Ultra-low Distortion Class-A Amplifier

A design using feedback to control the gain and the levels of voltage and current in the output stage

By L. Nelson-Jones, M.I.E.R.E.

There is in the design to be described nothing very revolutionary, but rather an attempt to get a little nearer to perfection, in the power amplifier section of an audio system. Like Mr. Linsley Hood[1], the author has long felt that the slight extra cost and power consumption that class-A implies, is well worth while, and that its advantages are not as marginal as has often been supposed.

The most often quoted advantage of class-A operation is the elimination of crossover distortion, but there are other factors other than this which give rise to distortion in a class-B stage, especially at the upper frequency limit of the audio range, among them hole storage and inequality of high frequency performance of the two halves of the output stage.

Circuit design

The perfect power amplifier will convert its input signal to a higher power level, which is an exact replica of the input. It will have zero output impedance, but will not be damaged by a short circuit of its output terminals. It will have a flat gain-frequency response over the whole of the audio band, but will not respond to frequencies greatly outside this band. It will give its full rated power over the whole audio band. It should preferably drive capacitive loads, so that it may be used with an electrostatic speaker. It should be driven from a signal source whose bandwidth does not exceed that of the power amplifier, so that on transients in particular the power amplifier is not required to produce an output in excess of its capabilities.

No mention has been made of the input impedance of such an amplifier, this is because whilst some prefer a voltage input (high impedance), others prefer a current input (low impedance), and there is in any case no magic in this aspect. The degree of input impedance only decides the design of the output stage of the pre-amplifier, and to some extent alters the problems of stray couplings in the leads between these two sections. With low impedance, hum pick-up is most likely to be due to magnetic induction in the wiring, whilst with high impedance, it will more likely be due to electrostatic causes. The author's preference is for a high input impedance, mainly because he has more experience with such circuits, and in addition most signal sources and test equipment are rated for voltage output rather than current.

Now to the actual design, and firstly to underline what J. L. Linsley Hood said in a recent article[1] – "…. the basic linearity of the amplifier should be good, even in the absence of feedback" so that the feedback is used to obtain the desirable attributes of a good amplifier and not to overcome the shortcomings of a poor design.

Output stage

The use of the simplest circuit is very desirable, if only because it reduces the number of components which can cause phase shift at the higher frequencies, with consequent difficulty in stabilization of the overall loop. In this respect Linsley Hood's circuit[1] is excellent, but the author has found that despite its good performance, the need to select the resistors in certain parts of this amplifier and its reliance on the stability of current gain of the output transistors to set the operating current, went very much "against the grain" after years of designing equipment for production runs.

In order to get a more acceptable overall loop gain, it was decided to use transistor pairs for both halves of the output stage, with the result that higher values of resistor may be used in the driver stage. Fig. 1 illustrates three possible output stages considered.

Fig. 1(a) uses complementary transistors and is truly symmetrical, but is not as efficient as that of (b) which has a lower saturation voltage for each half as well as local feedback through the common emitter resistor of the first pair of transistors. Fig. 1(c) is the commonly used quasi-complementary type of output stage, which is in effect one half of Fig. 1(a), together with half of Fig. 1(b). Using this arrangement it is necessary for the best results to include a diode in the emitter of the lower p-n-p transistor so that looking into the base of each half of the output stage the driving source sees two forward biased junctions having fairly equal transfer characteristics for each half. The use of such a diode is particularly necessary in class-B stages as discussed in a recent article[2] and a letter. The design described here uses the circuit of Fig. 1(c) mainly because of the better availability of n-p-n power devices.

In the three output stages of Fig. 1 box X is the source of bias for the output stage. To ensure true class-A operation, with repeatability of operation from one amplifier to another, it was decided to use feedback to control the operating current. To achieve this the circuit of Fig. 2 was evolved. It will be seen that two additional transistors Tr7, and Tr8 have been added, together with a current sensing resistor R11. The action of the circuit is to hold the current through the output pair such that the drop across R11 is equal to the forward bias requirements of Tr8 (approximately 500 mV). Any increase in the output stage current will cause Tr8 to pass a greater current, which in turn will increase the conduction of Tr7, thus reducing the potential difference between the bases of Tr3 and Tr5, i.e. the bias of the output stage, and hence reducing the current in this stage. The input to Tr8 is filtered to remove audio components, so that the control circuit establishes the correct mean current irrespective of the signal present. The RC filter used for this purpose (R10 C6) must have values such that adequate filtering is achieved, yet the drop in R10 must not be large or the current level of the output stage will vary with the current gain of Tr8. This effect can be minimized by the use of a high gain transistor for Tr8. The capacitor C6 will be operated with only 500 mV polarization, which is insufficient to maintain the characteristics of a normal aluminium electrolytic. To overcome this problem a "solid" tantalum capacitor is specified, whose dielectric film of tantalum pentoxide is permanent. "Solid" aluminium capacitors also exist such as Mullard C415 and C121. These are not to be confused with "dry" electrolytics, which are wet types with the electrolyte in the form of a paste, (as are almost all aluminium electrolytics currently in use).

The operation of the output stage, with the bias network included, is at first hard to understand, since it at first appears that the drive to the base of Tr3 is reduced by the presence of Tr7, whose collector-emitter impedance is fairly high. This reasoning ignores the effect of C3 and C5, which results in the drives to the bases of Tr3 and Tr5 being almost equal. At low frequencies the circuit works well without C5, but with increasing frequency, phase shift in the power stage results in slight side effects which can be removed by the use of C5. By connecting the capacitor between the base and collector of Tr7 its effective value as seen between the emitter and collector of Tr7 is multiplied by the gain of this transistor, and thus a value of 0.22 uF proved quite adequate. Alternatively to revert to a more conventional circuit Tr7 could by bypassed by a normal 250 uF 6V capacitor as shown dotted in Fig. 2, to ensure equal drive to both halves of the output stage, at all audio frequencies.

Input and driver stages

These follow the well known arrangement of p-n-p input stage, with n-p-n driver stage. The feedback is arranged to be 100% at d.c. by connecting the 3.3 kohm feedback resistor (Fig. 3) direct to the emitter of Tr1. This feedback is reduced at audio frequencies by the attenuator formed by the 3.3 kohm and 220 ohm resistors, but not at d.c. because of the 250 uF blocking capacitor.

The action of the d.c. feedback is to keep the midpoint of the output stage at a potential equal to the voltage at the base of Tr1 plus the base-emitter potential of Tr1 and the voltage drop in the feedback resistor (approximately 300 mV). Slight adjustment of the voltage of the bias chain feeding the base of Tr1 allows the mid-point of the output stage to be set for symmetrical clipping at the onset of overload. The mid-point level will vary slightly with temperature due to the 2 mV/ºC change in Vbe of Tr1, but this will be added to the effect of increase of current gain in the two input transistors, resulting in a drop in the collector current of Tr1 and hence a drop in the potential across the 3.3 kohm feedback resistor. However the total change over the range 0-40ºC is only some 200 mV, and is thus of little consequence, in relation to the level of 14 V.

Power supply

In order to ensure the greatest possible freedom from hum and similar problems it was decided that the extra cost of a fully regulated power supply was justified, in relation to the high performance being aimed at.

The series stabilizer is quite conventional except for the generation of the pre-regulator supply (+60 V). This supply is generated by a Cockroft voltage-doubler circuit which is connected to the main rectified supply, so that the outputs of both circuits add. The input (peak) voltage to the voltage doubler is only half that across the main bridge rectifier, since on negative half cycles, the arm of the bridge between the input to the voltage doubler and the 0 V line, is conducting, clamping the point near 0 V, whilst on positive half cycles it is non-conducting allowing this point to rise. The connection of the anode of D2 to the main rectified supply has the effect of increasing the voltage across the two capacitors by the voltage of the main supply; but does not affect the a.c. conditions in the circuit.

The main supply is a normal bridge rectifier with capacitance smoothing. The value of this capacitor is decided by the maximum permissible ripple, which in turn depends on the minimum mains voltage allowable and the minimum voltage across the regulator series transistors at which the regulator still retains full control.

The actual pre-regulator supply generated by the voltage-doubler circuit is used to supply a zener diode (6.8 V) connected to the regulated supply, thus making a d.c.-coupled bootstrap connection for the collector load of the amplifying stage of the regulator (Tr9), and giving a considerable increase in gain, within the regulator loop. The loop is stabilized by the 1200 pF capacitor across the base and collector of Tr9, and the output impedance rise that this causes at the higher frequencies, is removed by the connection of the 1250 uF capacitor across the regulated line, in accordance with normal practice in such regulators.

The performance of this regulator is excellent and the only additional smoothing needed is the 10 uF capacitor in the base bias network of Tr1. An output for the pre-amplifier and tuner etc. is available (via a low value decoupling resistor and a 1250 uF capacitor) at the input plug.

Overload protection

This is inherent in the action of the current control circuit, which prevents the output stage mean-current from varying. A full short-circuit can be sustained without damage. The current in the output stage remains correct as regards mean level but due to the high value of loop gain the current waveform becomes a square wave on heavy overload and as a consequence the dissipation in the current-sensing resistor doubles to approximately 1 W.

Frequency response

At low frequencies three capacitors determine the basic response. The input capacitor to the base of Tr1, the d.c. blocking capacitor of the feedback loop, in the emitter circuit of Tr1, and the capacitor feeding the load. The cut-off frequencies due to each alone, are 14, 3 and 8 Hz respectively. The combined effect was measured, and gave a "cut-off" at 15 Hz (-3 dB). In the author's opinion it is important that the main limitation of the bandwidth at low frequencies should be due to the input capacitor, so that the amplifier will not be overloaded by frequencies outside the useful audio-range. It is also important that the output capacitor is sufficiently large to allow the very low output impedance, obtained by high degrees of negative feedback, to damp the fundamental resonance of the loudspeaker cone. The values given are a good compromise, and provide an adequate bass response. For a lower cut-off, all three capacitors should be changed by the same factor.

No specific steps have been taken to limit the high-frequency response, which is found to be level to 15 kHz, - I dB at 54 kHz, and - 3 dB at 92kHz, above which it falls rapidly.

Noise and distortion

Clipping at the overload point is clean and symmetrical, as shown in Fig. 5(a) for a 1 kHz sinewave. The normal method of adjusting the bias of the amplifier is to adjust the "Set O/P Levels" control for symmetry of clipping, having previously set the supply regulator for a reading of + 28 V.

Distortion was measured - with some difficulty - at 1 kHz, when it was found that it was almost entirely 3rd harmonic in nature, and of very low level, only reaching 0.015% at the onset of clipping, so that at normal listening levels it would be quite insignificant.

Such a low level of distortion is not surprising when one considers the facts. The loop gain is measured as 4750 times, with the closed-loop figure of 16 times. The reduction in gain, and hence also in distortion, is therefore 297 times or -49.5 dB, implying a basic open-loop distortion of around 5%, a reasonable figure for a basically linear amplifier. The output of the amplifier operated under open-loop conditions at just under full output is shown in Fig. 5(b). The variation with output level of the distortion under closed-loop conditions is shown in the graph of Fig. 4(c).

Due to the use of a regulated supply the noise and hum levels are of a very low value. Hum components alone (50 and 100 Hz) are -83 dB relative to full output. Wideband noise, ignoring hum components, is approximately -100 dB below full output, rising very slightly if the input is open circuit. The result is a background level that is completely inaudible.

Response to square wave input, and to capacitive loads

The effect of capacitive loads is shown in Fig. 5(c) and 5(d). The capacitor was a 1 uF paper type, and little difference in waveform is noticeable, whether or not, the 8 ohm resistive load is connected in parallel. The ring frequency induced is at approximately 200 kHz for a 1 uF capacitor but reduces somewhat with larger values of capacitor. Fig. 5(e) shows the response to a steep input edge the total rise time is around 0.5 us, giving a slewing rate of 40 V/us. Fall time is similar.

Input impedance

Due to the high degree of series feedback employed, the input impedance is almost entirely that of the base bias network, i.e. the two 100 kohm resistors effectively in parallel. The value was measured and was found to be such, namely 50 kohm.

Current sensing resistor

It is desirable that this should be of a non-inductive type in order not to introduce high frequency effects, which might limit the available power at that end of the spectrum, and also cause stability problems in the loop. The requirement for a non-inductive resistor is more important in class B amplifiers, but is by no means unimportant in class A applications (see "Letters to the Editor", F. Butler and Arthur Bailey, Wireless World, December1966, pp. 611-614). The construction of the resistors used in the prototype is shown in Fig. 6. An alternative would be to use Eureka wire to connect the emitter of Tr4 to the remainder of the circuit, using a single straight length of a suitable gauge probably 26 s.w.g.). In this case the wire should be covered with high temperature sleeving, say silicone rubber, or glass fibre. The 1 kohm resistor feeding the base of Tr8 would then be connected direct to the emitter of Tr4.

Heatsinks

In the prototype, finned extruded aluminium heatsinks of approximately 4 in x 4 in are used for each of the output transistors. A similar heatsink is used for the series transistors of the regulator. In each case no insulation is used between the transistors and the heatsink, which is live to the collector in each case. This course of action was taken to maximize the efficiency of the heatsinks, and these must therefore be separately insulated from their mountings. The method used in the prototype is to cut slots in the edge of the heat sinks (0.25 in deep, 0.25 in wide), which then enable the heatsinks to be mounted on 4BA studding using Transiblocks, details of which are to be found in the constructional section below. Silicone grease is used to ensure a good thermal connection between the heat sink and the power transistors.

The amplifier must not be used in confined surroundings such that free air circulation is impeded, as some 60 W of heat have to be dissipated by the complete stack of heatsinks. The cabinet in which the amplifier is mounted should therefore be well ventilated, and in particular the author has found that a larger area of vent is required at the top of such a cabinet than at the bottom in order to stop the build up of a cushion of hot air at the top. The maximum rise in the centre of the heatsink stack, gives a case temperature for the power transistor which is approximately 40ºC above ambient. The junction temperature with the dissipation occurring in each transistor will be a further 20ºC higher in the worst case. Thus at 20ºC in free air the maximum junction temperature will be 80ºC, allowing a considerable amount of leeway for both raised ambient temperature and less than free air circulation. It is recommended that the maximum case temperature of the power transistors should not be allowed to exceed 100ºC in use, and in the cabinet in which it is to be mounted, so that a reasonable degree of reliability is achieved.

Adjustment of design for other than 8 ohm load

Referring to Fig. 2 again, we will first calculate the supply voltage required for any given load. (The number suffixes given refer to the transistor numbering in Fig. 2.)

Output voltage swing (pk-pk) = Vcc - { $V_{ce.sat3}$ + V_{be4} + $V_{ce.sat6}$ + (I + Î) * R11}

Also, power output (sinewave) = (output voltage swing)^2 pk-pk / (8 * Rload)

Since Vout (r.m.s.) = Vp-p / 2√2 (for a sinewave),

Vout (pk-pk) = √(8 * Rload * Pout) and therefore

Vcc = √(8 * Rload * Pout) + $V_{ce.sat3}$ + V_{be4} + $V_{ce.sat6}$ + (I + Î) * R11, minimum.

The standing current must exceed Vpk-pk / (4 * Rload) in order to achieve the required voltage swing, and for its satisfactory safety margin it should exceed Vcc / (4 * Rload).

Taking typical values for the circuit given using an 8 ohm load, and 10 W output level, we get:

Vcc = √640 + 0.25 + 1.0 + 0.5 + (0.90 + 0.79) * 0.56 = 28 V

Imin = 28 / (4 * 8) = 875 mA (a value of 900 mA being actually used.)

For a 3 ohm load and 10 W output we get figures of 19.5 V for Vcc, and 1.63 A for Imin. (Total power 31.8 W, 31.5% efficient).

For a 15 ohm load and 10 W output we get figures of 36V for Vcc, and 0.6 A for Imin. (Total power 21.5 W, 46.4% efficient).

From these figures it is apparent that the rise in Vce.sat and Vbe figures with the current used in a 3 ohm amplifier seriously reduces the overall efficiency. In the case of the 15 ohm load on the other hand, the efficiency is not far short of the theoretically possible figure of 50% for a class A stage. The efficiency of the 8 ohm stage is 39.8%.

Details of value changes for 3 ohm, and 15 ohm circuits are given with the constructional details below.

Performance of 8 ohm version

Output (at commencement of clipping)	10 W
Frequency response	36 Hz - 54 kHz (-1 dB)
	15 Hz - 92 kHz (-3 dB)
Power bandwidth	Full power 15 Hz - 30 kHz
	-3 dB (half power) at 60 kHz
Hum level	-83 dB relative to 10 W
Noise level	-100 dB relative to 10 W (ignoring hum components)
Rise time	0.5 us
Input impedance	50 kohm
Input sensitivity	0.56 V r.m.s. for 10 W (gain 16)
Open loop gain	4750
Feedback gain reduction	-49.5 dB (297 times)
Distortion	0.015% at 1 kHz, 10 W output (almost entirely 3rd harmonic)
	0.01% at 2.5 W
	0.005% at 350 mW
Channel separation	-43dB at 20 Hz rising to greater than
	-60 dB at 1 kHz and above

Construction details

Fig. 7 shows the construction of the underside of the chassis of the 10+10 W amplifier. The layout is shown in greater detail in the sketch of Fig. 8, the two amplifiers being constructed as mirror images, as can be seen in the photograph.

To avoid large circulating currents the loudspeaker return leads should be wired to the earth tags of their respective amplifiers, as shown in Fig. 8. The negative lead of the rectifier bridge should be connected to the same earth tag as the negative connection of the 5000 uF main smoothing capacitor, together with the negative connection of the second 50 uF smoothing capacitor of the voltage doubler.

Providing the layout given is followed, and the precautions listed over earth tags are followed, no problems should be encountered. Layout of the series regulator components is entirely non-critical and uses similar tag strips to those in the power amplifiers.

Fixed resistors – With the exception of the current sensing resistors R11, R11a and those marked with * in the circuit of Fig. 3, all resistors are solid carbon moulded 1/2 W 10%. All resistors marked * are 1/2 W 2% metal oxide (Electrosil TR5, Welwyn MR5, Radiospares "1/2 W oxide"). See Fig. 6 for details of the construction of R11.

Variable resistors – Both are wirewound Radiospares type "presets" (set +28 V and set output levels). Any good wirewound types such as those quoted of 1 W rating or above are suitable.

Non-electrolytic capacitors –

0.22 uF 160 V input capacitor Wima Tropyfol M (160 V) or Mullard C296AA/A220 K. Radiospares also make a suitable type 250 V PDC.

0.22 uF 20 V ceramic disc (base-collector Tr7). Radiospares 20 V discs, or use polyester 160 V type as above.

1200 pF tubular ceramic (1000 pF can be used). The capacitor used in the prototype is now obsolete; Radiospares suggest as alternatives "discs 0.001 uF" or "Hi-K 0.001 uF" (tubular).

0.1 uF 400 V (across bridge rectifier, necessary to prevent the generation of mains-borne interference due to hole storage effects in the rectifiers), Wima Tropyfol M(400 V), Mullard C296AC/A100K. Radio-spares 400 V PDC.

Electrolytic capacitors –

47 uF 6 V (base-emitter Tr8*).* This must be solid tantalum type. The Radiospares type used in the prototype is discontinued but is apparently identical to Union Carbide "Kemet E". Alternatives are S.T.C. 472/ LWA/401CA (metal case), S.T.C. TAG47/3 (3 V rating similar to Kemet E),. Mullard C421AM/BP47 (metal case), C415AP/C50 (50 uF, 6.4 V solid aluminium type).

10 uF 64 V (input bias chain) Mullard C426AR/H 10.

250 uF 25 V (feedback blocking capacitor) Mullard C437AR/F250.

250 uF 40 V (bootstrap capacitor) Mullard C437AR/G250.

1250 uF 40 V (across 28 V supplies) Mullard C431BR/G1250.

2500 uF 40 V (output capacitor) Mullard C431BR/G2500.

5000 uF 50 V (main smoothing) Daly type obtained from Electrovalue. Nearest Mullard type C432FR/G5600 (5600 uF 40 V).

50 uF 350 V (voltage doubler) Radiospares "tubes 50 uF 350 V". Alternative types of
not less than 100 V rating may be used.

Caution should be exercised in the selection of suitable types for the main smoothing capacitor because of the high ripple rating required. The Radiospares type "Cans 5000 uF 50 V" is not suitable on this account. The Daly type has a ripple rating of 4.3A.

Transformer – Radiospares "27 V rec trans" Prim. 0-100-115-205-225-245 V 50/60 Hz. Sec. 27 V at up to 3 A rectified d.c.

Fuse – 2A normal or 750 mA "anti-surge" delay type.

Semiconductors –

Tr1, Tr7	2N3702 (BCY7O)
Tr2, T9	BCI07 (BC 108 suitable for Tr9)
Tr3	2N2219
Tr5	2N2905
Tr4, Tr6, Tr11	2N3055
Tr10	2N3054
Tr8	BC168 (BCIO8)
D1	OA200 (HS1010, OA202)
D2, D3	RAS31OAF (Radiospares REC51A, 1N4005, BY103)
ZD1	ZF8.2 (Radiospares "MZ-E 8.2 V", Mullard BZY88-C8V2, Texas 1S2068A)
Rect. 1	Radiospares REC.40.5A bridge 200 V (p.i.v.)

Heatsinks –

Power transistors mounted on 5 Radiospares heatsinks, which are equivalent to "Marex" (Marston-Excelsior) type 10D - 4 in long. S.T.C. supply a similar type, code HSC4 and a clip for insulated mounting (but not as in photos) FP2551 (Electroniques). Heatsinks mounted on 4BA studding using four transiblocks per heatsink. Transiblocks are made by Industrial Instruments Ltd, Stanley Road, Bromley, Kent. Farnell Instruments Ltd (Industrial Supplies Division) also stock these items.

The TO-5 transistors (Tr3, Tr5), are fitted with cooling clips - Redpoint 5F, available from Electrovalue and Electroniques. A similar type - "Sinks TO-5" - is available from Radiospares.

Sundries –

Chassis size 7 in x 10 in x 2 in (sheet aluminium type).

The input socket is a 5 pin "DIN" audio connector. The loudspeaker sockets are Radiospares miniature non-reversible 2-way plugs, and sockets. Non-reversibility is essential to preserve the phasing of the outputs to the speakers. It is convenient to mount the fuseholder (Radiospares panel fuse holders or Belling-Lee L.1348, L.1382, L.1744) on a panel attached to the side of the mains transformer, with a strip on top of the transformer for connection of the mains lead, mains switch, etc., as shown in the photograph.

Modifications for 3 ohm output

R11 and R11a must be reduced to 0.31 ohm (5%) each. The mains transformer will require to be 21 V r.m.s. 3.5 A d.c. rectified rating. The output capacitor feeding the loudspeaker must be 5,000 uF 25 V. The 12 kohm resistor in the regulator will reduce to 7.5 kohm, and the 3.3 kohm resistor feeding the 6.8 V zener diode will reduce to 2-2 kohm. The main smoothing capacitor should be raised to 7,000 uF at not less than 30 V working. The collector resistors of Tr2 should be dropped from 820 ohm, 1.5 kohm and l.2 kohm to 470 ohm, 820 ohm and 680 ohm respectively.

Modifications for 15 ohm output

R11 and R11a, must be increased to 0.84 ohm (5%) each. The mains transformer must be 34 V r.m.s. 1.5 A d.c. rectified rating. The 12 kohm resistor in the regulator must be increased to 17 kohm which is not a standard value, alternatively the 4.7 kohm may be dropped to 3.6 kohm which is a standard value. The 3.3 kohm resistor feeding the 6.8 V zener diode should be raised to 3.9 kohm. The collector resistors of Tr2 may be raised if desired but this is not necessary. Tr9 must be BC107 since BCl08 has an inadequate voltage rating. Tr3 may be 2N2219A or 2218A which have a higher voltage rating than 2N2219. However if 2N2218 A is used then Tr5 should be changed to 2N2904, to preserve some equality of current gain. If a transistor tester is available then samples of 2N2219 may be selected for Vceo of above 40 V instead (normal minimum is 30 V).

It should be noted that the output to pre-amplifier and tuners will alter, being +19.5 V for the 3 ohm version, and +36 V for the 15 ohm version.

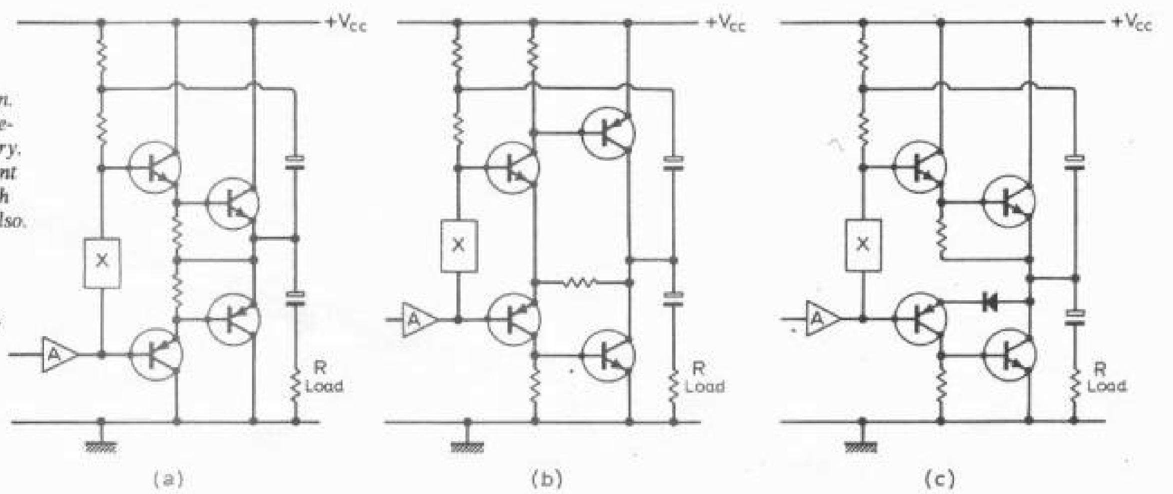
It is expected that the distortion of the 3 ohm version will be two to three times greater than that quoted for the 8 ohm version, with similar or slightly better figures for the 15 ohm version. In the author's opinion, since very few speakers deserving the title high-fidelity, have a 3 ohm voice coil, the 3 ohm version of the amplifier is not worth considering unless no other choice presents itself.

REFERENCES

1. J. L. Linsley-Hood, "Simple Class A Amplifier", Wireless World, April 1969.
2. I. M. Shaw, "Quasi-Complementary Output Stage Modification", Wireless World, June 1969.

Figures

Fig. 1. Possible output stages considered for class A operation. (a) Fully complementary symmetry. (b) More efficient arrangement with local feedback also. (c) Quasi-complementary output with equalizing diode.



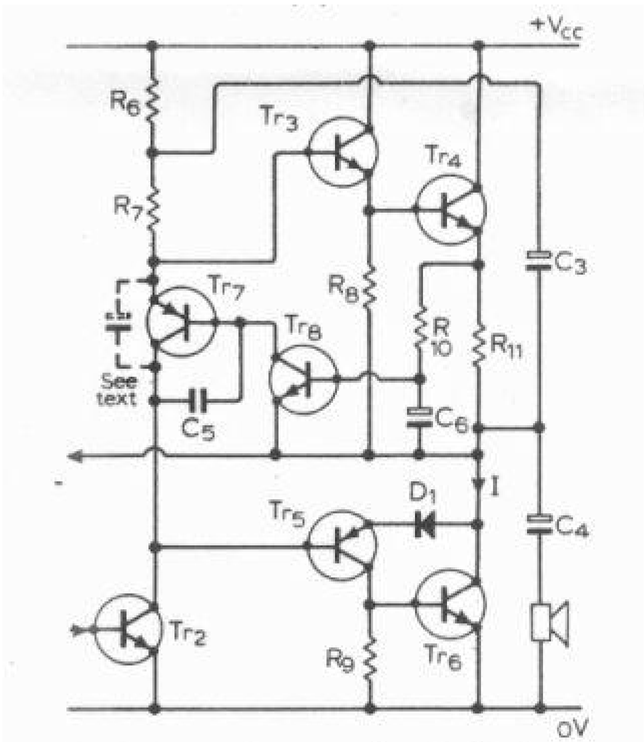


Fig. 2. Circuit chosen to allow feedback control of the operating current.

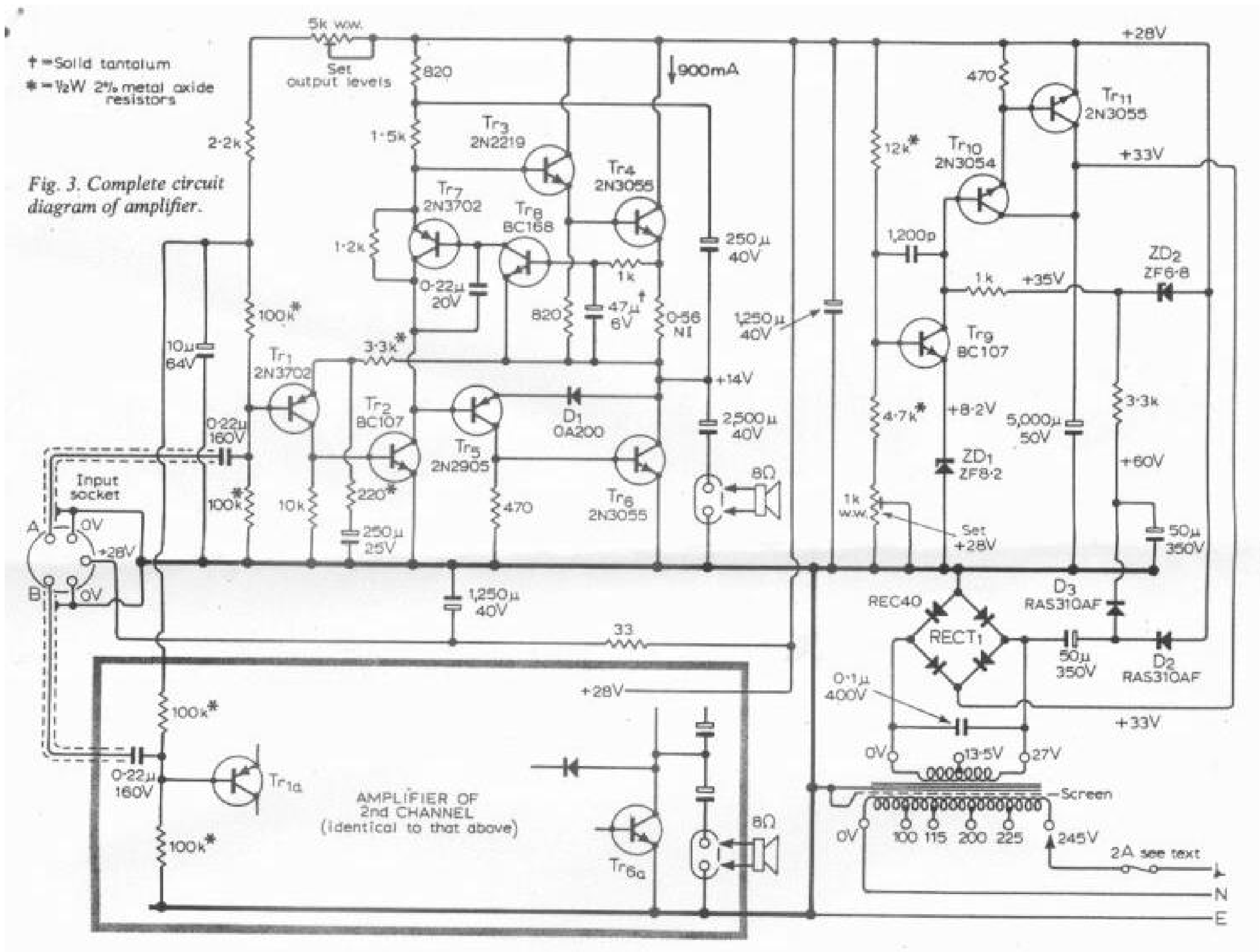


Fig. 3. Complete circuit diagram of amplifier.

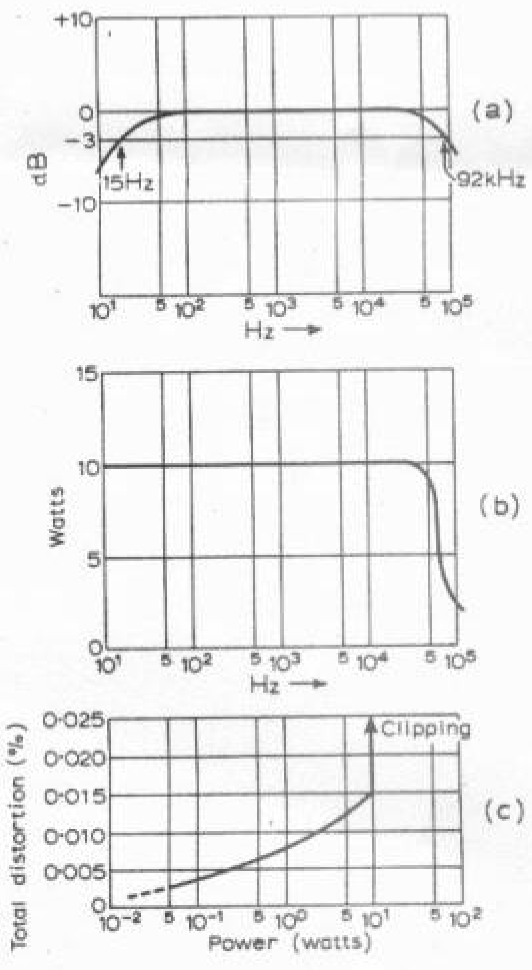


Fig. 4. Performance curves.

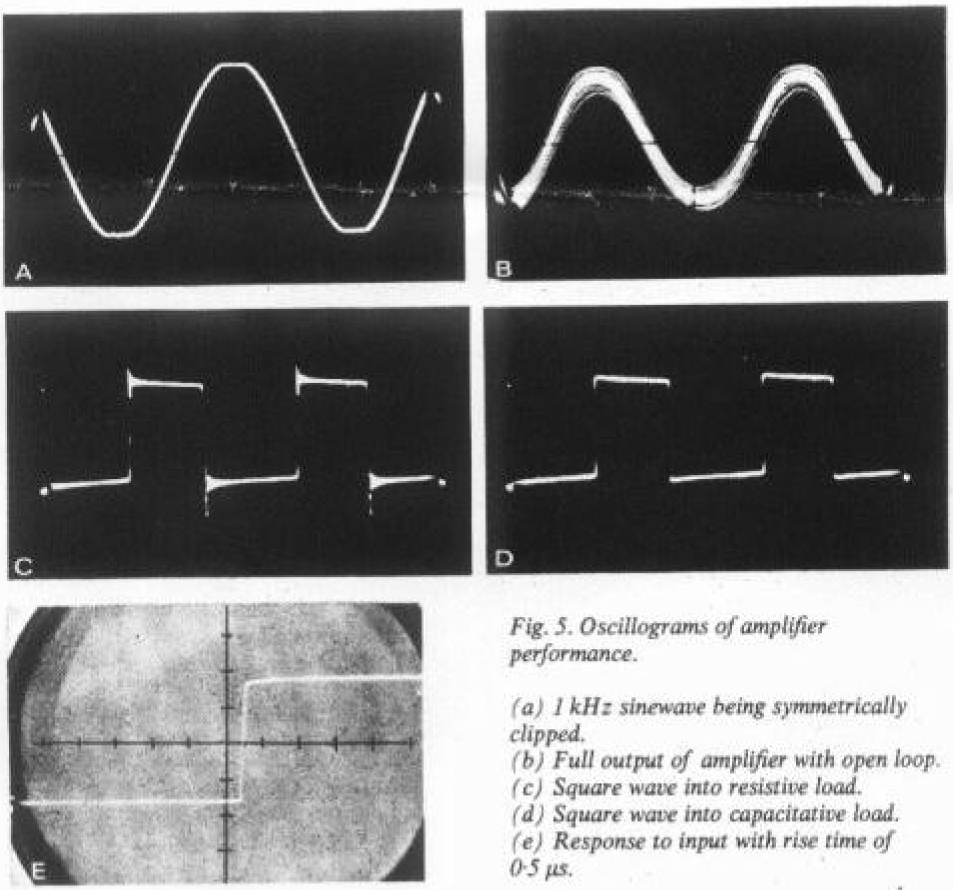


Fig. 5. Oscillograms of amplifier performance.

(a) 1 kHz sinewave being symmetrically clipped.
(b) Full output of amplifier with open loop.
(c) Square wave into resistive load.
(d) Square wave into capacitative load.
(e) Response to input with rise time of 0·5 μs.

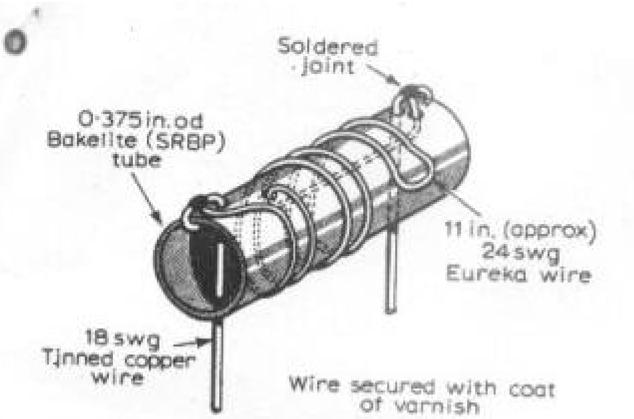


Fig. 6. Construction of 0·56 Ω 5% non-inductive resistors.

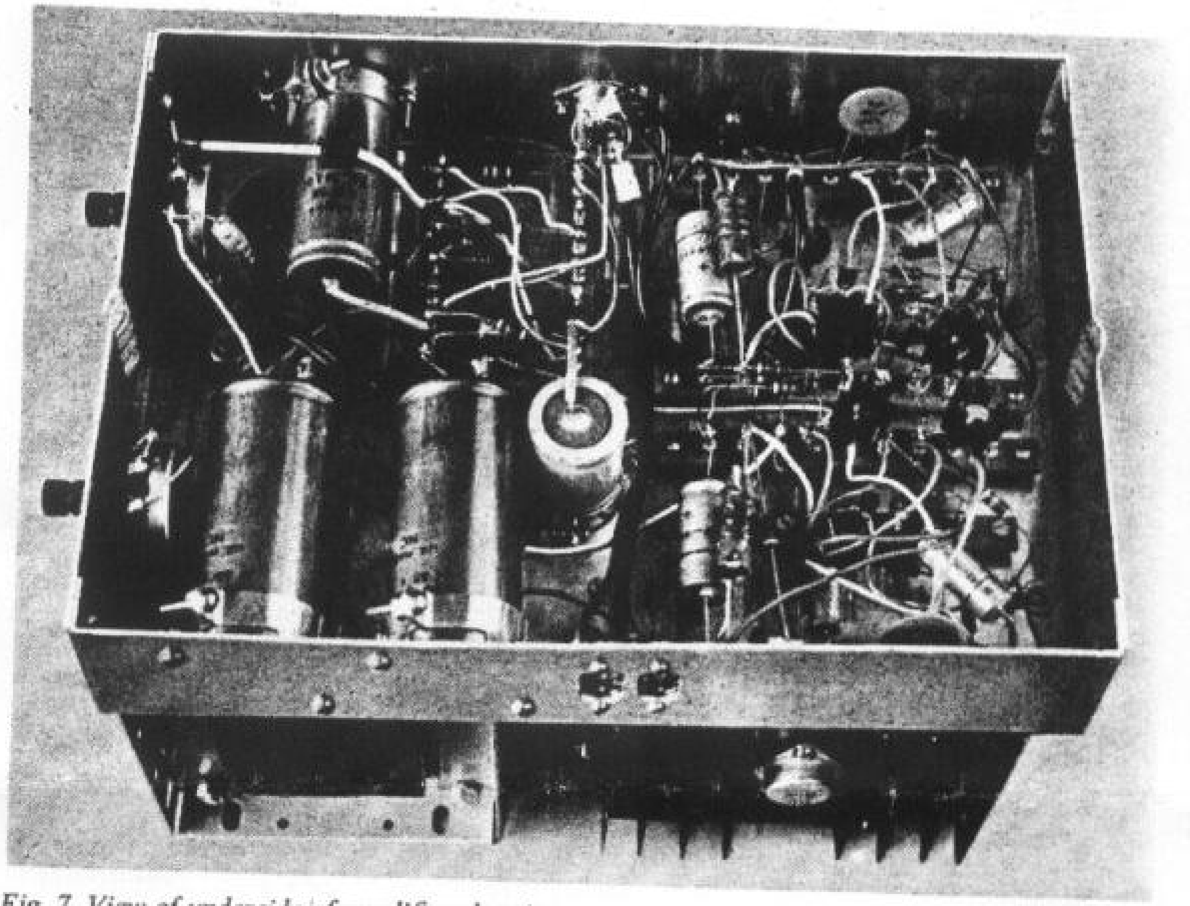


Fig. 7. View of underside of amplifier chassis.

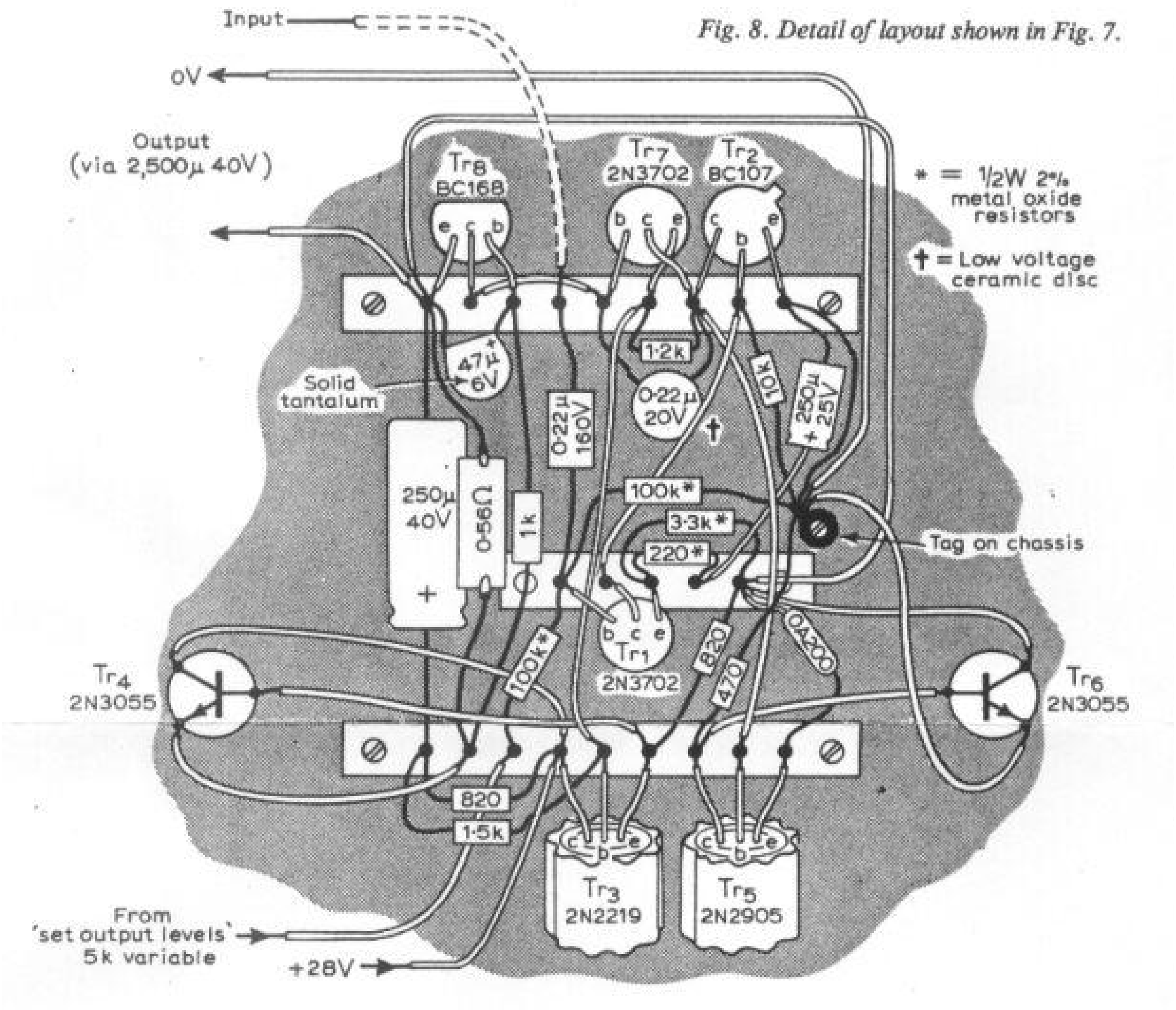


Fig. 8. Detail of layout shown in Fig. 7.