

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

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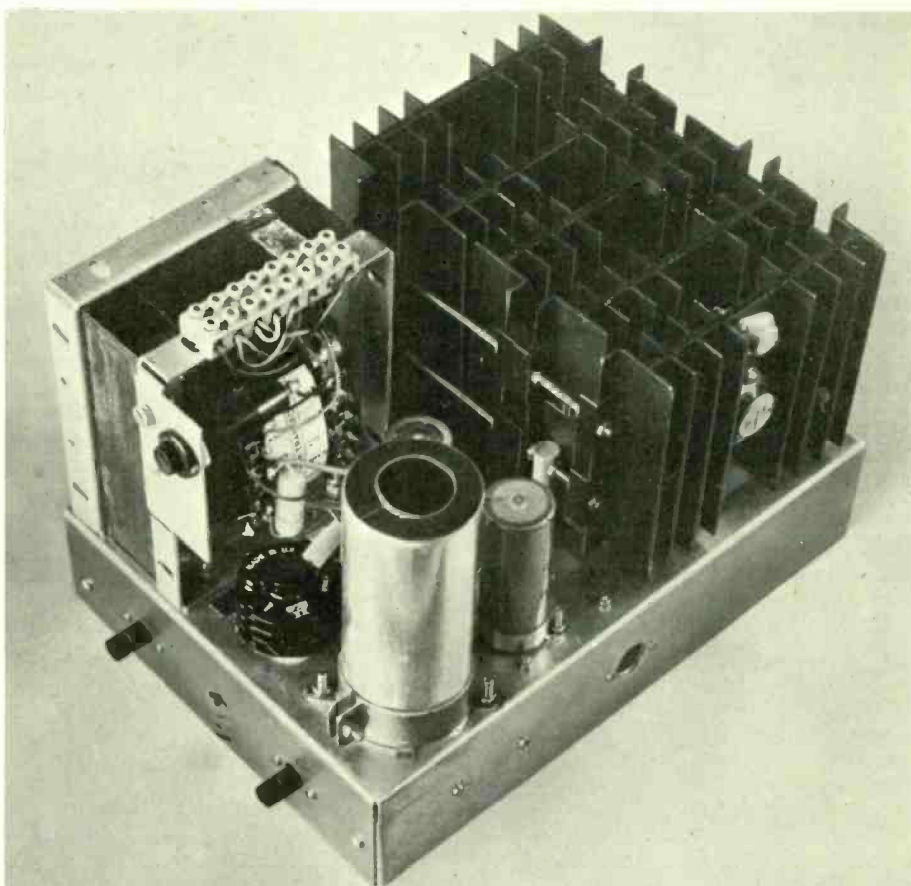
There is in the design to be described nothing very revolutionary, but rather an attempt to get a little nearer to perfection, in the power amplifier section of an audio system. Like Mr. Linsley Hood,¹ the author has long felt that the slight extra cost and power consumption that class A implies, is 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 cross-over 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 article¹ — "... 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.

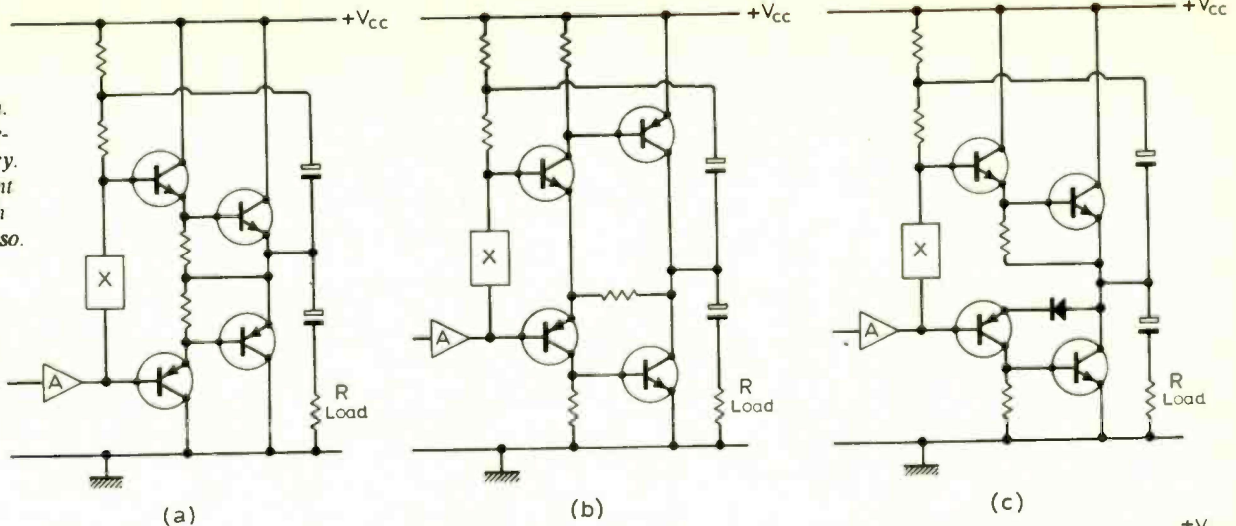
Output stage

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 circuit¹ 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

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.



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 article² and a letter³. 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 Tr_7 and Tr_8 have been added, together with a current sensing resistor R_{11} . The action of the circuit is to hold the current through the output pair such that the drop across R_{11} is equal to the forward bias requirements of Tr_8 (approximately 500 mV). Any increase in the output stage current will cause Tr_8 to pass a greater current, which in turn will increase the conduction of Tr_7 , thus reducing the potential difference between the bases of Tr_3 and Tr_5 , i.e. the bias of the output stage, and hence reducing the current in this stage. The input to Tr_8 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 ($R_{10} C_6$) must have values such that adequate filtering is achieved, yet the drop in R_{10} must not be large or the current level of the output stage will vary with the current gain of Tr_8 . This effect can be minimized by the use of a high gain transistor for Tr_8 . The capacitor C_6 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 Tr_3 is reduced by the presence of Tr_7 , whose collector-emitter impedance is fairly high. This reasoning ignores the effect of C_3 and C_5 , which results in the drives to the bases of Tr_3 and Tr_5 being almost equal. At low frequencies the circuit works well without C_5 , but with increasing frequency, phase shift in the power stage results in slight side effects which can be removed by the use of C_5 . By connecting the capacitor between the base and collector of Tr_7 its effective value as seen between the emitter and collector of Tr_7 is multiplied by the gain of this transistor, and thus a value of 0.22 μ F proved quite adequate. Alternatively to revert to a more conventional circuit Tr_7 could be bypassed by a normal 250 μ F 6 V 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 k Ω feedback resistor (Fig. 3) direct to the emitter of Tr_1 . This feedback is reduced at audio frequencies by the attenuator formed by the 3.3 k Ω and 220 Ω resistors, but not at d.c. because of the 250 μ F 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 Tr_1 plus the base-emitter potential of Tr_1 and the voltage drop in the feedback resistor (approximately 300 mV). Slight adjustment of the voltage of the bias chain feeding the base of Tr_1 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/ $^{\circ}$ C change in V_{be} of Tr_1 , 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 Tr_1 , and hence a drop in the potential across the 3.3 k Ω feedback resistor. However the total

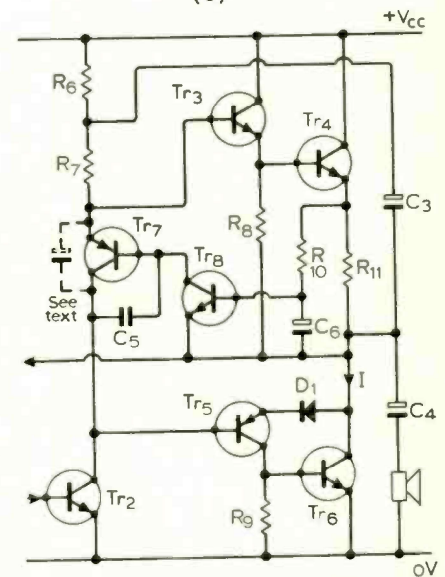


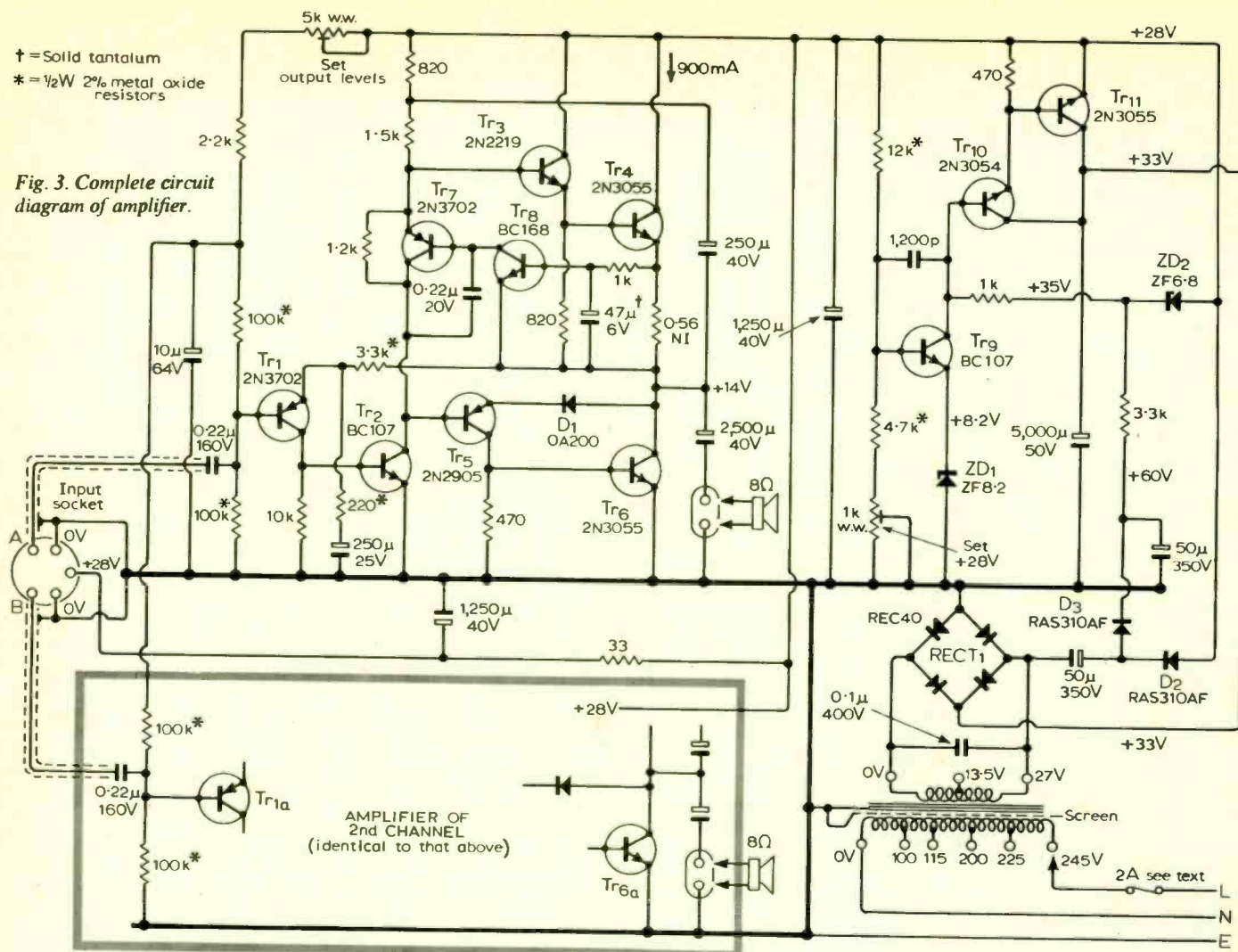
Fig. 2. Circuit chosen to allow feedback control of the operating current.

change over the range 0–40 $^{\circ}$ C 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 D_2 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 (Tr_9), 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 Tr_9 , and the output impedance rise that this causes at the higher frequencies, is removed by the connection of the 1250 μ F 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 μ F capacitor in the base bias network of Tr_1 . An output for the pre-amplifier and tuner etc. is available (via a low value decoupling resistor and a 1250 μ F capacitor) at the input plug.

Overload protection

This is inherent in the action of the current control circuit, which prevents the out-

put 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 Tr_1 , the d.c. blocking capacitor of the feedback loop, in the emitter circuit of Tr_1 , 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, -1 dB at 54 kHz, and -3 dB at 92 kHz, 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 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 μ F paper type, and little difference in waveform is noticeable, whether or not, the 8- Ω resistive load is connected in parallel. The ring frequency induced is at approximately 200 kHz for a 1- μ F 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 μ s, giving a slewing rate of 40 V/ μ s. 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-k Ω resistors effectively in parallel. The value was measured and was found to be such, namely 50 k Ω .

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*, December 1966, 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 Tr_4 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 k Ω resistor feeding the base of Tr_8 would then be connected direct to the emitter of Tr_4 .

Heatsinks

In the prototype, finned extruded aluminium heatsinks of approximately 4 in \times 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 heat

sinks. 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 heat sink stack, gives a case temperature for the power transistor which is approximately 40°C above ambient. The junction temperature with the dissipation occurring in each transistor will be a further 20°C higher in the worst case. Thus at 20°C in free air the maximum junction temperature will be 80°C, 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 100°C 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- Ω 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)

$$= V_{cc} - \{V_{ce \cdot sat3} + V_{be4} + V_{ce \cdot sat6} + (I + \bar{I}) R_{11}\}$$

Also, power output (sinewave)

$$= \frac{(\text{output voltage swing})^2 \text{ pk-pk}}{8R_{load}}$$

$$\text{Since } V_{out} \text{ (r.m.s.)} = \frac{V_{pk-pk}}{2\sqrt{2}}$$

(for a sinewave),

$$V_{out} \text{ (pk-pk)} = \sqrt{8R_{load} \cdot P_{out}}$$

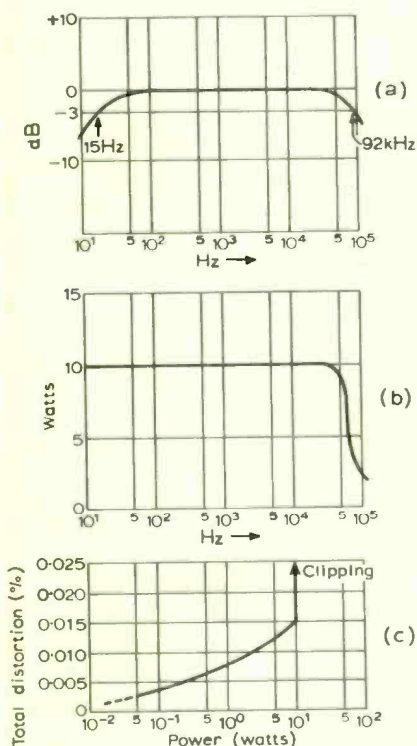


Fig. 4. Performance curves.

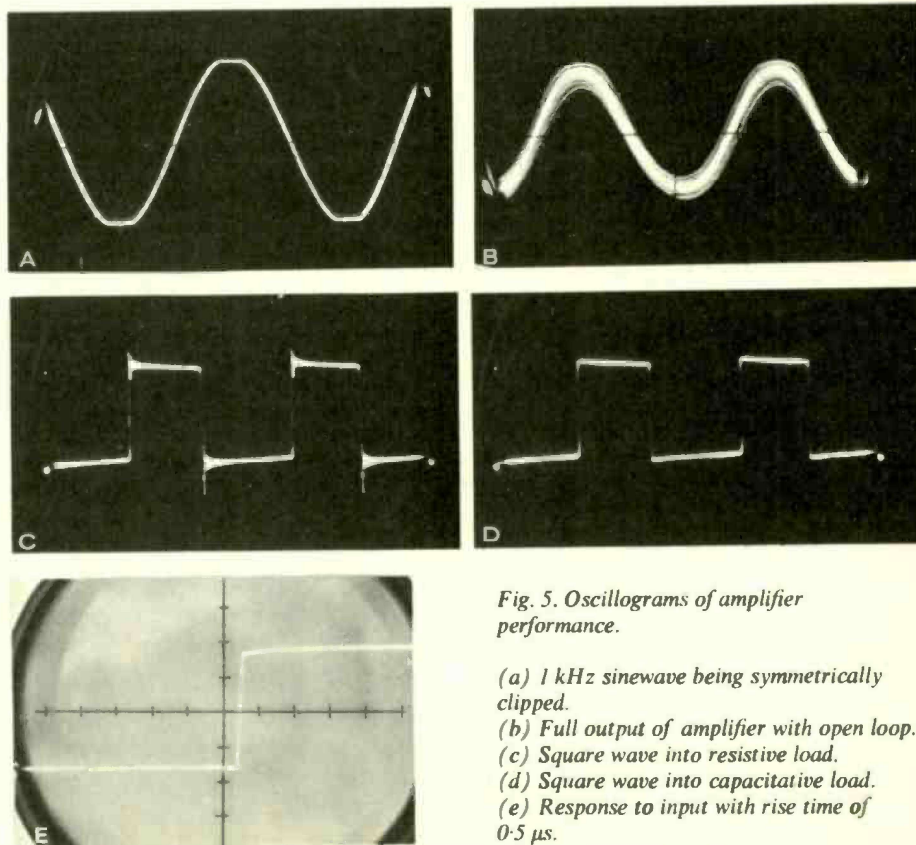


Fig. 5. Oscillograms of amplifier performance.

- (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 capacitive load.
- (e) Response to input with rise time of 0.5 μ s.

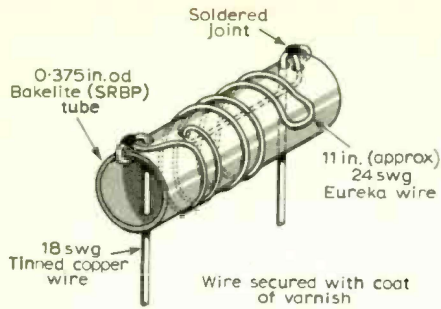


Fig. 6. Construction of 0.56 Ω 5% non-inductive resistors.

and therefore

$$V_{cc} = \sqrt{8R_{load} \cdot P_{out}} + V_{ce \cdot sat_3} + V_{be_4} + V_{ce \cdot sat_6} + (I + \hat{I})R_{11}, \text{ minimum.}$$

The standing current must exceed $\frac{V_{pk-pk}}{4R_{load}}$

in order to achieve the required voltage swing, and for its satisfactory safety margin it should exceed $V_{cc}/4R_{load}$.

Taking typical values for the circuit given using an 8-Ω load, and 10-W output level

$$\text{we get } V_{cc} = \sqrt{640 + 0.25} + 1.0 + 0.5 + (0.90 + 0.79)0.56 = 28 \text{ V.}$$

$$I_{min} = \frac{28}{4 \times 8} = 875 \text{ mA}$$

(a value of 900 mA being actually used.)

For a 3-Ω load and 10-W output we get figures of 19.5 V for V_{cc} , and 1.63 A for I_{min} . (Total power 31.8 W, 31.5% efficient.)

For a 15-Ω load and 10-W output we get figures of 36 V for V_{cc} , and 0.6 A for I_{min} . (Total power 21.5 W, 46.4% efficient)

From these figures it is apparent that the rise in $V_{ce \cdot sat_3}$ and V_{be} figures with the current used in a 3-Ω amplifier seriously reduces the overall efficiency. In the case of the 15-Ω 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-Ω stage is 39.8%.

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

Constructional 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 μF main smoothing capacitor, together with the negative connection of the second 50 μF 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.

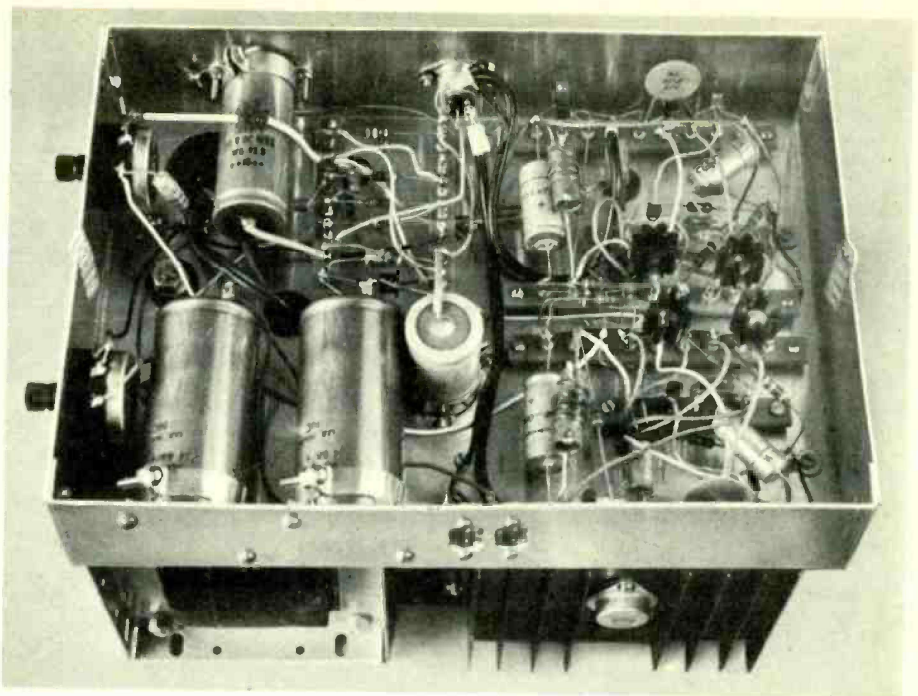


Fig. 7. View of underside of amplifier chassis.

Performance of 8-Ω version

Output (at commencement of clipping)	10 W
Frequency response	36 Hz–54 kHz (–1 dB)
Power bandwidth	15 Hz–92 kHz (–3 dB)
	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 μs
Input impedance	50 kΩ
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	–43 dB at 20 Hz rising to greater than –60 dB at 1 kHz and above

Fixed resistors

With the exception of the current sensing resistors R_{11} , R_{11a} and those marked with * in the circuit of Fig. 3, all resistors are solid carbon moulded $\frac{1}{2}$ W 10%. All resistors marked * are $\frac{1}{2}$ W 2% metal oxide (Electrosil TR5, Welwyn MR5, Radiospares “ $\frac{1}{2}$ W oxide”). See Fig. 6 for details of the construction of R_{11} .

Variable resistors

Both are wirewound Radiospares type “pre-sets” (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 μF 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 μF 20 V ceramic disc (base-collector

Tr_7). 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 μF” or “Hi-K 0.001 μF” (tubular).

0.1 μF 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. Radiospares 400 V PDC.

Electrolytic capacitors

47 μF 6 V (base-emitter Tr_8). 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 μ F, 6.4 V solid aluminium type). 10 μ F 64 V (input bias chain) Mullard C426AR/H10. 250 μ F 25 V (feedback blocking capacitor) Mullard C437AR/F250. 250 μ F 40 V (bootstrap capacitor) Mullard C437AR/G250. 1250 μ F 40 V (across 28 V supplies) Mullard C431BR/G1250. 2500 μ F 40 V (output capacitor) Mullard C431BR/G2500. 5000 μ F 50 V (main smoothing) Daly type obtained from Electrovalue. Nearest Mullard type C432FR/G5600 (5600 μ F 40 V). 50 μ F 350 V (voltage doubler) Radiospares "tubes 50 μ F 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 μ F 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 3A rectified d.c.

Fuse

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

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 (Tr_3 , Tr_5), are fitted with cooling clips—Redpoint 5F, available from Electrovalue and Electroniques. A similar type—"Sinks TO-5"—is available from Radiospares.

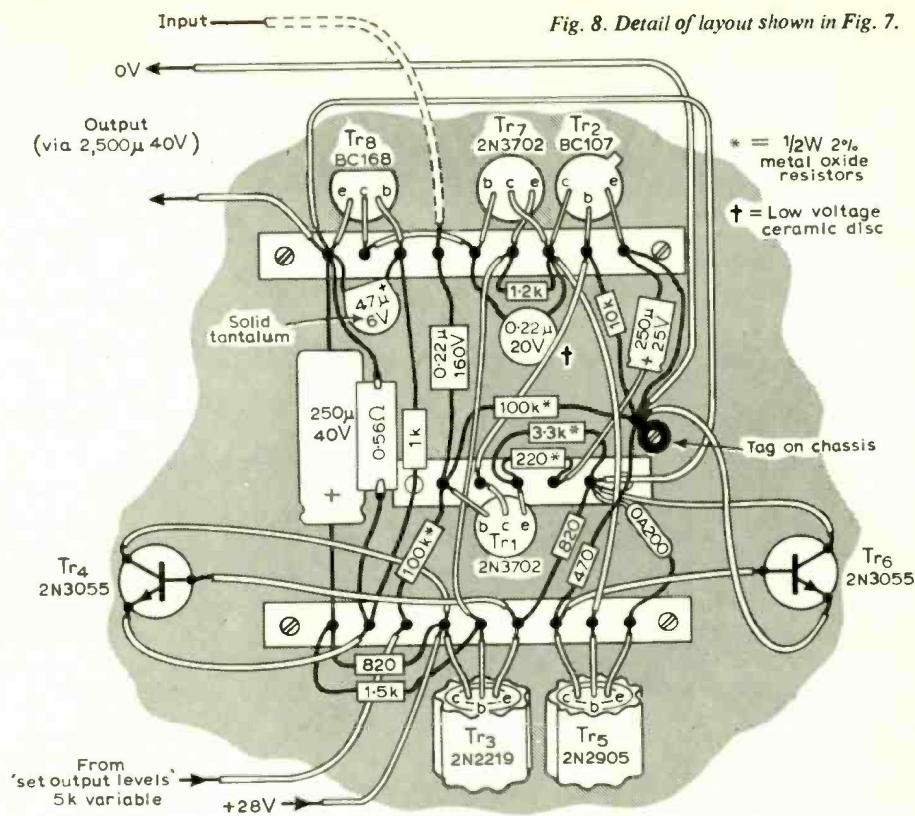
Sundries

Chassis size 7 in \times 10 in \times 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- Ω output

R_{11} and R_{11a} must be reduced to 0.31 Ω (5%) each. The mains transformer will require to



Semiconductors

Tr_1 , Tr_7	2N3702	(BCY70)
Tr_2 , Tr_9	BC107	(BC108 suitable for Tr_9)
Tr_3	2N2219	
Tr_5	2N2905	
Tr_4 , Tr_6 , Tr_{11}	2N3055	
Tr_{10}	2N3054	
Tr_8	BC168	(BC108)
D_1	OA200	(HS1010, OA202)
D_2 , D_3	RAS310AF	(Radiospares REC51A, 1N4005, BY103)
ZD_1	ZF8.2	(Radiospares "MZ-E 8.2 V", Mullard BZY88-C8V2, Texas IS2068A)
Rect. 1	Radiospares REC.40. 5A bridge 200 V (p.i.v.)	

be 21 V r.m.s. 3.5 A d.c. rectified rating. The output capacitor feeding the loudspeaker must be 5,000 μ F 25 V. The 12-k Ω resistor in the regulator will reduce to 7.5 k Ω , and the 3.3 k Ω resistor feeding the 6.8 V zener diode will reduce to 2.2 k Ω . The main smoothing capacitor should be raised to 7,000 μ F at not less than 30 V working. The collector resistors of Tr_2 should be dropped from 820 Ω , 1.5 k Ω and 1.2 k Ω to 470 Ω , 820 Ω , and 680 Ω respectively.

Modifications for 15- Ω output

R_{11} and R_{11a} must be increased to 0.84 Ω (5%) each. The mains transformer must be 34 V r.m.s. 1.5 A d.c. rectified rating. The 12-k Ω resistor in the regulator must be increased to 17 k Ω which is not a standard value, alternatively the 4.7 k Ω may be dropped to 3.6 k Ω which is a standard value. The 3.3-k Ω resistor feeding the 6.8-V zener diode should be raised to 3.9 k Ω . The collector resistors of Tr_2 may be raised if desired but this is not necessary. Tr_9 must be BC107 since BC108 has an inadequate voltage rating. Tr_3 may be 2N2219A or

2N218A which have a higher voltage rating than 2N2219. However if 2N2218A is used then Tr_5 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 V_{ceo} 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- Ω version, and +36 V for the 15- Ω version.

It is expected that the distortion of the 3- Ω version will be two to three times greater than that quoted for the 8- Ω version, with similar or slightly better figures for the 15- Ω version. In the author's opinion, since very few speakers deserving the title high-fidelity, have a 3- Ω voice coil, the 3- Ω 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.

London's New Colour TV Centres

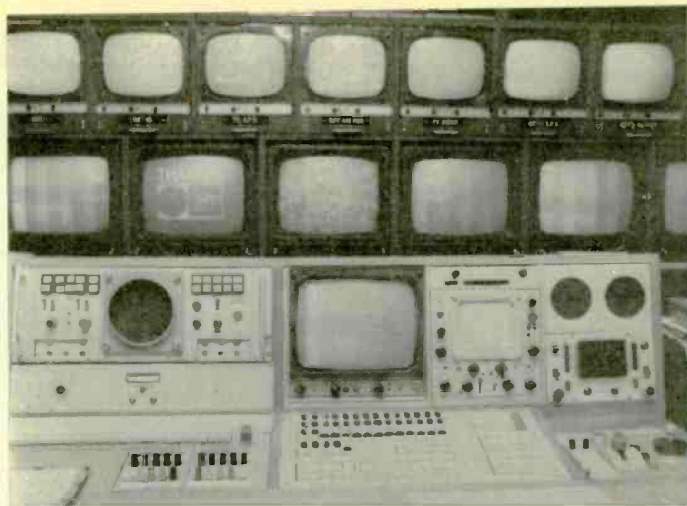
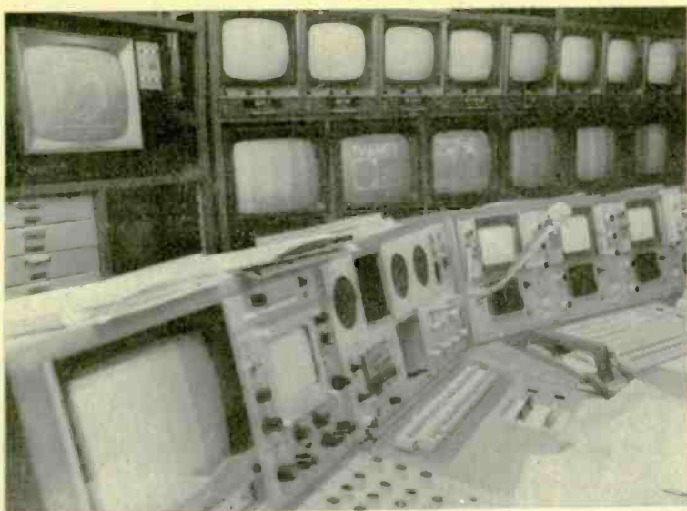
A pictorial look behind the cameras



One of two news studios (top left) at the recently built Television News Spur at the B.B.C. Television Centre. Amid the jungle of lights the four Mk. VII Marconi colour cameras can be seen. These are controlled remotely from control desks from which the operator can adjust pan, tilt, focus, zoom and camera height using simple potentiometers. Two banks of ten push-buttons, positioned one above the other, enable the operator to store up to twenty camera positions. Pressing any one of these buttons causes digital information describing the camera settings to be stored in a ferrite core store. Re-pressing the same button causes the camera to instantly take up the same position again. A "fader" control causes the camera to move between a position set up on one of the top row of buttons to a position which has been set

up on the bottom row of buttons. The camera control system was designed by Evershed Power-Optics Ltd. The news announcer sits in front of a screen which is saturated-blue in colour. The output of the blue gun of the main camera looking at the announcer can be made to switch an auxiliary camera the output of which is mixed with the main camera. If the auxiliary camera is looking at an outdoor scene, whenever the main camera is scanning the blue background the outdoor scene will appear on the screen. When the main camera scans the announcer very little blue signal will be picked up, the auxiliary camera will be switched out and the main camera will provide the vision signal. The effect on the television screen will be to have a picture of the announcer against a background of the outdoor scene. The sub-central apparatus room (top right)

which routes all the incoming and outgoing sound and vision signals to and from the B.B.C.'s news centre. In addition to this the C.A.R. provides communication facilities and can either route synchronizing pulses from the main television centre or generate its own for the rest of the news complex. The main sound routing system has 100 sources, any of which can be sent to any of 60 destinations. Remote controls also exist for the camera in the parliamentary studio. Part of the telecine area (bottom left) and one of the two telecine control desks (bottom right). Altogether there are nine 16-mm colour machines, two of which are multiplexed to deal with 8mm and super-8mm film from amateur sources, and two 16-mm monochrome machines. If necessary the colour quality of material from the telecines can be corrected.



The master control desk at Thames Television's new centre in Euston is boomerang-shaped and has positions for the lines engineer (top left), the engineer in charge who performs a quality control function (top right), and the network switcher (bottom left). The monitor bank facing the desk has a row of 14-in. monochrome monitors and a row of colour monitors underneath. These preview incoming sources, check the passage of signals

through the system and view the outputs. The lines engineer has a monitor-switching system controlling the input to an 11-in. Pye picture monitor and a 529 Tektronix 'scope. Communications and sound monitoring take up the rest of his desk. The central position for the engineer-in-charge has very comprehensive monitoring and switching facilities which include a vectorscope and a subcarrier phase meter (by Michael Cox Electronics)

seen below the vectorscope. The network switcher and the presentation mixer were built by Thames using E.M.I. vision matrices and Neve sound matrices. The presentation control room (bottom right) is separated from master control by a glazed screen so that visual contact can be maintained. The transmission controller sits centrally before the monitor bank, clocks, telephones and talkback keys on the desk before him.



ITN's new studios Wells St., London, were officially opened by the Queen on the 20th of November last year. The control room can be seen (left). Beneath the clock is the colour transmission monitor with colour preview pictures on either side. To the left of

the clock are the monitors for telecine and video tape recorders. Below the transmission monitor are the four studio camera monitors. Sitting from left to right: vision mixer, director, production assistant, and producer, rehearsing NEWS AT TEN. Far left



are the monitors for engineers controlling the quality of the picture. The 24-channel sound mixing and production desk in studio No. 1 is shown in the right photograph. This equipment, together with the turntables in the foreground, was supplied by Elcom.

80-metre S.S.B. Receiver

A limited coverage receiver of straightforward design for amateur use

by W. B. de Ruyter, PAOPRW

Since f.e.t.s are now available at low-cost it is possible to build a stable receiver with a performance similar to good valve receivers with the attendant advantages of low-power consumption and the absence of self-generated heat. The receiver described here operates on a 12-V supply and consumes only about 35mA.

Stability is such that the receiver stayed within 3Hz of zero-beat for several days when tuned to a standard frequency transmission. Detuning in the prototype due to supply voltage variation was about 50Hz/V making mobile operation using a good 12-V car battery possible. Due to the excellent square law characteristic of the f.e.t., cross-modulation properties are good. In a test, a 60mV unwanted signal spaced 100kHz from a weak wanted signal did not result in any harmful cross-modulation.

The sensitivity of the circuit depends almost entirely on the Q-factor of the input coil. It was noticed that practically no change in signal-to-noise ratio resulted when the aerial circuit was fed straight into the mixer instead of to the r.f. amplifier. However, the r.f. amplifier is needed to improve image rejection, reduce 455kHz interference and to provide adequate automatic gain control.

Circuit description

A block diagram is shown in Fig. 1 and the complete circuit diagram of the receiver is given in Fig. 2. The f.e.t./bipolar transistor r.f. stage, Tr_1 and Tr_2 , does not require neutralizing if due care and attention is taken with screening. Provided that the v.f.o. circuit is properly constructed, mechanical rigidity being important here, a good waveform and a stability approaching that of a crystal oscillator will be attained. The v.f.o. operating frequency is arranged to be 455kHz above the signal frequency (3.955 to 4.455 MHz).

All the r.f. coils employed in the prototype were of the type intended for valve trawler-band receivers for tuning between 60 and 180 metres.

The 4 to 40pF main tuning capacitor used in the prototype was salvaged from a Government surplus type 31 receiver and was complete with a 36:1 reduction gear box and trimmer capacitors. In fact constructors who are not too keen on "metal bashing" will find, as the author did, that the type 31 receiver cabinet makes an ideal case for the receiver described here.

The author considers that the money spent on the relatively

expensive mechanical filter is more than justified when looked at in terms of receiver performance. An added advantage is that i.f. alignment is reduced to trimming for maximum input to, and output from, the mechanical filter. The cascode i.f. amplifier is designed to properly match the mechanical filter and also incorporates the simple S-meter circuitry.

The use of a Colpitts oscillator for the b.f.o. eliminated the need for any coils in this part of the circuit. The b.f.o. operates below the bandpass of the mechanical filter.

A square law heterodyne detector is employed and it is necessary to adjust the i.f. output coil, L_s , for optimum reception quality.

After a d.c. coupled a.f. pre-amplifier stage, Tr_8 , the a.f. signal divides into two. One path is to a two stage f.e.t./bipolar a.f. amplifier via the a.f. gain control. This amplifier develops more than enough power to drive a pair of 150- Ω headphones. Some readers might prefer to incorporate a simple a.f. power amplifier for loudspeaker reception. The second path from the d.c. coupled a.f. pre-amplifier goes via an impedance converting emitter-follower, Tr_{10} , to the a.g.c. rectifier and smoothing capacitor. The a.g.c. performance is such that the heterodyne detector is not overloaded on even very strong signals. The switch S_1 is connected to the negative terminal of a suitable battery providing an r.f./i.f. manual gain control. The positive terminal of the battery is, of course, connected to earth (power supply negative).

The f.e.t. in the Tr_7 position, i.f. amplifier, must be selected for a certain value of pinch-off voltage, 3V being the target figure. It is best to obtain a good supply of these components so that suitable devices can be selected. A test circuit that will perform this task is given in Fig. 3; the meter will indicate pinch-off voltage. It is advisable to use an f.e.t. in the r.f. amplifier, Tr_2 , with a pinch-off voltage half a volt or so higher than the f.e.t. in the i.f. amplifier, Tr_7 . This will ensure that the a.g.c. cannot cut off the i.f. amplifier.

Construction

The author assumes that a type 31 receiver will be used as the basis for construction. The first step is to remove all the components from the chassis except the five-gang tuning capacitor and its associated reduction gearing. A small mA meter, which serves as the S-meter, is mounted in the position that was occupied

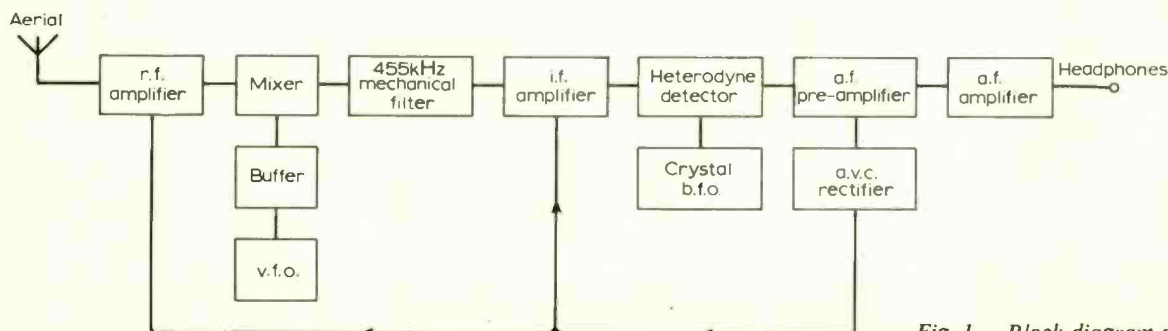


Fig. 1. Block diagram of the complete receiver.

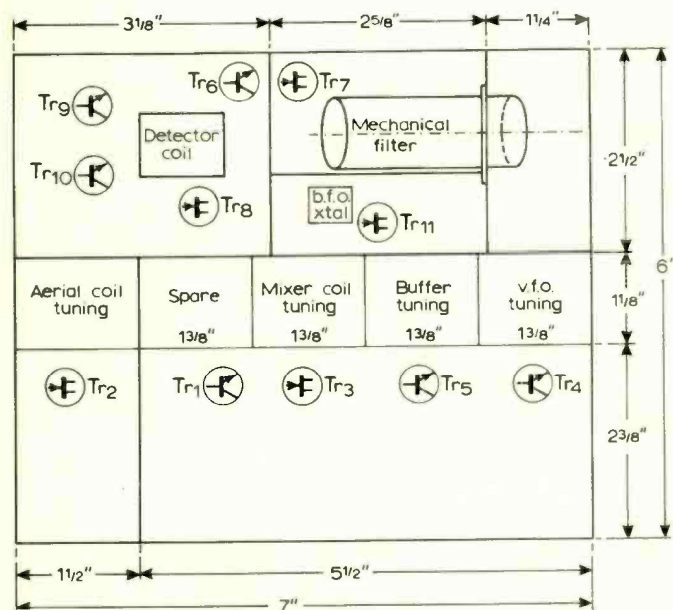
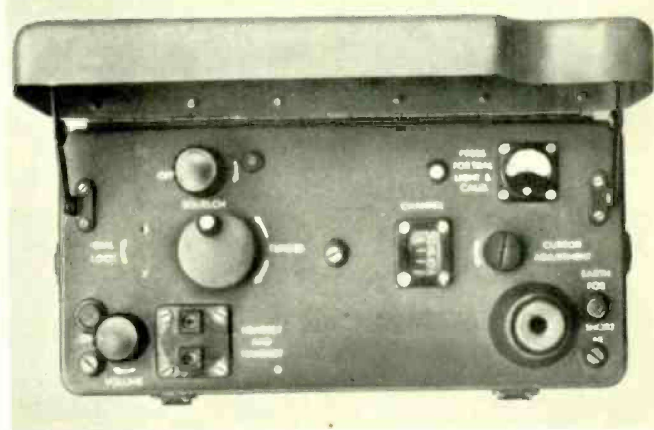
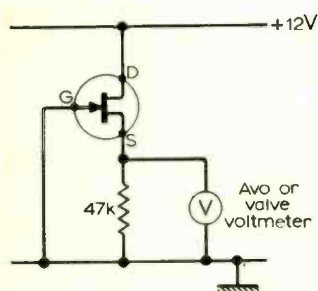
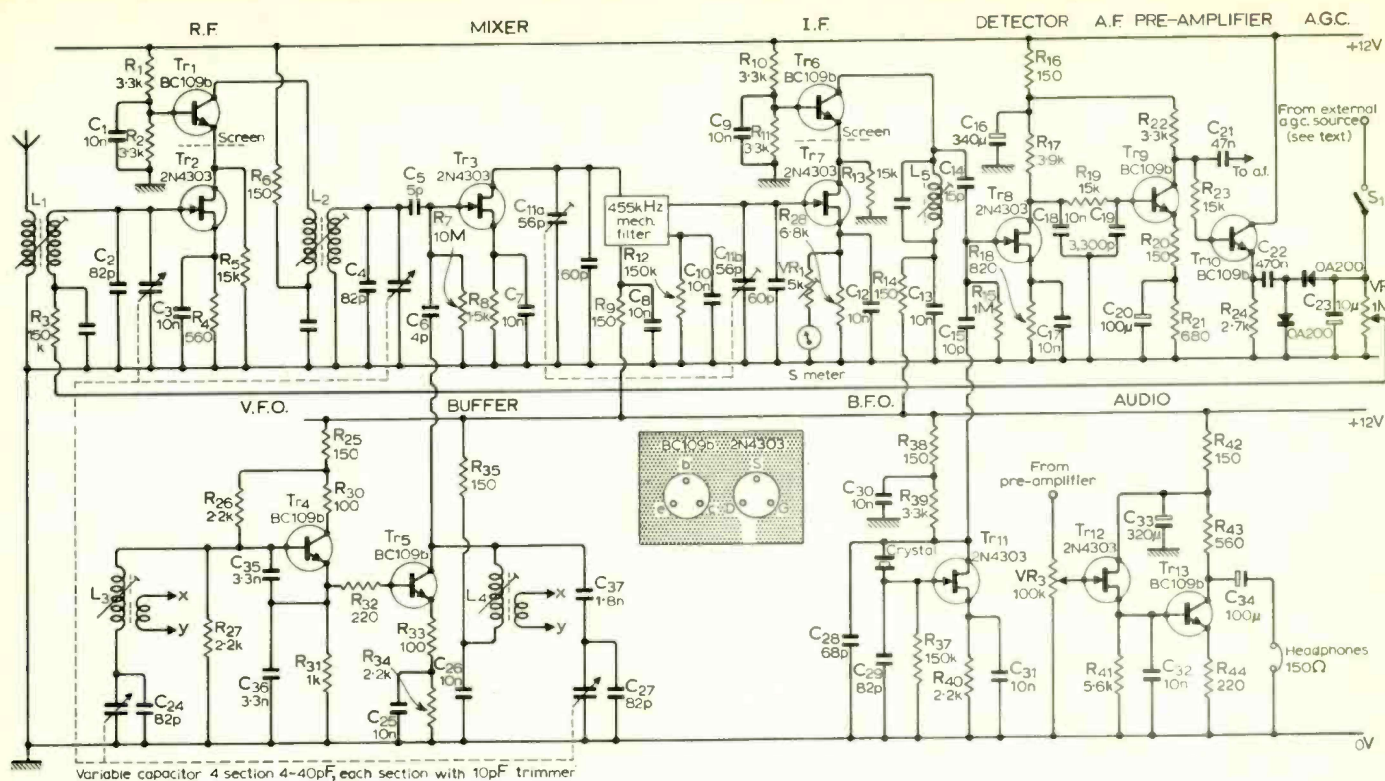


Fig. 5. Front view of the prototype.

by the dial-light knob, and the original squelch control knob becomes the a.g.c. control. The a.f. gain control is retained in its original position.

It was found that the 10-ft collapsible whip aerial supplied with the 31 set performed very well even without grounding the receiver.

The excessively large holes which now decorate the chassis are blanked-off with plates made from brass sheeting.

As previously stated any 60 to 180 metre trawler band coils can be used. The prototype employed Philips coils; type A3 125-34 for the aerial and mixer coils and type A3 125-68 for the v.f.o. and buffer. Only four of the sections of the five-section main tuning capacitor are used in the circuit; readers may find the fifth section useful for tuning a loop aerial.

The importance of rigid mechanical construction and good screening between stages cannot be overstressed as is normal with r.f. circuitry. It is a wise constructor who gives careful attention to these points. In particular excessive stray coupling between the input and output of the mechanical filter will seriously degrade the performance. Figs 4, 5 and 6 indicate the positions of the main components.

The first task is to check the source voltage of the f.e.t.s is