RADIOTRONICS

2

A Series on Tunnel Diodes

7: HIGH-CURRENT DEVICES

High-current tunnel diodes are best utilized as low-voltage inverters in circuits having low-impedance dc power sources.[39] They can be used for efficient inversion of the output of solar cells, thermoelectric generators, or thermionic converters. Other applications for these devices include overload detectors in dc and ac power supplies, pulse generators, high-speed switches, and oscillators.

High-current tunnel diodes are basically the same as conventional tunnel diodes in theory and construction. However, because of the high current involved, these diodes require a relatively large junction area. As discussed in the section on Characteristics, the peak current of tunnel diodes is directly proportional to junction area for a given crystal carrier concentration. A further difference of high-current tunnel diodes is the type of package used (some high-current devices generally use standard rectifier packages such as the DO-4 and the DO-8). In addition, if these diodes are operated at their maximum ratings, it is necessary that they be fastened to external heat sinks.

Conventional tunnel diodes and high-current tunnel diodes differ also in the value of the series resistance R_s. For high-current tunnel diodes, the series resistance should be kept as small as possible, generally in the order of 0.010 ohm, or less. If series resistance is not small, the efficiency of high-current tunnel diodes (especially when operated in inverter circuits) decreases rapidly. Fig. 87 shows the effect of an increase in series resistance on efficiency.

High-current tunnel diodes combine large total junction areas and rugged packages into units of great mechanical strength. In an experiment demonstrating this capability, a 10-ampere ger-manium unit operating in an inverter circuit was immersed completely into liquid nitrogen. Even at this extreme temperature ($-196°$ C), operation was successful. In fact, the inversion efficiency was improved considerably because of increasing peak-to-valley-current ratios at low temperatures.

Inverter Circuits

The two most common forms of inverter circuits for the conversion of low-voltage dc to higher-voltage ac (or dc) are shown in Figs. 88 and 89. The maximum conversion efficiency[40, 41] η for a diode operating in either circuit (if threshold operation is assumed) is given by

$$\eta \cong \frac{\left(\dfrac{I_P}{I_V} - 1 \right) \left(\dfrac{V_V}{V_P} - 1 \right)}{\left(\dfrac{I_P}{I_V} + 1 \right) \left(\dfrac{V_V}{V_P} + 1 \right)} \qquad (71)$$

As shown in Fig. 90, it is desirable to have large peak-to-valley current and voltage ratios. For germanium, typical current and voltage ratios are 10 to 1 and 3 or 4 to 1, respectively; these values correspond to efficiencies of 40 to 50 per cent. For GaAs, typical current and voltage ratios are 20 to 1 and 5 or 6 to 1, respectively. The voltage ratio depends primarily on the peak voltage V_P, which in turn depends on the series resistance R_s.

The actual peak and valley voltages, however, consist of two components, as follows:

$$V_P = V_P' + (I_P) (R_s)$$
$$V_V = V_V' + (I_V) (R_s) \qquad (72)$$
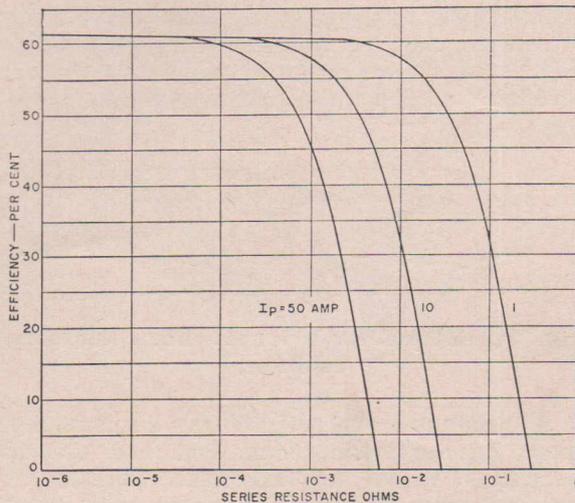
Fig. 87—Efficiency as a function of series resistance.

Fig. 89—Low-voltage dc inversion circuit.

where V_P' is the voltage across the junction itself. The inherent peak voltage V_P' of germanium diodes may be as low as 45 millivolts; inherent valley voltage V_V' may be in the range of 300 to 400 millivolts. These values correspond to a voltage ratio between 6 to 1 and 9 to 1.

As shown by the above equation, the value of R_s can have a major influence on the measured peak voltage (especially at high peak currents). For example, the typical series resistance of a one-ampere diode is 0.05 ohm. The value of $(I_P)(R_s)$ is then 1.0 x 0.05, or 50 millivolts, and the actual peak voltage is given by

$$V_P = 45 + 50 = 95 \text{ millivolts} \qquad (73)$$

Thus, the series resistance can reduce the voltage ratio approximately in half. The effect of series resistance on valley voltage, however, can usually be neglected because the term $(I_V R_s)$ is only one-tenth of the term $(I_P R_s)$.

If high conversion efficiencies are to be attained, the series resistance must be reduced as much as possible (see Fig. 87). Although efficiencies of approximately 50 per cent are presently realizable, this figure may be raised to 60 to 70 per cent by the further development of materials such as gallium arsenide. (Efficiencies greater than 60 per cent have already been obtained with experimental gallium arsenide high-current tunnel diodes.) Although both germanium and gallium arsenide are theoretically about equally well suited for high efficiencies, the greater peak-to-valley-current ratio and high-temperature capability offered by gallium arsenide may be advantageous for certain applications.
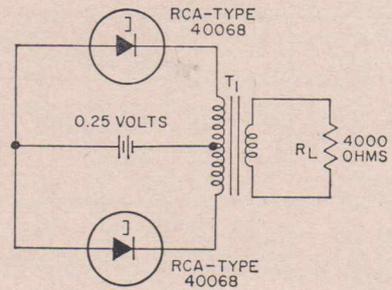
Although the efficiency of tunnel-diode inverters may seem low compared to the efficiencies of transistorized inverters, it must be remembered that transistors can only operate efficiently at relatively high voltages (one volt or higher). At lower voltage levels, tunnel diodes are considerably more effective, as shown in Fig. 91. At present the high voltages required for effective operation of transistorized inverters are obtained from a series arrangement of low-voltage dc sources (fuel cells, solar cells, thermoelectric generators). This method is not as reliable as the use of parallel supplies because the output of the entire series string is lost if only one cell opens.

The circuit shown in Fig. 88 is basically a relaxation oscillator. If the tunnel diode is biased in the negative region (for germanium diodes approximately 0.11 to 0.33 volts), oscillation having a trapezoidal waveform such as that shown

Fig. 88—Low-voltage dc inversion circuit.
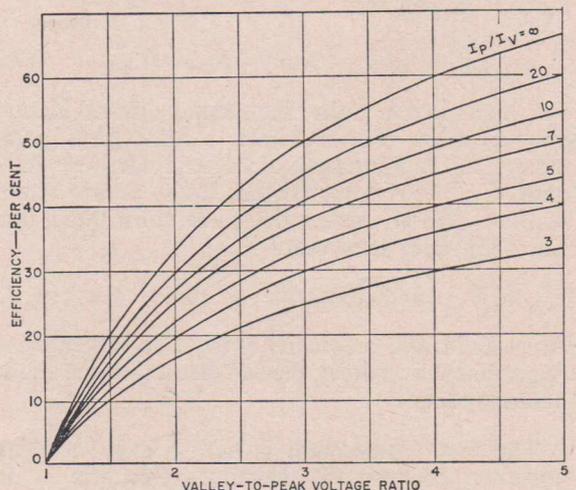
Fig. 90—Efficiency as a function of valley-to-peak voltage ratio.
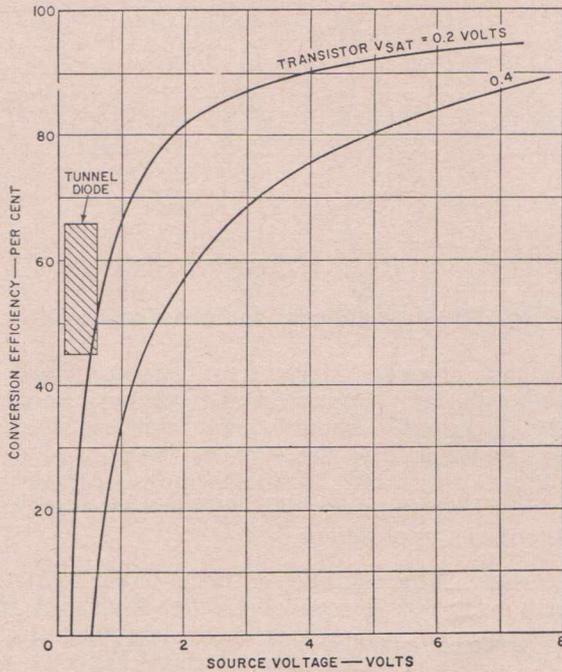
Fig. 91—Transistor and tunnel diode efficiency as a function of supply voltage.

in Fig. 92 occurs, provided that the following conditions are met: First, the total circuit resistance R_T, including diode series resistance, must be less than the absolute value of negative resistance of the diode at the operating point, as follows:

$$(R_T + R_s) < |R_j| \qquad (74)$$

Second, the total circuit series inductance L_T must be greater than the product of the total circuit resistance, the absolute value of the negative resistance of the diode, and the junction capacitance,[42] as follows:

$$L_T > (R_T + R_s)(|R_j|)(C_j) \qquad (75)$$

For high-current units, the second criteria is very easy to meet. For example, some typical parameters for a 5-ampere diode are: $(R_j) = 0.02$ ohm, $C = 0.011$ microfarad, $L_{CASE} =$ five nanohenries. When these values are substituted into Eq. (75), it is seen that

$$5 \times 10^{-9} > (0.02)(0.02)(1.1 \times 10^{-8} = 4.4 \times 10^{-12}.$$

As a result, the inequality is true, and the second requirement is met by the inductance of the diode package alone.

The first requirement is not so easy to meet, however, because the negative resistance is so small. For a 20-ampere germanium unit, the minimum negative resistance is about 0.006 ohm. As a result, the sum of the series resistance, the

power-supply resistance, the lead and contact resistances, and the transformer primary resistance must be less than 6 milliohms. Careful inverter design permits this requirement to be met.
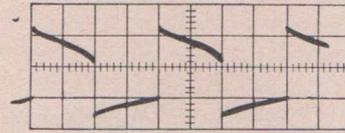

Fig. 92—Output for a tunnel diode relaxation oscillator.

A theoretical analysis[43] of the single-tunnel-diode inverter circuit under load and no-load conditions shows that maximum efficiency is obtained when the diode oscillates with equal positive and negative periods. An approximate equation for the frequency of a symmetrically oscillating loaded inverter is given by

$$f = \frac{1}{4L}\frac{(V_V - V_P)}{(I_P - I_V)} \qquad (76)$$

where L is the inductance of the primary winding of the loaded transformer.
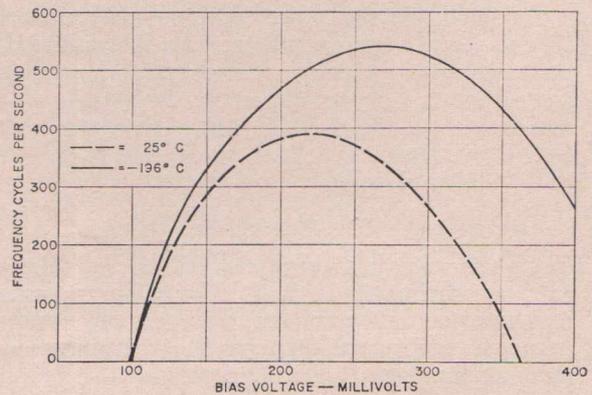
Fig. 93—Frequency as a function of bias voltage for type 40068 (resistance equals 1000 ohms).

The two-diode inverter circuit shown in Fig. 89 may be described as either parallel or push-pull. This circuit has three possible modes of operation which depend on the bias voltage, load resistance, and transformer-winding polarity and core material. If the transformer primary is wound in opposite directions on each side of the centre tap, a parallel mode of operation similar to that of the single-diode inverter mode occurs. Both diodes operate at the same frequency; the output waveform is similar to that shown in Fig. 92.

If the primary is wound in the same direction throughout, a push-pull arrangement results. The
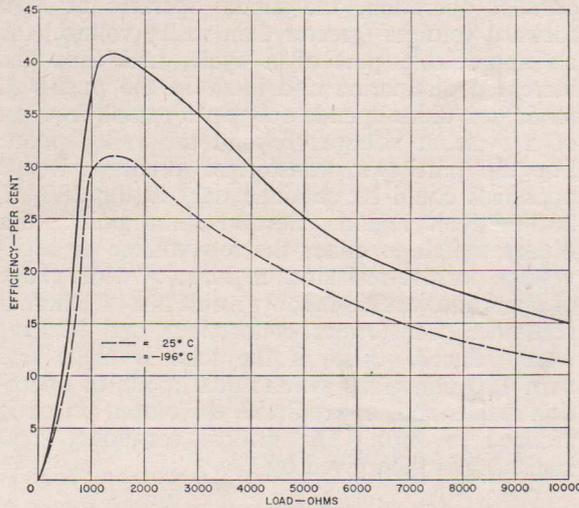
Fig. 94—Efficiency as a function of resistance for type 40068 (bias voltage is 240 millivolts).

Fig. 95—Frequency as a function of load resistance for type 40068 (bias voltage is 240 millivolts).

push-pull circuit configuration has two possible modes of operation. When the diodes operate in phase (symmetrically), efficiency is very low because the currents are "bucking" each other in the transformer primary. If the diodes operate 180 degrees out of phase (asymmetrically), the currents through the primary add to each other. In this mode of operation, conversion efficiency is high. The output waveform is symmetrical and approximately square, and its frequency is much lower than in the parallel mode of operation. Several excellent analyses and explanations of circuit operation for the push-pull inverter circuit are available in the literature.[44-47]

Experimental Circuits

In the inverter circuit shown in Fig. 88, the primary was constructed of six turns of heavy (No. 14) enamelled wire with a centre tap. Several hundred turns of fine wire were used for the secondary. The core material used was Carpenter "49", 0.014 inch. A regulated dc power supply was used as the energy source. Figs. 93, 94, 95, and 96 show the variations in tunnel-diode conversion efficiency and frequency as functions of bias or load resistance for two temperature ranges. Input dc power was measured with a dc ammeter and dc voltmeter. The voltmeter was connected directly across the diode. The transformer secondary was connected to a decade resistance box. Output power was measured with a true-rms-reading voltmeter, and computed by use of the following equation:

$$P_o = \frac{V_o^2}{R_L} \qquad (77)$$

The curves illustrate the increase of efficiency at −196 degrees Centigrade. This improved efficiency results because the peak-to-valley current

and voltage ratios increase considerably as the diode is cooled. For example, at room temperature the I_P/I_V ratio was 10 to 1; at liquid-nitrogen temperature it was 26 to 1. The shape of the curves is typical for all types of high-current tunnel diodes operated in a single-diode inverter circuit; however, different values of peak efficiency, frequency, and the like result when different transformers or diodes are used. Figs. 97 and 98 show the actual shape of the waveforms at three
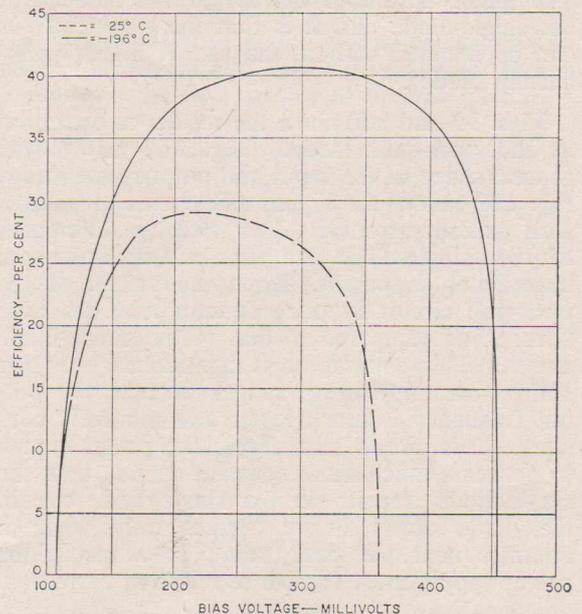
Fig. 96—Efficiency as a function of bias voltage for type 40068.
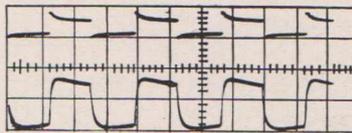
BIAS = 236 MILLIVOLTS
HORIZONTAL = 200 μSEC/DIV.
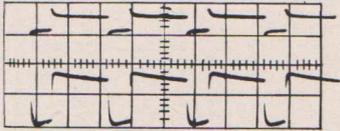
BIAS = 300 MILLIVOLTS
HORIZONTAL = 500 μSEC/DIV.
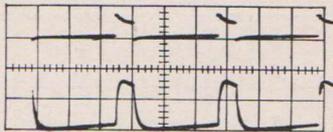
BIAS = 150 MILLIVOLTS
HORIZONTAL = 200 μSEC/DIV.

Fig. 97—Waveforms for the single-diode inverter.

different bias levels. The type of diode and transformer used also affect the exact bias voltage or load resistance at which the peak efficiency occurs.

The parallel mode of operation is very similar to single-diode operation. Curves for parameters such as efficiency and frequency have the same general shape as those for the single-diode circuit. The major advantage of the parallel mode over the single-diode circuit is that the output power can be doubled without the use of higher peak-current diodes.

Figs. 99 and 100 show the results of operation in the push-pull mode. Fig. 101 shows the general shape of the input and output waveforms. Fig. 102 shows the higher efficiencies obtainable with gallium arsenide. Two striking differences between single-diode and push-pull operation are immediately apparent. First, the output of the push-pull circuit is much squarer and has little "droop" as compared to that of the single-diode circuit. This squareness is continuous over the entire operating-voltage range. Second, the output frequency is much lower and relatively constant. This circuit also allows the power source to deliver a much more constant current than the single-diode circuit. In the single-diode circuit, the diode draws heavy current only when it operates near the peak point; as a result, the power supply must furnish a pulsating current.

In addition, the efficiency of the diode itself is materially greater in the push-pull circuit than in single or parallel operation. In single or parallel operation, the diode operates at high forward voltages (greater than valley voltage) for an appreciable part of a cycle, and thus has increased dissipation and loss. In the push-pull mode, the diode spends only a very small fraction of a cycle at voltages beyond the valley point. For the particular transformer used, push-pull operation could be obtained only in the 200- to 280-millivolt region. Beyond these points, the diodes switch to either the low-voltage or high-voltage positive-resistance regions. A wider range of operation would probably result if total external resistance (transformer, connections, and the like) were reduced. Also, if the load is made less than 190 ohms, the symmetrical mode of operation results and, as predicted, the output power is reduced to zero. The critical secondary load resistance is then given by

$$R_L = 2n^2 \left(\frac{[V_V + R_T I_V] - [V_P + R_T I_P]}{I_P - I_V} \right) \quad (78)$$

where R_T is the transformer primary-winding resistance and n is the transformer turns ratio. If a lower load resistance is used, the symmetrical mode of oscillation is destroyed, and no power output results.

Other Circuits

High-current tunnel diodes may be used to drive high-power transistors in either common-emitter or common-base configurations. For example, Fig. 103 shows the combined use of a
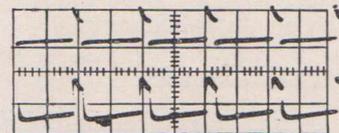
BIAS: 150 MILLIVOLTS
HORIZONTAL: 500 μSEC/DIV.
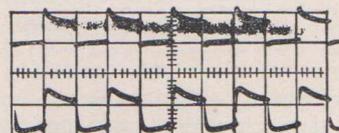
BIAS: 300 MILLIVOLTS
HORIZONTAL: 500 μSEC/DIV.
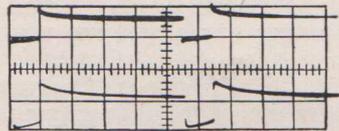
BIAS: 420 MILLIVOLTS
HORIZONTAL: 1 MS/DIV.

Fig. 98—Waveforms for the single-diode inverter. Vertical sensitivity: upper trace 0.5 V/Div, lower trace 50 V/Div.
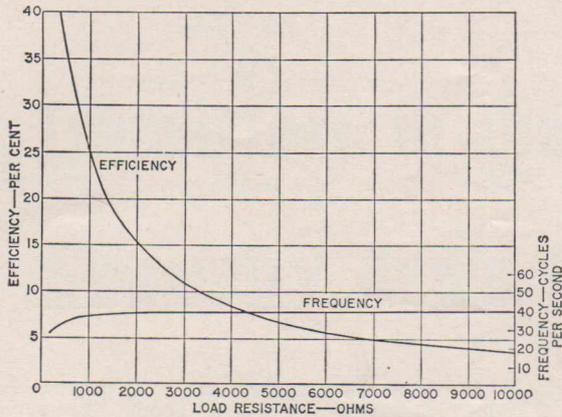
Fig. 99—Efficiency and frequency as a function of load resistance (load resistance is 1000 ohms).

germanium tunnel diode and a power transistor in a bistable switching circuit; in such circuits, the low rise time of the diode (as low as one nanosecond) can be used to drive transistors at rates much faster than many other methods presently available.

If silicon power transistors are used, their higher base-to-emitter voltage makes it necessary either to "bias up" the germanium tunnel diode, as shown in Fig. 104, or to use a gallium arsenide high-current tunnel diode. This application may be a very useful one for gallium arsenide diodes because their higher voltage swing could easily turn on a silicon transistor. At the same time, the base-to-emitter voltage of the transistor effectively clamps the gallium arsenide diode below
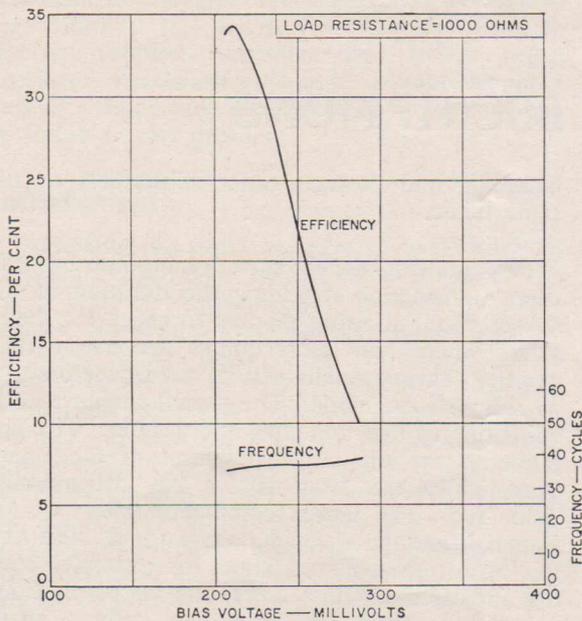
Fig. 100—Tunnel diode efficiency and frequency as a function of bias voltage.
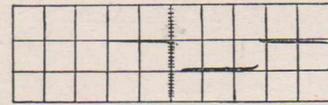
VERTICAL: 0.5V/DIV.
HORIZONTAL: 5MS/DIV.
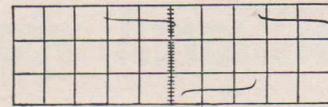
(b)

VERTICAL: 20.0V/DIV.
HORIZONTAL: 5MS/DIV.

Fig. 101—Inverter waveforms for two type 40068's operated in a push-pull circuit; (a) tunnel diode waveform, and (b) inverter output.

the point in the forward-voltage region at which the degradation mechanism of gallium arsenide becomes active.
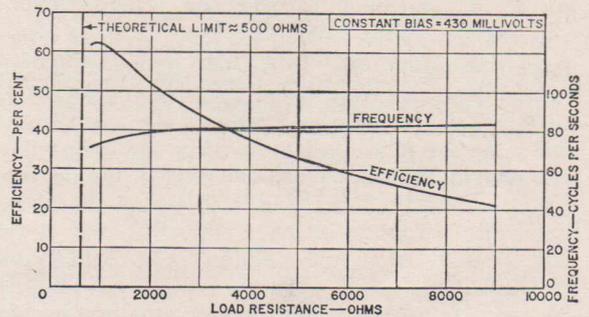
Fig. 102—Efficiency as a function of load resistance for six-ampere gallium arsenide tunnel diodes used in a push-pull inverter circuit.

Fig. 105 shows an example of a tunnel-diode overload-sensor circuit. This circuit uses high-current tunnel diodes and provides a fast-acting sensitive over-current detector which can be used to protect sensitive loads from current surges or overloads. Other circuit arrangements are pos-
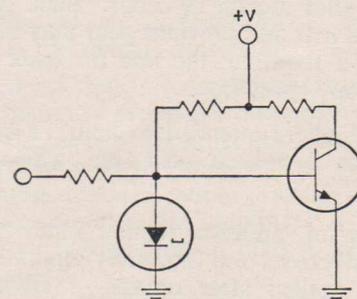
Fig. 103—Bistable tunnel-diode-transistor combination.
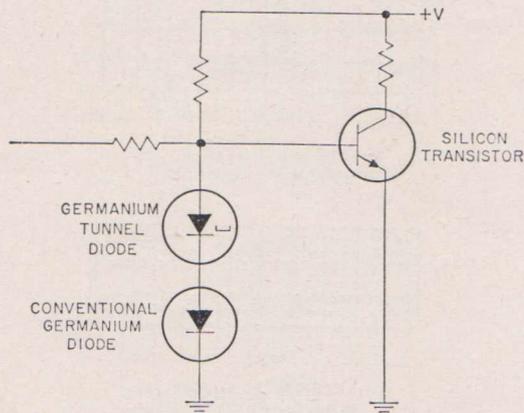
Fig. 104—Method for "biasing-up" tunnel diodes.
sible if the power supply is to be protected rather than the load.

Fig. 105—Tunnel diode overload sensor.

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41. Gevert, Wang and Broze, "Heat-to-AC Conversion Using Thermoelectric Generators and Tunnel Diodes", *Solid-State Electronics*, Vol. 3, 1961. In this paper, the equation for efficiency is given as

$$n = \frac{7}{8} \left[\frac{(I_P - I_V)(V_V - V_P)}{(I_P + I_V)(V_V + V_P)} \right]$$

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NOTICE RE SUBSCRIPTIONS

Transistors in Class B Complementary Symmetry Audio Frequency Amplifiers

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Summary

A description is given of an amplifier consisting of a power output stage in which n-p-n and p-n-p transistors are connected in a class B complementary-symmetry arrangement and a driver stage is directly coupled to the output. The use of complementary symmetry enables the development of power amplifiers which are transformerless and have a very good power frequency response, low weight and small size. Temperature stability has been a serious deterrent to directly coupled amplifiers and p-n-p, n-p-n transistor combinations, but by suitable application of temperature compensation a solution can be found to this problem.

Introduction

Following the rapid increase in availability of transistors after their development in 1948, the use of class B output stages in portable receivers became universal. The characteristics of class B operation which make it so important in this application are the relatively high power efficiency at high output power, the low quiescent power and the fact that the drain on the power source is proportional to the output power. The major disadvantage lies in the normal requirement to use input and output transformers with their problems of restricted frequency and phase characteristics, and for portable receivers, cost and weight. In addition, the secure mounting of transformers, which are often bulky, can be a serious problem in portable equipment where the use of printed wiring is common.

To overcome the use of the output transformer the single-ended class B push-pull type of output stage was developed using two identical output transistors and a driver transformer which has two equal but isolated secondary windings. The two output transistors are connected in cascade across the supply and for the same output requires twice the collector current, reducing the load impedance and enabling the voice coil of the speaker to be fed directly from the transistors and eliminating the output transformer.

The driving signals to each transistor are phased 180° apart and consequently two separate secondary windings are required on the driver transformer. The use of an n-p-n type of transistor in place of one of the p-n-p output transistors enables the phase of the driving signal to one transistor to be reversed, making both input signals of the same phase and allowing the use of one secondary winding on the driver transformer. This complementary-symmetry[1] configuration allows the two output transistors to be driven in parallel and to be coupled by a capacitor or directly to the driving transistor.

The general trend of development toward class B complementary symmetry output stages is indicated in the above introduction, and in the following section a description will be given firstly of the various types of output systems used generally in commercial audio amplifiers and secondly of the detailed operation of class B complementary symmetry.

1. Stanley, T. O., and Scott, T. M., "Design Considerations in Class B Complementary Symmetry Circuits", RCA Transistors I, 497.

Power Output Systems

Present output systems

The simplest form of output system consists of a single transistor operating under class A conditions. Two disadvantages follow:—
(1) the maximum power output for a given size transistor is low,
(2) the no load standing current is high.

The power required from the driver may be low enough to allow the use of a low level stage and the elimination of the driver coupling transformer. However, the use of an output transformer is essential since usually a high load impedance is required and the high standing current cannot be allowed to flow in the voice coil of the speaker.

Although class A operation has been used extensively in automobile receivers and to some extent in cordless mantel receivers, it is not attractive for portable receivers where power efficiency is important.

Class A power stages then have the following characteristics:—

Advantages

A low level driver stage is generally satisfactory.
The use of a driver coupling transformer is probably unnecessary.
Distortion, particularly at low power output, may be low.

Disadvantages

High current consumption in the idling condition.
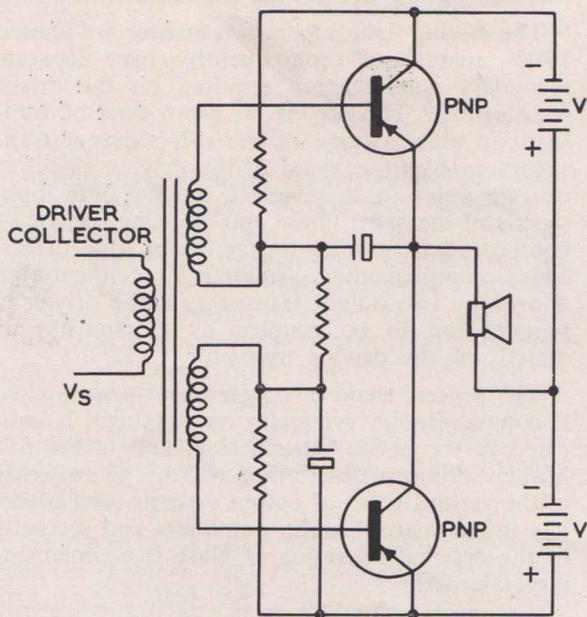
Fig. 1—Circuit diagram of a typical single-ended class B amplifier.

Low output power and an efficiency of approximately 45% maximum.
Extreme temperature sensitivity.
High operating cost for a given output.
Large size and weight due to the use of an output transformer.
Expensive due to the relatively poor ratio of power output to transistor dissipation.
Transistor dissipation highest on no-load conditions.
The output transformer restricts the use of feedback and limits frequency response.

The development of increased power from a given type of transistor necessitates the use of heat sinks or alternatively two transistors in parallel, but this has the same disadvantages as single class A. An improvement would be the use of two transistors in push-pull class A which would reduce distortion and ease the design of the output transformer. A very great improvement is obtained if two transistors are used under class B conditions[2]. This method of operation has the following features:—

Advantages

Low average current consumption, particularly when idling.
Efficiency considerably higher (approximately 70%).
Temperature stabilization simplified.
Operating cost considerably lower than class A for same output power.

Disadvantages

Dissipation peak is at about 50% of maximum unclipped output.
Driver transformer is essential.
Output transformer usually essential.
Frequency response dependent on transformers.
Size, weight and cost may be high, due to use of transformers.

Class B output stages have been used in almost all transistor receivers, due mainly to the power efficiency of the system. In an attempt to avoid the use of an output transformer some variations of the normal class B push-pull amplifier have been developed. In one, the speaker has a high impedance centre tapped voice coil in place of the transformer. A point worth noting here is that with the centre tapped speaker very small transformer action takes place between each half of the voice coil. Since each half operates for half of every cycle of the signal the whole voice coil is equivalent to approximately the impedance of one half effective over the whole cycle. Therefore, when measuring the power transferred to the

2. Light, L. H., "Transistor class B push-pull stages— Feedback arrangements in transformerless push-pull output stages", *Mullard Technical Communications,* **3,** No. 24, May, 1957; 98, 102.

speaker the approximate value will be given by measuring the rms value of the voltage across the whole voice coil and considering it to be acting across the impedance of half the total voice coil. The use of this method will result in general in an error of $+4\%$ to $+8\%$. The direct connection of the transistors to the voice coil eliminates the weight and poor frequency characteristics of the output transformer and obtains an overall cost improvement. These advantages are offset by the fact that since only half of the voice coil is effective at any instant the acoustic efficiency of the speaker is poor, and for a given electrical input power and magnet size the acoustic output will be approximately 3 db lower than that obtained using a conventional non-centre tapped voice coil. Another variation[2] is known as single-ended class B (refer to Fig. 1), where two similar transistors are connected in cascade across the supply, with the load connected between the common point of the transistors and the power supply. Only half of the supply voltage is available to each transistor and for the same output the collector currents must be doubled. A lower load impedance is therefore required and direct connection to a normal impedance speaker becomes possible. A driver transformer is still necessary, and since the collector of one transistor is connected to the emitter of the other, two separate secondary windings are necessary. This method has the following features:—

Advantages

Low average current consumption and efficiency as for normal class B operation.
Temperature stabilization relatively simple.
Higher negative feedback is possible with lower distortion.
Operating cost considerably lower than class A.
Output transformer unnecessary.
Weight, size and production cost can be reduced.

Disadvantages

The higher collector current swing may increase the non-linearity and nullify distortion gains.
The use of a driver transformer is essential.
A more complex and expensive driver transformer is necessary and it may further limit frequency response.
Low output systems may require the use of non-standard speaker voice coils.

Class B complementary symmetry

If an n-p-n transistor is placed in the position of the lower p-n-p transistor with the collector and emitter connections interchanged, as shown in Fig. 2, both transistors require an input signal of the same phase and their bases can be connected in parallel. They can then be driven either from a single secondary winding of a transformer
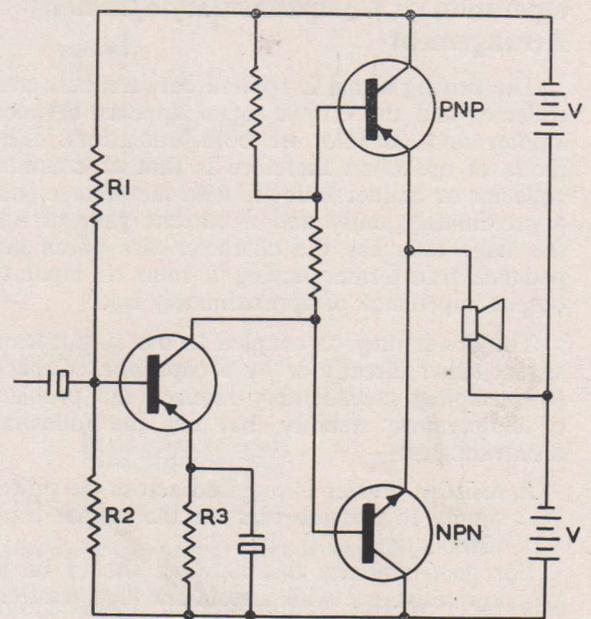
Fig. 2—Simplified circuit diagram of a complementary-symmetry class B amplifier.

or by networks in which the coupling is direct or provided by RC elements. Due to the symmetry about the common junction of the emitters this type of output stage is known as class B complementary symmetry pair, and is shown in Fig. 2[1, 3]. The features of this arrangement may be summarized as follows:—

Advantages

Low current consumption.
High efficiency when compared with class A.
Temperature stabilization relatively simple.
Driver transformer can be deleted with cost, space and weight savings.
Smallest possible unit size for a given output power.
Elimination of transformers allows higher feedback and also better frequency response.
Capable of lower distortion.

Disadvantages

Power sensitivity lower than conventional class B due to loss of driver transformer current gain.
Quiescent current for the driver may be higher than class B.
Overall efficiency, including power loss in the temperature stabilizing emitter resistors of the output transistors is approximately 55% (15% lower than for class B where the ratio of emitter resistance to collector load impedance is appreciably smaller).

3. Lindsey, J. E., and Wall, H. J., "Design considerations for direct coupled transistor amplifiers", *RCA Review*, Sept., 1958; 433.

Operation of Complementary-Symmetry Arrangement

The driving signal is applied between base and collector and the output signal appears between emitter and collector of both transistors. The mode of operation therefore is that of common collector or emitter follower with the voltage gain approximately unity and a current gain of h_{fe}; the stage thus has the characteristics of an impedance transformer having a ratio of input to output impedance of approximately h_{fe}.

The driver may be coupled to the output transistors either directly or by a capacitor. Capacitive coupling considerably reduces the problem of temperature stability, but has the following disadvantages:—

A resistive divider is required across the power supply to provide bias for the output transistors.

For good stability this network should be of low resistance with a resultant high wasteful bleed current.

An expensive, large value coupling capacitor is required.

Driver collector current may require to be increased over that required for direct coupling.

Bias network impedance must lower the load impedance for the driver and decrease gain.

Additional components required.

Advantage of space reduction due to the absence of driver transformer largely offset by size of the large coupling capacitor.

A consideration of the above points dictates the use of direct coupling, and this has been used for the circuits discussed in this paper. When direct coupling is used, temperature stabilization of the driver is essential, since otherwise the increase in driver collector current with temperature will move the operating point of the output transistors in opposite directions, thus unbalancing the distribution of supply voltage across the output transistors, causing increased distortion and loss of output power.

Referring to Fig. 2 it will be seen that if the bases of the two output transistors are connected together, there will be no forward bias voltage available between each base and its respective emitter, consequently in the quiescent condition collector current could not flow and low level operation would be very poor. This can be overcome by the insertion of a suitable resistive impedance between the two input bases and in series with the driver collector. This element can be a diode, a thermistor-resistor combination or a resistor depending on the application and type of temperature compensation required.

The driver collector load resistance can be supplied from either of two positions in the circuit, from the negative side of supply voltage or the active side of the output load. In the former case the driver collector voltage must vary sufficiently to produce at the base of each output transistor a peak signal voltage which is slightly greater than the peak signal voltage across the load. At full output this output signal voltage varies by almost half the supply voltage about the mean value, hence the driver collector voltage will vary from almost full supply potential to almost zero. Therefore the driver collector current must vary from near cut-off to twice the mean current, and any non-linearity of h_{fe} will be evident as distortion at high output levels. In addition the driver will require a high input signal and its sensitivity and input impedance will be low.

Connecting the collector of the driver to the active side of the output load applies a voltage having two components between the lower collector load and the output bases. One component is the direct voltage which is built up across the isolating capacitor and is constant, and the other an alternating voltage which is in phase with the driver collector voltage and almost equal to it in amplitude. In fact it differs from it only by the sum of the signal required between the base and emitter of each of the output transistors and the part of the output signal appearing across the emitter resistors (see Fig. 3). The net voltage which must be maintained by the driver across its collector load resistance which comprises the resistor R_4, R_5 and R_6 (Fig. 3) and the input impedance of the output transistors therefore, is this direct voltage (approximately 4.5 volts) and the small signal voltage required to drive the output transistors.

The direct component of this voltage is provided by the potential difference due to the zero signal collector current of the driver in its collector load resistance and the relatively small signal component by the swing of the collector current due to the driving signal.

Compared with the case where the driver load resistor is connected to the negative supply point the impedance presented to the driver by the bases of the output transistors is very much higher since the same collector voltage swing produces the very much smaller current swing represented by the signal current to both bases. Another way of looking at this is to consider each of the output transistors as an emitter follower. In the first case the ac load on the driver is the effective input impedance of the output transistor (high due to the negative feedback of the emitter follower) in parallel with the driver load resistor R_4, R_5 and R_6 shown in Fig. 3. In the second case the driver load is produced by the same value of input impedance in parallel with a resistance which is higher than R_4, R_5 and R_6 by a factor determined by the negative feedback
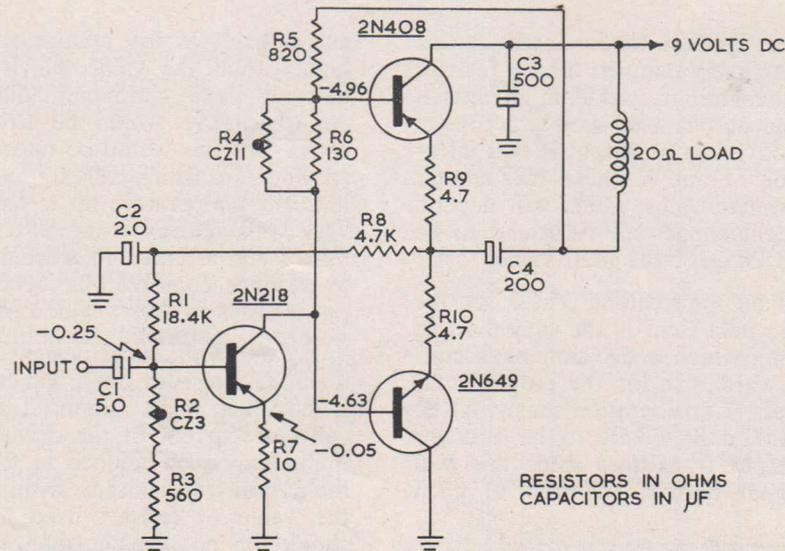
Fig. 3—Circuit diagram of a 300 mw complementary-symmetry class B amplifier.

due to the emitter circuit of the output transistors (see appendix).

In the practical case for the 300 mw circuit shown in Fig. 3 the effective load for the driver in the first arrangement is approximately 2000 Ω in parallel with 820 Ω (570 Ω) and in the second is 2000 Ω in parallel with 3800 Ω (1310 Ω (see appendix).

The higher effective ac load presented to the driver in this arrangement means that a smaller signal current is required from the driver and hence the gain of the driver stage is higher.

The output signal appears between the junction of the emitters of the output transistors, and the centre of the power supply. The current from the power supply is drawn in pulses which, because they flow only when the n-p-n transistor is conducting, have a repetition frequency half that met with in conventional class B. In addition, because only half the total supply voltage is effective over each half cycle the pulses have an amplitude twice that of the normal class B circuit. The impedance of the power supply therefore must be relatively low to avoid both a loss of output power due to poor regulation and excessive coupling between stages of the amplifier. The ac load then as far as the output stage is concerned can be placed between the common emitters and either side of the supply. In the complementary-symmetry arrangement discussed in this paper, use is made of a direct coupled driver with its collector load resistance connected to the high impedance side of the output, and hence the low impedance side of the output load must be connected to the negative side of the power supply. Since a direct potential difference of approximately 4.5 volts would exist across the output load a blocking capacitor must be used

to prevent the flow of a direct current. The value of the capacitor should be large so that its reactance at the lowest frequency to be amplified is small compared with that of the load. This capacitor is normally charged to half the supply voltage and as such acts as a power source for the p-n-p transistor when it is conducting. The reduction in the charge on the capacitor due to the load of the p-n-p transistor is restored during the other half cycle when the n-p-n transistor is conducting.

No class B amplifier is practical unless temperature stabilization is used, and in the complementary-symmetry arrangement this covers two sections of the amplifier. The output transistors must have their quiescent current stabilized to avoid excessive dissipation at high temperature. The driver transistor also must have its collector current stabilized against an increase in ambient temperature to avoid an alteration in the mean potential of the output bases with a resultant change in the distribution of the power supply voltage across the output transistor.

A resistor in the emitter circuit of each output transistor is a method of temperature stabilization which is simple and, depending on the design specifications, may be sufficiently effective[3, 4, 5].

As this resistor is increased the negative feedback is increased with a resultant improvement in temperature stabilization, reduction in distortion and increase in the amplitude of the input signal required for a given output. In addition, since the emitter resistance is part of the output load a smaller part of the total output power will

4. Tharma, P., "Compensation for changes in base to emitter voltage with temperature", *Mullard Technical Communications*, **3**, No. 24, May, 1957; 106.
5. Hunter, L. P., Handbook of Semiconductor Electronics, pp. 11-38.

be available to the speaker load. Since in this circuit with a capacitively coupled load a resistor must be used in each emitter, and as in all class B amplifiers, only one output transistor at a time is operating, the effective loss of output is that introduced by the value of one of these two similar resistors. The optimum value which will depend on the design specifications will be found to be between 10% and 20% of the load value.

Having selected an approximate value for the emitter resistor the next item is the operation of the driver, following which a decision on a suitable source of forward bias for the output transistors can be made. Consideration must first be given to the minimum h_{fe} likely to be encountered with the output transistors since this will determine the signal current required at each base.

The minimum driver collector current can then be determined such that the power required to drive the output stage will be available from the driver with low distortion. The dependence of overall distortion on driver collector current and driver load resistance for the 300 mw design treated in this paper is shown in Fig. 4.

Following the selection of the driver collector current, the value of the driver collector load and the voltage drop required for the forward bias of the output transistors can be obtained in this direct coupled arrangement by the use of one of three types of load in the collector circuit of the driver; (1) a germanium diode, (2) a thermistor-resistor combination and (3) a resistor. The diode is unsatisfactory since the bias that must be provided is the sum of bias voltages for the two transistors plus the sum of the potential drops
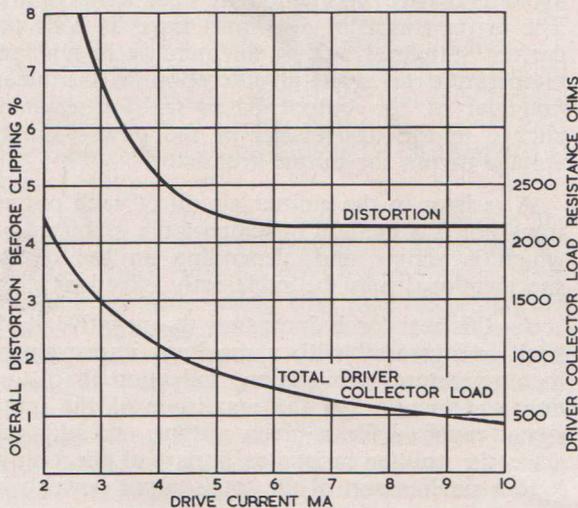
across each of the emitter resistors, and this is higher than the value which can be developed across a single germanium diode at low currents. An alternative would be to use two diodes in series, but this would be uneconomical. A design specification which calls for safe operation at high ambient temperature up to say 60°C can be met very satisfactorily using a combination of thermistor and resistor. A combination can be chosen to produce constant quiescent current in the output transistors over a wide range of temperature.

For most practical designs, however, the third method, a resistor only, will be satisfactory. This is possible, if as proposed in this design, the collector current of the driver is stabilized. The maintenance of balance in the drive to each of the output transistors to within 10% requires that the value of resistor used to develop the bias should be not greater than approximately 10% of the total driver collector load.

The bias supply for the driver can be obtained from two points, from the collector supply in the normal manner, or alternatively from the output emitter junction with a filter to remove signal feedback[2]. The latter method introduces a small amount of negative dc feedback which assists temperature stabilization, and with the assistance of a thermistor-resistor series combination from the driver base to earth, produces an almost constant collector current for a variation in ambient temperature of 20-60°C. Knowing the driver collector current and bias required for the output transistors, it is possible to determine the value of bias resistor which will produce the required output quiescent current. This current, if too low, will cause crossover distortion due to the non-linear transistor characteristics near cut-off, and if too high will reduce battery life. Since high temperatures normally will increase the quiescent current, causing a reduction in crossover distortion, and low battery voltage will reduce the quiescent current and increase the crossover distortion, the determination of the optimum quiescent current must be a compromise between the above conflicting factors.

The use of an emitter resistor in the driver stage will improve the temperature stability[3, 4] and if not bypassed to audio provides negative feedback and will reduce gain and improve distortion. The reduction in gain may be compensated by an increase in gain of the pre-driver stage due to the increased input impedance of the driver. The 10 Ω resistor in the design shown in Fig. 3 produced a drop in gain (feedback only) of approximately 6-8 db. An additional byproduct of the use of negative feedback in this way is the smaller variation in input impedance over each cycle of the signal. This of course appears as lower distortion in the driver stage. The leakage current I_{CBO} is the important factor in determining the temperature stability of driver and it has been

Fig. 4—Overall distortion and driver collector load resistance versus driver collector current for the circuit of Fig. 3.

found that for a leakage current greater than a certain value, sufficient correction cannot be obtained by a thermistor or thermistor-resistor network

Fig. 5 indicates the variation required in the driver bias resistor ($R_2 + R_3$) to maintain the driver collector current constant over the indicated changes in ambient temperature. The increase in the variation required in R_2 and R_3 as I_{CBO} increases can be clearly seen.

Careful consideration of operating conditions and requirements will enable the design of an amplifier capable of operation in ambient temperatures up to 60°C. The specified ratings of components will not be exceeded and the output power will be maintained with little increase in distortion.

Circuit Design Considerations

General

The design of a practical output stage entails the consideration of a number of factors. The designer must decide, in collaboration with all interested parties, upon a design which is functional, and yet can be produced at the lowest possible cost.

An important factor in determining cost is the maximum acoustic output power that is required from the receiver. Both the efficiency of the speaker and the output stage of the amplifier are important and the circuit designer must decide the contribution each must make in meeting the performance specification. It is possible, for example, to increase the acoustic power output from the receiver by using a speaker of higher efficiency.

The majority of portable receivers are required to produce an electrical power output of approximately 300 milliwatts and the minority represented by the small personal type an output in the order of 150 mw.

Two such amplifiers will be discussed in this paper. The first will be capable of developing 300 to 310 milliwatts, between 20° and 60°C, in a 20 ohm load, and the second, 150 to 160 milliwatts over the same temperature range into a 50 ohm load.

Circuit description, 300 mw amplifier

Fig. 3 shows the complete circuit of the 300 milliwatt amplifier, and the main features are as follows:—

(i) *Direct coupling*. Consideration of coupling methods given earlier in this paper indicates the use of direct coupling giving directly coupled feedback which assists temperature stabilization. This feedback loop is decoupled to eliminate any signal feedback and consequently does not reduce the amplifier sensitivity.
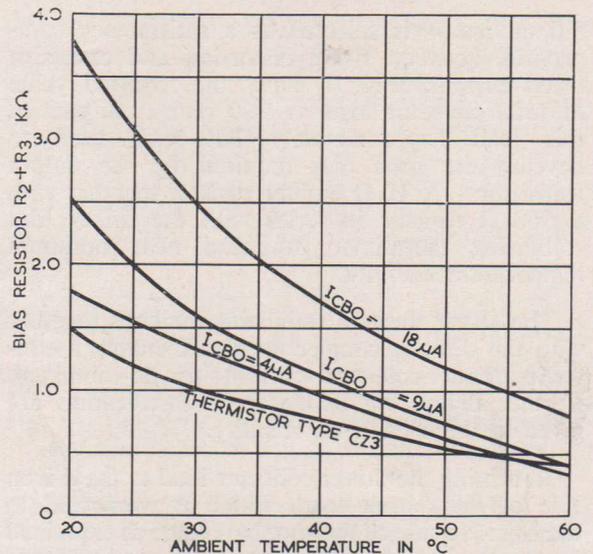
Fig. 5—Value of the bias resistor required to maintain a constant driver collector current, versus ambient operating temperature for three values of driver transistor cutoff current; the resistance versus temperature characteristic of a CZ3 thermistor is included on the same scales.

(ii) *Temperature stability*. One of the restrictions in the use of direct coupled amplifiers is the temperature stability problem associated with them. The preferred amplifier is usually the dc coupled type despite the fact that in general it is more complex and costly.

The temperature range over which the design must be stable will of course depend on the application. For equipment for use indoors the range of 20-40°C is satisfactory, for portable and auto receivers which can be used in direct sunlight a range of 20-65°C is desirable.

Various methods can be used to obtain the required stability, and in general all will reduce other performance characteristics as the stability factor is improved.

The increase in the collector current I_c of the output transistors with temperature due to the increase in I_{CBO} and the reduction in the forward bias V_{BE} required for a given I_c is partly compensated by the negative feedback due to the emitter resistors R_9 and R_{10} Fig. 3. Further compensation is possible if a thermistor is used as shown to control the bias in the base circuit.

(iii) *Driver amplifier*. The choice of the driver transistor depends upon the temperature stabilizing methods. An inspection of Fig. 5 indicates that I_{CBO} must not be greater than approximately 4 μa if a thermistor is to be used for compensation. This then dictates the use of a germanium alloy transistor of the rf type such as the 2N218 which has a maximum I_{CBO} of 5 μa.

(iv) *Driver operating conditions*. With the aid of the chart given in Fig. 4, a driver current of

5.0 milliamps is selected as a satisfactory compromise between rising distortion and excessive bleed current. Fig. 4 shows the required value of total collector load as 900 ohms. A part of this load (approximately 10%) is used to develop the total bias required by the output transistors. A 10 Ω emitter resistor together with a 560 Ω resistor in series with the driver bias stabilizing thermistor produced near optimum temperature stability.

The above thermal stabilizing methods together with the directly coupled feedback enable a constant driver collector current to be obtained. Further details on temperature stabilization are given in the section on stability.

Returning the driver collector load to the driven side of the output load, which is connected to the negative side of the supply voltage as explained previously, enables full use to be made of the driver capabilities.

(v) *Output stage operating conditions*. Both the 2N408 p-n-p germanium alloy transistor and its n-p-n complementary transistor type 2N649 may be fitted with cylindrical heat sinks. They then have the required dissipation rating up to 60°C to produce 300 milliwatts of power output. The effective collector supply is 4.5 volts, since only half of the supply is available to each transistor.

Considerations of transistor power dissipation and power output set the total output load at 25 ohms and this load is returned to the negative supply point as stated above. Connection of the load directly to the junction of the emitter necessitates the use of a large capacity blocking capacitor to isolate the supply voltage from the junction, otherwise the p-n-p transistor would be inoperative. Since this capacitor is in series with the load its impedance must be low compared with the load to avoid loss of output signal.

The capacitor is normally charged to a potential of 4.5 volts and provides the collector supply for the p-n-p transistor when it is operating. It is discharged slightly during this part of the cycle and is recharged to original condition when the n-p-n transistor is operating. Because the supply current pulse is present only when the n-p-n transistor is operating a bypass must be used to keep the supply impedance low. The use of a 500 μf capacitor will enable satisfactory operation down to a 9 volt battery life end point of 4.5 volts.

The total load for the output transistors (25 Ω) is made up of two parts—one the speaker voice coil and the other a resistor which is connected in series with each emitter circuit. As discussed earlier the value of this emitter resistor will be made equal to approximately 20% of the total

Table 1

All Measurements at 400 cps.

		Amplifier 1	Amplifier 2
Output for THD = 10%	mw	310	160
Output for THD = 5%	mw	250	130
Distortion before clipping	%	5	5
Distortion at 35 mw output	%	3	3
Sensitivity at full output	mv	50	20
Sensitivity for 50 mw output	mv	20	12
Input impedance at 20°C.	ohms	780	730
Input impedance at 60°C.	ohms	600	530
Load impedance	ohms	20	50
Quiescent current at 20°C.	ma	8	6·5
Quiescent current at 60°C.	ma	11	7·2
Current consumption at full output	ma	60	31
Sensitivity for 50 mw output*	mv	9	6·8

*With 500 microfarad capacitor in parallel with the driver emitter resistor R_7, Fig. 3.

load, i.e. 4.7 Ω. To allow for variation of h_{fe} in the driver transistor the value of R_1 (Fig. 3) should be adjusted to obtain 4.8 volts at the junction of the emitter resistors.

(vi) *Feedback*. Signal feedback has been provided by emitter resistors, 10 Ω in the driver and 4.7 Ω in the output transistors. Bypassing the driver emitter will require about 500 μf and increases the sensitivity by 6-8 db with a proportional increase in distortion. The output emitter resistors increase the input impedance of the output transistors to approximately 2000 Ω and the connection of the driver load to the high impedance side of the output load increases its effective impedance to approximately 3800 Ω producing a total collector load of approximately 1310 Ω.

The output impedance of the driver transistor operating in this arrangement with a 10 Ω emitter resistance is approximately 100 kΩ and has a negligible effect on the circuit impedances. The other feedback loop from output transistor emitter junction produces zero signal feedback but approximately 2 db of direct coupled feedback.

(vii) *Performance*. Details of performance are given in Table I under "Amplifier 1".

150 milliwatt amplifier

(i) *Differences between* 300 *and* 150 *mw designs*.

The circuit of the 150 milliwatt amplifier is similar to that of the 300 milliwatt amplifier shown in Fig. 3 except for the following modifications which are required for optimum results:—

(a) Lower maximum output is obtained by reducing the peak output currents. This is achieved by increasing the output load impedance from 20 Ω to 50 Ω.

(b) The lower output currents require less driving current and consequently the driver transistor collector current can be reduced from the original 5.0 milliamps to 3.5 milliamps for the same overall distortion.

(c) The modified driver collector current requires a higher value collector load. To maintain the 4.5 volts at the driver collector, the load resistor R_5 (Fig. 3) must be increased to 1200 Ω and to maintain the 3.5 milliamps collector current the driver forward bias must be reduced by suitable adjustment of R_1 in Fig. 3.

(d) Temperature stabilization can be simplified because the dissipation in the output transistor is lower and the omission of the thermistor R_4 in the output stage bias network with R_6 decreased to 110 Ω produces a total quiescent current change between 20°C and 60°C of less than one milliampere.

(ii) *Performance.* Details of performance are given in Table I under "Amplifier 2".

Conclusion

The operation of a p-n-p and an n-p-n transistor in the complementary-symmetry type of af amplifier has been described.

A basic amplifier of 300 mw output and the modifications required to limit its power output to 150 mw have been outlined. Both amplifiers are particularly suitable for low cost portable receivers due to the savings in cost, weight and size and the adaptability to printed circuit construction brought about by the absence of driver and output transformers.

Further Reading

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Wieneck, Z., "Complementary symmetry sliding class A audio system", *I.R.E. Transactions on Broadcast and TV Receivers,* Nov., 1961; 21.

Johnson, L. B., and Vermes, Miss P., "D.C. stabilization of junction transistors", *Mullard Technical Communications.* Vol. 3, No. 23, April, 1957; 68.

APPENDIX

The treatment given below assumes that the transistor characteristics, such as h_{fe}, h_{oe}, h_{re} and $r_{bb'}$ are relatively constant over the range covered by the operating conditions of the amplifiers. A constant voltage source driving the input transistor TR1 is assumed, from which will be derived the voltage gain of the amplifier. The
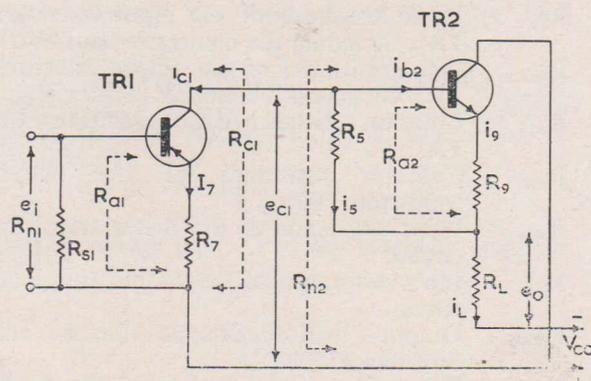

Fig. 6—The circuit of Fig. 3 simplified to enable easy derivation of the circuit characteristics.

speaker load, normally substantially resistive, will be assumed to be the impedance of the voice coil at 400 cps.

The symbols used in the following analysis are defined below:

A_v	Overall voltage gain of the amplifier.
e_{c1}	Peak incremental collector voltage across the load resistance R_{n2} of the driver transistor TR1.
e_1	Peak incremental input voltage.
e_o	Peak incremental output voltage across the load R_L.
G_{m1}	Effective transconductance of the TR1 circuit.
G_{m2}	Effective transconductance of the TR2 circuit.
h_{fe1}	Short circuit output forward current gain of the driver transistor TR1.
h_{fe2}	Short circuit output forward current gain of the output transistor TR2.
h_{oe1}	Open circuit output admittance of the driver transistor TR1.
h_{oe2}	Open circuit output admittance of the output transistor TR2.
h_{re1}	Open circuit input reverse voltage gain of the driver transistor TR1.
h_{re2}	Open circuit input reverse voltage gain of the output transistor TR2.
i_{b2}	Peak incremental base current of transistor TR2 due to the input voltage e_1.
i_{c1}	Peak incremental collector current of transistor TR1 as a result of e_1.
i_5	Peak incremental current in the load resistor R_5 of transistor TR1.
I_7	Steady emitter current of transistor TR1.
i_9	Peak emitter current of the transistor TR2 due to the base current i_{b2}.
i_L	Sum of the peak currents in the load R_L.
P_o	Power output level at which all practical measurements were made.
$r_{bb'1}$	Extrinsic base resistance of the driver transistor TR1.
$r_{bb'2}$	Extrinsic base resistance of the output transistor TR2.

R_{a1} Input resistance of the input transistor TR1, including the emitter resistor R_7.

R_{a2} Input resistance of the output transistor TR2, including the emitter resistor R_9.

R_{c1} Collector resistance of the driver transistor TR1.

R_{c2} Collector resistance of the output transistor TR2.

R_{n1} Input resistance of the driver transistor circuit.

R_{n2} Input resistance of the output transistor circuit.

R_L Output load resistance (voice coil impedance).

R_{s1} External source resistance of the driver transistor TR1.

Determination of Circuit Impedances

A simplified circuit of the amplifier is shown in Fig. 6. From Fig. 6:

$$e_{c1} - e_0 = i_5 R_5 \qquad \qquad \qquad (1)$$

$$= i_{b2} R_{a2} \qquad \qquad (2)$$

Also $e_0 = R_L(i_9 + i_5)$ by inspection ;

But $i_9 = i_{b2}(h_{fe2} + 1)$,

therefore $e_0 = (h_{fe2} + 1)i_{b2} + i_5 R_L \qquad (3)$

Then from equations (1), (2) and (3),

$$e_0 = i_{b2} R_L(h_{fe2} + R_{a2}/R_5 + 1) \qquad (4)$$

Equation (1) indicates that

$$e_{c1} = i_{b2} R_{a2} + e_0,$$

and from equation (4)

$$e_{c1} = i_{b2}\left[R_{a2} + R_L\left(h_{fe2} + \frac{R_{a2}}{R_5} + 1\right)\right] \quad (5)$$

(i) Input resistance of the Output Stage

$$R_{n2} = \frac{e_{c1}}{i_5 + i_{b2}}$$

Substituting for e_{c1} from equation (5), and using a relationship obtained from equations (1) and (2), namely :

$$\frac{i_{b2}}{i_5 + i_{b2}} = \frac{R_5}{R_5 + R_{a2}}$$

it can be proved that

$$R_{n2} = \frac{R_5}{R_5 + R_{a2}}\left[R_{a2} + R_L\left(h_{fe2} + \frac{R_{a2}}{R_5} + 1\right)\right] \quad (6)$$

(ii) Input resistance of the Output Transistor

It can be shown that [5,6] the input resistance R_{a2} (including the emitter resistance R_9) is given by :

$$R_{a2} = \left(h_{fe2} + 1\right)\left(R_9 + \frac{26\pi}{i_9}\right) +$$

$$r_{bb}{}'_2 - h_{re2}\left[\frac{h_{fe2} R_L}{1 + h_{oe2} R_L}\right] \quad \cdot\cdot \quad (7)$$

The internal emitter resistance $26\pi/i_9$ is an approximation, assuming a sine wave signal for the two output transistors. That is, the signal for each transistor is a half sine wave, neglecting any crossover distortion and quiescent current.

(iii) Input resistance of the Driver Transistor

Assuming that the emitter current of the driver transistor is large compared with the emitter signal current, a similar expression can be derived for the driver input resistance :

$$R_{a1} = \left(h_{fe1} + 1\right)\left(R_7 + \frac{26}{I_7}\right) +$$

$$r_{bb}{}'_1 - h_{re1}\left[\frac{h_{fe1} R_{n2}}{1 + h_{oe1} R_{n2}}\right] \quad \cdot\cdot \quad (8)$$

The values of h_{fe}, h_{re}, h_{oe} used in equations (7) and (8) will be modified by the unbypassed emitter resistors and the relationships are as follows :— [7].

$h'_{re} = h_{re} + h_{oe} R_e$, ($R_e$ is the external

$h'_{fe} = h_{fe}$,   emitter resistance).

$h'_{oe} = h_{oe}$.

Normally, at audio frequencies, $h_{oe}(R_e + r_e) \ll 1$, where r_e is the internal emitter resistance. The above equations include this approximation. The actual values used in equations (7) and (8) can be corrected if it is deemed that the correction will be significant.

(iv) Input resistance of the Driver Stage

The circuit input resistance is then the combined value of R_{a1} and R_{s1} in parallel, that is

$$R_{n1} = \frac{R_{a1} R_{s1}}{R_{a1} + R_{s1}} \quad \cdot\cdot \quad \cdot\cdot \quad (9)$$

Determination of Overall Voltage Gain

By their definition the following is known,

$$G_{m1} = \frac{i_{c1}}{e_1} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (10)$$

$$h_{fe1} = \frac{i_{c1}}{i_{b1}} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (11)$$

From these equations, and noting that $R_{a1} = \frac{e_1}{i_{b1}}$ by definition, it can be shown that :—

$$G_{m1} = \frac{h_{fe1}}{R_{a1}} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (12)$$

Also by definition,

$$G_{m2} = \frac{i_L}{e_{c1}} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (13)$$

and

$$h_{fe2} = \frac{i_9}{i_{b2}} - 1 \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (14)$$

6. Wilshire, H. R., "Transistors and their parameters", *Radiotronics,* **22,** No. 10, October, 1957; 159.
7. Hunter, L. P., "Handbook of Semiconductor Electronics", 11-26.

Now substituting equations (5) and (14) in equation (13) and knowing that

$$i_L = i_9 + i_5$$

then simplifying the result, it can be shown that

$$G_{m2} = \frac{h_{fe2} + \frac{R_{a2}}{R_5} + 1}{R_{a2} + R_L\left(h_{fe2} + \frac{R_{a2}}{R_5} + 1\right)} \quad (15)$$

The voltage gain A_v is given by,

$$A_v = \frac{e_o}{e_{c1}} \cdot \frac{e_{c1}}{e_i} \quad .. \quad .. \quad .. \quad (16)$$

and

$$\frac{e_o}{e_{c1}} = G_{m2} \frac{R_L}{1 + h_{oe2}R_L} \quad .. \quad .. \quad (17)$$

$$\frac{e_{c1}}{e_i} = G_{m1} \frac{R_c}{1 + h_{oe1}R_c} \quad .. \quad .. \quad (18)$$

where $R_c = \dfrac{R_{c1}R_{n2}}{R_{c1} + R_{n2}} \quad .. \quad .. \quad (19)$

Combining equations (17) and (18) in equation (16), and noting that in the practical amplifier considered here $1 \gg h_{oe2}R_L$ the following expression can be derived

$$A_v = \frac{G_{m1}G_{m2}R_cR_L}{1 + h_{oe1}R_c} \quad .. \quad .. \quad (20)$$

Evaluation of the 300 milliwatt Practical Amplifier

The values of all the amplifier components are as follows :

R_L	20 ohms.
R_5	900 ohms. (820 + 90 ohms).
R_7	10 ohms.
R_9	4·7 ohms.
I_7	5 milliamps.
i_9	70 milliamps peak.

$$[\sqrt{(2P_o/R_L)} - i_5]$$

P_o	50 milliwatts.

The transistor parameters are as follows :

h_{fe1}	80
h_{fe2}	80
h_{oe1}	33 x 10^{-6} mho.
h_{oe2}	33 x 10^{-6} mho.
h_{re1}	10^{-3}
h_{re2}	10^{-3}
$r_{bb'1}$	200 ohms.
$r_{bb'2}$	100 ohms.

Input resistance R_{n2} of the output transistor TR2 circuit

Substituting the appropriate values in equation (7), obtaining a value for R_{a2} and substituting this value in equation (6),

$R_{n2} = 1350$ ohms. 1300 ohms.*

Input resistance R_{n1} of the driver transistor TR1 circuit

R_{n1} can be derived by substituting the appropriate values in equation (8), obtaining a value for R_{a1}, and substituting this value in equation (9),

$R_{n1} = 780$ ohms. 780 ohms.*

Overall voltage gain A_v of the complete amplifier

Equation (20) indicates that

$$A_v = \frac{G_{m1}G_{m2}R_cR_L}{1 + h_{oe1}R_c},$$

and from equation (19), since $R_{c1} \gg R_{n2}$, then $R_c \doteqdot R_{n2}$. This modified equation (20) producing the following relationship :

$$A_v = \frac{G_{m1}G_{m2}R_{n2}R_L}{1 + h_{oe1}R_{n2}}$$

Substituting the circuit values in equations (12) and (15)

$G_{m1} = 61$ milliamps/volt.

$G_{m2} = 37$ milliamps/volt.

Then the above derived values of transconductance together with the circuit values of R_{n2} and R_L, are substituted in the gain equation, then

$A_v = 57$ 50*

Evaluation of the 150 milliwatt Practical Amplifier

The values of all the amplifier components are as follows

R_L	50 ohms.
R_5	1300 ohms.
R_7	10 ohms.
R_9	4·7 ohms.
I_7	3·5 milliamps.
i_9	45 milliamps.
P_o	50 milliwatts.

The transistor parameters are shown for the 300 milliwatt amplifier.

Repeating as for the 300 mw amplifier, but using the modified values shown above the following characteristics can be calculated :

Input Resistance R_{n2} of the TR2 circuit
$R_{n2} = 3050$ ohms.

Input Resistance R_{n1} of the TR1 circuit
$R_{n1} = 780$ ohms. 780*

Overall Voltage Gain A_v
$A_v = 142$ 132*

* **NOTE.** Values measured in the practical amplifier. (R_{n2} for the 150 milliwatt case was not measured). Differences between the measured and calculated performance characteristics are due to minor approximations used in derivation of equations and the sum of the accuracy tolerances in the measurements.

450-MILLIWATT COMPLEMENTARY-SYMMETRY AMPLIFIER

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Following the design of the 300 mw amplifier described in the previous article, a modified design has been developed producing 50% higher output power, and using a low-level amplifier before the driver to improve the overall sensitivity. This increase in output power has been obtained at the expense of a reduction in the maximum operating temperature from the original 64°C to 58°C. (A maximum ambient temperature of 60°C was quoted in the original article to provide a conservative safety margin.)

The ratio between the total maximum unclipped power output and the collector dissipation of each output transistor can be proved to be about 4.5:1 when the output stage operates under class B conditions. For the case of maximum transistor dissipation, which occurs when the output power is equal to approximately 50% of the maximum unclipped power output, the ratio of the maximum available unclipped power output to the maximum permissible dissipation for each transistor must be taken as 3.8:1.

In determining safe operating temperature, the latter of these two ratios is the important consideration. When the practical limitations are considered, and noting that the thermal resistance from junction to ambient of the transistor under consideration of 225°C/watt represents a permissible dissipation of approximately 4.5 milliwatts for each degree Centigrade, a total practical power output improvement of 17 milliwatts into the load is available for each degree Centigrade

+ TRANSISTORS FITTED WITH FLAG TYPE HEAT SINKS.   RESISTORS ARE 0.5W RATING.
STATED VOLTAGES ARE APPROXIMATE ONLY
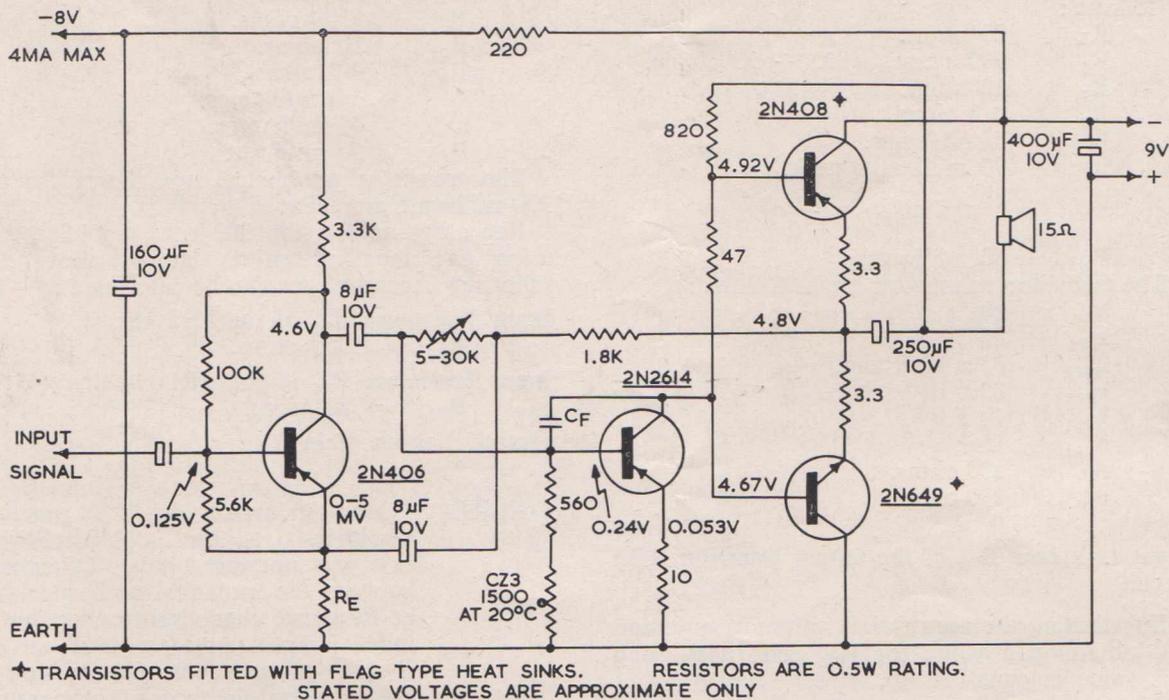
Circuit diagram of the improved amplifier discussed in this note.

that the maximum permissible ambient temperature is reduced.

The reduction of the maximum ambient temperature from 64°C to 58°C produces an improvement in output power of approximately 100 milliwatts. The maximum unclipped power output increases from 270 milliwatts to 370 milliwatts, and the output with 10% distortion from 300 milliwatts to 450 milliwatts. The calculated ambient temperature of 58°C represents the point at which the transistor's junction reaches the maximum allowable temperature of 85°C for the output transistors. The maximum permissible ambient temperature has been set at 55°C for this amplifier to provide a safety margin similar to that of the amplifier described in the previous paper.

The general circuit shown in the schematic diagram is similar in major details to the original circuit, but to obtain the increased output power the load impedance has been reduced to 15 ohms, together with a reduction from 4.7 ohms to 3.3 ohms of the emitter resistors of each of the output transistors. Due to the control of the driver transistor's collector current by the thermistor in the base circuit, the collector voltage is relatively constant. The use of a thermistor in the bias network of the output transistors has been found unnecessary for stable operation up to the maximum ambient temperature of 55°C.

An additional audio stage, capacitively coupled to the driver stage, has increased the sensitivity of the system from the original 20 millivolts to less than 1 millivolt for an output power of 50 milliwatts. This stage is operated in the common emitter mode, with a collector current of 1 milliamp, providing a power gain of 28 db and an overall power gain in the order of 88 db.

Negative feedback has been used to provide two other versions of the amplifier with lower distortion and sensitivity, but higher input impedance. Frequency response of the amplifiers without feedback normally extends from 100 cps to 14 Kc. All three amplifiers have been tailored to provide a response characteristic slightly broader than the typical passband of present-day portable receivers. This reduces any tendency towards high or low frequency instability due to coupling in a practical amplifier.

The table below tabulates the performance of the three designs. The first, No. 1, is similar to the amplifier in the original paper with negative feedback due to the unbypassed emitter resistor of the driver transistor. No. 2 amplifier has a feedback loop from the output to the emitter of the low level amplifier, producing an overall sensitivity of 1.6 millivolts and a higher input impedance than No. 1 amplifier. No. 1 and No. 2 amplifiers would be suitable for a typical transistor receiver employing a total of seven transistors. No. 3 amplifier has higher feedback and input impedance, but lower sensitivity and distortion, and would be suitable for any receiver employing a total of eight transistors. This latter combination should produce an overall sensitivity of about 30 μv/metre for a 6 db signal-to-noise ratio.

A capacitor of 400 microfarads is shown connected across the 9-volt supply, to provide maximum battery life. A saving in cost at the expense of a somewhat reduced battery life can be achieved by deleting this component.

The complete amplifiers have characteristics which are shown in the accompanying table, measured at 20°C, with a supply of 9 volts. The test frequency, where applicable, was 400 cps.

		AMPLIFIER		
		No. 1	No. 2	No. 3
Emitter resistor R_E—	ohms	0	2·2	5·6
Feedback capacitor C_F—	pf	470	1200	1200
Maximum ambient temperature—	°C	55	55	55
Output power for THD = 10%	mw	460	465	470
Output power for THD = 5%	mw	340	410	415
Output power before clipping—	mw	370	370	370
Distortion before clipping—	%	5·5	2·7	2·0
Sensitivity at full output—	mv	2·1	5·6	11
Sensitivity at 50 mw output—	mv	0·7	1·6	3
Input impedance—	ohms	800	2000	3700
Input current at full output—	ma	84	85	86
Input current at 50 mw output—	ma	31	31	31
Total quiescent current—	ma	8	8	8
Quiescent current at 55°C—	ma	20	20	20
Low frequency at 3 db down—	cps	100	50	30
High frequency at 3 db down—	kc	5·0	5·0	8·5

Three amplifiers have been shown, all producing an output power of 450 milliwatts at 10% distortion. The first two can be coupled to a suitable detector, or if improved performance is required, the third amplifier can be coupled to a high-level detector with an improvement in reproduction quality and sensitivity. The output system is single-ended, and uses a voice coil of standard impedance without matching transformer. Therefore, for the same acoustic output, it requires a drain on the power supply of approximately 70% of that required by the normal transformer-coupled class B circuit, and 50% of that required by the arrangement using a centre-tapped voice coil.
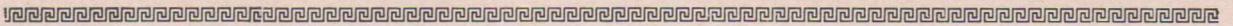
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