across this resistor through switch S1 and a series resistor (R48, R49, etc.).

It was found convenient to use a meter with a full-scale deflection of 200μA, the value of the series resistor being such that the meter indicated 0-200mA. At full output each cathode current is about 100mA to 125mA.   It may be preferred to substitute the individual cathode series resistors with a single resistor inserted between the meter and switch but the possibility of instability should be borne in mind as the individual resistors act as cathode circuit isolators.

The ultra-linear output transformer must have low leakage inductance between : primary and secondary ; half-primary and half-primary ; and each half-primary and its tap.   The absolute values of leakage inductance will depend upon the number of valves used but a 400W transformer used in the prototype had the following characteristics :

Primary Inductance :	4H
Leakage Inductances :	
Primary to secondary :	0·75mH
Half-primary to half-primary :	0·75mH
Each half-primary to tap :	1·5mH

To prevent ultrasonic oscillation, resistor/capacitor combinations are connected between each tap and the anode terminal of each half-primary.   In the prototype, C10 and R60 were also found to be desirable, the values used being 3500pF and 1kΩ respectively.

Negative feedback may be added to this amplifier in the normal way, from the secondary of the output transformer to the cathode circuit of the input valve. The values of the resistors R2 and R3 in fig. A-1 are determined by the amount of feedback and the ratio of the transformer.

Fig. A-2 illustrates a recommended layout for the output valves and is self-explanatory.   Adequate ventilation should be provided ; if in doubt, a temperature sensitive paint should be used (see footnote on page 9).

Power Supply.—The design of the power supply is an important factor in the satisfactory operation of an amplifier of this type.   The regulation should be good ; better than 10% with a current variation of 400-1200mA was obtained in the prototype.   This order of regulation was achieved by using the xenon filled rectifier GXU1 which, in this application, is considerably under-run at a PIV of 1kV as against the rated PIV of 10kV.   For up to six output valves, the smaller xenon rectifier GXU50 is suitable.

The circuit of a complete power supply is given in fig. A-3 and this provides the lower h.t. voltage required by the earlier stages as well as the grid bias supply.
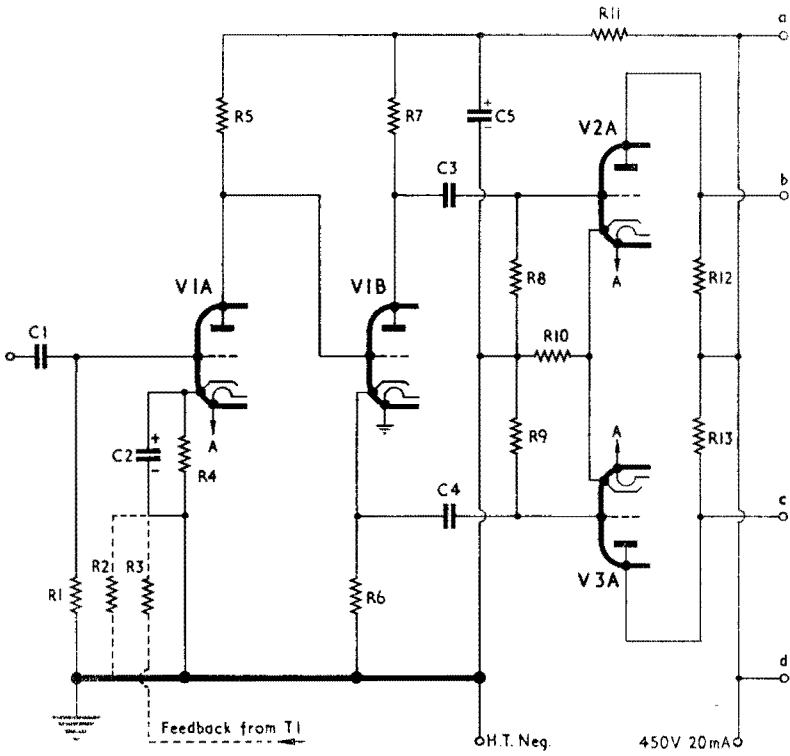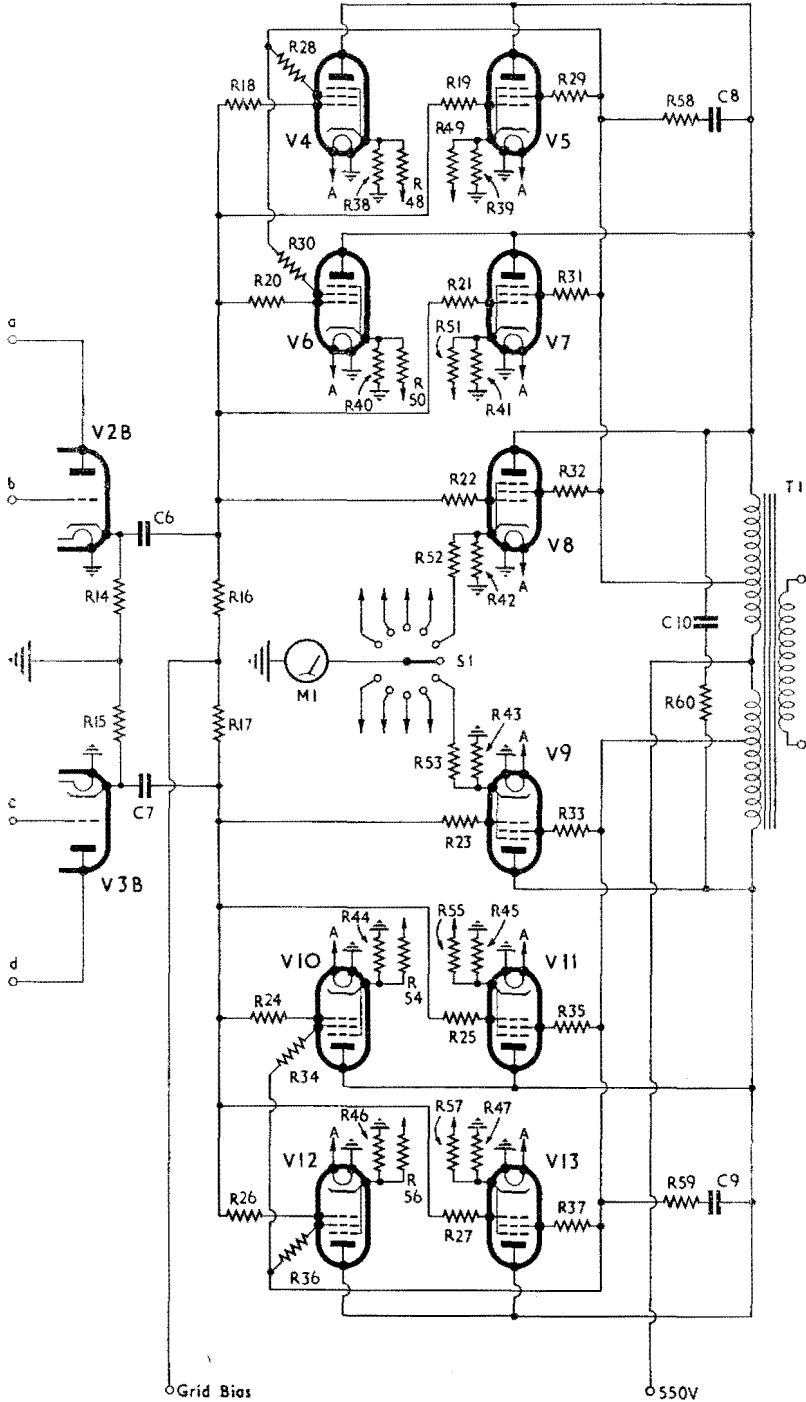
Fig. A-1. Circuit of the 400W multiple-pair amplifier, Component values are given on page 117.

A single inductance-input filter is shown and, with a smoothing capacitance of 150-200μF (obtained by series-connecting two larger capacitors), this should be satisfactory for most purposes. A further filter section may be inserted if desired. The smoothing inductor(s) should have a value of 2-3H and a d.c. resistance of about 25Ω.

TABLE A-I

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. A-1

$V_{a\ (b)\ (o)}$	570	V
$V_{a\ (b)\ (max\ sig)}$	530	V
$V_{a,\ g2\ (o)}$	565	V
$V_{a,\ g2\ (max\ sig)}$	525	V
$I_{a+g2\ (o)}$ (per valve)	35 to 60	mA
$I_{a+g2\ (max\ sig)}$ (per valve)	100 to 125	mA
$I_{a+g2\ (o)}$ (total)	450	mA
$I_{a+g2\ (max\ sig)}$ (total)	1200	mA
$p_{a+g2(o)}$ (per valve)	35	W
$p_{a+g2\ (max\ sig)}$ (per valve)	25	W
V_{g1} (approx)	—75	V
P_{out}	400	W
$R_{L\ (a-a)}$	1	kΩ
z_{out}	1·2	kΩ
D	5 to 7	%
$V_{in\ (rms)}$ to first stage	250	mV

If 10db of negative feedback is applied, the last three values become 400Ω, 2% and 750mV respectively.
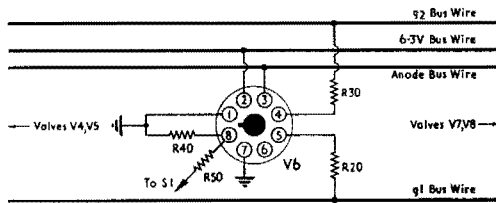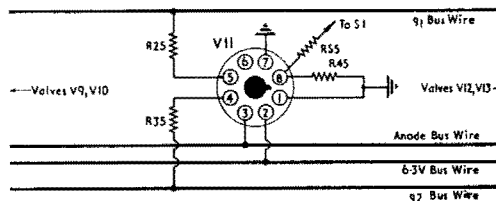

Fig. A-2. Recommended layout for the output valves of the 400W amplifier.

COMPONENT VALUES FOR FIG. A-1

KT88 400W AMPLIFIER

G.E.C. VALVES
V1-V3 B65/6SN7
V4-V13 KT88

RESISTORS
(20%, 0·25W unless otherwise shown)
R1 1MΩ
R2 ⎫
R3 ⎬ See page 113
R4 1kΩ
R5 100kΩ 0·5W
R6 15kΩ 0·5W ⎫
R7 15kΩ 0·5W ⎬ matched to 5%
R8 1MΩ
R9 1MΩ
R10 4·7kΩ 10% 0·5W
R11 33kΩ 0·5W
R12 220kΩ 1W ⎫ matched to 5%
R13 220kΩ 1W ⎬
R14 47kΩ 1W ⎫ matched to 5%
R15 47kΩ 1W ⎬
R16 47kΩ 10% 0·5W
R17 47kΩ 10% 0·5W
R18-R27 15kΩ
R28-R37 220Ω
R38-R47 Meter shunts, see p. 113

R48-R57 Meter series resistors, see p. 113
R58 1kΩ 2W
R59 1kΩ 2W
R60 See p. 113

CAPACITORS
C1 0·01μF
C2 50μF 12V
C3 0·01μF
C4 0·01μF
C5 8μF 450V
C6 0·25μF
C7 0·25μF
C8 1000pF
C9 1000pF
C10 See p. 113

MISCELLANEOUS
M1 200μA meter (see p. 113)
S1 1-pole 11-way switch b.b.m.
T1 400W Ultra-linear transformer
    1 kΩ anode-anode
    Primary inductance : ≮ 4H
    Leakage inductances :
    Prim.—sec. : ≯ 750μH
    ½ prim.—UL tap : ≯ 750μH

COMPONENT VALUES FOR FIG. A-3

POWER SUPPLY FOR KT88 400W AMPLIFIER

G.E.C. VALVES
V1-V4 GXU1
V5 U709 or U78/6X4

RESISTORS
R1 100kΩ 10% 0·5W
R2 100kΩ 10% 0·5W
R3 4·7kΩ 20% 2W
R4 22kΩ 20% 1W
R5 10kΩ wirewound
R6 47kΩ 10% 1W
R7 47kΩ 10% 1W
R8 10kΩ 20% 1W
R9 22kΩ 20% 1W
R10 1·5kΩ 20% 0·25W

CAPACITORS
C1 32μF 350V
C2 32μF 350V

C3 300μF 350V ⎫ each 100+200 dual
C4 300μF 350V ⎬
C5 0·01μF 500V
C6 2μF 250V
C7 2μF 250V

MISCELLANEOUS
L1 2H 1200mA 25Ω
T1 700V 1200mA*
T2 Bias transformer
    Secondaries
    150-0-150V 10mA
    6·3V 20A
T3 2·5V 5A
    2·5V 5A
    2·5V 10A (or 2×5A)
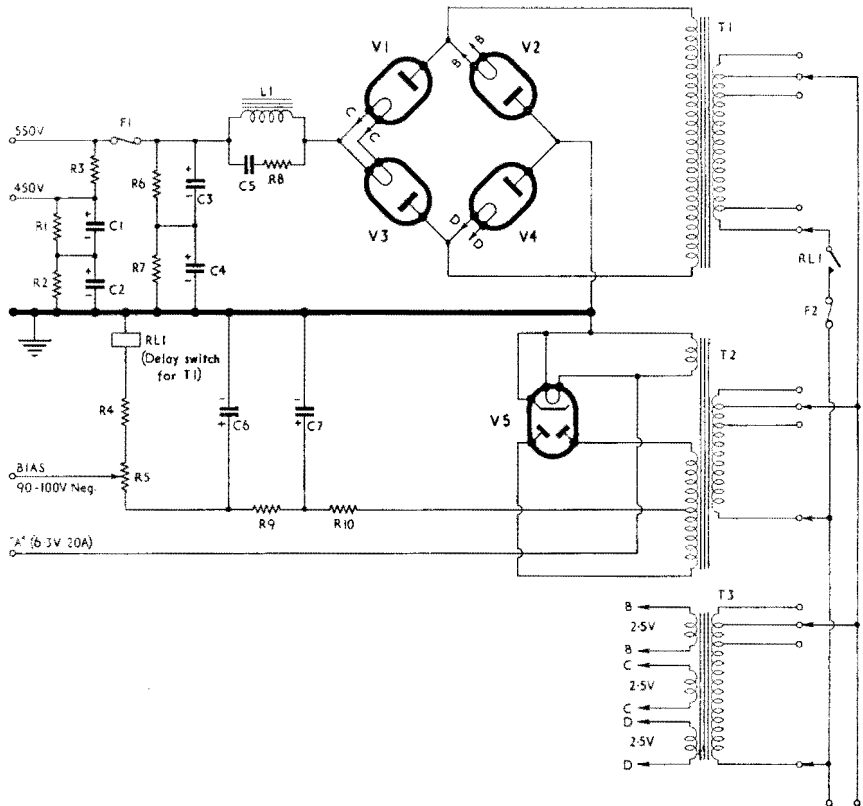F1 2A
F2 5A

    **10V taps on primary*

Fig. A-3. Power supply circuit for the 400W multiple-pair amplifier. The winding of T3 is simplified if four identical secondaries are used. To prevent mains-borne interference with nearby receivers, generated by the xenon rectifiers, a pair of capacitors (0·05μF, 700V a.c.) should be connected in series across the 700V winding and their junction connected to the negative (earth) line.

APPENDIX B

OUTPUT TRANSFORMERS AND STABILISATION

The output transformers used in the prototypes of the ultra-linear amplifiers described in this book were components of good quality made by various manufacturers. The leakage inductance was found to vary considerably, not so much between units by different makers, as between the dates of manufacture. The most recently made transformers incorporated improved methods of winding which gave very low values of leakage inductance. As a guide to this improvement, the values are tabulated below :

Relative Age of Transformer	Leakage Inductance (mH)			Comments
	P-S	$\frac{1}{2}$P-$\frac{1}{2}$P	$\frac{1}{2}$P-UL tap	
Early	20-25	20	50	a-g₂ capacitors required for stability
Medium	10-15	10	25	Stable without capacitors
Late	5-10	5	5	Stable with 30db of negative feedback

The capacitors mentioned in the last column are those shown connected between the anode and screen of each output valve in the ultra-linear amplifier circuit diagrams in previous chapters. They have a value of 1000-2000pF and are connected in series with resistors of 470-1500Ω.

All amplifiers and transformers were examined with a square wave generator and oscilloscope at a frequency of 5-10kc/s and it was found that the overshoot produced was roughly proportional to the leakage inductance. The earlier transformers produced an overshoot of about 25% and a ringing frequency below 40kc/s. In the later types the overshoot is reduced to less than 10% at a frequency above 100kc/s.

When negative feedback was applied it was found essential with the two earlier groups and desirable with the latest transformers to attenuate the high frequency response of the amplifier. The method of doing this is shown by the circuit of fig. B-1, in which the attenuation is achieved by a capacitor and resistor

in series between the anode of the first triode amplifier stage and earth.   A slightly different arrangement is necessary when the first stage consists of a balanced phase inverter such as the one used in the KT88 50W amplifier of fig. 5-2 (page 55).   In this, the attenuating components (C6, R10 and C7, R11) are incorporated in each half of the phase inverter in order to maintain the balance of the stage.   The principle is the same, however, and the following comments apply to both arrangements.

Referring now to fig. B-1, the optimum values of C1 and R1 will depend upon the degree of attenuation required but, with the latest transformers, values of 500pF and 10kΩ gave an attenuation of 3db at 25kc/s and 10db at 40kc/s. With these values, complete stability was obtained with feedback applied.   With earlier transformers the capacitor and resistor were required to have values up to 2000pF and down to 1kΩ.   All these values assume a valve impedance of about 10kΩ but if V1 in fig. B-1 is a pentode-connected Z729 with an anode load of 100-330kΩ, suitable values for C1 and R1 would be 50-200pF and 100-10kΩ respectively.

To ensure stability at very low frequencies, below the cut-off frequency of the output transformer, it is advisable to insert low frequency attenuation between the first and second stages within the feedback loop.   The recommended arrangement is shown in fig. B-1 and consists of a capacitor and shunt resistor in each grid line between the two stages, designated C4, R4 and C5, R5. Assuming the following grid resistors R6 and R7 to be each 1MΩ, values of 5000pF and 3·3MΩ will attenuate frequencies below 25c/s and will be satisfactory with an output transformer of high primary inductance.

Low values of transformer primary inductance will need correspondingly lower and higher values, respectively, of capacitors and resistors.   Values of 2000pF and 10MΩ will be found adequate for transformers of poor quality in terms of primary inductance.   Low frequency instability (" motorboating ") in amplifiers containing poor-quality output transformers may be eliminated by incorporating these networks.

COMPONENT VALUES FOR FIG. B-1

STABILISING NETWORKS

VALVES

V1 \
V2 ⎰ B65/6SN7

RESISTORS

(20%, 0·25W *unless otherwise shown*)
R1 1-10kΩ
R2 15kΩ   0·5W ⎫ matched to 5%
R3 15kΩ   0·5W ⎭
R4 3·3MΩ ⎫ see text above
R5 3·3MΩ ⎭

R6 1MΩ
R7 1MΩ

CAPACITORS

C1 500-5000pF (see text above)
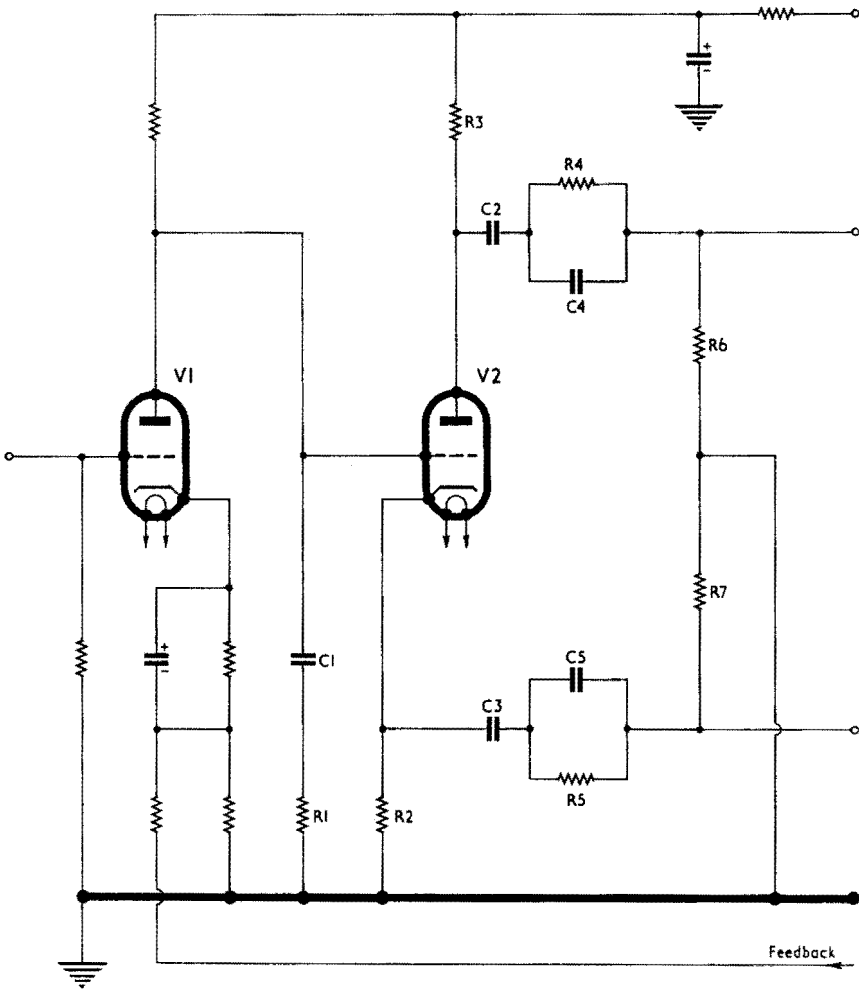C2 0·05μF
C3 0·05μF
C4 5000pF
C5 5000pF

Fig. B-1.   Stabilising networks in the first stages of a feedback amplifier.

APPENDIX C
VALVE RATINGS AND CHARACTERISTICS

AMPLIFYING VALVES

Type	Equivalents		Description	Cathode Type	Filament or Heater		Maximum Ratings			Characteristics						
	U.S.	European			Voltage	Current	V_a	V_{g2}	p_a(W)	V_a	V_{g2}	I_a	$-V_g$	μ	r_a(kΩ)	g_m(mA/V)
B65	6SN7GT	—	Double triode	Heater	6·3	0·6	250	—	2·5	250	—	—	8	20	7·7	2·6
B309	12AT7	ECC81	Double triode	Heater	12·6/6·3	0·15/0·3	300	—	2·5	250	—	—	2·5	55	10·0	5·5
B329	12AU7	ECC82	Double triode	Heater	12·6/6·3	0·15/0·3	300	—	2·75	250	—	—	8·5	17	7·7	2·2
B339	12AX7	ECC83	Double triode	Heater	12·6/6·3	0·15/0·3	300	—	1	250	—	—	2	100	62·5	1·6
B719	6AQ8	ECC85	Double triode	Heater	6·3	0·435	300	—	2·5	250	—	—	2	58	9·7	6·0
DA42	—	—	Class B triode	Heater	7·5	1·2	1250	—	40	1250	—	40	—	72	24·0	3·0
DA100	—	—	Class AB triode	Filament	6·0	2·0	1250	—	100	1000	—	100	—	5·5	1·41	3·9
DH77	6AT6	EBC90	Double diode triode ...	Heater	6·3	0·3	250	—	2·5	250	—	—	3	70	58·0	1·2
KT33C	—	—	Beam pentode	Heater	25/13	0·3/0·6	200	200	13	175	175	—	7	—	22·0	10·0
KT55	—	—	Beam pentode	Heater	52	0·3	400	300	25	200	175	125	—	—	5·0	19·0
KT61	6AG6G	—	Beam pentode	Heater	6·3	0·95	275	275	10	250	250	40	—	—	70·0	10·5
KT66	—	—	Beam pentode	Heater	6·3	1·27	500	400	25	250	250	—	15	—	22·5	6·3
KT88	—	—	Beam pentode	Heater	6·3	1·8	600	600	35	250	250	140	—	—	12·0	11·0
L63	6J5GT	—	Triode	Heater	6·3	0·3	250	—	2·5	250	—	—	8	20	7·7	2·6
LN309	—	PCL83	Triode-pentode {Triode : Pentode :	Heater	12·6	0·3	250 / 250	— / 250	3·5 / 5·4	250 / 165	— / 165	—	8·5 / 9	17 / —	7·7 / 45·0	2·2 / 4·7
N709	6BQ5	EL84	Output pentode	Heater	6·3	0·76	300	300	12	250	250	48	—	—	38·0	11·3
V1505	—	—	Class AB triode	Filament	14	6·5	3000	—	275	2000	—	150	—	16	2·0	8·0
Z729	6267	EF86	Low noise pentode ...	Heater	6·3	0·2	300	200	—	250	140	—	2	—	2·0	1·85

RECTIFIERS

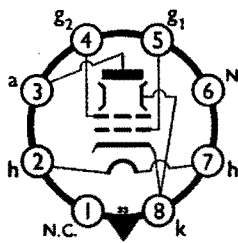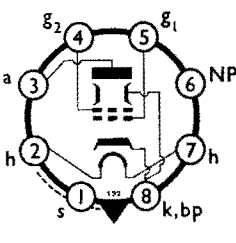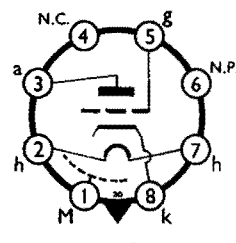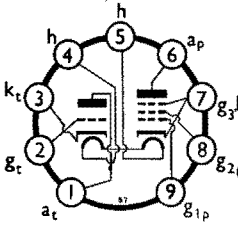| Type	Equivalents		Description	Cathode Type	Filament or Heater		PIV	$V_{a(rms)}$	$i_{a(pk)}$ (mA)	I_{out} (mA)	t_{sd} (seconds)
	U.S.	European			Voltage	Current					
GXU1	3B28	—	Xenon half-wave	Filament	2·5	5·0	10000	—	1000	250	5
GXU50	—	—	Xenon half-wave	Filament	4·0	3·0	5200	1750	1000	250	30
U19	—	—	Vacuum half-wave	Filament	4·0	3·3	7100	2500	1500	250	20*
U31	—	—	Vacuum half-wave	Heater	26·0	0·3	710	250	720	120	—
U50	5Y3G	—	Vacuum bi-phase half-wave ...	Filament	5·0	2·0	1000	350	370	120	—
U52	5U4G	—	Vacuum bi-phase half-wave ...	Filament	5·0	2·25	1430	500	770	250	—
U54	—	—	Vacuum bi-phase half-wave ...	Heater	5·0	2·8	1250	500	1500	250	—
U78	6X4	EZ90	Vacuum bi-phase half-wave ...	Heater	6·3	0·6	1250	325	210	70	—
U709	—	EZ81	Vacuum bi-phase half-wave ...	Heater	6·3	0·95	1000	350	450	150	—
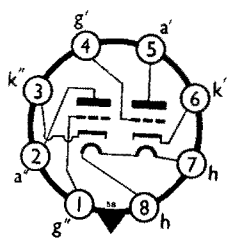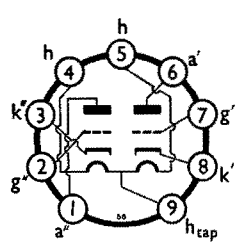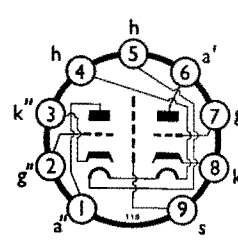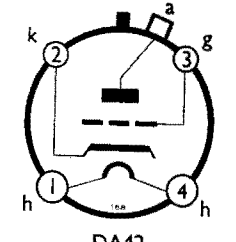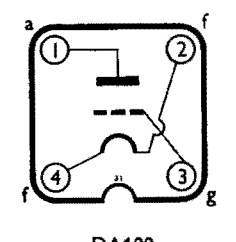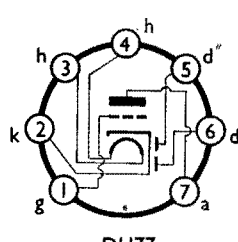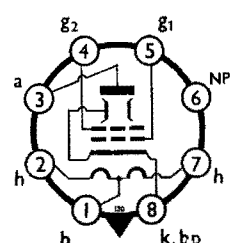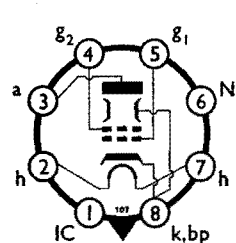
*Only when PIV exceeds 5kV.
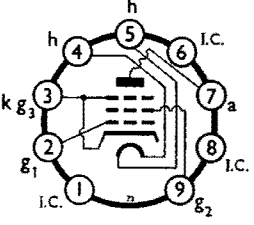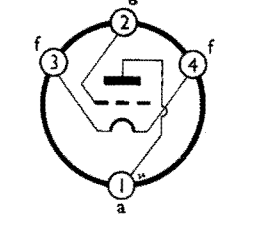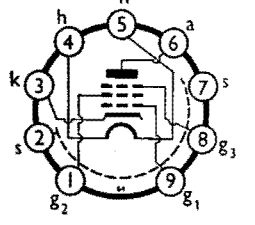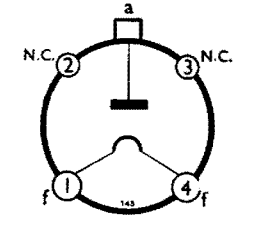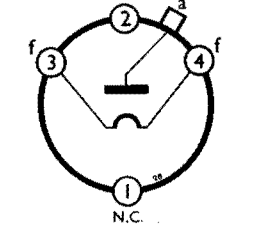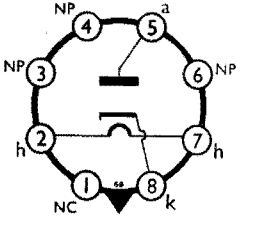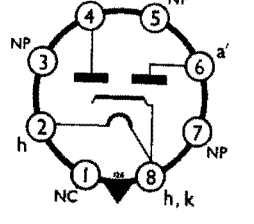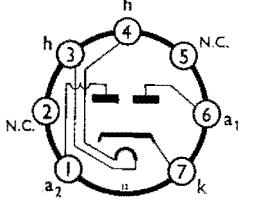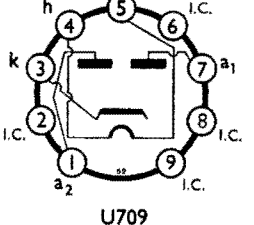
BARRETTERS (Base : E.S. ; Max. length : 123·5 mm ; Max. diameter : 57 mm)

303 : 0·3A, 86-129V.   **304** : 0·3A, 95-165V.   **305** : 0·3A, 40-90V.

B65
Base : Octal

B309, B329, B339
Base : B9A

B719
Base : B9A

DA42
Base : Med. 4-pin Bayonet
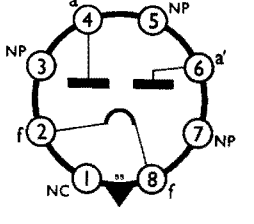
DA100
Base : Special

DH77
Base : B7G

KT33C
Base : Octal

KT55
Base : Octal

KT61, KT66
Base : Octal

KT88
Base : Octal

L63
Base : Octal

LN309
Base : B9A

N709
Base : B9A

V1505
Base : Special

Z729
Base : B9A

GXU1
Base : Med. 4-pin (UX4)

GXU50, U19
Base : B4

U31
Base : Octal

U50, U52
Base : Octal

U54
Base : Octal

U78
Base : B7G

*Viewed from
underside of base*

U709
Base : B9A

B4 :   *British 4-pin*
B7G: *All glass (Button)*
B9A: *All glass (Noval)*

APPENDIX D
ABBREVIATIONS

General Symbols and Abbreviations

A		Ampere.
	mA	Milliampere (Amperes $\div$ 1000).
	μA	Microampere (Amperes $\div$ 1 million).
a.c.		Alternating current.
A.S.		American standard (records).
b.b.m.		Break-before-make.
c/s		Cycles per second.
	Mc/s	Megacycles per second (c/s $\times$ 1 million).
	kc/s	Kilocycles per second (c/s $\times$ 1000).
C		Capacitance or capacitor.
CT		Centre tap.
D		Total harmonic distortion.
db		Decibel.
d.c.		Direct current.
ES		Edison screw (lamp cap).
F		Farad.
	μF	Microfarad (Farad $\div$ 1 million).
	pF	Picofarad ($\mu\mu$F) (Microfarad $\div$ 1 million).
g_c		Conversion conductance.
g_m		Mutual conductance (slope).
H		Henry.
	mH	Millihenry (Henry $\div$ 1000).
	μH	Microhenry (Henry $\div$ 1 million).
h.t.		High tension.
I		Current.
I_a, I_g, etc.		Static current through anode, grid, etc.
$I_{a+g2(o)}$		Combined anode and screen current at zero output.
$I_{a+g2(max\ sig)}$		Combined anode and screen current at maximum output.
I_h		Heater current.
i		Peak current (also instantaneous value of current).
$i_{a(pk)}$, $i_{g(pk)}$, etc.		Peak current through anode, grid, etc.
mA/V		Milliamperes per volt.
m.b.b.		Make-before-break.
mm		Millimetres.
P		Power (external to valve).
P_{out}		Output power.
p		Power dissipated inside valve.
p_a, p_{g2}, etc.		Power dissipated by anode, screen, etc.
p_{dr}		Driving power.
PIV		Peak inverse voltage.
p.u.		Gramophone pick-up.
R		Resistance (external to valve), also resistor.
R_a, R_{g2}, etc.		External load or series resistance for anode, screen, etc.
$R_{L(a-a)}$		Load impedance, anode-to-anode.
R_k		Cathode bias resistance.
r		Resistance inside valve.
r_a		Valve anode resistance ($\delta v_a / \delta i_a$).
r.m.s.		Root mean square.

r.p.m.		Revolutions per minute.
s		Second (time).
	ms	Millisecond (Second $\div$ 1000).
	μs	Microsecond (Second $\div$ 1 million).
s.w.g.		Standard wire gauge.
T		Temperature.
t		Time.
t_{sd}		Switching delay time.
UL		Ultra-linear.
V		Volt.
	kV	Kilovolt (Volt $\times$ 1000).
	mV	Millivolt (Volt $\div$ 1000).
	μV	Microvolt (Volt $\div$ 1 million).
V_a, V_{g2}, etc.		D.C. potential of anode, screen, etc.
$V_{a(b)}$		Anode supply voltage.
V_h		Heater voltage.
V_{in}		Input signal voltage (r.m.s.).
v		Peak voltage (also instantaneous value of voltage).
$v_{in(g-g)(pk)}$		Peak signal voltage between grids of push-pull valves.
W		Watt.
w.w.		Wirewound (resistor).
Z		Impedance (external to valve).
z		Impedance inside valve.
z_{in}		Input impedance.
z_{out}		Output impedance.
μ		Valve amplification factor (also sub-multiple : $\times 10^{-6}$).
Ω		Ohm.
	MΩ	Megohm (Ohm $\times$ 1 million).
	kΩ	Kilohm (Ohm $\times$ 1000).
$<$		Less than.
$\nless$		Not less than.
$>$		Greater than.
$\ngtr$		Not greater than.

Electrode and Other Valve Symbols (Based on B.S.1409 : 1950)

a	Anode.
bp	Beam forming plate(s).
d	Diode.
f	Filament (emitting).
g	Grid.
h	Heater (also hexode, heptode, etc.).
k	Cathode.
p	Pentode.
q	Tetrode.
s	Internal shield.
t	Triode (also target).

Where there are two or more similar systems within a valve (such as in a double triode) the different systems are identified by the use of primes attached to the electrode symbols, thus a' and a", g' and g", etc.

If a single valve system contains more than one electrode of the same type, they are identified thus : g_1, g_2, etc. counting from the cathode.