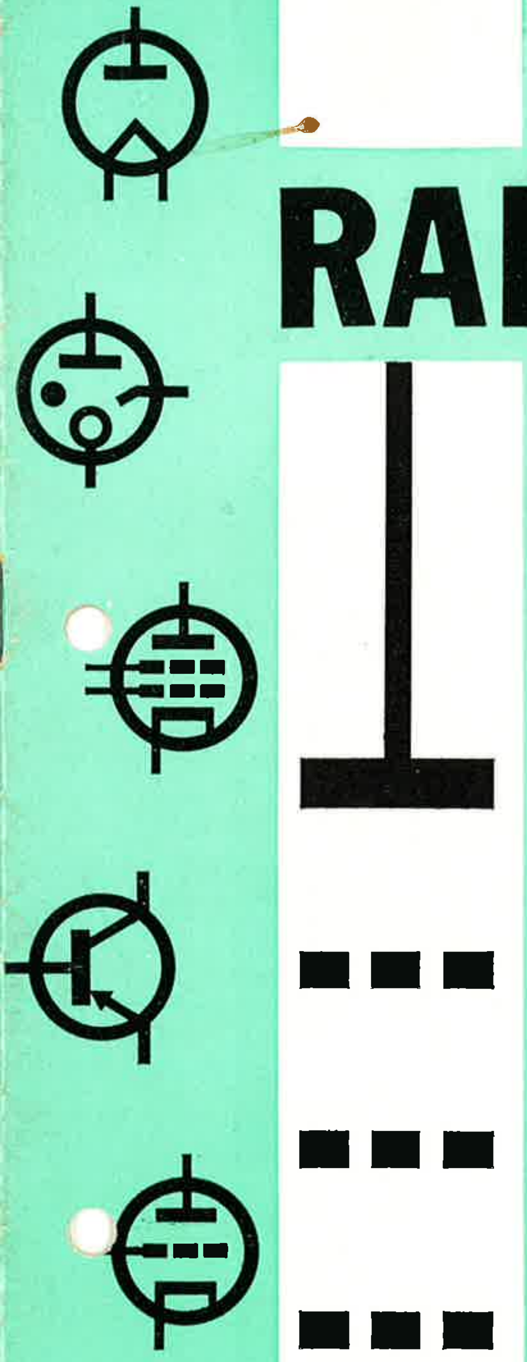
RADIOTRONICS



SPECIAL "TRANSISTORS IN HIFI" ISSUE

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IN THIS ISSUE

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TRANSISTORS IN HIFI

In November of last year, a lecture and demonstration were presented by invitation before the monthly meeting of the Sydney Division of the Institution of Radio Engineers (Australia), and they were intended to show the place of transistors in high fidelity sound reproducing equipment at that date. As will be explained later, the economic use of transistors in this type of equipment is now possible following the introduction of the Power Drift transistor by AWV.

The interest aroused by the lecture and demonstration has prompted us to reprint here the text of the lecture, details of the equipment demonstrated, and other material of interest. The lecture itself was divided into three parts, each delivered by a different speaker.

The attention of the audience was first invited to the Power Drift transistor and its advantages, by Mr. J. Hooke, Manager of the AWV Semiconductor Division. This was followed by a discussion of the practical application problems and some of the ways in which they are being solved, by Mr. H. Wilshire, of the AWV Application Laboratory. Mr. Wilshire also described the design of the experimental units which were on show, and which were demonstrated later in the evening. The lecture programme was concluded with a discussion of transistors in high fidelity along more general lines by Mr. B. Simpson, Editor of "Radiotronics", who subsequently conducted the practical demonstration of an experimental system.

The symposium is presented here essentially as delivered before the meeting, in the form of a complete article. As explained at the meeting the development of the equipment described is continuous. We have therefore brought up to date the circuits of the experimental units, and their performance figures and characteristics, these latter items being already better than those of the units originally demonstrated. Further development is still continuing, so that more may be heard of these units at a later date.

The enormous amount of work to be done in the development of transistorized high fidelity units by the industry in general, and the speed with which some of the answers appear, has persuaded us to this course of releasing data

The three speakers at the meeting reported here. From left to right, Mr. J. Hooke, Manager, AWV Semiconductor Division, Mr. B. Simpson, Editor of "Radiotronics," and Mr. H. Wilshire, AWV Application Laboratory.

prior to the completion of development. This has never in the past been our policy, but it is felt that in this case, there is a useful purpose to be served by so doing, provided it is clearly understood that these circuits are experimental.

As an example in point, we published in May of last year a 12 + 12 watt stereo amplifier, quite large numbers of which have been built and successfully put into service. Towards the end of last year the "Radiotronics" Laboratory completed an improved version of the amplifier, which would also be cheaper to build. Further developments and potential developments have already over-run this unit, with the result that it is now unlikely to be published until the new ideas are incorporated. So this is the way things are going; it is very exciting, but difficult to keep track of.

But more of all that later. The point here is to present to readers data promised at the lecture mentioned above, so here it is:

THE POWER DRIFT TRANSISTOR

By J. Hooke, B.E., B.Sc., Manager, AWV Semiconductor Division

Introduction

The Power Drift Transistor belongs to a new class of transistor in which are combined two well established, but very different, techniques. The technique of the high-power low-frequency transistor is combined with that of the low-power high-frequency drift field device. Just as the incorporation of a drift field in the base region gives many advantages to transistors operating at high frequencies, so it does in high power applications.

The construction differences between the Power Drift and the conventional alloy power transistor are illustrated in Fig. 1, p-n-p devices being shown in each case. The top part of the diagram represents a cross-section through the two transistors, while the corresponding impurity profiles are illustrated below. The donor density in the region of n-type material (the base) is plotted upwards, while the acceptor density in the p-type region (emitter and collector) is plotted downwards.

The alloy transistor is made by alloying an element from group three of the Periodic Table, typically indium, into a relatively lightly doped pellet of n-type germanium. The indium forms the collector and emitter zones, converting them to low resistivity p-type, while the original n-type material forms the base zone. It will be noticed that the base region is relatively wide and that it is uniformly doped.

The Power Drift transistor is made by diffusing a thin n-type base region into p-type germanium. The p-type material forms the collector and an alloyed connection serves to make contact to it. The p-type emitter is alloyed as before. Fig. 1 shows that the transistor made in this way differs from the alloy device as follows:

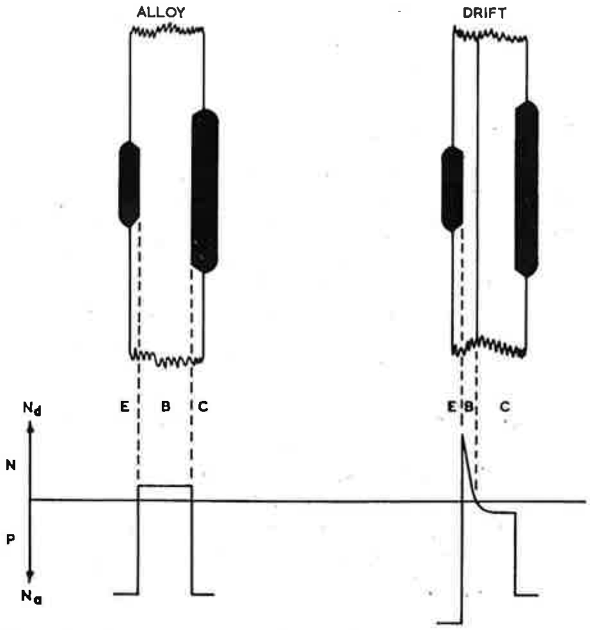

Fig. 1—Basic construction of the conventional alloy power transistor (left) and the power drift transistor (right).

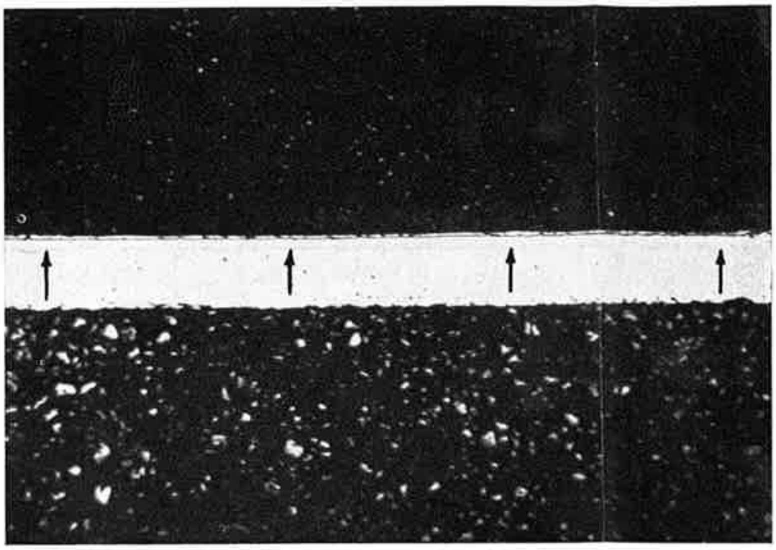

Fig. 2—Photo-micrograph of an actual cross-section of a power drift transistor. At the bottom of the photograph is seen the soldered connection to the collector, which consists of the light area below the faint line designated by arrows. Above that line is seen the base region, and then a very thin light area which is the emitter regrowth region. At the top is the emitter connection.

(i) the base is much thinner,

(ii) the donor density in the base is non-uniform, being high near the emitter junction and low near the collector junction,

(iii) the average donor density in the base is higher; e.g. the average base resistivity is lower,

(iv) the collector-base junction is graded rather than abrupt. Being formed by a diffusion process it is also considerably more uniform than an alloyed junction. This uniformity is brought out in Fig. 2, which is a photo-micrograph of an actual transistor cross section.

Characteristics

To appreciate fully the value of the new construction, it is necessary to consider some of the problems facing the designer of a conventional alloy power transistor. The main parameters which can be adjusted are the width and resistivity of the base region, and the area and geometry of the junctions. Conflicting requirements must be satisfied in adjusting these factors. For example:—

(i) If the device is to amplify high frequencies, the base width must be kept small; but this raises the base spreading resistance, which is undesirable,

(ii) A high collector to base avalanche break down voltage demands a high resistivity base; but for a high punch-through voltage the resistivity must be low,

(iii) For high current gain the base resistivity should be high; but the gain linearity is improved if the resistivity is low.

It is clear that the design must be a compromise, and even by careful juggling it is not possible to optimise for all requirements.

The value of the graded base contained in the Power Drift transistor is that it gives to the designer an additional degree of freedom, and permits many of the conflicting requirements to be met. We now consider these in turn.

Frequency Response

The frequency response can be extended at least tenfold, two factors being responsible. Firstly, the diffused base may be made substantially thinner without incurring other disadvantages, as we shall see later. The transit time of the charge carrier across the base region is proportional to W^2, W being the base width, so the frequency response rises rapidly as W is reduced.

Secondly, the graded impurity profile in the base region gives rise to a built-in electric field in the direction which aids the flow of holes across the base from the emitter to the collector. This field is considerably more effective in moving the charge carriers than is diffusion, a random and slow mechanism upon which the conventional transistor relies. For a given base width, the drift field can reduce the transit time by a factor of five or more, depending on the shape of the base impurity profile.

In a practical design, these two factors together raise the transistion frequency, f_T, from about 0.4 Mc for the alloy transistor to 5 Mc or more for the Power Drift.

The significance of this in high-fidelity amplifier service is considerable. This is apparent from

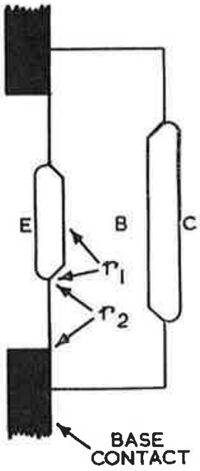

Fig. 3—Cross-sectional diagram of a transistor, showing the two components of the base spreading resistance.

the fact that at a frequency of 10 Kc, the magnitude of the current gain of a typical alloy power transistor falls to half its low-frequency value, accompanied by a phase shift of some 60°. It is phase shift which has so severely handicapped the conventional transistor, because it limits the amount of negative feedback which may safely be applied. This in turn makes very difficult the attainment of a high standard of linearity of gain and frequency.

The Power Drift transistor shows a substantial advantage since changes in magnitude and phase of the gain are negligible over a very wide range of frequencies. It is thus possible to apply a much larger amount of negative feedback and thus achieve high standards of linearity.

Voltage Ratings

Avalanche breakdown is a major limitation to the maximum voltage which may be applied across a p-n junction. In a practical abrupt-junction transistor, the collector-base breakdown voltage, $V_B \propto \sqrt{\rho_b}$, where ρ_b is the resistivity of the base in the vicinity of the collector junction. For high voltage it is thus desirable to set ρ_b as high as possible. This however lowers the voltage at which the collection junction depletion layer reaches across the base to the emitter junction. This punch-through voltage, V_{PT}, forms a second limit to the voltage, the maximum voltage rating of the transistor being V_B or V_{PT}, whichever is less.

It is thus clear that in a conventional alloy transistor, the requirements of V_B and V_{PT} conflict. This conflict is resolved in the Power Drift by employing a non-uniform distribution of impurities in the base layer. The resistivity is made low at the emitter junction, high at the collector junction. This reduces the emitter-base voltage rating, which is normally of little consequence; more important is that the collector-base voltage is raised. Also the punch-through problem disappears, since the depletion layer extends mainly into the collector and only little into the base.

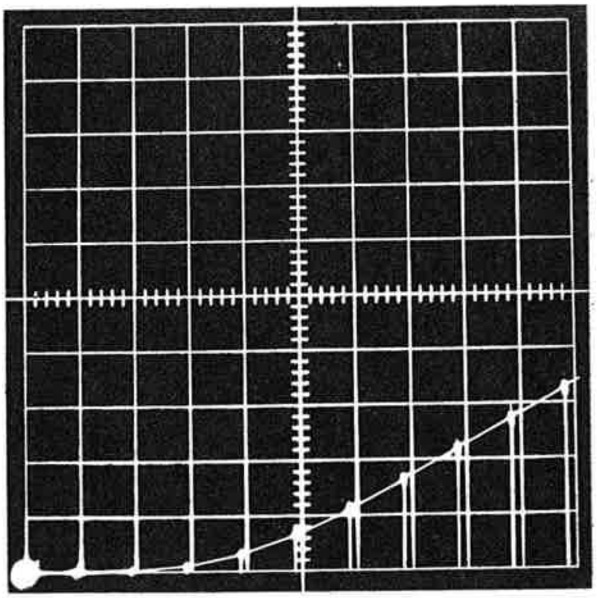

Fig. 4—Mutual conductance of a typical alloy power transistor. Collector current to 5 amperes is plotted vertically, with the base-to-emitter voltage plotted on the horizontal axis to 0.5 volts.

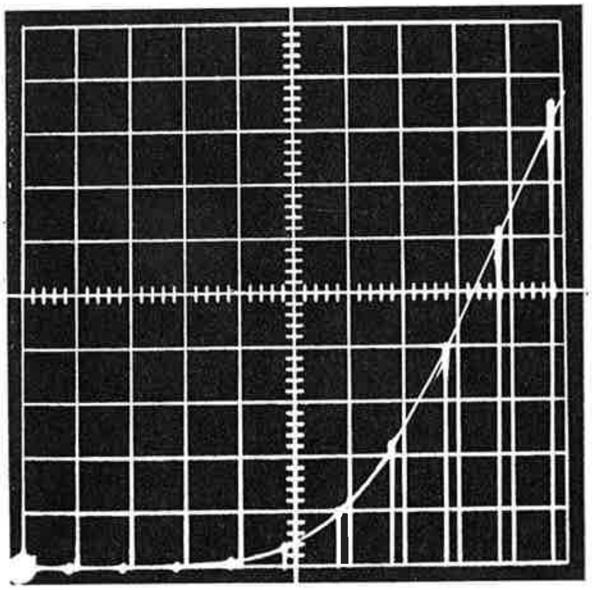

Fig. 5—Mutual conductance of a Power Drift transistor. Collector current to 5 amperes is plotted vertically, with the base-to-emitter voltage plotted on the horizontal axis to 0.5 volts.

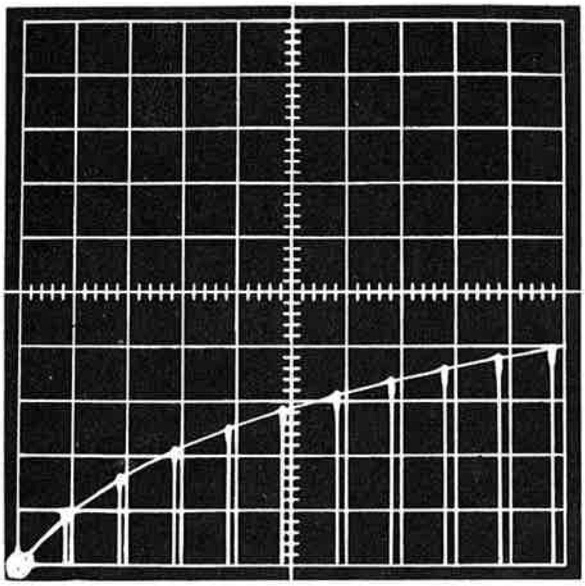

Fig. 6—Current gain of a typical alloy power transistor. Collector current is plotted vertically to 10 amperes, and the base current is plotted horizontally to 50 ma.

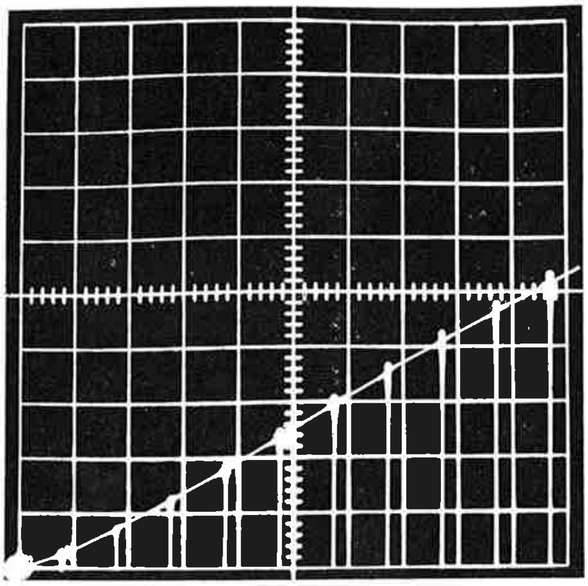

Fig. 7—Current gain of a Power Drift transistor. Collector current is plotted vertically to 10 amperes, and the base current is plotted horizontally to 50 ma.

Base Resistance

The base spreading resistance, which lies between the base contact and the intrinsic transistor, is an undesirable property of all transistors. Two components of base resistance may be distinguished, as in Fig. 3. Component r_1 lies under the emitter, while r_2 lies between the emitter periphery and the base contact. The base resistance $r_{bb}1$ is equal to the sum of r_1 and r_2. In an alloy power transistor, r_1 and r_2 have typical values of 40 ohms and 15 ohms respectively. Component r_1 is substantially independent of collector current, while r_2 falls rapidly as the collector current rises, owing to conductivity modulation of the base.

The lower average base resistivity of the Power Drift transistor is responsible for a reduction in both r_1 and r_2 by a factor of about 15 times. The lower base resistivity also reduces the conductivity modulation considerably, with the result that $r_{bb}1$ is nearly independent of collector current.

The value of such a reduction in base spreading resistance is well known in high frequency applications, since $r_{bb}1$ is a significant frequency limiting parameter. In high power audio applications, $r_{bb}1$ should be minimised, since the driving power dissipated in it is lost.

The improvement made possible by the Power Drift construction is reflected in the high values of g_m of the new device; values as high as 15 amps per volt achieved by the new device, compared with 5 amps per volt for an alloy transistor of similar geometry. Figures 4 and 5 illustrate this change.

Current Gain

Current gain is, of course, the key useful characteristic of the transistor. It is desirable first that the gain should be high, and second that it should remain constant as the emitter current is varied, i.e., that the current transfer characteristic should be linear. The commonly encountered fall in current gain, h_{FE}, at high currents is particularly undesirable in the present context because of the consequent additional negative feedback and increase in driving power which is required.

Two factors cause h_{FE} to fall at high emitter currents. First, the emitter efficiency falls because conductivity modulation of the base takes place at high currents.

We have

$$h_{FE} \propto \gamma = 1 - \frac{R_E}{R_B}\left[1 + \frac{3}{2} \cdot \frac{P_o}{N_d}\right]$$

where γ = emitter efficiency
R_E = sheet resistance of a slice of the emitter of thickness over diffusion length
R_B = base sheet resistivity

P_o = injected carrier density
$\overline{N}_d$ = average base doping density

The injected carrier density, P_o is proportional to emitter current. For a typical alloy power transistor operating at an emitter current of 5 amperes, the term within the brackets amounts to 100, but is less than 3 in the case of a Power Drift transistor of similar size. This improvement comes from the higher value of $\overline{N}_d$ of the Power Drift. It would of course be possible to make a conventional alloy transistor using a similar higher value of $\overline{N}_d$, but without a graded base, the corresponding avalanche breakdown voltage would be only about 10 volts.

At high currents, the transverse voltage drop in the base caused by the flow of I_B in r_1 (component of base resistance) becomes significant. It causes the centre of the emitter junction to be forward biased less than the periphery. The emitter current thus tends to crowd towards the periphery of the junction; this raises the current density and causes the current gain to fall still further. Here again the high base doping of the Power Drift shows to advantage, r_1 being reduced by an order of magnitude as discussed previously. The reduction in transverse voltage drop ensures that the emitter current is distributed over the emitter area far more uniformly.

The effectiveness of these measures in improving the linearity is shown in Figs. 6 and 7. The very substantial improvements in linearity and gain at high currents is apparent. Over the current range 0.5 to 5 amperes, the small-signal current gain typically varies by 20 db for the alloy transistor and by 3 db for the Power Drift.

Conclusion

The graded base has given a new degree of freedom to the design of power transistors, and has enabled many of the conflicting requirements of a transistor design to be reconciled. This new construction will open up new applications for transistors in many fields.

TRANSISTOR HIGH FIDELITY AMPLIFIERS

By H. Wilshire, A.S.T.C., S.M.I.R.E., A.M.I.E. (Aust), AWV Application Laboratory.

Introduction

A high output high performance amplifier has been developed using the new drift power transistors 2N2147 and 2N2148. The amplifier produces 35 watts for less than 0.3% total harmonic distortion with a power response extending from 2 cps to 100 Kc.

The following notes will briefly cover two main subjects, the characteristics of transistors important in the design of high performance audio amplifiers, and a description of the circuit and performance of the pre-driver, driver and output stages of the 35-watt amplifier.

At this point it should be made clear that the amplifier which is discussed here, and which will form one link in the chain of equipment required to produce high-quality sound from recordings, represents one step along a road leading to the faithful reproduction of sound. There are many roads and our work to date has indicated that each of them may lead to a different sort of compromise between conflicting sets of requirements.

High Frequency Phase Shift

One of the limitations of power transistors until recently has been the relatively low cut-off frequency, with the attendant high phase shift and restrictions in the use of feedback. The development of the Power Drift type with a gain-bandwidth product $f_T = 4$ Mc has greatly reduced this problem, and made it possible to design high power amplifiers with flat response extending to near 100 Kc.

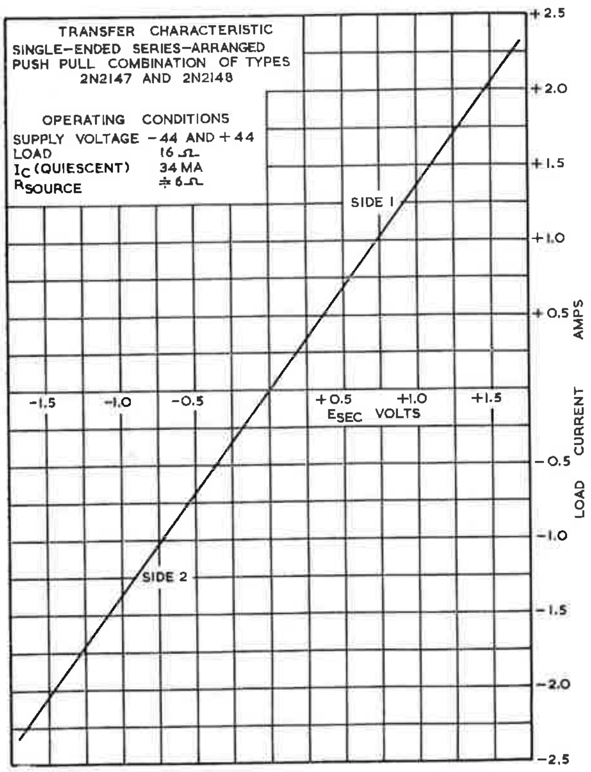

Fig. 8—Transfer characteristic of a set of 2N2147 and 2N2148 Power Drift transistors, operating under the conditions obtaining in the amplifier described herein.

Distortion

The causes of differences other than amplitude between the output and the input signals, normally referred to as distortion, can be grouped under the following three main headings:—

(i) non-linear transfer characteristics for each half of the push-pull stage,
(ii) asymmetry between halves of the push-pull stage,
(iii) cross-over transients caused by hole storage effects.

Of these, the first item (i) produces mainly odd-harmonic distortion and has a number of causes. The major factors are the non-linear input impedance of the transistor and its non-linear transfer characteristic. Each of these is a result of the variation of a number of parameters, e.g., Z_{in} in the common emitter configuration is given approximately by $Z_{in} = r_b + (h_{fe} + 1)\, r_e$, where Z_{in} is the input impedance, r_b is the base resistance of the "T" equivalent circuit, r_e is the emitter resistance of the "T" equivalent circuit, and h_{fe} is the common emitter current gain. Since r_b, h_{fe} and r_e all change to a varying degree with emitter current, the variation of Z_{in} can be very complex. The method of driving the base then must be very important in reducing this form of distortion.

Until recently, for common emitter circuits the use of a constant voltage source (low impedance) has been preferred to the constant current or high impedance source, since the transfer characteristic of collector current versus base-to-emitter voltage has been more linear. One writer[1] has recently drawn attention to the very linear transfer characteristics, particularly in the cross-over region, which can be obtained using a high source impedance drive. However, the amplifier described here uses a relatively low impedance base drive. The transfer characteristic for a set of 2N2147 and 2N2148 transistors operating under the conditions of the amplifier to be described later, is shown in Fig. 8.

The reduction of distortion due to asymmetry between halves of the push-pull stage, mostly even harmonics, is complicated, because although the parameters r_b, h_{fe} f_{hfe} can be matched at one or a few conditions, they each vary with I_E and should be matched over the full range of I_E experienced in the design.

The smaller the variation, the easier it is to match over the whole range of emitter current, and here the 2N2147, with its linear relationship between these characteristics and the emitter current, is of great assistance.

The third, and perhaps unexpected cause of disturbance in a transistor high-power wide-band audio frequency amplifier is the hole storage effect. It is found that as the speed of switching a transistor on and off increases, a point is found where the base and emitter currents will not exactly follow the switching signal. A current will flow in the base circuit after the base-to-emitter voltage has been reversed to switch the transistor "off". This current is in the reverse direction to the normal forward bias current, and is due to the minority carriers (holes in p-n-p type), which are in the base region and which must be removed before the base impedance can become high in the switched-off condition.

The higher the frequency, the more pronounced are these effects, since the magnitude of the hole storage effect is a measure of the switching time of the transistor. The rapid change in base and emitter current at the step, and the increased sharpness of the switch-off as a result of the hole-storage current, can induce ringing in any trans-

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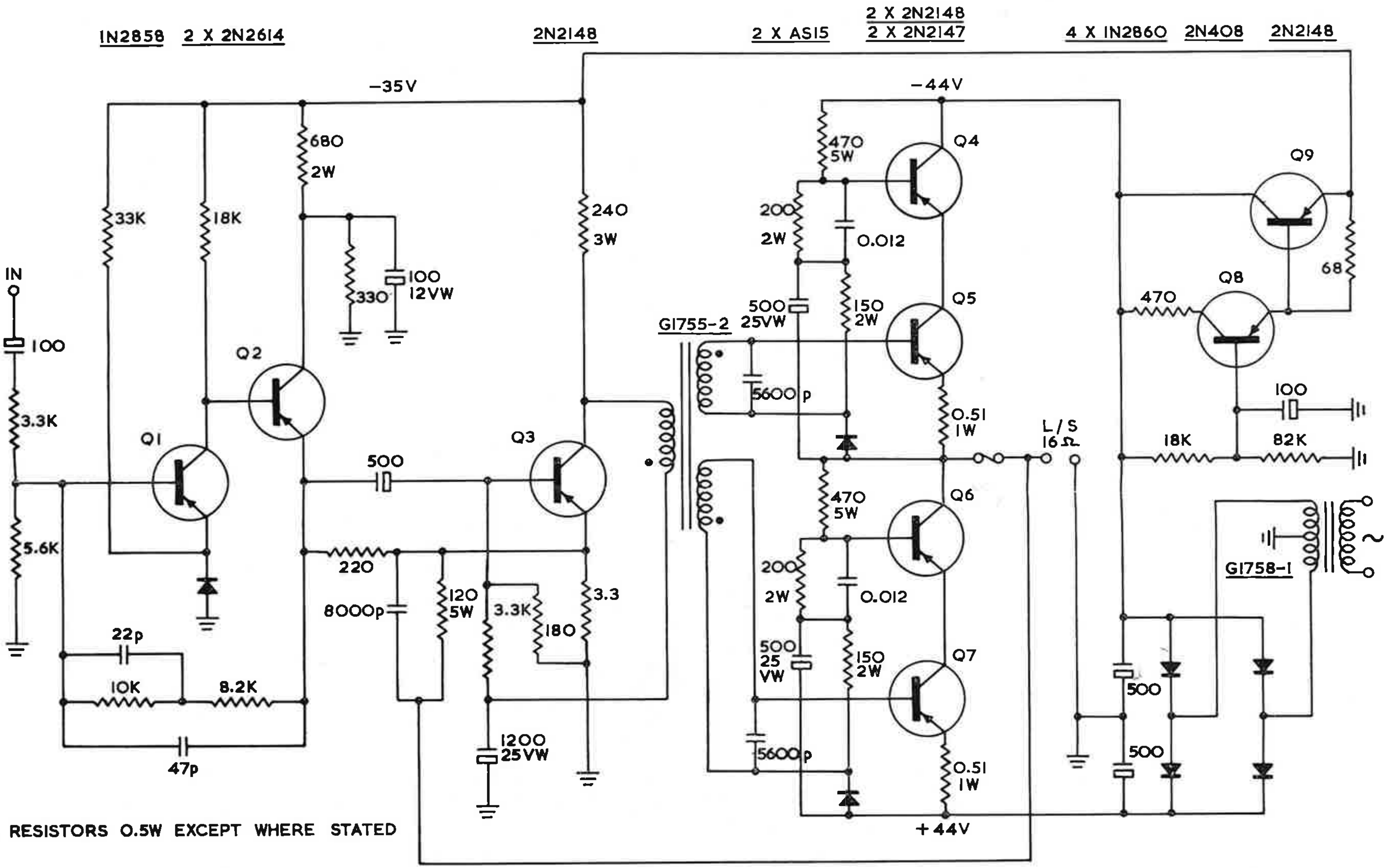
[1] "High Impedance Drive for the Elimination of Cross-over Distortion," I.R.E. Trans. on Audio, July/Aug. 62, 99. J. J. Faran & R. G. Fulks.



RESISTORS 0.5W EXCEPT WHERE STATED

Fig. 9—Circuit diagram of the main amplifier. In view of the experimental nature of this project, two main amplifiers each with its own power supply were used, instead of a two-channel unit with common power supply.

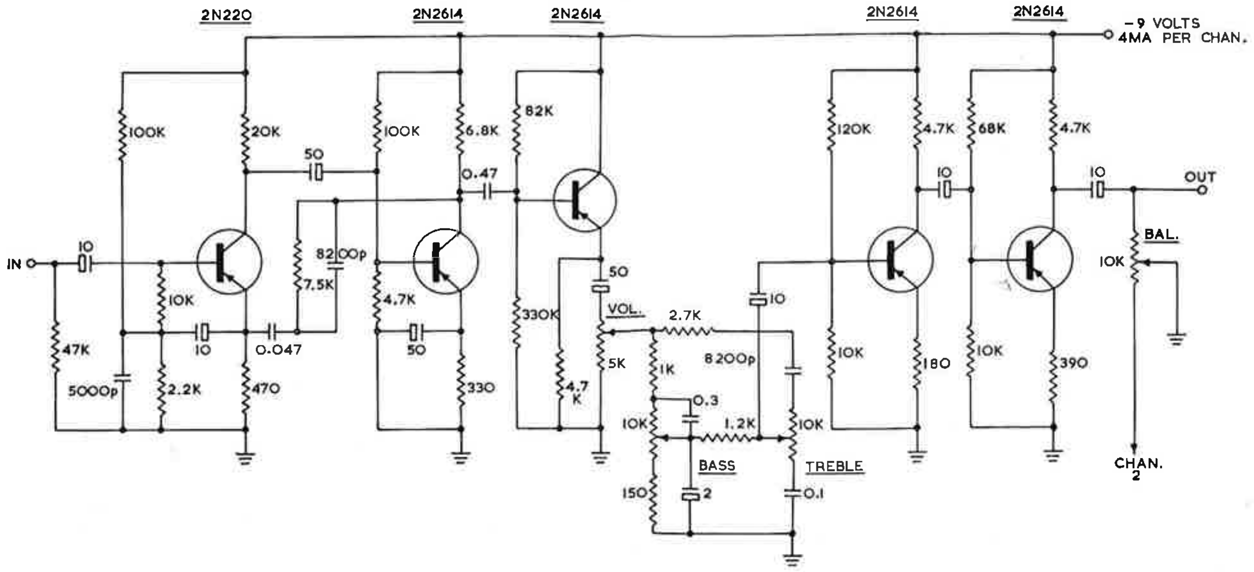


Fig. 10—Circuit diagram of the stereo control preamplifier used with the two main amplifiers. Being built specifically for this purpose, the preamplifier is without all the usual facilities for input selection, filters, etc., the addition of which would be a routine matter.

former in the associated circuit. The use of transistors with a high cut-off frequency and therefore lower hole-storage effects, and the reduction of leakage inductance in both input and output transformers, is required to minimise the disturbance.

It is worth mentioning at this point that when transistors are used in series to increase voltage or power ratings, their switching speeds are most important. As is the case with semiconductor rectifier diodes in series, they can switch with different time delays, and produce a very unequal division of the peak voltage across each. One or more of the following precautions must be taken; (a) ensure similar switching times by selecting transistors and/or the use of an appropriate circuit technique as discussed later, (b) suppress the resultant "spike" or (c) use devices with a breakdown voltage rating adequate for the worst combination of conditions.

Main Amplifier

The amplifier being described has been designed around these new Power Drift transistors, and is at a stage of development where comparison seems worthwhile with equipment using valves to give similar performance.

The main amplifier is based on a single-ended series-arranged push-pull Class B arrangement proposed by two RCA authors.[2] The circuit diagram is shown in Fig. 9.

Output Stage

Each half of the Class B stage consists of two Power Drift transistors in series across a relatively high dc supply voltage. The driven transistors Q5 and Q7 operate in the common emitter configuration with the other two (Q4, Q6) as common base amplifiers. The requirements for h_{FE} matching, linearity, breakdown voltage and saturation current are therefore much less stringent for half of the output transistors.

Other advantages of this arrangement include high attenuation of ripple. In normal push-pull circuitry, due to the impossibility of providing exact balance, an added hum ripple appears on the supply line. This can be fed back via the bias network to the input, and result in a hum output signal. In this arrangement the high output

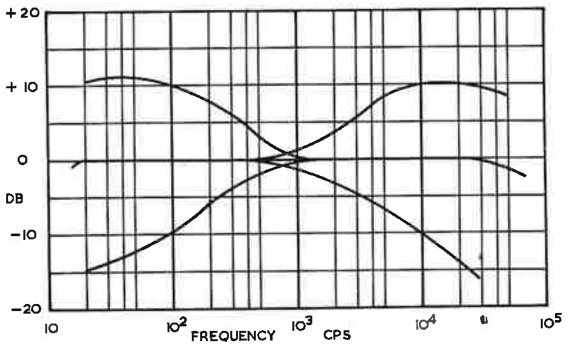


Fig. 11—Frequency response of the preamplifier unit, showing the operation of the bass and treble tone controls.

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[2] "An Ultra Low Distortion Transistorized Power Amplifier," I.R.E. Trans. BC and TV Receivers, July, 1961, H. M. Kleinman and C. F. Wheatley.

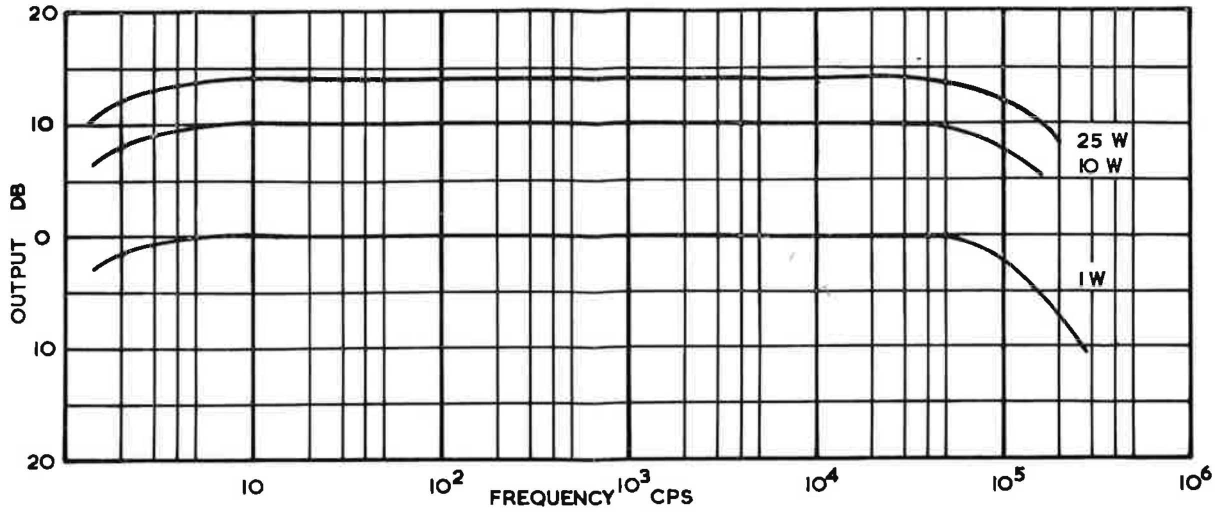


Fig. 12—Frequency response of the main amplifier, taken at output power levels of 1, 10 and 25 watts.

impedance of the common base transistors Q4, Q6 severely attenuates the ripple from the supply to the bias network. The high impedance in the emitter circuits of these transistors by negative feedback virtually eliminates ripple on the base of Q4, Q6. The 500 microfarad capacitors remove ripple from the base of Q5, Q7. The only remaining source of hum is due to the out-of-balance hum currents in the bias networks. As a result the power supply filtering is uncritical. A further advantage is that due to the two junctions in series, a high-supply-voltage low-current supply can be used.

Germanium diodes are used in the bias network for each of the common emitter transistors, partly to provide bias compensation for high ambient temperature, and partly to maintain the stability of the operating point with signal level and thus produce a better transient response.

In amplifiers of this power rating it is possible under certain circumstances for high peak voltages to be developed across the output transistor. The conditions causing the development of these high voltage "spikes" are the application to the amplifier of high level signals of low or high frequency, either momentarily or continuously. The momentary application can be caused by a noisy gain control, the inadvertent connection of the pick-up to the preamplifier with the gain control at the maximum gain position, or the connection of a measuring instrument to one of the preamplifier stages. In each case the signal level can be sufficient to drive the output stage to produce up to a 70-watt output which will be of square wave form. This will be true for signals of any frequency within the range of the system, and will produce a square wave output with a rise

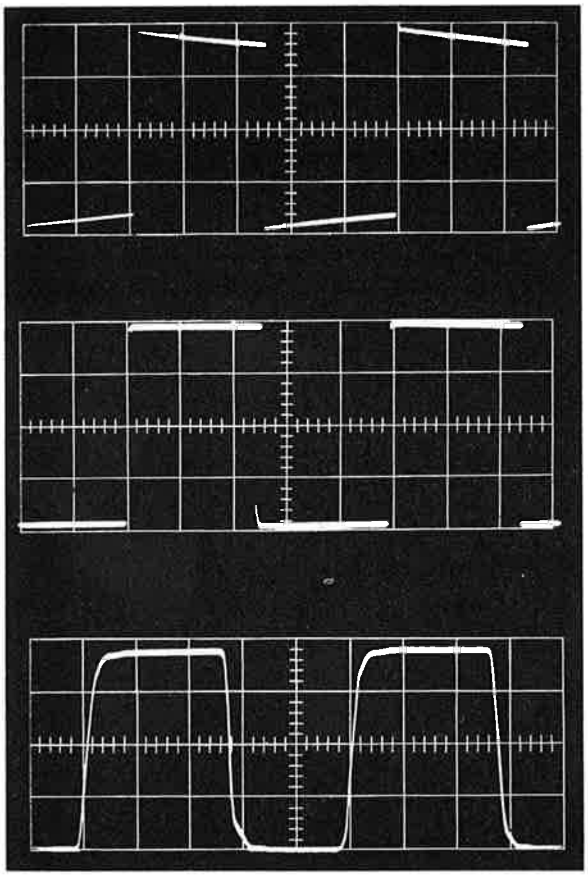


Fig. 13—CRO photographs of the square wave response of the main amplifier at frequencies of 100 and 1000 cps for a power output of 25 watts, and at a frequency of 20 Kc for a power output of 10 watts.

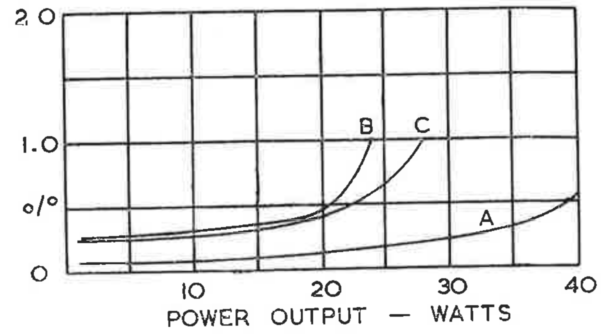


Fig. 14—Total harmonic distortion and intermodulation distortion of the main amplifier versus power output. Curve A shows the THD. Curve B shows the IM distortion with the standard amplifier as shown in the accompanying circuit diagram, whilst curve C shows IM distortion with the extra filter capacitor mentioned in the text.

time of approximately 4 microseconds. Under these conditions the relative switching times of the two series-connected transistors in each half of the amplifier are important.

In this amplifier the development of high peak voltages across any one of the output transistors is prevented by the connection of 12,000 picofarad capacitors across each of the base resistors of the common base transistors. These capacitors serve to "speed up" the application of the cutoff signal to the common base stages. A variation in the value of the "speed-up" capacitors will allow the time at which the common base transistors (Q4, Q6) cut off to be adjusted to occur before or after their respective common emitter drivers. If either of the capacitors is omitted, Q4 or Q6 will switch later than Q5 or Q7 respectively, with the result that a very high voltage "spike" can appear between the collector and emitter of Q4 and Q6. The conditions giving rise to this situation were mentioned previously.

Driver and Predriver

A Power Drift type is used as the driver, since power with low distortion at high frequencies is required. The coupling transformer is RC coupled to the driver, and uses grain-oriented silicon steel. The secondaries are bifilar wound between the halves of the primary. Negative dc feedback to the base of the driver provides very good stability for the operating point.

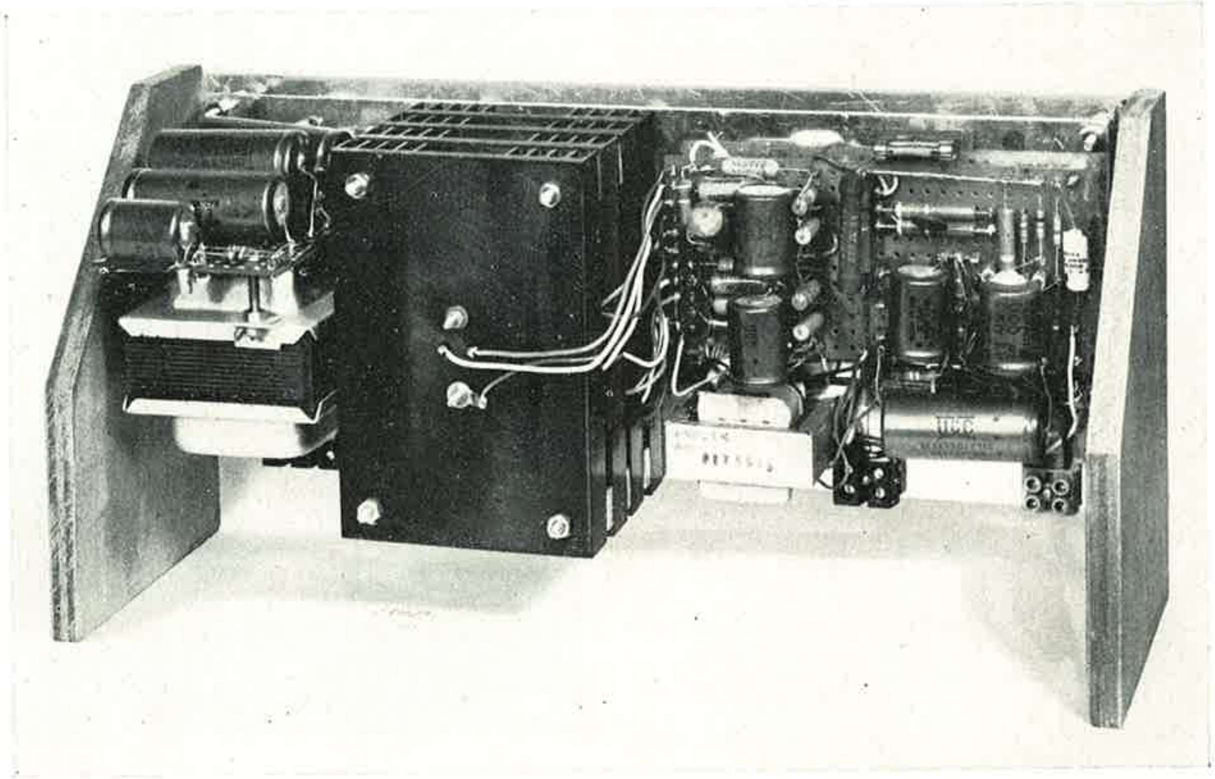


Fig. 15—A view of one of the two experimental 35 watt main amplifier units described herein. It should be noted that the method of construction and arrangement of the experimental units was dictated by convenience and ease of access rather than a conviction that the layout used was necessarily the best.

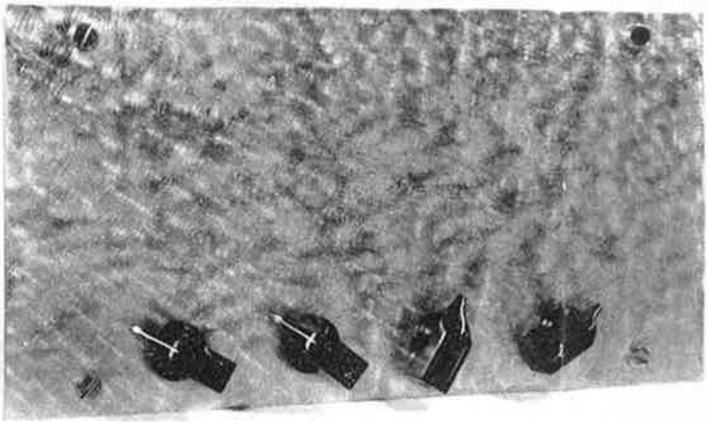


Fig. 16—View of the stereo control preamplifier. As mentioned in the text, this unit was constructed in "skeleton" form without switching and other control facilities.

Two direct-coupled stages are used in the predriver, with the new high-performance high-frequency low-level audio-frequency transistor 2N2613. Q2 operates as an emitter follower and provides a low impedance source for the relatively low input impedance of the driver. The silicon diode in the emitter circuit of Q1 provides very good stability for the operating bias of the two predrivers, and thus maintains low distortion.

Negative Feedback

Negative feedback totalling approximately 39 db is used around the driver and output stage. At 1000 cps, the voltage feedback between the load and the driver emitter provides approximately 20 db, while the driver emitter resistor accounts for approximately 18 db. At 20 cps, the combined feedback is 27 db, while at 100 Kc it is 15 db. The loop around the predriver from the emitter of the emitter follower to the base of Q1 provides at 1000 cps approximately 21 db of negative feedback.

Matching of Semiconductors

Whilst it must be emphasised that a high standard of performance is obtained from the amplifier when no precautions are taken to match the semiconductors used in the output stage, it has also been found that to optimise performance, particularly in the area of intermodulation distortion, some degree of matching is desirable.

The two type 2N2147 transistors used in the common emitter mode should be matched within 4 db at a collector current of 1 ampere for the best results. Matching of the two type 2N2148 transistors, used in the common base mode, is not required. It must be pointed out here that successful operation of high power transistors, such as those used in the output stage of this amplifier, relies upon correct fitting of the transistor on the heat sink, and silicone grease should always be used to reduce the thermal resistance between the transistor and the heat sink.

The two compensating diodes type AS15 have a bearing on the intermodulation distortion and the question of thermal stability. For the best results, these diodes should be matched in pairs for use in this amplifier, and should be so ordered. Further, flag-type heat sinks should also be ordered and used with them, and the heat sinks should be bolted to one of the heat sinks of the output stage; this latter requirement ensures the correct functioning of the diode as it holds the diode at essentially the same temperature as the output transistors.

Power Supply

Two full-wave circuits are used to produce separate negative and positive 44-volt supply voltages, which are each filtered by a single 500 microfarad capacitor. These two supplies are used for the output stage, while dc power for the driver and predriver stages is obtained from the —44 volt point through a "dynamic" filter using a 2N2148 and a 2N408. This arrangement provides a —35 volt supply with a ripple of about 4 millivolts, which represents about 60 db attenuation from the —44 volt point. The arrangement also introduces a time constant of approximately 2 seconds in the rate of rise of the —35 volt point when switched on. A 50-70 watt output "plop" on "switch-on" is thereby avoided.

Preamplifier

The circuit of the preamplifier is shown in Fig. 10. The unit was specifically prepared for use in the demonstration, where the only input requirement was that of a magnetic pickup. The circuit is therefore without all the input and mode switching facilities generally associated with such a unit, although the addition of the usual facilities would be only a routine matter.

The circuit of the preamplifier is conventional, and requires no special explanation. It provides an output of 300 millivolts for an input of 2.5 millivolts, and the first two stages incorporate equalization for the RIAA recording characteristic. Bass and treble tone controls and a balance control are included. The frequency response of the preamplifier and the effect of the tone controls are shown in Fig. 11. Total harmonic distortion in the preamplifier is less than 0.2%, and the noise figure is better than —60 db.

For the purposes of the demonstration, the collector supply for the preamplifier was derived from the filtered —35 volt supply in one of the main amplifiers, through a series resistor and decoupling capacitor.

Main Amplifier Performance

An appendix lists the main characteristics of the amplifier, whilst Fig. 12 shows the frequency response of the main amplifier for output powers of 1, 10 and 25 watts. In Fig. 13 are seen photographs of the face of an oscilloscope depicting the square wave response at 100 and 1000 cps for an output of 25 watts, and at 20 Kc for an output of 10 watts. Fig. 14 shows the total harmonic distortion and intermodulation distortion versus power output. Intermodulation distortion has been measured using 60 cps and 2000 cps signals mixed in the ratio of 4 : 1, and is expressed as a percentage of the lower (2000 cps) signal.

Two curves of intermodulation distortion are shown. One shows the effect of reducing the ripple on each of the 44 volt supply lines to the output stage, by the addition of an extra 4000 microfarad filter capacitor across the supply. When a signal consisting of two components, one of low and one of high frequency with a 4 : 1 difference in amplitude, is fed to the input of the amplifier, as is the case for the intermodulation measurement, the major part of the measured power output is due to the high-level low-frequency component. As the power output is increased under these conditions, clipping by the output stage will commence at a lower power output than is measured using a single frequency sine wave signal. At its commencement the clipping causes a very great increase in intermodulation distortion.

The presence of hum ripple in the supply voltage for the output stage means, of course, that the effective supply voltage varies at a 100 cps rate, by an amount equal to the ripple voltage. At high output powers, this causes 100 cps clipping of the mixed signal and causes a high inter-

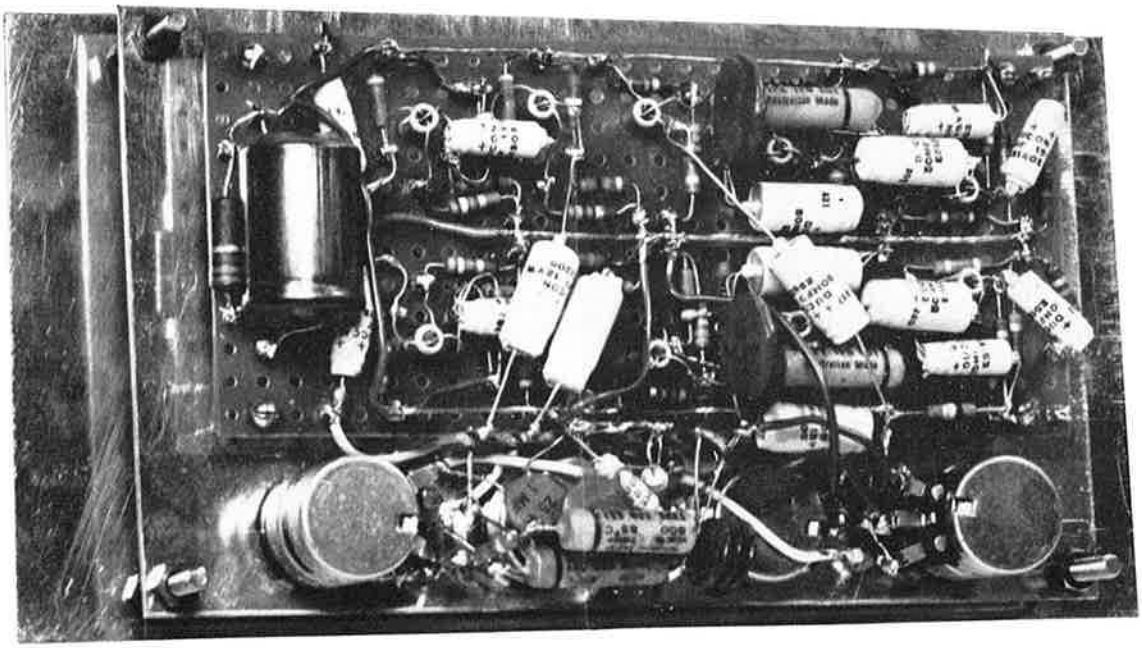


Fig. 17—A further view of the stereo control preamplifier, emphasising the small physical size possible with transistorized units.

modulation distortion. Curve C of Fig. 14 shows the reduction in distortion at high power obtained by adding a 4000 microfarad capacitor across the total supply voltage.

APPENDIX

MAIN AMPLIFIER PERFORMANCE SPECIFICATION

Nominal Power Output: 35 watts.

Total Harmonic Distortion: 0.16% at 25 watts, 0.3% at 35 watts, measured at 900 cps. (See diagrams.)

Intermodulation Distortion: 1.8% at 25 watts, with standard power supply; 0.78% at 25 watts, using additional filter capacitor. (See text and diagrams.)

Phase Shift: 3°.15′ at 20 cps, 4°.42′ at 20 Kc.

Frequency Response: 2 cps to 110 Kc +0 —3 db at 1 watt level, 10 cps to 35 Kc ±0.5 db at 25 watts level. (See diagrams.)

Hum and Noise: —73 db relative to 25 watts output.

Sensitivity: 120 mv rms at 900 cps for 25 watts.

Output Impedance: 1.1 ohm.

Damping Factor: 14.5.

TRANSISTORS IN HIFI

By B. J. Simpson, Editor, "Radiotronics."

The introduction of the transistor something over ten years ago has revolutionised large sections of the electronics industry. During this time, there has inevitably been a great deal of comparison between these new devices and the more-familiar electron valve. Industry has now settled down in the situation where there is a choice of two basic types of active device, each having its own respective merits and advantages.

It is not proposed at this time to offer a direct comparison of electron valves and transistors, or to advocate the use of one against the other. Such factors must be decided by the user in the light of the circumstances surrounding his design. It may however be appropriate at this time to enumerate the outstanding areas in which the use of semiconductors has attractions for the designer, and to clear up some of the misconceptions on these points.

The advantages offered by semiconductors are in general well known, but may be restated here for the sake of completeness. They are:

(i) Small size, and weight which, when taken in conjunction with other factors, offer savings in manufacturing costs, and open the way to styling advantages.

(ii) The absence of the heater requirement, and the often lower power drain of a transistorized stage versus the cathode power drain of an electron valve, lead to reduce chassis dissipation, supporting a reduction in unit size.

(iii) Virtual absence of microphony, and high resistance to shock and acceleration, leading to greater reliability.

(iv) Reduction in auxiliary hardware, such as valve sockets, and an increase in suitability for printed circuit techniques. There is a potential saving here both in cost of components and in assembly costs.

These advantages just enumerated apply in a general way to all cases. There are however, further advantages, either actual or potential, to be gained from the use of semiconductors in audio amplifiers in particular. Foremost among these are the following:

(i) Use of the single-ended series-arranged push-pull circuit allows us to set up a power amplifier stage which needs no output transformer, and which can be used with loudspeakers of conventional impedances.

(ii) It is much easier with semiconductors to set up circuits with direct coupling. This not only improves the low frequency performance of the unit, but also allows the use of higher levels of negative feedback, leading to an improvement in overall performance.

Until the recent introduction of the Power Drift transistor, the application of transistors to high performance amplifiers was not successful, except at a cost that was uneconomic, and/or by the use of silicon transistors. The greatly increased linearity of the Power Drift transistor has opened this field up, and it is rapidly being exploited, particularly in USA.

At the same time, several misconceptions have arisen about these applications, both as regards performance and cost. Performance of course, speaks for itself, and it is in the region of cost that some reconsideration may be required. Unfortunately at this stage, there is not available the large fund of cost data that we have available on units using electron valves, so that a direct comparison cannot be made. Any comparison is made more difficult by the fact that amplifier specifications vary so much, and who can decide the weighting factor to be placed against the various figures quoted?

The experience of readers, however, will allow those who examine the circuits and specifications of the units we are describing, to form some opinion of the cost of the units, and to compare them with units of similar performance using electron valves. In "Radiotronics" of May 1962, there was featured a further unit with an output of 12 watts per channel, which could also be assessed in this way. I feel that, even in this early stage of the development of transistorized high fidelity amplifiers, costs stand comparison.

Further, these units are admittedly in the early stages of development. Development is proceeding on these and other units, and there is little doubt that there is opportunity for further notable improvements. It is obvious that the situation with respect to transistorized equipment will improve, both in regards to performance and cost.

It is important to remember always the basic requirements for high quality reproduction. As a rule, the list is headed by harmonic distortion, although we know that intermodulation distortion is far more important, and is more directly related to subjective evaluation of a unit.

Frequency response must be as flat as possible, and should extend at low power levels at least two octaves on either side of the accepted audio spectrum of 15 cps to 20 Kc. The development of high output powers at the higher frequencies is not required, and may be undesirable.

Output impedance should be low and the damping factor high, to quote perhaps stylised conditions. There is great variation of opinion on this matter, and it seems clear that with the sophisticated speaker enclosures we are using today, which have high acoustic damping, the need for high electro-magnetic damping is less desirable.

Noise is a very important factor, and transistorized equipment often has a decided advantage here, particularly where the source impedance is below about 5000 ohms. The use of balanced and hum-resistant output stage configurations, and the use of dynamic filters for the earlier stages, virtually removes hum from many transistor units, leaving only white noise. It is a fact that higher levels of white noise can be tolerated than noise with a large content of one or more discrete frequencies. The low hum situation is also helped by the absence of heaters and wiring, although with very low noise valves available, this may not apply.

Summarising, it is clear that the door to transistorized high fidelity equipment has now been opened, and that transistorized equipment is going to be with us in ever larger quantities.

Development over the next year or so will bring us some fine equipment of very high performance alongside and in competition with valve equipment. The surprising thing about all this is that in the short time suitable transistors have been with us, they have made such a large advance against the entrenched position of the electron valve, with years of development experience behind it. If the promise inherent in this situation is fulfilled, then the future looks bright indeed for both the audiophile and the industry as a whole.

APPENDIX

Several questions have been received concerning the equipment, other than the AWV amplifiers, used in the demonstration. A complete list is therefore given here below. It must be stressed that the choice of equipment, whilst obviously guided by the desire to use only units of very high quality, should not be interpreted necessarily as an exclusive endorsement of the units, as there are doubtless other units of equipment of comparable quality which could have been used instead.

AWV experimental 35 watt main amplifier. Two units were used.

AWV experimental stereo control preamplifier.

"Thorens" TD124 turntable unit.

"Thorens" BTD12S pickup arm.

"Shure" M337 cartridge. The 7-mil radius unit was used as some of the records we wished to play have already been played on various other installations.

"Leak" "sandwich" loudspeaker system, consisting of "sandwich" woofer together with a tweeter and cross-over arrangements. Two such units were used for stereo.

GERMANIUM P-N-P DRIFT-FIELD TRANSISTORS 2N2147, 2N2148

The 2N2147 and 2N2148 are diffused-collector, graded-base power transistors of the germanium p-n-p type intended for use in high fidelity audio amplifiers and other units where wide frequency range and low distortion are required. These transistors utilize a combination of alloying and diffusion techniques to provide a built-in accelerating field in the base region. This field makes possible a wide frequency response and a linearity of characteristics not available in conventional power transistors.

The 2N2147 is intended for use in class B amplifier service, and has specially controlled breakdown voltage and collector saturation current characteristics to assure dependable performance in this type of service. The 2N2148 is intended for use as a class A power amplifier in driver stages of high-power high fidelity audio amplifiers, and in the output stages of moderate-power audio amplifiers. It can also be used in class B power amplifier service.

The average dissipation for both types is 12.5 watts at a mounting flange temperature of 81° C. These types have a high dc beta, typical figures being 150 for the type 2N2147 and 90 for the type 2N2148, whilst the gain-bandwidth products are typically 4 Mc and 3 Mc respectively. These features, together with the low base resistance and linear gain characteristics, make these types eminently suitable for the applications mentioned.

MAXIMUM RATINGS, AF POWER AMPLIFIER SERVICE

Absolute Maximum Values

	2N2147	2N2148	
Collector-to-base voltage	−75	−60	volts
Collector-to-emitter voltage	−50	−40	volts
Emitter-to-base voltage*	−1.5	−1	volts
Collector current	−5	−5	amp
Base current	−1	−1	amp
Transistor dissipation:			
Average, for mounting flange temperatures up to 81° C†	12.5	12.5	watts
Peak, determined by operating conditions, see Figs. 1 and 2.			
Temperature range:			
Storage	−65 to + 100° C		
Operating (junction)	−65 to + 100° C		

* This rating may be exceeded provided the combined dissipation in the emitter and collector does not exceed the maximum dissipation rating for the transistor.

† Above 81° C, derate linearly at 0.66 watt per degree Centigrade.

CHARACTERISTICS RANGE VALUES FOR EQUIPMENT DESIGN

At a mounting-flange temperature of 25° C

Characteristics: Common-Emitter Circuit, Base Input Unless Otherwise Specified	Symbols	TEST CONDITIONS						LIMITS						Units
		DC Collector Voltage (volts)		DC Emitter Voltage (volts)	DC Collector Current (ma)	DC Base Current (ma)	DC Emitter Current (ma)	Type 2N2147			Type 2N2148			
		V_{CB}	V_{CE}	V_{EB}	I_C	I_B	I_E	Min.	Typical	Max.	Min.	Typical	Max.	
Collector-to-Base Breakdown Voltage	BV_{CBO}				-10		0	-75	-	-	-60	-	-	volts
Collector-to-Emitter Breakdown Voltage	BV_{CEO}				-100	0		-50	-	-	-40	-	-	volts
Emitter-to-Base Breakdown Voltage	BV_{EBO}				0		-2.5	-1.5	-	-	-1	-	-	volts
Collector Cutoff Current	I_{CBO}	-40					0	-	-	-1	-	-	-1	ma
Collector Cutoff Saturation Current	I_{CBO} (sat)	-0.5					0	-	-	-70	-	-	-100	μa
Emitter Cutoff Current	I_{EBO}			-1.5 -1	0 0			- -	- -	-2.5 -	- -	- -	- -2.5	ma ma
DC Forward-Current-Transfer Ratio	h_{FE}		-2		-1000			100	150	-	40	80	-	
Gain-Bandwidth Product	f_T		-5		-500			-	4	-	-	3	-	Mc
Thermal Resistance (Junction-to-Case)	θ_{JC}							-	-	1.5	-	-	1.5	°C/watt
Base-to-Emitter Voltage	V_{BE}		-10		-50			-	-0.24	-	-	-0.26	-	volt

OPERATING CONSIDERATIONS

Attention is drawn to the fact that the maximum ratings quoted for these transistors are established under the absolute-maximum rating system. They are limiting values, and must not be exceeded.

In the design of amplifier circuits using the 2N2147 and 2N2148, it is extremely important to assure that the operating characteristic for either type does not, under any foreseeable combination of operating conditions, extend outside the region of reliable operation shown in Figs. 1 and 2. Even momentary excursion of the operating characteristic outside this region, or momentary operation of the transistor above any of its maximum ratings can result in permanent damage to the device.

To assure that the 2N2147 and 2N2148 are operated at all times within the regions shown in Figs. 1 and 2, the circuit designer should take into account the possible effects of the following factors:

(i) phase shift due to the circuit capacitances and/or loudspeaker resonance,

(ii) parasitic oscillations, such as "ringing" caused by excessive or improperly neutralized feedback,

(iii) high line voltage,

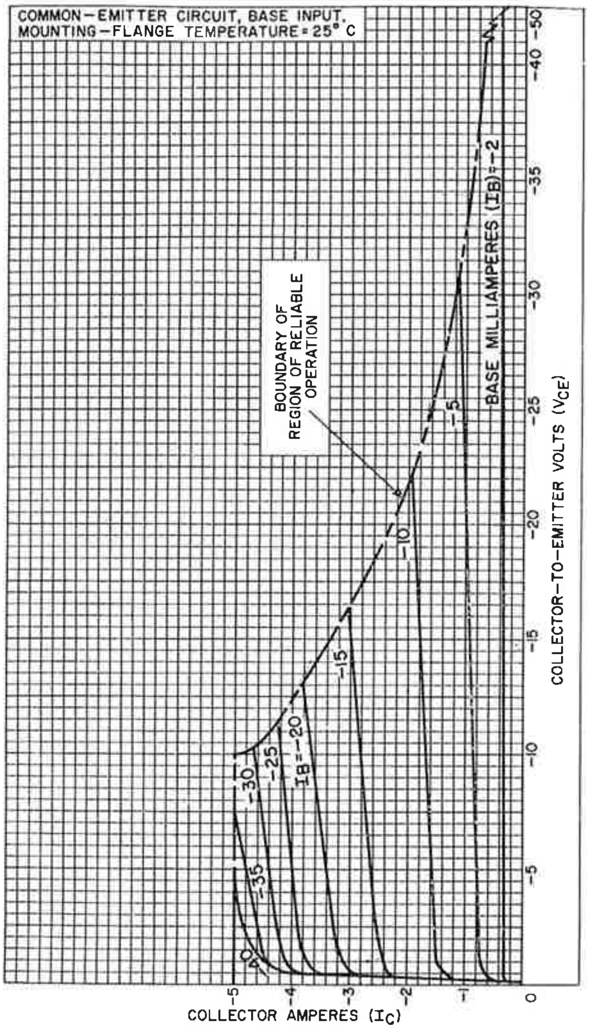
(iv) variations in loudspeaker impedance,

(v) overdriving of transistors,

(vi) non-sinusoidal signal waveforms.

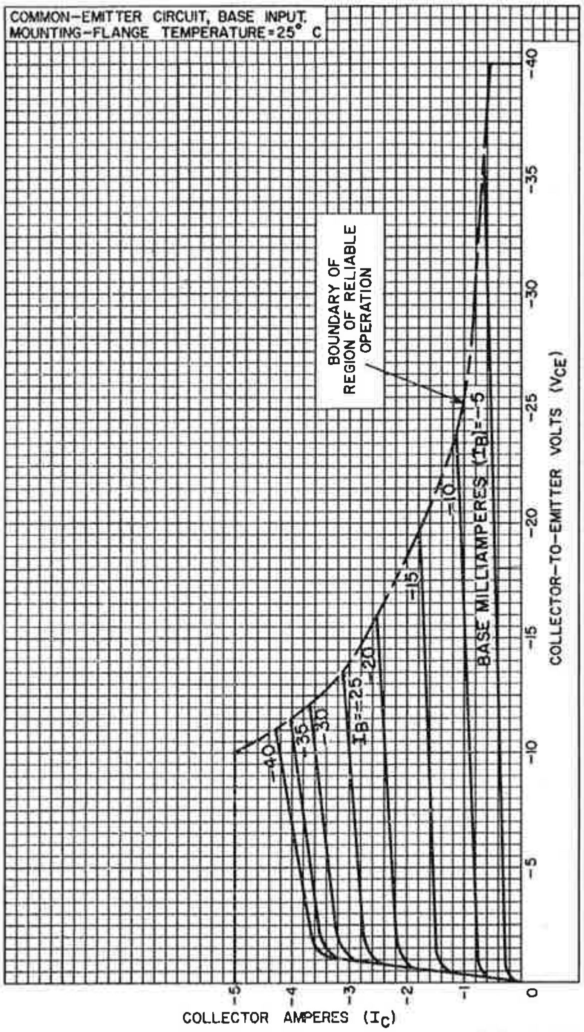
Any one of these factors, or a combination of them, can change the character and value of the transistor load sufficiently to cause operation outside the region of reliability shown in Figs. 1 and 2.

The methods of connecting the transistor into the circuit, the provision of heat sinks, and the reduction of thermal resistance between the heat sink and the transistor, should follow the procedures for power transistors already published in these pages.



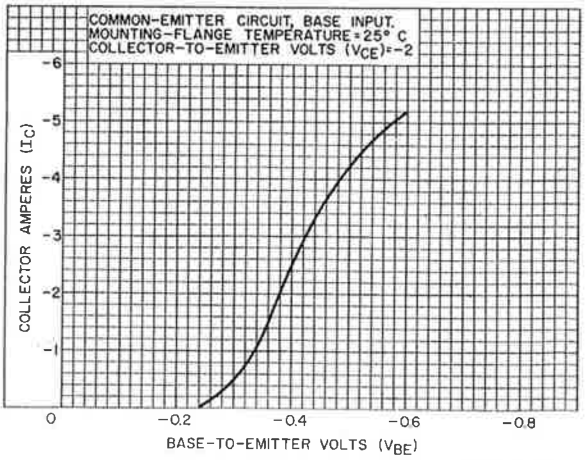
92CM-11331

Fig. 1—Typical collector characteristics for type 2N2147



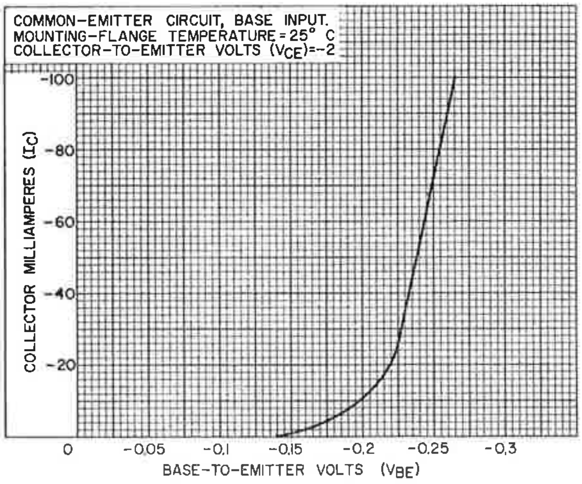
92CM-10999R2

Fig. 2—Typical collector characteristics for type 2N2148



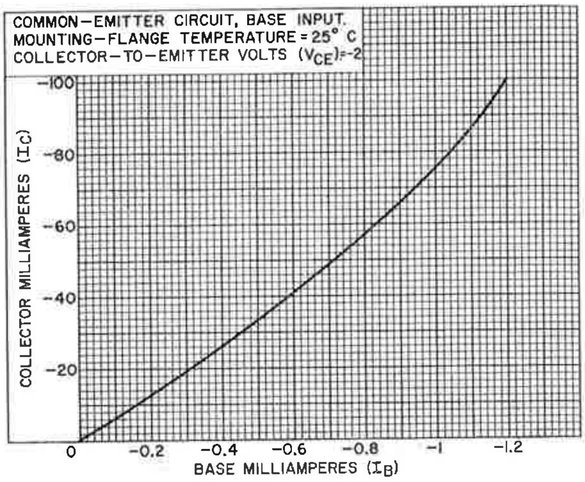
92CS-11324

Fig. 3—Typical characteristics for type 2N2147



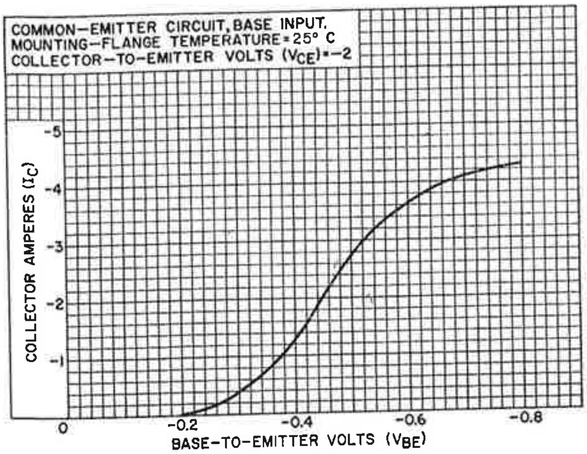
92CS-11326

Fig. 4—Typical characteristics for type 2N2148



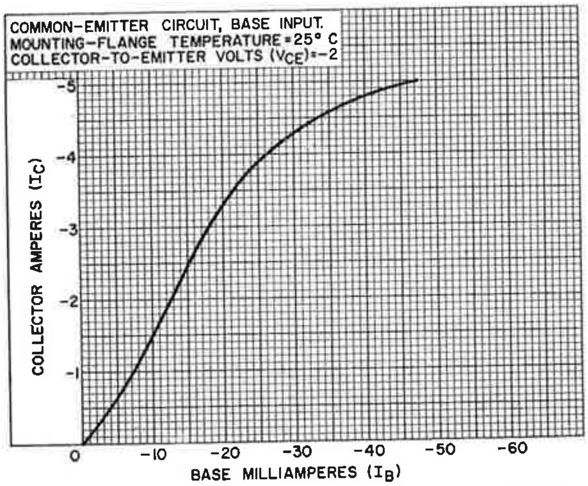
92CS-11834

Fig. 5—Typical characteristic for type 2N2147



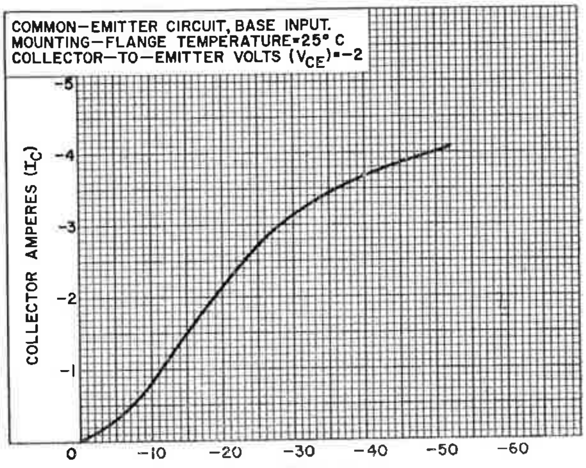
92CS-10995

Fig. 6—Typical characteristic for type 2N2148



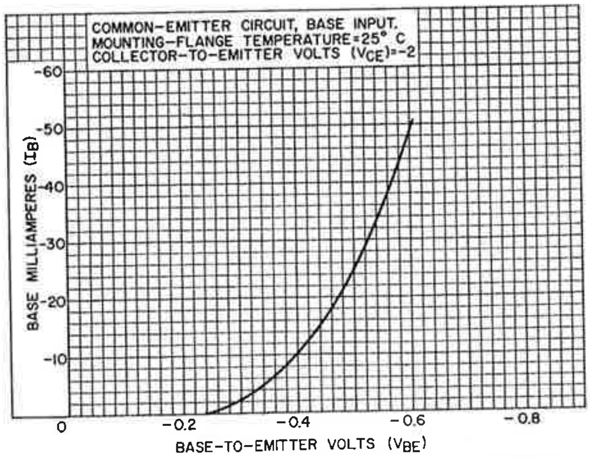
92CS-11325

Fig. 7—Typical current transfer characteristic for type 2N2147



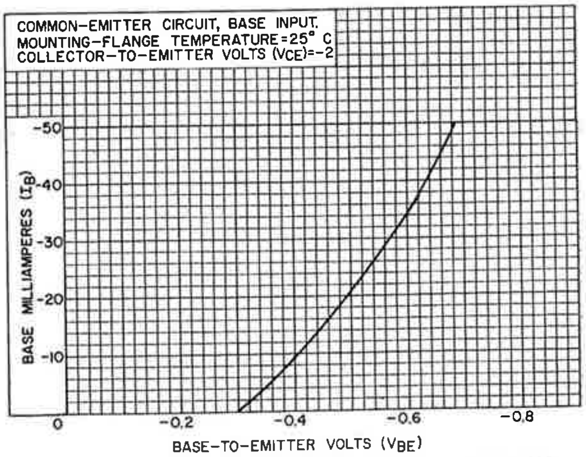
92CS-10993

Fig. 8—Typical current transfer characteristic for type 2N2148



92CS-11327

Fig. 9—Typical input characteristic for type 2N2147



92CS-11329

Fig. 10—Typical input characteristic for type 2N2148

ANNUAL INDEX, 1962

GERMANIUM P-N-P TRANSISTORS

2N2613, 2N2614

The 2N2613 and 2N2614 are alloy-junction transistors of the germanium p-n-p type intended primarily for use in small-signal and other low-power stages of high-quality audio amplifiers. Both types feature a high small-signal beta, typically 125 at a collector current of 0.5 ma for the 2N2613 and 160 at 1 ma for the 2N2614; minimum values are 60 and 100 respectively.

These types have a high cutoff frequency of 10 Mc typical, and uniform gain characteristics over the entire audio frequency spectrum.

The 2N2613 is a low-noise type, specifically designed for use in the input and low-level stages of equipment having stringent performance requirements at low idling-current levels of the order of 0.3 to 0.7 ma. In addition the 2N2613 features low saturation currents to assure excellent thermal stability at low current levels.

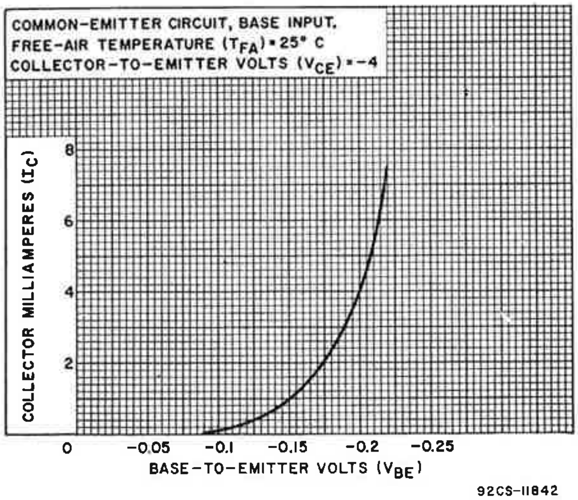


Fig. 1—Typical transfer characteristic for type 2N2613

MAXIMUM RATINGS

Absolute Maximum Values

	2N2613	2N2614	
Collector-to-base voltage	—13	—20	volts
Collector-to-emitter voltage	—10	—15	volts
Emitter-to-base voltage	0.5	1	volt
Collector current	—10	—50	ma
Emitter current	10	50	ma
Transistor dissipation:			
For free-air temperatures up to 55° C	100	100	mw
Derating factor	2.2	2.2	mw/° C
Temperature range:			
Storage		—65 to +100° C	
Operating (junction)		+100° C	

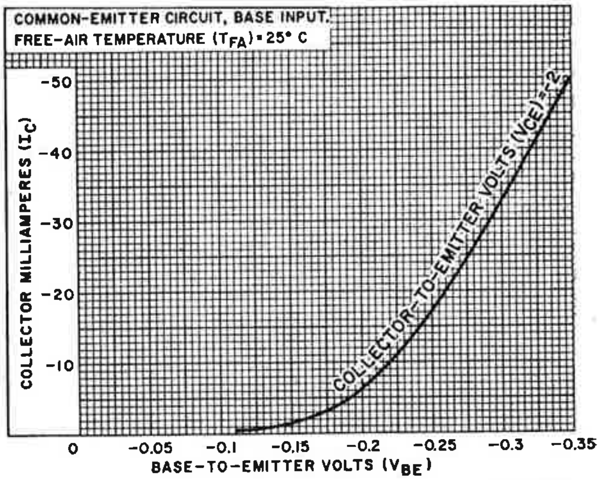
ELECTRICAL CHARACTERISTICS

At a free-air temperature of 25° C

Characteristics	Symbols	TEST CONDITIONS						LIMITS						Units
		DC Collector-to-Base Voltage V_{CB}	DC Collector-to-Emitter Voltage V_{CE}	DC Emitter-to-Base Voltage V_{EB}	DC Emitter Current I_E	DC Collector Current I_C	DC Base Current I_B	Type 2N2613			Type 2N2614			
		Volts	Volts	Volts	Ma	Ma	Ma	Min.	Typ.	Max.	Min.	Typ.	Max.	
Collector-Cutoff Current	I_{CBO}	-6 -12			0 0			– –	– –	-4 –	– –	– –	– -6.5	μa μa
Emitter-Cutoff Current	I_{EBO}			-0.5 -1		0 0		– –	– –	-12 –	– –	– –	– -12	μa μa
Collector-to-Emitter Breakdown Voltage	BV_{CEO}					-2.5	0	-10	–	–	-15	–	–	volts
Collector-to-Base Breakdown Voltage	BV_{CBV}			-2		-0.05		-13	–	–	-20	–	–	volts
Small-Signal Forward-Current Transfer Ratio (Measured at 1 kc)	h_{fe}		-4 -12			-0.5 -1		60 –	125 –	– –	– 100	– 160	– –	
Small-Signal Forward-Current Transfer-Ratio Cutoff Frequency	f_{hfb}		-4 -12			-0.5 -1		– –	10 –	– –	– –	– 10	– –	Mc Mc
Extrinsic Base-Lead Resistance Measured at 20 Mc	$r_{bb'}$		-4 -12			-0.5 -1		– –	300 –	– –	– –	– 300	– –	ohms ohms
Noise Figure: Reference Signal Frequency = 1 kc Circuit Bandwidth = 1.1 kc Generator Resistance = 1000 ohms	N_F●		-4			-0.5		–	–	5●	–	–	–	db
Equivalent RMS Noise Input Current over the frequency range 20 to 20000 cps: External Base-to-Emitter Resistance = 50000 ohms	i_N●		-4			-0.5		–	–	0.001●	–	–	–	μa
Collector-to-Base Feedback Capacitance	$C_{b'c}$		-4 -12			-0.5 -1		– –	14 –	– –	– –	– 8	– –	pf pf

● If the device is operated at free-air temperatures above 25° C, the maximum value for this characteristic may be higher than that shown.

In a common emitter type circuit, the 2N2613, operating with an idling current of 0.5 ma, can provide a noise figure of 5 db maximum, a small-signal beta of 125 typical, and a typical cutoff frequency of 80 Kc. Under these conditions, the 2N2613 can deliver an output signal current of 125 microamps with a signal-to-noise ratio of better than 60 db, for a 1 microamp input signal current from a high-impedance source. These performance characteristics make the 2N2613 particularly suitable for use in the input stages of high fidelity preamplifiers and similar applications where devices having exceptionally low noise characteristics are required.



92CS-11843

Fig. 2—Typical transfer characteristic for type 2N2614

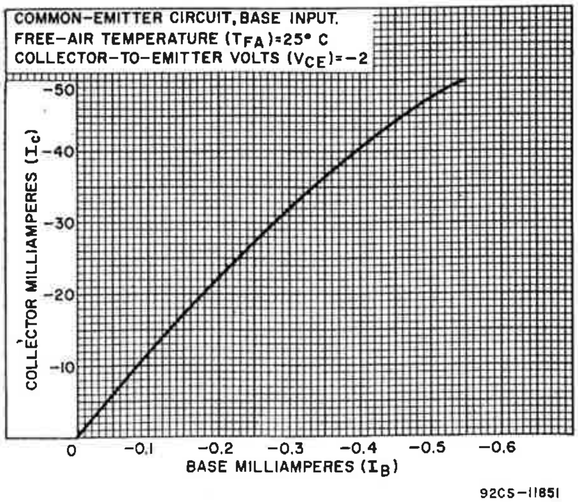


Fig. 3—Typical transfer characteristic for type 2N2614

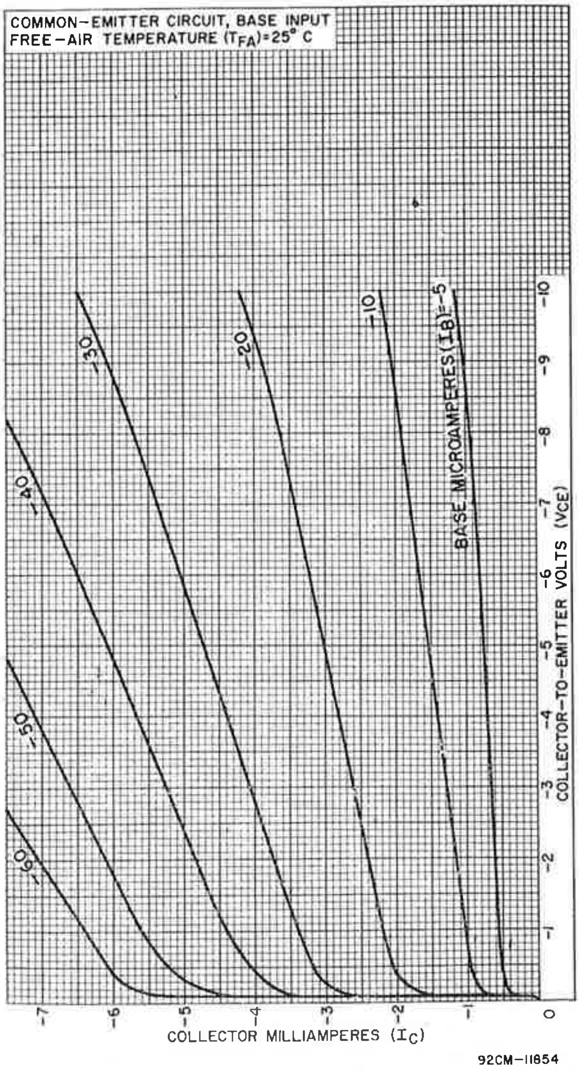


Fig. 4—Typical collector characteristics for type 2N2613

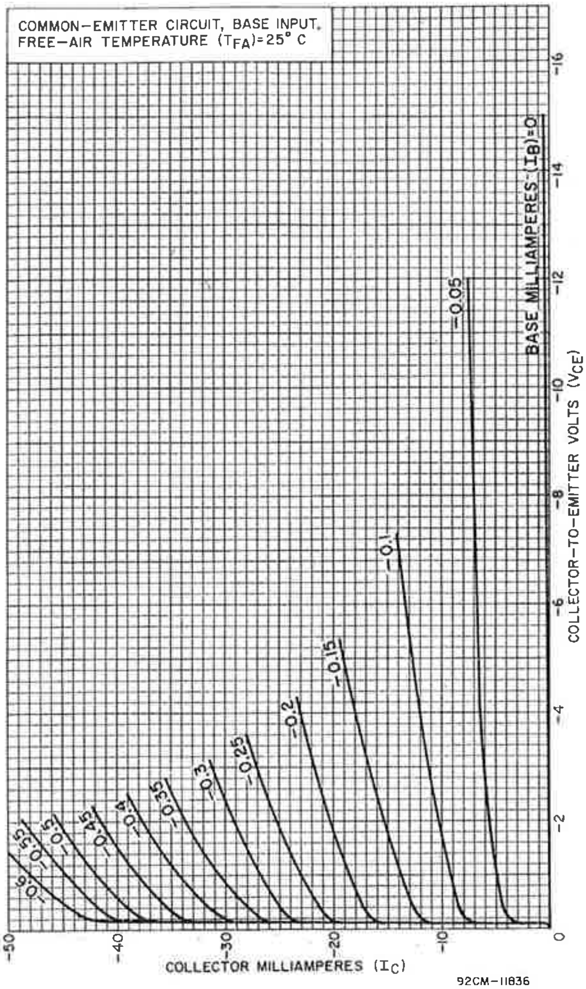


Fig. 5—Typical collector characteristics for type 2N2614

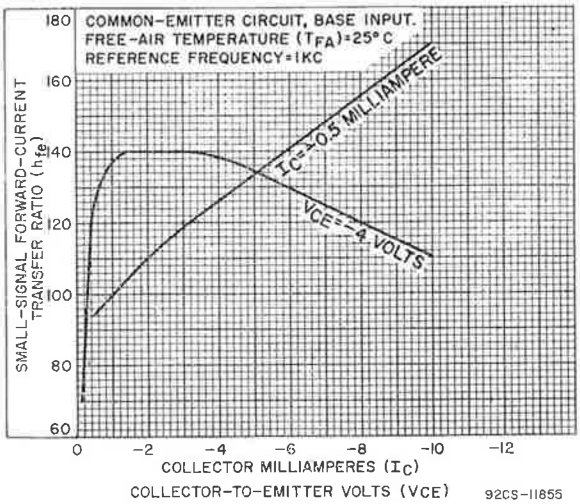


Fig. 6—Typical small-signal beta characteristics for type 2N2613

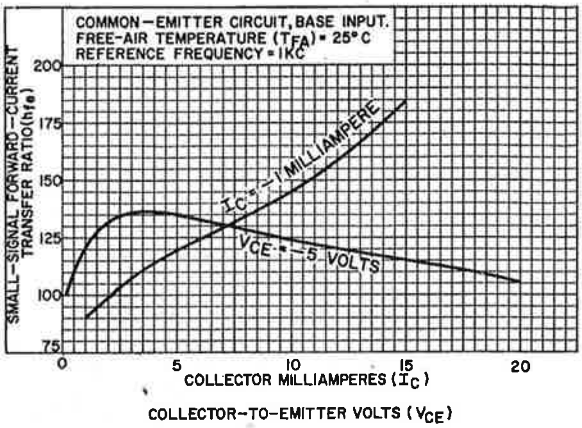


Fig. 7—Typical small-signal beta characteristics for type 2N2614

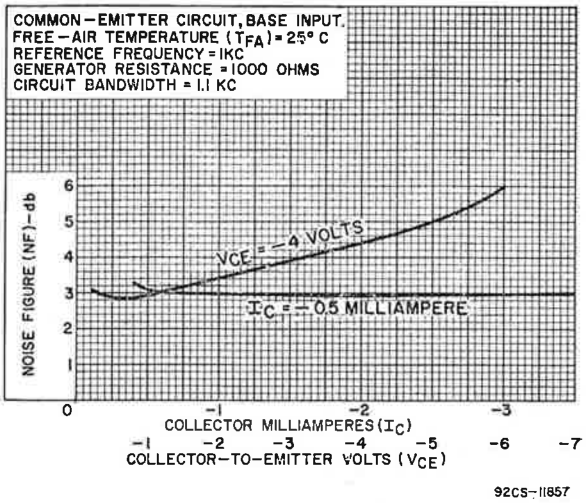


Fig. 8—Typical noise characteristics for type 2N2613

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Development of the experimental main amplifier described in these pages is now felt to have realised the potential of the configuration and components used. A finished version of the amplifier will now be prepared, and will be published in these pages in due course.

Editor .... **Bernard J. Simpson**

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