# A Dual Channel Transistor Power Amplifier*

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***Summary*—This paper describes the theory and performance of a dual channel transistor power amplifier capable of developing 30 watts in each channel with less than 1% intermodulation distortion. Full power is developed at frequencies below 20 cycles per second. Detailed transformer construction and amplifier assembly information is presented along with a list of parts and purchasing information.**

THE BASIC circuit of the amplifier[1,2] described in this paper was presented at the 1957 IRE National Convention. It has been redesigned to take advantage of transistor improvements and price reductions that have been made during the intervening years. In this paper it is presented with a power capability of 30 watts in each channel. The amplifier can be built as a single channel amplifier or as a double channel amplifier one channel at a time. In either case, the two channel power transformer, which is the only element common to the two channels, should be used to facilitate future conversion to two channel operation.

The basic power amplifier circuit shown in Fig. 1 has three common emitter stages and is capable of accepting in excess of 25 db of over-all voltage feedback. In this circuit a low level *n-p-n* transistor TR1 is direct coupled to the driver *p-n-p* transistor TR2. The driver transistor is in turn coupled through the driver transformer *DT* to the output stage *p-n-p* transistors TR3 and TR4. An output autotransformer *OT* is used to match the load $R_L$ to the output transistors. With this arrangement the collector of TR4 is at approximately the same dc potential as the emitter of TR1, and a direct coupled over-all feedback network, consisting of a resistor $R_{fb}$ and a capacitor $C_{fb}$ in parallel, is connected between them. The resistor controls the low and middle frequency overall feedback while the capacitor controls the high-frequency over-all feed-back. Obviously the same general arrangement could be used if TR1 were a *p-n-p* transistor and TR2, TR3, and TR4 were *n-p-n* transistors.

The 10 K resistor $R_1$ completes an internal negative feedback loop that involves TR1, TR2, and the driver transformer. The input must be increased by roughly 10 db to compensate for the effect of this resistor. The two resistors $R_2$ also provide negative feedback in this loop, but the capacitor $C_2$ limits their effect to the very low-frequency range. One of these resistors $R_2$ also appears

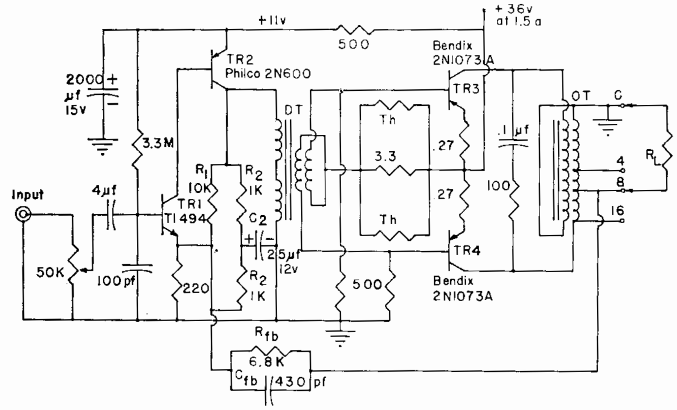


Fig. 1—Power amplifier circuit.

as a damping resistor across the primary of the driver transformer at higher frequencies.

Conduction transfer occurs in both the input and output of the output stage, and in order to avoid conduction transfer notches,[3] the secondaries of the driver transformer *DT* and the two windings of the output autotransformer *OT* are bifilarly wound.

The driver transformer is one of the major reactive components in the feedback loop and it is desirable that it have low primary-secondary leakage reactance. The low primary-secondary leakage reactance is attained by winding the primary in two equal sections with a bifilar secondary sandwiched between them. The primary of the driver transformer carries dc current and this must be taken into account in its design.

The output transformer winding consists of a relatively small number of turns of parallel wires suitably insulated from each other. In general just two parallel copper wires with Heavy Formvar insulation are adequate. The ground return is used as the common output terminal and taps are brought out on one section of the bifilar winding to match 4, 8, and 16 ohm loads. The laminations in this transformer should be fully interleaved.

The operating points of TR1 and TR2 are controlled by the resistor $R_B$ in the base circuit of TR1. This resistor should be chosen to produce 11 volts between the emitter of TR2 and ground when the supply voltage is 36 volts. This automatically sets the collector current of TR2 at about 50 ma and this is adequate to drive the output transistors to saturation. Normal operation produces very little temperature rise in TR1 and TR2

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[1] A. B. Bereskin, "A high power high quality transistor audio power amplifier," 1957 IRE National Convention Record, pt. 7, pp. 149–161.

[2] A. B. Bereskin, U. S. Patent No. 2,932,800; April 12, 1960; assigned to the Baldwin Piano Company, Cincinnati, Ohio.

[3] A. P.-T. Sah, "Quasi-transients in class B audio-frequency push pull amplifiers," Proc. IRE, vol. 24, pp. 1522–1541; November, 1936.

so that stabilization beyond the effect of the 500 ohm dropping resistor is not necessary.

The output transistors TR3 and TR4 operate over a relatively wide temperature range and require bias stabilization. The bias is provided by the voltage developed across the 3.3 ohm base return resistor and the dc resistance of the driver transformer secondary windings by the current flowing through the two 500 ohm resistors connected between the two bases and ground. Temperature compensation is provided by the two thermistors *Th* connected in parallel with the 3.3 ohm base return resistor. These thermistors have a resistance of 5 ohms at 25°C and a −3.9 per cent/°C temperature coefficient of resistance. They are thermally coupled to the collectors of TR3 and TR4. Additional bias stabilization is obtained from the two 0.26 ohm resistors in series with the emitters of TR3 and TR4 at the expense of approximately 6 db decrease in the loop gain.

The 100 pf capacitor connected between the base of TR1 and ground provides stabilization for frequencies in excess of 100 kc. The series combination of 0.1 $\mu$f and 100 ohms connected between the collectors of the two output transistors makes the amplifier open circuit stable. With some loudspeaker loads this combination may be removed, while with others it is essential in order to keep the system from oscillating at frequencies in excess of 100 kc.

The power supply used with this amplifier is shown in Fig. 2. It consists of one power transformer *PT* with two single phase full wave bridge rectifiers. Simple choke input filters are used. Fuses are provided in the ac line and also in the individual dc channels. Pilot lights have been incorporated in the ac and dc sections of the power supply.

The low level frequency response characteristics in in Fig. 3 were obtained in each case with the input potentiometer set for one half the output, at 1 kc, that resulted when it was set at its maximum position. This is not necessarily the same position in each case since the shunting effect of the amplifier input varies with the amount and type of feedback used. The value of 0.774 volts was chosen as the 0 db reference because it corresponds to 1 mw at 600 ohms and is the reference used on the scale of the VTVM.

Data for curve *A* was obtained with all local and over-all feedback removed except for the 0.27 ohm emitter resistors in the output stage. In this case response is within 3 db of the 1 kc value between 100 cycles and 15 kc. At the low end the response is dropping off at 9 db/octave while at the high end it is dropping off at about 13 db/octave.

Adding the 10 K resistor $R_1$ changed the response to that of curve *B*. The input signal required was 11 db higher than in the previous case. The low frequency end of this curve has the same shape as the previous curve but has been displaced toward the left by about $1\frac{1}{2}$ octaves. The high frequency end of the curve remains

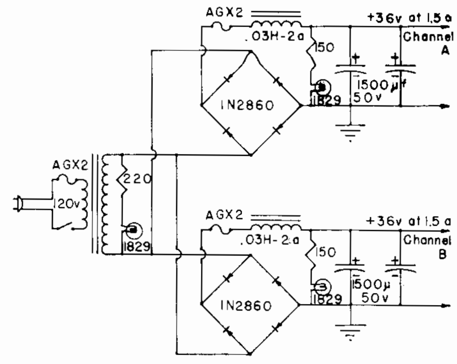
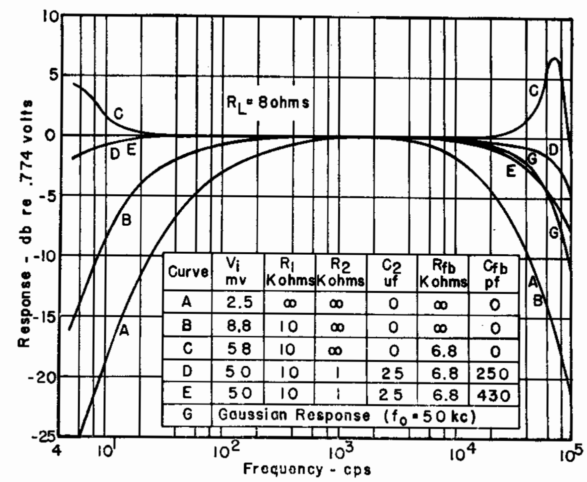



Fig. 2—Power supply.




| Curve | $V_i$ mv | $R_1$ Kohms | $R_2$ Kohms | $C_2$ uf | $R_{fb}$ Kohms | $C_{fb}$ pf |
|---|---|---|---|---|---|---|
| A | 2.5 | ∞ | ∞ | 0 | ∞ | 0 |
| B | 8.8 | 10 | ∞ | 0 | ∞ | 0 |
| C | 58 | 10 | ∞ | 0 | 6.8 | 0 |
| D | 50 | 10 | 1 | 25 | 6.8 | 250 |
| E | 50 | 10 | 1 | 25 | 6.8 | 430 |
| G | Gaussian Response ($f_0$ = 50 kc) | | | | | |

Fig. 3—Low level frequency response.

unchanged showing that the drop off in high frequency response, up to 100 kc, is due almost entirely to the output stage which is not affected by $R_1$. The response is now within 3 db of the 1 kc value between 25 cycles and 15 kc.

The further addition of $R_{fb} = 6.8$ K between the 8 ohm output tap and the emitter of TR1 results in the response shown by curve *C*. The input had to be increased by an additional 16.5 db to obtain this curve. The response is now essentially flat between 20 cycles and 20 kc, but there are undesirable humps at both the low- and high-frequency ends. These could result in both low and high-frequency instability.

The low-frequency response was corrected by inserting resistors $R_2$ and capacitor $C_2$ to reduce the low-frequency loop gain. The high-frequency response was corrected by using a 250 pf capacitor for $C_{fb}$. This combination gave rise to curve *D*. The shunting effect of $R_2$ across the primary of *DT* reduced the middle frequency loop gain so that the input signal required was now only 15 db above that of curve *B*.

Uniform frequency response at high frequencies is not as important as good transient response. To insure good transient response, $C_{fb}$ was changed to 430 pf because it was determined experimentally to be the value necessary just to eliminate ringing with a 5 kc square wave. The resulting response characteristic is shown as curve *E*. Good transient response is insured if the frequency response is Gaussian

$$\left( A = A_0 e^{-.346\left[\frac{f}{f_0}\right]^2} \right)$$

and to see how closely the experimental curve approached this response a curve for Gaussian response with $f_0 = 50$ kc has been drawn in as curve *G*. Within the limitations of its shape, curve *E* appears to be as good an approximation to curve *G* as could be expected. Response characteristic data was also obtained for 20 per cent and 80 per cent settings of the input potentiometer. For these two cases the low frequency response was identical to curve *E* and the high frequency response was slightly higher than curve *E*, being everywhere within $\frac{1}{2}$ db of curve *G*.

The 5 kc square wave response of the amplifier is shown in Figs. 4(a)–(c). Fig. 4(a) is for 20 volts, 2 volts, and 0.2 volts peak to peak signal across an 8 ohm resistive load. Slight dissimilarities are evident between the leading and trailing edges of the square wave. The dissimilarities are due to the fact that, since class *B* operation is employed, only one output transistor is working at a time. There are bound to be differences in the frequency response of individual transistors so that the best value of $C_{fb}$ for one transistor would not necessarily be best for the other unit. The value of $C_{fb}$ used is therefore a compromise between the best leading and trailing edges of the 5 kc square wave. Fig. 4(b) is for an RCA 515S2 loudspeaker, housed in a 10 cubic foot base reflex cabinet, connected to the 16 ohm tap of the amplifier. Fig. 4(c) is for a GE S1201D-7 loudspeaker connected to the 8 ohm tap of the amplifier. This speaker was housed in a simple wall baffle. The RCA 515S2 is a compound speaker having low and high frequency units with a very simple crossover network. With this speaker, the amplifier is stable without the 0.1 $\mu$f-100 ohm combination connected between the two collectors. The GE S1201D-7 speaker is a single cone unit and requires the 0.1 $\mu$f-100 ohm combination for high-frequency stability. If the 0.1 $\mu$f-100 ohm combination can be removed, additional high frequency power is available for the load. All tests were made with this combination connected.

The harmonic distortion of the amplifier is shown in Fig. 5 for various frequencies and various values of power. The apparent increase in distortion at very low power levels is part fact and part fiction. The fictional part is due to the fact that the harmonic distortion meter cannot distinguish between residual hum and noise and the harmonic distortion due to the application of a signal. The residual hum and noise is 74 db below 32 watts. For the 0.5 watt output condition this corresponds to 0.16 per cent which must be combined with

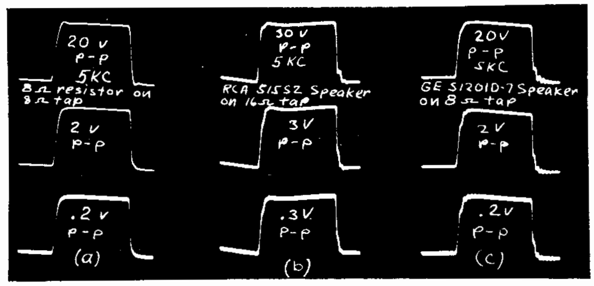



Fig. 4—5-kc square wave response.

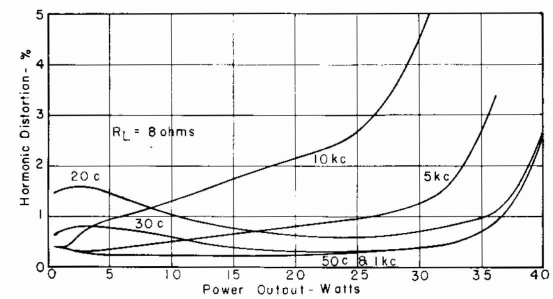



Fig. 5—Harmonic distortion characteristics.

the 0.12 per cent distortion in the signal generator. The measured distortion for this condition is 0.40 per cent for frequencies of 50 cycles and above. The factual part of the distortion increase can be attributed to the fact that for low signal levels, with class *B* operation, the curved lower portion of the transfer characteristic is used. This results in both higher inherent distortion and lower loop gain. The lower loop gain in turn reduces the effectiveness of the over-all feedback. In any case the distortion is low enough to classify this as a good 30 watt amplifier.

The curves of intermodulation distortion, shown in Fig. 6 for 80 per cent 70 cycle signal and 20 per cent 5 kc signal, confirm the fact this is an excellent 30 watt amplifier. It should be noted that the intermodulation distortion is plotted as a per cent of the smaller of the two signals.

The output impedance of the amplifier is shown as a function of frequency in Fig. 7. Over most of the frequency range the output impedance is roughly one fourth of the nominal tap impedance. As explained in a former paper,[1] the nominal tap impedance is not determined by maximum power considerations but by power transistor dissipation, the power capability of the driver transistor, and the rectifier supply current and voltage rating. Load impedances other than the nominal values may be used on any of the taps within the other limitations specified above.

The feedback circuit elements were worked out empirically for one channel, and the same nominal values were then applied to the other channel. The only exception was the value of $C_{fb}$ which was carefully adjusted for best 5 kc square wave response. The value of $C_{fb}$ was

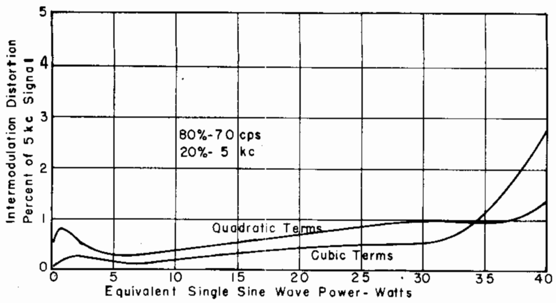



Fig. 6—Intermodulation distortion.

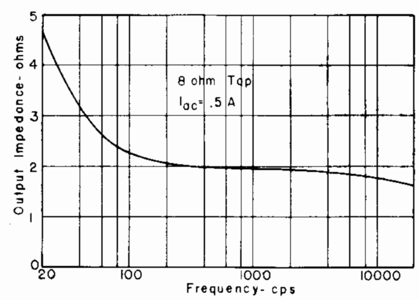



Fig. 7—Output impedance characteristic.

20 pf higher for the second channel than for the first one. A spot check was run on all performance characteristics with substantial agreement between the two channels.

## Thermal Design

In order to achieve good electrical performance with a transistor amplifier, strong consideration must be given to its thermal design. For ideal class $B$ operation, maximum device dissipation occurs when the peak signal is 63.6 per cent of the supply voltage. At this condition the dissipation per device is 0.1015 $V^2_{CC}/R_L$. For a supply or voltage of 36 volts and an equivalent load resistance of 16 ohms, this corresponds to 8.25 watts per device.

Ideal class $B$ operation is too severe for small signal operation so that a small amount of positive bias is generally necessary. If the zero signal current is 100 ma, maximum collector dissipation could easily be 10 to 12 watts per transistor. This power must be dissipated without permitting the collector temperature to exceed 100°C and preferably without adding to the temperature of the lower level transistors and the power rectifiers.

A 17×10×4-inch 14 Ga aluminum chassis was used to investigate the temperature rise with various transistor mounting schemes. Silicone grease was used at all thermal contacts and the power dissipation per unit was 10 watts.

1) The four transistors were uniformly spaced on a 4×17-inch side of the chassis and electrically insulated but thermally coupled to the chassis with anodized aluminum wafers. The temperature rise at the transistor base was 29°C while the coolest spot on the chassis was 10°C above ambient.

2) Each of the four transistors was attached directly to a 3×4×3/16-inch aluminum plate and the aluminum plates were mounted on one 17×4-inch side of the chassis. A sheet of 0.004-inch bond paper was placed between the aluminum plates and the chassis to supply thermal coupling but electrical insulation. The temperature rise at the transistor base was 25°C while the coolest portion of the chassis was 8.5°C above ambient.

3) Each of the transistors was mounted directly on a Delco 7270725 heat sink. Two of the heat sinks were attached with nylon insulators to each of the 10×4-inch sides of the chassis with 1-inch clearance between the chassis bottom and the heat sink bottom to permit free circulation of air. The temperature rise at the transistor base was 22°C and the coolest portion of the chassis was 1.5°C above ambient.

4) The heat sinks of part 3, with transistors attached directly to them were attached back to back, two in a row, and mounted vertically $1\frac{1}{4}$ inches above the 10×17-inch side of the chassis. The heat sinks were insulated from each other with nylon insulators. The temperature rise of the transistor base was 22°C as before but there was no noticeable temperature rise on any of the 4-inch sides of the chassis.

This last method was considered to be the best thermal design since it kept the transistor base temperature as low as any of the other methods and did not produce any chassis temperature rise to interfere with the heat sinking of the driver transistor and the current rating of the power rectifiers. A 14×10×3-inch 16 Ga aluminum chassis was now adequate since it had sufficient surface area for components and was relieved of the need of acting as a power transistor heat sink.

Heat sinking of the 2N600 driver transistors is particularly important since they are being operated at about 0.5 watts collector dissipation. These transistors were stud mounted to $\frac{1}{8}$-inch brass plates that were cut and drilled to fit the anodized aluminum power transistor insulating wafers. The back of the brass plates were lapped with fine sand paper to eliminate high spots and burrs. Two of these assemblies were thermally coupled but electrically insulated from one of the 10×3-inch sides of the chassis with the anodized aluminum wafers.

The TI 494 transistor is silicon and dissipates only 6 mw so that it does not require heat sinking. The 1N2860 rectifiers operate within a chassis maintained essentially at room temperature and can therefore operate at full rating. These rectifiers along with the pilot lights and the 500 ohm bias and dropping resistors are mounted at the chassis end most distant from the 2N600 transistors to minimize their effect on the operation of these transistors.

The thermistors used for bias stabilization of the power transistors normally come as disks with wires soldered to the two sides of the disk. To obtain good thermal coupling between the thermistors and the power transistor heat sinks they were modified by removing one of the wires and soldering them flat to an $\frac{1}{8}$-inch brass plate similar to the one used for the 2N600 transistors. This operation was performed on a hot plate that just barely brought the brass plates to solder melting temperature and was then turned off to avoid over heating the thermistors. Each of these brass plates was then thermally coupled but electrically insulated from the corresponding heat sink by an anodized aluminum wafer. Excellent temperature control was obtained in this manner.

## Transistor Complement

It was mentioned previously that the transistor complement could contain either *p-n-p* or *n-p-n* transistors in the output stage. From a practical point of view, however, only the *p-n-p* transistors need to be considered since *n-p-n* transistors of this dissipation rating are available only in silicon and are quite expensive. This dictates that the driver transistor be of the *p-n-p* type and the low level transistor be of the *n-p-n* type.

Since the low level and driver transistors are physically much smaller than the output transistors, it is to be expected that types will be available which will have considerably better frequency response than that of the output transistor. This permits the high-frequency response to be limited mainly by the output transistors and leads to a more gradual reduction in response and variation in phase than if several stages were contributing to the drop off simultaneously.

The original amplifier[1] used Delco 2N174 transistors with a beta cutoff frequency of 7 kc. This is still a very fine transistor for the purpose. All recorded tests on the present amplifier were made with Bendix 2N1073A transistors. The second channel had one 2N1073A and one 2N1073, selected for high breakdown voltage, which had reasonably well matched transfer characteristics. Matching of the transfer characteristics serves to keep down the distortion at low signal levels. On the other hand the use of the 0.27 ohm emitter resistors serves to reduce the need for transistor matching. A judicious balance between these two conditions must be achieved.

The manufacturer specifies an alpha cutoff frequency of 1.5 mc and a low current beta of 50 for the 2N1073A transistor. This corresponds to a beta cutoff frequency of about 30 kc. Additional transistors that might be considered for the output stage are the RCA 2N1906 and the TI 2N1046 series.

A Sylvania 2N68 transistor with a beta cut off frequency of 12 kc was used in the original amplifier.[1] The Philco 2N600 transistor, with a beta cutoff frequency of about 100 kc is a much better choice at the present time. Very few other transistors will satisfy the driver requirements as well as the 2N600.

Any high frequency *n-p-n* transistor will satisfy the requirements of the low level stage. The relatively inexpensive TI 494 transistor, with an alpha cutoff frequency of 20 Mc, satisfies these requirements admirably and removes the need to design additional temperature stability into this stage.

## Iron Core Components

The Stancor C2685 filter chokes used in the power supply are the only iron core components available commercially at the present time. Some transformers relatively close to the power transformer may be found commercially, but the driver and output transformers will have to be fabricated. The windings involved are relatively simple and should not cause an undue amount of trouble to anyone really interested in completing the construction of this amplifier.

1) The power transformer uses a $1\frac{3}{4}$-inch stack of EI-$1\frac{3}{8}$-inch laminations. The type of laminations used is not critical. The primary contains 310 turns of No. 20 HF wire while the secondary contains 114 turns of No. 17 FV wire. The laminations on this transformer may be interleaved three at a time if desired.

2) The output transformer uses a $1\frac{1}{2}$-inch stack of EI-$1\frac{1}{8}$-inch grain oriented laminations with Mil-T holes. The winding consists of 212 bifilar turns of No. 20 HF wire with taps at 106 and 150 turns on one file only. This coil must be tightly wound to fit in the window space available. This transformer uses 103 lamination sets which must be completely interleaved.

3) The driver transformer uses a $\frac{7}{8}$-inch stack of EI-$\frac{7}{8}$-inch laminations. The laminations of this transformer must not be interleaved. The E sections should be butted against the I sections without any additional spacers. The type of laminations used here is not critical either. All of the wire used on this transformer is No. 24 FV. The primary consists of two 250 turn sections connected in series. The secondary consists of 90 bifilar turns sandwiched between the two sections of primary. The end of one of the bifilar secondaries is connected to the start of the other one to form the center tap.

For transformer winding hints and a comparison of iron characteristics the reader is referred to an earlier paper.[4] The *H* holes described in that paper correspond to what is now called a Mil-T hole.

## Chassis Layout

The chassis top layout is shown in Fig. 8. The power transformer was the only iron core component mounted on top of the chassis. This was necessary because this transformer was too large to fit within the 3-inch chassis depth. The chassis space around the heat sinks is relatively clear in order to provide free flow of cooling air.

[4] A. B. Bereskin, "Build it yourself," IRE Student Quarterly, pp. 15–37; September 1956.

The heat sinks themselves are $1\frac{1}{4}$ inches above the chassis to permit air to get into the "smokestack" region between the two back to back transistors. The heat sink tops, which are the tallest components on the chassis, are $4\frac{1}{4}$ inches above the surface of the chassis. The input potentiometers were intended only for channel balancing purposes and were therefore placed relatively inaccessibly below the right hand heat sinks close to the input phono jacks.

The chassis bottom layout is shown in Fig. 9 in spread-out fashion. A considerable amount of equipment has been assembled in a relatively small space but all components are accessible for ease of servicing. Transistors TR1 are not shown on this diagram since they are soldered directly to the terminal strip on the far right.

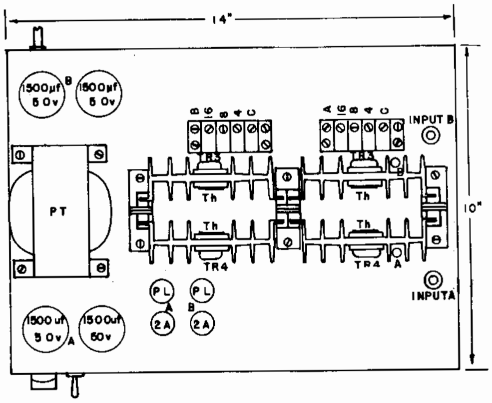



Fig. 8—Chassis top layout.

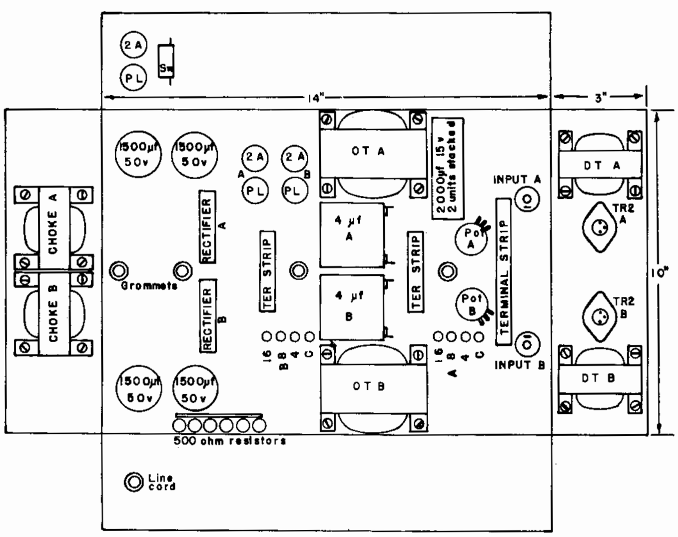



Fig. 9—Spread out view of chassis bottom layout.

It will be noticed that the driver transformer cores have been placed at right angles to all other cores and relatively distant from each other to avoid hum pickup and crosstalk. The residual hum of the amplifier is 74 db below 32 watts while the crosstalk is 70 db below 32 watts.

## Initial Adjustment

When this amplifier is first turned on, the ac input should be controlled with a Variac and brought up to operating level gradually while the current in the output stage is observed. In order to avoid excessive voltage on TR2, the resistor $R_B$ should be initially set at 1 megohm and should then be increased until the emitter of TR2 is at 11 volts when the dc supply is 36 volts. The output transistor bias and balance can be checked by measuring the voltage across the 0.27 ohm resistors. This voltage should correspond approximately to 100-ma current flow for both transistors. If substatial unbalance exists, the 500 ohm bias resistors should be modified slightly or different transistors should be used. The use of transistors other than those specified may lead to substantially different values of bias and feedback components.

Reducing the input coupling capacitor below 4 $\mu$f is not recommended since it leads to low frequency instability when the input potentiometer is set at zero. If the input capacitor is of the metal can variety, its capacitance to the can will be about 100 pf, and then the 100 pf input circuit capacitor may be omitted. If paper tubulars are used, the 100 pf capacitor must be used.

## Summary

The amplifier described in this paper has been employed in a home stereo system in conjunction with a transistor phono preamplifier[5] developed previously. The two units are compatible and provide excellent reproduction of the recorded signal. The performance characteristics presented in this paper amply testify to the quality of the amplifier. A photograph of the complete amplifier, which weighs 28 pounds, is shown in Fig. 10.

## Where to Buy

The net prices shown in the "List of Material" were obtained from a late catalog but may be subject to some variation depending on where the material is purchased. Most of these items are available from your local radio parts or electrical supply distributor. If you live in an area that does not have these distributors the parts may be purchased from one of the several mail order houses that operate on a national scale.

The iron core material may be purchased from some local transformer manufacturer or from Thomas and Skinner, Inc., 1120 East 23 St., Indianapolis 7, Ind. The Thomas and Skinner price list specifies a minimum order charge of \$10.00 and a minimum item charge for two or more items of \$5.00 per item, so that it would be desirable for two or more people to pool their lamination orders to meet these minimums.

[5] A. B. Bereskin, "A transistorized stereo preamplifier and tone control for magnetic cartridges," IRE Trans. on Audio, vol. AU-8, pp. 17–20; January–February, 1960.

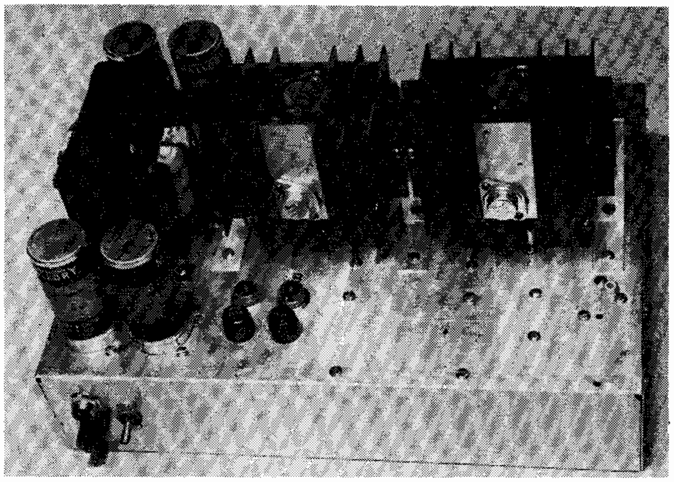


Fig. 10—Complete amplifier.

The Fenwell, ZB05J1 thermistor has been discontinued since this paper was written. The Fenwell KDO5L1 thermistor is available and will work equally well. Stocks of the ZB05J1 may still be available at some of the parts distributors.

LIST OF MATERIAL

| Quantity | Item | Total Net Cost |
|---|---|---|
| 1 | 14×10×3-inch 16 Ga aluminum chassis | $ 2.78 |
| 4 | 2N1073A Bendix transistors | 29.40 |
| 2 | 2N600 Philco transistors | 6.90 |
| 2 | TI 494 Texas Instrument Transistors | 5.50 |
| 8 | 1N2860 RCA Silicon Rectifiers | 5.92 |
| 4 | 2B05J1 or KD05L1 Fenwall Thermistors | 5.00 |
| 4 | Transistor mounting kits with anodized aluminum washers | 1.80 |
| 4 | No. 7270725 Delco Heat Sinks | 3.00 |
| 8 | No. 7269634 Delco insulating spacers for heat sinks | .40 |
| 4 | 1500 $\mu$f-50 volt electrolytic capacitors | 12.08 |
| 2 | 2000 $\mu$f-15 volt electrolytic capacitors | 4.30 |
| 2 | 25 $\mu$f-12 volt electrolytic capacitors | 1.88 |
| 2 | 4 $\mu$f-150 volt metallized paper capacitors | 7.74 |
| 2 | .1 $\mu$f-200 volt paper capacitors | .48 |
| 3 | Fuse mounts | 1.23 |
| 3 | Pilot light assemblies | 2.13 |
| 3 | 1829 pilot lights | .95 |
| 1 | SPST power switch | .52 |
| 1 | Line cord | .21 |
| 2 | Phono jacks | .32 |
| 2 | 50 K linear potentiometers | 1.82 |
| 2 | 4 screw-terminal binding posts | .68 |
| 6 | 500 ohm-10 watt wire wound resistors | 3,36 |
| 2 | 100 ohm-10 watt wire wound resistors | 1.12 |
| 1 | Tube Transistor Silicone Compound | 1.47 |
| | Assorted resistors, capacitors, grommets, terminal strips, and hardware | |
| 2 | C2685 Stancor Filter Chokes | 6.96 |
| $5\frac{1}{2}$ lb. | EI-$1\frac{3}{8}$-inch HS Radio 6AAS M-15 Laminations (26 Ga) | 1.76 |
| 7 lb. | EI-$1\frac{1}{8}$-inch Mil-T hole Orthosil 3XAAS M-6 Laminations (29 Ga) | 3.14 |
| $2\frac{1}{2}$ lb. | EI-$\frac{1}{6}$-inch HS Radio 4AAS M-19 Laminations (26 Ga) | .78 |
| 2 lb. | No. 20 HF wire | 2.40 |
| 1 lb. | No. 17 FV or HF wire | 1.14 |
| 1 lb. | No. 24 FV wire | 1.34 |

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