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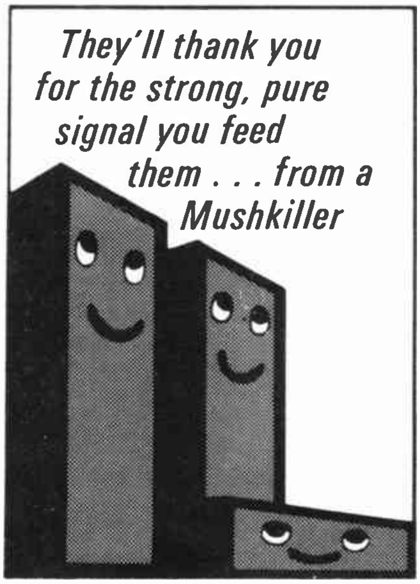

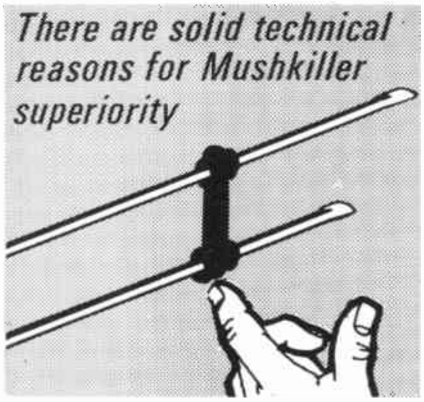

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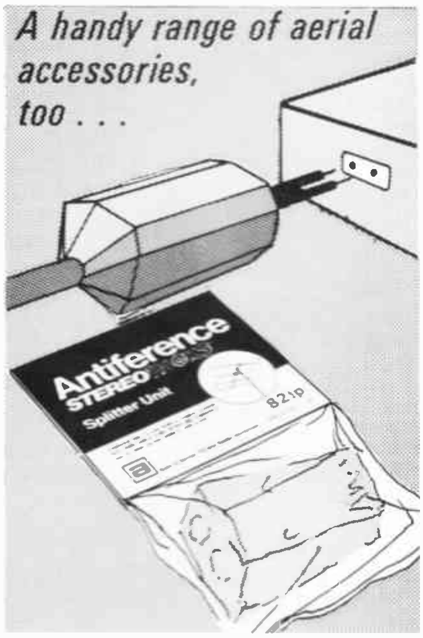

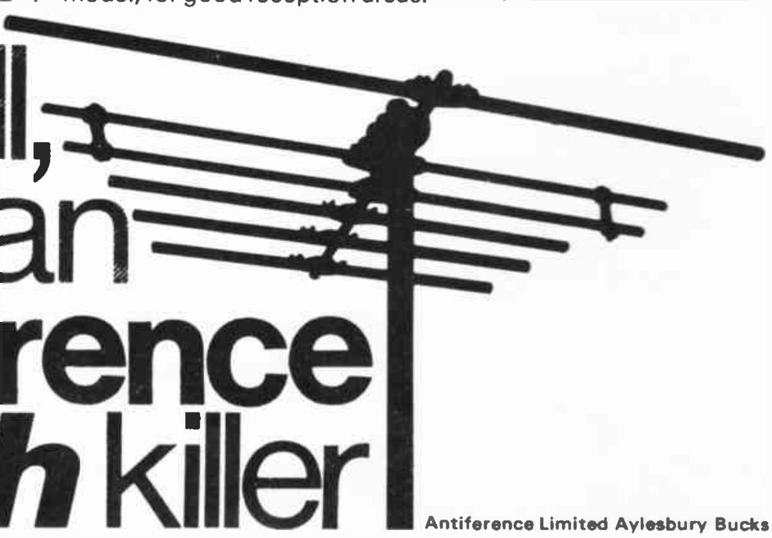

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High Fidelity Designs is a collection of the most popular audio constructional articles published in Wireless World during the last few years. It covers the whole range of equipment, from signal sources to speakers and headphones, and from it can be selected a system suitable for most requirements.

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From Hacksaw to Haydn

THE preoccupations of electronic engineers cover the majority of human activities—from farming to brain surgery, from submarines to space vehicles, from concertos to computers. But whatever their principal field of endeavour, common ground to them all in their leisure moments is the faithful reproduction of sound.

Over the years we have seen a stream of refinements to the recording and playback system until the hi-est of currently-available fi is sufficiently like the original sound to satisfy all but those who are unwilling to suspend even a fraction of disbelief. And many of these refinements are the result of spare time investigation.

The cost of a comprehensive system—turntable and pickup, tape deck, tuner, amplifier and loudspeakers—can be anything from £150 to £2000. At the lower-cost end of the range, it must be obvious that the reproduced sound is not as lifelike as it would be in the higher-cost equipment. In addition, the lower the cost of a system, the higher the proportion of the price seems to be devoted to external non-essentials.

It makes sense, in this situation, to investigate the possibilities of "rolling one's own" and it is with this in mind that we have collected a selection of input units, amplifiers and speakers that have appeared in our journal over the last few years. With this equipment, the highest quality of reproduction can be secured. It can be built to suit the owner's requirements exactly with no bowing in the direction of mass production techniques or a manufacturer's idea of good appearance. Cost will be appreciably lower than in an equivalent-quality commercial unit.

Even if the reader does not wish to build the equipment described, the collection will be a very useful description of modern techniques in high-quality sound reproduction. Our authors are in the forefront of development and their designs can be accepted as being the best available.

Enjoy the book—and please try not to lose sight of the objective. Music is for listening to, not for testing audio equipment. Do not let the percentages interfere with the Purcell.

High-quality Tape Recorder

1. Specification and design

by J. R. Stuart, B.Sc.*

Tape recorder construction has received relatively little attention over the years, and presumably one reason is the apparent complexity of the circuitry and alignment, compared with other items of domestic audio equipment. Two tape recorders have been described in these pages in a period of ten years, whereas several power amplifiers have been described in the last few months.

In view of the large interest in the construction of domestic audio equipment, it was decided to produce a design for a tape link which would be simple and cheap to build and easy to set up.

Reel-to-reel or cassette?

Continuing tape recorder development has resulted in commercial machines, using standard reeled tape, which give excellent performance at low speeds with or without crossfield bias.

Probably the most significant developments have been the large improvement in high-speed tape copying techniques, widespread acceptance of the Dolby noise reduction process, and the rapid growth of interest in four-channel stereo. These combine to create a situation in which tape will take over from disc as the major programme source particularly as no compatible coding can record four independent channels on a disc—although it can be done at the expense of crosstalk.[1]

It is now possible to manufacture a cassette tape to run at 1⅞ i.p.s. which, with Dolby, gives a performance better than disc. However, at present no cassette tape transport is available which can offer the necessary low wow and flutter performance nor the retrieval capability of a high-quality deck of the conventional form.

The choice of a conventional deck for this design was made without hesitation, for the use of such a machine will not decline

*Marconi Instruments Ltd.

TABLE 1
Specification of the complete recorder

Bandwidth measured at −9 dB. C.C.I.R. replay;

15 i.p.s.	25 Hz	− 30 kHz	±1 dB
7½ i.p.s.	25 Hz	− 17 kHz	±1 dB
3¾ i.p.s.	25 Hz	− 11 kHz +1	−2 dB
1⅞ i.p.s.	25 Hz	− 6 kHz +1	−3 dB

Distortion (at 1 kHz)

7½ i.p.s.	0 dB	1%	third harmonic	(reference level)
	+2.3 dB	2%	third harmonic	
15 i.p.s.	0 dB	0.7%	third harmonic	
	+2.3 dB	1.5%	third harmonic	

Dynamic range

56 dB	15 i.p.s. and 7½ i.p.s.	(weighted)
54 dB	3¾ and 1⅞ i.p.s.	(weighted)

Crosstalk
−60 dB mono
−45 dB stereo
Amplifier hum and noise
below −66 dB
Input sensitivity
7 mV rms into 45kΩ or 600Ω
or 25 mV rms into 150kΩ or 600Ω
or 250 mV rms into 1.5MΩ or 600Ω
Output
25mV rms. Output Impedance < 100Ω
and 250mV rms. Output Impedance < 100Ω
Peak-programme meter
Switchable to measure record, replay and bias levels.
Cost
£20 + £64-68 for the Brenell MK 6 deck

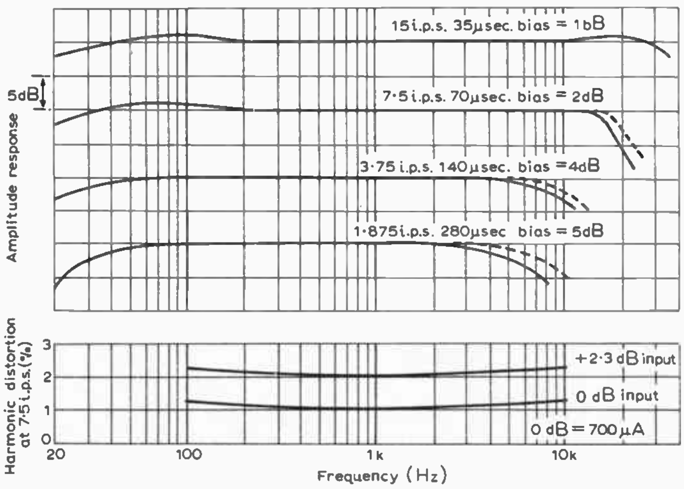


Fig. 1. Maximally flat frequency response.

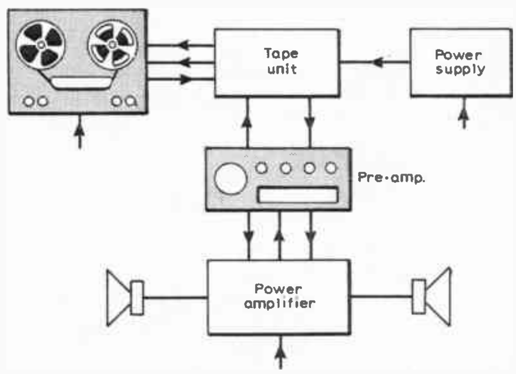


Fig. 2. Expected arrangement of the tape unit.

when live recordings are made for either amateur or professional applications, particularly those requiring editing. Further, many decks of this type are in use and may be adapted to this design.

Crossfield biasing was not considered in view of the extreme mechanical problems this would create for the constructor.

This tape recorder has been designed around the Brenell Mk 6 deck. Brenell Engineering Ltd have agreed to make this deck available in the required form.

Evolving a specification

Table 1 shows the performance of this tape recorder for the conditions described and Fig. 1 shows the frequency response for constant current record, C.C.I.R. play back at 7½ i.p.s. and 15 i.p.s. adjusted for a maximally flat response.

Equalization is described in detail later along with the corresponding setting-up and performance details.

In evolving a design the primary considerations were

(a) simplicity of design consistent with high performance

(b) non-critical construction
(c) the use of readily available components
(d) a minimum number of adjustments (the circuits deliberately leave very few parameters undefined and all calibration can be done with a multi-meter, although the full procedure is described)
(e) design flexibility to enable ready extension to four-channel and cassette applications when such decks are available.

The unit described is a mains-powered tape link and is intended for use with an existing audio system of pre- and power-amplifiers, and mixer if required. Such a recorder receives its signal from the pre-amplifier or mixer and replays through the same system. Three tape heads are fitted to allow simultaneous recording and playback; this affords better performance and much extended monitoring facilities.

The unit is readily compatible with the designs published in *Wireless World;* in particular the signal levels have been chosen to match the Bailey[2] and Nelson-Jones[3] pre-amplifiers. Fig. 2 shows the expected arrangement.

It was decided that the standard tape recorder should be a stereo unit capable of recording or replaying mono on either of the channels, with extensive monitoring facilities.

In addition to the considerations above, the particular performance parameters are cost, bandwidth, dynamic range and simplicity, and to achieve a good overall performance these must be carefully examined at each stage of the design.

To achieve simplicity it has been necessary to produce non-critical alignment with the full manufacturers' spread of devices, and the construction is no more complex than a power-amplifier. A block diagram of the tape unit is shown in Fig. 3.

Bandwidth

The bandwidth of a tape recorder is determined by the tape transport mechanism at low frequencies, and at high frequencies, to a first order, by

(a) recording speed

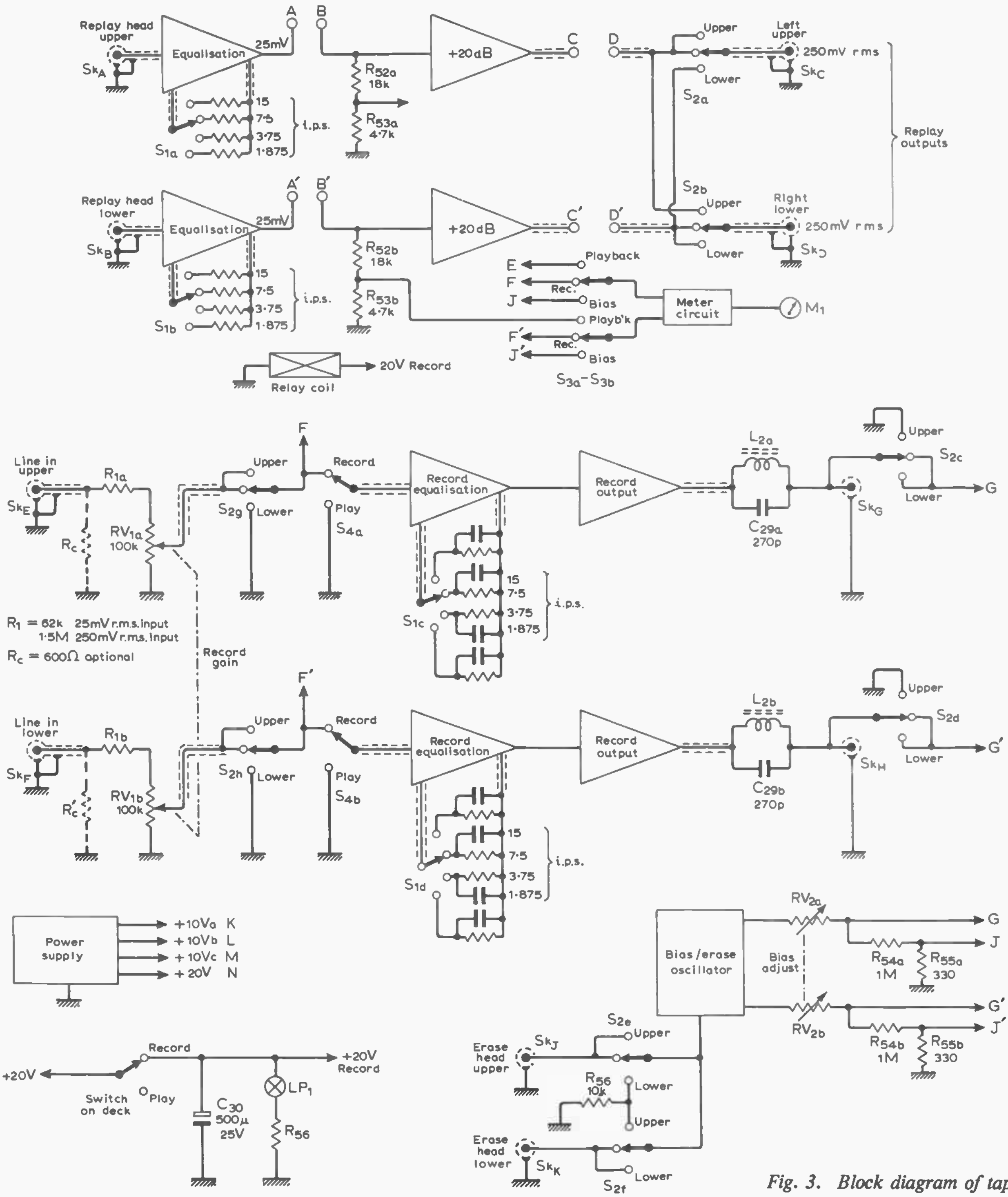


Fig. 3. Block diagram of tape unit.

(b) h.f. bias level
(c) replay head gap
(d) alignment of record and replay gaps
(e) equalization standard (I.E.C., N.A.B., D.I.N., C.C.I.R., N.A.R.T.B.)
(f) magnetic domain size on tape
(g) head losses (copper, and iron, and leakage).

Second order effects include the recording-head gap.

The Brenell Mk 6 uses Bogen heads which have a hyperbolic face to ensure good tape-head contact. They also have pressure pads which nevertheless seem to allow good low-frequency response as is seen from Fig. 1.

In a given system the parameters which the designer may control are a, b, and e, and to some extent d.

Great care must be exercised in producing a bandwidth specification; it seems dangerous to rely as much as we do on these figures. The problem is that in most cases it is the published specification for bandwidth and noise which sells a tape recorder. The author feels that it is of limited value to reject a model with an upper −2dB point of 15 kHz in favour of one which has the same point at 22 kHz; the reasons are as follows.

The sensitivity of the human ear at 17 kHz is a mean of 10 dB below 4 kHz at listening level of 60 phons, and the 1% duration peak content in an orchestral piece at 15 kHz is 10 dB below 500 Hz[4]. It would seem that a variation of ± 2dB at 20 kHz should have little effect, particularly as the threshold of hearing at 20 kHz is at a loudness of 80 phons (Robinson & Dadson) and in the upper octave just noticeable distortion is greater than 1 phon.

The ear is however sensitive to transient 'slewing-rate' and to inharmonic products.

No recording system can easily retain the phase information required to reproduce the transient information in the way required; however a lot can be done to reduce the intermodulation products which are generated in the upper band. It seems evident that the perceived difference between the systems of different bandwidth, is due to distortion produced by the method of bandwidth reduction, causing intermodulation products to appear in the region 1–6 kHz, with obvious effect. Because the major control of bandwidth of a tape recorder is the high-frequency pre-emphasis, and since harmonic products in the upper octave are not retained, the intermodulation products here, and the bandwidth, are determined by the recording characteristic.

Dynamic range

In a well designed tape recorder the dynamic range is determined by the tape and defined by tape overload and inherent background noise.

Sources of noise in the recorder are the amplifiers (more than 10 dB below tape noise in this design) and recorded noise by the bias and erase waveforms. In order to minimize this the erase waveform must be very pure and free from even-order harmonics.

Another source of noise is hum. However, careful power supply design and overall construction have reduced basic amplifier hum to less than −80 dB. The hum level appears far below the amplifier noise, and is inaudible in the author's set-up at a gain setting equivalent to 40W at a distance of 6 feet from the speakers.

Two-track operation was chosen to give a maximum dynamic range, however the Brenell Mk 6 deck is available with four-track heads and these may be used with no circuit modification giving about 3 dB less dynamic range.

Power supply

It was intended that the recorder should obtain raw d.c. from the power amplifier with which it is used, and a regulator is used to derive the system rail of +20 V. In case this power is not available a simple supply will be described in Part 2.

Choice of devices

The R.C.A. integrated circuit quad-amplifiers CA 3048 and CA 3052 were chosen for this design—which uses one of each. In

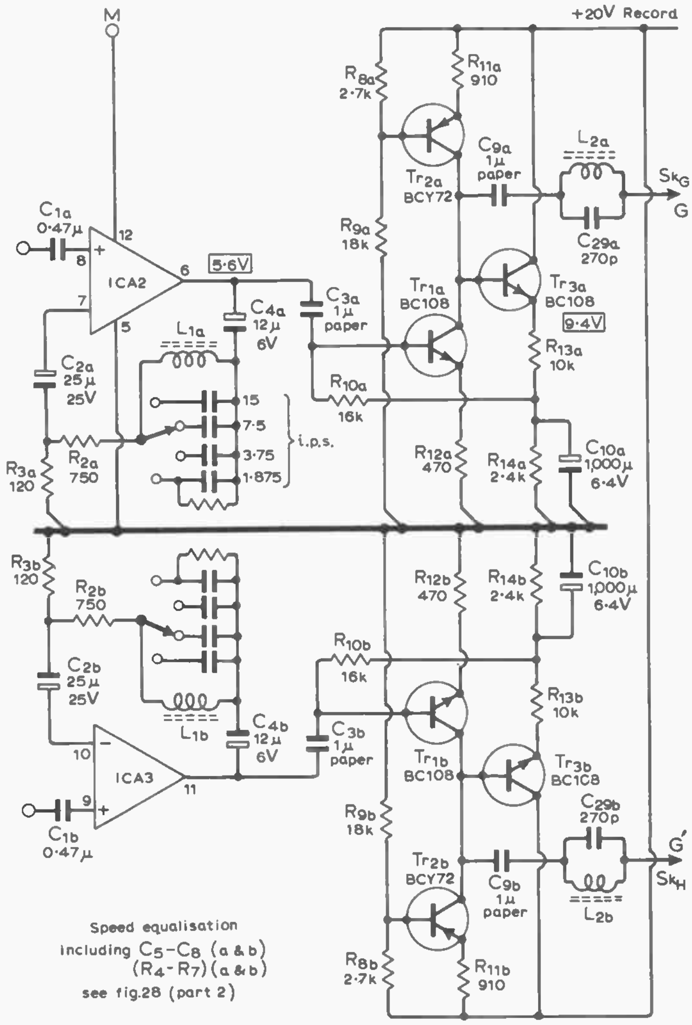


Fig. 5. Circuit diagram of recording stage.

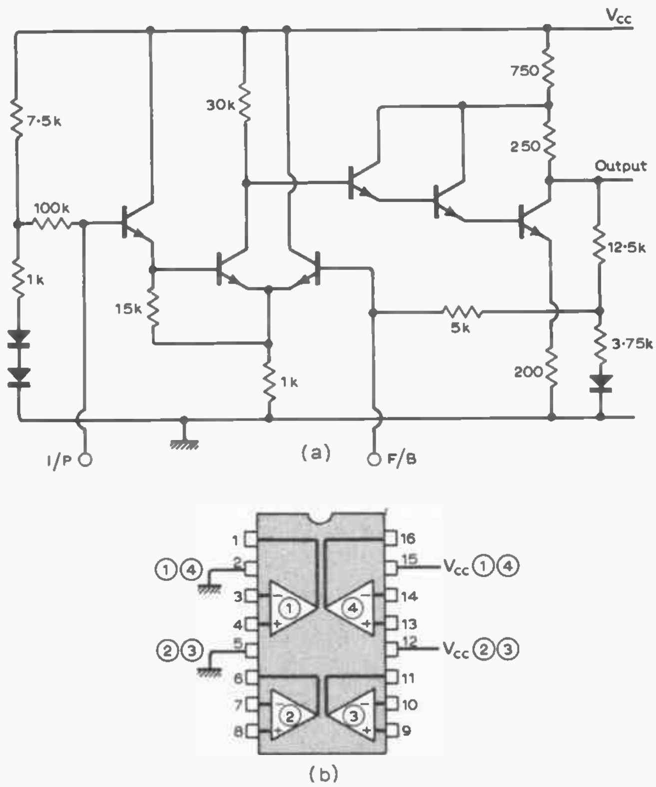


Fig. 4. Data on the i.c. linear amplifiers type CA3048 and CA3052; (a) the circuit diagram of each amplifier section (¼i.c.); (b) pin connections; and (c) performance details of each amplifier.

the author's experience they have a highly predictable, reliable performance and offer a saving of a very large number of discrete components. Although there is no reduction in cost, the reliability of one of these chips for home construction, when compared with the minimum equivalent of twelve transistors and associated components, is high. The circuits should be carefully checked however for the cost of mistakes could be higher. Fig. 4 (a) & (b) show the circuit diagram and specification for these devices. The transistors chosen are cheap silicon-planar devices of ready availability.

Recording section

The essential recording function is to produce a residual flux/input voltage transfer function which is linear with respect to amplitude variations.

In the mid-band residual flux relates linearly to applied flux, which is in turn proportional to the current flowing in the recording head windings, and so it follows that the recording current should be proportional to the signal voltage.

It is also necessary to modify the amplitude/frequency response of the recording stage to obtain the optimum bandwidth as described earlier.

The recording amplifier falls readily into two sections, namely the equalization and output stages.

Fig. 5 shows the circuit diagram for the stereo recording section. Reference to Fig. 3 shows that the record gain control is placed at the input to the equalization stage, i.c.A 2&3, to maintain optimum conditions of dynamic range and distortion. S_2 g & h, and S_4 a&b direct the input signals according to the selected function.

The open loop gain of i.c.A 2 & 3 is set to 45 dB by R_3, and the low-frequency gain of this stage is

$$\frac{R_2 + R_3}{R_3} \approx 7.25$$

This implies a sensitivity of 7 mV r.m.s. for 0 dB output level. Here 0 dB output was set for a flux density of 32 millimaxwells per millimetre of tape at +1 dB bias and $7\frac{1}{2}$ i.p.s.

The parallel tuned circuit formed by L_1, and C_{5-8}, increases the gain at the resonant frequency by an amount determined by R_{4-7}. Several combinations of frequency and boost may be used and these will be described in Part 2. Fig. 6 shows the frequency response of the equalizing stage when set for maximally flat response as in Fig. 1. This rising gain at high frequencies compensates to some extent for the losses in the recording head and tape, and ensures a 'constant induction' characteristic. Noise and distortion are both very low in this stage, distortion at 1 kHz is less than 0.01% at rated input, and the noise is more than 70 dB down.

As the CA 3052 amplifier can give 2V r.m.s. output with 0.65% distortion open loop, this equalizing stage is capable of producing 32 dB boost with less than 0.1% distortion. Because the recording head is a non-ideal inductor, it is an interesting problem to produce a 'constant current' drive at all frequencies in the pass-band; this implies an amplifier whose voltage gain is proportional to the head impedance.

A large number of designs have appeared, to produce this constant current drive for the head, and indeed to arrange this drive with a good 0 dB overload margin, to allow pre-emphasis, is quite difficult.

The Brenell Mk 6 deck is fitted with a Bogen UK202B record head, which has an inductance of 120 mH at 1 kHz and requires a recording current of 110 μA to induce a remanent flux of 32 mMx/mm; this head achieves its maximum impedance of 10 kΩ at about 14 kHz. Without pre-emphasis then the voltage across the head will be 1.1 V r.m.s. and as the output amplifier can provide 5.5 V r.m.s. across the head at this frequency the minimum pre-emphasis which can be applied to allow no overload at the 0 dB level is 14 dB at 14 kHz. It is worthwhile investigating the power-frequency spectrum of the signal source, as many music sources have maximum peaks at 15 kHz 10 dB below 500 Hz[4]. Thus if necessary a further 10 dB boost could be applied with less than 1% duration overload at these frequencies.

Wherever possible the nature of the pre-emphasis has been designed to accept a 0 dB signal without overload. If this is not the case the amount of overload is stated. Traditionally 'constant

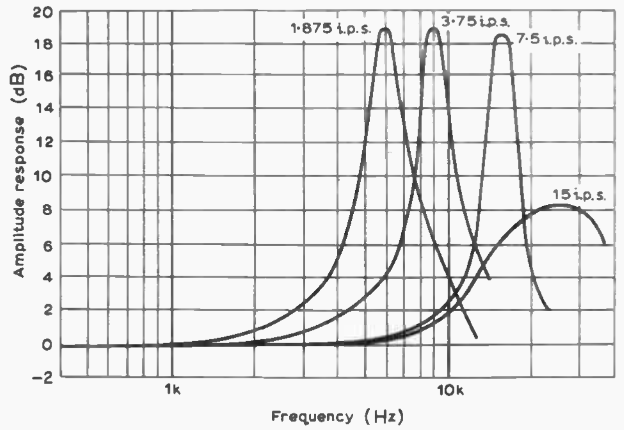


Fig. 6. Frequency response of recording pre-emphasis.

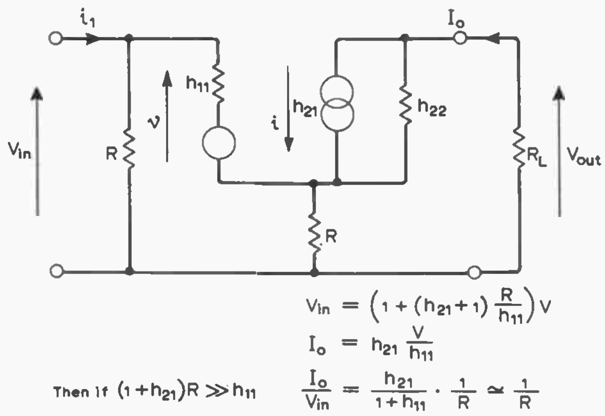


Fig. 7. Mid-band small-signal equivalent circuit of recording output amplifier.

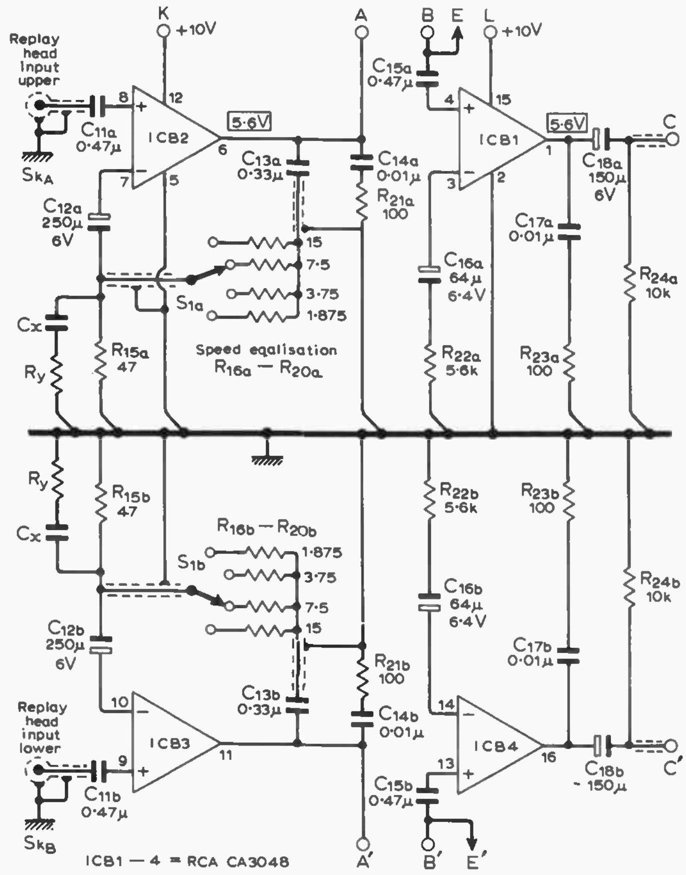


Fig. 8. Circuit diagram of replay amplifier.

TABLE 2 **(Replay equalization details.)**

speed	standard	time constants (μs) t1	t2	gain @ 1kHz (approx)	Cx μF	Ry μF	Rp	Rq	Rs	Rt	Rm kΩ	Cz
15 i.p.s. 38 cm/sec	CCIR/DIN IEC94* BSI (1970)†	35	∞	10	—	—	56	—	—	—	∞	s/c
	NAB, IEC (USA)	50	3180	12	—	—	100	—	—	—	9.5	10μ 6V
7½ i.p.s. 19 cm/sec	CCIR/DIN IEC94 (GB) BSI (1970)	70	∞	13	0.5	22	—	160	—	—	∞	s/c
	NAB, IEC (USA)	50	3180	12	0.5	22	—	100	—	—	9.5	10μ 6V
	IEC (FRANCE)	50	∞	12	0.5	22	—	100	—	—	∞	s/c
3¾ i.p.s. 9.5 cm/sec	CCIR	140	∞	15	1.0	22	—	—	390	—	∞	s/c
	BSI (1970) } IEC94 (GB)	90	3180	14	1.0	22	—	—	220	—	9.5	10μ 6V
	IEC94 (EUR)	140	3180	15	1.0	22	—	—	390	—	9.5	10μ 6V
		or 90	∞	14	1.0	22	—	—	220	—	∞	s/c
1⅞ i.p.s. 4.75 cm/sec	CCIR	280	∞	20	1.5	22	—	—	—	820	∞	s/c
	BSI (1970) }	120	1590	15	1.5	22	—	—	—	330	19	10μ 6V
	IEC94	or 120	∞	15	1.5	22	—	—	—	330	∞	s/c

* IEC94 inc. G8, USA, FRANCE.

† BS 1568 (1970). NOTE: Measurements on this unit used the CCIR replay time constants, and were made before the publication in September of 8S 1568 (1970).

current' was obtained by generating a very large signal voltage, and then swamping the head impedance with a large series resistance. Although simple to implement with valves this technique is inefficient and inelegant, although there are no problems with bias rejection.

Others have made use of the high intrinsic collector impedance of a transistor, notable examples being P. W. Blick[5], J. B. Watson[6], and G. Wareham[7].

Certainly the best method of ensuring accurate 'constant-current' drive is to include the head in the feedback loop of a high gain amplifier. However, this gives rise to considerable problems of bias rejection, and for this reason this technique was not employed in the basic recording unit. It will however be described in Part 3.

The circuit developed for this recorder is simple but effective. Tr_1 is a common-emitter amplifier with local feedback in the emitter and this stage is biased by the current source Tr_2; this gives a high output impedance, and the load seen by Tr_1 is essentially the recording head. Fig. 7 shows the equivalent circuit of the output

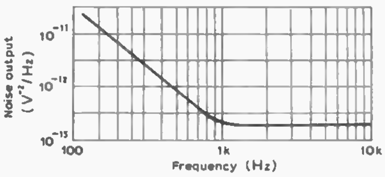


Fig. 9. Output noise spectrum of replay amplifier.

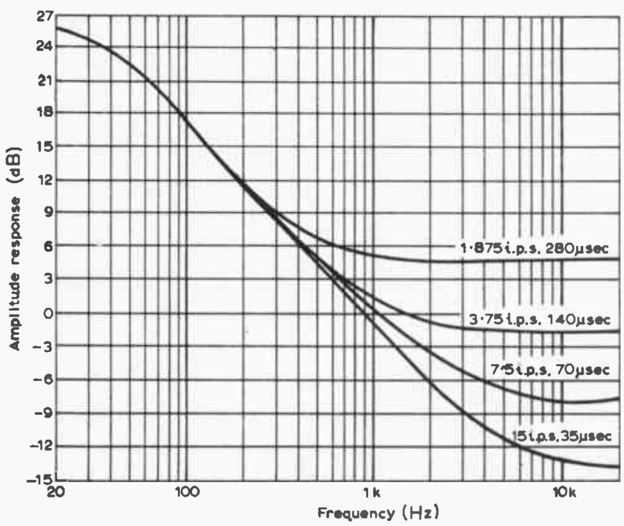
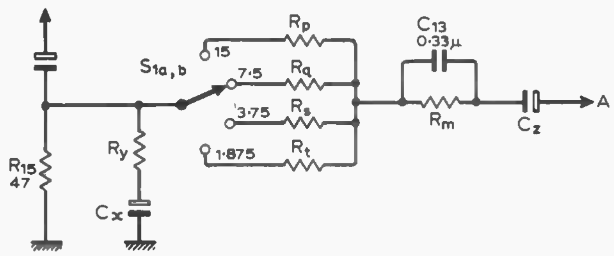


Fig. 10. Replay frequency response.



Equalization circuit referred to in Table 2.

stage for small signals at mid-band. The trans-conductance is given by $I_0/V_{in} = 1/R_{12}$. Tr_3 is an emitter-follower stage arranged to set the d.c. conditions in the amplifier. The d.c. stability is excellent, and substituting for Tr_1, transistors with h_{FE} between 30 and 475, causes a variation of only 200 mV on the standing d.c. level at the collector of Tr_1. Beware of measuring this with a meter of less than 10 MΩ resistance. The measured output impedance at 1 kHz is 420 kΩ, falling to 390 kΩ at 20 kHz. Maximum output is 5.6 V r.m.s. and clipping occurs symmetrically at an output of 18 V pk-pk.

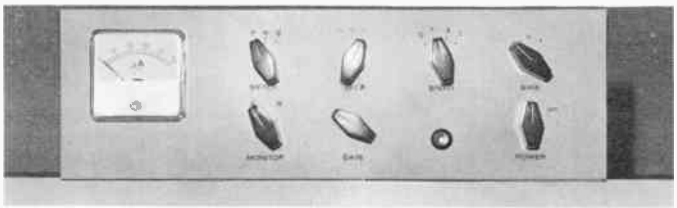
The frequency response measured with a 2.2 kΩ load was flat between 30 Hz and 100 kHz with −3 dB at 10 Hz and 220 kHz. At rated output the distortion in the current waveform in the head at 1 kHz was 0.01%.

It is strongly recommended that capacitors C_1, C_3 and C_9 be paper or polyester. In particular any leakage in C_9 would cause a permanent polarization to build up in the recording head, degrading the performance. To avoid large currents flowing in the head during switch-on the d.c. level at the output rises slowly and the h.f. bias is arranged to decay slowly after switch-off to demagnetize the head.

Replay

Fig. 8 shows the circuit of the replay amplifier. It is arranged as an equalization stage and a 20 dB gain stage to raise the output level to 250 mV r.m.s. The input from the UK 202B replay head is 2mV r.m.s. for a 1 kHz tone recorded at 32 mMx/mm at 7½ i.p.s.

Careful power supply design has enabled a hum level of −80 dB to be achieved with a very low crosstalk. The amplifier crosstalk measured was −74 dB at 1 kHz, and −65 dB at 10 kHz, for rated output; distortion is less than 0.01% and is predominantly 2nd harmonic; the overload capacity is 17 dB at 1 kHz with 7½ i.p.s. equalization. To obtain the best signal to noise ratio in this amplifier the CA 3048 amplifier is used; it has a tighter noise specification than the CA 3052 and is slightly more expensive. The measured noise was 66 dB below 0 dB level with 7½ i.p.s. C.C.I.R. replay equalization in a 20 kHz band and Fig. 9 shows the spectral density of the noise output of the replay stage.



R_{15} sets the open loop gain of the i.c. amplifier to 55 dB and the replay characteristic is determined by the equalizing network. This is shown in detail in Table 2, along with the values for the various standards. Fig. 10 shows the frequency response of the amplifier for the C.C.I.R. replay time constants. C_x and R_y can be added to lift the response at high frequencies. This is discussed fully in Part 2. As the power supply voltage at the i.c. terminal 12 rises slowly at switch-on, the charging current for C_{11} through the head is less than 1 μ A, and if C_{11} is a paper or polyester capacitor there should be no problem with polarization of the head. However, routine demagnetization will always be essential for high-quality work. The possibility of head magnetization is the only disadvantage with integrated or bipolar devices. An f.e.t. input would certainly eliminate the problem, but the circuit shown is far more convenient and this current has been reduced to an acceptable level. Those interested in the reduction of head polarization should refer to an article by P. F. Ridler[8].

In Part 2 next month the design will be concluded and constructional detai's given.

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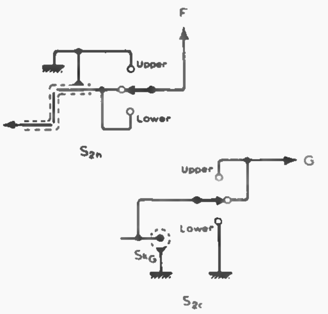
CORRECTIONS TO HIGH QUALITY TAPE RECORDER

The following amendments should be made to Part 1.

In Fig. 1 the bias levels given in the upper graph should have minus signs in front of them and refer to the 0dB = 700µA in the lower graph.

In Fig. 3 $S_{2}h$ and $S_{2}c$ should be drawn as below.



The mid position of S_2 is the stereo mode.

'. . the just noticeable difference is greater than 1 phon.'

In Fig. 4 the i.c. is viewed from above.

In Fig. 7 the final equation should read

$$\frac{I_o}{V_{in}} = \frac{h_{21}}{1+h_{21}} \quad \frac{1}{R} \quad \frac{1}{R}$$

In Fig. 9 the ordinate should be V^2/Hz. On p.10 col.2, line 16 should read '24V r.m.s. at 0.7mA'. In the captions to Figs. 13 and 14 'connect R to A' should read 'connect R_{56} to A'. Table 5 is referred as Fig. 28 on Fig. 5. In Fig. 15 the erase output should go to Skj and Skk—the first k is drawn as a 2. Table 2 gives R_y as µF instead of Ω

The components list gives the Plessey core number for T_1 as 905/1/01613/008µe; this should be 905/1/01613/108µe.

It is becoming difficult to obtain the Plessey cores in small numbers, however the requirement can be met from the new Mullard range as below:

L_1 6.25mH Plessey 905/1/01581/006µe; μ_{eff} 220, 41 turns/mH or Mullard LA1225 *and* LA1274, both numbers.

L_2 10.6mH Plessey 905/1/01581/009; μ_{eff}63, 84.5 turns/mH or Mullard LA1416 *and* LA1339.

T_1 Plessey 905/1/01613/108µe; μ_{eff}300, 32 turns/mH. Nearest equivalent: Mullard LA1219 *and* LA1275.

If the Mullard core is used for T_1, C_{29} should be 400pF beehive trimmers.

The tape heads specified were ½-track Bogen UK202B record and replay and UL290 erase; quarter-track heads UK 207B can be used without modification, although it is better to make C_{28} 100pF beehive trimmers.

The old quarter-track heads UK207 require more bias than the UK202B; for this the bias windings will need to be 120+120 turns and the oscillator run from about 11V, with C_{29} adjustable. The erase heads are all good substitutes in stereo.

High-quality Tape Recorder

2. Construction

by J. R. Stuart, B.Sc.

Signal monitoring takes two forms in this recorder—A-B signal switching, and a peak-programme meter to read signal and bias levels.

Experience has shown that the most effective meter has a fast attack and slow decay time. This stabilizes the pointer and gives a pessimistic reading, reducing the risk of overload while maintaining wide dynamic range.

Probably the most difficult transient a music signal could provide would be of 25 μ s duration, being one half cycle at 20 kHz! Although this is unlikely the meter described was designed to attempt to cope with this. A circuit diagram is shown in Fig. 11.

For simplicity only positive-going peaks are read. It is a simple matter to extend the circuit to read positive and negative peaks, but no instances have been found to suggest that this should be necessary. The author has used a meter of this type for several years without problems.

As the input signal to the recorder is balanced, either from a programme source or mixer, it is not necessary to have two expensive meter movements. Instead a ganged record gain control is used and the meter circuit of Fig. 11 indicates on a logarithmic scale the peak value of whichever channel is the greater at any instant.

Two i.c. amplifiers, A1 & 2, are used to raise the OdB input signal to 1V r.m.s. These low output impedance amplifiers charge

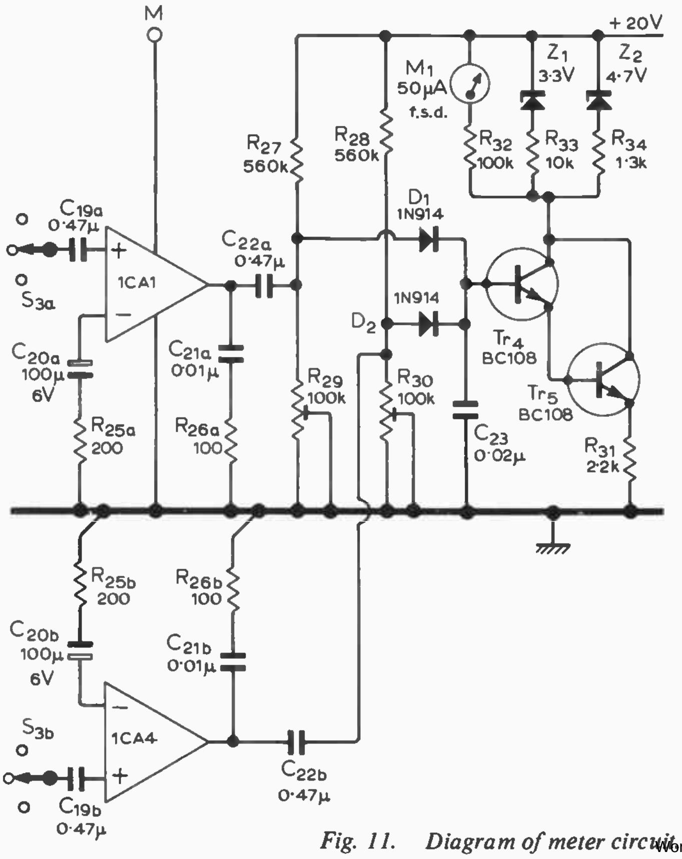


Fig. 11. Diagram of meter circuit.

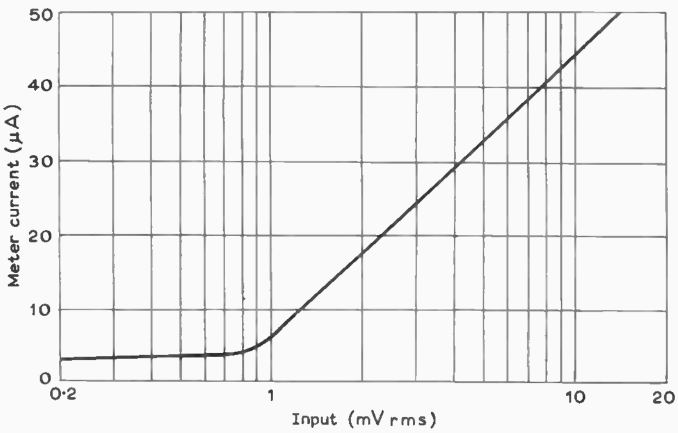


Fig. 12. Response of meter circuit.

capacitor C_{23} to the peak positive value via D_1 or D_2 and as the forward resistance of the diodes is current dependent, the attack time will depend upon the amplitude difference between successive peaks, with a minimum value of around 20 ms. The decay time is determined by the rate at which C_{23} is discharged by the high input impedance of the amplifier formed by Tr_4 and Tr_5. This impedance is nearly $\beta_4\beta_5R_{31}$ and the measured decay time is around 2s. Two zener diodes are included in the collector network to give a three-slope approximation to a logarithmic response; this approximation is good as can be seen in Fig. 12. Table 3 gives the meter calibration. The linearity is good and the meter response to continuous sine-wave is within $\pm$ 0.5dB from 10 Hz to 10 MHz.

To set-up the meter, switch on with the wipers of R_{29} and R_{30} set to ground, slowly rotate R_{29} until 2 μ A flow in the meter and then rotate R_{30} until the meter reads about 3 μ A. The circuit is now set and the calibration is determined by R_{25} and R_{31}.

As would be expected the super-alpha pair Tr_4 and Tr_5 exhibit some temperature sensitivity but the author has found no problems: R_{29} and R_{30} need never be touched after the initial set-up, provided $C_{22a,b}$ are paper or polyester capacitors and D_1 and D_2 silicon planar diodes. Small drifts in the standing current through the meter do not affect calibration, which is highly predictable.

Two schemes have been included in the basic recorder for A-B monitoring and the one chosen will be determined by the pre-amplifier with which the recorder is used. The essential difference is that one scheme includes the A-B switch in the recorder, in the other this switching is a pre-amplifier function.

TABLE 3 Meter calibration details

calibration current μA	calibration dB
50	+6
38.5	0
27	−6
16	−12
4.5	−18

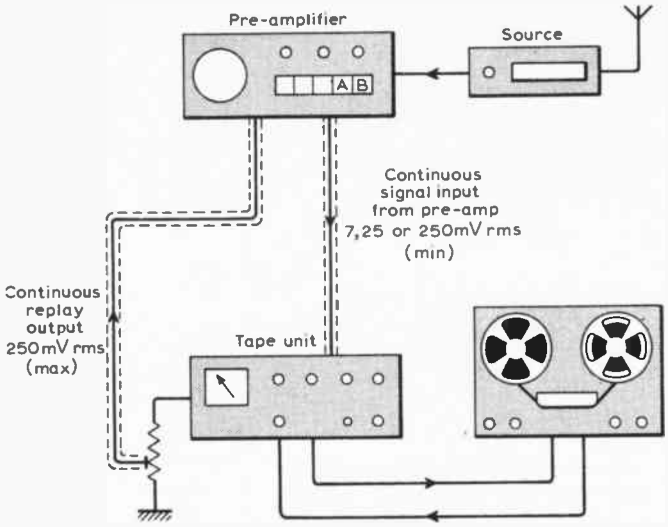


Fig. 13. A-B monitoring arrangement type 1. Input to top of RV_1 is 7mV. For output of 250mV r.m.s. link A + B, C + D; A′ + B, C′ + D′. For output of 25mV r.m.s. link A + D, A′ + D′; connect R to A.

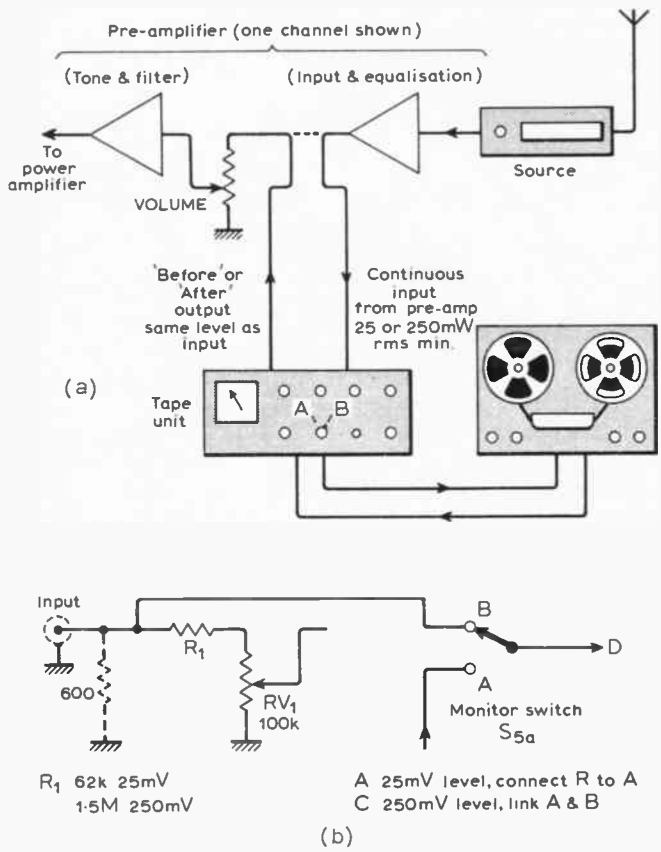


Fig. 14. A-B monitoring arrangement type 2. (a) The pre-amp is modified by breaking the lead shown dotted and re-routing as indicated. (b) Wiring for one channel.

Some amplifiers (Quad 33 and Leak for example) provide tape A-B switching. The scheme is shown in Fig. 13.

If this facility is not available then the scheme shown in Fig. 14 may be used; the input signal is passed to the recorder, which then either routes it back to the pre-amplifier 'B', or replays the signal on tape 'A'. The levels are arranged to be the same in this design, and the A-B function may be carried out at a 0dB level of 25 mV r.m.s. or 250 mV r.m.s. as shown in Figs 13 and 14, with reference to Fig. 3.

Bias and erase oscillator

As the bias network and erase heads present a reactive load to an erase oscillator it is a difficult problem to ensure that the calibration of the bias and erase current are retained on switching from stereo to single track recording. The most direct method of ensuring well defined currents is to use a master oscillator and two current mode output stages, this is described in Part 3; however the increased complexity and cost were not considered worthwhile in this recorder, in view of the excellent results which can be obtained, with care, from a multivibrator.

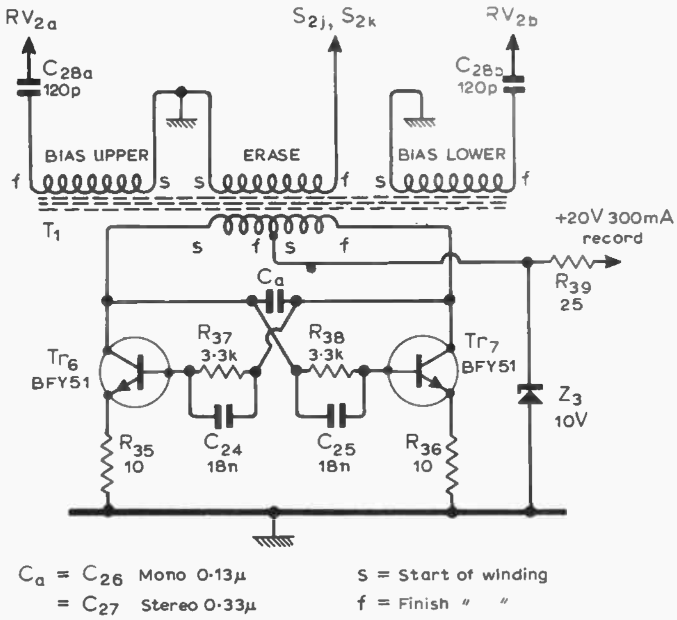


Fig. 15. Circuit diagram of erase oscillator.

Fig. 15 shows the circuit diagram of the oscillator. Although it does not operate in Class D[9] it is of a current switching type, the current being determined by the reflected load on the transformer primary and the supply voltage. Amplitude of oscillation is determined by Z_3. The Brenell Mk6 deck is fitted with Bogen UL290 erase heads. These require 70 V r.m.s. at a current of 45 mA (100 kHz) the equivalent loss resistance is 10 kΩ, giving a power of 500 mW per channel. Bias requirement for the recording head is 24 V r.m.s. at 9.7 mA for each channel.

In order to minimize interference from Droitwich transmissions, or harmonics of 38 kHz from stereo decoders, the oscillator was designed to run at 107 kHz. This frequency is set by the total effective primary inductance of T_1 and C_a; which is switched to keep the frequency constant for mono or stereo recording. On mono a 10 kΩ load is switched across the erase head to keep the bias and erase currents unchanged, see Fig. 2. L_{2a} and L_{2b} form high impedance tuned rejectors to keep the h.f. bias out of the recording stage, some 100 mV pk-pk of bias waveform appears at C_9.

The bias current is set by the capacitors C_{28a} and $_b$, and the ganged potentiometer RV_2 gives +3, −5 dB variation in the bias current, around 0.7 mA. As the loaded Q of the tuned circuit is around 30 the third harmonic distortion should be less than 0.4%. In fact because the driving waveform on the transistor bases is not square the distortion is about 0.1%. This is quite adequate for a bias waveform and the author can detect no increase in the noise on virgin tape when the oscillator is switched on, in fact the increase is just less than 1 dB. C_{30} ensures that the bias decays slowly to demagnetize the recording heads. Winding details for all coils are given in Table 4.

Power supply

During playback the quiescent power requirement is about 60 mA at +20V and this rises to about 500 mA during recording. Fig. 16 shows the circuit of a regulator which will accept 30-60V d.c.

TABLE 4 Winding details for inductors

L_1	100 turns 36 s.w.g. enamel covered wire.
L_2	275 turns 36 s.w.g. enamel covered wire.
T_1	Primary 5 + 5 turns 26 s.w.g. enamel covered wire, bifilar wound. Erase winding 100 turns 34 s.w.g. enamel covered wire. Bias winding 100 + 100 turns 36 s.w.g. enamel covered wire, bifilar wound. Separate each winding with one layer of Sellotape or similar.

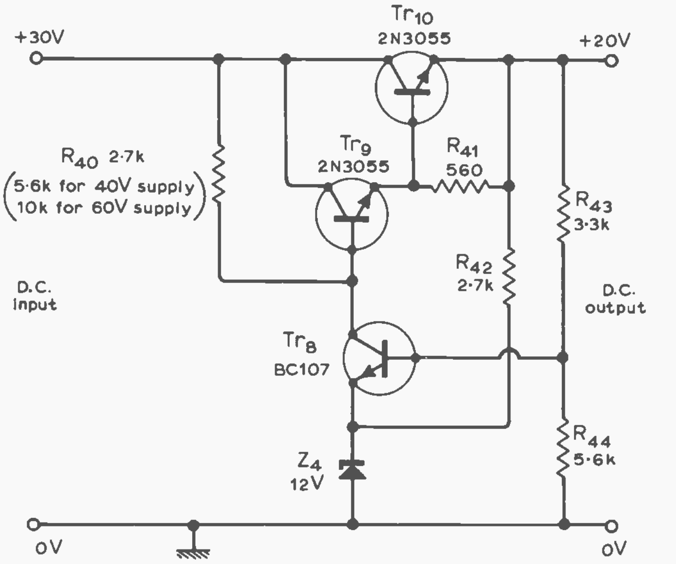


Fig. 16. Power supply input regulator. R_{40} must be altered as shown if higher supply rails are employed.

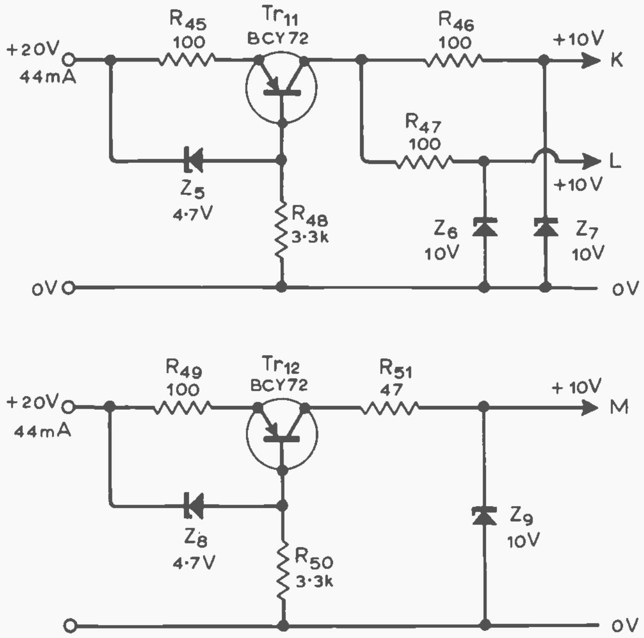


Fig. 17. 10V supply regulators.

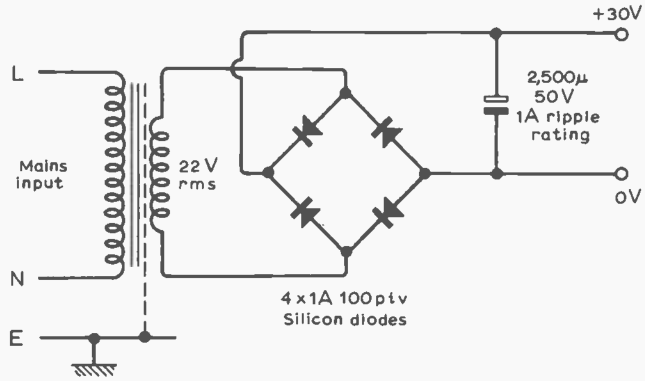


Fig. 18. Main supply circuit.

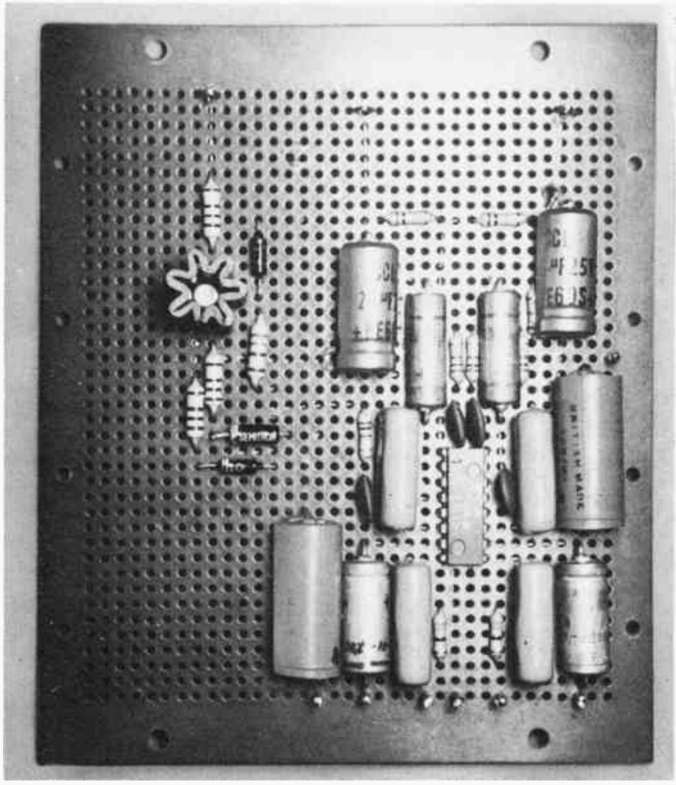


Fig. 19. Replay amplifier board.

input to give a 20V supply for the recorder. Tr_{10} should be mounted on a heatsink, or bolted to the chassis, with mica insulation to allow a safe operating temperature.

The integrated circuits use three separately derived +10V supplies to ensure stability and low crosstalk; the circuit is shown in Fig. 17. Tr_{11} and Tr_{12} are arranged as current sources. This affords protection to the i.c. and the effect of the low zener slope resistance with the high impedance of the current source is to reduce ripple by more than 66 dB. Tr_{11} and Tr_{12} should be fitted with cooling clips as they dissipate 160 mW each. Fig. 18 shows a simple power supply which may be used if d.c. is not available from the power amplifier.

Construction

The recorder unit circuits are built into an enclosed metal case. This affords rigidity and adequate screening. If a mains transformer is required it is recommended that this is not mounted in the case, but away from the unit and deck to maintain the low hum figures. A front panel carries all the controls detailed in Fig. 3 and the recording-level meter.

The circuits are built up on three Lektrokit boards 4in × 4¾in. The first board, which is shown in Fig. 19, carries the replay circuits including i.c.B and the regulator Tr_{11} which derives the +10V supplies for K and L in Fig. 8. (Pt. 1).

No special precautions need to be taken with the integrated circuits; the networks C_{14}, R_{21} and C_{17}, R_{23} reduce the gain above 200 kHz. Stability is independent of layout.

Fig. 20 shows the second board which carries the meter circuits, record equalization and output stages, including i.c.A. Layout is again non critical.

The third board shown in Fig. 21 holds the erase oscillator, bias rejection and the regulator Tr_{12} to power i.c.A. In the author's version the 20V regulator of Fig. 17 was external, however there is plenty of room to mount it on this third board. All equalization components are mounted on the switch S_1. To obviate any earthing or hum-loop problems the course adopted was to common on each board all earth and supply points and arrange only one connection to be made to these two rails. This means that the earth and supplies are common to the two channels but only one earth and supply lead is used per board. All signal leads in the unit are screened as indicated in Fig. 3 and the braid is earthed at only one end, that being the source. A wiring diagram is shown in Fig. 22.

It is not necessary to screen the circuits from each other. The erase board and the replay amplifiers should be arranged to be as

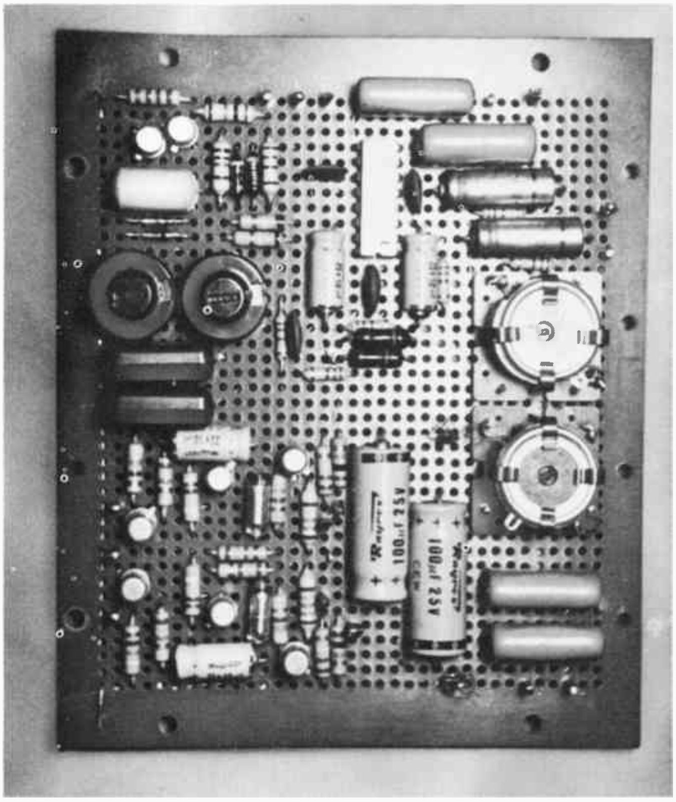


Fig. 20. Recording and meter drive circuits.

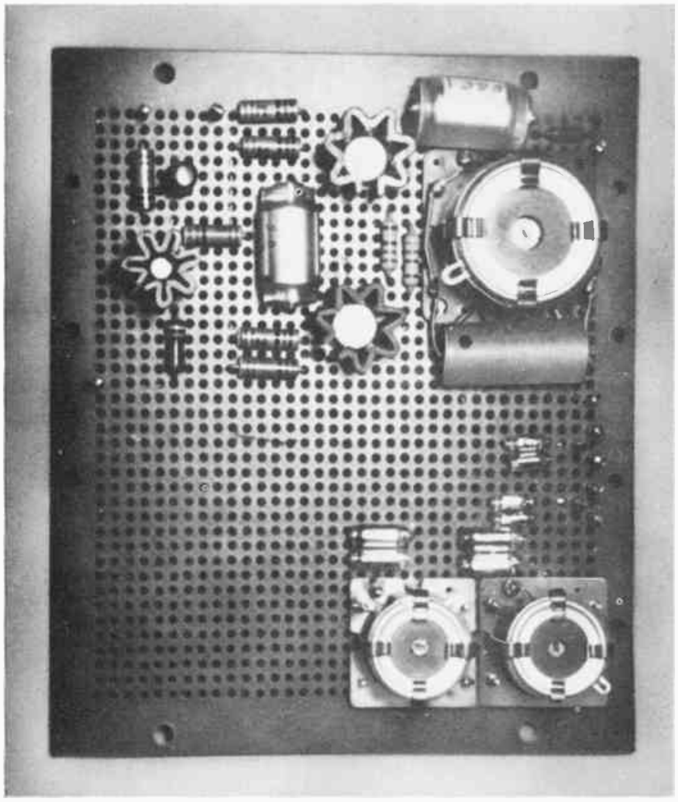


Fig. 21. Construction of erase oscillator.

far apart as possible but otherwise the layout is a matter of convenience. Inputs and outputs use Belling-Lee coaxial sockets for economy and good reliability. The Brenell deck supplies coaxial plugs on the record and replay heads, however the erase leads will have to be fitted with coaxial plugs.

Switching

Switching is carried out at the front panel and on the deck, the only switch which is repeated is the speed change. This was done to avoid the inevitable hum pick-up which would occur switching signals at the deck, and also it would not be possible to cater for the four speeds as the Brenell Mk6 deck uses a three-way switch with two capstans to provide $1\frac{7}{8}$, $3\frac{3}{4}$, $7\frac{1}{2}$ or $3\frac{3}{4}$, $7\frac{1}{2}$ and 15 i.p.s.

The record-play function is only required to switch the inputs to the recording equalization and to provide a 20 V rail to the recording circuits. For simplicity the Brenell deck is modified to give this +20 V rail when it is switched to record—which has an interlock button; the signal switching is then performed by a small low current relay, S_4.

This modification is necessary because in standard form the record-off-play switch on the deck is arranged to short the erase heads on all but the record position. This is not desirable as it would inhibit the decaying bias waveform. Fig. 23 illustrates the modification. The spare wafer is used to switch the +20V supply.

Testing and setting-up

It is recommended that the circuits are tested individually before final power supply connections are made. D.C. levels should be checked using a good multimeter; check all supplies, and the output of each i.c. amplifier, which should be around +5.6V in each case. The d.c. conditions of the recording output stage can be checked by measuring the potential on the emitter of Tr_3. This should be about +9.4V. If any discrepancy is noted, switch off immediately and check the circuit. Set up the meter as described. This may now be used when adjusting the bias rejection; otherwise use a high input impedance millivoltmeter or an oscilloscope. The bias rejectors L_2 allow a 5% adjustment of inductance and this should be more than enough to accept the expected uncertainty of oscillator frequency. Loose wiring or winding on L_2a or $_2b$ may move the resonant frequency, in which

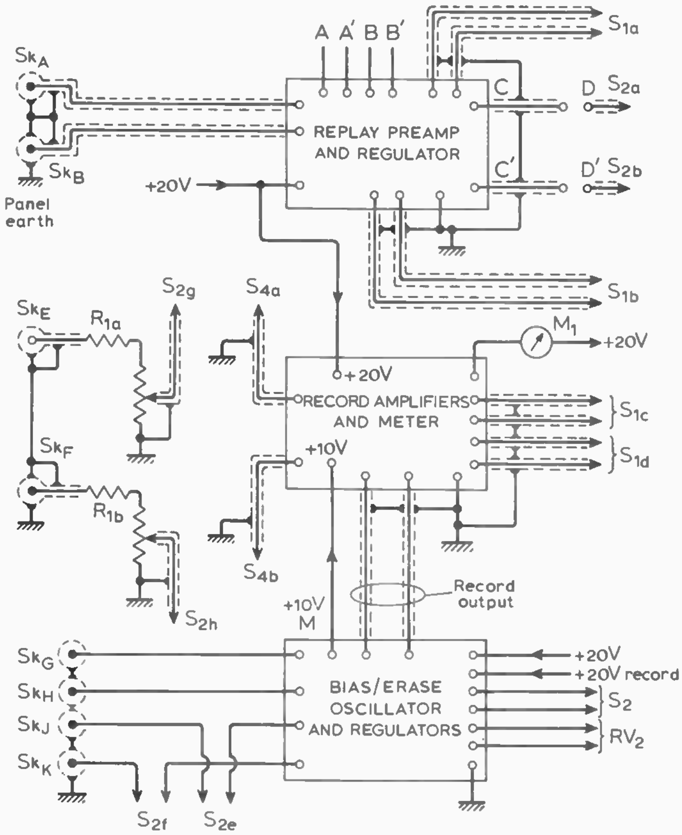


Fig. 22. Overall wiring diagram.

TABLE 5 Recording pre-emphasis component valves

speed ips	boost frequency kHz	amount of boost dB	C_1 nominal μF	R* kΩ	overload margin of boost dB
15	28	8.6	0.01	2.4	+5
7.5	15.5	18.25	0.02	22	−4
3.75	9	19.0	0.05	27	−5
1.875	6.5	19.0	0.08	27	+1

* depends on final Q of L_1

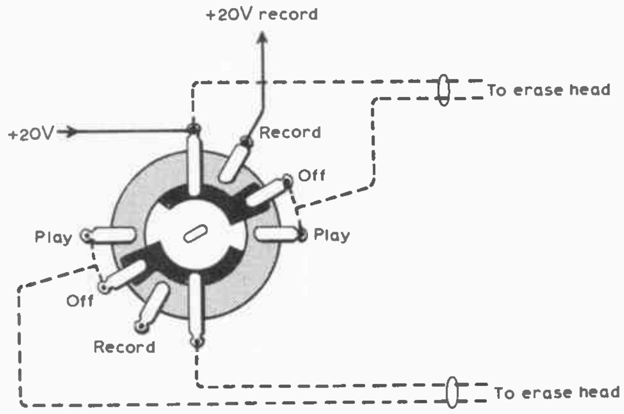


Fig. 23. Switch modification for Brenell Mk 6 deck. Remove wires shown dotted and add the wires shown solid.

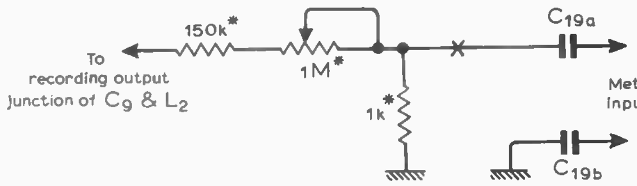


*Fig. 24. Arrangement of meter to set up bias rejection. Begin with the potentiometer set to maximum, then reduce as null is approached. Components marked * are required only for setting up.*

case *C* will have to be adjusted to find the null. Switch to record and using the arrangement of Fig. 24, or by another means, adjust L_2a and $_2b$ for a minimum bias voltage at the recording amplifier—with all erase and record heads connected. The meter may now be wired, according to Fig. 3. Check that bias is present at the correct level and that about +3 −5dB variation is available. The recorder is now set up and ready to operate.

Choice of recording standard

Several replay standards can be used and these are shown in Fig. 10. The choice will depend upon the tapes to be replayed. All the measurements given here use either the C.C.I.R. standard or C.C.I.R. with high frequency lift. On replay some high-frequency boost can be applied by connecting *Cx* and *Ry* of the values given; note that this network must be switched for each speed requiring an extra two poles on S_1. If tapes recorded to a different standard are to be used this lift must not be applied; however for recordings made on this machine the boost will compensate replay head losses at high frequencies and enable a true 'constant-induction' recording characteristic to be used, with less chance of high frequency overload.

It must be understood that having chosen a replay characteristic, then although a good approximation can be made, the amount of recording pre-emphasis which must be applied depends upon the tape used, the bias level—which incidentally will be different for each speed—and the form of response which is required, see Part 1.

Two examples are given here, maximally flat response with C.C.I.R. playback, maximally flat response with C.C.I.R. and high frequency lift. These two give low intermodulation distortion at high frequencies.

The component values are given in Table 5. The first response is shown in Fig. 1 and the dotted curve gives the response with high frequency lift on playback. With careful construction there is no reason why these specifications should not be met.

REFERENCE

9. Baxandall, F. J., "Transistor LC Sine-wave Oscillators", I.E.E. paper 2 978 E, February 1960.

Component List

Capacitors

C_1*	0.47μ paper†	C_{16}*	64μ 6.4V
C_2*	25μ 25V	C_{17}*	0.01μ disc ceramic
C_3*	1μ paper	C_{18}*	150μ 6V
C_4*	12μ 6V	C_{19}*	0.47μ paper
C_5*	0.01μ disc ceramic	C_{20}*	100μ 6V
C_6*	0.02μ disc ceramic	C_{21}*	0.01μ disc ceramic
C_7*	0.05μ disc ceramic	C_{22}*	0.47μ paper
C_8*	0.08μ disc ceramic	C_{23}	0.02μ disc ceramic
C_9*	1μ paper	C_{24}	18n polystyrene
C_{10}*	1000μ 6.4V	C_{25}	18n polystyrene
C_{11}*	0.47μ paper	C_{26}	0.13μ 2% paper
C_{12}*	250μ 6V	C_{27}	0.33μ 2% paper
C_{13}*	0.33μ paper	C_{28}*	120p polystyrene
C_{14}*	0.01μ disc ceramic	C_{29}*	270p polystyrene
C_{15}*	0.47μ paper	C_{30}	500μ 25V

*two required

†or plastic in each case

Semiconductors

ICA	CA3052	supplied by A. Marshall	Tr_{11}	BCY 72	with cooling clips
ICB	CA3048		Tr_{12}	BCY72	
Tr_1*	BC108		$D_{1,2}$	1N914	
Tr_2*	BCY72		Z_1	3.3V 400mW	
Tr_3*	BC108		Z_2	4.7V 400mW	
Tr_4	BC108		Z_3	10V 1W	
Tr_5	BC108		Z_4	12V 400mW	
Tr_6	BFY51	with 5 cooling clips	Z_5	4.7V 400mW	
Tr_7	BFY51		Z_6	10V 400mW	
Tr_8	BC107		Z_7	10V 400mW	
Tr_9	2N3055	plus insulation	Z_8	4.7V 400mW	
Tr_{10}	2N3055		Z_9	10V 400mW	

* two required

Resistors All ¼W 5% carbon unless otherwise stated.

R_1*	see Fig. 3	R_{21}*	100	R_{41}	560
R_2*	750	R_{22}*	5.6k	R_{42}	2.7k
R_3*	120	R_{23}*	100	R_{43}	3.3k
R_4*	2.4k	R_{24}*	10k	R_{44}	5.6k
R_5*	22k	R_{25}*	200	R_{45}	100
R_6*	27k	R_{26}*	100	R_{46}	100
R_7*	27k	R_{27}	560k	R_{47}	100
R_8*	2.7k	R_{28}	560k	R_{48}	3.3k
R_9*	18k	R_{29}	100k pre-set	R_{49}	100
R_{10}*	16k	R_{30}	100k pre-set	R_{50}	3.3k
R_{11}*	910	R_{31}	2.2k	R_{51}	47
R_{12}*	470	R_{32}	100k	R_{52}*	18k
R_{13}*	10k	R_{33}	10k	R_{53}*	4.7k
R_{14}*	2.4k	R_{34}	1.3k	R_{54}*	1M
R_{15}*	47	R_{35}	10 ½W metal oxide	R_{55}*	330
R_{16}*	see Fig. 8.	R_{36}	10 ½W metal oxide	R_{56}	to suit bulb
R_{17}*		R_{37}	3.3k	RV_1	100k log 2-gang
R_{18}*		R_{38}	3.3k	RV_2	50k lin. 2-gang
R_{19}*		R_{39}	25 5W w/w		* two required
R_{20}*		R_{40}	2.7k		

Inductors

L_1 Plessey 905/1/01581/006 μe 18mm pot core with base and clips, or Mullard LA2532. Two required.

L_2 Plessey 905/1/01581/009 μe 18mm pot core with base and clips, or Mullard LA2538. Two required.

T_1 Plessey 905/1/01613/008 μe 26mm pot core with base and clips, or Mullard LA2332. One required.

Switches etc.

S_1	4-pole 4-way (minimum, see text)	make-before-break
S_2	9-pole 3-way	make-before-break
S_3	2-pole 3-way	break-before-make

S_4 low-current relay 2-pole change over.

S_5 required for A-B monitoring arrangement type 2. 2-pole 2-way break-before-make

Bulb (6V 40mA: R_{56}=330)

Lektrokit board ×3 plus pins.

High-quality Tape Recorder

3. Extensions and modifications

by J. R. Stuart, B.Sc.

The variable high-frequency bias allows optimum recordings to be made with a variety of tapes and speeds, and it is a simple matter to reset any bias condition with the meter. Although the A–B monitoring allows an excellent attempt to be made by ear, it is not always straightforward to discover the required bias initially. In particular, if the recording is to be replayed on another machine, it may be necessary to bias for maximum sensitivity, minimum distortion, or some arbitrary standard.

The normal criterion for low tape speeds is to increase the bias until the sensitivity at 1 kHz is 1 dB below maximum. To allow easy setting of this bias current many high-quality recorders include a 1 kHz reference oscillator.

Such an oscillator would either be an *RC* arrangement with amplitude definition a¬d stabilization provided by a thermistor or field-effect transistor, or a current switching *LC* oscillator[9] of the type shown in Fig. 25. The output of this oscillator is well defined by the dynamic impedance of the tuned circuit and the tail current. However the values of *L* and *C* required do not lead to accurate prediction of the frequency of oscillation.

To calibrate the recorder using a reference, switch the meter to record and set the input level to −6 dB, then, while recording, switch the meter to replay and adjust the high-frequency bias for the required sensitivity. Note the bias voltage.

It is probable that in a large number of applications a simple stereo signal is not available. This will certainly be true of live or special-effect recordings, and for these a linear mixer is essential.

Fig. 26 shows a mixer which could be built as part of the recorder unit. For simplicity a CA 3048 integrated quad amplifier has been used, giving two inputs per channel. By extrapolation, further addition of i.c.s will give the required number of inputs.

The i.c. should be powered by a regulator identical to that shown in Fig. 17. Output *M* is satisfactory providing that it is not intended to cascade amplifiers in the same chip—to do so may cause low-frequency instability, so the dual output regulator (*K*, *L*) should be used.

This mixer is intended for a 250 mV rated output, with a 12 dB overload margin. R_x defines the gain of each mixer stage and a range of values is given in Fig. 26. However, if at any time high sensitivity is required, better noise performance would result from a lower gain mixer feeding the 25 mV input.

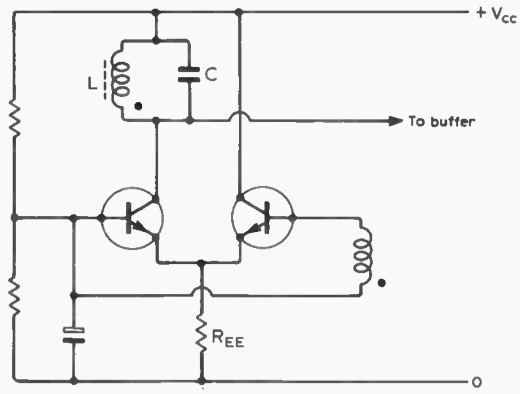
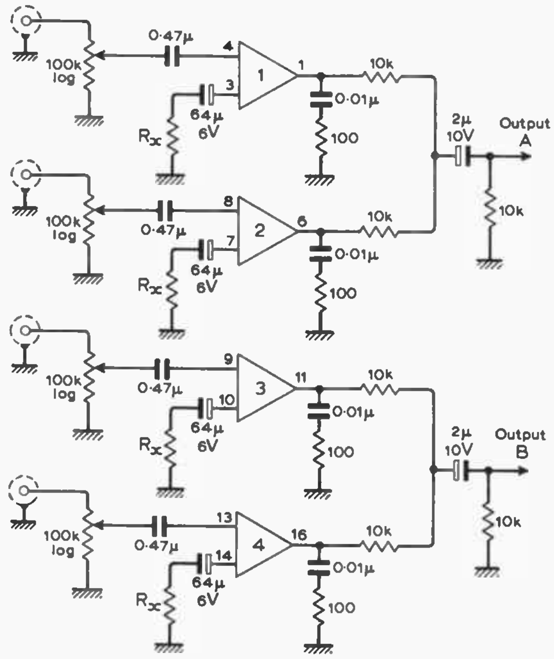


Fig. 25. A current-switching oscillator.



R_x	Stage gain (dB)	Mixer gain (dB)
∞	12	6
5·6k	20	14
2·2k	26	20
1k	30	24
560	36	30
390	40	34
56	50	44

Fig. 26. Circuit of a linear mixer.

Superimposition was at one time a common facility on good quality recorders. However, this is extremely unsatisfactory as each recording erases to some extent the high frequency information of the previous recordings. By rearranging the track-switching and making use of the mixer and the logarithmic meter, signals may be superimposed by recording from one track to another. This allows the quality of the initial signal to be maintained through several superimpositions. For this two switches replace S_2, one for record and the other for replay.

Discrete component version

Some constructors may prefer to build discrete amplifiers in place of the integrated version recommended and described in parts 1 and 2. This could be suitable for a mono record-only machine where all replay equalization is performed in the pre-amplifier.

A discrete-component replay amplifier is shown in Fig. 27 and the circuit values for equalization are given in Table 6.

Transistors Tr_{13-15} form a direct-coupled triple with a mid-band open-loop forward voltage gain of around 80 dB; the closed loop

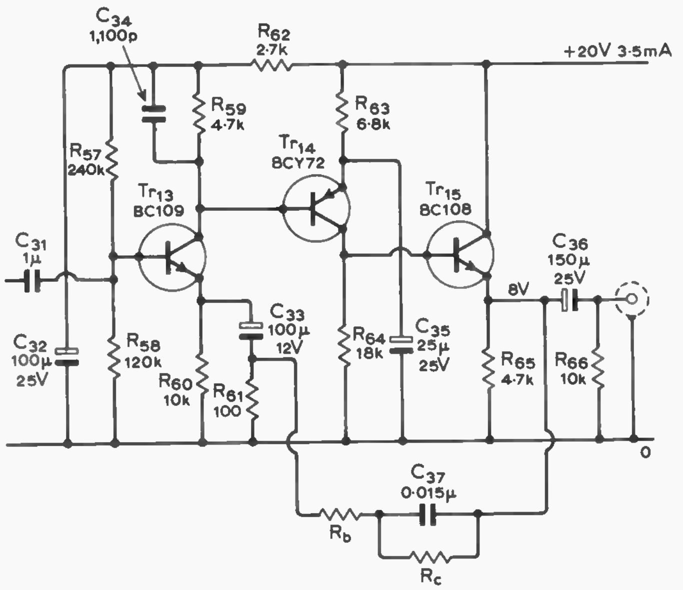


Fig. 27. A discrete-component replay amplifier.

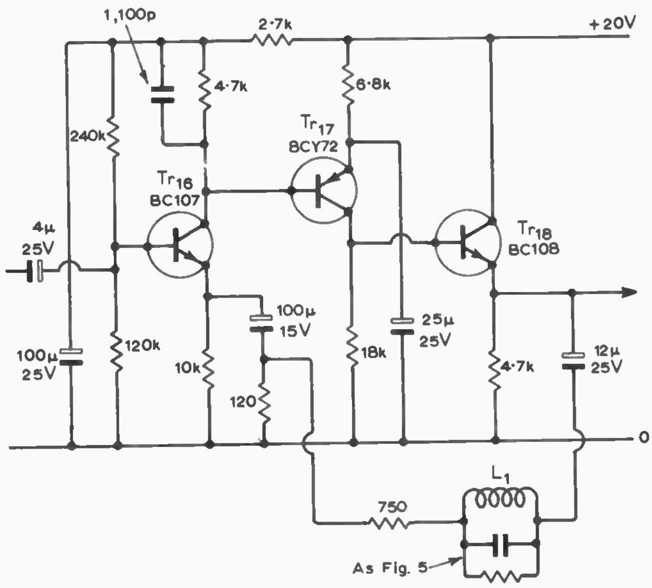


Fig. 28. Recording equalized amplifier.

gain has been arranged to give an output of 250 mV r.m.s. at the rated input, with a signal-to-noise ratio of 70 dB.

Capacitor C_{31} should be paper or plastic to ensure low leakage and avoid polarization of the head, C_{34} stabilizes the loop at high frequencies and the maximum output is 4·5 V r.m.s.

There should be no discernible differences between the performance of this amplifier and the integrated version.

The recording pre-emphasis pre-amplifier can be replaced by the amplifier shown in Fig. 28. This is very similiar in performance to that of Fig. 27 and the equalization components will be identical to those used in the integrated version. By replacing the equalization network with a 17 kΩ resistor the amplifier of Fig. 28 will have a forward gain of 140, to drive the meter.

Record output

It was stated in part 1, that the best method of ensuring a constant-current recording characteristic, is to include the head in the feedback loop of a high-gain amplifier. Such an arrangement is shown in Fig. 29. The performance of this circuit is excellent. Measured total harmonic distortion in the current waveform was less than 0·01% at an output of 140 μA r.m.s.

However the problem of bias rejection is considerable and it is strongly recommended that only an experienced constructor, with access to a good oscilloscope, should attempt this type of output stage. The problem arises because the rejection must take place at an input, where only 50 mV r.m.s. bias will switch the amplifier output between the rails.

Erase and bias oscillator

Although the oscillator described in part 2 performs very well on mono or stereo, a direct method of ensuring that bias and erase current calibration is retained in all modes, is to employ a separate output stage for each erase head, synchronizing these outputs by a master oscillator.

Considerable thought was given to the output stage. Class A and B were ruled out directly because of cost, and as it is extremely

TABLE 6. Equalization details for Fig. 27

time constants μs	R_b Ω	R_c Ω
35+ ∞	2·2 k	∞
50+3180	3·3 k	220 k
70+ ∞	4·7 k	∞
50+ ∞	3·3 k	∞
140+ ∞	9·1 k	∞
90+3180	5·6 k	220 k
140+3180	9·1 k	220 k
90+ ∞	5·6 k	∞
280+ ∞	18 k	∞
120+1590	8·2 k	110 k
120+ ∞	8·2 k	∞

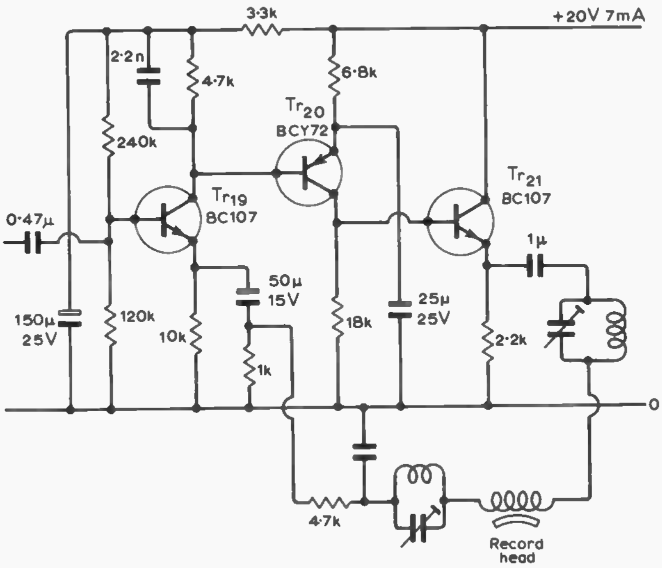


Fig. 29. A feedback recording output stage.

difficult to predict the performance of a class C amplifier, a current switching design was evolved. Fig. 30 shows an erase oscillator of this type; only one output stage has been drawn but several may be driven from the master oscillator without modification.

Transistors Tr_{22} and Tr_{23} form an emitter-coupled multivibrator which runs at 93 kHz. Tr_{24} is a buffer, the output of which is arranged to have a positive maximum a few hundred millivolts above V_{ref}.

Frequency of oscillation is stabilized by the two zener diodes, and the long-term drift is less than 0·1%. A current, defined by V_{ref} and R_m, is switched alternately between Tr_{25} and Tr_{26} and its magnitude must be arranged so that these transistors nearly saturate at the required output level.

In order that the bias waveform will decay slowly at switch-off, the time constants are arranged so that the multivibrator continues to oscillate on frequency until C_m has been discharged, allowing an exponential decay in the output current.

The transformer primary must have a high unloaded Q, and to achieve low distortion the loaded Q factor must be at least 10. The amplitude of the nth harmonic in the output for an ideal current-switching operation is

$$\frac{100}{(n^2-1).Q_l}\ \%$$

where Q_l is the loaded Q factor. A good L/C ratio is necessary to allow reasonable loading of the tuned circuit. In Fig. 30 measured values for Q were 90 unloaded and 30 loaded, however the final values are considerably affected by construction, in particular long cables connecting the oscillator to the head can radically modify the levels.

Modification for alternative heads

The designs described in these articles were intended to be used with the Brenell Mk 6 deck, which incidentally, uses the same heads as the Mk 5 range. However a large number of readers may possess perfectly good decks which have recording, replay and erase heads whose parameters are very different from the Bogen heads.

It is expected that a large variety of heads can be accommodated with a few component changes, the critical parameters for the various heads are as follows:

(a) recording—a.f. current (μA)
—bias current and voltage
—bias frequency
—inductance
(b) replay —playback level at 1 kHz, 7·5 i.p.s. and 32 mMx/mm
(c) erase —voltage and current.

In Fig. 7 the transconductance of the output stage was expressed as $1/R_{12}$. Thus the input sensitivity can be deduced for any output current, and by calculation the constructor can decide whether or not sufficient output voltage swing is available. The recording sensitivity may be deduced as the pre-emphasis low-frequency gain is 7·25.

The only modification to the replay amplifier would be to adjust the forward gain to change the sensitivity from 2 mV.

As the open loop gain of the input i.c. is only 50 dB it is not advisable to attempt to increase the closed-loop gain by more than 6 dB although it may be reduced by some 10 dB. Any further adjustment should be made in the gain stage by adjustment of R_{22}, as described in Fig. 30.

The closed loop gain G of the amplifier shown in Fig. 31 is given by

$$G = \frac{R_c}{R_d} \cdot \frac{\dfrac{R_b + R_c + R_d}{R_c} + j\omega t_1}{1 + j\omega t_2}$$

if $A \gg G$

where t_1 is the upper time constant $= C_a(R_b + R_d)$ e.g. 70 μs, 140 μs
and t_2 is the lower time constant $= C_a R_c$ e.g. 3180 μs.
The appropriate equalization values may thus be determined.

It is not so simple to calculate the component changes to the erase oscillator.

Ensure that Tr_6 and Tr_7 are allowed to saturate. If this is not the case excessive power will be dissipated probably resulting in device failure. Beware also of raising the supply voltage above 15 V as the theoretical peak collector potential could be $\pi \times$ supply voltage.

Mono and four-channel

To construct a single channel version of the recorder it is necessary only to re-arrange the i.c. amplifiers for one i.c., and to modify the erase oscillator. The author suggests that i.c. amplifiers 2 and 3 be used for the replay section and 1 and 4 for recording pre-emphasis and meter circuits. A block diagram is given in Fig. 32. For those wary of modification, the erase oscillator can be built in standard form with C_{26} and R_{56} permanently wired in. See Figs 3 and 15. Otherwise R_{56} may be omitted, along with one bias winding, and the circuit operated from a lower supply—around 7 V.

Only one bias chain will be used in the meter; thus R_{28}, R_{30} and D_2 are omitted, and the current will be set to 3 μA by R_{29}.

At the time of writing the author knows of no source of decks fitted with four-track heads, for four channel recording, however it is straight forward to multiply the circuitry to cater for this—at any point in the future the replay and recording amplifiers can be duplicated, but the erase oscillator must be replaced by a design similar to Fig. 30, or by a more powerful version of Fig. 15. There are no strong arguments for re-arranging the i.c.s. The CA 3048 lends itself to a four channel cassette replay system, although at present no deck of suitable quality is available.

The author thanks Brenell Engineering Ltd, for valuable assistance given during the development of this recorder.

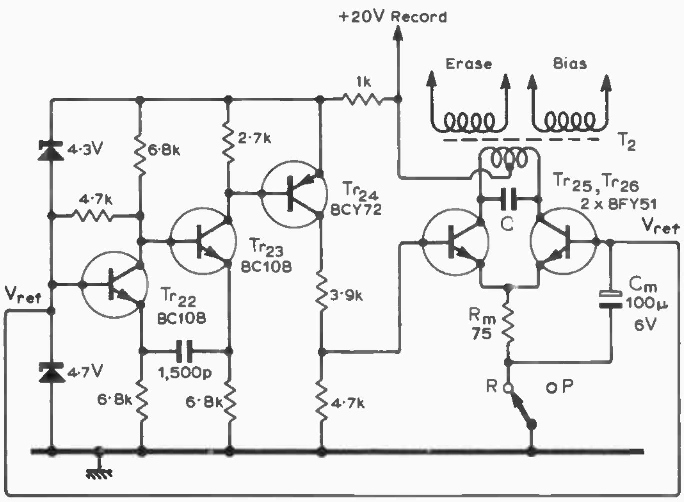
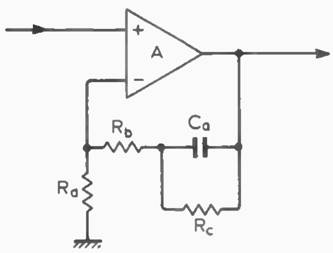


Fig. 30. Circuit diagram for an erase-bias oscillator.



$$G = \frac{A}{1 + A\beta} \simeq \frac{1}{\beta} \text{ when } A \gg G$$

$$\beta = \frac{R_a(1 + j\omega C_a R_c)}{R_c + (R_b + R_a)(1 + j\omega C_a R_c)}$$

Fig. 31. Replay amplifier equivalent circuit.

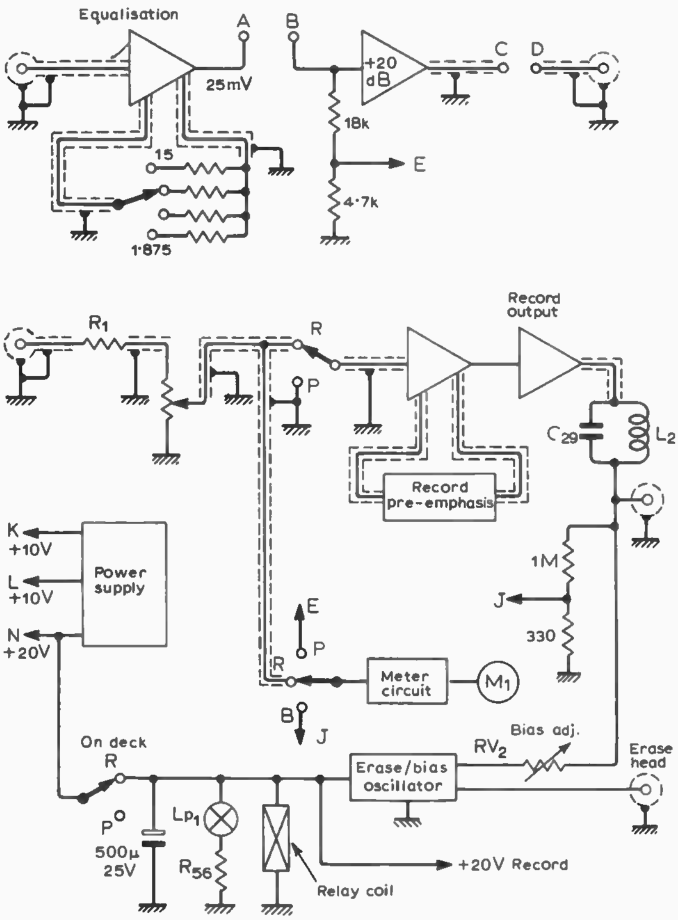


Fig. 32. Block diagram for mono.

Turntable Design for Home Construction

by R. Ockleshaw

In three articles the author describes a turntable, pickup arm and wow and flutter meter for home construction. The turntable, believed to be the first complete design for home construction and described in this issue, has a rumble level of −36dB relative to 1cm/s r.m.s. recorded velocity and peak wow and flutter of ±0.25%. Ready-turned parts are available for those without access to a lathe. Detailed drawings show how the parts are made and assembled and the article also shows how the mechanical filtering system is derived. Cost of the turntable and pickup arm is between £20 and £25. The second article will describe the construction of the pickup arm and the third will show how to check turntable performance and describes a novel wow and flutter meter using a phase-locked loop.

Although several designs for pickup arms have been published, I do not know of a constructional project that included a turntable. Perhaps the reason for this is obvious—it is mechanics on the grand scale, not normally suitable for the amateur with a limited range of tools.

To produce certain parts for this project within a satisfactory tolerance a lathe has to be used, but I have been careful to ensure that lathe-work can be accomplished on medium-sized or even small machines, the type most likely to be available. Provided one has the basic ability to use a lathe, gaining access should be easy. Model societies may have one—almost certainly some of the members—local schools or colleges that run evening metal-work courses might be persuaded to allow the occasional use of a lathe, or may even encourage it. Those that do may charge a nominal fee. There are only three parts for which the use of a lathe is essential. Use of a lathe for certain other parts simplifies their manufacture, but if made by other means will only affect the finish and not the performance. (Ready-made turned parts can be obtained from the address given in the parts list.)

Filtering wow and rumble is discussed towards the end of this article where it may be better appreciated in the light of practical knowledge.

Styling is functional and in keeping with modern design, and performance, in relative terms, should satisfy all but the most critical. Although the design contains most features normally desired in a 'transcription' unit it is essentially simple both in concept and in detail.

The unit features a 10-in diameter machined cast-aluminium turntable driven by a self-starting synchronous motor through a resilient rubber belt. There is a choice of two speeds using a simple manual change. The matching pickup arm is protected against vibration and acoustic pick-up by integral mechanical filters. A pickup-arm lowering device is featured. Provided instructions are followed, wow and flutter will be 0.5% pk-pk and rumble −36dB relative to a recorded velocity of 1cm/s r.m.s.

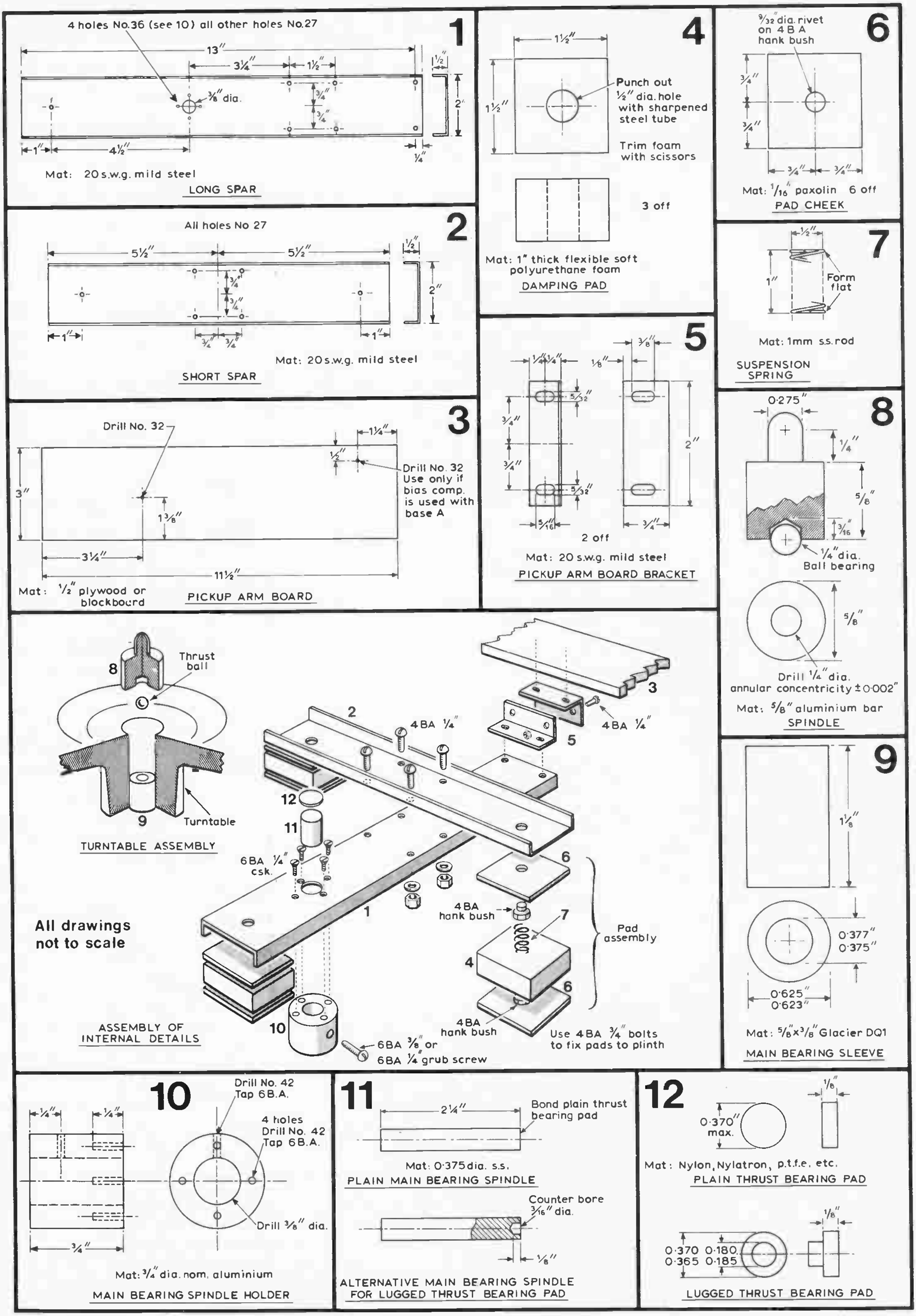
Construction

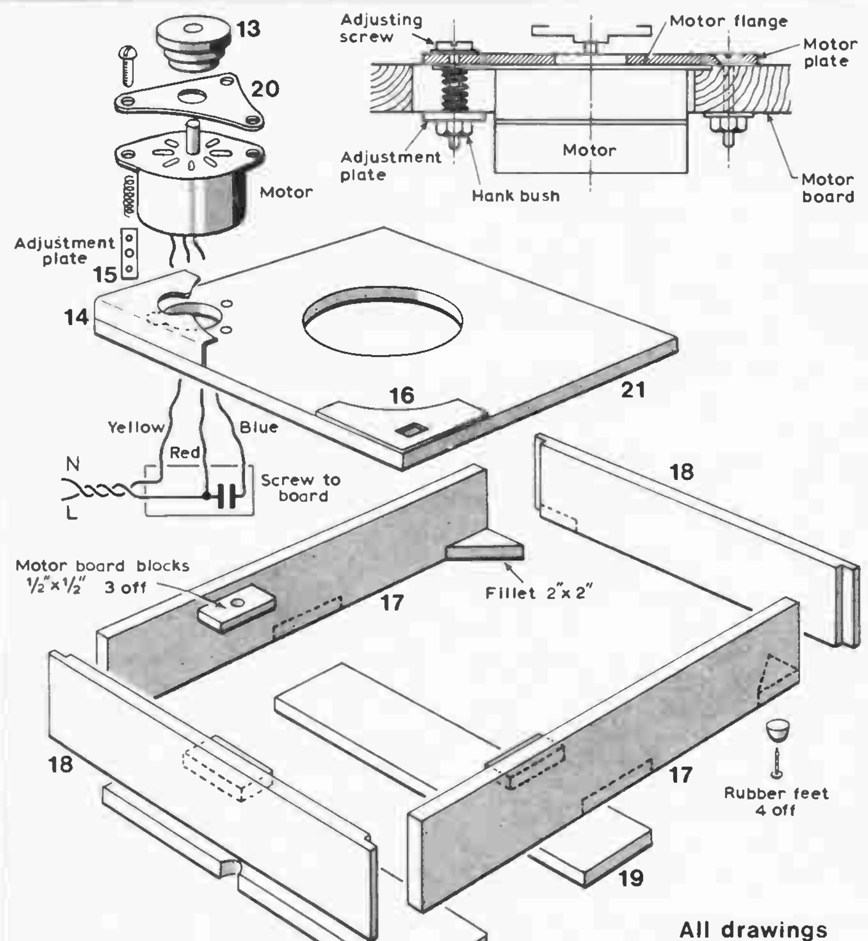
Motor board. The motor board should be to hand when assembling the plinth as it can be used as a jig to ensure that the plinth is square. Make from ½-in plywood, or blockboard, the apertures being cut by jig-saw, coping saw, etc. In following the accompanying diagrams, note that the area around the aperture for the switch body is recessed to accommodate the switch mounting plate and screws. Veneer or paint the top surface. Other ideas are matt-black Formica, or if you paint the plinth, a contrasting colour. Finish also the exposed edge adjacent to the pickup-arm board.

Plinth. Make the three duplicated parts in the plinth from ½-in high-density plywood to the dimensions indicated and assemble as shown together with fillets and blocks. Use a good-quality synthetic glue (Evode Resin W, Cascamite, etc.) for the joints, these being held by veneer pins (use motor board as jig) while the glue sets. Rubber feet, obtainable from hardware or do-it-yourself shops, can be screwed or glued underneath. Finish as desired.

Next, make the motor plate, adjustment plate and pickup-arm board. Although a captive nut (hank bush) is specified for the adjustment plate, in practice a shakeproof washer and nut may be a useful alternative. Evostik impact adhesive can be used to fix the motor to the motor plate.

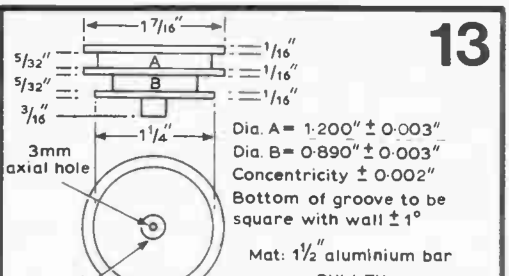
Turntable. The 3-lb turntable is a faceplate for an industrial sanding machine. It is produced by Picador Engineering Co. Ltd, a well-known firm of tool makers, and can be obtained through tool shops. It is essentially a sandcasting and may have slight casting flaws when received but these will be generally of little consequence. Complete the turntable by inserting the main bearing sleeve (see later) thrust bearing assembly (see later) and turntable mat. As a cheaper alternative to the mat, glue six rubber pips (mine came from a moulded rubber car mat) onto the turntable—preferably into drilled recesses.


1
4 holes No.36 (see 10) all other holes No.27
13″
3¼″
1½″
⅜″ dia.
¾″
¾″
½″
2″
1″
4½″
¼″
Mat: 20 s.w.g. mild steel
LONG SPAR
2
All holes No 27
5½″
5½″
½″
¾″
¾″
2″
1″
1″
¾″
¾″
Mat: 20 s.w.g. mild steel
SHORT SPAR
3
Drill No. 32
1¼″
½″
Drill No. 32
Use only if
bias comp.
is used with
base A
3″
1⅜″
3¼″
11½″
Mat: ½″ plywood or
blockboard
PICKUP ARM BOARD
4
1½″
1½″
Punch out
½″ dia. hole
with sharpened
steel tube
Trim foam
with scissors
3 off
Mat: 1″ thick flexible soft
polyurethane foam
DAMPING PAD
5
⅜″
¼″
¼″
⅛″
5/32″
¾″
¾″
2″
5/32″
5/16″
¾″
2 off
Mat: 20 s.w.g. mild steel
PICKUP ARM BOARD BRACKET
6
9/32″ dia. rivet
on 4 B A
hank bush
¾″
¾″
¾″
¾″
Mat: 1/16″ paxolin 6 off
PAD CHEEK
7
½″
1″
Form
flat
Mat: 1mm s.s. rod
SUSPENSION
SPRING
8
0·275″
¼″
⅝″
3/16″
¼″ dia.
Ball bearing
⅝″
Drill ¼″ dia.
annular concentricity ±0·002″
Mat: ⅝″ aluminium bar
SPINDLE
9
1⅛″
0·377″
0·375″
0·625″
0·623″
Mat: ⅝″x⅜″ Glacier DQ1
MAIN BEARING SLEEVE
8
Thrust
ball
9
Turntable
TURNTABLE ASSEMBLY
2
4BA ¼″
3
4BA ¼″
5
12
11
6BA ¼″
csk.
1
6
4BA
hank bush
7
4
6
Pad
assembly
4BA
hank bush
10
6BA ⅜″ or
6BA ¼″ grub screw
Use 4BA ¾″ bolts
to fix pads to plinth
All drawings
not to scale
ASSEMBLY OF
INTERNAL DETAILS
10
Drill No. 42
Tap 6B.A.
¼″
¼″
4 holes
Drill No. 42
Tap 6B.A.
Drill ⅜″ dia.
¾″
Mat: ¾″ dia. nom. aluminium
MAIN BEARING SPINDLE HOLDER
11
2¼″
Bond plain thrust
bearing pad
Mat: 0·375 dia. s.s.
PLAIN MAIN BEARING SPINDLE
Counter bore
3/16″ dia.
⅛″
ALTERNATIVE MAIN BEARING SPINDLE
FOR LUGGED THRUST BEARING PAD
12
⅛″
0·370″
max.
Mat: Nylon, Nylatron, p.t.f.e. etc.
PLAIN THRUST BEARING PAD
⅛″
0·370 0·180
0·365 0·185
LUGGED THRUST BEARING PAD

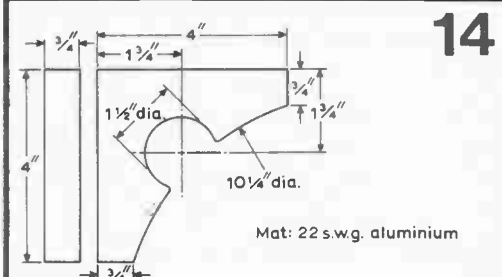


ASSEMBLY OF PLINTH

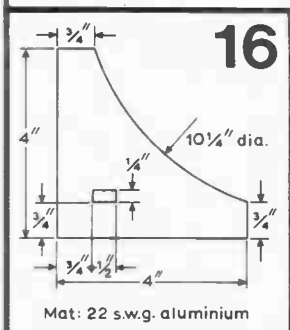
All drawings not to scale



PULLEY

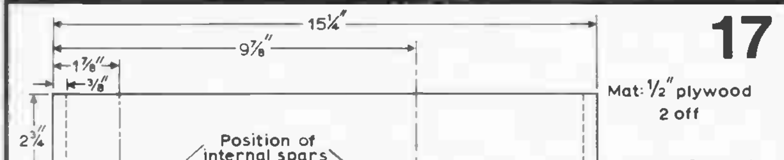


PULLEY COVER

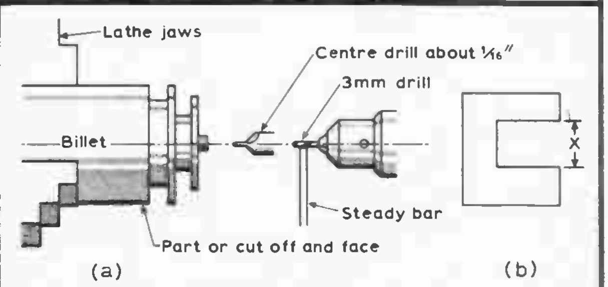


FACIA PLATE

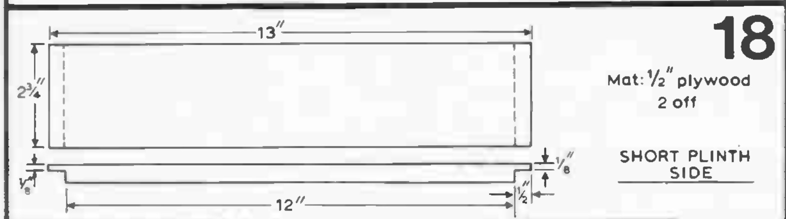
ADJUSTMENT PLATE



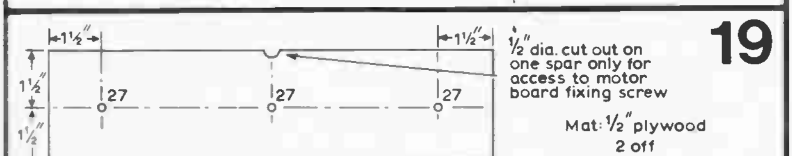
LONG PLINTH SIDE



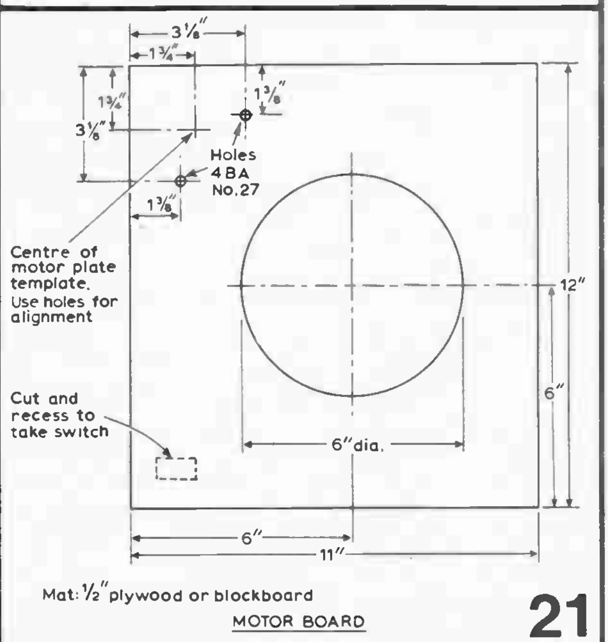
TURNING THE PULLEY



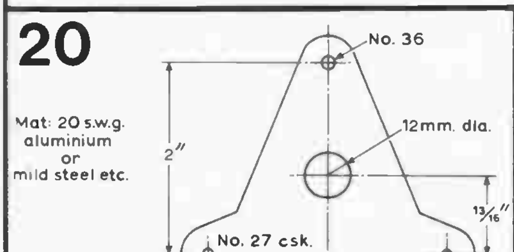
SHORT PLINTH SIDE



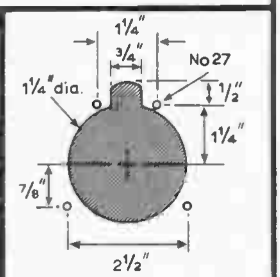
PLINTH SPAR



MOTOR BOARD



MOTOR PLATE



MOTOR PLATE TEMPLATE

(The inner three pips in the photograph are for 7-in discs.)

Main bearing spindle. Silver steel is supplied to a high degree of tolerance: ±0.0005-in of the nominal size. It is ground to this accuracy giving a finish considered more suitable for bearing surfaces than a turned one. After cutting to size, face the ends on a lathe. If a lathe is not available, rough file one end and grind with a rolling jig and oil stone.

Harden the spindle by heating to cherry red and then quenching in water. Polish with fine emery cloth and finally metal polish while rotating the spindle in the lathe or drill chuck. Carefully preserve the spindle surface prior to hardening as any attempt to remove blemishes by polishing may prove futile. These blemishes will cause excessive bearing wear. Round off top edges to prevent the bearing sleeve being scored when placing the turntable over it.

It is more important that the top surface of the nylon thrust-bearing pad is square than the top of the spindle. In this case bond the nylon thrust bearing pad to the top of the spindle with Araldite after hardening and grind square if necessary.

Main bearing sleeve. The main bearing sleeve is made from a p.t.f.e.-compound dry bearing material. This retains most of the desirable features of p.t.f.e. but is much easier to machine. The material is quite soft and convenient to use, if a trifle dirty, but care should be exercised to prevent bearing surfaces being damaged. This compound, in common with many plastics, has a high coefficient of linear expansion. Indeed, p.t.f.e. at 20°C exhibits a so-called phase change resulting in a sharp dimensional increase, which may be used to advantage. The bearing material is supplied as a tube with nominal $\frac{3}{8}$-in inside and $\frac{5}{8}$-in outside diameter. As the inside hole is a little too small, use successive reamers to bore the material to correct size. Hold the reamers in a tail-stock chuck (the work in the main chuck of course) and rotate the work by hand or at least at a very low speed. Then carefully turn down the outside of the bearing sleeve to the correct size.

Fit the bearing sleeve into the turntable. At this stage the main bearing spindle may still not fit the hole or it may be too stiff. This may be because of the squeezing action of a tight fit of the sleeve into the turntable. In this case carefully run-through the reamer to bring it to the correct size. If the tolerance of the main bearing spindle is on the high side this still may not be sufficient to give a perfectly free-running bearing. (Test at normal room temperatures, not straight from the lathe or after being held in the hand.) To give increased bearing clearance place the turntable together with the fitted bearing sleeve in a refrigerator for a few minutes, followed by reaming. Use a reamer in reasonable condition—it is important that the surface should be free from scores. Always feed the reamer with a rotating action.

Thrust-bearing spindle. Construction of the thrust-bearing spindle is more straightforward. No tight tolerances apply except that it is as well to check that the spindle fits the hole in the middle of a new record as these rapidly wear. The thrust ball may be held in the assembly by overswaging with a centre punch or with Araldite. The thrust-bearing ball should rest on the exact centre of the nylon thrust bearing pad to reduce rumble. No difference in performance should result if this part is made in other materials, like nylon or p.v.c.

Long and short spars. Bend the two spars from 20-s.w.g. mild steel (galvanized or passivated if possible) to give adequate stiffness, If a bending machine is not available careful work with a mallet and vice should be adequate remembering that this part is unseen and finish is unimportant.

Pulley. The pulley is possibly the most difficult part to make. The tolerances given must be strictly adhered to if performance is not to be impaired. Measuring the internal diameter of the grooves when turning the pulley is facilitated by using a simple gauge (see drawing 'Turning the pulley'), the groove being too narrow for a normal micrometer. (Set dimension X with a template.) To start the hole in the pulley, use a centre drill and ensure the drill does not deviate inside the pulley by using a bar to steady the drill.

The pulley cover (drawing 14) can alternatively be made without the angled sides, instead using $\frac{1}{2}$-in thick wood for the two sides (as in the photograph). By making the height slightly greater than the pulley height, the $1\frac{1}{2}$-in dia. cut can be avoided.

Damped suspension pads. For the suspension pads, three springs are required, made by winding 13in of 1-mm silver-steel wire on a $\frac{1}{4}$-in dia. former. After removing from the former, even out the spring and form as shown. Compress with fingers until it 'bottoms'. After removing pressure it should be 1-in long. To give a clean $\frac{1}{2}$-in hole in the pads to take the springs, punch out using a sharpened piece of thin-walled $\frac{1}{2}$-in stainless-steel pipe (as used in the lifting device). Make the pads larger than necessary and trim on assembly. Use Evostik for bonding.

Assembly

Assemble the main bearing spindle with its bottom flush with its holder. Use the grub screw to lock it in place temporarily as adjustment will follow. Place the suspension cruciform comprising the long and short spars, with the pickup-arm board fixed by brackets, in the plinth before screwing on the motor board from underneath. Additional pad cheeks can be used as packing to make the turntable top parallel with the motor board.

Assemble the bonded motor plate and motor to the motor board together with the adjustment plate and spring. (The spring must be strong enough to allow the end of the motor plate to rise above the board.) Screw the capacitor holder to the underside of the motor board.

Wire the motor to the switch before screwing the motor board to the plinth, earthing the motor casing. Assembling the pulley to the motor may require a little persuasion as it has been designed as an 'interference' fit. Heat the pulley in hot water for a few minutes—expansion should then allow a fit. Do not ream the hole to size and do not unnecessarily pull off the pulley once fitted as this may cause enlargement of the hole and consequent slipping. Lower or raise the main bearing spindle to align the top of the pulley with the top of the turntable.

Designing to avoid rumble

Rumble generally is generated by two sources in a turntable unit—the motor and turntable main bearing—and may be described as noise the spectral content of which lies within the range of about 10 to 200Hz. Apart from a comparatively small amount of noise due to mechanical displacement in the motor bearings, its contribution arises from 'stepping' or 'cogging'—the tendency to rotate in discrete steps rather than in a uniform way. If a synchronous motor is held loosely in the hand and allowed to rotate this 'cogging' vibration is felt quite strongly. Thus if the motor is coupled to the turntable it must always be through some kind of mechanical vibration filter—for example a resilient belt or rubber-tyred wheel.

Unfortunately it is more difficult to mechanically filter rumble generated by the turntable main bearing. This kind of rumble is generally random (except perhaps when the turntable is blessed with ball bearings) unlike the discrete motor vibrations that are related to mains frequency. It is caused by imperfections on the bearing surfaces.

Motor rumble

In this design, vibrations from the motor can be transmitted to the turntable by two paths—the drive belt and the motor mountings—and a mechanical filter is necessary in both paths (Fig. 1).

The electrical equivalent—Fig. 2—shows the motor as a two-dimensional

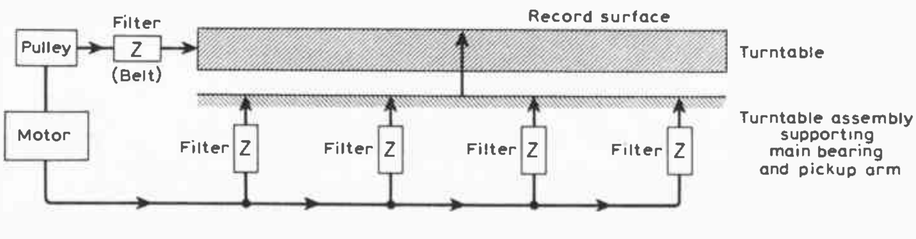


Fig. 1. Rumble from the motor is transmitted to the turntable by two paths (motor mounting and pulley) which must be separately filtered.

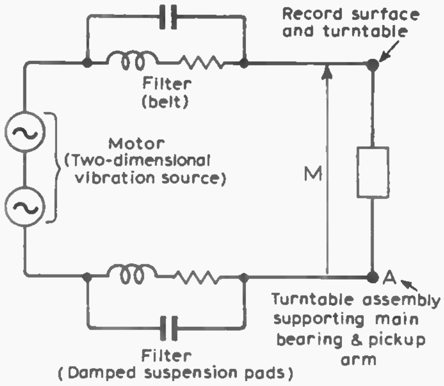


Fig. 2. Electrical equivalent of Fig. 1. Motor produces rumble through the belt in a lateral plane and through the mounting in a vertical plane. The lateral vibration can produce vertical vibration if the turntable is not stiff enough.

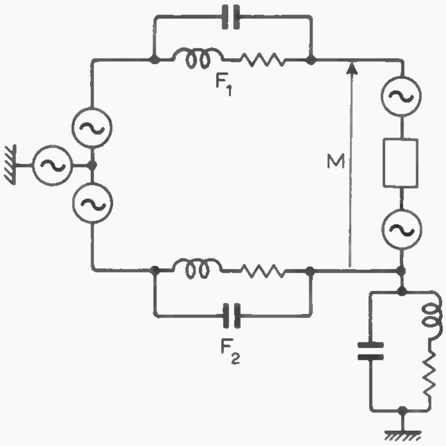


Fig. 5. In the belt-driven unit, F_2 can be made effective enough to eliminate F_3 and the motor casing can be grounded.

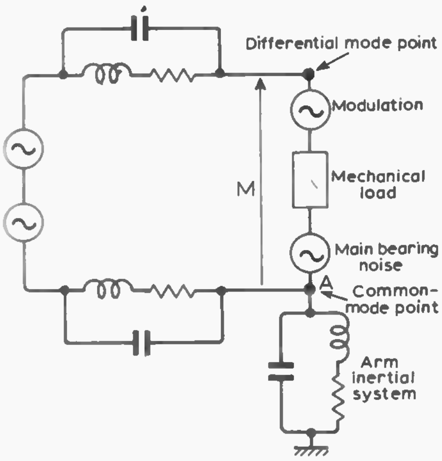


Fig. 3. Inertial grounding effect of pickup arm creates common and differential-mode points. Common-mode response can be eliminated by grounding point A.

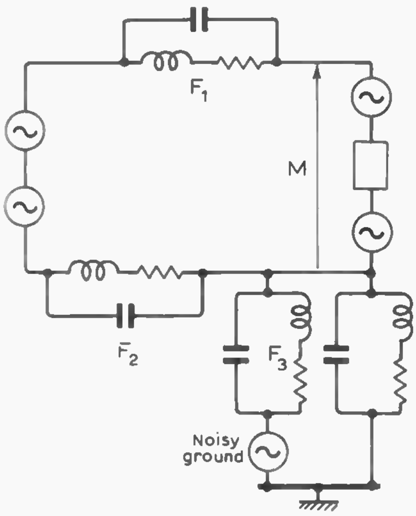


Fig. 4. Mounting a turntable unit on a 'noisy ground'—a non-rigid shelf for example—introduces noise and one solution is to use a 'lockable in transit' type of spring mounting for F_3.

vibration source. Relative to the ground plane, vibration through the turntable mountings will be in a vertical plane while that from the belt will be predominately in a lateral plane. However, this lateral vibration can give rise to vertical vibrations if the turntable is not sufficiently stiff or adequately damped.

The main responses of a stereo cartridge will be at 45° to the plane of this vertical displacement and will thus reproduce a component of any vertical displacement relative to the pickup-arm mounting point. This is why it is always desirable to mechanically couple the pickup arm as close to the turntable as possible to reduce as much as possible any differential movement.

But, however much we may reduce the differential displacement, the common-mode displacement may still have an effect on the pickup output because the pickup is essential an inertial system and will therefore have a common-mode response. What response it does have obviously depends on the design of the pickup arm.

We therefore should modify the electrical model to that of Fig.3 and it is obvious that any common-mode response can be eliminated by grounding (literally!) at point A. Unfortunately, if the ground is noisy due to the unit being placed on a shelf, then the pickup will produce an output in sympathy with the displacement. This is one common cause of acoustic feedback.

Many turntable units are grounded at this point but the manufacturers are always careful to ensure that effective filtering eliminates any displacement that might excite the common-mode response of the pickup arm Fig. 4.

The design of belt-drive units is not so mechanically restrictive as some jockey-wheel-driven types, and allows the use of more efficient filters for F_1 and F_2 (Fig. 4)—the belt and turntable suspension pads respectively—and it becomes possible to combine F_2 and F_3 and ground the motor casing (mechanically). The resulting, chosen, arrangement is shown in Fig. 5.

Main bearing rumble

The prime cause of bearing rumble is imperfection of bearing surfaces. The

Parts List

All parts, including cover, for both pickup arm and turntable are available as raw materials or ready-turned where appropriate from Longdendale Technological Products.

Part	description/material	source
motor	Berger RSM 50/8	Longdendale Technological Products (LTP)
turntable	Picador 10-in sanding disc $\frac{5}{8}$-in shaft	most good tool shops
belt	rubber	LTP
main bearing sleeve	Glacier DQ1 $\frac{5}{8} \times \frac{3}{8}$-in nom.	Glacier Bearing Co. Ltd.
main bearing spindle	$\frac{3}{8}$-in dia. silver steel	sold in most tool shops & engineers' suppliers in 13-in lengths
suspension springs	1-mm dia. silver steel	" " "
motor alignment spring	$\frac{3}{4} \times \frac{1}{4}$-in o.d.	LTP
damping pads	1-in thick polyurethane foam	upholstery dealers
long and short spars	20-s.w.g. mild steel	ferrous metal dealers
turntable mat	rubber	Metrosound, C. Watts radio hobby shops
pad cheeks	$\frac{1}{16}$-in Paxolin	radio hobby shops
pulley	$1\frac{1}{2}$-dia., 2-in long aluminium bar	non-ferrous metal dealers
facia plate, pulley cover, motor plate, adjustable slate	20-s.w.g. aluminium	non-ferrous metal dealers
switch	2-pole c/o slider	Radiospares etc.
thrust bearing pad	$\frac{1}{4}$-in dia. $\times$ $\frac{1}{2}$-in long nylon	LTP
thrust ball	$\frac{1}{4}$-in dia. ball bearing	cycle shops, motor accessory shops
spindle	$\frac{5}{8}$-in dia. p.v.c. rod $\times$ $1\frac{1}{2}$-in	LTP
main bearing spindle holder	$\frac{3}{4}$-in dia. $\frac{3}{4}$-in long aluminium rod	non-ferrous metal dealers

Miscellaneous

Captive nuts (hank bushes, six 4BA one 6BA), screws (six 4BA $\frac{1}{4}$-in cheesehead, four 4BA $\frac{3}{4}$-in countersunk head, four 6BA $\frac{1}{4}$-in cheesehead, six 4BA $\frac{1}{2}$-in cheesehead), nuts, rubber feet. From usual sources or LTP.

obvious approach is to make them as perfect as possible. This is quite reasonable but there is a limit and mechanical filtering techniques must be used to reduce the effect further.

It is well known that the less dense a material the greater is the attenuation to the passage of sound and noise. Here in lies the key. The bearings should be made of a material with a density that is as low as possible. The energy then generated, which can be quite high, suffers a great deal of attenuation in its passage to the stylus. The reproduction of rumble can also be affected by resonance phenomena in the pickup arm, about which more later.

Most modern plastics fall into the low-density category but not all are suitable bearing materials. Of those that are, nylon and p.t.f.e. are most common. Unfortunately p.t.f.e. is virtually impossible to machine, nylon is difficult but machinable, and p.v.c. while not an ideal material from the wear viewpoint is easier still to machine. Better still are some compound materials that have p.t.f.e. as a base. They retain all or most of the desirable properties but are easy to machine. They are known under proprietary names like Glacier DQ1, used in this design.

Wow and flutter

Wow is caused by slow variations in record speed, flutter by fast variations. Like rumble, it cannot be eliminated completely, merely reduced to an acceptable level. In the simple arrangement of a slow-speed motor directly driving a revolving turntable, wow and flutter could be caused by sticky main bearing, belt slip, motor cogging and pulley eccentricity. (Wow can also be caused by a badly eccentric or warped record but this is outside our control.) But note that only one of these points really indicates a design problem, that is motor cogging, and this is really tied up with rumble. If the belt is an efficient filter this source of flutter is eliminated.

The most usual way of preventing wow and flutter is to prevent the remaining three imperfections occurring due to bad manufacture, or dirt and grease being smeared on the belt, and to provide a turntable with a high inertia. Care must be taken that any turntable inertia is not overcome by too tight a coupling of the drive motor. This most certainly would occur if a synchronous motor was used with a rim-drive jockey-wheel system. The speed of a synchronous motor is fixed and does not depend on the load as with an asynchronous motor.

Pickup Arm Design for Home Construction

by R. Ockleshaw

The pickup arm described is designed to accompany the turntable detailed in our last issue. It includes an optional bias compensator and lift mechanism. Mechanical resonance is damped by a flexible coupling between counterweight and arm. A further article will describe how to check performance of the turntable using a test record and novel wow and flutter meter.

Design of pickup arms has been well described. The articles* published in *Wireless World* May and June 1966 contain all the information required to design an arm for minimum distortion due to lateral tracking errors. In the present design, note has also been taken of the opinions of J. Walton on pickup-arm design.†

Briefly, one should try to avoid a system reproducing frequencies generally below the limits of audibility, because they may produce a disturbing Doppler effect on some loudspeaker systems whose acoustic impedance at these frequencies is low.

*J. K. Stevenson, 'Pickup arm design', *Wireless World* vol. 72 1966 pp. 214-8 and 314-20.
†J. Walton, 'Turntable rumble and pickup arm design', *Wireless World* vol. 68 1962 pp. 435-7.

Also, vibrations of the turntable and pickup-arm suspension should not cause excitation of the pickup arm, however damped.

A pickup arm has a natural period of oscillation of $T = 2^{\pi}(MC)$ where M is the effective mass of the pickup arm and C is the compliance of the pickup cartridge. Mechanical impedance moves from a low to a high value around the resonant frequency peak—Fig. 1. Below the resonant frequency, because the mechanical impedance of the arm is low in comparison with the mechanical impedance of the pickup cartridge armature, the output from the pickup will be severely attenuated. Thus the arm acts like a high-pass filter, rejecting frequencies in the rumble range. The cut-off can be quite sharp but its value as an active part of a system is lost if different cartridges of varying compliance are fitted. Consequently my approach is that it is always better to ensure that rumble is reduced as much as possible at source and not rely entirely on the impedance characteristics of the arm. Damping the resonant peak is important too as the coincidence of some discrete vibration with the high-impedance resonant peak of an undamped arm may cause excitement which could damage the disc groove. This design is damped by ensuring that the counterweight is flexibly coupled to the arm. This effectively spoils any modes of mechanical resonance.

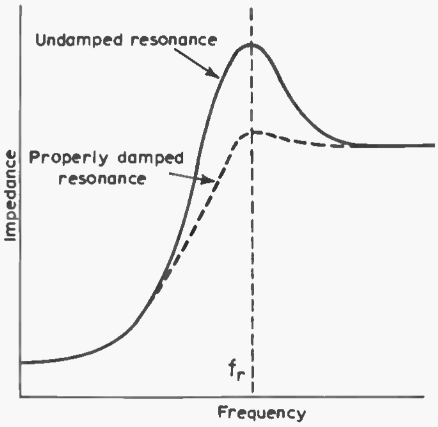


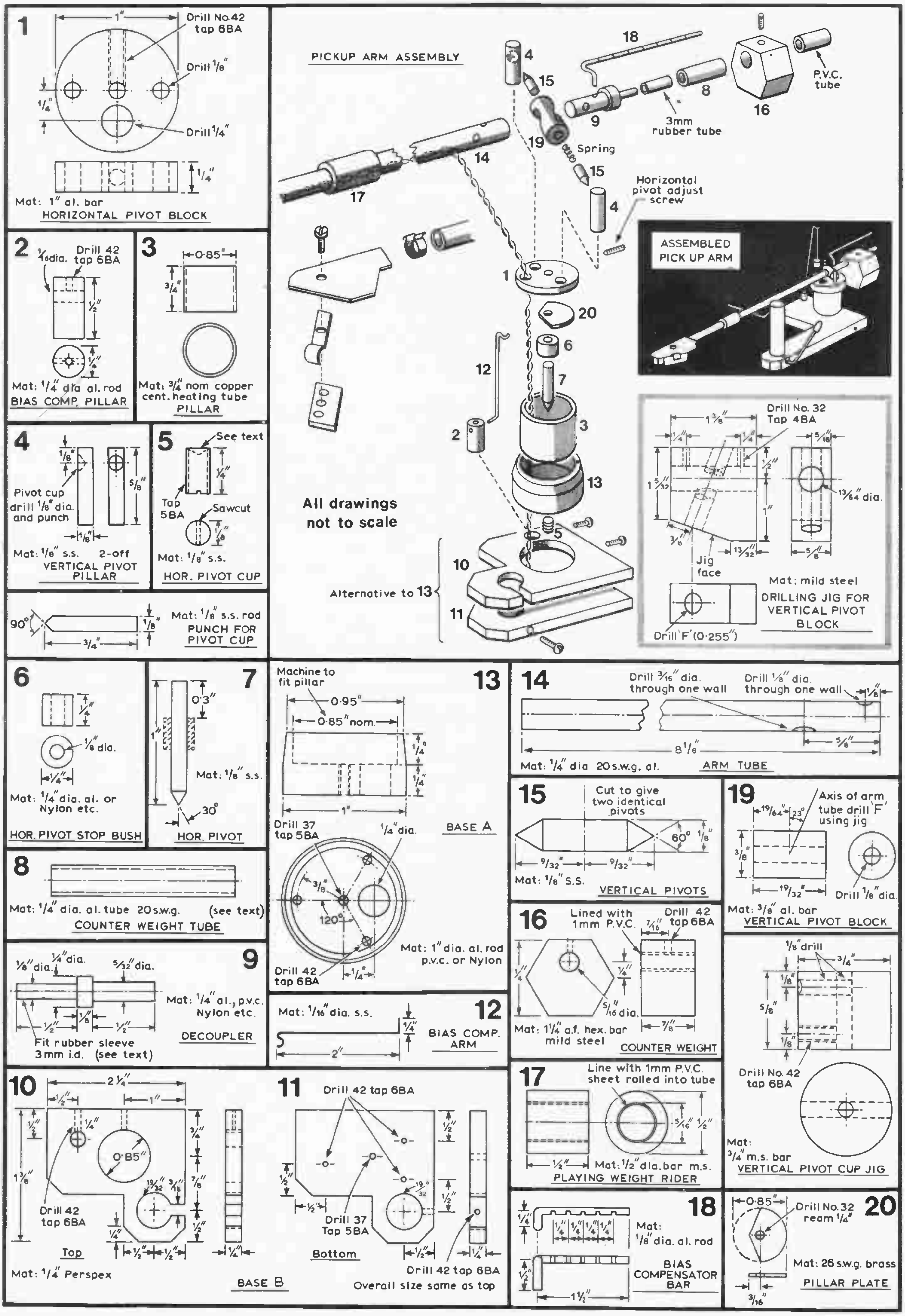
Fig. 1. Pickup arm resonance must be damped to allow for different cartridges. In this design damping is achieved with plastic 'decoupling' between balance weight and arm.

Record warp causes large vertical pickup-arm movements and it is important that the stylus remains normal to the record surface. Making the vertical pivot axis normal to the axial line of the cartridge, as in this design, gives a better approximation to correct movement than making the axis normal to the whole arm.

Construction is described in the drawings and in the supplementary notes which follow. The material for the counterweight is steel, but this can be replaced by any high-density material such as brass—though the dimensions may have to be changed to maintain the correct weight. When making the decoupler, which fits into the counterweight tube, ensure the wide end is a comfortable push fit into the arm tube. Fit a 3-mm internal dia. rubber sleeve over the smaller end and push into the counterweight tube, checking that the tube does not touch the decoupler.

The vertical pivot block is drilled at an angle to accept the arm tube. This is a difficult operation in practice without the aid of a jig and so a suitable design is shown. The material required is a 1-in length of ¼-in dia. aluminium bar which is inserted

In this photograph, the pickup arm has a different shell to that shown in the diagrams. A drawing showing how to make this version — heavier, though possibly aesthetically more acceptable — is available from the editorial offices.



1
Drill No.42 tap 6BA
Drill 1/8"
Drill 1/4"
Mat: 1" al. bar
HORIZONTAL PIVOT BLOCK
2
1/16 dia.
Drill 42 tap 6BA
Mat: 1/4" dia al. rod
BIAS COMP. PILLAR
3
0·85"
Mat: 3/4" nom copper cent. heating tube
PILLAR
4
Pivot cup drill 1/8" dia. and punch
Mat: 1/8" s.s. 2-off
VERTICAL PIVOT PILLAR
5
See text
Tap 5BA
Sawcut
Mat: 1/8" s.s.
HOR. PIVOT CUP
90°
Mat: 1/8" s.s. rod
PUNCH FOR PIVOT CUP
PICKUP ARM ASSEMBLY
18
P.V.C. tube
16
3mm rubber tube
Spring
Horizontal pivot adjust screw
ASSEMBLED PICK UP ARM
All drawings not to scale
Alternative to 13
Drill No. 32 Tap 4BA
13/64" dia.
Jig face
Mat: mild steel
DRILLING JIG FOR VERTICAL PIVOT BLOCK
Drill 'F' (0·255")
6
1/8" dia.
Mat: 1/4" dia. al. or Nylon etc.
HOR. PIVOT STOP BUSH
7
0·3"
Mat: 1/8" s.s.
30°
HOR. PIVOT
13
Machine to fit pillar
0·95"
0·85" nom.
Drill 37 tap 5BA
1/4" dia.
BASE A
120°
Mat: 1" dia. al. rod p.v.c. or Nylon
Drill 42 tap 6BA
14
Drill 3/16" dia. through one wall
Drill 1/8" dia. through one wall
8 1/8"
Mat: 1/4" dia 20 s.w.g. al.
ARM TUBE
15
Cut to give two identical pivots
60°
Mat: 1/8" S.S.
VERTICAL PIVOTS
19
Axis of arm tube drill 'F' using jig
23°
Drill 1/8" dia.
Mat: 3/8" al. bar
VERTICAL PIVOT BLOCK
8
Mat: 1/4" dia. al. tube 20 s.w.g. (see text)
COUNTER WEIGHT TUBE
9
1/8" dia.
1/4" dia.
5/32" dia.
Mat: 1/4" al., p.v.c. Nylon etc.
DECOUPLER
Fit rubber sleeve 3mm i.d. (see text)
12
Mat: 1/16" dia. s.s.
BIAS COMP. ARM
16
Lined with 1mm P.V.C.
Drill 42 tap 6BA
5/16" dia.
Mat: 1 1/4" a.f. hex. bar mild steel
COUNTER WEIGHT
1/8" drill
Drill No. 42 tap 6BA
Mat: 3/4" m.s. bar
VERTICAL PIVOT CUP JIG
10
0·85"
Drill 42 tap 6BA
Top
Mat: 1/4" Perspex
BASE B
11
Drill 42 tap 6BA
Drill 37 Tap 5BA
Bottom
Drill 42 tap 6BA
Overall size same as top
17
Line with 1mm P.V.C. sheet rolled into tube
Mat: 1/2" dia. bar m.s.
PLAYING WEIGHT RIDER
18
Mat: 1/8" dia. al. rod
BIAS COMPENSATOR BAR
20
0·85"
Drill No.32 ream 1/4"
Mat: 26 s.w.g. brass
PILLAR PLATE

into the jig. Lock it into position by two 4BA screws. Using an F(0.255in) drill, pierce the aluminium bar by inserting the drill into the hole in the jig face with the jig held in a vice. After piercing, shorten the pivot block to the dimensions given.

A jig is also used to make the vertical pivot pillars. Hold the pillar in the jig while preforming the cup with a ⅛-in dia. drill. The pillar should not be removed from the jig, however, before the pivot cup is formed using the punch shown. Heat the punch to cherry red, quench and polish. After punching, likewise harden the pivot cups. Form the horizontal pivot cup in the same way, harden both pivot and cup, and finally polish the pivot.

Two versions of pillar base are shown. Use version A—best made on a lathe—if the lift mechanism is not required. Base B accepts both the lift mechanism and bias compensator pillar. Bond the two parts of base B after they have been made with Evostik and spray if desired.

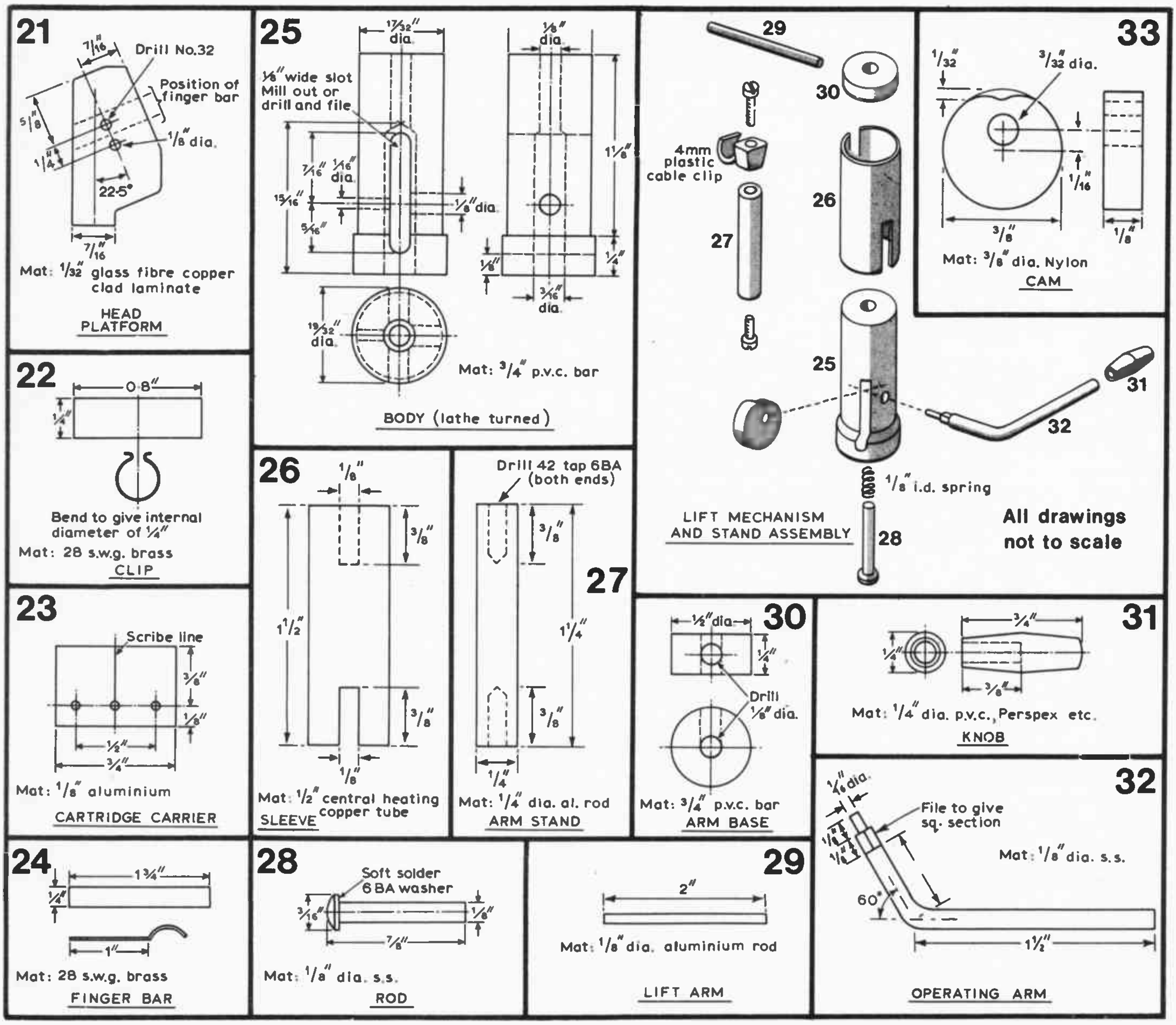
Assembly

Once the vertical pivot block and decoupler are assembled on and in the arm tube

Parts list

All turntable and pickup arm parts are available from Longdendale Technological Products.

part	description/material
arm tube	¼-in dia. × 20 s.w.g. Al tube (12-in)
vertical block	⅜-in dia. × 1½-in Al bar
decoupler	¼-in dia. × 1-in Al bar
bias compensator bar	⅛-in dia. × 3-in Al bar
horizontal pivot block & base A	1-in dia. Al bar (2-in)
horizontal pivot stop	¼-in dia. Al bar
horizontal pivot and cup, vertical pivot & pillar	⅛-in dia. silver steel (13-in)
pillar plate, finger bar & clip	28 s.w.g. brass or copper
pillar	¾-in dia. nom. copper central-heating tube (2-in)
head platform	$\frac{1}{16}$-in copper-clad laminate
base B	¼-in Perspex sheet
counterweight	1¼-in a.f. mild steel hex. bar (1-in)
cartridge carrier	⅛-in Al
playing weight rider	¾-in mild steel bar
vertical pivot loading spring	from Longdendale Technological Products
bias compensator pillar	¼-in dia. Al rod
bias compensator arm	$\frac{1}{16}$-in dia. s.s. (13-in)
socket-head grub screw	6BA × ¼-in (6 off)
pickup-arm wire nylon thread	about 18-in
bias compensator weights	appropriate lengths of ¼-in dia. brass rod
lift mechanism	
body & arm base	2-in × ¾-in p.v.c. bar
lift and operating arm & rod	⅛-in silver steel (7-in)
cam	⅜-in nylon
sleeve	½-in nom. copper central heating tube (1½-in)
spring	⅛-in i.d. × ¼-in long from Longdendale Technological Products
knob	¼-in dia. p.v.c., Perspex etc. (1-in)



respectively, use the vertical pivot block as a jig to complete the $\frac{1}{8}$-in dia. axial hole through the arm tube and decoupler. A small amount of Araldite or Evostik ensures a permanent assembly. Now insert the spring and two pivots into the axial hole of the pivot block as shown.

Bond the vertical pivot pillars into the pivot holder with Araldite with the cups accurately aligned inwards. After setting, insert the vertical pivot block between the pillars by squeezing the pivot loading spring in the pivot block over the pivots. This is a tricky operation requiring a little patience and, hopefully, only one spring! The resulting pivot should be completely free from sticking and quite stable.

Bond the horizontal pivot-stop bush to the horizontal pivot after it has been hardened and polished. Insert the square-cut end through the $\frac{1}{8}$-in hole in the pillar plate. Assemble the base to the pillar.

Fix the vertical pivot pillar holder on to the horizontal pivot by the grub screw. Screw the horizontal pivot cup to the pillar base until the bush tightens against the top of the pillar. Slacken off $\frac{1}{4}$ turn and lock with cellulose paint. Adjust the vertical pivot block to give a clearance of about 0.025in.

Wiring should present no problem if it is done before the arm is fitted to the pickup-arm board. Remember to mark one of the wires at both ends for identification. It may help if a piece of stiffer wire is threaded first so it can be used to pull both of the coaxial wires through at once. The two wires can be terminated on a small tagboard underneath the pickup-arm board or on to a plinth-mounted socket.

Performance of the arm is improved by using the bias compensator. Possibly the best way of setting up the compensator, for a spherically-tipped stylus at least, is with an unmodulated disc. But be prepared for some experimentation.

Setting up the arm

A jig for assembling the arm to the pickup-arm board is shown in Fig. 2. It should be used with the turntable in place, the small hole being placed over the spindle. The other end should be slipped over the pickup-arm pillar. The arm's position should then be selected and marked.

Effective arm length should be nine inches — i.e. the distance from stylus tip to centre line of vertical pivots. To do this slide the head of the arm either forward or backward along the arm tube. The overhang is designed to be 0.625in and is measured as the distance the stylus overhangs the centre of the turntable. Using the adjusting screw on the head, adjust offset angle to give zero tracking angle—i.e. angle of stylus to groove at a distance of 2.4in (2.375) from the turntable centre and then at a distance of 4.6in (4.606) from the turntable centre. There should be very little difference in tracking angle. If it is discernible check the positioning of the arm base, the effective length and overhang.

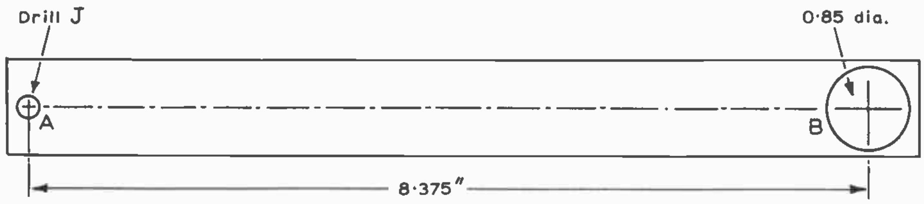


Fig. 2. When turntable and pickup arm are assembled place hole A over spindle and hole B over pickup arm pillar. Draw round the base to mark selected position.

Calibration

The playing weight rider can be omitted, in which case the playing weight must be set up each time using a suitable balance. If the rider is used the arm can be calibrated against either a 'pressure' gauge or a set of weights. In either case stick a piece of plasticine to the cartridge platform. Its weight is not important but it should be roughly equal to the weight of a cartridge—say 6 or 7g.

If you use a pressure gauge, adjust the counterweight to balance the arm with the rider as close to the pivots as possible. Moving the rider away from the pivots will unbalance the arm and increase the playing weight. Relate distance from the pivots to playing weight using the pressure gauge.

If you use weight, stick four 1-g weights to the plasticine (assuming a maximum playing weight of 4g). Adjust the counterweight to balance with the rider close to the pivots. Remove one of the weights and move rider away from pivots to re-balance. Mark the arm. Repeat this procedure removing one weight at a time until all have been removed. Half-gram markings can be inserted by interpolation as the scale will be linear.

F.M. Stereo Tuner

High-performance design for home construction

by L. Nelson-Jones, F.I.E.R.E.

In recent years there have been a number of developments in the components field, particularly in semiconductor devices, that have led to great improvements in the design possibilities for f.m. broadcast receivers. In particular these have been the advent of the dual-gate m.o.s.f.e.t., integrated-circuit i.f. amplifiers and demodulators, ceramic filters and improved variable-capacitance diodes. This two-part article describes an f.m. tuner design using these devices, discusses the advantages of the devices and gives constructional and alignment details. It does not attempt to be all-embracing and there will doubtless be some who disagree with the author's views of the current scene. It is hoped however that they do show some ways in which f.m. tuner design is currently evolving.

The work is the result of many months of design and measurement on five prototypes, so that results are not based on a one-off, and should be reproducible by readers who wish to copy the design. The receiver was designed to achieve in a relatively simple way a performance equal to the better examples of the commercial models available, but at a much lower price. (Total material cost comes out at about £11.) Comparison with the figures given in a recent *Wireless World* survey of commercial tuners (September 1970) suggest this aim has been achieved. The performance of the tuner under normal conditions of use has been excellent. One of the units is in use in Blandford Forum in Dorset—very much a fringe area—and gives noise-free reception from the Isle-of-Wight transmitters, including the new local station Radio Solent.

The design for the front-end of the f.m. receiver is shown in Fig. 1. Both r. f. amplifier and mixer stages use dual-gate f.e.ts with gate protection diodes. In the r.f. stage the upper gate is decoupled and acts as a screen between drain and gate 1, much as the g_2 electrode of a thermionic valve does. In the mixer stage this second gate is used as the injection point for the local oscillator voltage. There is not the same need for a screen between drain and gate 1 in a mixer stage as the drain load is not tuned to either the signal or oscillator frequencies. There is therefore little or no gain at signal frequencies to cause oscillation provided care is taken with the layout, particularly the length and placing of leads.

The magnitude of the local oscillator injection at gate 2 will affect the mixer gain and the spurious signal response characteristics of the mixer stage. This local oscillator voltage will be higher than in transistor tuners using bipolar devices by up to an order of magnitude and for the circuit conditions used a value of

Fig.1. Front-end of receiver using dual-gate f.e.ts.

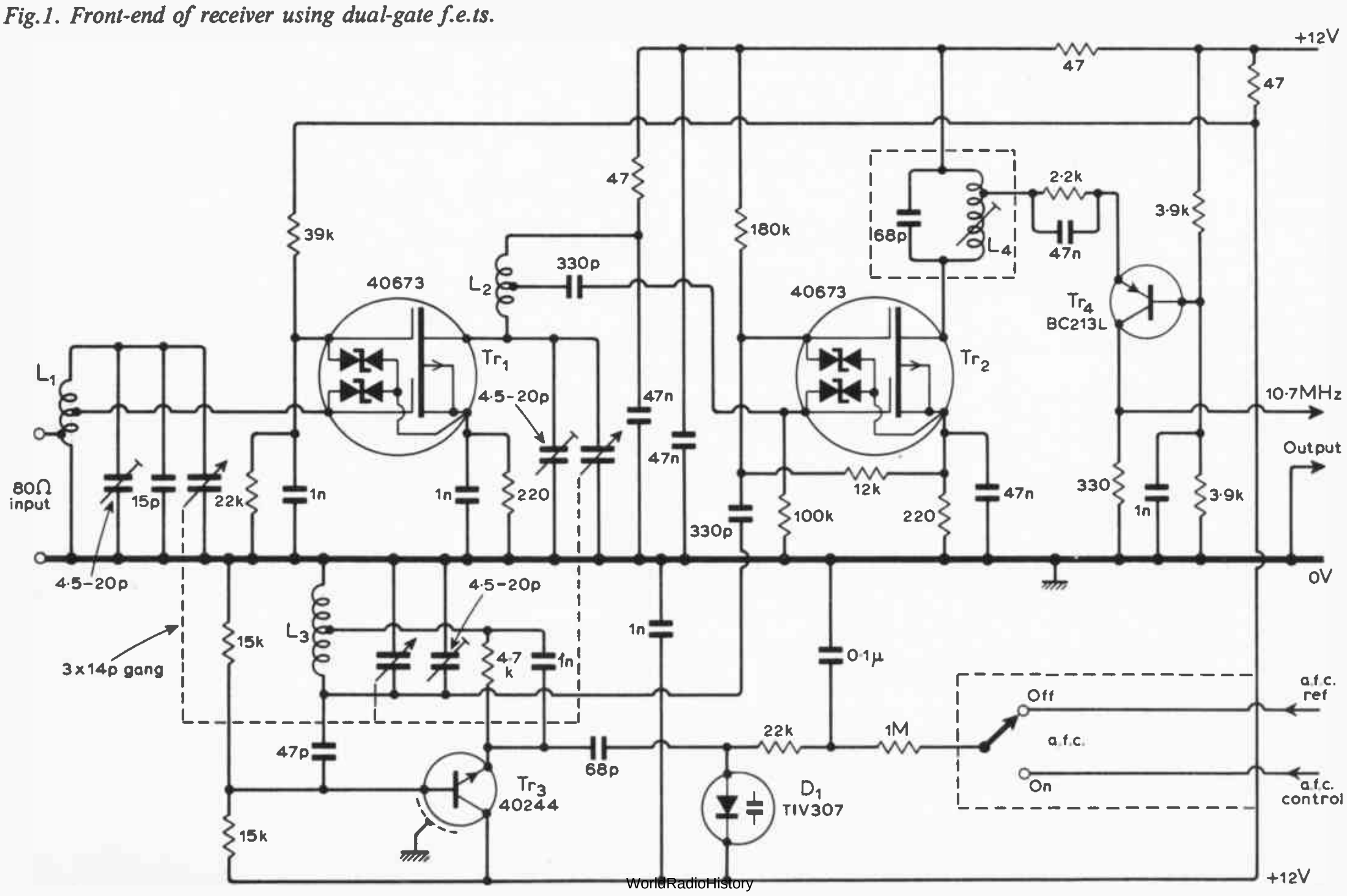
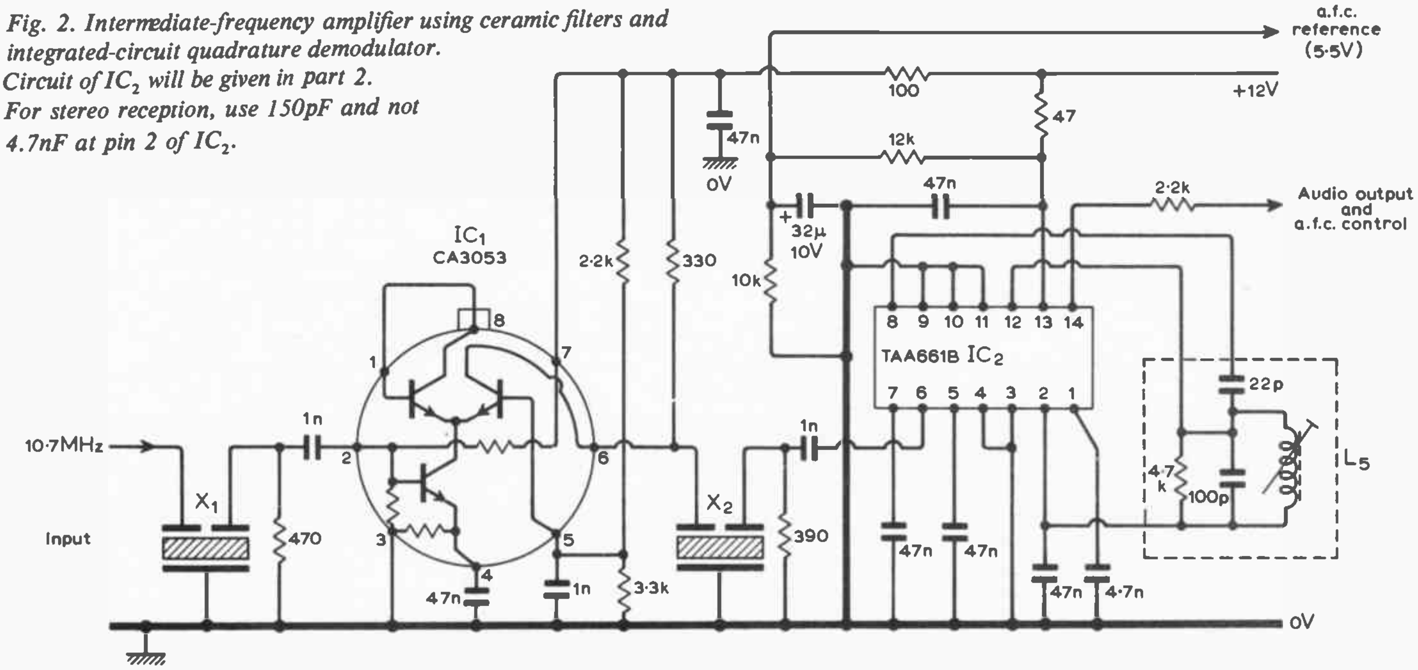


Fig. 2. Intermediate-frequency amplifier using ceramic filters and integrated-circuit quadrature demodulator. Circuit of IC_2 will be given in part 2. For stereo reception, use 150pF and not 4.7nF at pin 2 of IC_2.



500mV r.m.s. gives a reasonable mixer gain without too high a spurious signal response. In fact higher levels have been used without great trouble from spurious signals. Far greater trouble can be caused by lack of screening, leading to i.f. harmonics being picked up by the front end—especially with the high sensitivity of this tuner and, because of its small size, the close proximity of the front-end and the i.f. amplifier.

The oscillator is a conventional Hartley circuit with the ground point moved to give a grounded-collector design. There is no particular advantage to be gained in using an f.e.t. in this stage so that the cheaper bipolar device is preferred. Automatic frequency control is applied by the variable-capacitance diode D_1 coupled to the emitter of the oscillator stage. A resistor of 22 kΩ prevents the decoupling capacitor of 0.1μF from shorting out the oscillator voltage. The 0.1-μF capacitor together with the 1 megohm resistor form a low-pass filter to prevent audio voltages in the a.f.c. voltage from reducing the modulation of the carrier by audio frequency modulation of the local oscillator.

The a.f.c. can be switched out of operation by connecting the diode to a constant reference voltage from the i.f. amplifier. Diode D_1 is returned to the 12-volt supply line of the oscillator so that the a.f.c. control voltage changes the diode reverse voltage in the correct direction to reduce any oscillator drift. An increase in local oscillator frequency increases the intermediate frequency, which in turn leads to a rise in the output potential of the demodulator of the i.f. integrated circuit IC_2 (Fig. 2). As the diode is connected to the +12V supply, this increase reduces the reverse bias across D_1, increasing the diode capacitance and reducing the local oscillator frequency to correct its drift.

The mixer has a grounded-base stage feeding the 330-ohm resistor needed to correctly terminate the first filter unit X_1 (Fig. 2). This resistor also makes a convenient low-impedance output point from the front-end. A cheap p-n-p bipolar device is more than adequate for this position, because in a grounded-base configuration the requirements in respect of high frequency or noise performance are not stringent. The working Q of the tuned circuit is less than 20 so that tuning is not critical, and it is set to maximize gain in the usual way.

The supply for gate 2 of the r.f. stage is derived from the decoupled oscillator supply rather than from the top of L_2. This is brought about purely by layout convenience on the printed wiring board and, as gate 2 is additionally decoupled, has no effect on the performance.

I.F. amplifier

Two ceramic filter units X_1 & X_2 are separated by a buffer amplifier (IC_1) of moderate gain (about 20dB). The reason for this moderate gain is that it is desirable to place the filters as early as possible in the i.f. amplifier so that as successive stages limit with increasing signal level there is no change in bandwidth. This would be fully achieved if the whole of the i.f. gain were after the filter, and provided the mixer did not limit.

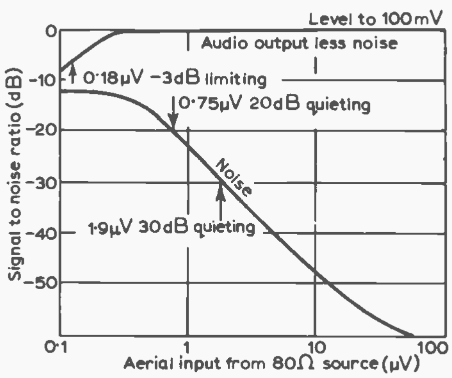


Fig. 3. Graph of low-level performance shows sensitivity of 0.75μV (for ±75kHz) for 20dB quieting. Above 50uV input, signal-to-noise ratio is better than 60dB.

In practice it is not possible to achieve this ideal, but the compromise of using only moderate gain before the second filter unit is a reasonable one and does not give rise to any undue increase in bandwidth over the normal range of signal levels. The performance obtained is a great improvement over normal bipolar i.f. amplifiers using discrete components with several double-tuned i.f. transformers. In such an amplifier the selectivity of the transformers is gradually lost starting at the output end as successive stages limit and the overall selectivity can leave a lot to be desired at high signal levels.

The first integrated circuit is a long-tailed pair circuit, used as a cascode amplifier by ignoring one of the top pair of transistors and driving into the long-tail transistor. The input impedance of this stage is suitable for the ceramic filter unit so far as resistance is concerned, but is above the maker's recommendations so far as input capacitance is concerned. For this reason the resistor terminating the filter X_1 is raised to 470 ohms, which compensates for the increased capacitance loading of the stage in restoring the top of the filter characteristic to reasonable flatness.

The load of the cascode stage is a 330-ohm resistor—to drive the filter X_2 from the correct source impedance. This low value results in the stage gain being low, especially when the loading effects of the filter are accounted for, so that although the slope of the cascode stage is around 100mA/V the overall gain from the output of X_1 to the input to X_2 is only a little over 20dB.

The input impedance of the IC_2, around 2 kΩ, is not very much greater than 330 ohms so that a terminating resistor of 390 ohms is used at the output of X_2. The value of the feed capacitor to the 'quadrature' tuned circuit L_5 is larger than the maker's recommended value of 18pF.

The value is not very critical in practice and up to 47pF has been used with little change in performance once the circuit had been retuned. It is likely that with large departures from the recommended value there will be an increase in distortion, but no appreciable effect is likely to stem from the increase to 22pF, a value more readily available than 18pF.

One possible reason for the apparent insensitivity to the value of this capacitor is that the quadrature drive voltage is derived from a resistive tap on the load resistor of the final stage of the i.f. amplifier of IC_2. The value of the impedance at this tap is fairly low, and as the coupling to the circuit is increased by increasing the value of the capacitor, the damping effect of this resistor increases, lowering the Q of the quadrature circuit, and compensating—at least to some extent—for the increased drive and tending to restore the correct phase relationship.

The audio output from pin 14 is taken via a 2.2-kΩ resistor to provide additional overload protection.

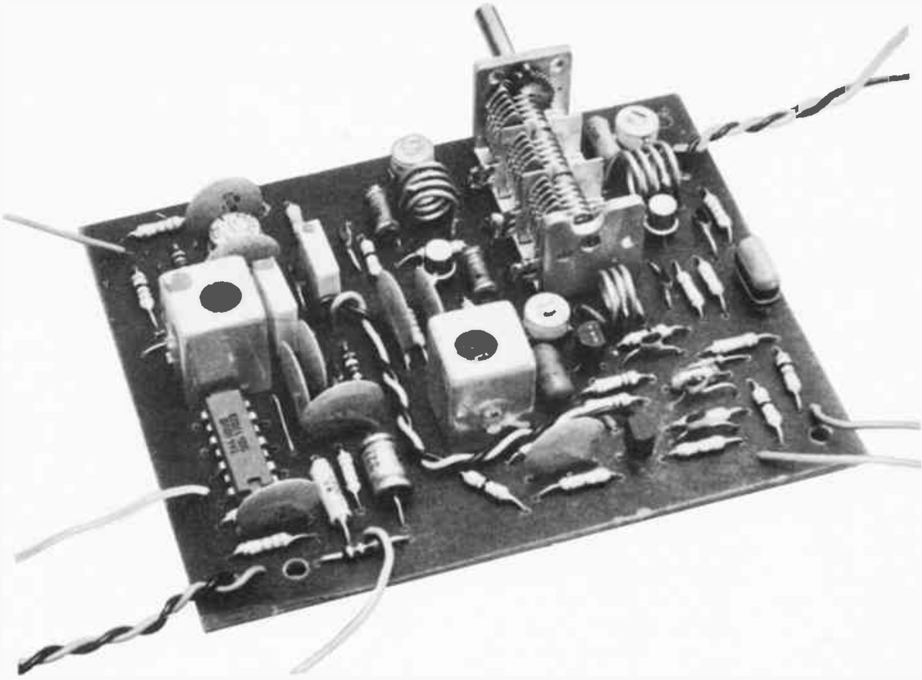
The reference used when a.f.c. is not required is derived from a potentiometer across the supply. The values chosen give a voltage close to that obtained at pin 14 of IC_2 (when there is either no signal or the signal is at centre frequency). The output level at pin 14 of IC_2 stays fairly close to around 46% of the supply voltage over the range 10-16 volts, and thus a simple potential divider is adequate for this reference voltage as this will also provide such a percentage of the supply voltage.

Because the output at pin 14 is a percentage of the supply voltage it is essential that the supply to the tuner be very well smoothed if hum and noise on the output is to be prevented. This is also important as the a.f.c. diode is returned to the 12-volt supply, and supply rail hum will therefore produce frequency modulation of the intermediate frequency whether or not a.f.c. is switched on.

The requirement for a ripple-free supply was one of the reasons for the choice of a 12-volt supply, so that adequate resistive smoothing could be used with a large reservoir capacitor. The drain of the tuner is almost constant and fairly well defined at a little under 35mA so that simple resistive dropping is satisfactory. An important point is that with such a network it is essential that the supply to the resistor be switched to disconnect the receiver, and not the supply to the receiver from the reservoir capacitor. In this latter case the capacitor would charge up to the full supply voltage with the receiver off, and on switching on the receiver would momentarily receive the full supply, which might well be enough to permanently damage the active devices of the receiver, especially IC_2. For supplies much above, say, 36 volts some form of simple stabilizer would be preferable.

Performance

The low-level performance of the tuner is shown in Fig. 3. At signals above 50μV



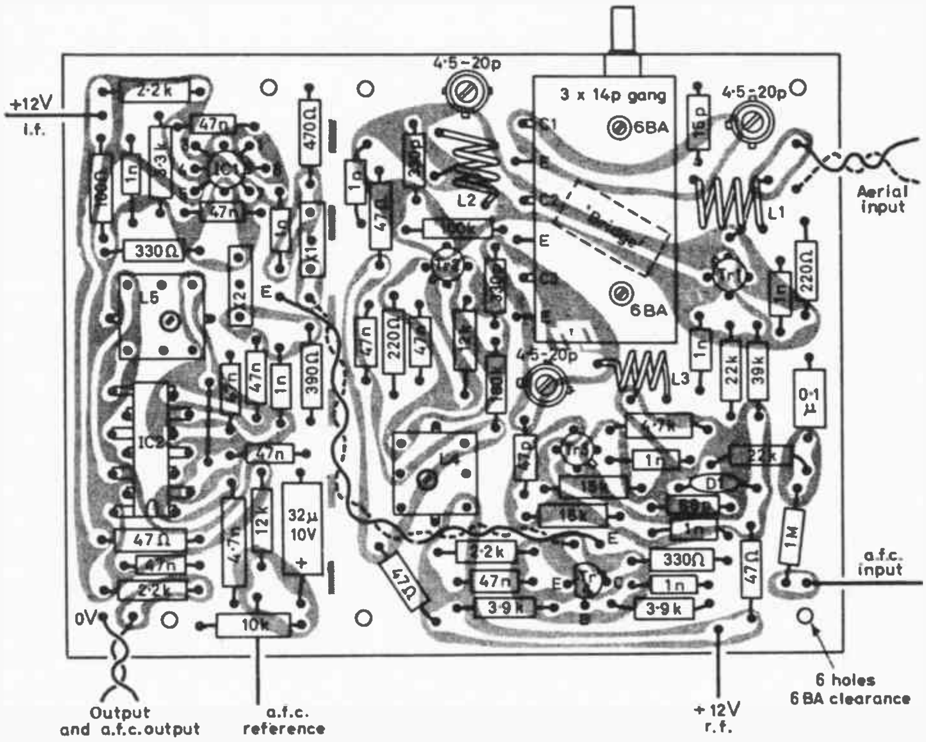


Fig. 4. Metal bridge—covering lead from Tr_1 drain to its tuned circuit—under the tuning capacitor is essential to maintain stability. Bridge, which can be tinplate, is detailed in Fig. 7. Complete tuner is screened by fitting into a die-cast box.

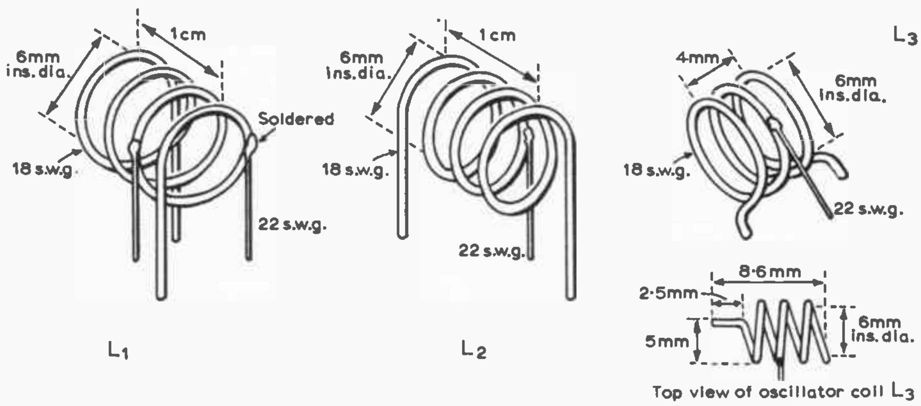


Fig. 5. Self-supporting r.f. coils, wound with slightly stretched wire, are shaped on a ¼-in dia. rod.

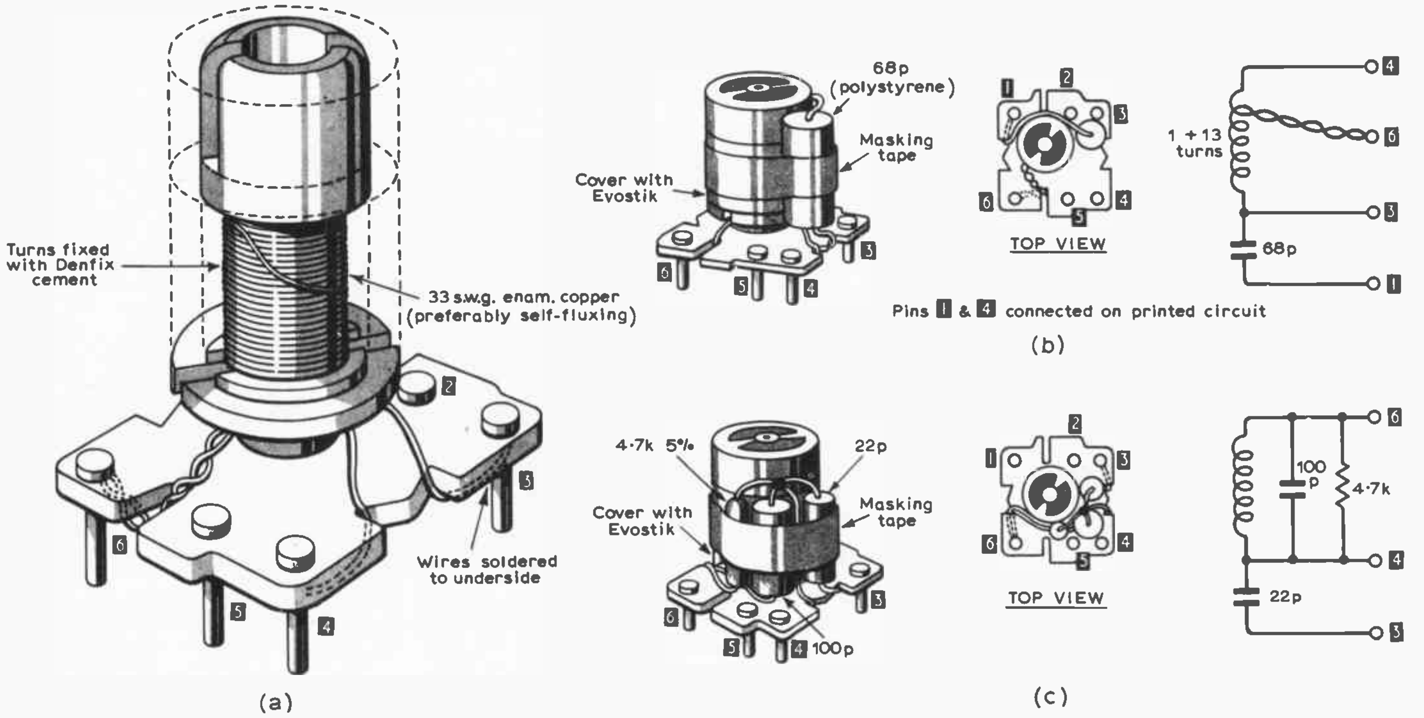


Fig. 6. Two screened i.f. coils are wound with 33-gauge enamelled wire and secured with a little Denfix cement. Capacitors are fixed with masking tape.

the signal-to-noise ratio is better than 60dB, and the 20 and 30dB quieting figures are 0.75 and 1.9μV respectively. Limiting (−3dB) of the demodulated audio signal (single-tone filtered from noise) occurs at only 0.18μV so that at all usable signal levels the i.f. section is limiting.

Unfortunately no signal generator was available which had an output with amplitude modulation-free from f.m., and it has therefore not been possible to check the a.m. rejection of the receiver. However, it is expected that the result will be close to the figure quoted for the i.f. integrated circuit (IC_2) at moderate signal levels, with an improvement when the first i.f. amplifier limits. The figure quoted for the TAA661B is 40dB at 10mV input, equivalent to around 10μV input at the aerial. Performance is summarized in the table on this page.

General layout

The tuner is constructed on a single-sided printed circuit board and is divided into two areas. The front-end and the i.f. amplifier are laid out separately, side by side, on a printed circuit board about 10 × 8 × 5cm overall, and in such a way that they may be separated if desired.

The complete tuner is enclosed in a screened box to cut down on spurious responses due to radiation from the i.f. amplifier, and to reduce local oscillator radiation to a minimum. This screening is especially necessary with this design because of its very high sensitivity.

The dial drive system suggested gives a scale length of 13.7cm with a reasonably linear frequency change over the centre 80% of this scale. A cord drive system is used which has the advantage of retaining the pointer at both ends, thus eliminating the problem of sliding friction at one end of the pointer, and giving a much smoother drive.

The overall layout of the components is shown in Fig. 4 as seen from the components side of the board, and also in the oblique view.

In construction keep leads short and if possible test all components on a bridge before fitting them on the board, as this can save ruining a good p.c. board should the tuner not work first time. It cannot be emphasized too strongly that such component checks can save much wasted money and temper. It is also vital to check that the components are correctly located, the diode connected with the right polarity, and that there are no breaks or shorts on the 'track' of the p.c. board. This latter point is of importance on such a small board with roughly 200 component holes, as tracks are necessarily fine and gaps small. A watchmaker's eyeglass has been the constant companion of the author during the construction and design of this tuner.

Coil construction

The r.f. coils are all made from 18-gauge tinned copper wire and are self-supporting. Taps are made by soldering leads of 22-gauge tinned copper direct to the turns of the coils--Fig. 5. The coils were made by winding the wire on a ¼-in rod such as a drill shank. The wire should be straightened by placing one end of a length in the vice and pulling the other end until the wire stretches very slightly, when the wire will have lost all kinks.

The coil wires should be a firm push fit in the board, if undue strain on the copper foil of the board is to be avoided when adjusting the coils. At all cost avoid the wires being loose in the board before soldering and if necessary apply the minimum of Araldite epoxy adhesive around the 18-gauge coil wires on the components side of the board after soldering in position. The joints should then be quickly reheated with the soldering iron to cause the Araldite to run into the holes in the board, thus securing the coils rigidly. After using such an adhesive the board must be left in a slightly warm place, e.g. an airing cupboard, for 24 hours to ensure that the adhesive has set hard.

The coils should be mounted on the board in the positions shown in Fig. 4, and with the turns of the coils nearest the board surface 2.5mm clear of the board. In the case of the oscillator coil this must also be 3.5mm clear of the rear face of the tuning capacitor. It is best to adjust the coils before soldering them into the circuit board to minimize subsequent adjustment, and the overall coil lengths given (over the outside of the end turns) are close to the

Performance

Sensitivity	
−3dB limiting	0.18μV
20dB quieting	0.75μV
30dB quieting	1.9μV
Spurious response	
image rejection	−70dB
i.f. rejection	−85dB
other unwanted signals	−94dB
Audio output	0.5V r.m.s. for ±75kHz
*Capture ratio**	2dB approx.

* Difficult to measure. There does not appear to be much dependence on signal level provided the signals are reasonably above noise level—a result to be expected from the very low level at which limiting starts. Figures varying from 1 to 4dB were measured at various signal strengths on repeated measurements. In general a signal 10dB below produced no noticeable effect on the demodulated output.

final adjusted lengths required for correct tracking.

An alternative method of mounting is to open up the main coil mounting holes in the board and insert 'eyelets' which are big enough to allow the 18-gauge coil leads to pass through them. The eyelets are riveted into the board so that the strain is removed from the copper track. Overall connection is obtained by soldering the track to the eyelet and to the coil leads. All the main pads for the ends of the coils are large enough to allow this to be done.

The two screened i.f. coils are both constructed on Neosid coils type NS/E3, and both are wound with 33-gauge enamelled wire, preferably of the self-fluxing variety. The two coils are wound as shown in Fig. 6 and the turns secured in place for stability with a minute quantity of Denco Denfix polystyrene cement. (Do not use modelmaker's polystyrene cement as some varieties can have high loss factors.) It is essential to use the least possible quantity as the bobbin of the coils is made from polystyrene loaded with iron dust and is very easily dissolved by this cement. When the cement has dried, place the ferrite sleeve over the coils, ensuring that the leads are well pressed down in the slot at the base of the bobbin so that this ferrite sleeve does not scratch the wires.

Next push on the polythene retaining disc. Secure the ferrite sleeve to the coil former with a smear of Evostik latex-resin contact adhesive around the join between the sleeve and the coil former near the base. When dry connect the capacitors (and the resistor in the case of L_5). Tape these components to the former as shown to hold them clear of the coil can. Make sure all leads are well clear of the can when this is slipped over the coil. Next fix the core into the coil. The ferrite core cuts its own thread into the polythene top retainer. Set the top of the core level with the top of the polythene retainer–Fig. 6b. This should be close to the final adjustment position. In the case of L_4 the capacitor is connected to the coil on one side via the printed circuit, which connects pins 1 and 4, thus placing the 68-pF capacitor across the coil.

Fitting components to the board

Due to the small size of the board and components it is absolutely essential to use a soldering iron with a small tip which is adequately hot and clean. Lead lengths must be kept short and the components close to the board. This is especially important with the ceramic disc decoupling capacitors (1 and 47nF types). The transistors should be pushed down onto the board until the body is 2.5mm above the board; pushing the transistors closer than this will strain the leads unnecessarily. This rule applies also to IC_1. The second i.c. should be placed down on the board until the shoulders on each lead contact the board; the body of the i.c. will then be just clear of the board.

There will be some difficulty in locating the polarity of the diode due to the small size of this device. If in doubt check it with an ohmmeter on the ohms × 1 range, when the cathode will be the one connected to the positive lead of the multimeter when the meter shows conduction. (On a multimeter the polarity of the leads on the resistance ranges is the opposite to that shown on the meter panel so that the red positive lead is negative, and when connected to the cathode results in the diode being forward biased.)

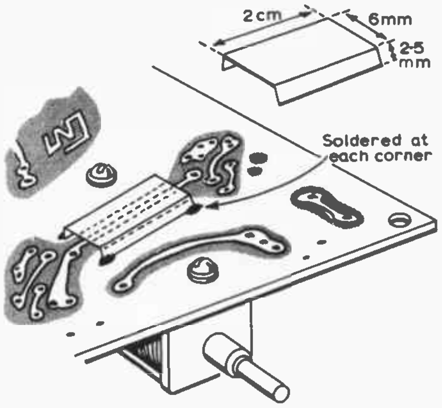


Fig. 7. Metal bridge—covering lead from Tr_1 drain to its tuned circuit—under the tuning capacitor is essential to maintain stability.

There are two wire links on the board, one on either side of L_5 in the i.f. section. These links may be either 22-gauge tinned copper or 1-024 p.v.c. covered wire. Connections to the 3-gang capacitor are by similar wires. The capacitor is secured to the board by two 6BA screws not longer than 4.5mm of thread.

The link from the r.f. to i.f. sections is by a twisted pair of 1-024 p.v.c. insulated wire as shown in Fig. 4 and in the oblique view. Take care to see that this is correctly connected, i.e. the live lead of the pair connects between the collector of Tr_4 and the input to X_1.

The two screened coils L_4 and L_5 are soldered into circuit after being pushed well down on the circuit board so that when the can is placed over the coil it just rests on the polythene retainer at the top of the coil, while also just contacting the p.c. board. The can is put on the coil after soldering the coil into the board, and the two can tags are then also soldered to the board.

There is one component not shown on the circuit because it is not a circuit component. This is a 'bridge' across the lead from the drain of Tr_1 to its tuned circuit under the tuning capacitor. This bridge continues the earth plane as well as screening the lead. It is essential to use this bridge to maintain stability in the r.f. amplifier. The bridge is necessary because of the layout limitations set by the capacitor having its connections on only one side of the body, and the need to keep the coils well spaced to obtain good stability and keep oscillator radiation low.

The dimensions and location of this bridge are shown in Fig. 7. The bridge may be made of any metal that will not corrode but will solder. Tinplate was used in the original units. Take care not to short the wire to the centre-section stator of the variable capacitor at the end of the bridge nearest to L_2.

If the ceramic filters need removing from the board, take care not to apply pressure when applying heat to the connections, otherwise the component will be damaged. Remove solder first with a desoldering tool or with copper braid.

In the photograph the alternative type of oscillator transistor is shown. If this type is used then the fourth connection on the transistor won't exist—used to earth the can on the TO-72 type specified (40244). Only the three connections nearest the 3-gang capacitor will then be used. In addition, an extra lead is shown in the photograph adjacent to the integrated circuit IC_2 that is not shown in Fig. 4. This lead connects to pin 14 of this i.c. to control the a.f.c. However, as the output lead (via the 2.2-kohm resistor) will

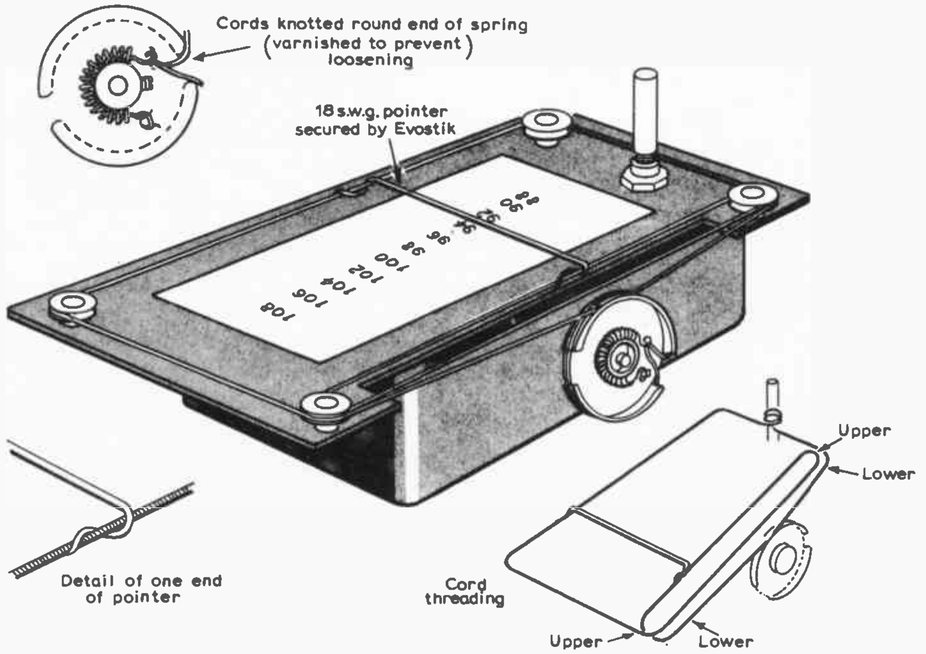


Fig. 8. Suggested cord system eliminates pointer friction and can be mounted in any plane.

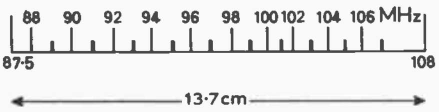


Fig. 9. Typical scale graduations for the band 87.5 to 108 MHz.

normally have a d.c. blocking capacitor in series, it will be at the same potential as pin 14, and is therefore suitable for the same purpose.

Screening of the tuner

It is essential to screen the tuner to avoid instability in the i.f. section due to pick-up, particularly from the output lead. Although the de-emphasis capacitor removes most of the 10.7-MHz signal and its harmonics, the output lead can still have sufficient of these signals present to cause spurious whistles when tuning if this lead is anywhere near the r.f. section. A great improvement results from connecting a capacitor of 470pF from the output to ground which, with the 2.2k Ω series resistor, removes these high-frequency signals sufficiently to make the position of this lead less of a problem. (In stereo applications the capacitor used will probably need reducing to 100-200pF to avoid attenuating the stereo switching waveform unduly, and if a long screened lead is used for the output the capacitance of this lead should be enough by itself.)

Prototypes were fitted into an ITT die-cast case with internal dimensions 12 × 9.5 × 2.5cm and is a tight fit with regard to height. Connections to the tuner should be made through insulated feed-through terminals close to the board so that all leads are as short as possible inside the screened box. A slot will have to be cut in the side of the box to allow the tuning capacitor shaft to pass through. The author found it easiest to fit the tuner board to the lid of the box, and to use the box as a cover, with holes drilled in line with the trimmer capacitors and L_4 and L_5 to enable final alignment with the box closed.

The board is mounted by four 6BA screws as shown in Fig. 4 and must be spaced about 5-8mm from the surface of the screened box, to prevent the track-side of the board from shorting on the box. Extra nuts or spacers may be used to achieve this spacing.

Dial drive system

Fig. 8 shows the layout of the suggested cord drive system which eliminates the problem of pointer friction, and is suitable for mounting in any plane. The parts required are made by the manufacturers of the 3-gang capacitor with the exception of the pointer, made of 18-gauge tinned copper, or similar, and the cord. Typical scale graduations for the 87.5 to 108-MHz band are shown to scale in Fig. 9.

To be concluded with a discussion of devices used and alignment methods.

Parts list

Set of parts is available from Integrex Ltd, P.O. Box 45, Derby DE1 1TW.

Inductors
See illustrations and text for winding details
L_4 & L_5 inductor assemblies are Neosid NS/E3

Dial components (all Jackson)
¼-in brass pulleys, type 4879 (4 off)
¼-in brass pivots, type 4539 (4 off)
brass spacers, type 4880 (4 off)
type 'G' drive spindle, type 5080
drive drum 2.5cm dia. 4-mm bore

Variable capacitors
3 × 14pF tuning capacitor, part no.5560/3/14 (Jackson)
4.5 to 20pF trimmer (3 off) (Piher make from Henry's Radio or Rosenthal type STSE-7, N750 from Radio Resistor Co. or type 7S-Triko 02 from Steatite Insulations)

Fixed capacitors
1nF disc ceramics, 50V, 1-cm mounting centres (10 off)
0.1μF, 16V (Mullard type C280)
32μF, 10V (Mullard type C426)
22, 68 (2 off) & 100pF, 160V, polystyrene
15, 47 & (2 off) 330pF ceramic tubular or disc, or polystyrene mounting centres 1cm
4.7nF miniature tubular polystyrene 1.65-cm mounting centres. Use 150pF for stereo reception

Resistors
Miniature carbon film type, ⅛ watt ±5% tolerance (Mullard)

Active devices
40673 (RCA 2 off). In mixer stage, lack of gate protection diodes may be acceptable, in which case types 40604 or 3N141 can be used. If the risk of not using diodes is acceptable for r.f. stage, types 40603 or 3N140 can be used. N.B.: retain protective spring until power is applied
40244 (RCA), alternatively TI409 (TO-92) —now avaialable as TIS64 (TO-18), Texas
BC213L (Texas) or BCY70
CA3053, 3028A or 3028B (RCA)
TAA661B (SGS)
TIV307 (Texas)

Also needed are
printed circuit board (available drilled, solder coated and with component locations)
ceramic filters type FM-4 (Vernitron). Order as pair with same colour coding (orange-10.625 MHz, yellow-10.6625 MHz, green-10.700MHz, blue-10.7375MHz, violet-10.775MHz)
trimming tool for L_4 and L_5 cores
nylon cord
die-cast box
Denfix cement
Denfix cement (from Home Radio)
Evostik latex-resin impact adhesive

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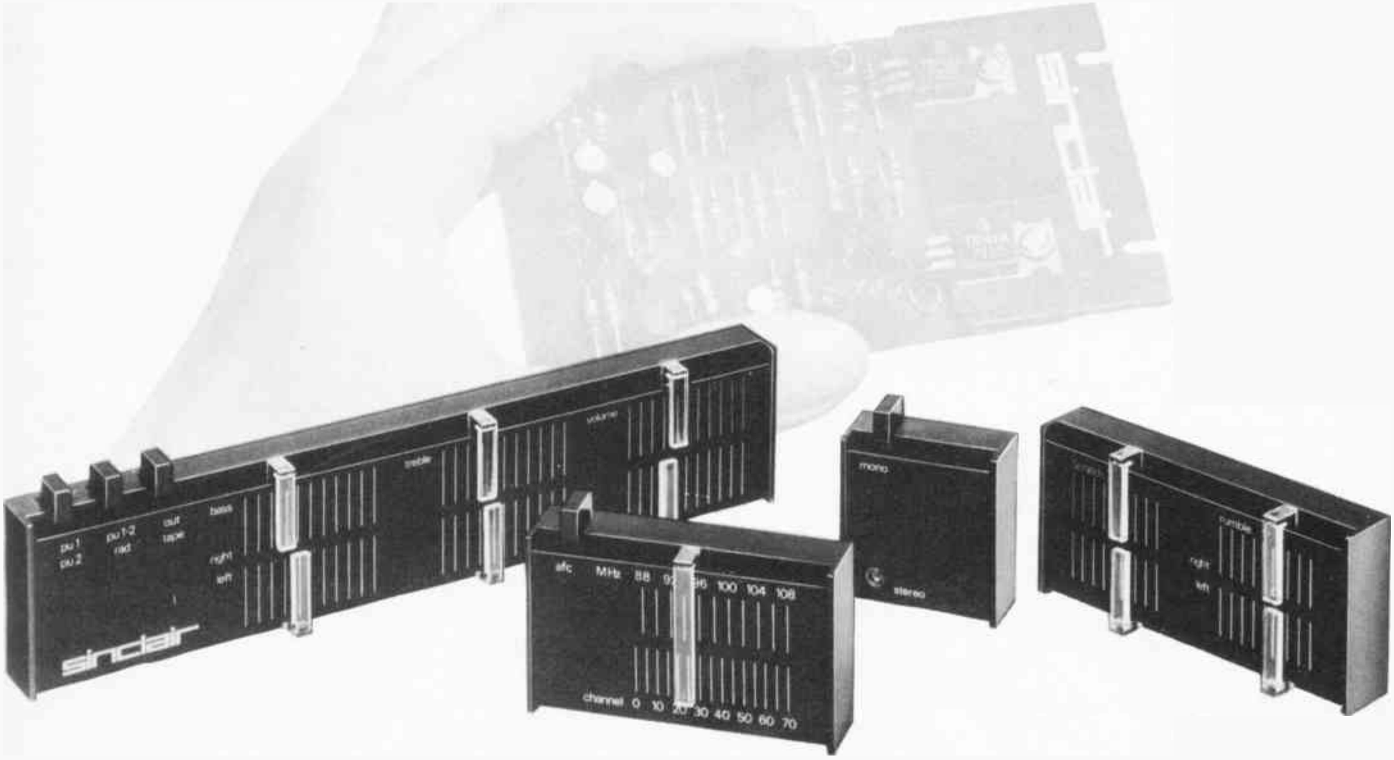
Sinclair Project 80



Stereo 80 Control Unit Size – 260 × 50 × 20mm (10¼ × 2 × ¾ins) **Finish** – Black with white indicators and transparent sliders **Inputs** – Magnetic pick-up 3mV RIAA corrected; Ceramic pick-up 350mV Radio 100mV; Tape 30mV **Signal/noise ratio** – 60db **Frequency range** – 20Hz to 15KHz ±1dB; 10Hz to 25KHz±3dB **Power requirements** – 20 to 35 volts **Outputs** – 100mV+AB monitoring for tape **Controls** – Press button tape radio and P.U. Sliders on each channel for volume bass treble R.R.P. (add £1·19 V.A.T.) **£11.95**

Project 80 FM Tuner Size – 85 × 50 × 20mm (3½ × 2 × ¾ins) **Tuning range** Dual varicap – 87·5 to 108MHz **Detector** – I.C. balanced coincidence **One I.C. equal to 26 transistors Distortion** – 0·2% at 1KHz for 30% modulation 4 pole ceramic filter in I.F. section **Aerial impedance** – 75 Ω or 240-300 Ω **Sensitivity** – 5 microvolts for 30dB S/N ratio **Output** – 300mV for 30% modulation **Power requirements** – 25 to 35 volts R.R.P. (add £1·19 V.A.T.) **£11.95**

Project 80 Stereo Decoder Size – 47 × 50 × 20mm (1⅞ × 2 × ¾ins) **One 19 transistor I.C. Channel separation** greater than 30dB **Power requirements** – 25V **Output** 150mV per channel R.R.P. (add 74p V.A.T.) **£7.45**

Active Filter Unit Separate controls on each channel. **Size** – 108 × 50 × 20mm (4¼ × 2 × ¾ins) **Voltage gain** – minus 0·2dB **Frequency response** – 40Hz to 22KHz controls minimum **Distortion** – at 1KHz – 0·03% using 30V supply **H.F. cut off** (scratch) – 22 KHz to 5·5KHz. 12dB/oct. slope **L.F. cut off** (rumble) – 28dB at 20Hz, 9dB/oct. slope R.R.P. (add 69p V.A.T.) **£6.95**

Z.40 Power Amplifier Size – 55 × 80 × 20mm (2⅛ × 3⅛ × ¾ins) 9 transistors **Input sensitivity** – 100mV **Output** 18 watts RMS continuous into 4 Ω (35V) **Frequency response** – 30Hz-100KHz±3dB **S/N ratio** – 64dB **Distortion** – at 10 watts into 8 Ω less than 0·1% **Power requirements** – 12 to 35 volts; built-in protection against overload. R.R.P. (add 54p V.A.T.) **£5.40**

Z.60 Power Amplifier Size – 55 × 98 × 15mm (2⅛ × 3⅞ × ⅝ins) 12 transistors **Input sensitivity** – 100-250mV **Output** – 25 watts RMS continuous into 8 Ω (50V). **Distortion** – typically 0·03% **Frequency response** – 15Hz to more than 200KHz±3dB **S/N ratio** – better than 70dB **Built-in protection** against transient overload and short circuiting **Load impedance** – 4 Ω min. safe on open circuit R.R.P. (add 69p V.A.T.) **£6.95**

Power Supply Units **PZ.8** Stabilised. Re-entrant current limiting makes damage from overload or even direct shorting impossible. Normal working voltage (adjustable) 50V. R.R.P. £7·98+79p V.A.T. Without mains transformer **PZ.6** 35V. stabilised R.R.P. £7·98+79p V.A.T. **PZ.5** 30V unstabilised R.R.P. £4·98+49p V.A.T.

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Project 80 gives you choice from a range of 9 different modules for combining in a variety of ways to suit your requirements. The Stereo 80 is a versatile pre-amp control unit designed to meet all domestic hi-fi requirements including tape monitoring, high sensitivity magnetic cartridge input, and of course, individual slide controls on each channel for precise output matching. By separating the F.M tuner and stereo decoder, useful economies can be effected where stereo radio reception is not needed. Two power amplifiers – Z.40 (18 watts RMS continuous into 4 ohms using 35V) and Z.60 (25 watts RMS continuous into 8 ohms using 50V) are available with choice of 3 different power supply units. The PZ.8 with its virtually indestructible circuitry is particularly recommended. For the final word in system building, the Active Filter Unit puts the finishing touch of quality to what are easily the world's most technically advanced hi-fi modules. Any further units likely to be added to Project 80 range will be compatible with those already available.

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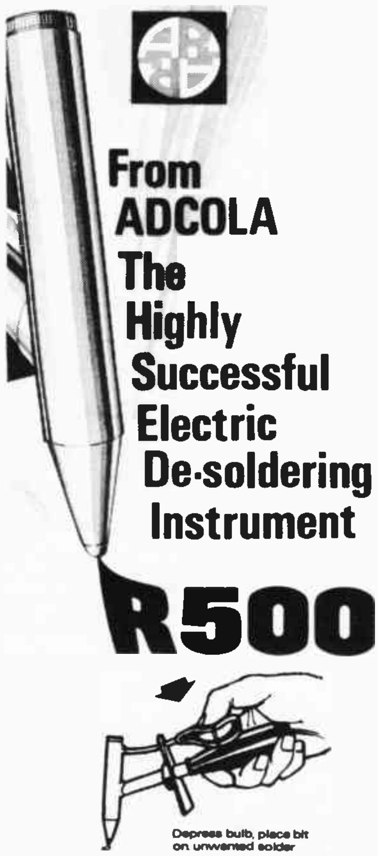

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F.M. Stereo Tuner

2—Further details of high-performance design for home construction

by L. Nelson-Jones, F.I.E.R.E.

This sensitive f.m. tuner design, described in last month's issue, has a performance equal to the better examples of commercial tuners, but at a much lower cost. Full constructional details were given in Part 1 and this article discusses in detail some of the devices used—especially the dual-gate m.o.s.f.e.t., integrated circuit demodulator and ceramic i.f. filters—and concludes with alignment instructions.

The dual-gate m.o.s.f.e.t. is not to be confused with the type of junction f.e.t. which has two gate connections, usually one to the gate and the other to the substrate, as this has gates effectively in parallel. The dual-gate m.o.s.f.e.t. has gates effectively in series so that it can be likened to the multi-grid valve or a cascode stage and like these devices has the advantage of very low feedback capacitance from output to input. It has also the same advantages as single-gate m.o.s.f.e.t.s namely, good signal handling, low noise, and high input impedance. Fig. 9 shows the likeness of the dual-gate m.o.s.f.e.t. to a cascode stage, and its construction. The drain current of a dual-gate m.o.s.f.e.t. is a function of both gate potentials, and this enables gate 2 to be used for gain control in the case of r.f. amplifiers, or for injection of local oscillator voltages in the case of mixer stages. Type 40673 is very similar to the 3N140 but in addition has full protection of both gates by pairs of zener diodes between each gate and the source (and substrate) electrodes. These diodes are clearly of minute proportions—they add only a fraction of a picofarad to the gate capacitances. The breakdown of these diodes is around ±10 volts, so that normal signal levels do not cause conduction. But the diodes will conduct long before the gate breakdown voltages are reached, and, provided the resultant currents are adequately limited by the circuit values, no harm will result to the gates.

Apart from the obvious advantage of two controlling gates, the great advantage of the second gate is that it acts as a 'guard ring' between the drain, and gate 1. The result of this guard ring action is a typical drain-to-gate 1 capacitance of 0.02pF (with a maximum for the 40673 and 3N140 types of 0.03PF). This low value of feedback capacitance enables such a device to give up to 28dB of power gain at 100MHz, without need for neutralization, but in practice a gain of 20dB is a more realistic figure for an r.f. amplifier at this frequency. This ensures a high margin of stability, which together with the superior signal handling qualities of the m.o.s.f.e.t. make this a very easy device to use for r.f. amplification in an f.m. tuner.

Integrated circuit i.f. amplifiers

Integrated-circuit i.f. amplifiers have been available for some time now in various forms, from the simple differential pair and the cascode stage, up to relatively complex circuits such as that used in the receiver described (TAA661B). There are now a number of these more complex circuits available, nearly all of which use a product detector for demodulation. Examples of these are the Sprague ULN-2111, Plessey SL432A, and the SGS TAA661B. Fig. 10 shows the circuit of the TAA661B, together with the basic external connections.

Gain is provided by three stages, each of which is a non-saturating differential amplifier followed by an emitter follower.

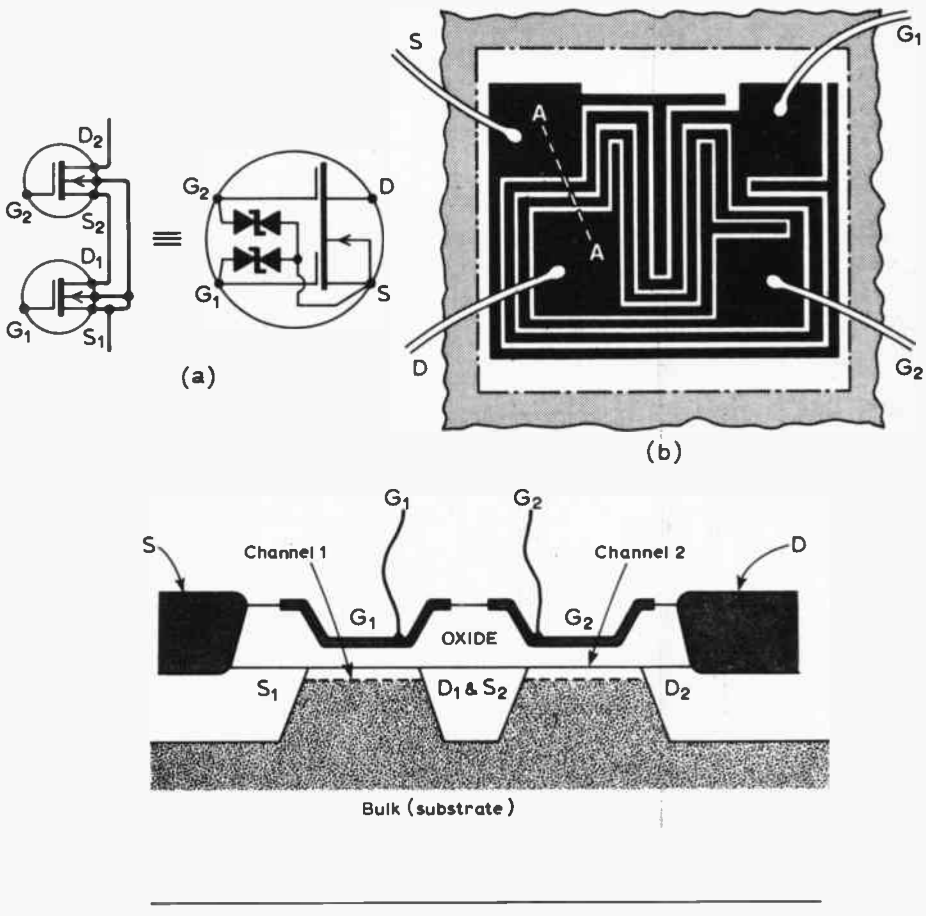


Fig. 9. Symbolic representation of dual-gate m.o.s.f.e.t. (a) showing similarity to cascode stage. Plan view (b) shows complete separation of gate 1 from drain-by-gate 2; (c) shows section across A-A. Bi-directional zener diodes conduct at around ±10V preventing gate breakdown (type 40673 only).

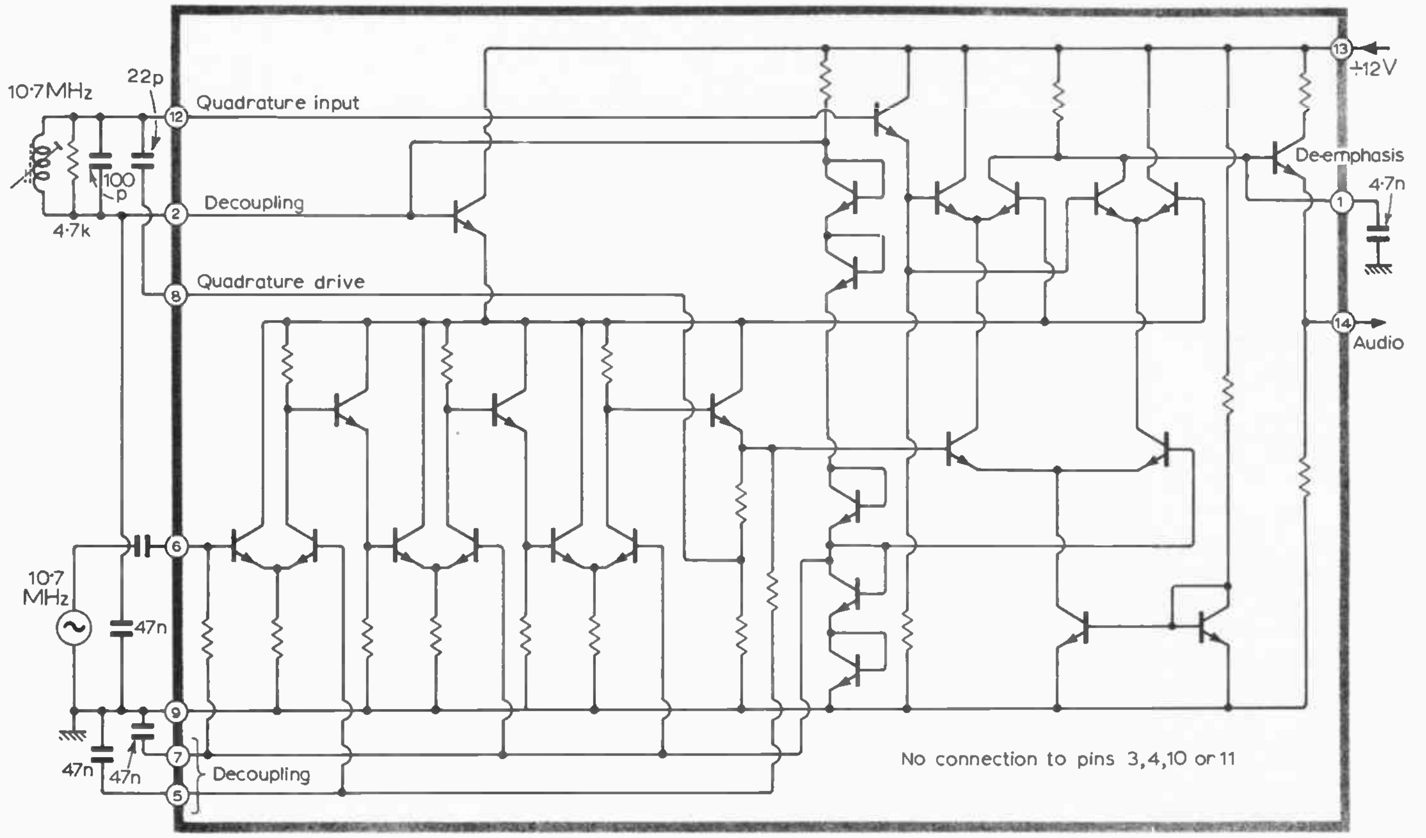


Fig. 10. Demodulation and i.f. amplification are performed in this single-chip integrated circuit (TAA661B). Phase-sensitive detector consists of 'tree' of differential pairs with constant-current tail (to right of bias chain), fed with a phase reference provided by tuned circuit and with signal to lower pair.

Overall d.c. feedback is applied so that the output level at the third emitter follower is kept equal to that of the base of the input transistor. This voltage is set at approximately 1.4 volts by the bias chain of five diodes which has two outputs, equal to two and five 'diode voltage drops'. The higher of these voltages is used to control the main supply line of the amplifier stages via an emitter follower. This supply line is therefore at approximately 2.8 volts—five 'diode drops' less the base-diode drop of the emitter follower.

The detector consists of a 'tree' of differential pairs with a constant-current source in the common 'tail' connection. This constant-current source is a 'current mirror' circuit where the constant current is equal to the current feeding the second transistor, strapped as a diode. The current mirror principle is based on the fact that two equal transistors with equal base-emitter voltages will also have equal collector currents. This principle may be extended so that two transistors of an integrated circuit having different areas (but otherwise similar) will have collector currents equal to their areas when used in such a circuit.

The detector acts as a phase-sensitive full-wave rectifier, with a phase reference provided by a tuned circuit driven from a tap on the load of the final emitter follower of the amplifier. The lower two transistors of the tree are driven by the signal from

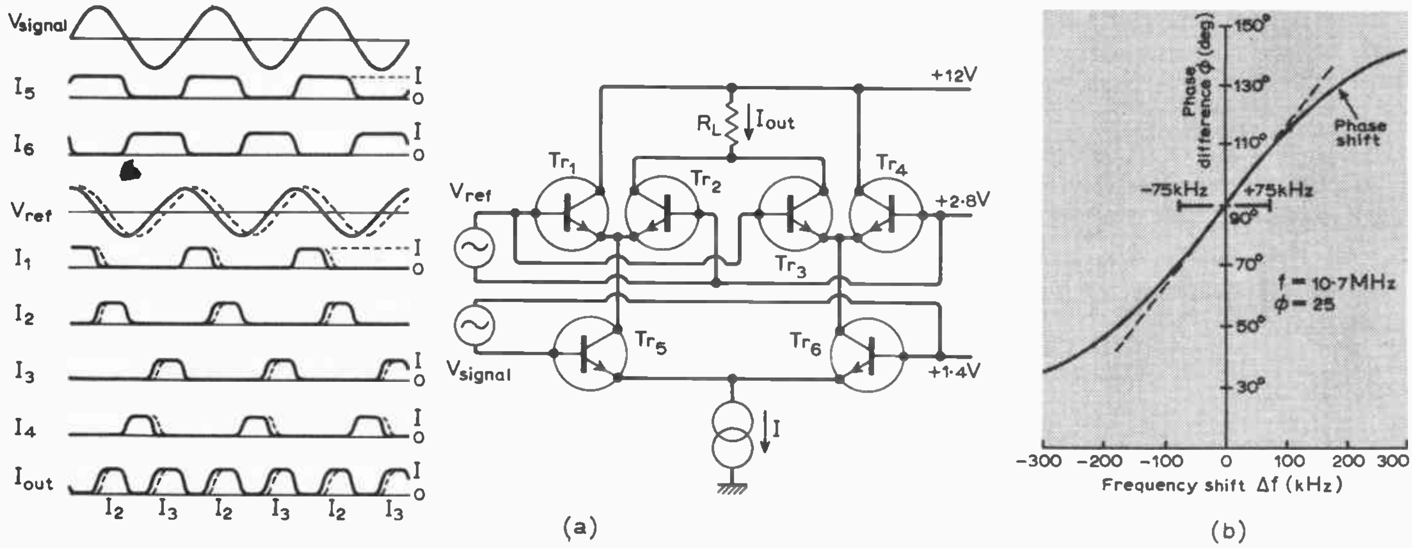


Fig. 11. At resonant frequency of tuned circuit—see Fig. 10—signal applied to lower differential pair is in quadrature with reference from tuned circuit fed to upper pairs, and current divided equally between each of the upper pairs (a). When signal frequency deviates, phase difference between the two signals increases or decreases (b), changing proportion of current through each half of both upper pairs.

the amplifier, and as the two bases of the pair are at equal d.c. potential—due to the overall 100% d.c. feedback over the amplifier—the collector currents of these two transistors become square waves at the carrier frequency, at all signal levels above the limiting threshold of the amplifier chain. The two upper pairs of transistors are fed by the reference voltage from the tuned circuit and like the lower pair the bases of both pairs are at equal d.c. potential.

One base of each pair is connected to the supply line of the amplifier, while the other is fed by an emitter follower, biased via the tuned circuit from the same potential as the base of the emitter follower controlling the supply line of the amplifier. The voltage across the tuned circuit is approximately 300mV peak-to-peak with full limiting so that these upper pairs of transistors are also fully switched at the reference frequency. At the resonant frequency of the reference circuit, the signal voltage applied to the lower pair of transistors is in quadrature with the signal from the tuned circuit to the upper pairs of transistors, due to the loose coupling of the tuned circuit via the 22-pF capacitor. Thus at resonance the current square wave through each half of the lower pair of transistors will divide equally between each of the upper pairs, because the two signals are in quadrature, and the transition of the reference waveform takes place midway through each half cycle of the current square wave supplied by the lower pair. Action is shown in Fig. 11a.

As the frequency departs from the centre frequency of the tuned circuit, the phase difference between the two signals decreases or increases, depending on the direction of the frequency shift, so that the proportion of the current passing through each half of each upper pair changes, Fig. 11b. The collectors of the upper pairs are connected so that the pair which have an increase in current for an increase in frequency are connected together, as are those having a decrease. One pair of collectors is connected to a load resistor, and the other pair direct to the 12-volt supply. The load resistor drops approximately 6.0 volts at the centre frequency so that the output level is typically +5.5 volts, at the emitter of the emitter-follower output stage.

De-emphasis is arranged by a capacitor connected to the base of the emitter follower. Alternatively a separate de-emphasis network can be connected to the output in the usual way, with a much smaller value of capacitor connected to pin 1. A capacitor connected to this pin is still essential to preserve overall stability by by-passing the r.f. voltages present at this point. A similar reduction is necessary if the output is applied to a stereo decoder. A value of 150pF is suitable in either case.

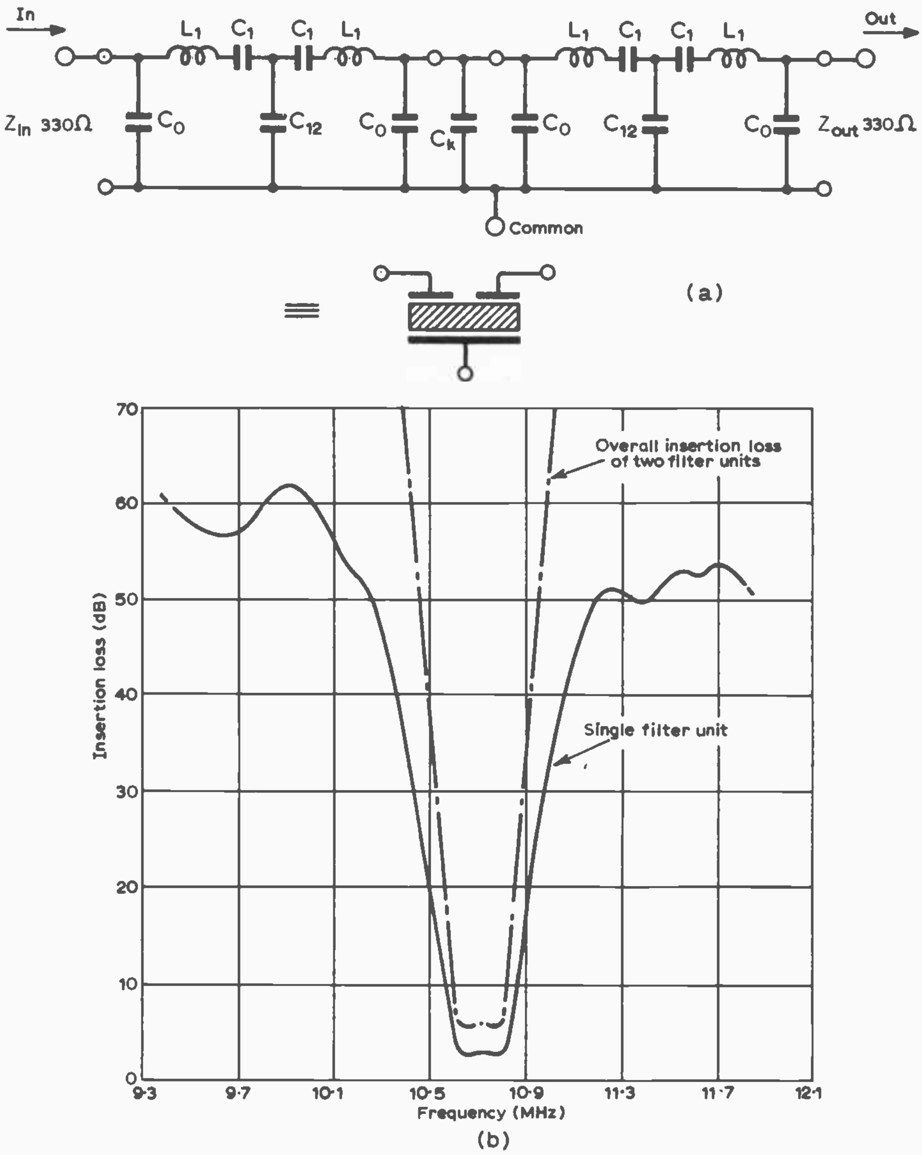


Fig. 12. Ceramic resonrtor used is equivalent to two 2-pole filters coupled by capacitor C_k (a) and has selectivity shown at (b). Two cascaded filters give a bandwidth of 220kHz at 3dB down and 560kHz at 60 dB down.

Ceramic i.f. resonators

There are a number of ceramic resonators on the market and they take different physical forms. Some are similar to the type of filter common in communications receivers where high degrees of selectivity are required, and these have tuned circuits at input and output with one or more resonators between. Others use only ceramic resonators, with perhaps coupling capacitors, and have a family resemblance to the type of crystal filter used in v.h.f. communications receivers for high degrees of i.f. selectivity at 10.7MHz. The type of filter used in the tuner consists only of a single ceramic resonator which by the layout of its electrodes performs the function of a multi-section filter with a bandpass characteristic.

Such filters are now also being made in quartz for v.h.f. communications receivers and can equal the performance of much more complex multi-element filters, despite their relative simplicity. This excellence of performance is true also of the ceramic type, where the device used has a performance slightly better than a multi-element device of otherwise similar characteristics, both in respect of selectivity and passband loss. Due to its greater simplicity it is also much cheaper, and smaller.

The equivalent circuit of the filter (Vernitron FM-4) is of two 2-pole filters coupled by an additional capacitor *Ck* as shown in Fig. 12. Physically all the elements are on a single ceramic substrate. The overall response is equivalent to two critically-coupled bandpass circuits in cascade. Figure 12b shows the typical selectivity of such a single unit (solid curve) consisting of a single substrate multi-pole filter with the equivalent circuit of Fig. 12a. The broken curve shows the result of using two such complete resonator units (with a suitable buffer stage between) to obtain higher selectivity. The resultant performance is more than adequate for f.m. broadcast reception, with 3-dB bandwidth of typically 220kHz and 60dB bandwidth of around 560kHz. Ripple in the pass-band is quoted as not exceeding 1dB (2dB for two stages).

Measurements confirm these figures for typical pairs of filter units in a practical amplifier. These resonators cannot be coupled directly to one another in normal use or the balance of the response curves will be upset, resulting in a highly asymmetric response—the use of driving or load impedances noticeably different from the 330-ohm design impedance will upset the degree of coupling in individual sections. This relatively low impedance of 330 ohms is perhaps one of the drawbacks of this type of ceramic resonator, although most such filters have impedances in the same region. In practice, however, the reduction in gain due to the use of such low-impedance loads in the amplifier chain is not too serious, especially as an additional low-gain buffer stage between the filter sections is needed to avoid interaction of the filters.

The most serious loss of gain from this low load impedance would occur in the mixer stage, where with a typical dual-gate f.e.t. stage as described, the voltage gain of the mixer would be reduced to a little below unity with such a load. In the tuner design the mixer is therefore modified to use a tuned load with a grounded-base buffer stage feeding a 330-ohm resistive load to which the first filter is connected. This ensures a true 330-ohm source for the filter, and results in an overall mixer gain of 24dB. The mixer load circuit is designed to work at only a moderate Q so that the tuning of this circuit is not highly critical.

Due to production tolerances the ceramic resonators are graded into frequency bands and appropriately colour coded to indicate their exact frequency tolerance. For the type used there are five groups covering a total spread of 150kHz (at 37.5-kHz intervals) around 10.7MHz. In a receiver using two such filters, both must be of the same colour group to achieve a satisfactory result. (Details of these groupings were given in the parts list.)

Variable-capacitance diodes

In the past few years some improvements have been made in the parameters of variable-capacitance diodes for tuning. These improvements have given diodes a higher Q and a wider variation in capacitance for a given voltage change. Many tuners now use such diodes exclusively for tuning the r.f. circuits, and they are becoming common in u.h.f. television tuners. The great advantage of these diodes for tuning is that the r.f. circuitry can be made very compact, thus minimizing pick-up and easing screening problems, and making circuit location independent of dial mechanism location. The main disadvantage, so far as the average constructor is concerned, is that both availability and price are at a disadvantage compared with a normal tuning capacitor at the present time.

Another common use of these diodes is a.f.c., and here the requirements are not nearly so severe as only a small variation of capacitance is required. Availability of diodes for this purpose with a smaller change of capacitance with voltage is fairly good and prices are moderate. Such a device is the Texas TIV307, whose capacitance versus voltage curve is shown in Fig. 13 (as measured by the author on three samples). This device could also be used for tuning purposes—it has just adequate capacitance variation without

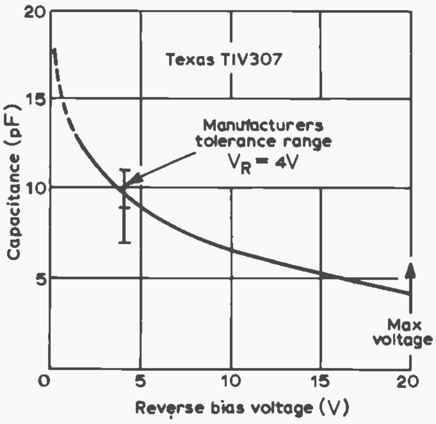


Fig. 13. Capacitance-voltage characteristic for a.f.c. diode.

using too high or too low a control voltage. Too low a control voltage is undesirable, especially in the oscillator stage, where harmonic generation and conduction of the diode become a problem at low bias voltage. Too high a voltage can be a problem either because the diode will not withstand it, or because the voltage is simply not available.

The smallest capacitance swing is in the oscillator circuit (from 87.5 + 10.7 = 98.2 to 108 + 10.7 = 118.7MHz) which has a frequency ratio f_{max}/f_{min} of 1.22, and a capacitance ratio of 1.22^2 :1 or 1.46 : 1. The r.f. circuits need a capacitance swing of $(108/87.5)^2 = 1.53:1$. A swing of from 2.5 to 7 volts would give such a change if the only capacitance were the diode. But there will always be 10-15pF of general circuit capacitance, so that a diode change from 10 to 20pF at least is necessary (or just over 2:1 variation in the diode). With some care in circuit layout and a change in L/C ratio the TIV307 could just give this swing (or its companion with higher capacitance TIV308–12pF at −4 volts), especially if a higher supply voltage than the 12 volts used in the design were available. It is the author's intention at a later date to design a diode-tuned receiver, but in the present design a normal tuning capacitor is used mainly on the grounds of cost and the difficulty of obtaining diodes in suitably matched triplets.

Alignment of tuned circuits

The tuned circuits must be aligned in reverse order, that is starting at L_5 and working back to L_1. By far the easiest way of aligning the i.f. section is to use a wobbulator centred on 10.7MHz and having a sweep frequency of 50Hz, with a peak-to-peak deviation of 1 to 2MHz. Fig. 14 shows the response of a correctly aligned i.f. amplifier and demodulator: The y-axis is the output of the tuner (1 volt/division) and the x-axis is the modulation voltage (75kHz/division). The display shown is for a moderate input, but is well into limiting. Apply the wobbulator input to L_2 via a capacitor at about 1mV level from 80 ohms. The core of L_4 is easily set for maximum gain by looking at the noise amplitude at either side of the display. As the core is moved the noise is first greatest on one side and then at the other as the resonant frequency of L_4 moves across the band. Set the core to mid-way between the positions giving maximum noise on either side.

If a wobbulator is not available then at least a signal generator must be used. Connect as for the wobbulator above to L_2 via a capacitor and apply a level of around 1mV from 80 ohms. Connect a centre-zero meter of around ±3V full scale between the output and the a.f.c. reference lead (preferably better than 10kohm/volt sensitivity). Rock the tuning of the signal generator back and forth around 10.7MHz while adjusting the core of L_5 until the positive peak excursion is equal to the negative peak excursion—Fig. 14. If the signal generator calibration is fine enough the tuning of L_5 is finally set for best linearity, plotting output voltage against frequency. If the core of L_5 is far from the correct setting, the output may be a totally positive or totally negative excursion, with no S-shape.

If a centre-zero meter is not available the 10-volt range of a multi-meter may be used connected between the output and earth. When the signal generator is far off tune the reading of the output level should be around 5.5volts (supply at 12 volts). This is equivalent to the zero centre reading using the centre-zero instrument as above. The tuning of L_5 is now set for equal deflections about 5.5 volts.

To set L_4 the signal generator is set slightly to one side of the centre frequency and the level dropped until the meter indication begins to change (i.e. drops below limiting level). Adjust L_4 to make good this change (i.e. to increase signal strength) reducing the signal generator level to keep the i.f. stages below limiting level. Continue the process until no further improvement can be made. Alternatively, the core of L_4 may be tuned for maximum noise output with no signal generator connected or with generator switched off.

While aligning the i.f. section it is a help to have the oscillator out of action to

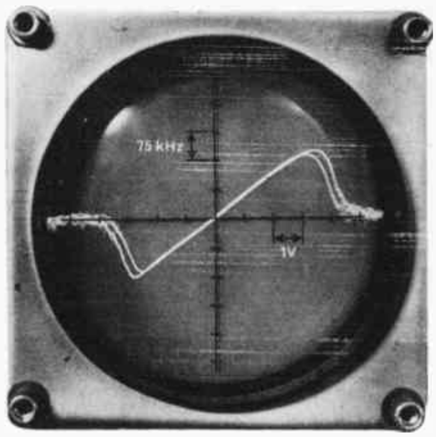


Fig. 14. Response of correctly aligned i.f. amplifier and demodulator. L_4 is set for maximum gain midway between positions giving maximum noise, either side of the display.

prevent spurious responses from i.f. harmonics. This is most easily achieved by shorting out L_3 with a single crocodile clip across the ends of the coil to connect together the end turns.

It is not sufficient to set the output level to 5.5. volts (or zero with respect to the a.f.c. reference) with an input of 10.7MHz when tuning L_5 for two reasons. First the majority of signal generators, even of very high quality, are not accurate enough to ensure a symmetrical S-shaped characteristic, and, secondly, the ceramic resonators are not necessarily peaked at 10.7MHz. If the frequency of the generator is known to within about 10kHz or better and is set to the i.f. indicated by the ceramic resonators colour code (see parts list), then L_5 may be set initially in this way. But the symmetry of the S-shaped characteristic of the detector should still be checked after setting L_4, and any slight correction made as appropriate to the core position of L_5.

Align the r.f. section in the usual way for superheterodyne receivers—set the oscillator so that the correct span of input frequencies is covered, and then adjust the r.f. circuits to track correctly.

To adjust the oscillator set the tuning capacitor to maximum capacitance and the signal generator to 87.5MHz. Adjust the variable capacitor next to the oscillator coil to receive the 87.5-MHz signal. Set the tuning capacitor to minimum capacitance and the signal generator to 108MHz. Now adjust the trimmer capacitance again noting which way this adjustment is to tune in the 108-MHz signal. (With this type of capacitor, maximum capacitance is with the silvering on the top disc towards the centre connecting pin of the capacitors three pins, and minimum is 180° from this position i.e. farthest from the middle pin.)

If the capacitance setting needs reducing at 108MHz then increase the value of L_3 by squeezing the coil to bring the turns closer together. Re-adjust the capacitor to bring the receiver back to tune at 108MHz and return to 87.5MHz and maximum capacitance of the tuning capacitor. If the trimming capacitor now needs decreasing in capacitance then the coil inductance has been increased too much.

An alternative method, possibly quicker, is to set the 108-MHz end using only the trimmer capacitor and then find to what frequency the low-end is tuned, without altering the trimmer capacitor. If this is below 87.5MHz, reduce the inductance by opening out the turns; if it is above 87.5MHz, close up the turns. Set the frequency and tuning again to 108MHz and reset the trimming capacitor. Return to the low end and again check the frequency the receiver is set to; continue this process until on reaching the low end the receiver is set to exactly 87.5MHz.

Having set the span of the oscillator, the two r.f. coils and their trimmer capacitors need adjusting to complete the alignment. This is possibly the most difficult part of the alignment procedure because of the high sensitivity of the receiver. Perhaps the simplest method is to dispense with the signal generator altogether at this point and to tune for maximum noise, with the signal generator switched off but still connected. Tracking of the r.f. coils may be set in a similar way to the oscillator coil.

Set the tuning capacitor to minimum capacitance, and tune for maximum noise using the trimmers. Set the tuning capacitor to maximum capacitance and, in the manner used for the oscillator, check whether the trimmer capacitance needs increasing or decreasing to tune for maximum noise at the low-frequency end of the dial. If the capacitance needs increasing, squeeze the coil turns closer together; if the trimmer needs decreasing in capacitance then open out the coil turns slightly. Return to the minimum value of the tuning capacitance and repeat the process from the beginning, and continue to do so until both r.f. coils need no change of tuning of the trimmer on going from one end of the dial to the other.

Further reading

The publications listed may interest those wishing to pursue various design aspects.

"MOS field-effect transistors", RCA product guide MOS160A

Data sheets on devices 40673, 3N140, 3N141, 40603, 40604 (RCA)

Data sheets on devices CA3028A, 3028B, 3053 (RCA)

"Understanding and using the dual-gate m.o.s.f.e.t." RCA application report ST-3529

"Application of dual-gate m.o.s. field-effect transistors in practical radio receivers," RCA application report ST-3486

"Integrated-circuit frequency modulation i.f. amplifiers", RCA application report ICAN-5380

"Integrated circuits for f.m. broadcast receivers", RCA application report ICAN-5269

"Use of 10.7MHz ceramic coupled-mode filters in linear i.c. i.f. strips", Vernitron application report

Data sheet on ceramic filter FM-4, bulletin 94033 (Vernitron)

Data sheet on TAA661B (SGS)

Addresses

RCA (GB) Ltd, Lincoln Way, Windmill Road, Sunbury-on-Thames, Middx

SGS Ltd, Aylesbury, Bucks

Vernitron Ltd, Thornhill, Southampton SO9 1QX, Hants

F.M. Tuner Design — 12 Months Later

by L. Nelson-Jones, F.I.E.R.E.

The author of the articles on the stereo f.m. tuner design (April, May 1971) looks back on the 12 months since the design was published and gives some hints on getting the best from the design. Cures are listed for the small number of troubles – out of nearly 2,000 tuners for which parts have been sold – experienced by readers. Test voltages are also given as well as dial mechanism and stereo decoder mounting details. A solid-state tuning indicator is described separately on pages 182 and 183.

This tuner has been made by many of the readers of *Wireless World* and, inevitably, a few of these tuners have given trouble. Most of these troubles have been traced to component faults, or errors in construction, but two faults in particular have recurred in a number of cases. One of these is concerned with oscillator tracking, and the other an instability, apparently in the r.f. section, but which is very difficult to track down.

Oscillator tracking

In many cases it has not been possible to get the oscillator to quite reach 108-MHz coverage while at the same time covering down to 87.5MHz, because the oscillator coil could not be closed up enough without shorting. The first attempt to increase the inductance of the oscillator coil was to reverse the direction of winding thus giving about an extra half turn. This proved to be too big an increase resulting in the coil not being able to be reduced in inductance sufficiently even with the turns highly extended. The next move was to increase the diameter of the coil some 15%, but this too proved non-effective due to the close proximity of the tuning capacitor body, and in fact an appreciable drop in inductance was noted. Finally it was found that the very small increase in inductance needed could easily be obtained by standing the coil further away from the printed circuit board. Instead of mounting the coil so that the turns are 2.5mm clear of the board, as described in the original article, it should be mounted with 5 to 6-mm clearance. The coil should now track easily with the turns spaced from one another by about half a wire diameter. The r.f. coils both track normally with the turns spaced by approximately one wire diameter. In each case this is about the 'natural' length of the coils.

In existing tuners it will be necessary to remove the oscillator coil and fit a new one to effect this cure as it is unlikely that the existing coil will have sufficient surplus lead length. Take great care when removing the leads from the board. The simplest way in my experience is to cut the coil off leaving a reasonable lead length, and then to remove each lead separately, being very careful to pull the lead through the board from the component side (never from the copper circuit side). This minimizes the risk of 'pad lifting'. The holes are then cleared of solder, being careful not to increase the hole diameters, and a new coil fitted in the normal manner but with the increased clearance. Fig. 1 shows the amended mounting details for the coil.

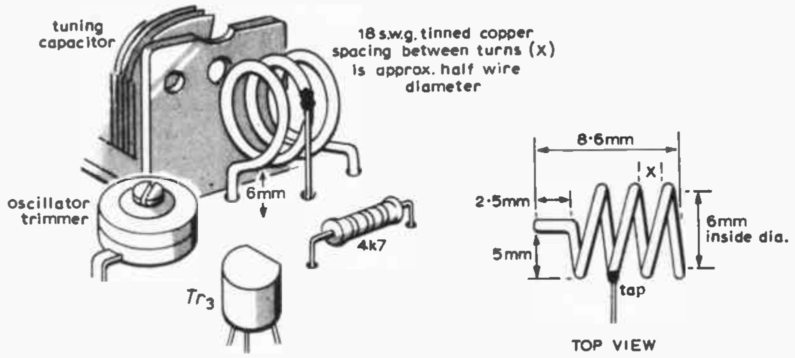


Fig. 1. For those who cannot tune to 108MHz, inductance of oscillator coil can be increased by moving it further away from the printed board.

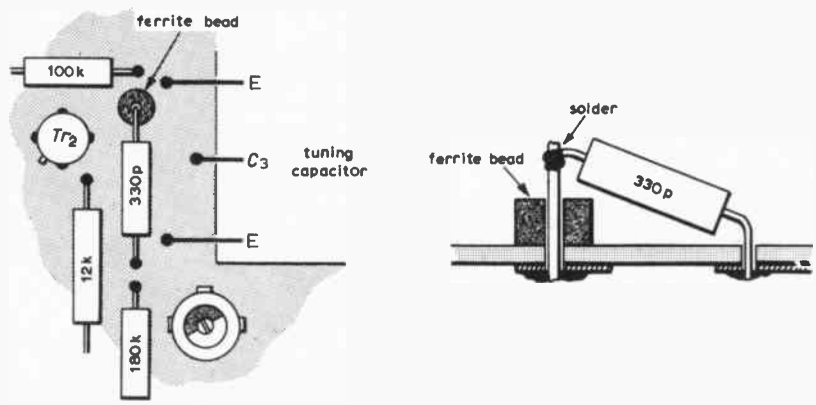


Fig. 2. In some cases unwanted oscillation at around 500MHz has caused odd effects, and has been cured by adding a ferrite bead as shown (see also photograph).

Instability

This fault has proved a very elusive one which nearly always shows up only at the upper end of the frequency coverage. It can also show up as an excessive noise level, with more than a reasonable number of spurious responses, but with the tuner working quite normally on strong signals. Most readers deduced that the mixer was unstable, others that it was the r.f. stage, and some that the oscillator was squegging, which can give a similar effect. In fact it seems likely that all three stages are involved, as the oscillation is almost certainly not at the normal carrier frequencies for which the receiver was designed but around 500MHz.

It is extremely hard to track down the exact mechanism, but it appears that it is usually associated with f.e.ts having higher than average gain in this u.h.f. region. An f.e.t. differs from a bipolar device in that the drop off in gain with frequency is much more sudden. A device such as those used in the tuner r.f. and mixer stages can have barely reduced gain up to say 450MHz, above which the gain rapidly reduces so that little gain exists at say 600MHz. For this reason I believe that only a few tuners have this trouble, and this is confirmed by the fact that of those with the trouble that I have examined, all had f.e.ts in the r.f. stage with higher than usual I_{dss} currents, and with these devices one would usually expect a higher gain at u.h.f.

The path for the feedback appears to be via the g_2–g_1 capacitance of the mixer, via the oscillator layout to the r.f. stage input, through this stage back to g_1 of the mixer and thus full circle. This view of the mode of oscillation is backed by a number of facts, such as the fact that the oscillation frequency is unaffected by the receiver tuning but is affected by the voltage applied to the 'varicap' diode, and that a ferrite bead in series with the oscillator feed to g_2 of the mixer effectively suppresses the oscillation. I suspect that the inductor forming the u.h.f. resonant circuit is formed by the earth plane of the printed board.

The cure for this problem is thus a very simple one and the location of the ferrite bead is shown in Fig. 2. It is mounted on the lead of the 330-pF oscillator feed capacitor. It may be mounted on either end of the capacitor, but is more easily mounted on the end shown. To mount the ferrite bead (Mullard FX1115 or Radiospares antiparasitic beads) on a tuner already constructed, the end of the capacitor is unsoldered from the board, a tinned copper wire is soldered in place of the capacitor lead in the board, and the ferrite bead placed on this lead. The capacitor is then soldered to the wire above the bead (Fig. 2).

Oscillator temperature coefficient

A small number of readers have complained of excessive drift of the tuner, even with the a.f.c. switched on. It is suspected that the oscillator may have had a faulty component in these cases, but a tuner subjected to a large change in temperature on warming-up (for instance a tuner in use in the same cabinet as a valve amplifier) may give rise to some problems. The main component causing the drift is the oscillator trimmer capacitor which has a temperature coefficient of up to −1000 p.p.m./deg C. Most of the oscillator components have measurable temperature coefficients, but the trimmer is much the highest.

Unwanted oscillation at u.h.f. may be cured with a ferrite bead over the lead of the oscillator output capacitor.

It must not be forgotten that the tuner must be operated from a stable supply as the tuner has quite a high voltage-versus-frequency coefficient—330kHz/V without a.f.c., and 60kHz/V with.

Measurements have been made on a tuner fitted with a trimmer having a lower temperature coefficient of N470 (i.e. −470 p.p.m./deg C). This trimmer gives rather better drift figures: −6kHz/deg C with a.f.c. and −16kHz/deg C without a.f.c. These figures are quite acceptable in normal circumstances.

Thus for readers having trouble with drift the change to the N470 device, which is of the same type as that used at present but of 3.5 to 13pF range, will make a considerable improvement.

For those requiring a still greater freedom from drift the use of an air dielectric or p.t.f.e. trimmer will give an improvement. Jackson type 5440/PC/PT/14.0 should be adaptable to the task although its pins will need spreading slightly, and the stator and rotor connections will be reversed so that on tuning with a screwdriver the placing of the screwdriver in the slot will pull the oscillator off tune and the screwdriver must be removed before re-checking the alignment.

My advice is therefore to leave well alone unless drift is a problem, in which case an investigation should be made to ascertain the component responsible in case there is a faulty item. Should the trimmer be the cause as indicated then a change to a type having a better temperature coefficient may be the cure. The most useful tool to ascertain approximately the relative temperature coefficients of the various components is a miniature soldering iron, which may be held close to components to heat them up, in conjunction with a tuning meter of approximately 2–0–2V f.s.d. The drift of the tuner is easily seen on such a meter which has a deflection constant of approximately 100kHz/V. A 'freezer' aerosol is not

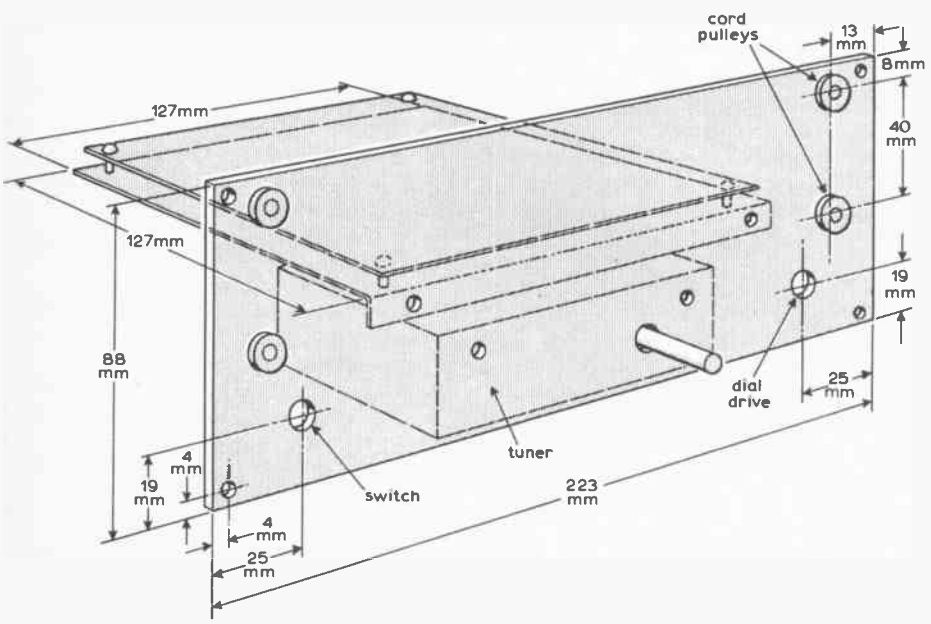


Fig. 3. New dial mounting system applicable when using stereo decoder (a) and front panel cut-outs for tuning meters (b).

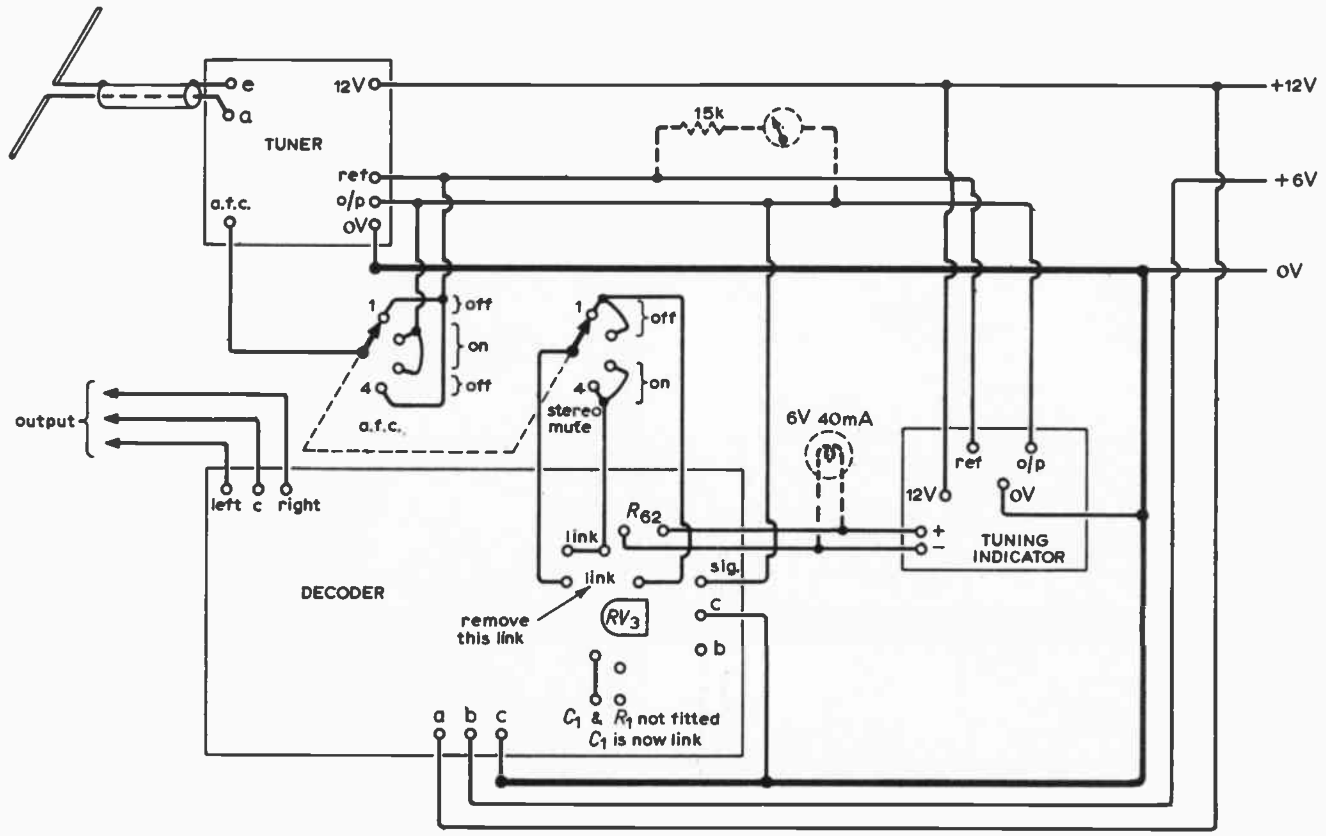


Fig. 4. Wiring diagram for connecting decoder and tuning indicators.

satisfactory as the large amount of condensation produced causes erroneous results due to the effect of the moisture on the stray capacitances.

Mounting the tuner and decoder

In the original article I suggested a method of mounting the tuner which is quite satisfactory. However, many readers will want to take advantage of the increasing amount of stereo broadcasting now available, and to receive these transmissions a decoder is needed. The design by Portus and Haywood (September 1970 issue) works well with the tuner, and I have therefore designed a new dial mounting system for use with this decoder and either a moving-coil meter or solid-state tuning meter. This chassis system is shown in Fig. 3.

The interconnection of the tuner and decoder is a simple matter provided the decoder has been assembled for single polarity supplies. Using supplies of +6V and −6V would mean having three separate supplies for tuner and decoder together. The connections are shown in Fig. 4. This also shows the connections needed to mute the decoder in the event of excessive noise on stereo transmissions, thus gaining the 20dB or so extra signal-to-noise ratio of a mono signal. There are two possibilities for muting the decoder, (a) the decoded outputs may be shorted together or (b) the decoder may be disabled by reconnecting the junction of R_{46} and R_{48} of the Portus and Haywood decoder to 0V (connection C) instead of to the collector of Tr_{13}. This point of the circuit is easily accessible and the resistor junction is disconnected from Tr_{13} by removing a link on the board adjacent to RV_3, replacing it with two leads to each end of the previous link. A third lead is connected to the short link adjacent to the positive end of D_4, and next to the link removed. These connections are shown in Fig. 4.

Either a three- or four-way switch may be used for a.f.c. and decoder muting, depending whether muting is required with or without a.f.c. This is the last position of the four-way switch, and is not strictly necessary. However, for completeness the four-way switch is shown.

All parts for this dial system and the stereo decoder mounting are available from Integrex Ltd, P.O. Box 45, Derby, TE1 1TW, including a set of printed scales. The ferrite bead is also available.

Finally, a list of alterations and corrections to the original article, some of which have already been published.

- A capacitance of 15pF is now recommended for the oscillator base capacitor, not 47pF.
- Decoupling of r.f. stage is 1nF, not 47nF (top of L_2).
- Fig. 2 caption last line should, of course, read '4.7nF at pin 1 of IC_2', not pin 2.
- In Fig. 2, the 2.2kΩ resistor at pin 5 of IC_1 should go directly to the +12V supply, and not via the 100-Ω resistor.
- In Fig. 6(c) the coil (L_5) should have 10 turns.
- The components list should have listed 1nF, nine off, and 47nF, nine off, disc ceramic capacitors.
- Type 3N201 m.o.s.f.e.t. can be used in place of the 40673.
- Texas types TI409 or TIS64 can be used in place of the 40244 transistor, now obsolete.
- Ferranti transistor type ZTX500 can be used in place of BC213L.
- ITT diode type BA110 can be used in place of TIV307.

Test voltages for f.m. tuner

test point	voltage relative to 0-V line (V)	comments
Tr_1 gate 1	0	
gate 2	4	
source	1	approx. 0.7 to 1.5V spread
drain	11.5	
Tr_2 gate 1	0	
gate 2	1.7	approx. 1.0 to 2.0V spread
source	1	approx. 0.7 to 1.5V spread
drain	11.7	
Tr_3 emitter	5.4	
base	6	
collector	12	
Tr_4 emitter	6.6	
base	6	
collector	0.75	
IC_1 pin 1	no connection	
2	3.6	
3	0	
4	2.9	
5	7	
6	9.7	
7	11.5	
8	no connection	
IC_2 pin 1	5.8 (nominal)	depends on tuning (4 to 7.5V)
2	3.5	
3 & 4	0	
5, 6 & 7	1.35	
8	0.13	
9, 10 & 11	0	
12	3.5	
13	11.5	
14	5.2 (nominal)	depends on tuning (3 to 7.5V)
a.f.c. reference	5.2	

Measurements assume a meter of 20kΩ/V and a +12-V supply.

Fitting a moving-coil or solid-state tuning indicator

Fitting a tuning indicator to the f.m. tuner is a simple matter using a centre-zero meter of 100–0–100μA sensitivity (Fig. 1). With a correctly-aligned tuner the d.c. output level and a.f.c. reference voltages are equal when the tuner is on tune, and differ when it is off tune. Thus as a station is tuned in the meter first deflects to one side as the station is reached, then deflects back through zero as the station's frequency is passed, finally falling back close to zero when well off tune again. If a.f.c. is switched on the above effect is modified. On tuning towards a station the meter suddenly snaps to a reading as the a.f.c. captures the carrier. This reading may be either side of zero depending on how fast one tunes to the station. It is then only necessary to set the meter to zero for correct tuning, as the tuning cannot remain on any section of the S-shaped demodulator characteristic except the correct central portion, due to a.f.c. action. On tuning off the station with a.f.c. connected, the meter will deflect to a maximum in the appropriate direction and then snap near to zero as the a.f.c. 'loses' the station.

In designing the tuner I deliberately did not use a very strong a.f.c. control loop so that tuning could be done with or without a.f.c. In normal use it is often easiest to leave the a.f.c. permanently on, as tuning with moderate a.f.c. is so simple, especially for the less-skilled members of the family.

The cut-out for the dial panel is shown in Fig. 2.

Solid-state indicator

The recent improvement in the availability of light-emitting diodes at acceptable prices makes their use for indicators in electronic equipment very attractive. Fig. 3 shows a simple circuit using these diodes. Correct tuning (i.e. equality of output and a.f.c. reference voltages), is indicated by equality of the light output from the diodes D_1 and D_2. The diode D_3 and its associated resistors are for use as a stereo indicator lamp. The values chosen produce the approximate equivalent of a 6-V 40-mA lamp, and were chosen to match the requirements of the stereo decoder by Portus and Haywood (September 1970 issue).

The action of the circuit of Fig. 3 needs little explanation. It is a long-tailed pair which with equal voltages at the two bases will pass equal currents through the two diodes. When the base voltages differ the current through the long-tail resistor divides unequally between the two transistors, so that when the input to Tr_1 is approximately 1V below the input to Tr_2 then most of the current flows through Tr_2 and D_2. When the input to Tr_1 is approximately 1V above the input to Tr_2 then most of the current flows through Tr_1 and D_1.

Fig. 4 shows the relationship between the potential difference of the bases of Tr_1 and Tr_2 and the current through the two diodes as measured in the circuit of Fig. 3. The difference in peak currents off tune is of no practical importance as the visual difference is not great, and in any case there is no basis for comparison as either one light is on, or the other. In use the correct tuning point is easily found. Because the input impedance loads the output of the tuner, it is important to ensure that the input impedance of this tuning indicator is sufficiently high and linear, over a range of approximately ± 1V about the centre. Full ± 75-kHz deviation is equivalent at the tuner output to ± 0.7V. This requirement for reasonable input impedance linearity has been achieved in this circuit by degeneration in the emitters of Tr_1 and Tr_2, the resistor values being chosen to obtain the correct sensitivity.

The input resistance of the circuit is

$$\beta_1\left[r_{e1}+R_2+\frac{(R_4+r_{e2}+R_5/\beta_2)R_3}{R_4+r_{e2}+R_5/\beta_2+R_3}\right]+R_1.$$

Assuming minimum current gain for the

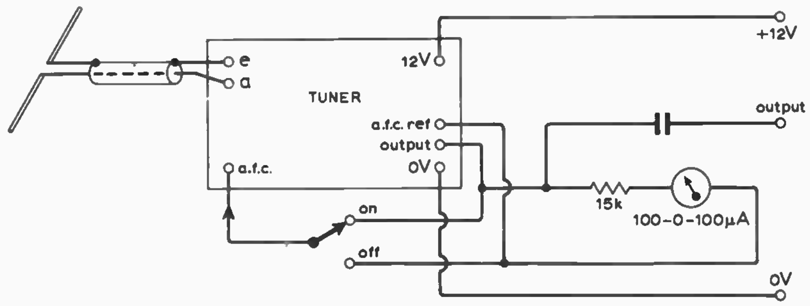


Fig. 1. Adding a moving-coil indicator is simple. With a.f.c. on and with a station tuned in, one merely sets the meter for zero indication.

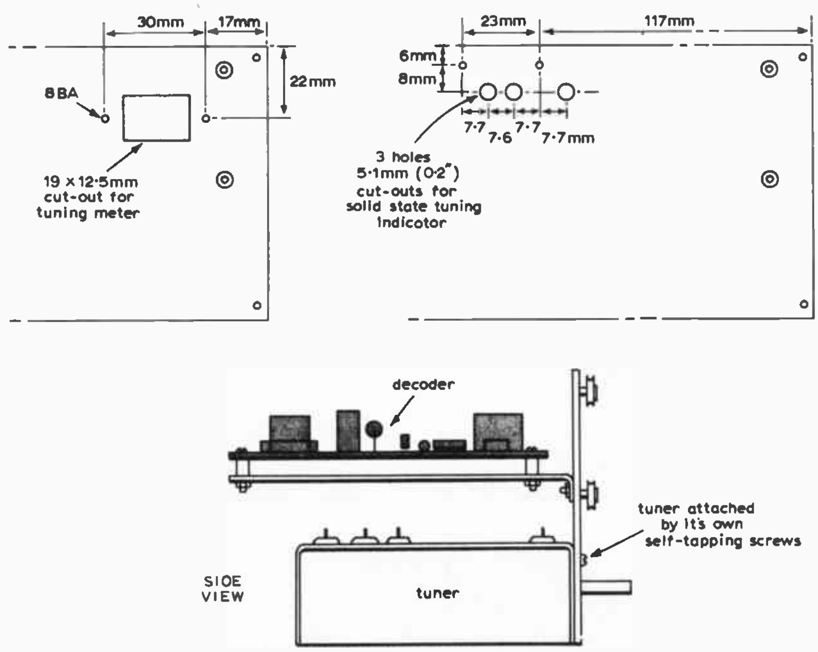


Fig. 2. Cut-outs for the two alternative tuning indicators.

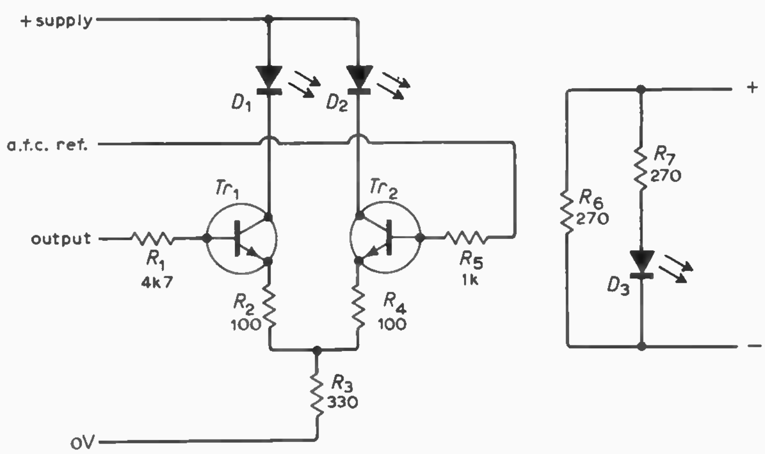
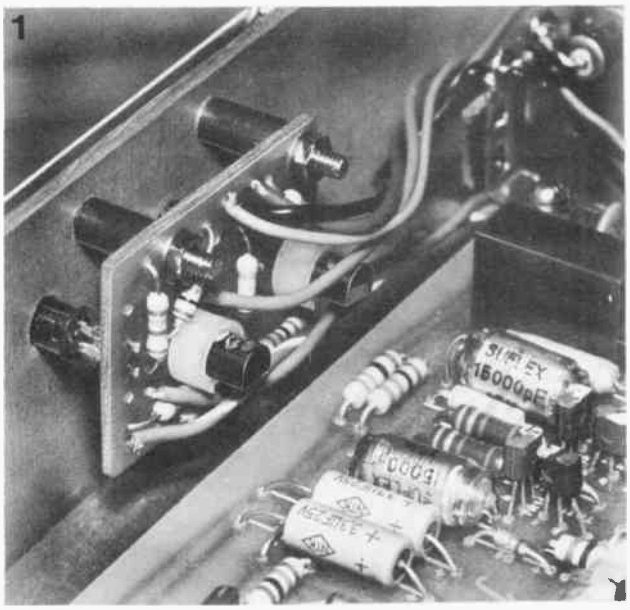
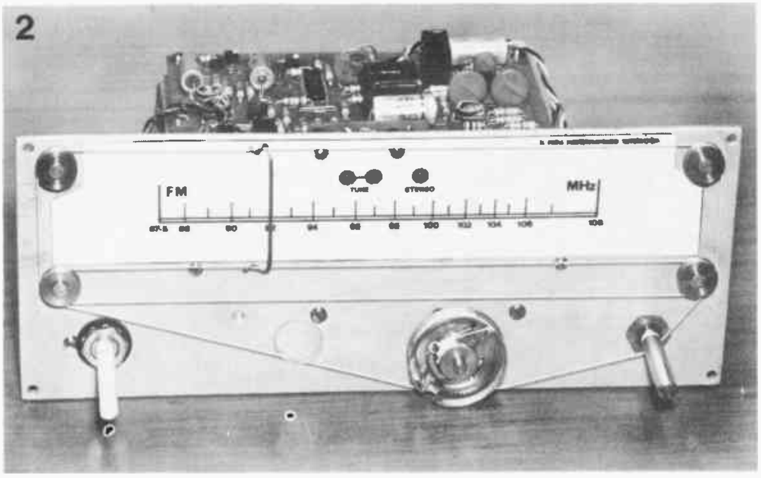
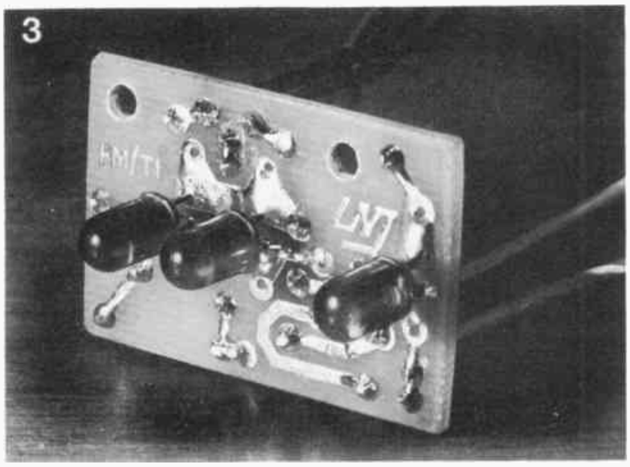
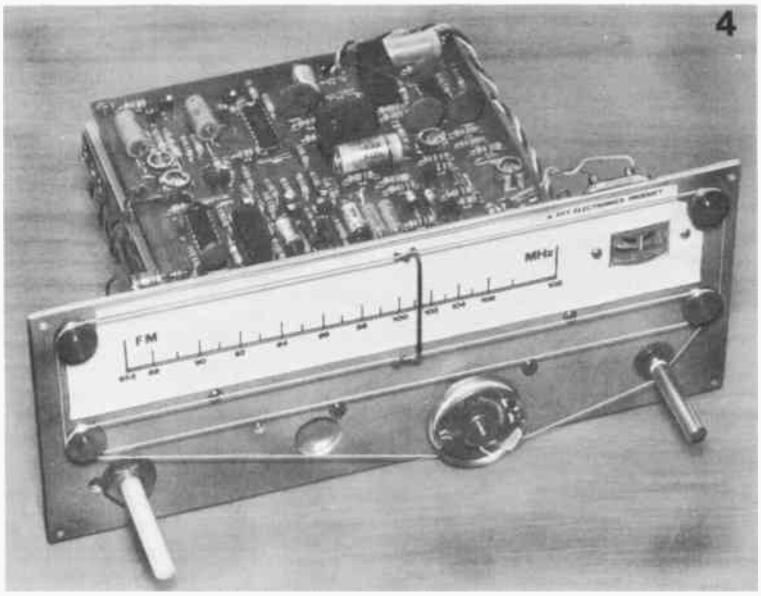


Fig. 3. Author's suggested circuit for solid-state indicator gives both tuning and stereo indication. Diode cathode is identified by the short lead and orange spot.









1 and 2: Mounting of solid-state indicator p.c. board. 3: Diodes mounted on p.c. board. 4: Front panel with moving-coil indicator.

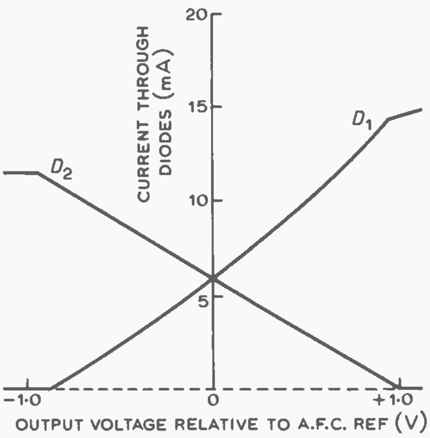


Fig. 4. Difference in off-tune current between the two diodes is not important as the lamps are never both fully on.

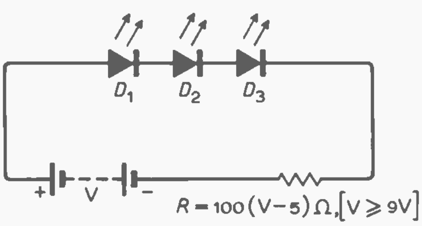


Fig. 5. The two diodes with most nearly equal brilliance can be selected for the tuning indicator when operated in series.

transistors of 220 and the values of Fig. 3 we get values of 47.35kΩ for V_{in} 0.7V above a.f.c. reference, and 46.50kΩ for V_{in} 0.7V below. The attenuation ratio formed by the output impedance of the tuner (2.2kΩ) and the tuning indicator circuit therefore varies by less than 0.08% over this span. Even if it were not for the need to linearize and raise the input impedance of the tuning indicator, R_2 and R_4 would still be needed, because with these two resistors effectively removed (by connecting together the two emitters of Tr_1 and Tr_2), the complete span from D_1 full on to D_2 full on is only ±0.2V. This sensitivity is too high for easy tuning and considerable flicker of the diodes is caused by the modulation of the carrier.

The resistors R_1 and R_5 are chosen to equalize the resistance seen by the two transistors, to minimize offset due to base current. These base resistors also ensure that the circuit does not become a resonant line oscillator when long leads are connected to the circuit. With modern planar transistors it is all too easy to get such effects if precautions are not taken.

Construction

The prototype tuning indicator is shown above, with the board mounted behind the tuning dial of a tuner. (A photograph shows the unit removed from the dial to show the mounting of the diodes.) The board is designed to take the type of diode having two pins on 2.5mm (0.1in) centres, and these are mounted on the circuit side. If desired, the board can be mounted remote from the diodes with leads from the board to the diodes.

Lamp matching. As there is some variation of brilliance of the diodes with identical currents from one lamp to another, it is desirable that the two lamps D_1 and D_2 be approximately matched for brilliance. This is easily done by temporarily connecting the lamps in series and passing a current through all three lamps at once (Fig. 5). The two lamps having most nearly equal brilliance are used for D_1 and D_2.

Avoid excessive heat or mechanical force on the leads of these lamps as they are easily damaged by heat, by the nature of the materials used and of their small size.

Components. Resistors should be $\frac{1}{8}$-watt, 5% tolerance and of the carbon film type. Transistors should have a $V_{cb} \geq 20V$ and an $h_{FE} \geq 220$ at I_C of 5mA, e.g. BC109, BC169, BC184L. Diodes are Hewlett Packard 5082–4440. All parts for the tuner, decoder, and indicator are available from Integrex Ltd, P.O. Box 45, Derby, DE1 1TW.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Ingenious tuning indicator

Readers may be interested in what I consider to be an improved solid-state tuning indicator for use with Mr Nelson-Jones' f.m. tuner.

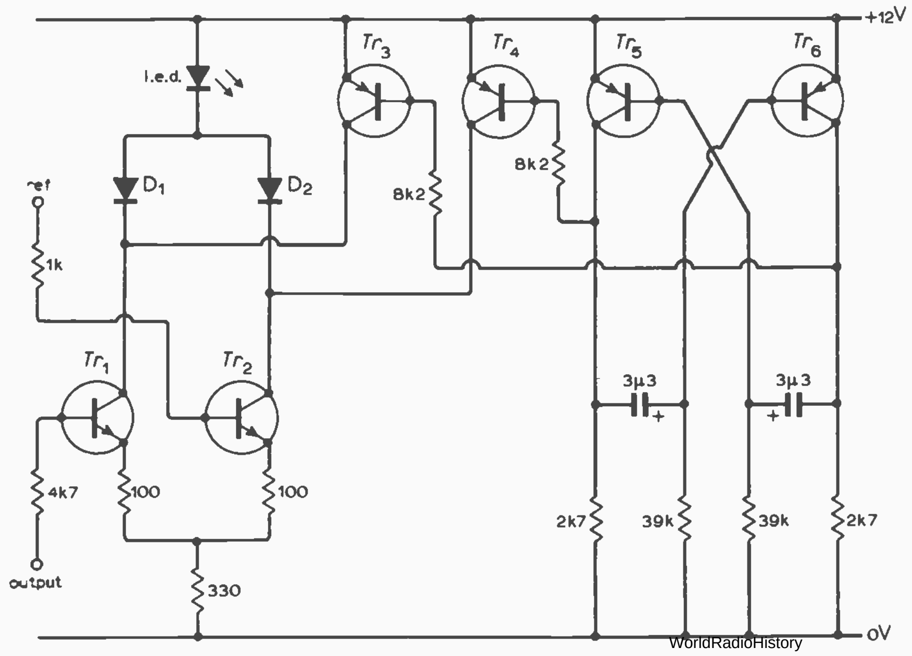
The circuit retains the basic long-tailed pair amplifier Tr_1, Tr_2 etc., as suggested by Mr Nelson-Jones, with the addition of a pair of switches, Tr_3 and Tr_4, driven alternately by an astable multivibrator Tr_5, Tr_6 etc.

When one of the switches say Tr_3, is "on" it shunts the collector current of Tr_1 to the supply rail; D_1 serving to isolate the switch from the l.e.d. Current through the l.e.d. passes through D_2 to the collector of Tr_2. The result is that the l.e.d. gives an indication of the current through Tr_2 on one half cycle of the multivibrator, and that through Tr_1 on the next.

When a station is correctly tuned the l.e.d. gives a steady light but any off-tuning is indicated by the diode blinking. The correct tuning capacitor position can be found far more accurately than with the two-light system, and probably with a similar accuracy to that obtained by using a centre-zero meter.

Component values indicated are those used in my prototype, the timing components giving a blink rate of the order of 10Hz, which I found to be suitable. Too slow a blink rate fails to give a fast enough response to a tuning correction. Diodes D_1 and D_2 were 1544; transistors 3, 4, 5 and 6 were ZTX500, but many other types are suitable.

M. Gavins,
Cheadle,
Cheshire.



The author replies: I read Mr Gavins' letter with interest — it is certainly a most ingenious method of indication, not only for tuning indicators but for many other purposes. I would point out, however, that it does suffer from the difficulty of not indicating the direction of out of balance.

The circuit as shown by Mr Gavins can surely be simplified by removing Tr_3 and Tr_4 together with the two 8.2k Ω resistors, and replacing them with two diodes from the collector of Tr_5 (and Tr_6) to the collector of Tr_1 (and Tr_2) (anodes to transistors 5 and 6, cathodes to 1 and 2). If in the circuit shown, Tr_5 and Tr_6 are silicon alloy types then all is well, but if planar types are used something must be done to reduce the excessive reverse *Veb* voltage to which these transistors will be subjected with a 12-V rail. Probably the simplest way is to connect two 2.2k Ω resistors from the collectors of Tr_5 and Tr_6 to the 12-V line, thus forming a potentiometer across each transistor and limiting the voltage on each to about 5.4V, which is usually below the zener breakdown of a base emitter diode. It may then be necessary to increase the two capacitors to 6.8µF, or thereabouts

My experiments with the tuning indicator circuit as published in the April 1972 issue indicate that the circuit provides a more than accurate enough method, as does the moving-coil meter. The tuner has a very wide discriminator characteristic and either tuning indicator will place the receiver on tune with the required degree of accuracy for stereo reception. Again my congratulations to Mr Gavins on an ingenious approach.

L. Nelson-Jones.

F.M. tuner instability

I was pleased to see Mr Nelson-Jones' article "F.M. Tuner Design — 12 months later" in the April issue.

This looked as if the answer to all the instability problems experienced was at hand. Various attempts to cure the instability in my particular tuner have been made, without success. The performance when tuned to the local f.m. stations has been excellent, but between stations the noise level has been high and television signals have been very troublesome.

The effect of adding the ferrite bead as suggested by Mr Nelson-Jones has only increased the number of spurious responses.

A determined attempt was then made to isolate the instability. The following additions were made to the circuit board:—

- An 80-ohm resistor was fitted between ground and aerial input tap of L_1 to terminate the aerial feeder cable. Instability appeared to be affected by moving a hand along the feeder, tending to indicate a standing wave on the feeder.
- The bypass capacitor of gate 2 of Tr_1 was increased by adding a 0.01 µF capacitor.
- The bypass capacitor across the 220-ohm source resistor of Tr_1 was increased by adding a 0.005 µF capacitor.
- The bypass capacitor decoupling the top end of L_2 was increased by adding a 0.01 µF capacitor.
- The bypass capacitor decoupling the + 12-V line supplying Tr_3 local oscillator was increased by adding a 0.01 µF capacitor.

All capacitors were disc ceramic and were soldered in place on the underside of the circuit board. The original components, on top of the board, were not disturbed. The result of these modifications was stable performance with much lower interstation background noise and the ability to peak the r.f. and mixer tuned circuits from 87.5 to about 100 MHz.

There is, however, still instability at about 100MHz. This appears to be oscillation in the r.f. stage; running a hand along the aerial feeder cable still has an effect on the instability in this range.

Have other readers or Mr Nelson-Jones any further views on the causes and cures for instability in this tuner?
Roy Ellis,
Sutton Coldfield,
Warwicks.

The author replies: There has recently been some evidence of the trouble Mr Ellis describes, and this has been traced to u.h.f. oscillation of the r.f. amplifier stage. The reason why the addition of four capacitors did not completely remove the trouble is that the inductance of disc capacitors of this type is still not low enough for the frequency at which these oscillations take place, which can be up to about 1GHz. Why these oscillations in some receivers are now more common is not clear but it may well be that the manufacture of the dual gate f.e.ts has improved to give a higher cut-off frequency. Certainly the Texas 3N201 seems more prone to this trouble than the 40673, although on paper they are near identical products. Two readers have written to recommend a simple cure for the fault and I have myself done a deal of work to trace the cause; all three of us agree on the cure, though we use rather different values of component.

This cure is achieved by inserting an impedance in the feed to g_1 of the r.f. stage Tr_1. I put in a 22-ohm resistor, one reader a 1 kΩ resistor and another a ferrite bead. The first and last will have very similar effects since the bead injects an effective resistance in the g_1 lead of around 20-30 ohms. The 1kΩ resistor is rather too high and will result in some loss of gain. The use of a resistor here will lower the Q of the input circuit slightly as it is in series with part of the total capacitance, but this effect has been found to be very slight. The resistor is the simplest to fit and should take the place of the link from g_1 of Tr_1 to the coil L_1, immediately next to Tr_1. The reason why the aerial feeder seems to take an active part is that the coupling of the feeder at the oscillation frequency is nothing like a good match, and hence gives rise to standing waves, and indeed the feeder may be the main resonant circuit element. This last point is backed up by the fact that with a direct 82-ohm resistor across the terminals of the tuner, or an attenuator box connected on short leads to the tuner, the effect is often absent. I am indebted to B. Brook, of Huddersfield, and J. R. Hooper (G3PCA), of Ilford, for their most helpful letters on this matter.

To sum up. The resistor fitted to replace the g_1 (Tr_1) to L_1 link should be of 22-47 ohms and may be a Mullard CR25 carbon film ⅛watt, or any carbon, metal or oxide film resistor of similar or smaller size. My sincere apologies to any readers who have suffered from this puzzling and very frustrating fault.
L. Nelson-Jones.

F.M. tuner and stereo

I was most impressed with the f.m. tuner design by L. Nelson-Jones, published in the April *Wireless World*, but I feel compelled to query the tuner's suitability for stereo reception or, to be more specific, the suitability of the FM-4 filters.

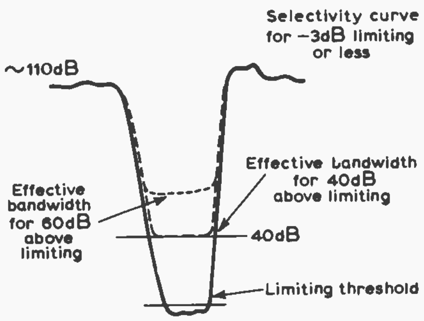
I think it is now generally accepted that, for good stereo performance, an i.f. bandwidth of 250-300kHz is required (the good old rule of thumb formula 2 (f_m+f_d) for bandwidth in f.m. systems gives a required bandwidth of 256kHz with a deviation of 75kHz and maximum baseband frequency of 53kHz). However, the manufacturer's data shows that the 3dB bandwidth of a single FM-4 could be as narrow as 200kHz and, if one was unlucky enough to obtain two filters which were at the minimum end of the bandwidth specification spread, the tuner would have a 6dB bandwidth of 200kHz, which would be ideal for mono, but virtually useless for stereo.

It could well be that the specification tolerances are wider than the actual production spread and the problem would then not arise, but I do feel that some assurance to this effect should be given, either by the author or by Vernitron, especially as the text of the article implies that the prototypes of the tuner have not actually been tried with a stereo decoder.
K. Clayson,
Redhill,
Surrey.

The author replies:

The quick answer is, yes the tuner is certainly suitable for stereo use, and since the script was originally written has been tested on stereo transmissions, on which it performs well.

The reason why would be more apparent had the second half of the article been available to Mr. Clayson, in that this contains the response curve of the i.f. amplifier together with a more thorough treatment of the mode of operation of the tuner. The figures of the requirements of a stereo tuner and of the performance of the FM-4 filters given by Mr. Clayson are essentially correct, but he has overlooked one important point, namely the good limiting performance of the tuner.



This results in an effectively wider bandwidth as shown in the graph, so that at the minimum signal strength at which stereo reception would give anything like an acceptable signal-to-noise ratio, the effective bandwidth is approximately 400 kHz, or more. It was partly for this reason that the limiting threshold of 0.18μV was set so low.
L. Nelson-Jones.

F.M. tuner bandwidths

L. Nelson-Jones, in his reply to K. Clayson (July issue) about f.m. tuner bandwidths, exposes a problem which I believe to be the subject of much woolly thinking. If this system were to be examined with a slowly sweeping input frequency, the response curves would indeed be as he indicates, the effective bandwidth being increased by the limiter; but then if only a slowly changing input frequency had to be dealt with a bandwidth equal to the peak-to-peak deviation would in any case be sufficient.

A normal f.m. signal, however, requires a larger bandwidth because it contains many sidebands extending beyond the ± 75 kHz region. So long as the signal passes through a linear system, we are equally entitled to think of it either in simple terms as a single sinewave of varying frequency or as a sum of carrier and sidebands. When it encounters a limiter, however, the sideband picture at once leads us into very deep water. The mind boggles at the prospect of analysing the response of a highly nonlinear system to an input which contains many frequencies, and we are forced back to the simple picture. I believe that the need for more than 150 kHz bandwidth can be explained in terms of this model by considering the phase/frequency response of the i.f. amplifier; which is, of course, closely related to the amplitude/frequency response. It then appears that the limiter will be of no value in correcting for an inadequate amplifier bandwidth.

Anyone who wishes to apply a rigorous mathematical analysis is welcome to do so, but perhaps it would be better for someone with the right facilities to test the matter experimentally. Before starting, though, we should question anew just what bandwidth an f.m. system does require, and what effects an inadequate bandwidth is likely to produce, particularly with a stereo signal.
Richard G. Mellish,
Heriot-Watt University,
Currie,
Midlothian.

F.M. Tuner Design — Two Years Later

Varicap diode tuning and lower gain modifications

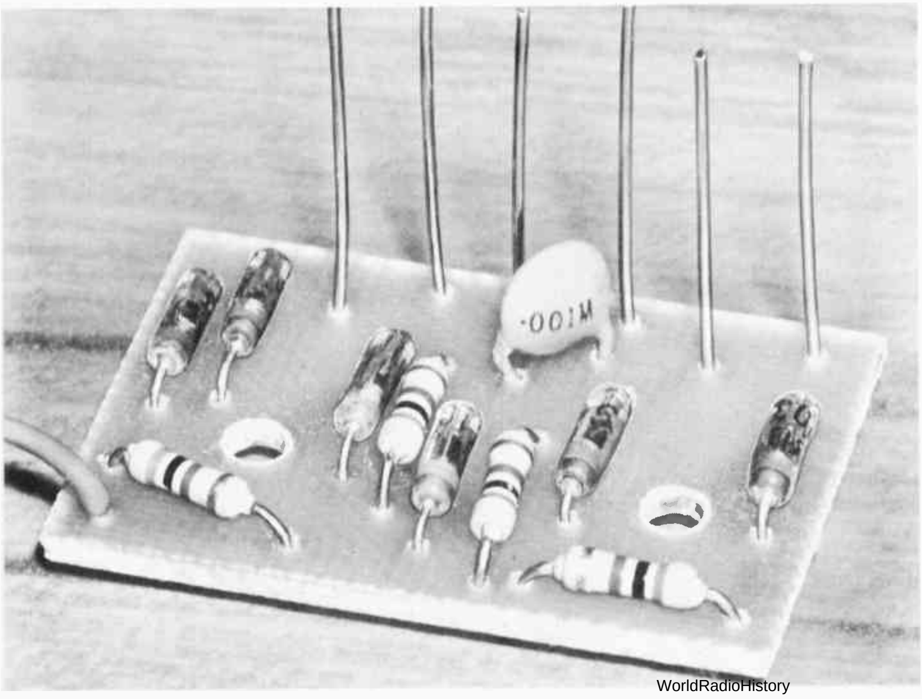
by L. Nelson-Jones, F.I.E.R.E.

Many readers have asked the author for details of a modification to the f.m. tuner published in April and May 1971 (and April 1972) which will enable it to be voltage tuned. The variable-capacitance diode design subsequently produced and described in this article covers the same tuning band of 87.5 to 108MHz. Also described is a simple modification to reduce the gain of the original tuner.

In recent years a number of types of silicon variable-capacitance diodes have appeared on the market, some having very high ratios of minimum to maximum capacitance. These diodes make use of the capacitance of the depletion layer in a reverse biased diode. All semiconductor diodes exhibit a voltage dependence of reverse biased capacitance, but the degree of variation and loss factor depend on the construction of the diode. Normal signal diodes, although exhibiting the effect, do not have a sufficient variation for tuning the f.m. band. The 87.5-108MHz band needs a minimum capacitance change of the order of 3:1. This apparently large ratio is necessary because of the relatively high stray capacitances in the circuit. In a tuner designed solely for varicap tuning these strays can be reduced somewhat, but here we are concerned with replacing a normal tuning capacitor without altering the layout.

For the design described below the BB103 diode has been chosen and the capacitance/voltage relationship for these diodes is shown in Fig 1. A wide capacitance range can be covered by the diodes (almost 5:1) if the bias can be allowed to drop to a value of some −0.3 volts. However, the range required can be obtained between approximately −2 and −27V. A low reverse bias has to be avoided in a tuner using f.e.ts since the oscillator voltage injected is high compared with a bipolar design. The diodes must not be allowed to conduct or the law will be spoilt in the oscillator tuning section by the resultant build up of charge at the low frequency end of the band where the diode bias is lowest. To assist in the application of the bias and to obtain a suitable value of capacitance, two diodes are used in each tuning section (see Fig. 2) The diodes are "back-to-back" and the bias is applied to the centre point. In the case of the oscillator and aerial input tuned circuits, the whole circuit is at ground potential and therefore only a high value feed resistor is needed to supply the centre point of the diodes with the necessary reverse bias. The value of the resistor must be high enough to avoid unnecessary reduction of the Q of the tuned circuits. A value of from 100k to 1MΩ is suitable. The tuned circuit coil in the drain circuit of Tr_1 is at the drain potential and a blocking capacitor has therefore to be added in this circuit. This in turn means the addition of a second high value resistor to ensure that the bias is applied to both diodes.

The varicap diodes and associated components are assembled on a small p.c.b. which takes the place of the tuning capacitor. The components are all mounted on one side of the board in the usual manner together with connection leads of 22s.w.g. tinned copper wire and the p.c.b. is then mounted on spacers. Components on the underside use the original mounting holes provided for the tuning capacitor. An external lead from this sub-board is then taken to the tuning control unit,



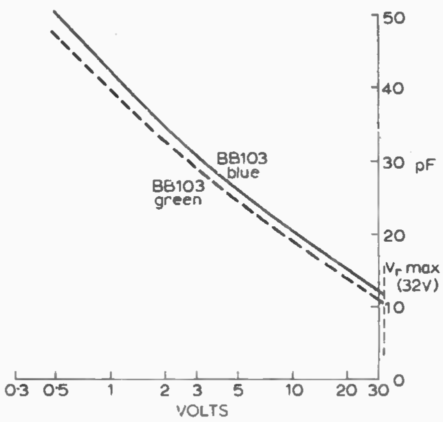


Fig. 1. Capacitance versus voltage relationship for the variable-capacitance diode alternatives.

Varicap board showing the diodes, resistors and capacitor, together with the connecting wires. The diode cathodes face away from the six connecting wires.

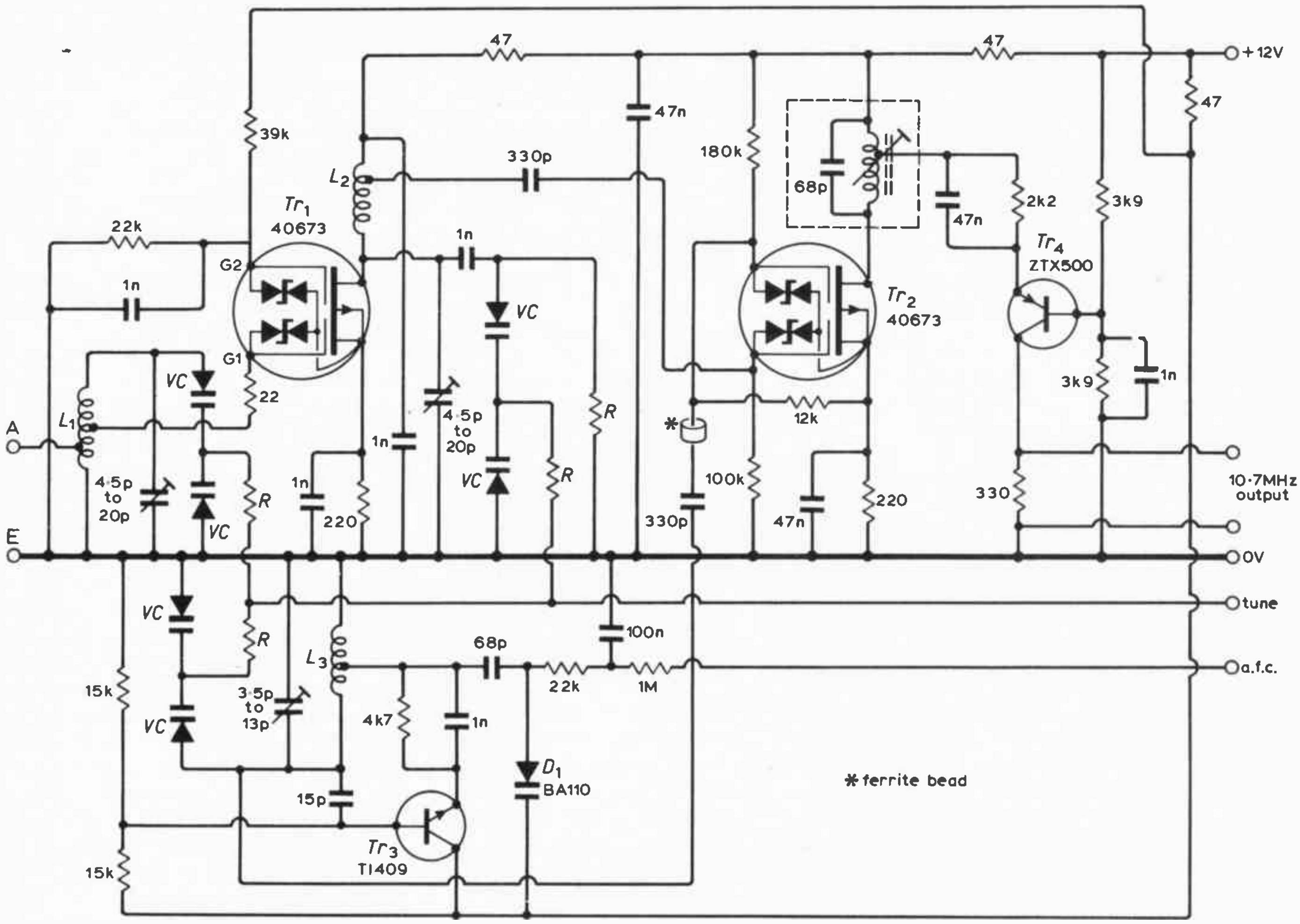
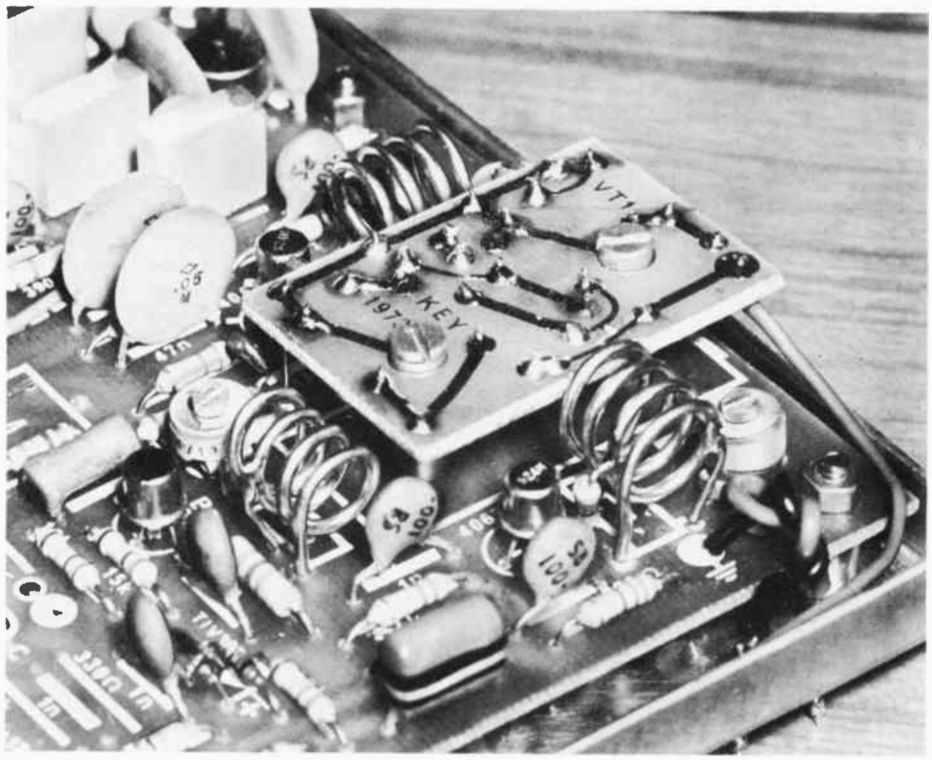


Fig.2. Revised front end of the tuner showing the replacement of tuning capacitors by varicap diodes which provide voltage tuning.

which is described below. This new p.c.b. is shown mounted on a tuner. A photograph shows the underside of the board before mounting on the tuner.

The only other changes needed to the tuner for varicap diode tuning is that the coils need to be replaced with slightly modified versions (shown in the photograph). The coils are all made from 18 s.w.g. tinned copper wire, but L_1 and L_2 are of a different diameter to the original coils. These two coils are now wound on a $\frac{5}{16}$in former. Inductor L_1 is otherwise as the original design, but on L_2 the tap has been moved to the end turn nearest the supply, and physically adjacent to the 100kΩ resistor. This change reduces the interaction between the tuning of the mixer and oscillator tuned circuits when trying to "track" the tuner, and although not great this reduction in interaction does make tracking much easier. Gain is not affected to any noticeable degree. The oscillator coil L_3 remains at $\frac{1}{4}$in internal diameter, but differs in construction. The direction of winding has been reversed, giving approximately an extra half turn. A 22Ω anti-parasitic resistor prevents the front-end "taking-off" at u.h.f.[1] This "vertical" resistor replaces the tap connecting L_1 to $G1$ of Tr_1 and is shown in the photograph just behind the 0.001μF capacitor in the foreground. Coils are 0.1in from the surface of the board as in the

General view of the varicap board fitted to the tuner.



original design. It must be emphasized that these new coil designs only apply to the varicap design.

Tuning control units

A tuner control unit is used with the varicap tuner which is similar to those now being fitted to most single standard u.h.f. television sets.

This particular unit has six selector buttons. Pressing any button releases the previously selected button, and turning the selected button adjusts the tuning. In these units six potentiometers with graded tracks are wired with all tracks in parallel, but with the sliders selected by the action of pushing a button. The grading of the track is such that an approximately linear relationship between potentiometer movement and frequency is achieved when used with a varicap tuner. Various manufacturers fit other extra facilities; for instance the tuning control unit has a switch made when any button is held fully in, so that when tuning a station by rotation of the control knob, and pressing it at the same time, the a.f.c. control can be shorted out. The method of connection of this switch to the tuner is shown in Fig. 3. Many of these tuning control units have such a switch, though not necessarily ganged to the selector mechanism. It is often an additional button at one end. Various values of total resistance are available but a common value is 100kΩ per track, giving a six-button unit a resistance of around 16kΩ with the six tracks in parallel.

Power supplies

The varicap tuner needs a highly stable noise free supply of around 30V to supply the tuning control unit, and the author has therefore devised a suitable circuit, Fig. 4, giving regulated supplies of 30V (nominal), and 12V. An optional 6V supply (stabilized) is also included for use with the Portus and Haywood decoder[2]. The Motorola MC1310 integrated circuit decoder also works well with the tuner and may be run off the stabilized 12V rail. The circuit of the 30V stabilizer uses the SGS TBA271 regulator, which is itself an integrated circuit, but is equivalent to a low temperature coefficient high voltage zener diode. For as good a regulation as possible the device is fed by a constant-current circuit, consisting of a transistor with emitter resistor and the base-emitter voltage stabilized by a zener diode. The temperature coefficients of the forward bias of the transistor and the zener diode are approximately equal and therefore cancel. The use of this constant current feed together with the low slope resistance of the TBA271 gives the main smoothing action of the supply, but this is augmented by the addition of a capacitor to reduce still further the ripple and noise. This capacitor is used together with the series impedance of one of the two pre-set potentiometers (used to set the span of the tuning control voltage) as a simple *RC* filter. The result is a supply with less than 100μV pk-to-pk noise. The 12V supply uses an SGS TBA625B integrated circuit series regulator. This gives a very simple circuit and is current overload protected above 100mA load.

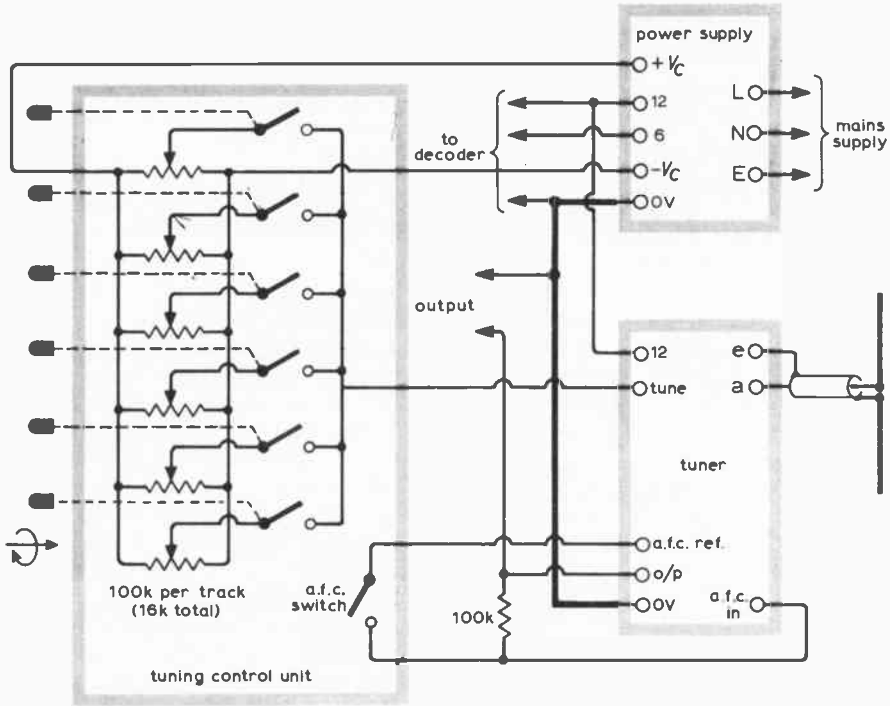
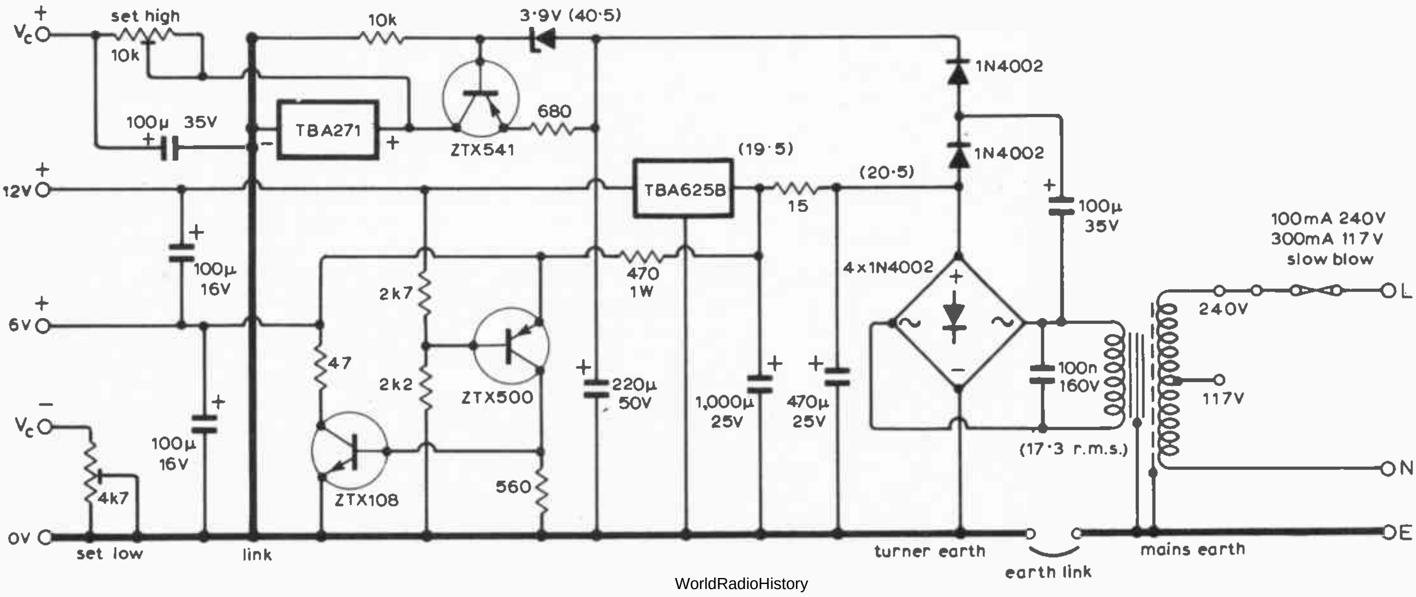


Fig.3. Method of connecting the switches of the tuner control unit to the tuner.

Fig.4. Stable noise free power supply giving regulated 30V and 12V and an optional stabilized 6V supply. Voltage levels in volts are shown in brackets at appropriate points on the diagram.



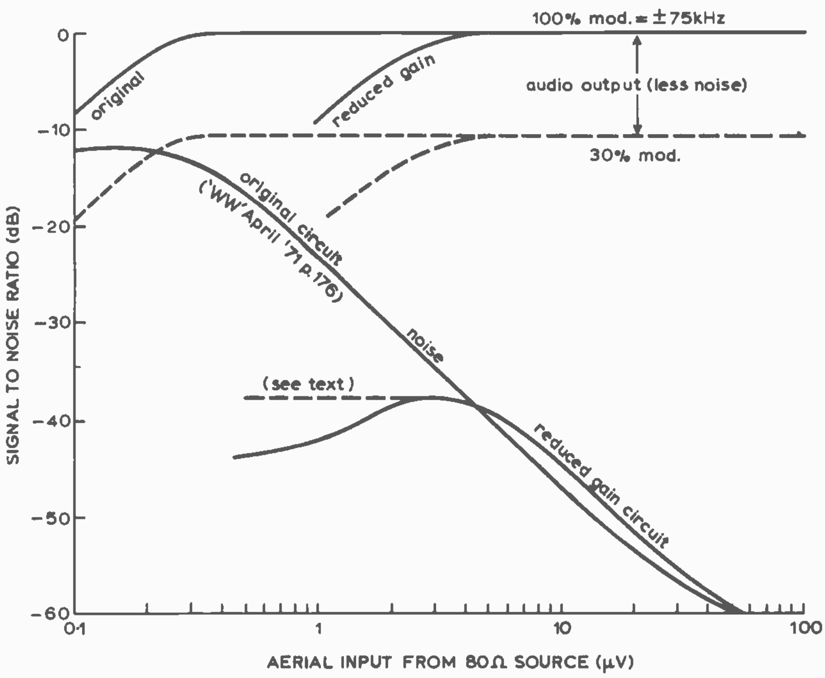


Fig. 7. Noise performance of the modified receiver compared with the original curves of the April 1971 article.

The rectification circuit for the 30V supply may seem a little odd at first until it is realized that it is a simple voltage doubler but "hung" on top of the rectified supply for the 12V regulator. The voltage doubler only gets half-wave pulses from the side of the bridge rectifier to which it is connected, but the output of the doubler (when added to the main supply) results in a supply approximately twice that of the main supply, without the added complexity of an additional winding on the transformer.

If the 6V supply is omitted then the capacitor between the 6V output and the 0V line should be omitted and replaced by a wire link so that the upper 100μF capacitor is connected to the 0V line.

Varicap setting-up procedure

The first stage in this procedure is to reduce the number of variables by setting the voltage limits of the tuner unit. The limits are set using the two variable resistors of the power supply circuit "set low" and "set high". The two controls interact to some extent and the adjustment must be repeated several times until both are correct. The "set low" control is set for a potential of +2.2V with respect to the 0V line at the bottom extreme of the tuner, and the "set high" control is set for +27.0V at the upper extreme of the tuner control unit. (If BB103-Blue diodes are used in the tuner then the "set low" should be set to +2.7V and the "set high" to +29.0V.)

Following this step the normal alignment procedure is followed, and this is easiest if the first two buttons of the tuning control unit are set to the two extremes of voltage. With the lowest voltage selected the inductors are set, and with the highest voltage selected the trimmer capacitors are set. Initially, it is easier to set the oscillator range in this way and then concentrate on the aerial and r.f. tuning, followed by a final check of all three tuned circuits.

It is important to state that in some tuning control units on the market there is a fixed resistor in series with the low voltage end of the potentiometers. It is therefore important to check the voltage actually available from the tuning control unit at the actual extremes of travel.

The relationship between voltage and frequency is shown in Fig.5, the voltage being plotted on a logarithmic basis.

Performance

Measurements on the prototype tuner indicate that the change to varicap tuning has had no effect on sensitivity or signal to noise ratio, though since the alterations were not made to the original prototype, it is not possible to be absolutely firm on this point. Certainly the performance is well up to the expected level and differs little from the original.

Components

Varicap diodes	Siemens type BB103 Green or Blue (all six diodes must be of the same colour selection. With this proviso matching should not be necessary).
Resistors *(R)*	any value between 100k and 1MΩ $\frac{1}{8}$ or $\frac{1}{4}$W carbon film.
Capacitor	0.001μF ceramic disc, preferably 500V low leakage type.
Mounting	$\frac{3}{8}$in 6 BA spacers (2) and $\frac{3}{4}$in × 6 BA cheese head screws (2) plus washers and nuts (2 each).

All components for tuning board, control unit and power supplies, are available from Integrex Ltd, P.O. Box 45, Derby, DE1 1TW. S.A.E. for details.

Lower gain version

The author has had a number of enquiries regarding the sometimes very high noise level on some tuners when tuning between stations. The optimum level of overall gain practical in a tuner with a given noise factor has been investigated. The conclusion has been that, although helpful to the long distance listener, the full gain of the original tuner does not add greatly to the practical usefulness of the receiver for listening to normal broadcast programmes at reasonable signal strength. Removing the first i.f. stage and coupling the mixer directly to the first i.f. filter provides the necessary simple modifications.

The actual noise level of a receiver is related not only to the noise factor of the receiver, but also to its gain, as even the quietest receiver will be noisy if it has sufficient gain. This is because the "noise factor" is merely a measure of the ratio of the noise output of the actual receiver, compared to that of an ideal receiver. This ratio is expressed usually in decibels. The source in both cases is resistive, and equal to the normal source impedance (in the case of this receiver an 80Ω source).

Since the tuner has a number of stages all subject to normal tolerances, it has been found (not unexpectedly) that some receivers are relatively quiet and others almost reach limiting on their own noise. Only ±1dB variation per stage means a variation of overall gain between receivers of 4 : 1, and ±2dB (roughly) ±20%)

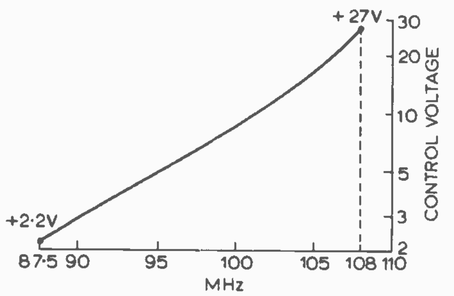


Fig.5. Voltage available from the tuning control unit as a function of frequency.

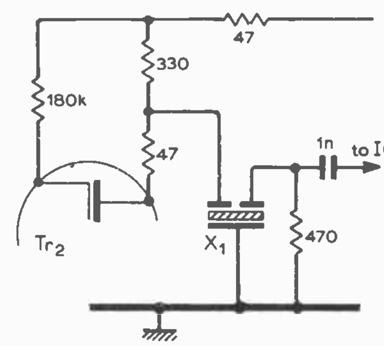


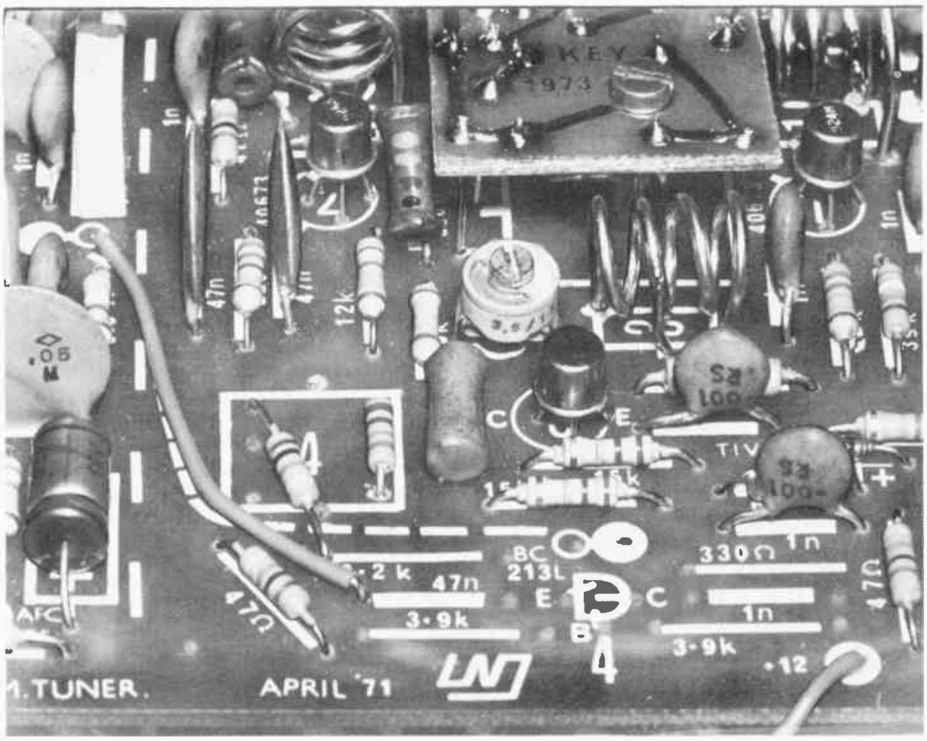
Fig.6. Arrangement after removal of the first stage of the i.f. amplifier for the lower gain version.

will give about 16 : 1 so the apparent differences in noise are hardly surprising.

A "noisy" receiver (but with a noise factor of approximately 3dB), was therefore taken and the first stage of the i.f. amplifier (Tr_4) was removed leaving the arrangement shown in Fig.6. The actual modification is shown (see photograph) where the blank component locations are clearly seen together with the two added resistors and the new link between r.f. and i.f. sections (now only a single wire). The 47Ω resistor lies diagonally across the location of the previously fitted L_4 and the 330Ω resistor is on the right of the outline of L_4.

The noise performance of the modified receiver is shown in Fig.7 which also shows the original curves of the April 1971 article. The curves should merge, but it must be remembered that the figures were measured on two different receivers, and the 2dB difference in the noise levels between 5 and 50μV is therefore quite a good agreement between measurements taken nearly two years apart. The noise level below 3μV should flatten off, not drop as shown. The reason for this is that a slight offset in the i.f. amplifier demodulator stages can result in one sided clipping of low noise levels and this is evident in the output waveform without any injected signal. The effect disappears as soon as the signal level rises appreciably above the basic noise level.

It has been found in this reduced gain version that the interaction between the tuning of the mixer and oscillator tuned circuits is greater though the reason is not immediately obvious. The effect is reduced to normal proportions by moving the tap on L_2 by one turn to the end turn nearest the 100kΩ resistor and Tr_2, as recommended in the varicap tuned version of the receiver. The curves of Fig.7 were taken with this change to L_2.



View of the area where the missing and additional components are located in the lower gain version.

The actual 30dB quieting level of the modified receiver is lower than that of the original receiver according to the curves of Fig.7 (about 1.1μV against 1.9μV), which only proves that anything taken out of context can prove almost anything. The performance should be more than adequate for all normal listening, and the author has received RADIO Bristol in Bournemouth fairly regularly on a simple loft mounted dipole.

References

1. Letters to the Editor, *Wireless World*, July 1972, p.318.
2. Portus, R. T. and Haywood, A. J. "Phase-locked Stereo Decoder", *Wireless World*, Sept. 1970, pp.418-422.

Phase-locked Stereo Decoder

Improved channel separation and low distortion obtained using an inductorless circuit

by R. T. Portus, A.M.I.E.E. and A. J. Haywood*, A.M.I.E.E.*

During a stereo broadcast the f.m. carrier transmitted by the B.B.C. is composed of three parts, as shown in Fig. 1. The components are:
1. Left plus right (L + R) forming the compatible mono signal containing frequencies up to 15 kHz.
2. Left minus right (L − R) which amplitude modulates a 38 kHz carrier. (The carrier is suppressed to better than 1% in order to make full use of the maximum deviation available at the transmitter.)
3. A low level "pilot tone" at half the carrier frequency, i.e. at 19 kHz, whose zero crossing points are coincident with those of alternate cycles of the 38 kHz suppressed subcarrier. This tone is used as a reference to regenerate the suppressed subcarrier at the receiver.

Stereo decoders

There are many forms of decoders,[1] the most popular being the switching type where the 19 kHz pilot tone is filtered by a tuned circuit, frequency doubled (to 38 kHz) and used to switch the complex signal—as shown in Fig. 2. Appendix III explains why complete channel separation is not achieved by such simple switching.

Limitations of tuned-circuit decoders

Poor separation results if the derived 38 kHz switching signal is modulated by any extraneous signal. This means that all the stereo information has to be removed by the tuned circuit. Also, if the output from the 19 kHz tuned amplifier contains any frequencies which pass through subharmonics of 38 kHz, the frequency multiplication stage will produce a modulation of the recovered 38 kHz. This in turn will produce sum and difference frequencies ("birdies") when switching the incoming complex signal. From the above considerations it may be seen that a high-Q tuned circuit is required. On the other hand, if the Q is made very large the phase of the recovered subcarrier becomes extremely drift sensitive. Any change in phase will reduce the separation. The effect of uncompensated phase error on separation is as follows.

* Rolls-Royce Ltd.

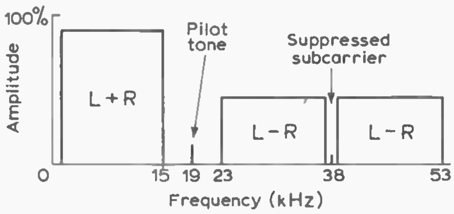


Fig. 1. Frequency spectrum of a stereo multiplex signal.

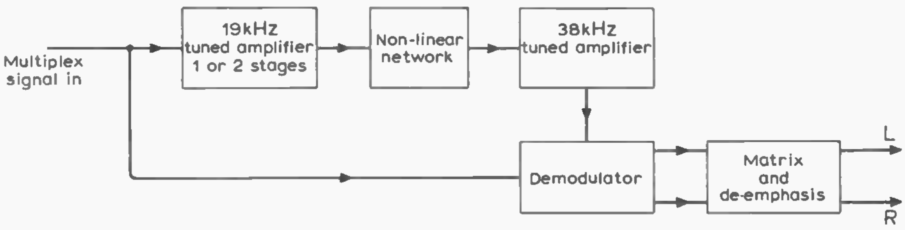


Fig. 2. Block diagram of switching stereo decoder.

Phase error at 19 kHz (deg.)	Separation (dB)
1	82·5
2½	54·5
5	42
10	30
15	23

Fig. 3 shows the change for small changes in tuned circuit L or C with various values of Q.

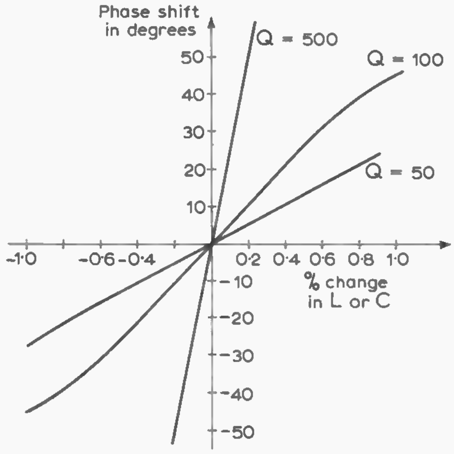


Fig. 3. Phase changes due to variation of L or C in a parallel tuned circuit.

Phase-locked decoder

Because of the shortcomings of tuned decoders the authors decided to use a phase locked loop to regenerate the 38 kHz suppressed subcarrier. A phase-locked loop is used to lock a locally generated signal in phase with an input signal whose frequency is liable to vary. Such a system may be given a very narrow bandwidth so that noise components of the input signal will not affect the loop operation.

If a locally generated 38 kHz signal is binary divided the remaining 19 kHz may be phase locked to the "pilot tone". This means that alternate zero crossings of the 38 kHz signal are coincident with those of the 19 kHz pilot tone. The locally generated 38 kHz signal is therefore in phase with the suppressed subcarrier.

The performance of such a system may be made superior to that of decoders using tuned circuits for the following reasons:
1. A phase-locked oscillator is a closed-loop system and so changes in component values, due to ageing, temperature etc., are corrected. Other systems have no comparison between input and output, and so errors may only be reduced by careful matching, the use of high quality components and accurate setting up.
2. The generation of "birdies" is very much reduced because the loop is given a low bandwidth, ensuring that phase modulation of the 38 kHz switching signal can only occur at low frequencies.
3. The setting up of a phase-locked loop is a simple matter, a d.c. voltmeter being the only instrument used. For other decoders an oscilloscope is usually required.

Performance of phase-locked decoder

The decoder (British patent application No. 35600/69) was tested using a Radiometer stereo signal generator, and the following specification obtained.

Frequency	Separation
80 Hz	28 dB
1 kHz	45 dB
10 kHz	40 dB

In a decoder fitted with a variable matrix the separation was 56 dB at 1 kHz.

Distortion at 1 kHz and with full modulation is 0·3% (predominantly 2nd harmonic), and the distortion introduced by crosstalk at 1 kHz is 67 dB down at full modulation. The mono and stereo gains are within 1 dB of each other. Cancellation is 45 dB. The input voltage range is 130 mV r.m.s. to 1·3 V r.m.s., and the output voltage 250 mV r.m.s.

The separation at 80 Hz could be improved by using matched components in the p.s.d. stage, but directional information is negligible at this frequency so that component matching is hardly worthwhile.

Operation of a phase-locked loop

The phase-locked loop (Fig. 4) can be looked upon as a servo-amplifier in which the 19 kHz pilot tone is used as an input reference phase, and the servo loop used to control a local oscillator in a fixed phase relationship. If there is a fixed phase between two signals then they must be at the same frequency. A phase-locked sub-carrier may be regenerated by deriving the 19 kHz from 38 kHz by use of a binary divider. Alternate zero crossings of the 38 kHz signal are therefore held locked to the incoming 19 kHz.

The operation of the loop is as follows. The balanced phase sensitive detector (p.s.d.) produces a d.c. output proportional to the difference in phase between the input frequency (f_{in}) and a locally derived frequency (f_o), plus higher frequency components produced by the chopping of the input signal. For a balanced p.s.d. the lowest frequency present is that produced by the highest audio frequency (15 kHz) beating with the 19 kHz chopping signal. The difference signal is 4 kHz (19 kHz − 15 kHz). This error signal is then amplified and low pass filtered by the error amplifier. The filtering removes the high frequency signals produced by the balanced modulator. The error amplifier output is used to control a voltage controlled oscillator (v.c.o.).

If the p.s.d. is not perfectly balanced small low-frequency audio signals will be present at the p.s.d. output. If the frequency of these signals approaches the loop bandwidth they will not be completely filtered out by the error amplifier. The resulting "ripple" on the error amplifier output will cause "jitter" on the v.c.o. output (i.e. phase modulation).

Loop operation is such that a change in phase between (f_{in} and f_o (e.g. due to drift)

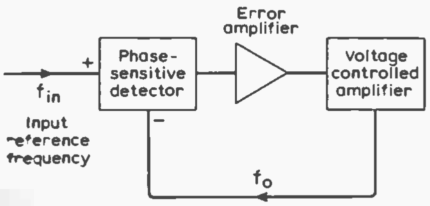


Fig. 4. Block diagram of phase-locked loop.

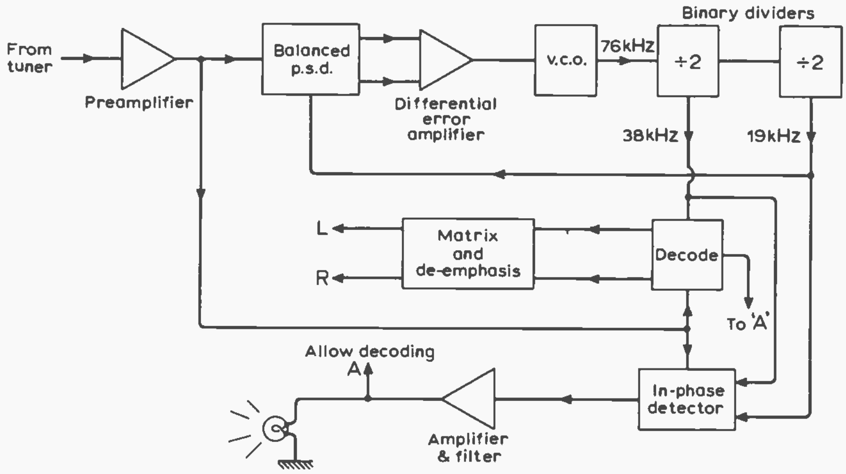


Fig. 5. Complete stereo decoder system built round the phase-locked loop.

causes the output of the p.s.d. to drive the v.c.o. in such a direction as to hold the phase of f_{in} locked to that of f_o.

The system behaves as a servo-loop with a bandwidth determined by the time constant of the filter in the error amplifier.

When deciding on loop bandwidth two points have to be considered. These are "jitter" on the v.c.o. and the time required to pull into lock. The first point is important because phase jitter will reduce channel separation. Jitter is minimized by having a low loop bandwidth.

On the other hand if loop bandwidth is made too small the loop will take a long time to pull into lock.

A compromise must therefore be made and a bandwidth of 25 Hz was found to be adequate.

Theory and design equations for the loop are given in appendices I and II.

Practical decoder system

In Fig. 5 the pre-amplifier acts as a buffer on the tuner output. Its gain is made variable so that a fixed level of pilot tone is presented to the loop independent of the tuner output level.

The phase detector is balanced, and the error amplifier is made differential in order to maintain balance. The output from the error amplifier constitutes the control voltage of the voltage controlled oscillator (v.c.o.) which is free running at approximately 76 kHz.

By the use of binary dividers square-wave outputs are obtained at 38 kHz and 19 kHz. The former output drives the decoding transistors and the latter the loop p.s.d. transistors. The decoder outputs are matrixed and de-emphasized to provide the left and right channel outputs.

On acquiring lock an output is obtained from the in-phase detector. This output is amplified and filtered, and used to allow decoding to begin, and to drive a stereo indicator lamp, if required.

Circuit description

The complete decoder circuitry is shown in Fig. 6. Tr_1 and Tr_2 comprise a pre-amplifier whose output level is adjusted by RV_1 to set the p.s.d. gain constant. The p.s.d. consists of R_5, R_7, Tr_3 and Tr_4, and these transistors are used in the inverted mode to minimize offset voltage. Since the bases of Tr_3 and Tr_4 are driven alternately at 19 kHz the p.s.d. and differential error amplifier IC_1 form a double-sideband suppressed-carrier modulator, the output of which is modified by the error amplifier response.

Audio signals close to the loop bandwidth are attenuated by C_3. The phase shift due to C_3 ($10\frac{1}{2}°$) is corrected for in the matrix which is fully described in a later section.

The error amplifier IC_1, type U6E7709393 or similar, is a low cost differential inte-integrated operational amplifier. This is chosen for its high gain and low input offset voltage. At ±6 V supply the typical open loop gain is 72 dB.

C_4 and C_6 provide an additional h.f. roll-off beyond the loop unity gain frequency so that loop stability is not affected.

The amplifier response is tailored to give an overall loop bandwidth of 25 Hz with an error response damping factor of 0·707. The d.c. gain is chosen from a consideration of the static error due to v.c.o. drift. The error is $<1°$ at 19 kHz.

RV_2 provides an adjustment for input offset voltage, which would otherwise appear as a phase error. The low input offset voltage temperature drift of IC_1 ensures good phase stability.

The v.c.o. (Tr_6 and Tr_7) is a conventional astable multivibrator with the timing voltage of the bases controlled by IC_1. It is arranged to free run at approximately 76 kHz. RV_4 provides a fine frequency control by adjustment of the mark/space ratio.

R_{25} and D_1 ensure that the v.c.o. will not operate above 80 kHz, by clamping the error amplifier input to approximately +0·6 V.

The 76 kHz signal is then applied to IC_2 which is a dual D-type SN7474 or similar binary divider. IC_2 will then provide a 38 kHz and a 19 kHz switching signal with a 1 : 1 mark/space ratio. The 19 kHz signal is applied to the loop p.s.d. transistors Tr_3 and Tr_4, thus closing the loop. The 38 kHz signal is applied to the decoding transistors Tr_{11} and Tr_{12}.

The bases of all the switching transistors are pulled negative to remove stored charge and to hold the transistors off in the presence of audio signals.

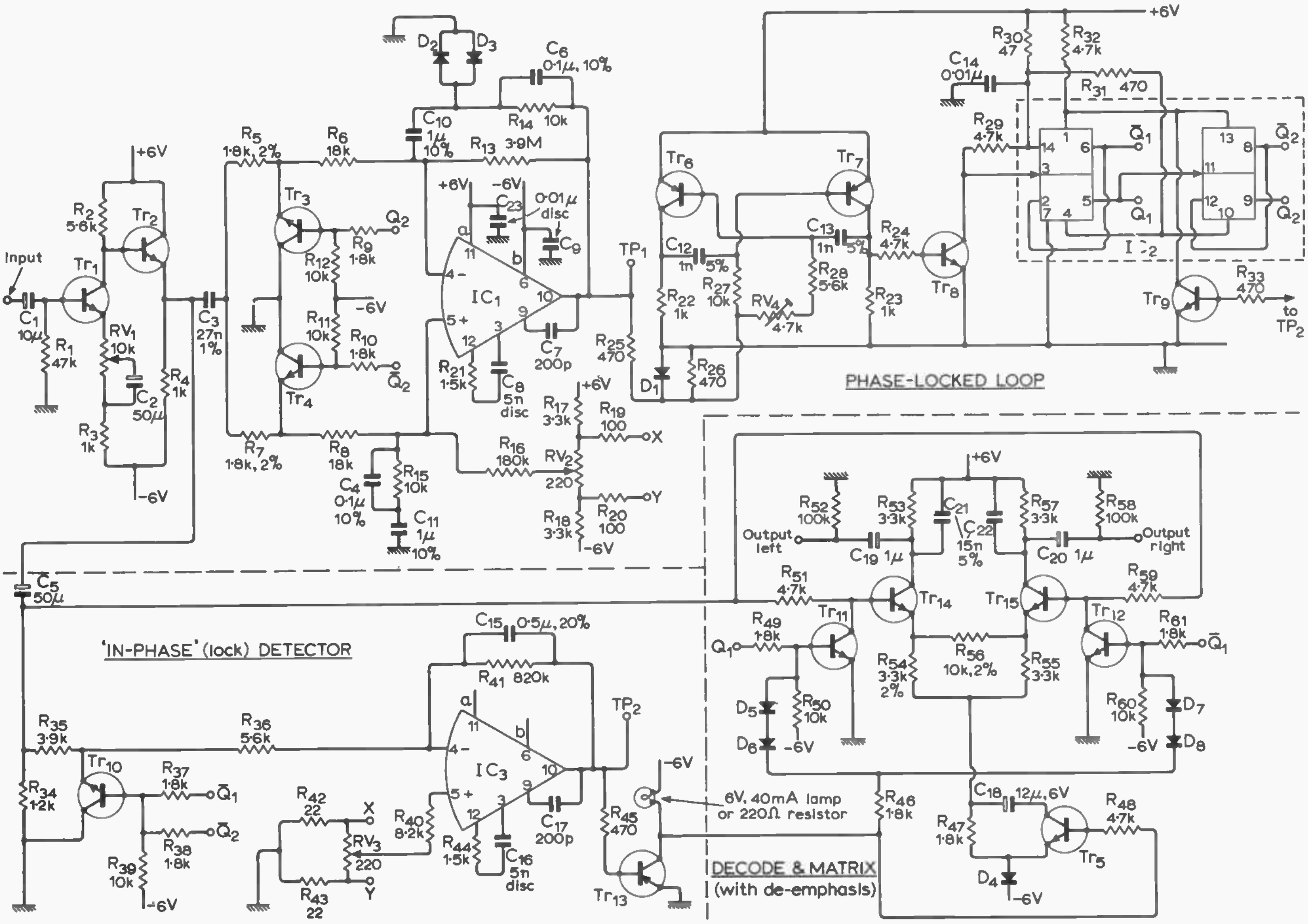
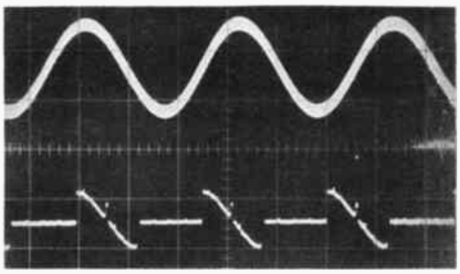


Fig. 6. Circuit diagram of decoder. Integrated circuits IC_1 and IC_3 are type U6E770 9393 or equivalent, and IC_2 type SN7474N or equivalent. Transistors can be BC108 or equivalent for n-p-n types, and ZTX500 or equivalent for p-n-p. Diodes are 1S44 or equivalent. Resistors can be ¼W 5% tolerance unless specified otherwise.

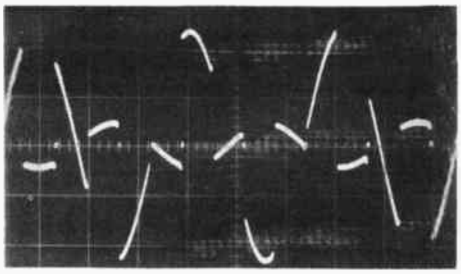
The action of D_2, D_3 is to speed up lock-in at switch-on for outputs greater than ± 0.6 V. The diodes shunt current away from C_{10} and effectively increase loop bandwidth. Between ± 0.6 V the diodes have no effect.

In-phase detector. Transistor Tr_{10} is driven by both binary dividers so that it is off for ¼ cycle only of the 19 kHz. Thus a pulsed d.c. component is produced when the loop is locked. This signal is filtered and amplified by IC_3 and used for the following purposes:

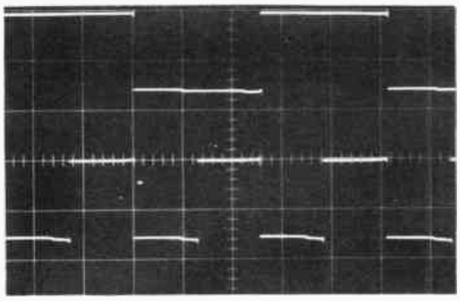
1. To turn on a stereo indicator lamp, if required.
2. To allow decoding to begin in Tr_{11} and Tr_{12}. If the decoder is allowed to switch during mono transmission, supersonic noise around 38 kHz will be heterodyned into the audio bandwidth causing a deterioration in s/n ratio. Also from some transmitters a 23 kHz low-level tone is present during mono transmission. This tone would produce a "beat" at 15 kHz in a free running decoder. For this reason the v.c.o. is not



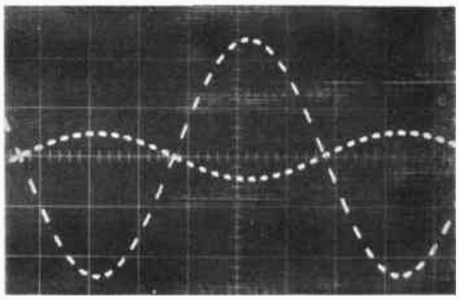
Upper: pilot tone + 150 Hz (A = B).
Lower: emitter of Tr_3.



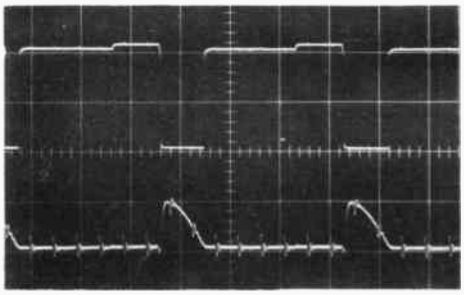
Matrixing of 15 kHz signal (A = B). No de-emphasis. Pilot omitted.



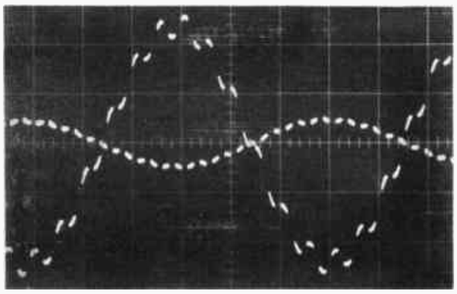
Upper: 19 kHz binary output.
Lower: 38 kHz binary output.



Matrixing of 1·5 kHz signal (A = B). No de-emphasis. Pilot tone omitted.



Upper: Tr_{10} base drive.
Lower: Tr_{10} emitter.



As above with 19 kHz pilot tone included.

Fig. 7. Oscillograms of typical waveforms generated by the locked-loop.

allowed to operate above 80 kHz, i.e. the binary output cannot exceed 20 kHz.
3. To ensure that the l.h. and r.h. channels do not reverse in the presence of noise etc. If the loop initially locks in the wrong direction i.e. l.h. and r.h. channels are reversed, a positive output will be produced by the in-phase detector, Tr_9 will be turned on thus clearing the binaries, and reversing l.h. and r.h. This will also be the case if a line transient flips one binary only. In the writers' experience the loop has never locked in the wrong direction. Typical waveforms of the loop in operation are given in Fig. 7.
4. To keep the mono and stereo gains approximately constanf via Tr_5.

Matrix. Once the signal has been decoded by Tr_{11} and Tr_{12} it is matrixed and de-emphasized via Tr_{14} and Tr_{15}. Matrixing is necessary in any switching decoder as may be seen from the theory given in appendix III.

Power supply. The circuit of Fig. 6 is shown for a supply of ± 6 V. Fig. 8(a) shows a suitable low cost regulator using a miniature mains transformer.

Single supply operation is accomplished from an unregulated (15 V minimum) supply by altering the "earth" rail from the 0 V to the -6 V line as shown in Fig. 8(b).

In this case there are two minor circuit alterations: capacitor C_1 is reversed; and resistors R_{52} and R_{58} are returned to the former -6 V rail i.e. instead of the former 0 V rail. The signal "earth" line becomes, of course, the former -6 V line.

Setting up procedure

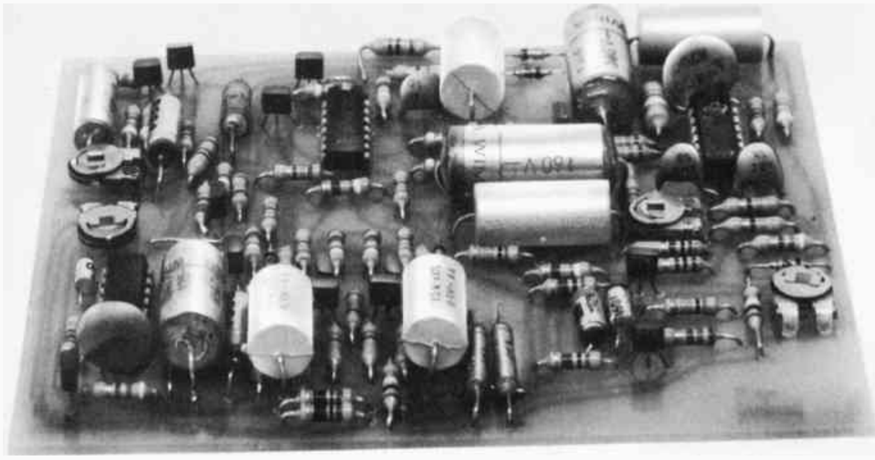
The decoder is set up using a d.c. voltmeter. The stages are as follows:
1. With no input signal adjust RV_2 and RV_3 for 0 V at test points TP_1 and TP_2 respectively.
2. Apply a stereo signal and adjust RV_1 until a negative voltage 1·5 V appear at TP_2 (Adjust RV_4 if necessary).
3. Adjust RV_4 to bring TP_1 back to zero.
4. Adjust RV_1 to bring TP_2 to $-1·5$ V.

Operation

Several decoders have been built and tested by the authors, using both single and dual supply operation.

Over the last year they have proved to be reliable and have required no adjustment beyond the initial setting up procedure.

The high inherent separation, in particular at high frequencies, produces a marked improvement in performance over that obtained with more conventional decoders.



Decoder components mounted on a printed circuit board approximately 100 mm × 125 mm.

Notes

This decoder design arose from work done on a phase-locked loop frequency multiplier made to improve the resolution of an engine tachometer.

The authors are grateful to J. W. Hill and D. L. Lynas for the use of equipment and for helpful criticism, and thank Rolls-Royce Ltd for permission to publish.

A kit of parts including a printed circuit board is available, mail order only from Integrex Ltd, P.O. Box 45, Derby, DE1 1TW. The price is £8 19s 6d plus 2s 6d for carriage and packing. This printed circuit board is adapted to single or dual supply options. The size of the board is 100 mm × 125 mm (approx.) and it may be seen from the photograph that a high packing density has been achieved. To simplify assembly, component positions have been marked and the track is roller tinned. A power supply kit is also available.

REFERENCE

1. "Receiving Stereo Broadcasts", *Wireless World*, September 1966.

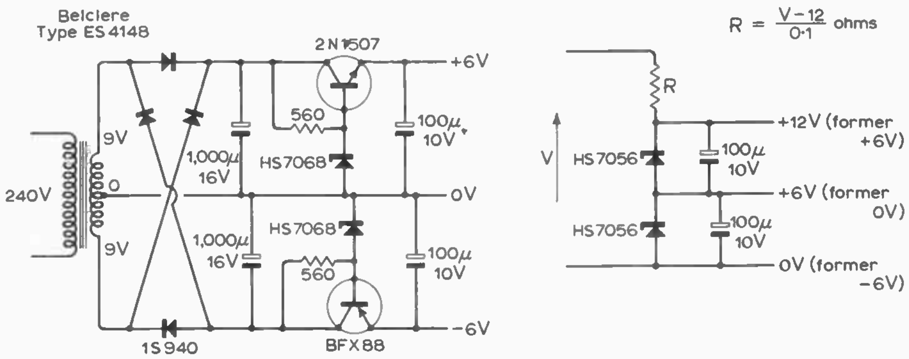


Fig. 8. Decoder power supply arrangements. (a) Stabilized supply for ± 6 V. (b) Supply obtained from an unregulated 15 V (or greater) d.c. source.

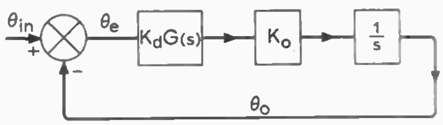
Appendix I

Loop Theory

The constants used are defined as follows:

K_d = phase detector gain constant in volts/radian.
K_o = v.c.o. gain constant in radians/s/V ($= 2 \times$ Hz/V).
$G(s)$ = error amplifier transfer function
θ_{in} = input reference phase angle.
θ_o = locally generated phase angle
θ_e = phase angle error ($= \theta_{in} - \theta_o$).

The loop block diagram is redrawn below.



The integration term (1/s) is introduced by the conversion of the frequency of the v.c.o. to its phase.

By inspection of the diagram

$$\frac{\theta_o}{\theta_{in}} = \frac{K_oK_dG(s)\,1/s}{1+K_oK_dG(s)\,1/s} = \frac{K_oK_dG(s)}{s+K_oK_dG(s)} \quad (1)$$

The error amplifier has a response that can be shown to be

$$G(s) = \frac{A(1+sT_2)}{1+s(T_1+T_2)} \quad (2)$$

where

$$A = \frac{R_1}{R_3}; \quad T_1 = CR_1; \quad T_2 = CR_2$$

Substituting (2) in (1) we get

$$\frac{\theta_o}{\theta_{in}} = \frac{[K_oK_dA/(T_1+T_2)](1+sT_2)}{s^2+\dfrac{sK_oK_dAT_2}{T_1+T_2}+\dfrac{K_oK_dA}{T_1+T_2}} \quad (3)$$

Where it is assumed $K_oK_dAT_2 \gg 1$.
This is of the form

$$\frac{\theta_o}{\theta_{in}} = \frac{2\eta\omega_n s+\omega_n^2}{s^2+2\eta\omega_n s+\omega_n^2}$$

where using servo terminology ω_n = natural (undamped) frequency of the loop and η = loop damping factor.

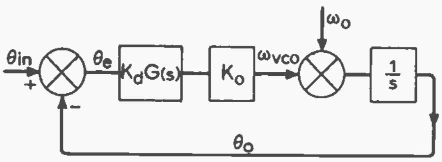
$$\text{Thus}\quad \omega_n = \left(\frac{K_o K_d A}{T_1+T_2}\right)^{\frac{1}{2}} \qquad (4)$$

$$\eta = \frac{T_2}{2}\left(\frac{K_o K_d A}{T_1+T_2}\right)^{\frac{1}{2}} = \frac{T_2}{2}\omega_n \qquad (5)$$

Static phase errors

Since the phase loop contains an integrator term then the steady state phase error will be reduced to zero if the v.c.o. free-running frequency is the same as that of the input frequency.

The response of the loop to a disturbance causing drift in the v.c.o. free running frequency can be obtained from the figure below.



where

$\omega_{v.c.o.}$ = free running v.c.o. frequency

ω_o = disturbance (or drift) frequency.

It can be seen that

$$\frac{\omega_{v.c.o.}}{\omega_o} = \frac{1/s.K_o K_d G(s)}{1+1/s.K_o K_d G(s)} \qquad (6)$$

By referring $\omega_{v.c.o.}$ back to an equivalent θ_e we get

$$\frac{\theta_e}{\omega_o} = \frac{1/s}{1+1/sK_o.K_d.G(s)} = \frac{1}{s+K_o K_d G(s)} \qquad (7)$$

For a step input of $\omega_o = \Delta\omega_o$ then

$$\theta_e.s = \frac{\Delta\omega_o}{s}.\frac{1}{s+K_o K_d G(s)} \qquad (8)$$

Applying the final value theorem we get

$$\text{steady state phase error} = \frac{\Delta\omega_o}{K_o K_d A}$$

$$(\text{since } G(0) = A) \qquad (9)$$

This phase error is required in order to pull the v.c.o. away from its free-running frequency. By making $K_o K_d A$ sufficiently large the phase error can be made as small as required.

For those who wish to pursue phase-locked loop theory a very good account is given in "Phaselock Technique" by Floyd M. Gardner, published by John Wiley and Sons Inc.

Appendix II

Loop equation constants

P.S.D. gain constant. For a phase error γ, when the loop is locked the mean output voltage from the phase detector, V_o, is given by:

$$V_o = \frac{V_{in}}{\pi}\int_{\frac{\pi}{2}+\gamma}^{\frac{3\pi}{2}+\gamma} \sin\theta\, d\theta$$

where $V_{in}\sin\theta$ is the input signal.

Thus $|V_o| \approx \frac{2}{\pi}\times V_{in}\times\gamma$ where γ is small.

Thus, if we choose an input signal of 200 mV peak to peak amplitude:

$$K_d = 63{\cdot}7 \text{ mV/radian}$$

or $$\underline{K_d = 63{\cdot}7\times 10^{-3} \text{ V/radian}}$$

Steady state error. From equation (9) it may be seen that the steady state error is given by:

$$\text{Error} = \frac{\Delta\omega_o}{K_o K_d A}$$

where $\Delta\omega_o$ is the frequency difference of the free running v.c.o. and the pilot signal, K_d is the phase detector gain constant and K_o is the v.c.o. gain constant.

Now

$K_d = 63{\cdot}7\times 10^{-3}$ V/rad and

$K_o = 1{\cdot}2\times 10^3 \times 2\pi$ rad/s/V (measured)

For the v.c.o., the frequency drift is approx. 0·075%/°C. It follows that for a steady state error of 1° over 20°C temperature change.

$$\underline{A = 214}$$

From equation (4)

$$\omega_n = \frac{K_o K_d A}{T_1+T_2} \qquad T_1+T_2 = 4{\cdot}16 \text{ s}$$

From equation (5)

$$\eta = \frac{T_2\omega_n}{2} \;\therefore\; T_2 = 9 \text{ ms}$$

Appendix III

Matrixing in switching decoders

In an f.m. stereo broadcast the instantaneous deviation of the transmitter is given as:

$$f = 0{\cdot}9\left[\tfrac{1}{2}(A+B)+\tfrac{1}{2}(A-B)\sin\omega_s t + 0{\cdot}1\sin\frac{\omega_s}{2}t\right]$$

where $\frac{\omega_s}{2\pi} = 38$ kHz

The pilot tone is $0{\cdot}1\sin\frac{\omega_s}{2}t$.

In a switching decoder the input signal is treated as if it were time division multiplexed. The signal is multiplied by a square wave:

$$= \left(\frac{1}{2}+\frac{2}{\pi}\sin\omega_s t+\frac{2}{3\pi}\sin\omega_s t+\ldots\right)$$

Suppose there is phase error ϕ existing between the 38 kHz carrier used for modulation and the local 38 kHz decoding signal.

The output signal will be proportional to V_1 where:

$$V_1 = [(A+B)+(A-B)\sin\omega_s t]\left[\frac{1}{2}+\frac{2}{\pi}\sin(\omega_s t+\phi) + \frac{2}{3\pi}\sin(3\omega_s t+\phi)+\cdots\right]$$

$$\left(\text{neglecting } \sin\frac{\omega_s}{2}t\right)$$

$$= \left(\frac{1}{2}+\frac{1}{\pi}\cos\phi\right)A+\left(\frac{1}{2}-\frac{1}{\pi}\cos\phi\right)B$$

plus modulation around ω_s, $3\omega_s$, etc, which is of the form:

$$V_1 \propto (A+\Delta B)$$

where Δ is dependent on the phase shift ϕ.

If the decoding signal is shifted by 180°, the output signal may be shown to be proportional to V_2 where: $V_2 \propto (B+\Delta A)$.

The purpose of matrixing is to subtract ΔV_2 from V_1 and vice-versa.

$$\text{i.e. } V_1-\Delta V_2 \propto A+\Delta B-\Delta B-\Delta\Delta A$$
$$\propto A-\Delta\Delta A$$

It may be seen that signals from channel A only are present in one output and from channel B in the other.

Thus, if the phase shift ϕ is known, and hence Δ, complete separation of an ideal broadcast is achieved.

Phase-locked-loop Stereo Decoder I.C.

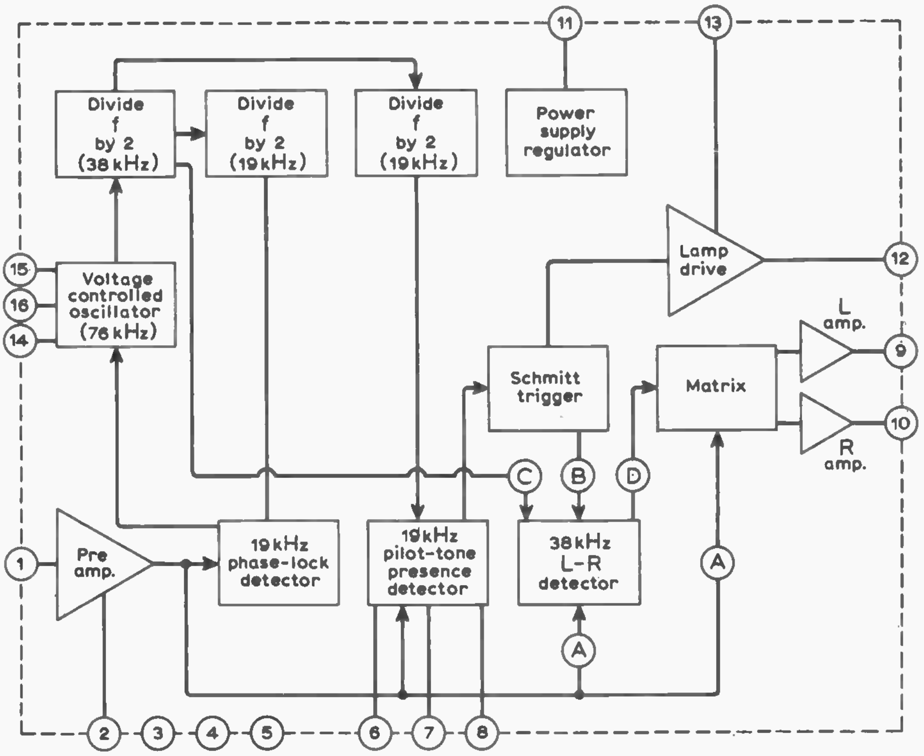
Build a high-performance decoder with the minimum number of components

It is possible to make a high-performance phase-locked-loop stereo decoder with just sixteen components and a printed circuit board. Only one coil is required and only one adjustment is necessary. The major component in the decoder is an integrated circuit (CA3090Q), containing 126 transistors, which has just been introduced by R.C.A.

A block diagram of the i.c. is given in Fig. 1. The composite output signal from the discriminator of an f.m. receiver is applied to pin 1 of the i.c. where it is amplified for distribution to other parts of the chip. The phase-locked-loop consists of a voltage controlled oscillator (v.c.o.), two divide-by-two stages and a phase comparator (phase-lock detector). An inductor and a capacitor connected to pins 15 and 16 give the v.c.o. a natural centre frequency of 76kHz. This 76kHz signal is divided by four in two cascaded divide-by-two stages to provide a 19kHz reference for the phase-lock detector. The phase-lock detector compares the locally generated 19kHz signal with the incoming 19kHz pilot tone and provides an output to alter the operating frequency of the v.c.o. if there is any difference. The bandwidth of this loop—which may be likened to a servo system—is determined by an *RC* network connected to pin 14.

The whole purpose of the loop is to regenerate the 38kHz sub-carrier which is suppressed at the transmitter before the signal is transmitted. The 38kHz sub-carrier is necessary to demodulate the composite stereo signal and the action of the loop ensures that the regenerated sub-carrier is very closely related in phase to the transmitted 19kHz pilot tone.

When the v.c.o. is running at exactly the right frequency the output from the phase-lock detector is zero so it is necessary to provide a second detector, to sense the presence of the pilot tone, in order that the chip can distinguish between a stereo and a mono signal—the pilot tone is not present on a mono signal.



(A) Composite signal (B) Stereo enable signal (C) Stereo gating signal (D) Difference signal

Fig. 1. Block diagram of the CA3090 integrated circuit which forms the major part of a phase-locked-loop stereo decoder.

This detector is called the pilot presence detector and it is driven by a second divide-by-two stage operating from the chip's 38kHz line. The resulting 19kHz signal is compared with the composite input signal and if a pilot tone is present the pilot presence detector trips a Schmitt trigger. The sensitivity of the pilot presence detector is set by a resistor connected between pins 7 and 8. With the value shown in Fig. 2, a 4mV input signal (pin 1) will be sufficient to operate the Schmitt trigger. If greater sensitivity is required the resistor can be replaced with a 4.7mH coil in series with 15nF capacitor across pins 7 and 8. The Schmitt trigger will then operate at 3.3mV (off at 2mV) and an improved overload characteristic is obtained as a by-product. An *RC* combination connected to pin 6 is a filter for the pilot presence detector.

When the Schmitt trigger operates it lights the stereo indicator lamp via an integral driver amplifier and informs the left/right channel detector that a stereo signal is being received and switches the whole chip to stereo operation.

The left/right channel detector uses the 38kHz sub-carrier (stereo gating signal), generated by the phase-locked-loop, and the composite input signal to produce a stereo difference signal which drives the matrixing circuits. The matrix extracts the left and right channel outputs from the composite input signal in the normal way and after amplification the left and right channel outputs appear at pins 9 and 10.

Practical notes

The complete circuit diagram is given in Fig. 2 and little need be said about it as the purposes of most of the components have already been described. The capacitors C_1 and C_2 provide the necessary de-emphasis and the two 10kΩ resistors are the collector loads of the 'open ended' channel amplifier output transistors.

The stereo indicator lamp can be a light-emitting diode as shown or a normal filament lamp which may be connected in place of the light emitting diode and 680Ω series resistor provided that the lamp does

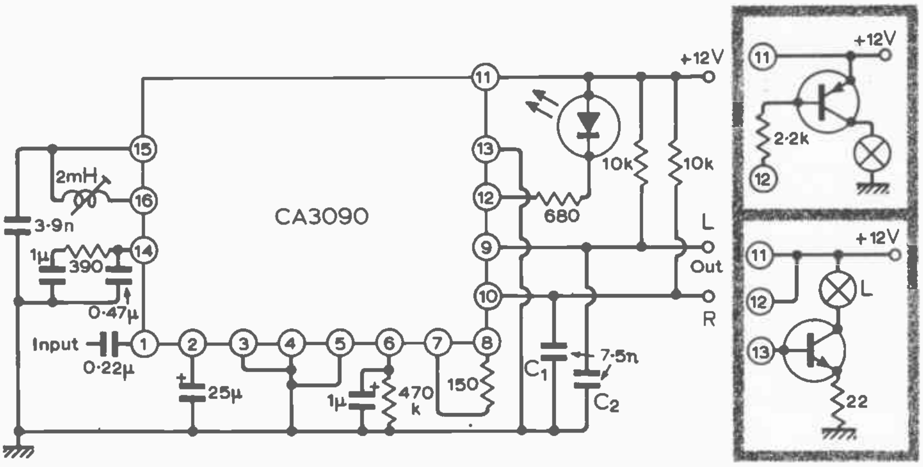


Fig. 2. Additional components required to complete the decoder. For operation in the UK (50µs de-emphasis) change the value of the 7.5nS capacitors to 5nS.

counting the turns, until the stereo indicator lamp goes out. Set the core at a point midway between the points where the lamp came on and went off.

Alter the potentiometer setting so as to reduce the input to the decoder and extinguish the stereo indicator lamp. Rock the core of the 2mH coil about its centre position to see if the indicator lamp lights. If not, slightly increase the potentiometer setting and rock the core again. The correct position for the coil's core is the one that lights the lamp with the minimum input signal.

R.C.A. manufacture two versions of the decoder i.c. One is in a staggered 16-pin dual-in-line package which is used in the illustrated printed circuit board and is called type CA3090Q, the second—type CA3090E— is electrically identical and is housed in a conventional 16-pin dual-in-line package. The i.c. is available from R.C.A. distributors, price £4.20.

not consume more than 14mA at 12V. If a higher current lamp is used an outboard driver transistor must be added. The inset shows circuits using either a p-n-p or an n-p-n transistor. The transistor type is not critical provided that it can handle the lamp current. For instance, a 40mA, 12V, lamp could be used if it were driven by a BC108 (use the n-p-n circuit in this case). However, the maximum lamp current—whatever the transistor used—should not exceed 100mA because the drive is limited to 14mA. Anyway who wants to use a searchlight to indicate that a stereo signal is being received!

The decoder can be built on the printed circuit board shown in Fig. 3 full size, or 'pin-board' construction can be employed. The 2mH coil can be obtained from Harrogate Radio Ltd., 2/3 Sykes Grove, Harrogate, Yorks., price 15p including postage, etc. Ask for type 87BN135BX2. The prototype used a coil of American origin. The type we have specified in fact contains two coils so for this application use coil pins 3 and 4 only. A slight alteration to the printed circuit board may be necessary. Alternatively use any 2mH coil which allows a ±25% adjustment.

When connecting the decoder to the discriminator output of a receiver care should be taken to ensure that the receiver's de-emphasis network is disconnected. The decoder will accept inputs between 40 and 400mV. If the discriminator of your receiver provides an output higher than this use a potentiometer of about 100k to reduce the signal. Make sure your receiver has enough bandwidth for stereo operation.

Two methods may be employed to set-up the decoder both of which are extremely simple. If you have access to a digital frequency meter connect it to pin 15 of the i.c. and adjust the core of the 2mH coil to give 76kHz. This adjustment is done when there is no input to pin 1.

The second method of adjustment does not require the use of any test equipment. Connect the decoder to a receiver via a 100kΩ potentiometer and tune in a stereo broadcast. Start with the core of the 2mH coil fully out and the potentiometer set to give maximum input to the decoder. Screw in the core of the 2mH coil until the stereo indicator lamp lights; continue turning the core in the same direction,

Typical Decoder Specification

Input impedance	50kΩ
Channel separation	40dB
Channel balance (mono)	0.3dB
Mono gain	6dB
Stereo/mono gain	0.3dB
Indicator lamp turn-on voltage*	4mV
Capture range (deviation from 76kHz centre frequency)	±10%
Distortion	
2nd harmonic	0.35%
3rd, 4th and 5th harmonic	0.1%
19kHz rejection	35dB
38kHz rejection	25dB
Input voltage range	40 to 400mV
Supply voltage	12V
Supply current (lamp off)	22mA
Operating temperature range	−40 to +85°C

*For improved pilot sensitivity and overload characteristics replace the 150Ω resistor between pins 7 and 8 with a coil of 4.7mH in series with a capacitor of 0.015µF.

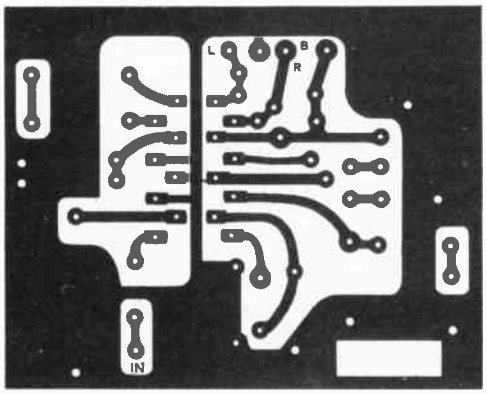


Fig. 3. Prototype printed circuit board layout shown actual size.

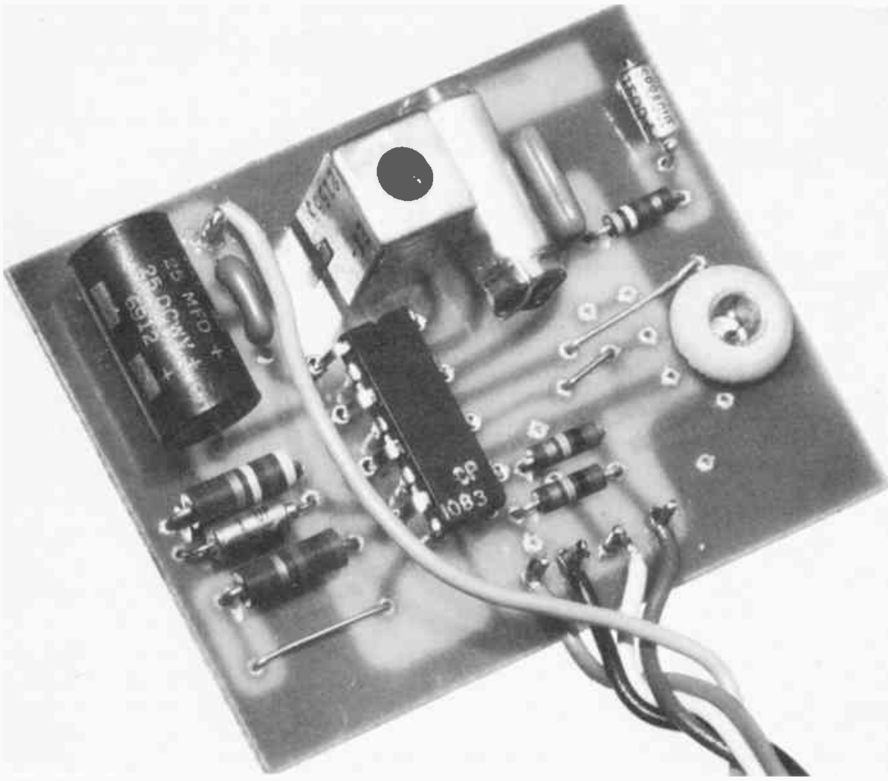


Fig. 4. Photograph of the prototype. Because this is a demonstration model built by R.C.A. some of the components shown in Fig. 2 are not included.

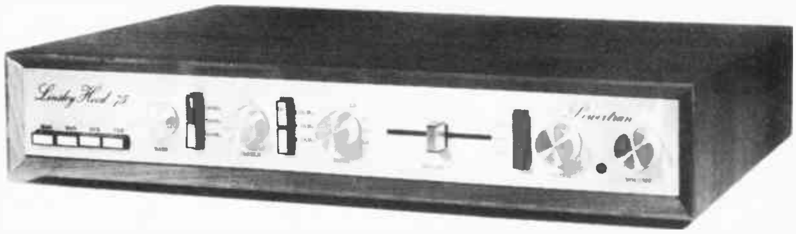
AMPLIFIER KITS OF *Distinction*

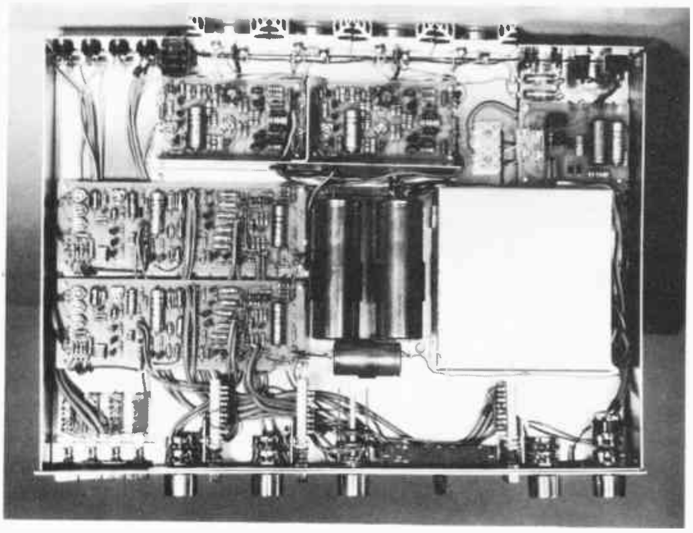
75 WATTS/CHANNEL

Hi-Fi News Linsley-Hood 75 W Amplifier Mk. III Version (modifications as per Hi-Fi News April 1974)

DESIGNER APPROVED KIT

In Hi-Fi News there was published by Mr. Linsley-Hood a series of four articles (November 1972 — February 1973) and a subsequent follow up article (April 1974) on a design for an amplifier of exceptional performance which has as its principal feature an ability to supply from a direct coupled fully protected output stage, power in excess of 75 watts whilst maintaining distortion at less than 0.01% even at very low power levels. The power amplifier is complemented by a pre-amplifier based on a discrete component operational amplifier referred to as the Linlac which is employed in the two most critical points of the system, namely the equalisation stage and tone control stage, positions where most conventional designs run out of gain at the extremes of the frequency spectrum. Unusual features of the design are the variable transition frequencies of the tone controls and the variable slope of the scratch filter. There is a choice of four inputs, two equalised and two linear, each having independently adjustable signal level. The attractive slimline unit pictured has been made practical by highly compact P.C.B.'s and a specially designed Toroidal transformer.





* Full circuit description in handbook *
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10	Toroidal transformer complete with magnetic screen/ housing primary: 0-117-234 V, secondaries: 33-0-33 V, 25-0-25 V.	9.15
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2N3906	0.20	BC126	0.15	BFY50	0.20	SN72748P	0.58		
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2—Practical low-impedance circuits

by B. J. C. Burrows, B.Sc.

This article gives full circuit details of an economy and a high-performance pre-amplifier which use a new design principle to provide optimum performance from stereo and mono ceramic cartridges.

Many ceramic cartridges are capable of a very high standard of performance—but this is seldom realized in practice. This is because conventional pre-amplifiers cannot cope satisfactorily with the wide range of electrical parameters encountered in different makes of ceramic cartridge.

The two factors that cause the problems in pre-amplifiers for piezo-electric cartridges are (i), self capacitance, and (ii), the degree of built-in mechanical equalization. In conventionally designed circuits using high-value load resistances ($1-2M\Omega$), the pickup self-capacitance has a profound effect on low-frequency performance and hence on the rumble performance. Fig. 1 shows curves of output voltage against frequency for two well known pickups when operated into a conventional pre-amplifier with $2M\Omega$ input impedance. These show that the overall frequency response is far from flat.

Typical pickups vary in capacitance from 200pF to greater than 1500pF, and with manufacturing tolerances plus the uncertain nature of the lead capacitance an overall variation of 180pF to >2000pF is possible. To obtain good l.f. performance with 180pF needs a loading resistance of $18M\Omega$ (not $1-M\Omega$ as commonly provided). If $18M\Omega$ were used with a pickup of 2000pF the bass turnover frequency would be 4.5Hz! This of course would result in very objectionable rumble and l.f.

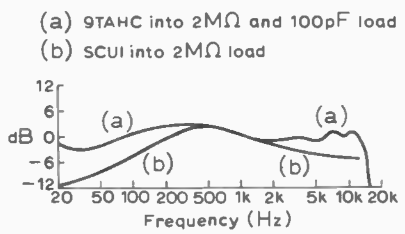


Fig. 1. Voltage/frequency curves of two well-known ceramic cartridges when used with a conventionally-designed pre-amp with $R_{in}=2M\Omega$, and a flat frequency response.

arm resonance† problems.

Conventional pre-amplifier designs do not allow for built-in mechanical equalization which varies from one pickup to another, and unfortunately the usual type of tone controls are not suitable for providing the necessary correction.

We can draw up a list of performance characteristics which an ideal pre-amplifier should possess:

(1) l.f. performance independent of cartridge capacitance;
(2) accurate rumble filtering independent of cartridge capacitance;
(3) means of correcting for variability in mechanical equalization (i.e. some form of 'tone balance' control).
(4) ability to cope with pickups of widely differing output voltages.

To these may be added: low noise, low distortion, good overload capability, built-in tone controls, etc.

Economy pre-amplifier

The complete circuit of the economy design is given in Fig. 2 for a positive h.t.

†See Appendix II.

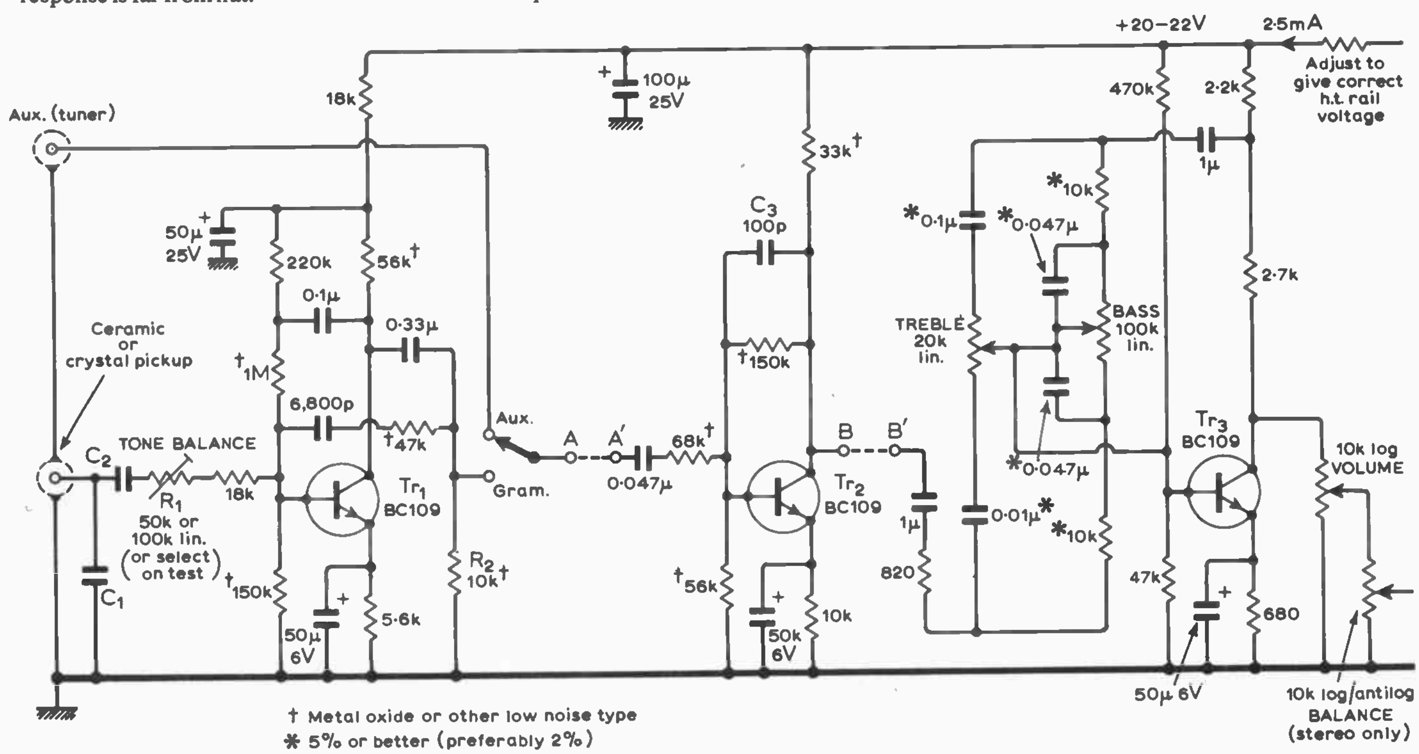


Table of values for C_1, C_2 & R_1 in economy circuit.

Cartridge type	C_1	C_2	R_1 (optimum value)	Comment
Decca Deram, Goldring CS91E	3.3nF	0.1µF	18–27kΩ, 56kΩ	low output
Goldring CS90, Sonotone 9TAHC	3.3nF	0.1µF	56kΩ, 22kΩ	medium output
Connoisseur SCU1	3.3nF	0.1µF	0	medium output
B.S.R. SC5M, Acos GP94/1, Garrard KS40A	10nF	6.8nF	22–56kΩ	high output

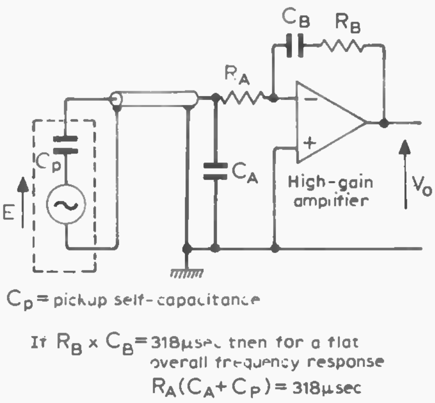


Fig. 3. First-stage design of equalization circuit.

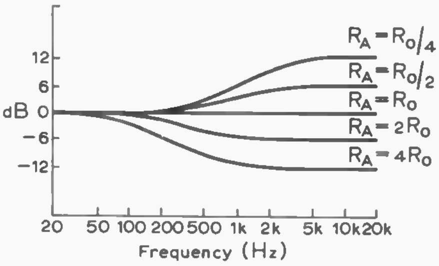
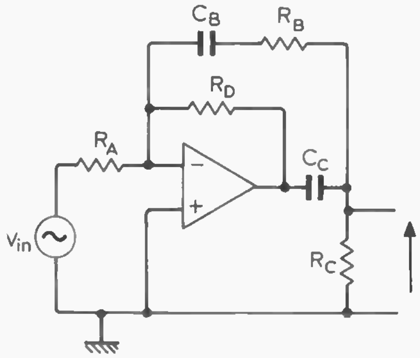


Fig. 4 Operation of tone-balance control, R_A in Fig. 3.



Design formulae for Q = 1

(1) Choose R_C

(2) Make R_B several times R_C

(3) $C_B = \dfrac{1}{2\pi f_1 R_B}$

(4) $C_C = \dfrac{1}{2\pi R_C}\left(\dfrac{1}{f_0} - \dfrac{1}{f_1}\right)$

(5) $R_D = R_B\left(\dfrac{(C_C R_C + C_B R_B)^2}{C_C R_C C_B R_B} - 1\right)$

Fig. 5. Baxandall bass lift-and-cut circuit.

rail system. A negative h.t. rail version is given in Appendix I. For normal use connect A to A' and B to B' and use full circuit. For ultra-economy operation with any of the pickups except the Deram or CS91E, the second stage may be omitted by connecting A direct to B' and omitting the intervening circuitry associated with Tr_2. Thus a very good, yet simple, gramophone amplifier may be built by using only Tr_1 and Tr_3 directly connected into an amplifier with 100mV sensitivity for full output.

Design principles of equalization stage

Last month the merits of the shunt feedback (or virtual earth) amplifier were mentioned as being very suitable for ceramic pickup equalization. Further, it was shown that loading the pickup with a low impedance had no effect on its internal e.m.f. In the present design, then, the effects of the variability in capacitance have been eliminated by swamping the pickup in every case with a shunting capacitor of 3.3nF or more. An input resistor of 75kΩ then gives an input time constant of 318µs (equivalent to 500Hz); to match this, the feedback circuit has a time constant of 318µs also (see Fig. 3); the complete circuit has a *flat* frequency response:

$$\frac{V_O}{E} = \text{constant} = \frac{R_B}{R_A} = \frac{C_P + C_A}{C_B}$$

If any one of the components suffixed A or B is made variable, a 'tone balance' type of control is achieved in a much simpler manner than circuits described previously[1]. R_A is the best one to vary and provides performance variation as in Fig. 4. The value of R_A to give an overall flat frequency response is termed R_0. In practice only values of R_A between R_0 and $R_0/4$ are needed to fully correct all ceramic pickups for their lack of complete mechanical equalization, e.g. the Sonotone 9TAHC pickup needs $R_A = R_0/1.8$ and the Connoisseur SCUI needs $R_A = R_0/4$.

With an infinite gain amplier in Fig. 3, overall gain is flat down to d.c. theoretically. This is no use in audio work because of rumble and the l.f. arm resonance. Some form of rumble filtering is essential and may be built into the equalization stage by using the circuit due to P. J. Baxandall[2]. The essence of this circuit is in Fig. 5, and its performance in Fig. 6.

Economy pre-amplifier specification

rated output	500mV r.m.s.
distortion (1KHz)	0.1% at maximum recorded level
noise	below audibility at normal listening level
hum	depends on layout and h.t. decoupling
overload capacity	>6dB above maximum recorded level
sensitivity	full output for pickup with 50mVcm/sec
sensitivity is reduced by	raising C_1 and lowering C_2 to keep C_1 $C_2/(C_1=C_1) \approx$ 4000pF
input impedance	not applicable (68kΩ for aux input connected as shown)
disc equalization	in conjunction with the better ceramic pickups can be adjusted to flat ±1.5dB 30Hz—10KHz. Low-frequency performance independent of pick-up capacitance.
rumble filter	18dB/oct, f_0=50Hz independent of pick-up capacitance
low-pass filter	fixed, C_3=100pF gives f_{-3dB}=12KHz. Scale C_3 up in proportion for low f_{3dB}
tone controls	h.f. about ±14dB; l.f. about ±14dB.
current consumption	≈2.5mA.

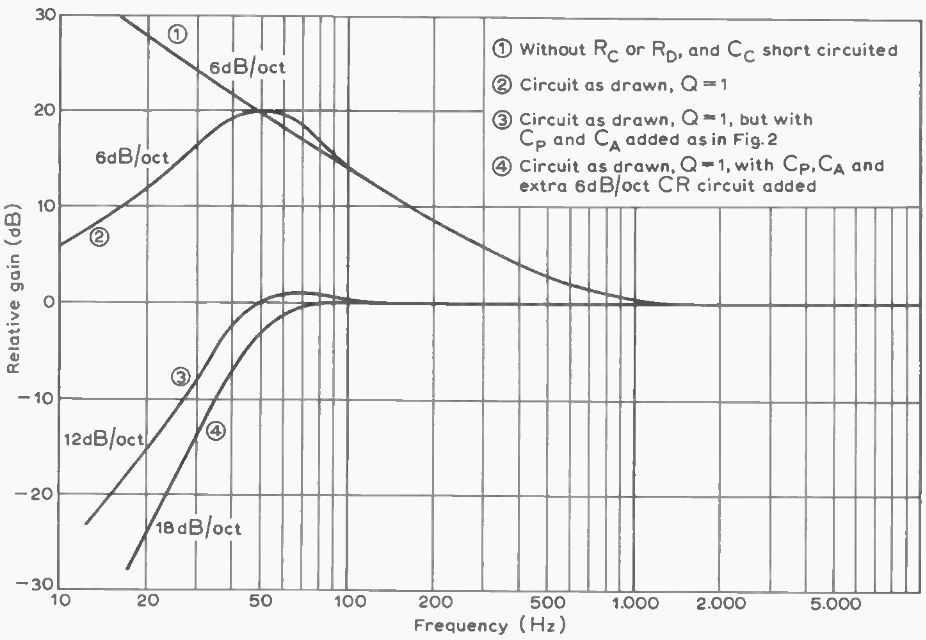


Fig. 6. Performance of circuit of Fig. 5 with f_0=50Hz and f_1=500Hz.

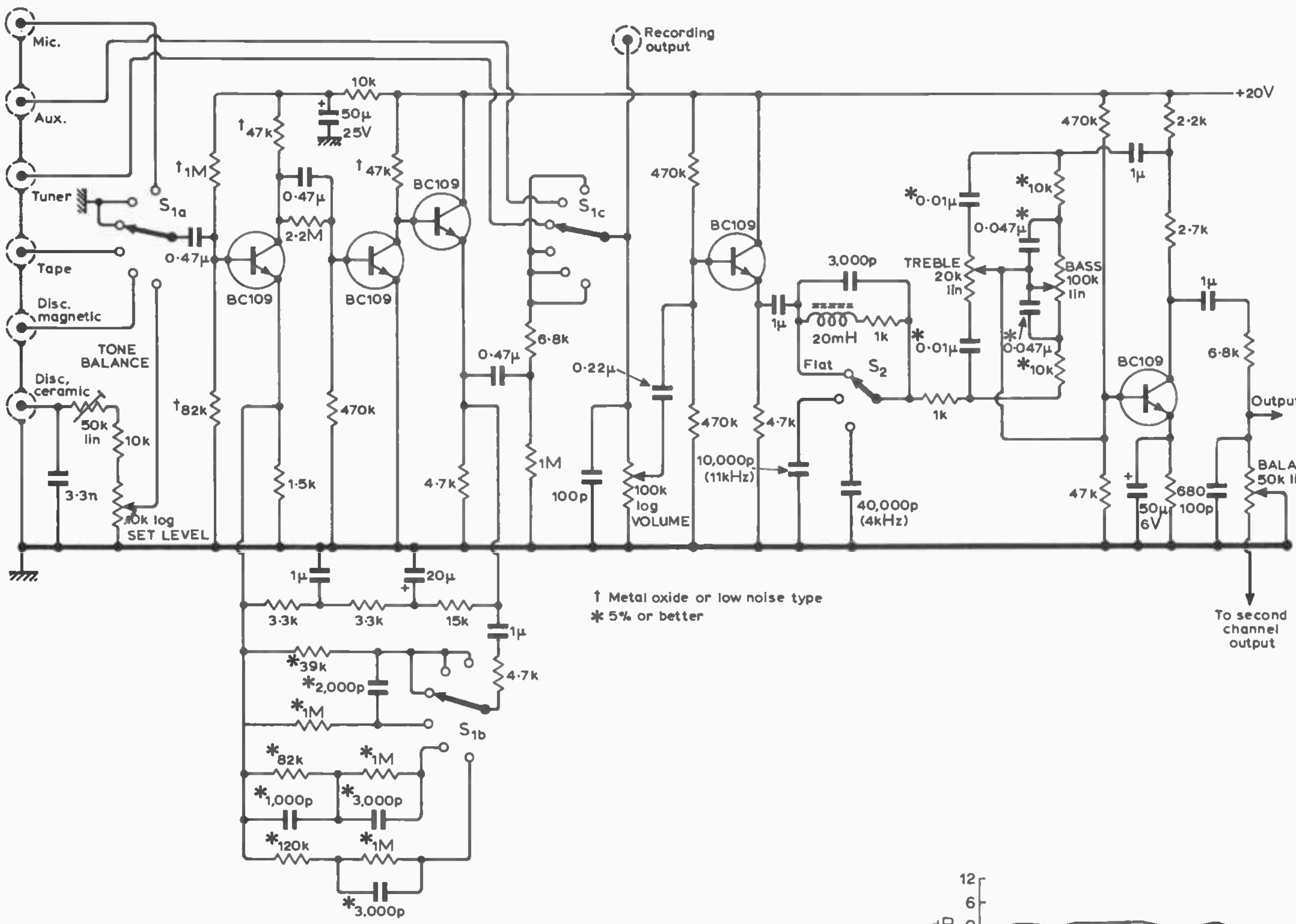


Fig. 7. Complete circuit of one channel of 'Bailey pre-amplifier'. No circuit changes are required for different ceramic pickup cartridges, only adjustment of 'tone balance' and 'set level'.

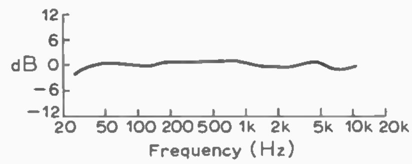


Fig. 8. Measured voltage/frequency curve for a 9TAHC operating into an 'economy design' circuit with $R_A=R_0/1.8$. The curve for the SCU1 would be just as flat, but with $R_4=R_0/4$.

If a further high-pass *RC* filter is added,

$$f_0=\frac{1}{2\pi R_C}$$

where a flat response to nearly 50 Hz is achieved with a rapid turnover to a slope of 18dB/octave to attenuate rumble. Finally, with R_A adjustable, the tone balance facility is still retained as with the basic circuit of Fig. 3. It is common to design rumble filters with cut-off frequencies much lower than 50Hz; but, to achieve adequate attenuation at 25Hz—a common frequency of the l.f. arm resonance—a high value of f_0 is required. The actual circuit of Fig. 2 achieves −28dB at 15Hz and −15dB at 25Hz. In practice this is very satisfactory.

The economy-design pre-amplifier closely matches the theoretical performance of Figs. 4 and 6 and provides excellent bass, good balance and excellent freedom from rumble. As shown in the table relating to the main circuit, the only circuit changes needed to accommodate different pickups are for curbing those with a very high output voltage with a capacitive divider. In connection with the table of values given for the input capacitors it is very important to stress that the values given must be used as specified and that the manufacturers' recommendations regarding load impedance and equalization must be totally ignored. This circuit has been specifically designed to take care of all the loading, matching and equalization factors and *no further components are needed.*

High-performance pre-amplifier specification

rated output	500mV r.m.s.
harmonic distortion	0.02% at rated output
noise	−60dB all inputs −80dB for tuner and aux inputs
hum	negligible with good layout
overload capacity	23dB over whole audio range, infinite for tuner and aux
sensitivity	tuner 250mV aux 250mV disc magnetic 3mV disc ceramic 20mV tape 4mV mic 10mV
input impedance	tuner, aux 60-100KΩ disc magnetic 47KΩ disc ceramic frequency dependent tape, mic 47KΩ
disc equalization	magnetic—RIAA to within ± 1dB ceramic—can be adjusted to give flat response ± 1½dB l.f. response independent of cartridge capacitance
tape equalization	7½i.p.s. with R_{FB} = 39KΩ 15i.p.s. with R_{FB} = 18KΩ 3¾i.p.s. with R_{FB} = 82KΩ
rumble filter	modified design giving higher cut off frequency; response at 25Hz is −15dB
low-pass filter	switched, flat or cut off at any frequency from 4 to 11KHz (see Ref. 7)
tone controls	Baxandall type treble ±16dB at extreme bass ±20dB at extreme
current consumption	7mA

The economy circuit as described fulfils all the design criteria enumerated earlier except for the slight inconvenience of changing two capacitors if pickups of widely differing output voltages are exchanged. The noise performance is very good with all the cartridges listed apart from two (the CS91E and Deram) with which it is satisfactory for everything but the most exacting requirements.

High performance pre-amplifier

This is based on the Bailey[3] design of 1966 but with all the subsequent modifications to improve the filter[4] and tone control[5] circuits, plus the addition of a complete ceramic-pickup equalizing circuit achieving the same performance with ceramic cartridges as the economy pre-amplifier. The complete circuit is given in Fig. 7, which also incorporates one further modification to raise the cut-off frequency of the rumble filter in accordance with the design philosophy discussed in Appendix II. Equalization for magnetic pickups has been retained and is selected by the input selector switch. The 'set level' control needs a mention. To avoid overloading the input stage, adjust the set level control with any particular

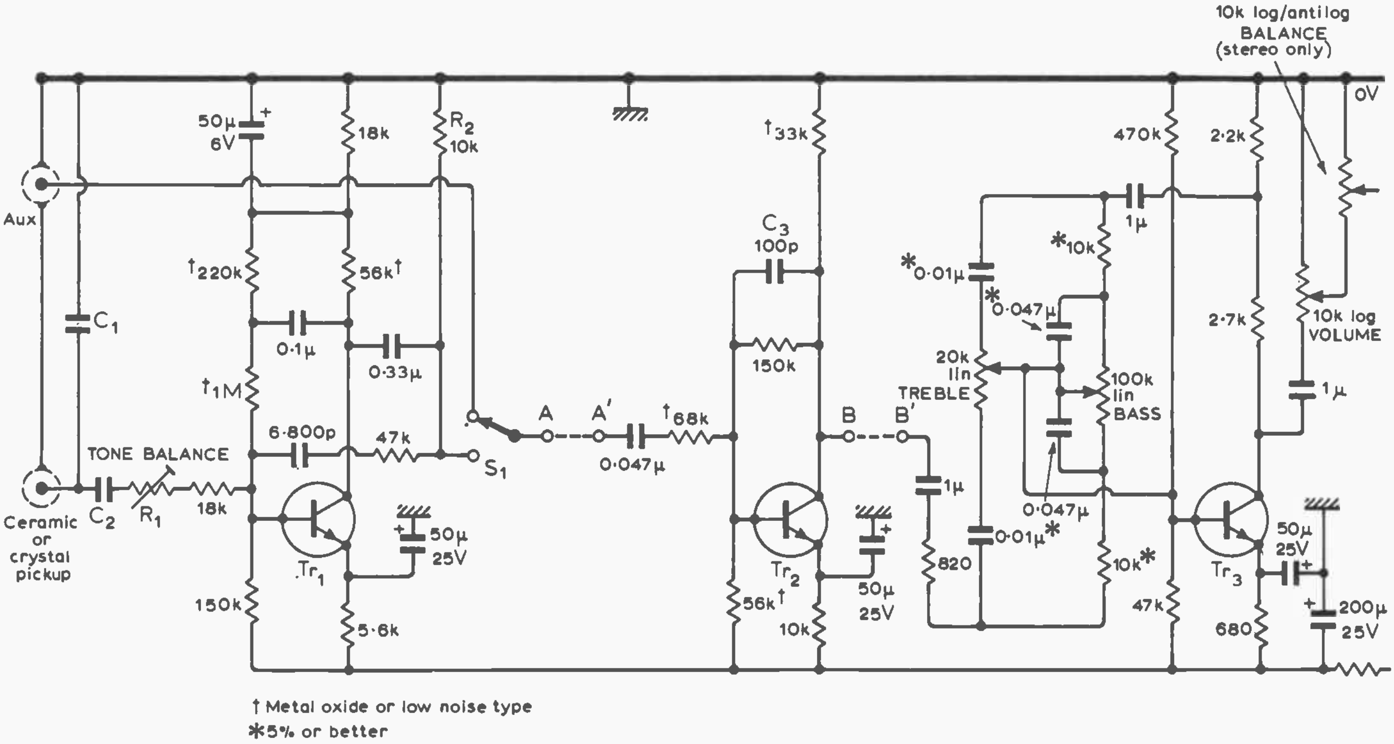


Fig. 9. Economy circuit arranged for negative h.t. rail. For values of C_1, C_2, and R_1, see table earlier.

cartridge to give comfortable listening level with the main volume control at about half of its maximum rotation. This control need be only a preset with screwdriver slot adjustable from the back of the pre-amplifier. The tone balance could be the same, or it could be brought out as a front panel control, or as a skeleton pot mounted internally or even a 'select-on-test' fixed resistor.

On paper, the specification of the high performance pre-amplifier looks most impressive, but subjectively the economy version is very good indeed, and both represent a considerable improvement on conventional designs in that reproducible low-frequency performance, effective rumble filtering independent of pickup capacitance, and a simple means of correcting for partial mechanical equalization have been incorporated. Fig. 8 in conjunction with Fig. 1 gives a comparison of the performance of the Sonotone 9TAHC and Connoisseur SCU1 using conventional loading (2MΩ plus flat amplifier), compared with the measured results on the author's 9TAHC using the economy circuit.

The calculated performance of the Connoisseur SCU1 with $R_1 = R_0/4$ is a straight line coincident with the 0dB line on Fig. 8, although in practice there would be a variation of up to ± 1dB about the 0dB line.

Modifications to provide a similar standard of performance with the Dinsdale Mark I and Mark II pre-amplifier circuits were incorporated in a previous article[6].

Appendix I

Alteration of economy circuit for negative h.t. rail operation, e.g. from a germanium-transistor amplifier like the Dinsdale Mark I or II, is basically to return all electrolytic capacitors to the positive potential rail, viz. the earth line (see Fig. 9). There are no modifications to circuit values apart from the voltage rating of the electrolytics.

Appendix II

Arm resonance (l.f.) is the tendency toward damped oscillation at a low frequency and is exhibited by most pickup arms. It has the effect of greatly increasing the cartridge output voltage at or near the resonant frequency. The frequency, f_{lf}, is normally in the range 10-25Hz, so its effect is to greatly increase rumble. The frequency of the oscillation is:

$$f_{lf} = \frac{1}{2\pi\sqrt{MC}}\text{Hz}$$

M is the mass of cartridge plus effective mass of arm measured at cartridge.
C is the compliance of stylus cantilever suspension. With M in grams, C is in cm/dyne.
With modern high compliance cartridges it is desirable to keep M very low—hence lightweight headshells—to make f_{lf} as high as possible. Generally speaking the lower the frequency of resonance the higher the Q, and vice versa. But a higher resonant frequency is more trouble electrically. A low-frequency high-Q resonance causes mechanical difficulties—the pickup tends to leave the record surface when excited. A resonance at 25Hz is acceptable mechanically if the Q is low enough and its electrical effects can be removed with a steep slope filter. Below this resonant frequency the cartridge output voltage falls off very sharply indeed (24dB/octave) thus providing the required severe attenuation of sub-audio frequencies.

With regard to pre-amplifier design, the point to note is that the highest amplitude rumble components will be at, or near, the l.f. arm resonance. A filter in the pre-amplifier should ideally provide 12dB or more of attenuation at 25Hz, yet not interfere with l.f. audio response. A cut off frequency of 50Hz with slope approaching 18dB/octave is a very good compromise since it causes very little error in the R.I.A.A. equalization, yet gives −15dB at 25Hz and −25dB at 15Hz.

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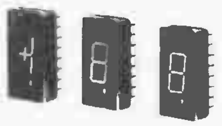

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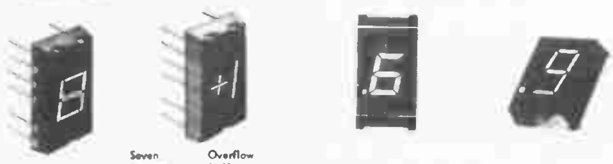

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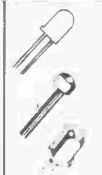

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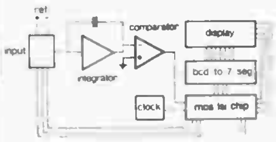

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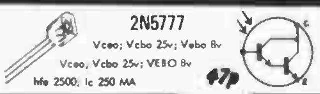

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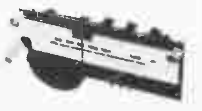

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BA110	45p	OA79	10p
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BB105B	42p	Z5142	33p
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AC128	13p	BC108	11p	BC214L	13p	BF184	28p
AC176	15p	BC109	12p	BC268	15p	BF185	28p
AC187	22p	BC117	22p	BC407	16p	BF194	15p
AC187K	20p	BC147	10p	BCY70	17p	BF195	17p
AC188	22p	BC148	10p	BCY71	22p	BF196	16p
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	£1.30	£1.65	£1.65	£2.42
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	£1.30	£1.65	£1.65	£2.42

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200v	MCR120	51p	106B	53p	2N4442	£1.06
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200	IN4003	9p	IN5402	17½p	BYZ13	58p	BYX61 -200	£3.76
400	IN4004	9p	IN5404	22p	BYZ12	66p	BYX61 -400	£4.29
600	IN4005	11p	IN5406	26p	BYZ11	73p	BYX62 -600	£4.29
800	IN4006	13½p	IN5407	28p	BYZ10	85p		
1000	IN4007	16½p	IN5408	31p				

TRIACS

4A		
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50v	2N6069	52p
100v	2N6070	58p
200v	2N6071	63p
400v	2N6073	68p
600v	2N6075	£1.49

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Output stage using complementary transistors

*by Arthur R. Bailey**, *M.Sc., Ph.D., M.I.E.E.*

It is only recently that matched complementary output transistors, capable of high dissipation, have been available at a reasonable price. In the past this has had the effect of concentrating high power amplifier design into two main streams. The first uses a driver transformer with a pair of identical output transistors in a series connection. The use of a driver transformer is undesirable mainly on account of the cost, as the bandwidth of a well designed component may well extend from the sub-sonic region up to several megahertz. Nevertheless a circuit that does not require the use of such a component will obviously be at an advantage.

The alternative circuit that has been used by many designers is the quasi-complementary output stage. In this design identical output transistors are used and a complementary pair of driver transistors is arranged so as to give phase-inversion to the bases of the two output transistors. These two circuits are shown in Figs. 1(a) and 1(b) respectively. A correctly designed fully complementary output stage (Fig. 1(c) shows the basic arrangement) is capable of better performance than either of these common circuits and the reasons for this will be examined.

Compared with the quasi-complementary amplifier, the transformer-driven amplifier has the great advantage that the input impedances to the two sides of the output circuit are identical. This means that if a suitable quiescent current is used in the output transistors, cross-over distortion will be almost completely absent.

The quasi-complementary amplifier, however, gives greater overall distortion even if identical output transistors are used. This increase is due to the different input impedances of the two halves of the output stage in the quasi-complementary circuit. In the upper half of Fig. 1(b) the input impedance is due to two emitter-base junctions in series, whereas in the lower half the signal feeds into only one transistor. The effect of this is an extremely marked asymmetry between the input impedances of the upper and lower halves of the output stage. Unfortunately the two input impedances cannot be equalized by the use of a series resistor as the curvature of the two stages is completely different. This dissimilarity of curvature can be seen in Figs. 2 and 3, these being the transfer characteristics of the upper and lower halves of an output stage using matched transistors.

*University of Bradford

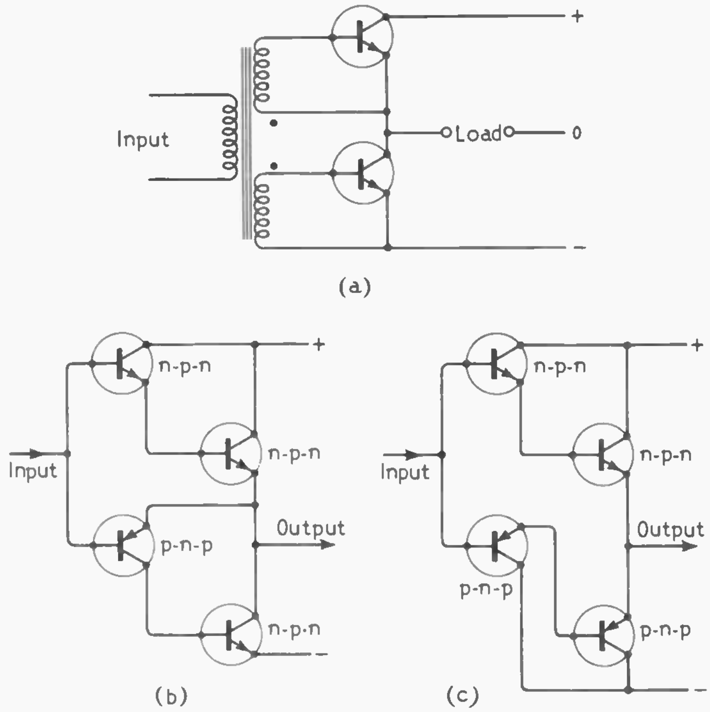


Fig. 1. Direct-coupled output stages: (a) with driver transformer; (b) quasi-complementary; (c) fully complementary.

The dissimilarity in input impedance is most marked at low values of collector current. Hence in the case of a class B output stage there is an abrupt change in slope at the cross-over point, giving rise to the well known phenomenon of cross-over distortion. This distortion may not be particularly serious when measured on an r.m.s. basis, but as it unfortunately occurs mainly within a small part of the overall output swing, the peak value of the distortion can be surprisingly high. Also the distortion does not normally decrease appreciably as the output swing is reduced, since the effect is occurring at small signal levels. The overall effect is quite serious, therefore, and the ear seems to be very sensitive to such types of distortion.

This then is perhaps the reason why two amplifiers may sound quite different even though their 'paper' performance may be identical on the basis of normal amplifier measurements. Very few valve amplifiers suffer from cross-over distortion, and this may be the reason why the best valve amplifiers are difficult to evaluate on subjective tests. Certainly there are much greater subjective differences between the performances of current transistor amplifiers.

If cross-over distortion is present it would appear that the common 0.1 per cent harmonic distortion rule for an acceptable limit at peak output is no longer valid, and at least one manufacturer is working on the basis of far lower distortions being necessary.

There appear to be two ways of tackling this problem. The first is to use a larger value of overall feedback so as to reduce the effect to inaudible proportions. The main drawback with this

Specification

Sensitivity	1.0 volt for 30 watts into 8-ohm load 0.8 volt for 20 watts into 16-ohm load
Rise time	approximately 0.7 microsecond
Distortion	below 0.1% over the whole of the audio-frequency range at rated power outputs
Load stability	unconditional
Abnormal load protection	provided adequate heat sinks are used the amplifier will not be damaged by operation into incorrect loads
Noise	better than 80 dB down on full power output
Hum	depends on layout if stray hum fields exist. Negligible hum in output if normally smoothed supplies are used.
Distortion generated	predominantly third harmonic, cross-over distortion being absent.

method is that high values of overall feedback make the amplifier closer to instability, and it may be difficult, if not impossible, to achieve a reasonable stability margin. Stability may then be obtained by decreasing the cut-off frequency of a stabilizing step-network, but this has the effect of decreasing the available power at high frequencies as well as degrading the distortion characteristics at high frequencies.

Complementary Symmetry Output Stage

In view of these considerations the author decided that the best line of approach was to use a fully symmetrical output based on complementary transistors. With such a symmetrical system, there is no difference between the input impedances in the upper and lower halves of the circuit. From the basic circuit in Fig. 1(c) it will be seen that both halves of the circuit have the same input impedance characteristics because of their identical configurations. By a suitable choice of standing quiescent current, cross-over distortion can be reduced to levels where it is extremely difficult to detect. This absence of cross-over distortion means that perfectly satisfactory results will be obtained if the overall distortion factor of the amplifier is similar to that commonly found in valve amplifiers, i.e. about the 0.1 per cent mark. In fact lower distortions than this are possible while maintaining both unconditional load stability and good high-frequency performance.

During the development of this amplifier it was discovered that the overall performance was not as good as might have been expected from the output stage characteristics. This distortion increase was traced to the common-emitter amplifier stage that drives the output stages. This is transistor Tr_3 in the complete amplifier circuit shown in Fig. 4. The effect was found to be caused by 'Early effect'[1], the high collector voltage swing modulating the gain of the stage. In fact the overall distortion was approximately three times that which would have been expected. As this effect depends entirely on the design of the transistor in use, it was necessary to select a suitable transistor type for this position in the amplifier. This source of distortion seems to have been largely overlooked in the past, but it is obviously a possible source of extremely bad distortion. In addition, the high-frequency distortion was found to increase more rapidly than was expected and this was traced to the modulation of the collector-base capacitance of this transistor. The high collector voltage swing was causing non-linear capacitive feedback, and this in turn was increasing the high-frequency distortion. Again the only cure is by transistor selection. The type used appears to be the best currently obtainable, and the distortion introduced by these effects is below that of the output stage proper, over the whole of the audio-frequency range.

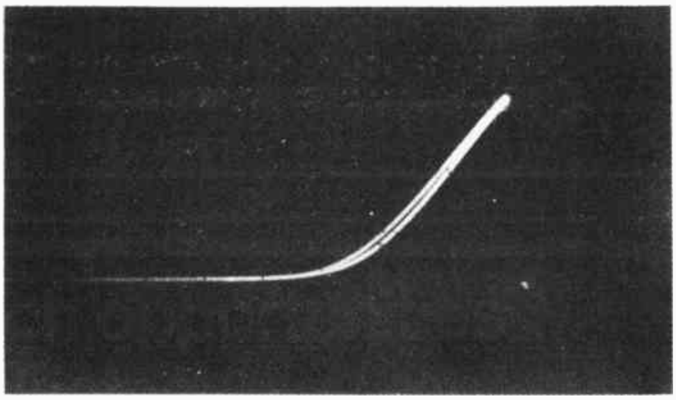


Fig. 2. Transfer characteristic of upper half of Fig. 1(b).

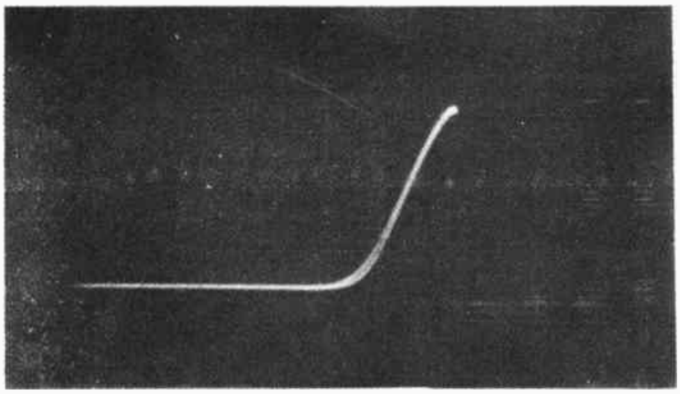
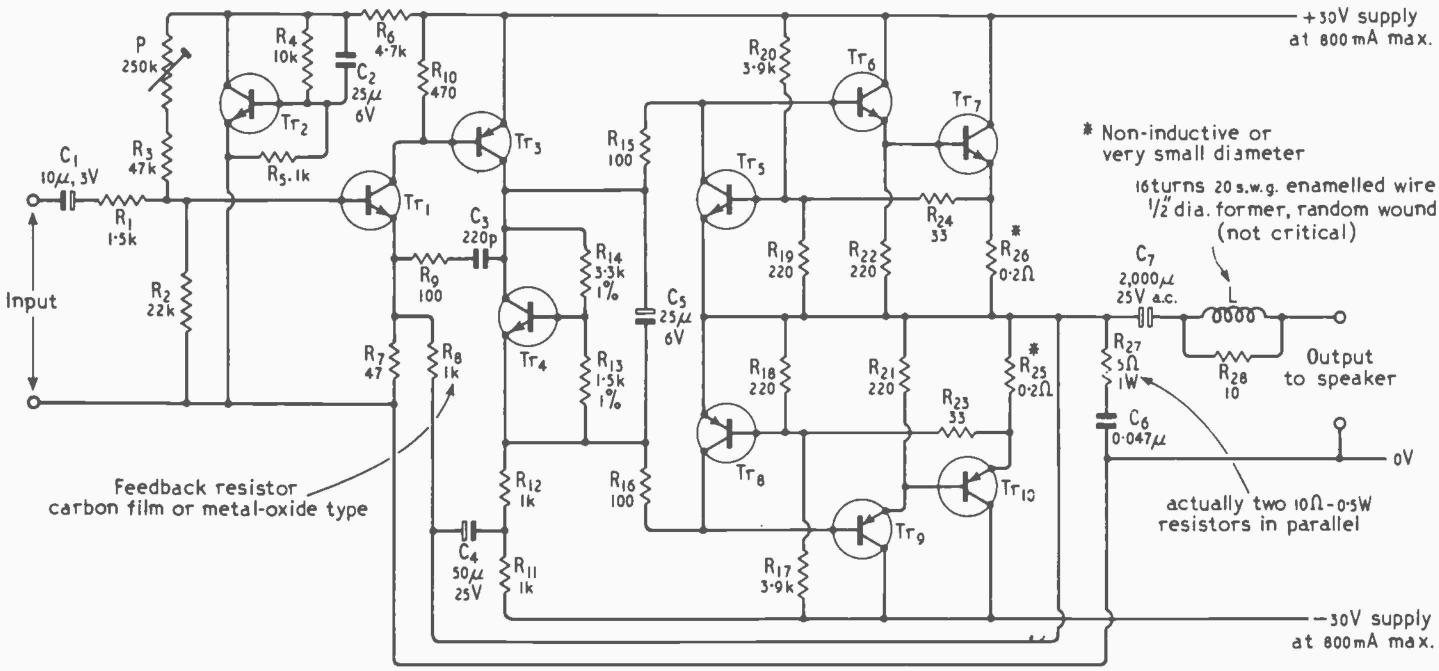


Fig. 3. Transfer characteristic of lower half of Fig. 1(b).

For low distortion at high frequencies, it is essential that the transistors should have as high a cut-off frequency as possible. Planar transistors are used in all but the output stage to give this bandwidth. The output transistors used have a cut-off frequency of several megahertz and this enables low distortions to be obtained at 20 kKz at full power output.

The design of the remainder of the amplifier circuit is fairly straightforward. The input stage is a common-emitter amplifier, but the current and voltage swings associated with it are very small, so there is little difficulty in the operation of this stage. To correct for the emitter-base voltage change of this input stage with temperature, a transistor is used to regulate the base supply

Fig. 4. Circuit of complete power amplifier. The transistors used are: Tr_1—40361 (R.C.A.); Tr_2—BC109 (Mullard); Tr_3—40362 (R.C.A).; Tr_4—BC107 (Mullard); Tr_5—BC125 (Fairchild); Tr_6—40361 (R.C.A.); Tr_7—MJ481 (Motorola); Tr_8—BC126 (Fairchild); Tr_9—40362 (R.C.A.); Tr_{10}—MJ491 (Motorola). Note that C_7 is a reversible electrolytic and could be made up of two 4000-μF polarized electrolytics connected 'back-to-back'.



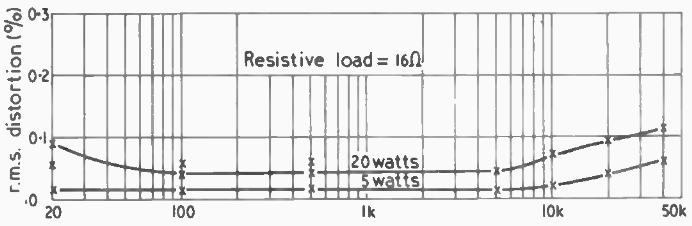


Fig. 5. Distortion characteristics of amplifier with 16-ohm load.

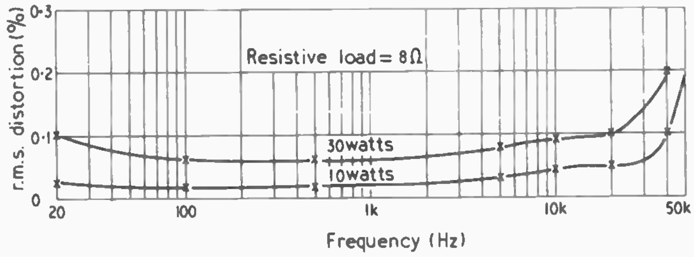


Fig. 6. Distortion characteristics of amplifier with 8-ohm load.

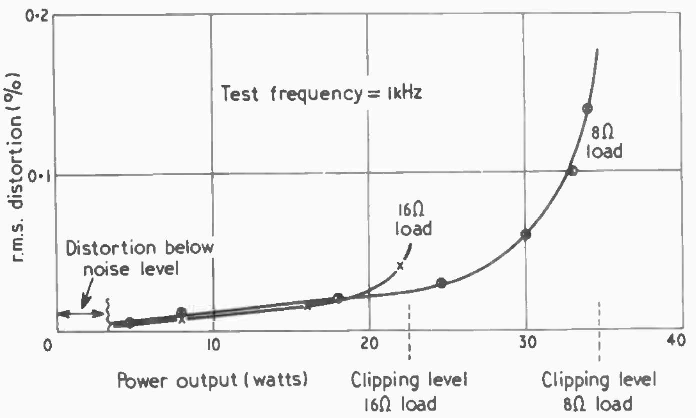


Fig. 7. Variation of distortion with output power level.

current. This transistor, Tr_2, operates as a rather crude temperature-sensitive zener diode and also as a hum filter. The net effect is to stabilize the d.c. base current of the input transistor, the supply voltage to the base of this transistor decreasing with increased temperature. This stabilization of the d.c. operating conditions enables the amplifier to deliver full output over a wide temperature range.

The bias for the driver and output transistors is produced by means of a transistor, Tr_4, rather than a string of diodes as is commonly used. This is mounted in the heat sink of one of the output transistors, being as close to the output transistor as possible. This method of compensation works extremely well, and the transistor type is not critical provided a silicon one is used. The standing current in the output stage can easily be adjusted to its correct value (which is not critical) by slightly adjusting the ratio of the two resistors in the base circuit of the transistor.

For full power output from the amplifier the d.c. potential existing at the output of the amplifier proper should be as low as possible. This can be adjusted by the potentiometer in the base of Tr_1. If this is not done the amplifier will not be able to swing equally in the two output polarities.

The quoted figures for the amplifier were obtained using regulated supplies. Unless the amplifier is to be called on to deliver large sustained outputs, this is not really necessary. On the other hand, reduced mains voltage will severely restrict the power output of an amplifier with unregulated supplies. Commercially, a thyristor regulated supply is being utilized, and this has the two advantages of small heat dissipation and saving in components.

Constructional Points

The overall bandwidth of the amplifier is extremely wide and the stabilizing step-network necessary only becomes operative in the ultrasonic region. Equally the inductor in series with the output lead, which improves the stability with capacitive loads, need have only a very small inductance. This wide bandwidth gives exceptional high-frequency performance as can be seen from the distortion figures in Figs. 5, 6 and 7. Unfortunately, however, wideband amplifiers are very susceptible to layout, particularly common coupling leads. Provided lead lengths are kept *very* short there should be no difficulty, but the author experienced tremendous variations in high-frequency stability when 'rats-nest' construction was used. For this reason the safest course is to use a printed circuit, so that the strays can be kept to a minimum. The design of a suitable board along with its component layout is shown in Figs. 8 and 9. The performance details given were measured using this particular layout. The leads to the output transistors should be as short as possible, preferably no longer than 3 to 4 inches. The size of the heat sinks for the output transistors is a matter of personal choice, the author having used sinks of finned aluminium about 4 in. by 4 in. square. This size is not really necessary for high-fidelity use, and sinks of half this size would be adequate provided that extended periods of testing were not undertaken.

The overall performance of the amplifier is very good, considerably better in fact (on paper) than the best valve amplifiers. Unfortunately, listening tests have shown that the performance of the amplifier is only slightly, if any, better than the best valve amplifiers. Extensive listening tests indicate only a very slight improvement in audible results, the subjective effects being almost identical. It would therefore appear that any further improvement will be of no real benefit for high-fidelity applications, the main need for work here definitely being in the field of loudspeakers, discs, etc.

Owing to the absence of cross-over distortion, the distortion at low levels is very difficult to measure and the curves appear in Fig. 7. The wide bandwidth can be seen from the curves in Figs. 5 and 6, where it will be observed that the amplifier will deliver full power output from 20 Hz to 20 kHz with less than 0.1 per cent of distortion. Indeed it is possible to obtain about 15 watts of power at 200 kHz. The square-wave tests are far better than with any known valve amplifier. Even with pure capacitive loads there is no tendency whatever towards instability. The waveforms are shown in Figs. 10, 11, 12, and 13.

The protection circuits of the amplifier operate very satisfactorily, short-circuits and 50 microfarad capacitors giving no distress to the amplifier whatever. One word of caution is

Fig. 8. Layout of suitable printed-circuit board, actual size. (Courtesy Radford Audio Ltd.)

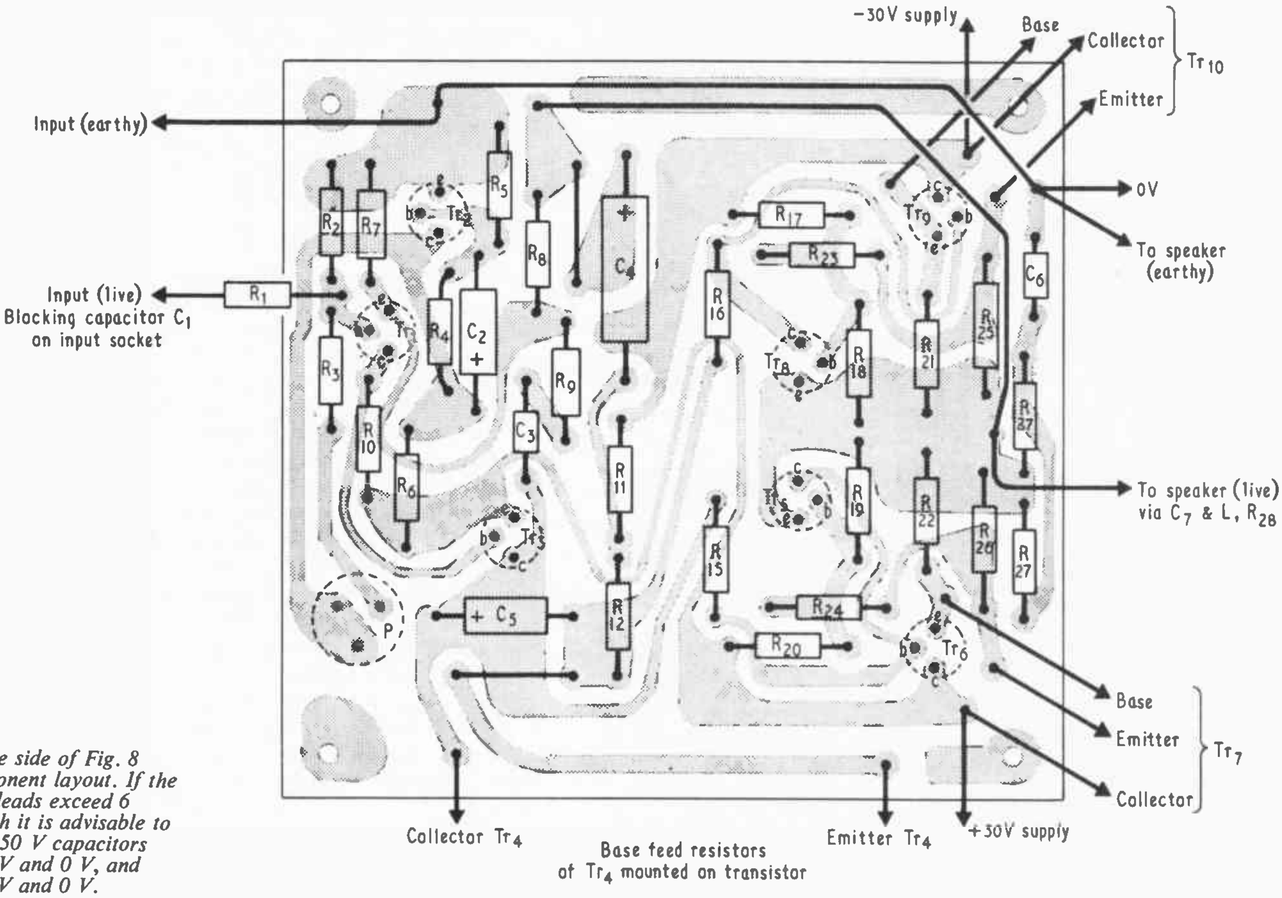


Fig. 9. Reverse side of Fig. 8 showing component layout. If the power supply leads exceed 6 inches in length it is advisable to connect 8 μF, 50 V capacitors between +30 V and 0 V, and between −30 V and 0 V.

necessary, however; extended tests on low impedance reactive loads and short-circuits can cause high junction temperatures in the output transistors because of the finite heat-sink size. Unless one uses very large heat sinks, it is therefore undesirable to run the amplifier at full drive for extended periods when applying such abnormal load conditions. If 16-ohm load operation only is to be used, then the emitter resistors in the output stage can be increased up to 0.4 ohm, with a corresponding halving of the transistor dissipation under abnormal load conditions.

The specification is shown on page 61. The overall sensitivity may be either doubled or halved by doubling or halving the value of the 1000-ohm feedback resistor. This has the effect of increasing the sensitivity at the expense of distortion if the increased amplification is felt to be necessary. With the increased feedback the overall distortion is halved, and even with this value of overall feedback the amplifier is still unconditionally load stable.

When the amplifier is operated in the reduced feedback condition for 500-millivolt sensitivity, the author cannot hear any difference in performance as compared with the halved distortion characteristic obtained with the 2-volt sensitivity. It appears therefore that no further improvement in amplifier performance will be detectable until other limiting factors are greatly improved. In fact the author has a sneaking suspicion that this may be the end of the road so far as amplifier design for sound reproduction is concerned, further improvements being limited to power and cost.

Reference

1. 'The Transistor' by E. Wolfendale. Heywood & Co., London (1963), p. 24.

Fig. 10. Square-wave response, 1 kHz and 8-ohm load.

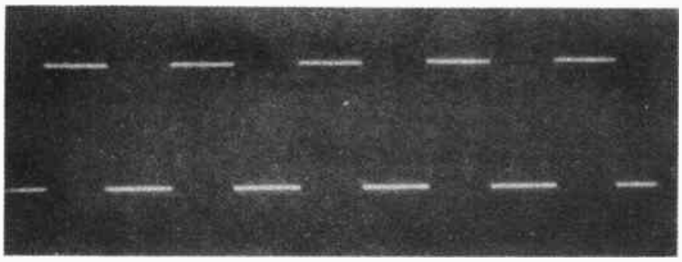


Fig. 12. Square-wave response, 10 kHz and 0.1-μF load.

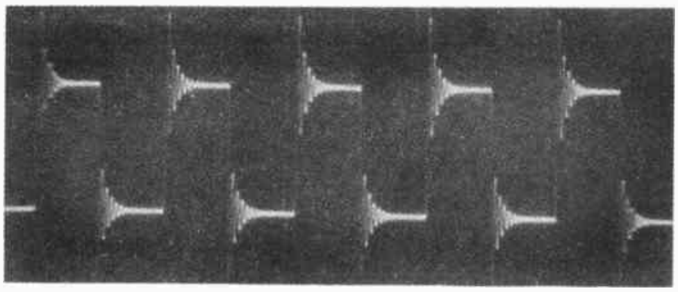


Fig. 11. Square-wave response, 50 kHz and 8-ohm load.

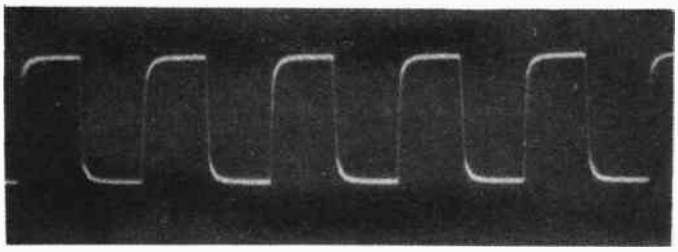
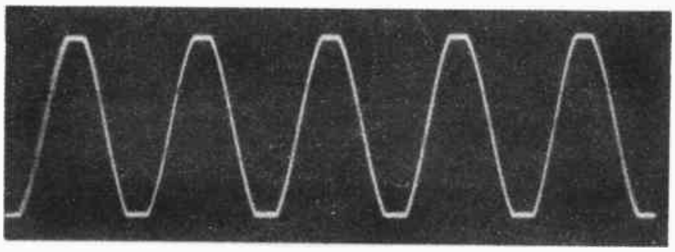


Fig. 13. Overdrive with sine-wave input, showing clean limiting 1 kHz and 8-ohm load.)



30-watt Amplifier Modification

Operation from a single supply rail

by Arthur R. Bailey *M.Sc. (Eng.), Ph.D., M.I.E.E.*

The 30-watt amplifier circuit published in the May 1968 edition of *Wireless World* utilized two power supplies and a non-polarized output capacitor. By suitable modifications to the circuit it is possible to use a single 60-volt supply rail and use a more standard polarized output capacitor. The overall cost then is appreciably reduced as compared with the original circuit. Also if it is contemplated to use a stabilized power supply, only one stabilizer has to be built.

The modifications are as follows:

1. The 0 and −30-volt amplifier connections are strapped together. This common connection goes to the negative side of the 60-V rail, and the previous +30-V lead goes to the positive side of the 60-V supply.

2. The output capacitor is now a 2000μF 50V d.c. electrolytic with its positive side connected to the amplifier output.

3. The feedback resistor R_8 is replaced by a 1.5-kΩ 1-W type, and its associated resistor R_7 is increased to 68 ohms.

4. To maintain d.c. balance in the amplifier, R_6 is increased to 10 kΩ and R_4 to 22 kΩ.

5. Owing to the changes produced in 4 above, C_2 is now 10μF and 12 volts d.c. working.

The d.c. conditions are now adjusted by the potentiometer P so that the amplifier side of the output capacitor C_7 is exactly half-way between the two supply lines in potential. The amplifier performance is not affected appreciably by these modifications—if anything the overall performance is slightly better.

The design of a suitable power supply is shown in Fig. 2. The thermistor is included to prevent the charging current in the output capacitor C_7 becoming dangerously large at switch-on and thus endangering loudspeakers. If a stabilized supply is used, then it can be 'slugged' so as to give a slow rise in output voltage at switch-on; this making the thermistor unnecessary.

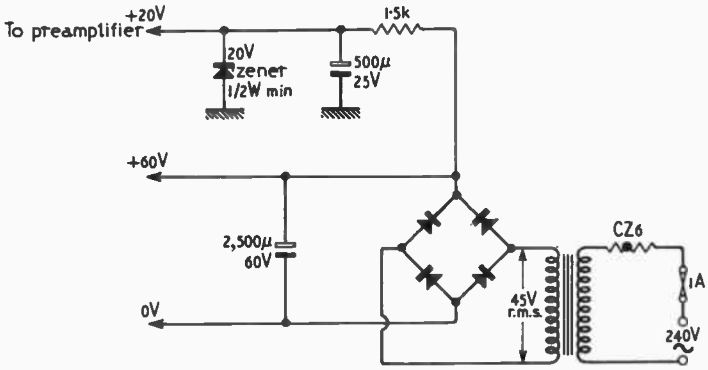
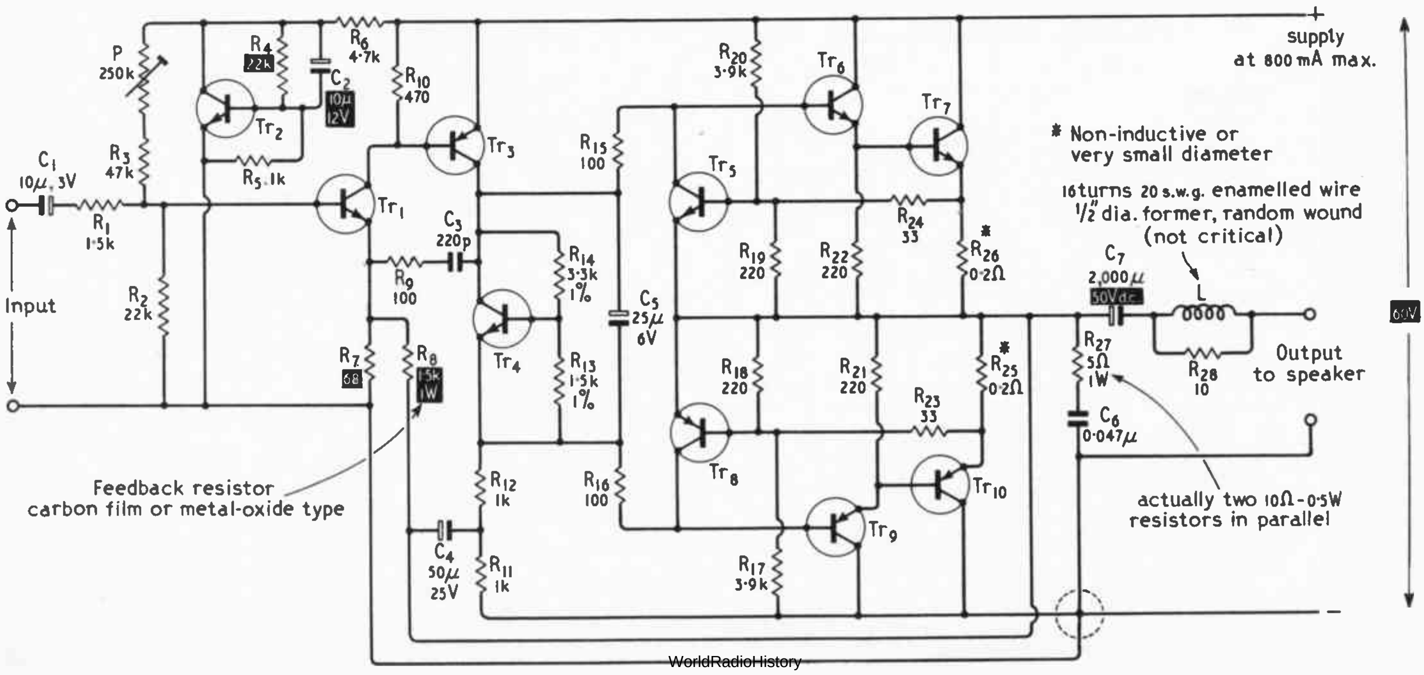


Fig. 2. A suitable power supply for the modified circuit.

Fig. 1. Modified circuit of complete power amplifier. The transistors used are: Tr_1—40361 (R.C.A.); Tr_2—BC109 (Mullard); Tr_3—40362 (R.C.A.); Tr_4—BC107 (Mullard); Tr_5—BC125 (Fairchild); Tr_6—40361 (R.C.A.); Tr_7—MJ481 (Motorola); Tr_8—BC126 (Fairchild); Tr_9—40362 (R.C.A.); Tr_{10}—MJ491 (Motorola). Changes of values are indicated by black squares.



Modular Pre-amplifier Design

Optimally designed stages that may be used separately or in several different combinations

By J. L. Linsley Hood, M.I.E.E.

The type of distortion introduced by a class A transistor amplifier operating at low signal level will be predominantly second harmonic and inoffensive to the ear. Although harmonic distortion is a convenient thing to measure, and makes a reasonable yardstick for comparative purposes, at low levels its presence is less important than that of the intermodulation effects which it causes. When a complex signal is transmitted through a non-linear element, intermodulation products between the separate components of the signal are formed, and these are readily apparent in the final audible result as a 'blurring', and loss of separate identity, of the individual components which make up the whole. A measure of this is the ease (or difficulty) in distinguishing the words of a choral performance in the presence of an orchestral background, or in identifying the presence and nature of individual instruments in a large orchestra.

John Laurence Linsley Hood, born in 1925, was educated at Reading School, Acton Polytechnic, the Royal Technical College (Glasgow) and, after the war, at Reading University. In 1942 he joined the G.E.C. Research Laboratories at Wembley, working on magnetron development as junior member of a team. In 1943 he joined the R.A.F. in aircrew but was transferred to work on radar. He subsequently worked with T.R.E. (Malvern) overseas. After a return to university he joined the Windscale Research Laboratories of the Atomic Energy Authority. He has been in charge of the electronics team in the research laboratories of British Cellophane Ltd. since 1954.

Measurements by a number of workers[1] have indicated that the magnitude of intermodulation products can be much greater than that of the total harmonic distortion level, and the non-linearities which are likely to be of most importance in this respect are those at the low- and high-frequency ends of the audible range.

At the moment, the performance of audio amplifiers is much superior in this respect to that of f.m. transmissions, tape recordings, disc replay systems or loudspeakers. However, advances in the manufacturing techniques of gramophone records, pickup cartridges and loudspeakers have allowed a continuing improvement in the performance of these in harmonic and i.m. distortion, and it is clear that any amplifier design offered at this time should have a very high standad of performance if it is to remain of continuing value over the next decade.

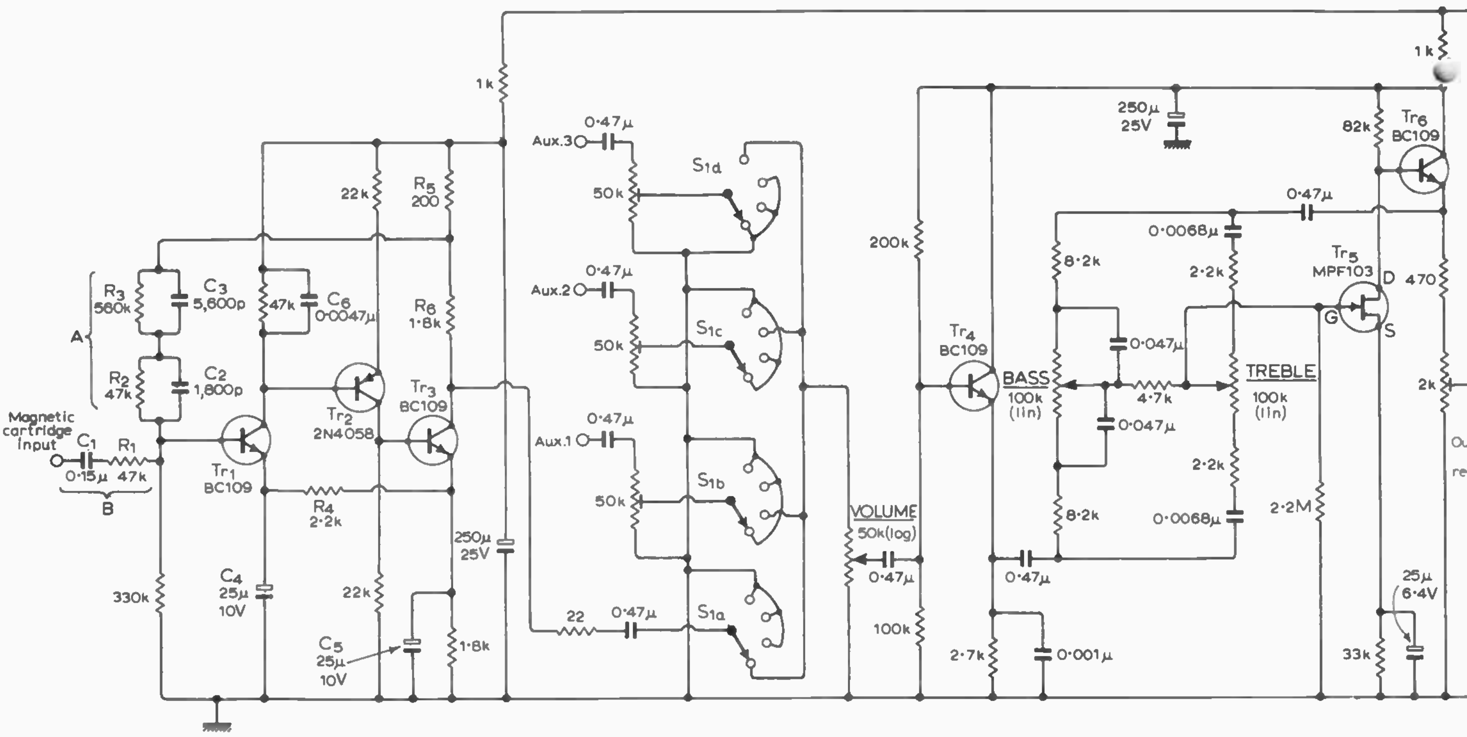


Fig. 1. A likely combination of stages.

The author has designed a range of high-quality pre-amplifier stages. Each stage performs its required operation with negligible noise and distortion. When joined together, as for example in Fig. 1, the total harmonic distortion level is below 0.1% over the frequency range 20 Hz-20 kHz, at any tone-control setting, and for up to 2V r.m.s. output. Each stage is capable of operating on its own and has an output impedance low enough for screened cable inter-connections to be made without high frequency loss.

Magnetic pickup equalization circuit

The required R.I.A.A. replay characteristic can be approximated by several different circuit arrangements. The most straightforward from the point of view of performance calculation is that shown in Fig. 2, employing a simple phase-inverting amplifier stage. If the gain of amplifier M is high enough, point Z becomes a virtual earth (see Appendix I), and the input impedance of the circuit equivalent to that of the input network B. The load resistance required by the pickup cartridge, usually 47-50 kΩ, is provided by a suitable choice of R_1. With resistor R_2 equal to R_1, stage gain is given by $\frac{R_4 + R_5}{R_5}$ at the mid-point frequency (usually 1 kHz) if the impedance of C_2 is large, and that of C_3 small, in relation to R_2. Since the voltage output to be expected from most good quality magnetic pickup cartridges is in the range 4-10 mV for a 5 cm/sec recorded velocity, a gain of 10 is adequate for this stage. The required replay frequency-response curve shown in Fig. 3 can be obtained by a suitable choice of C_2 and C_3. Since the two networks A and B determine the frequency response of this circuit, it is apparent that substitution of these can be made to provide a wide range of different performance characteristics without alteration to the circuit of the amplifier unit M.

The final circuit can be seen at the front of Fig. 1. Because phase inversion between input and output is required, and because the necessary gain is higher than can be obtained from any single transistor arrangement, a triplet circuit has been used. Tr_1 and Tr_3 are high-gain, low-noise voltage-amplifying stages, and Tr_2 is a phase and voltage transformation stage allowing the input transistor to be used in its most linear region. The output transistor has a low collector load resistance, to reduce distortion to the lowest possible level.

D.c. working-point stability is ensured by d.c. negative feedback through R_3 and R_2 to the base of Tr_1, and through R_4 to the emitter circuit of the same transistor. The circuit R_4, C_4 and

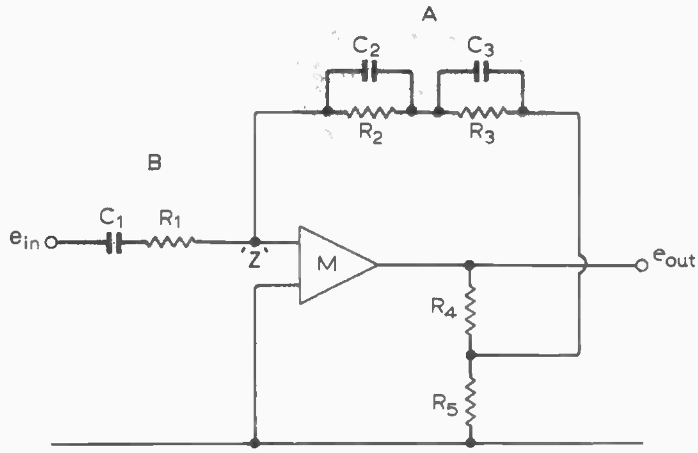


Fig. 2. Phase-inverting amplifier stage used to obtain R.I.A.A. replay characteristic.

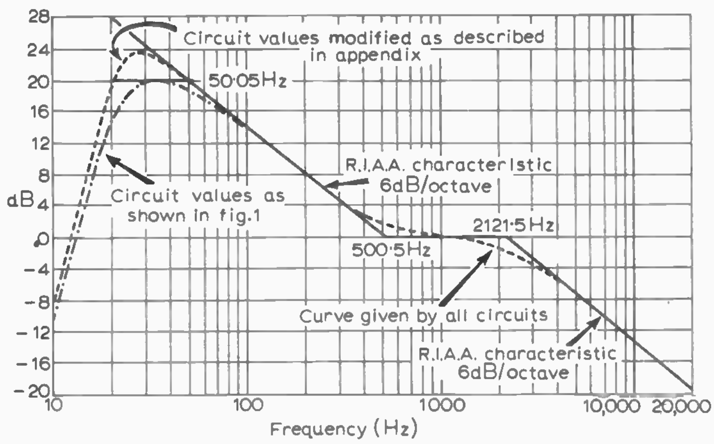
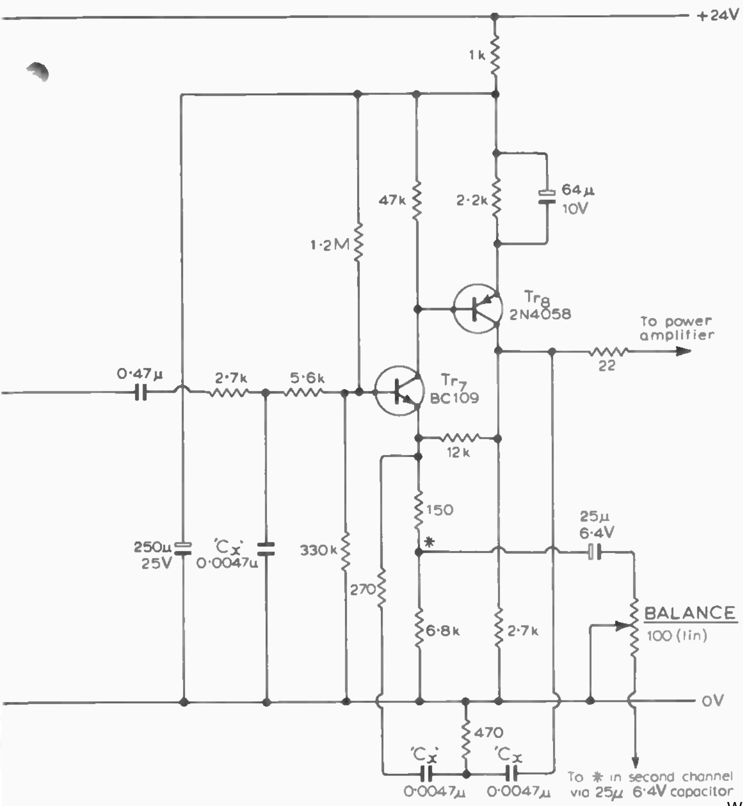


Fig. 3. Required R.I.A.A. frequency-response curve and circuit approximation to this.

C_5 also provides the feedback path necessary, in conjunction with the input capacitor C_1, to provide an 18 dB/octave steep-cut rumble filter, with a turn-over frequency of 25 Hz (see Appendix II), and an ultimate attenuation of more than 40 dB at 8 Hz.

Capacitor C_6 provides phase correction, and is essential for a clean square-wave response, and freedom from transient ringing, when used with a capacitive load.

The response of this circuit is particularly good, and it can deliver up to 1 volt output with distortion less than 0.02% from 100 Hz to 10 kHz.



Stages for ceramic cartridge equalization

Fig. 4 is an impedance conversion stage contributing less than 0.05% distortion at 1 kHz and having a flat response from 35 Hz to greater than 200 kHz, with 18 dB/octave roll-off below 35 Hz. This simple stage may be directly substituted for the magnetic cartridge stage of Fig. 1.

Alternatively, should it be required that the pre-amplifier be able to cope with inputs from both magnetic and ceramic cartridges, then switchable equalization networks for A and B can be provided. These are shown in Fig. 5. When used with a ceramic cartridge the output voltage is from 50 to 200 mV. To preserve the required shape of the rumble filter characteristic it is necessary to alter the values of C_4 and C_5 from 25 μF to 12.5 μF. The pre-amp response is then as shown in Fig. 5, curve 1.

The performance of many ceramic pickup/amplifier combinations is disappointing in comparison with that obtainable from a good magnetic cartridge with a similar amplifier. This is sometimes due to the mismatching between cartridge and amplifier, or through inadequate input impedance provision (in the modification shown in Fig. 5 this is 4.4 M Ω), or due to the failure of the piezoelectric element within the cartridge to provide the required equalization for the 12 dB fall in voltage output anticipated when a recording having R.I.A.A. velocity characteristics is replayed on a displacement sensitive device. In the latter case, a very considerable improvement in the relative performance of the ceramic cartridge may be obtained by shunting part of the input resistor in the input network B by a small capacitor. Curves 2 and 3 in Fig. 5 show partial and complete correction respectively.

Tone-control stage

The tone-control stage is of conventional type, and uses a negative feedback system derived from the design due to Baxandall[2]. However, it differs from normal practice in that a junction field-effect transistor is used as the active element. Field-effect transistors have both lower noise levels and better linearity than bipolar transistors, and in this type of circuit the high input impedance results in negligible loading of the tone-control network. The stage gain needed in this circuit requires a high value drain load resistor, and the f.e.t. must therefore be followed by an emitter-follower to provide the low output impedance desired for easy interconnection of the separate units.

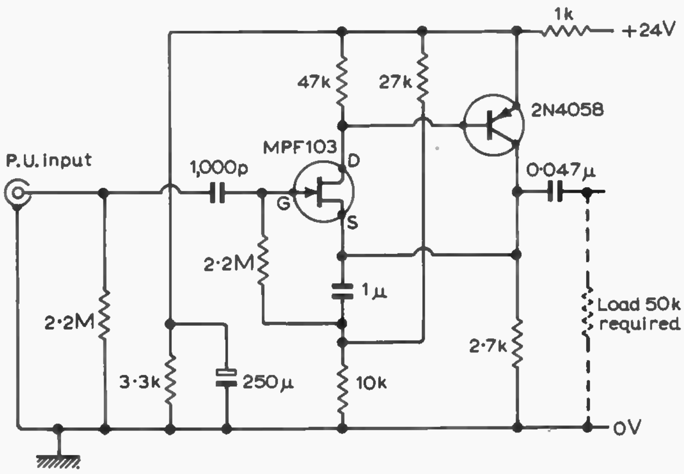


Fig. 4. Impedance conversion stage for use with ceramic cartridge. This may be directly substituted for the magnetic cartridge stage at the front of Fig. 1.

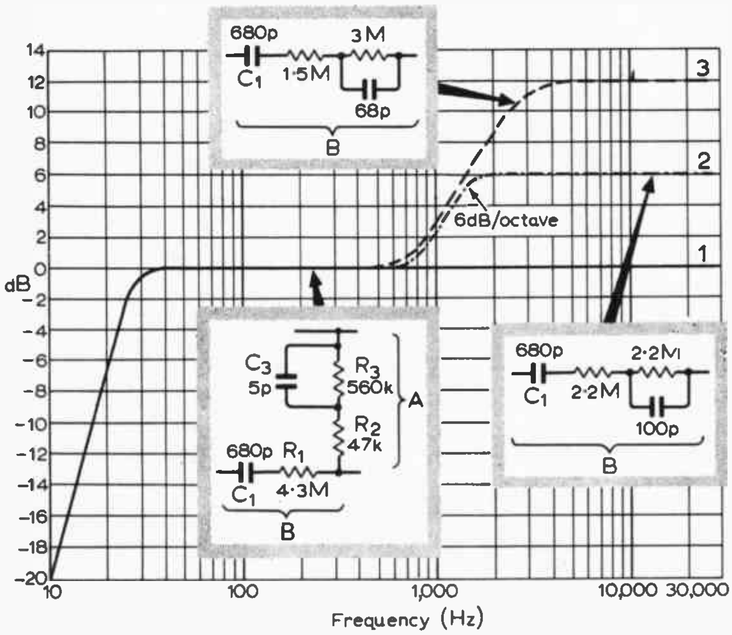


Fig. 5. Changes in equalization networks A and B of the magnetic cartridge input stage allowing direct use of ceramic cartridge. Components for network A are the same for the three curves show.

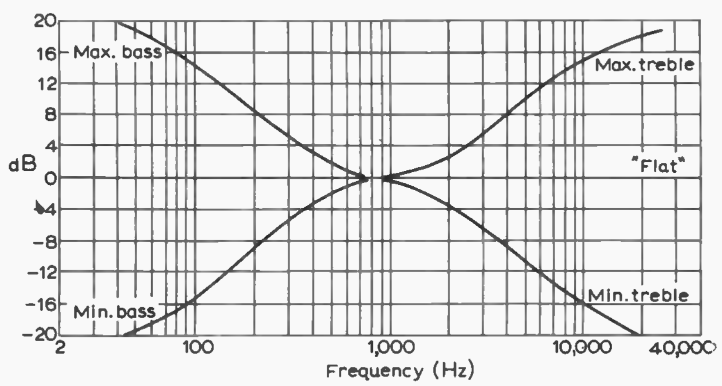


Fig. 6. Gain/frequency characteristics of the tone control stage.

If the feedback tone-control network is to perform satisfactorily, both the input and output impedances seen by the network at its ends must be low in relation to the network input impedance when the sliders of the potentiometers are at the position nearest to the point being measured. Some form of impedance conversion circuit is therefore also needed between the volume control and the tone-control circuit. An emitter follower is also used at this point. The 0.001μF capacitor in the emitter circuit of Tr_4 is to avoid the possibility of high frequency parasitic oscillation occurring if long screened leads are used to connect the base of Tr_4 to the volume control.

The input to this section is taken through a switch from the gramophone pre-amplifier section, and other inputs provided with preset gain-equalization potentiometers. The switch is arranged to earth the inputs not in use, to minimize break-through between programme channels.

The gain/frequency characteristics of the stage are shown in Fig. 6.

Low-pass filter circuit

The voltage amplifying stage preceding the main amplifier should include a steep-cut low-pass filter that can be set to remove unwanted high frequencies. This can be done either by a suitable *LCR* filter arrangement, or by an active filter giving an equivalent performance without the use of inductors. The circuit arrangements available for low-pass active filters are shown in Fig. 7. (b) is the well known circuit arrangement first employed in an audio amplifier design by P. J. Baxandall[3], and (d) is the unity gain rearrangementof this circuit introduced by Sallen and Key[4]. The frequency response of all of these circuit arrangements is similar, *mutatis mutandis*, to that shown in Fig. 8, and the circuit should be preceded or followed by a simple *RC* filter if the type of response shown in the dotted line is required.

For a given overall stage gain, type (b) gives a much better distortion factor near the region of cut-off than (a), and (c) is marginally better than (b) when used with non-linear amplifier elements. The particular advantage of (c), however, is that it can be used conveniently with a very low-distortion two-transistor circuit.

The final stage, with the filter circuitry, is shown in Fig. 1. As a matter of practical convenience, the component values of this circuit have been chosen so that the required low-pass response is obtained when all of the capacitors 'C_x' are of equal value to each other. The frequency response obtained with a given value of 'C_x' can be found from Fig. 9. The user can interpolate between these to obtain turn-over frequencies at any points to suit his own requirements. If a ganged selector switch is employed to give a range of turn-over frequencies, the switch arms (moving contacts) should be connected to the junction of the resistors in the *RC* filter and to the 470 Ω resistor in the main filter network. In Fig. 1 the 0.0047 μF capacitor for 'C_x' results in response being 3 dB down at about 18 kHz. With good quality programme sources this is a recommended capacitor value.

With capacitors of zero value, the response of the circuit is flat to about 100 kHz. The user should, however, arrange for the response to fall off above 25 kHz. (It is unlikely that the listener will find anything to gain from the parts of the sonic spectrum beyond this point.)

The optimum performance of this particular type of circuit arrangement is obtained when the overall gain is about 50 with feedback. A 20-40 mV input is therefore adequate for this stage for the output voltages required.

The distortion level of this circuit is less than 0.03 % at 2 volts r.m.s. output or less, at any frequency within the pass band. The output impedance is less than 150 ohms over the range from 20 Hz to the cut-off frequency selected.

It is convenient, for several reasons, to operate at the 60-100 mV level through the tone-control stages. At this output voltage level the distortion introduced by an *RC* coupled f.e.t. stage is less than 0.1% even without feedback, so that the maximum 'lift' settings of either 'bass' or 'treble' controls cannot give rise to unacceptable levels of distortion. It is also large enough for the noise and inevitable 50 Hz pickup to be unobtrusive. Some attenuation is therefore desirable between the tone control unit and the steep-cut filter circuit. This is obtained by the preset 2 k Ω potentiometer in the tone control circuit, which provides a convenient means for setting the overall gain of the amplifier system, and also as a coarse 'balance control' in a stereo system. Fine balance between channels is obtained by adjusting the 100 Ω balance potentiometer in the output stage. This alters the stage gain over the ratio 6 : 10.

Constructional notes

The constructional technique used by the author in building the prototype of this amplifier is similar to that used in the 10-watt class A design described in *Wireless World* in April 1969, with the separate units laid out in mirror image form, as a stereo pair on a single 4 in. × 4¾ in. s.r.b.p. pin board. Two units of each type can be accommodated on each board, laid out more or less in the form of the circuit diagram (or its mirror image).

In general, reasonable care should be taken to separate input from output leads, and where the boards are to be mounted as a group within the same box, it would be wise to interpose a sheet metal screen between them.

The units are separately decoupled by 250 μF capacitors from a common 24-volt line, derived from a zener diode stabilized *RC* filter power supply. This supply is separate from the main amplifier, and a 30 mA output is ample. Details of a suitable power supply are given in Fig. 10. The expected working voltage on each of the unit sub-rails is about 15 volts.

Apart from the input transistor in the gramophone pre-amp unit (Tr_1) for which the BC109 is to be preferred, there is no particular reason why any modern silicon planar types should not give an indistinguishable performance. For example, the n-p-n types could be 2N3904, BC107/8/9, 2N3707 or BC184Ls. Similarly, the p-n-p types could be 2N4058, 2N3906 or BC214Ls.

Although, in many cases, the use of ¼ watt resistors is sufficient, it will probably be found simpler to use ½ watt units throughout. 5% tolerance carbon film resistors are to be preferred.

The author has mounted the gramophone pickup equalization circuit in a separate small diecast box, immediately under the gramophone turntable unit, so that the leads from the gramophone are taken at a low impedance from the output of this unit. This has been very effective in reducing the hum picked up on the output leads to an imperceptible level.

Appendix I

The use of 'virtual earth' (null seeking) amplifier circuit arrangements is superficially ill-advised with input elements such as pickup cartridges, because it appears that as the operating frequency is increased, the input half of the balancing limbs will also change, with a resultant change in the gain of the circuit. In particular, a magnetic pickup cartridge may have an inductance of some 300-800 mH and the impedance of this will exceed that of the input circuit in the range 12-20 kHz. This should clearly reduce the gain of the system by reducing the ratio of A to B.

However, on reflection, it can be seen that the amplifier operates as a null generating device, sensitive only to the current flowing in the input circuit to the 'virtual earth'. As the operating frequency increases, so the current flow through R_1 will decrease, but so it would in any case, regardless of the amplifier, were the element simply connected across network B at the load recommended by the cartridge manufacturers (at these frequencies the impedance of C_1 can be ignored), and the voltage across R_1 measured by a perfect voltage amplifier. The decrease of current input into a given resistive load from a source having series inductance is simply an unfortunate fact of life, from which one cannot escape, whatever one's technique of measurement, and high impedance voltage amplifiers connected across the load, or low impedance current amplifiers connected in series with it, are alike in this respect, except that with transistors the latter are a bit easier to contrive. The same argument is also applicable, in the appropriate context, to high impedance capacitative elements such

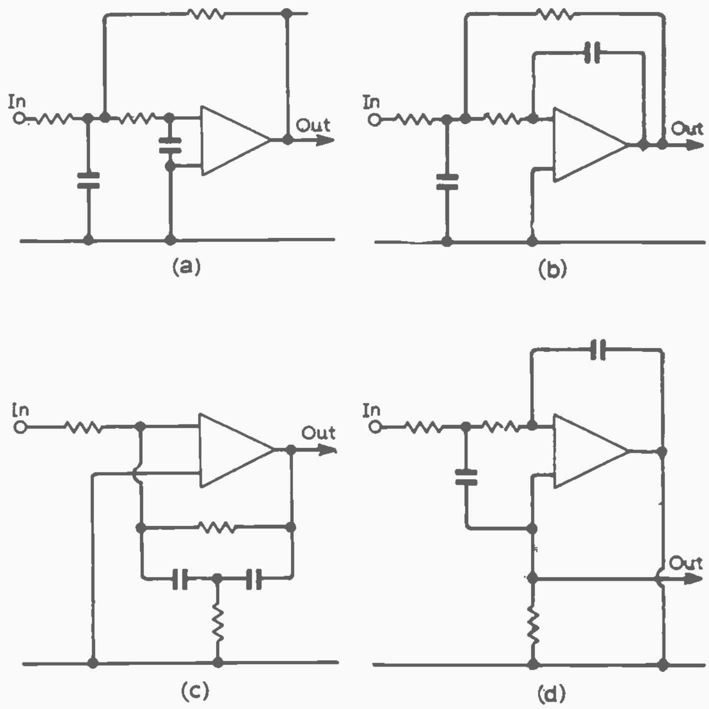


Fig. 7. Circuit arrangements for active low-pass filter design.

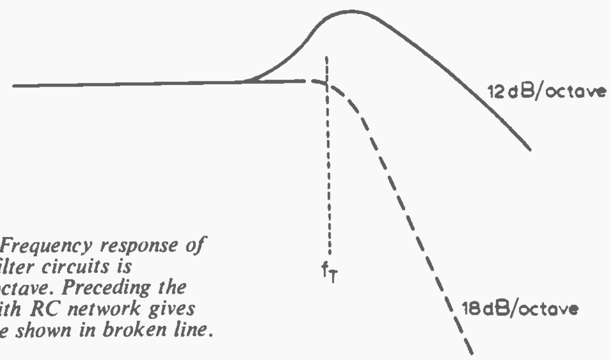


Fig. 8. Frequency response of active filter circuits is 12 dB/octave. Preceding the filter with RC network gives response shown in broken line.

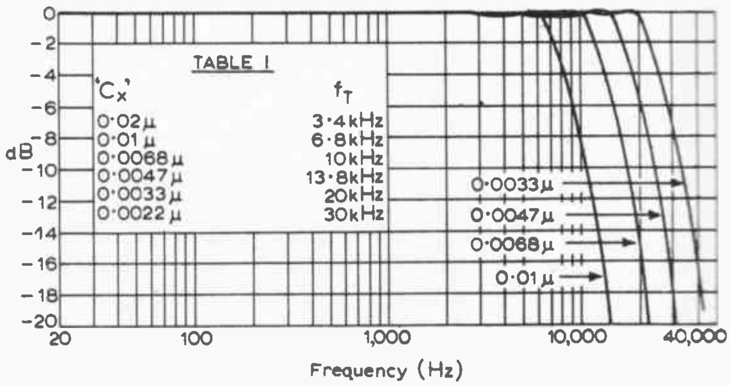


Fig. 9. Graph and table of turn-over frequencies for different value of 'Cx'.

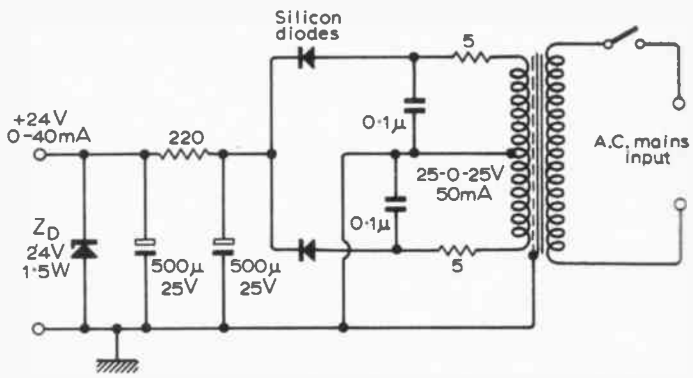


Fig. 10. Suitable power supply for any combination of stages.

as piezo-electric pickup cartridges. Once again, the voltage amplifier and current amplifier see the same phenomena in identical form. The necessary, and inevitable, corrections can be accomplished simply by the tone-control settings.

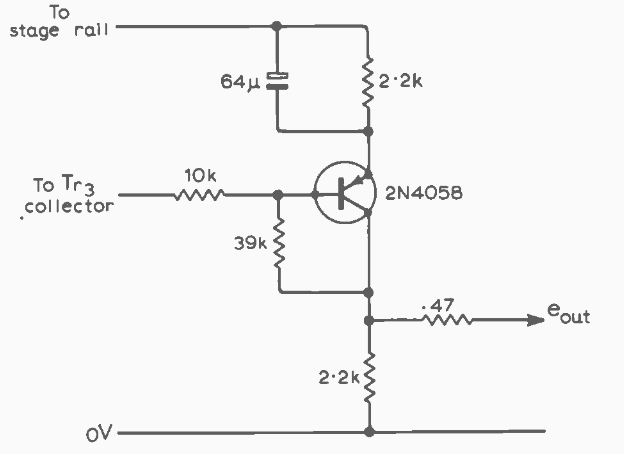


Fig. 11. Floating emitter collector-follower circuit referred to in Appendix II.

Appendix II

Although the R.I.A.A. replay characteristic suggests an approximately flat velocity response from 20 Hz-50 Hz, this would effectively imply recording bass lift in this region, and the author suspects that this is not done, a constant modulation characteristic being used instead. The author has therefore, for his own use, modified the values of the feedback elements as follows: R_5-470 ohms; R_6-1.5 k ohms; C_1-0.47 μF; C_x-6800 pF; and C_6-6800 pF. These changes maintain the velocity response flat down to 25 Hz, with a rapid rumble attenuation below this frequency. Unfortunately the mid point gain of the circuit is reduced to 5, and some additional amplification is therefore needed if it is desired to avoid working with the tone control circuit at the 20 mV level. The simple floating emitter collector-follower circuit of Fig. 11 is therefore interposed, without coupling capacitors, between the output series resistor and the collector of Tr_3. The distortion contributed by this is less than 0.05%.

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1. Langford-Smith, F., 'Radio Designers Handbook', Vol. 4, ch. 72.
3. Baxandall, P. J., 'Gramophone and Microphone Pre-amplifier', *Wireless World*, October 1952.
3. Baxandall, P. J., 'Gramophone and Microphone Pre-amplifier', *Wireless World*, January 1955.
4. Sallen, R. P. and Key, E. L., *I.R.E. Trans. Circuit Theory*, March 1955, pp. 74-85.

Letters to the Editor

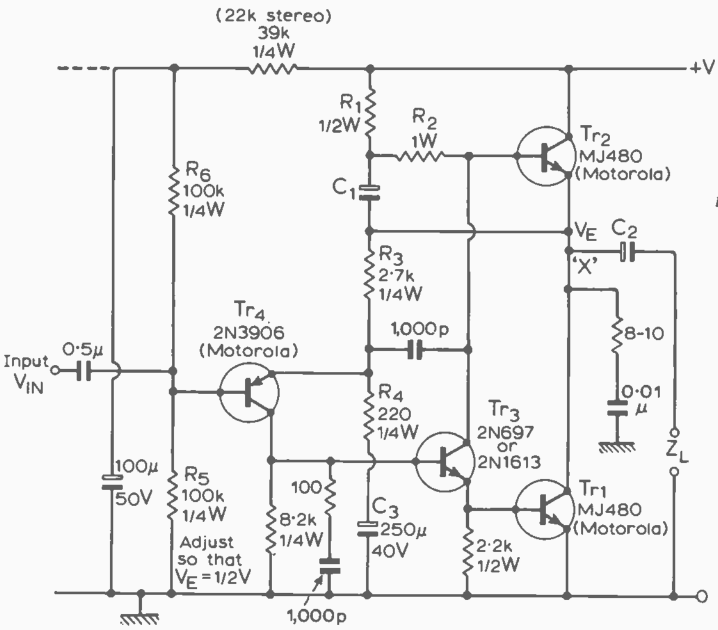
Linsley Hood class A amplifier

Recent measurements on this amplifier have indicated that the gain and power bandwidths of this design, using the component layout shown on page 75 of your issue, are wider than indicated by the Figs. 4 and 5 of the article. The apparent fall-off in gain beyond about 100 kHz was, in fact, due to shortcomings in the measurement apparatus, and measurements made with better equipment suggest that the —3 dB points for voltage gain are above 1.5 MHz although power output falls beyond 200 kHz.

Since the output is in phase with the input, it is necessary to take care that the output leads and output capacitor are not close to the input. (A 2-inch separation will be adequate for normal lead lengths.) However, an additional point must also be noted. If a capacitative load is connected with *short* leads between the output and the earth line near the input connection, the potential developed along the earth line, due to its inductance, can inject an in-phase signal, and thereby cause instability, in the MHz region. To avoid this possibility, it is recommended that the earthy lead to the loudspeaker terminal be returned to the earth line at the same point as the emitter of Tr_1. The inclusion of a small r.f. choke (25 turns of 26-28 s.w.g. wire wound round the outside of a 10-ohm 1-watt resistor is ideal) between the output (point 'X') and C_2 will also prevent this possibility of trouble.

In practice, with the components and layout suggested, the inductance of the normal 12 to 18 inches (or more) of loudspeaker connecting lead prevents instability with capacitative loads, so this should be only of academic interest.

As an alternative, it is possible to reduce the r.f. response of the amplifier to give a smooth, 6 dB/octave, roll-off beyond 50 kHz—which removes much of the need for care in the layout of the components, without detriment to the harmonic distortion in the audible range, and without any audible alteration to the performance—by connecting a 1,000 pF capacitor between the collector of Tr_3 and the emitter of Tr_4; a 1,000 pF in series with 100 ohms between the collector of Tr_4 and earth; and a 0.01 μF in series with a 8 to 10 ohms between the output ('X') and earth. (It should be noted that either all of these components should be added or none at all, they are not alternatives.) If the r.f. response is reduced in this manner, the use of a series r.f. choke would be unnecessary.



Mr. Linsley Hood's amended circuit of his class A amplifier originally described in the April 1969 issue.

A series of measurements has also been made, using the amplifier design exactly as described in the article (without r.f. chokes or other modifications) to determine the voltage waveform produced, actually across the loudspeaker, with a square wave input to the amplifier. It was found, in practice, with several different loudspeaker systems, that the output waveform was virtually identical to that obtained with an equivalent resistive load—photographs of which were reproduced in the April issue. It was, in fact, the discovery that a good square wave was reproduced up to the 1 MHz limit of the generator in use which prompted a reassessment of the r.f. response of the amplifier. The absence of any overshoot or significant ringing also provides confirmation of the stability of the amplifier under practical conditions.

A correspondent has reported that this design has been up-rated successfully to 15 watts into a 15-ohm load, to give a direct power equivalent to the Williamson amplifier, using 2N3055 output transistors with a 43-volt supply (1.1 amp per channel), and rather larger heat sinks. There would seem no good reason why this could not also be done using MJ481s.

J. L. Linsley Hood,

Simple Class A Amplifier

A 10-W design giving subjectively better results than class B transistor amplifiers

by J. L. Linsley Hood, M.I.E.E.

During the past few years a number of excellent designs have been published for domestic audio amplifiers. However, some of these designs are now rendered obsolescent by changes in the availability of components, and others are intended to provide levels of power output which are in excess of the requirements of a normal living room. Also, most designs have tended to be rather complex.

In the circumstances it seemed worth while to consider just how simple a design could be made which would give adequate output power together with a standard of performance which was beyond reproach, and this study has resulted in the present deisgn.

Output power and distortion

In view of the enormous popularity of the Mullard '5-10' valve amplifier, it appeared that a 10-watt output would be adequate for normal use; indeed when two such amplifiers are used as a stereo pair, the total sound output at full power can be quite astonishing using reasonably sensitive speakers.

The original harmonic distortion standards for audio amplifiers were laid down by D. T. N. Williamson in a series of articles published in *Wireless World* in 1947 and 1949; and the standard proposed by him, for less than 0.1% total harmonic distortion at full rated power output, has been generally accepted as the target figure for high-quality audio power amplifiers. Since the main problem in the design of valve audio amplifiers lies in the difficulty in obtaining adequate performance from the output transformer, and since modern transistor circuit techniques allow the design of power amplifiers without output transformers, it seemed feasible to aim at a somewhat higher standard, 0.05% total harmonic distortion at full output power over the range 30 Hz-20 kHz. This also implies that the output power will be constant over this frequency range.

Circuit design

The first amplifier circuit of which the author is aware, in which a transformerless transistor design was used to give a standard of performance approaching that of the 'Williamson' amplifier, was that published in *Wireless World* in 1961 by Tobey and Dinsdale. This employed a class B output stage, with series connected transistors in quasi-complementary symmetry. Subsequent high-quality transistor power amplifiers have largely tended to follow the design principles outlined in this article.

The major advantage of amplifiers of this type is that the normal static power dissipation is very low, and the overall power-conversion efficiency is high. Unfortunately there are also some inherent disadvantages due to the intrinsic dissimilarity in the response of the two halves of the push-pull pair (if complementary transistors are used in unsymmetrical circuit arrangement) together with some cross-over distortion due to the low current non-linearity of the I_c/V_b characteristics. Much has been done, particularly by Bailey[1], to minimize the latter.

An additional characteristic of the class B output stage is that the current demand of the output transistors increases with the output signal, and this may reduce the output voltage and worsen the smoothing of the power supply, unless this is well designed. Also, because of the increase in current with output power, it is possible for a transient overload to drive the output transistors into a condition of thermal runaway, particularly with reactive loads, unless suitable protective circuitry is employed. These requirements have combined to increase the complexity of the circuit arrangement, and a well designed low-distortion class B power amplifier is no longer a simple or inexpensive thing to construct.

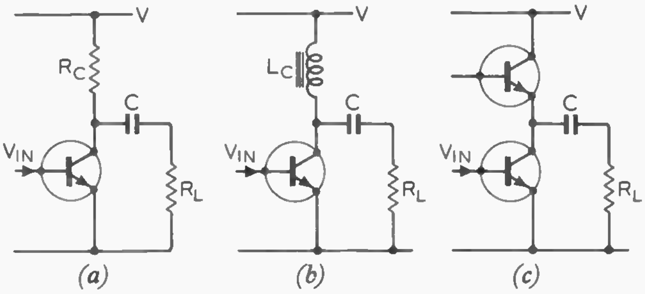


Fig. 1. Basic class A circuits using (a) load resistor R_c giving power conversion efficiency of about 12%, (b) l.f. choke giving better efficiency but being bulky and expensive, and (c) a second transistor as collector load.

An alternative approach to the design of a transistor power amplifier combining good performance with simple construction is to use the output transistors in a class A configuration. This avoids the problems of asymmetry in quasi-complementary circuitry, thermal runaway on transient overload, cross-over distortion and signal-dependent variations in power supply current demand. It is, however, less efficient than a class B circuit, and the output transistors must be mounted on large heat sinks.

The basic class A construction consists of a single transistor with a suitable collector load. The use of a resistor, as in Fig. 1(a), would be a practical solution, but the best power-conversion efficiency would only be about 12%. An l.f. choke, as shown in Fig. 1(b), would give much better efficiency, but a properly designed component would be bulky and expensive, and remove many of the advantages of a transformerless design. The use of a second, similar, transistor as a collector load, as shown in Fig. 1(c), would be more convenient in terms of size and cost, and would allow the load to be driven effectively in push-pull if the inputs to the two transistors were of suitable magnitude and opposide in phase. This requirement can be achieved if the driver transistor is connected as shown in Fig. 2.

This method of connection also meets one of the most important requirements of a low distortion amplifier—that the basic linearity of the amplifier should be good, even in the absence of feedback. Several factors contribute to this. There is the tendency of the I_c/V_b non-linearity of the characteristics of the output transistors to cancel, because during the part of the cycle in which one transistor is approaching cut-off the other is turned full on. There is a measure of internal feedback around the loop Tr_1, Tr_2, Tr_3 because of the effect which the base impedance characteristics of Tr_1 have on the output current of Tr_3. Also, the driver transistor Tr_3, which has to deliver a large voltage swing, is operated under conditions which favour low harmonic distortion—low output load impedance, high input impedance.

A practical power amplifier circuit using this type of output stage is shown in Fig. 3.

The open loop gain of the circuit is approximately 600 with typical transistors. The closed loop gain is determined, at frequencies high enough for the impedance of C_3 to be small in comparison to R_4, by the ratio $(R_3 + R_4)/R_4$. With the values indicated in Fig. 3, this is 13. This gives a feedback factor of some 34 dB, and an output impedance of about 160 milliohms.

Since the circuit has unity gain at d.c., because of the inclusion of C_3 in the feedback loop, the output voltage, V_e, is held at the same potential as the base of Tr_4 plus the base emitter potential of Tr_4 and the small potential drop along R_3 due to the emitter current of this transistor. Since the output transistor Tr_1 will turn on as much current as is necessary to pull V_e down to this value, the resistor R_2, which together with R_1 controls the collector current of Tr_2, can be used to set the static current of the amplifier output stages. It will also be apparent that V_e can be set to any desired value by small adjustments to R_5 or R_6. The optimum performance will be obtained when this is equal to half the supply voltage. (Half a volt or so either way will make only a small difference to the maximum output power obtainable, and to the other characteristics of the amplifier, so there is no need for great precision in setting this.)

Silicon planar transistors are used throughout, and this gives good thermal stability and a low noise level. Also, since there is no requirement for complementary symmetry, all the power stages can use n-p-n transistors which offer, in silicon, the best performance and lowest cost. The overall performance at an output level of 10 watts, or any lower level, more than meets the standards laid down by Williamson. The power output and gain/frequency graphs are shown in Figs. 4-6, and the relationship between output power and total harmonic distortion is shown in Fig. 7. Since the amplifier is a straightforward class A circuit, the distortion decreases linearly with output voltage. (This would not necessarily be the case in a class B system if any significant amount of crossover distortion was present.) The analysis of distortion components at levels of the order of 0.05% is difficult, but it appears that the residual distortion below the level at which clipping begins is predominantly second harmonic.

Stability, power output and load impedance

Silicon planar n-p-n transistors have, in general, excellent high frequency characteristics, and these contribute to the very good stability of the amplifier with reactive loads. The author has not yet found any combination of *L* and *C* which makes the system unstable, although the system will readily become oscillatory with an inductive load if R_3 is shunted by a small condenser to cause roll-off at high frequencies.

The circuit shown in Fig. 3 may be used, with very little modification to the component values, to drive load impedances in the range 3-15 ohms. However, the chosen output power is represented by a different current/voltage relationship in each case, and the current through the output transistors and the output-voltage swing will therefore also be different. The peak-voltage swing and the mean output current can be calculated quite simply from the well-known relationships $W = I^2R$ and $V = IR$, where the symbols have their customary significance. (It should be remembered, however, that the calculation of output power is based on r.m.s. values of current and voltage, that these must be multiplied by 1.414 to obtain the peak values, and that the voltage swing measured is the peak-to-peak voltage, which is twice the peak value.)

When these calculations have been made, the peak-to-peak voltage swing for 10 watts power into a 15-ohm load is found to be 34.8 volts. Since the two output transistors bottom at about 0.6 volt each, the power supply must provide a minimum of 36 volts in order to allow this output. For loads of 8 and 3 ohms, the minimum h.t. line voltage must be 27 V and 17 volts respectively. The necessary minimum currents are 0.9, 1.2 and 2.0 amps. Suggested component values for operation with these load impedances are shown in Table 1. C_3 and C_1 together influence the voltage and power roll-off at low audio frequencies. These can be increased in value if a better low-frequency performance is desired than that shown in Figs. 4-6.

Since the supply voltages and output currents involved lead to dissipations of the order of 17 watts in each output transistor, and since it is undesirable (for component longevity) to

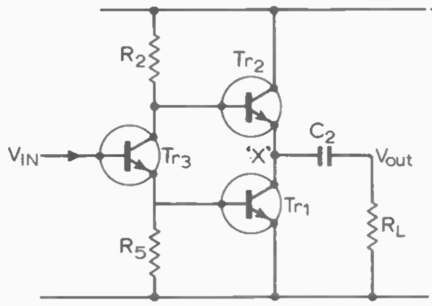


Fig. 2. Arrangement for push-pull drive of class A stage.

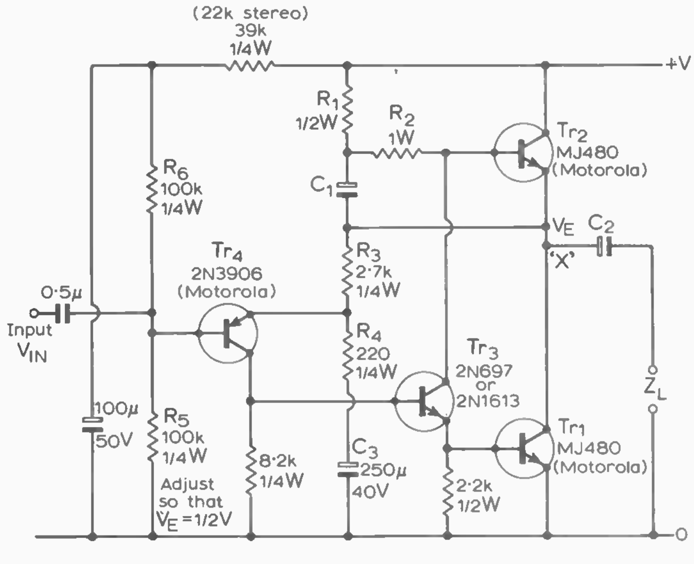


Fig. 3. Practical power amplifier circuit.

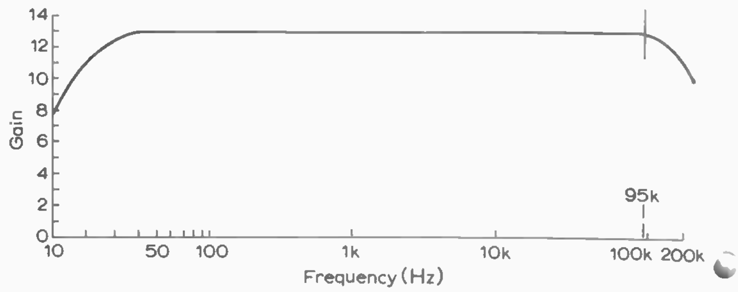


Fig. 4. Gain/frequency response curve of amplifier.

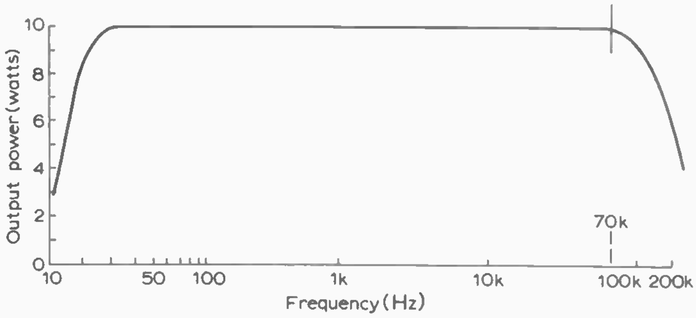


Fig. 5. Output power/frequency response curve of amplifier.

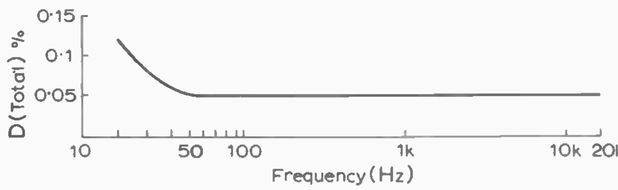


Fig. 6. Distortion/frequency curve at 9W.

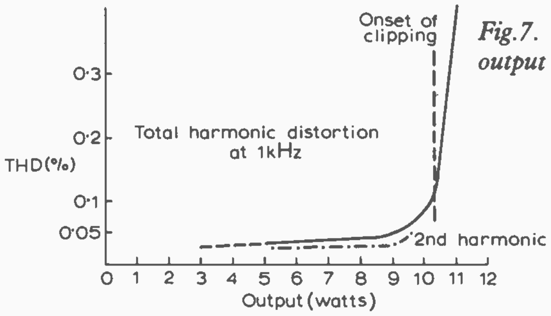


Fig.7. Distortion/ output power curve.

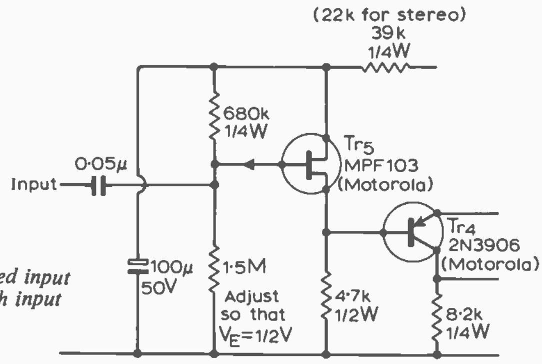


Fig. 8. Modified input circuit for high input impedance.

permit high operating temperatures, adequate heat sink area must be provided for each transistor. A pair of separately mounted 5 in. × 4 in. finned heat sinks is suggested. This is, unfortunately, the penalty which, must be paid for class A operation. For supplies above 30 V Tr_1 and Tr_2 should be MJ481s and Tr_3 a 2N1613.

If the output impedance of the pre-amplifier is more than a few thousand ohms, the input stage of the amplifier should be modified to include a simple f.e.t. source follower circuit, as shown in Fig. 8. This increases the harmonic distortion to about 0.12%, and is therefore (theoretically) a less attractive solution than a better pre-amplifier.

A high frequency roll-off can then be obtained, if necessary, by connecting a small capacitor between the gate of the f.e.t. and the negative (earthy) line.

Suitable transistors

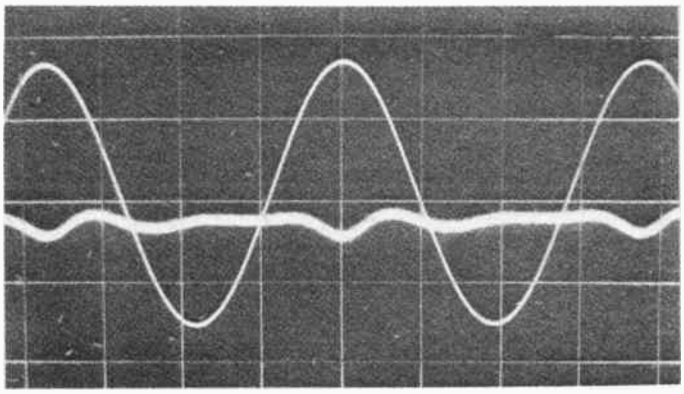
Some experiments were made to determine the extent to which the circuit performance was influenced by the type and current gain of the transistors used. As expected the best performance was obtained when high-gain transistors were used, and when the output stage used a matched pair. No adequate substitute is known for the 2N697/2N1613 type used in the driver stage, but examples of this transistor type from three different manufacturers were used with apparently identical results. Similarly, the use of alternative types of input transistor produced no apparent performance change, and the Texas Instruments 2N4058 is fully interchangeable with the Motorola 2N3906 used in the prototype.

The most noteworthy performance changes were found in the current gain characteristics of the output transistor pair, and for the lowest possible distortion with any pair, the voltage at the point from which the loudspeaker is fed should be adjusted so that it is within 0.25 volt of half the supply line potential. The other results are summarized in Table 2.

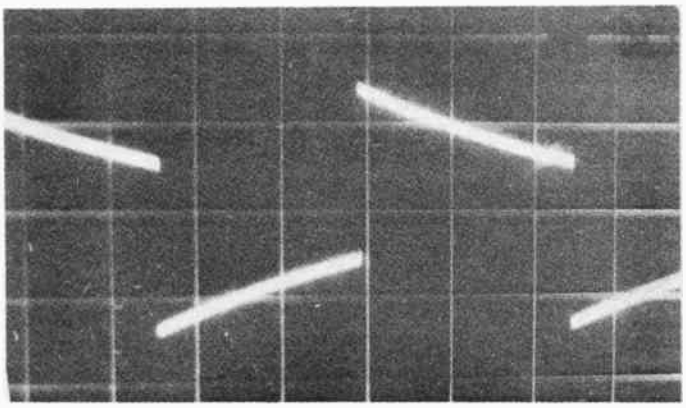
The transistors used in these experiments were Motorola MJ480/481, with the exception of (6), in which Texas 2S034 devices were tried. The main conclusion which can be drawn from this is that the type of transistor used may not be very important, but that if there are differences in the current gains

Table 1. Summary of component combinations for different load impedances.

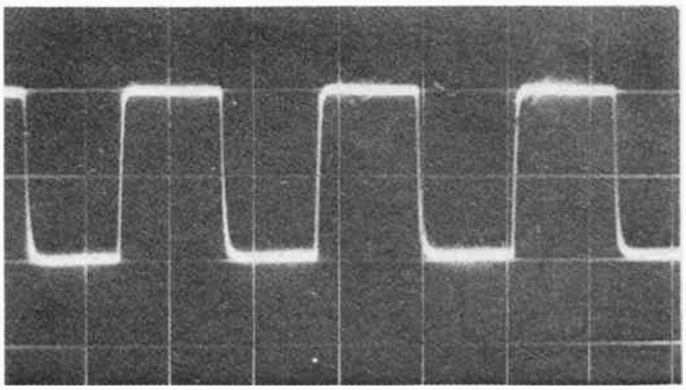
Z_L	V	I	R_1	R_2	C_1		C_2		V_{IN} (r.m.s.)
3Ω	17V	2A	47Ω	180Ω	500µ	25V	5000µ	25V	0.41V
8Ω	27V	1.2A	100Ω	560Ω	250µ	40V	2500µ	50V	0.66V
15Ω	36V	0.9	150Ω	1.2kΩ	250µ	40V	2500µ	50V	0.9V



Sine wave performance at 1kHz. 9 watts, 15 ohm resistive load. Fundamental on scale of 10 V/cm. Distortion components on scale of 50 mV/cm with r.m.s. value of 0.05%.



Square wave response at 50 Hz.



Square wave response. Scale 10 V/cm. Frequency 50 kHz. 15 ohm resistive load.

of the output transistors, it is necessary that the device with the higher gain shall be used in the position of Tr_1.

When distortion components were found prior to the onset of waveform clipping, these were almost wholly due to the presence of second harmonics.

Constructional notes

Amplifier. The components necessary for a 10 + 10 watt stereo amplifier pair can conveniently be assembled on a standard 'Lektrokit' 4 in. × 4¾ in. s.r.b.p. pin board, as shown in the photographs, with the four power transistors mounted on external heat sinks. Except where noted the values of components do not appear to be particularly critical, and 10% tolerance resistors can certainly be used without ill effect. The lowest noise levels will however be obtained with good quality components, and with carbon-film, or metal-oxide, resistors.

Power supply. A suggested form of power supply unit is shown in Fig. 9 (a). Since the current demand of the amplifier is substantially constant, a series transistor smoothing circuit can be used in which the power supply output voltage may be adjusted by choice of the base current input provided by the emitter follower Tr_2 and the potentiometer VR_1. With the values of reservoir capacitor shown in Table 3, the ripple level will be less

Table 2. Relation of distortion to gain-matching in the output stage.

	Current gain (Tr_1)	(Tr_2)	Distortion (at 9 watts)
1.	135	135	0.06%
2.	40	120	0.4%
3.	120	40	0.12% (pair 2 reversed in position)
4.	120	100	0.09%
5.	100	120	0.18% (pair 5 reversed)
6.	50	40	0.1%

Table 3. Power-supply components.

AMPZ$_L$	I_{OUT}	V_{OUT}	C_1	$Tr_{1/2}$	MR1	T_1	
15Ω	1A	37V	1000μ 50V	MJ480 2N697	5B05	40V	1A
2x 15Ω	2A	37V	5000μ 50V	MJ480 2N697	5B05	40V	2A
8Ω	1.25A	27V	2000μ 40V	MJ480 2N697	5B05	30V	1.25A
2x 8Ω	2.5A	27V	5000μ 40V	MJ480 2N697	5B05	30V	2.5A
3Ω	1.9A	18V	5000μ 30V	MJ480 2N697	5B05	20V	2A
2x 3Ω	3.8A	18V	10,000μ 30V	MJ480 2x2N697	7B05T	20V	4A

than 10 mV at the rated output current, provided that the current gain of the series transistors is greater than 40. For output currents up to 2.5 amps, the series transistors indicated will be adequate, provided that they are mounted on heat sinks appropriate to their loading.

However, at the current levels necessary for operation of the 3-ohm version of the amplifier as a stereo pair, a single MJ480 will no longer be adequate, and either a more suitable series transistor must be used, such as the Mullard BDY20, with for example as 2N1711 as Tr_2, or with a parallel connected arrangement as shown in Fig. 9(b).

The total resistance in the rectifier 'primary' circuit, including the transformer secondary winding, must not be less than 0.25 Ω. When the power supply, with or without an amplifier, is to be used with an r.f. amplifier-tuner unit, it may be necessary to add a 0.25 μF (160 V.w.) capacitor across the secondary winding of T_1 to prevent transient radiation. The rectifier diodes specified are International Rectifier potted-bridge types.

Transistor protection circuit

The current which flows in the output transistor chain (Tr_1, Tr_2) is determined by the potential across Tr_2, the values of R_1 and R_2, and the current gain and collector-base leakage current of Tr_2. Since both of these transistor characteristics are temperature dependent the output series current will increase somewhat with the temperature of Tr_2. If the amplifier is to be operated under conditions of high ambient temperature, or if for some reason it is not practicable to provide an adequate area of heat-sink for the output transistors, it will be desirable to provide some alternative means for the control of the output transistor circuit current. This can be done by means of the circuit shown in Fig. 10. In this, some proportion of the d.c. bias current to Tr_1 is shunted to the negative line through Tr_7, when the total current flowing causes the potential applied to the base of Tr_6 to exceed the turn-on value (about 0.5 volt). This allows very precise control of the series current without affecting the output power or distortion characteristics. The simpler arrangement whereby the current control potential for Tr_7 is obtained from a series resistor in the emitter circuit of Tr_1 leads, unfortunately, to a worsening of the distortion characteristics to about 0.15% at 8 watts, rising to about 0.3% at the onset of overload.

Performance under listening conditions

It would be convenient if the performance of an audio amplifier (or loudspeaker or any other similar piece of audio equipment) could be completely specified by frequency response and harmonic distortion characteristics. Unfortunately, it is not possible to simulate under laboratory conditions the complex loads or intricate waveform structures presented to the amplifier when a loudspeaker system is employed to reproduce the everyday sounds of speech and music; so that although the square wave and low-distortion sine wave oscillators, the oscilloscope, and the harmonic distortion analyser are valuable tools in the design of audio circuits, the ultimate test of the final design must be the critical judgment of the listener under the most carefully chosen conditions his facilities and environment allow.

The possession of a good standard of reference is a great help in comparative trials of this nature, and the author has been fortunate in the possession, for many years, of a carefully and expensively built 'Williamson' amplifier, the performance of which has proved, in listening trials, to equal or exceed, by greater or lesser margins, that of any other audio amplifier with which the author has been able to make comparisons.

However, in the past, when these tests were made for personal curiosity, and some few minutes could elapse in the transfer of input and output leads from one amplifier to the other, the comparative performance of some designs has been so close that the conclusion drawn was that there was really very little to choose between them. Some of the recent transistor power amplifier circuits gave a performance which seemed fully equal to that of the 'Williamson', at least so far as one could remember during the interval between one trial and the next. It was, however, appreciated that this did not really offer the best conditions for a proper appraisal of the more subtle differences in the performance of already good designs, so a changeover switch was arranged to transfer inputs and outputs between any chosen pair of amplifiers, and a total of six amplifier units was assembled, including the 'Williamson', and another popular valve unit, three class B transistor designs, including one of commercial origin, and the class A circuit described above. The frequency response, and total harmonic distortion characterisics, of the four transistor amplifiers was tested in the laboratory prior to this trial, and all were found to have a flat frequency response through the usable audio spectrum, coupled with low harmonic distortion content (the worst-case figure was 0.15%).

In view of these prior tests, it was not expected that there would be any significant difference in the audible performance

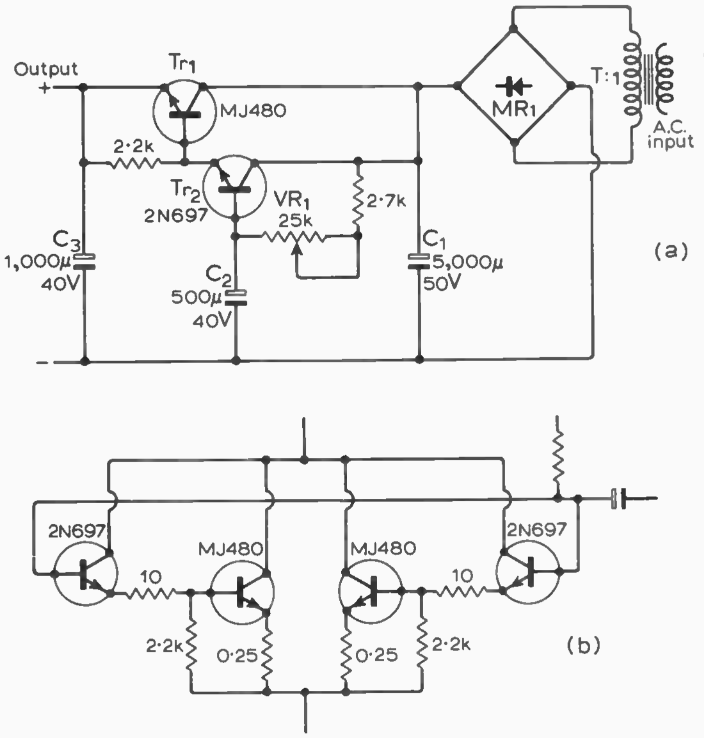


Fig. 9. (a) Power supply unit, and (b) parallel connected transistors for high currents.

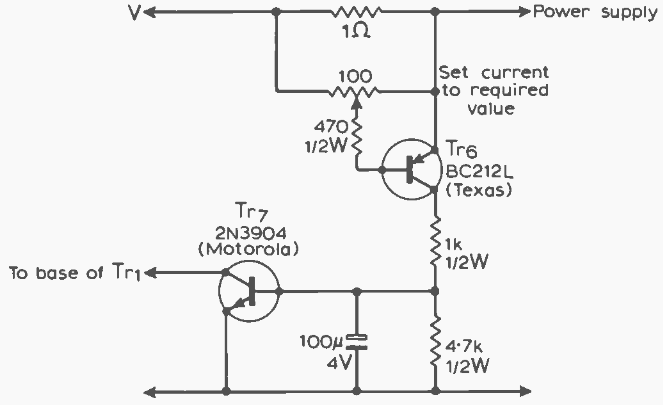


Fig. 10. Amplifier current regulation circuit.

of any of the transistor designs, or between them and the valve amplifiers. It was therefore surprising to discover, in the event, that there were discernible differences between the valve and the three class B transistor units. In fact, the two valve designs and the class A transistor circuit, and the three class B designs formed two tonally distinct groups, with closely similar characteristics within each group. The 'Williamson' and the present class A design were both better than the other valve amplifier, and so close in performance that it was almost impossible to tell which of the two was in use without looking at the switch position. In the upper reaches of the treble spectrum the transistor amplifier has perhaps a slight advantage.

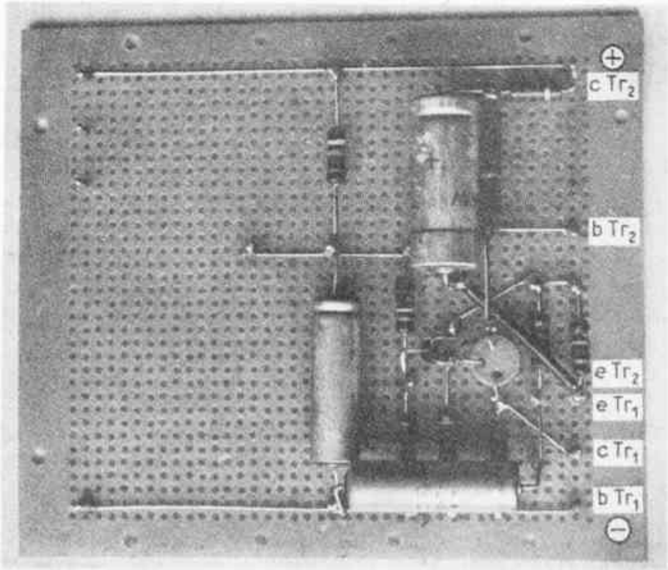
The performance differences between the class A and the class B groups were, however, much more prominent. Not only did the class A systems have a complete freedom from the slight 'edginess' found on some high string notes with all the class B units, but they appeared also to give a fuller, 'rounder', quality, the attractiveness of which to the author much outweighs the incidental inconvenience of the need for more substantial power supply equipment and more massive heat sinks.

Some thought, in discussions with interested friends, has been given to the implications of this unlooked-for discovery, and a tentative theory has been evolved which is offered for what it is worth. It is postulated that these tonal differences arise because the normal moving-coil loudspeaker, in its associated housing, can present a very complex reactive load at frequencies associated with structural resonances, and that this might provoke transient overshoot when used with a class B amplifier, when a point of inflection in the applied waveform chanced to coincide with the point of transistor cross-over, at which point, because of the abrupt change in the input parameters of the output transistors, the loop stability margins and output damping will be less good. In these circumstances, the desired function of the power-amplifier output circuit in damping out the cone-response irregularities of the speaker may be performed worse at the very places in the loudspeaker frequency-response curve where the damping is most needed.

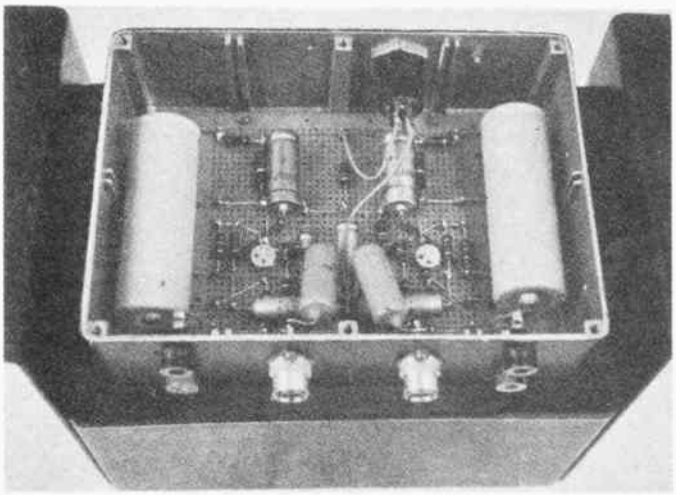
It should be emphasized that tbe differences observed in these experiments are small, and unlikely to be noticed except in direct side-by-side comparison. The perfectionist may, however, prefer class A to class B in transistor circuitry if he can get adequate output power for his needs that way.

Listener fatigue

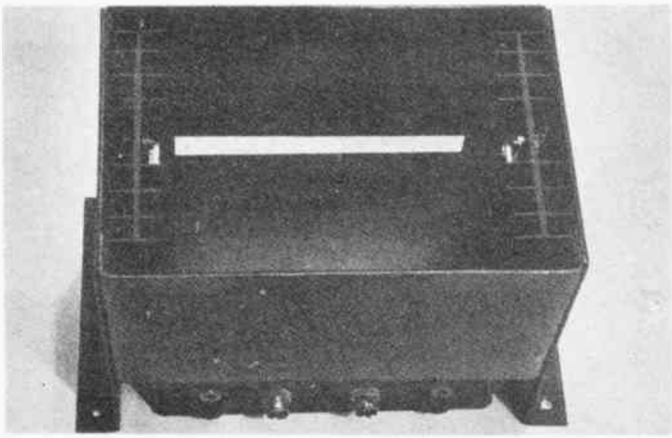
In the experience of the author, the performance of most well-designed audio power amplifiers is really very good, and the differences between one design and another are likely to be small in comparison with the differences between alternative loudspeaker systems, for example, and of the transistor designs so far encountered, not one could be considered as unpleasing to the ear. However, with the growing use of solid-state power amplifiers, puzzling tales of 'listener fatigue' have been heard among the *cognoscenti*, as something which all but the most expensive transistor amplifiers will cause the listener, in contradistinction with good valve-operated amplifiers. This seemed to be worth investigation, to discover whether there was any foundation for this allegation.



Layout of single channel of 10 + 10 watt amplifier on standard 4in × 4¾in 'Lektrokit' s.r.b.p. pin board.



Underside of completed amplifier, with base cover removed, showing external box-form heat sink.



Looking down on the completed amplifier.

In practice it was found that an amplifier with an impeccable performance on paper could be quite worrying to listen to under certain conditions. This appears to arise and be particularly associated with transistor power amplifiers because most of these are easily able to deliver large amounts of power at supersonic frequencies, which the speakers in a high quality system will endeavour to present to the listener. In this context it should be remembered that in an amplifier which has a flat power response from 30 Hz to 180 kHz, 90% of this power spectrum will be supersonic.

This unwanted output can arise in two ways. It can be because of wide spectrum 'white noise' from a preamplifier with a significant amount of hiss—this can happen if a valve preamplifier is mismatched into the few thousand ohms input impedance of a transistor power amplifier, and will also cause the system performance to be unnaturally lacking in bass. Trouble of this type can also arise if transient instability or high frequency 'ringing' occurs, for example when a reactive load is used with a class B amplifier having poor cross-over point stability.

Reference
1. Bailey, A. R., 'High-performance Transistor Amplifier', *Wireless World*, November 1966; '30-Watt High Fidelity Amplifier', May 1968; and 'Output Transistor Protection in A.F. Amplifiers', June 1968.

Letters to the Editor

Minimum standing current in class A

I read the article by Mr. J. L. Linsley Hood, in the April issue of *W.W.* with great interest, and the results he has achieved are certainly excellent for so simple a design. However I think he has done the cause of class A amplifiers a slight disservice, by the figures for minimum standing current (Table 1, p. 73) which are considerably above the true minimum figures, a fact which I have checked in practice.

The value of standing current required to just enable the output to swing to the points where one or other of the output transistors is bottomed depends on the desired output power and load impedance. These factors also dictate the minimumsupplyvoltage. These values of voltage are $V_{supply} = 2\sqrt{2 P_{out}.R_{load}} + 2 V_{ce\text{-}sat}$. (from $P_{out} = V^2/R$ where V is the r.m.s. output). Likewise the peak current may be

determined $I_{pk} = \sqrt{\dfrac{2P_{out}.R_{load}}{R_{load}}}$

which equals $\sqrt{\dfrac{2\,P_{out}}{R_{load}}}$

Now the standing current need be only half this value, since the change in load current is due to the upper transistor cutting off and the lower transistor doubling its current (or vice versa), a total change of current which must flow through the load of twice that originally flowing in the transistors, i.e. the standing current. Evaluating these for 10 watts and 8-ohm load we get (ignoring $V_{co\text{-}sat}$.) figures of 25.3 volts and 790 mA giving a total dissipation of exactly 20 watts, which is in agreement with the theory for such a 'perfect' stage (i.e. 50% efficiency). In practice we must add the figure of 2 $V_{ce\text{-}sat}$ as stated by Mr. Linsley Hood. We thus get figures of 26.5 V and 990 mA. The corresponding figures for a 10-watt output are 16.7 V at 1.29 A (3ohm), and 360 V at 580 mA (15 ohm). In practice we must also add an allowance to the current to provide for the variation which will take place due to temperature, etc. This excess current is especially necessary in the design described, where the only thing having any appreciable control of the standing current is the current gain of the output pair. I feel that this point was not sufficiently emphasized, or for that matter, the fact that to use any given pair of output transistors, and obtain the correct standing current needs quite a range of the total value of $R_1 + R_2$, from the values quoted of 100 + 560 ohms (for the 8-ohm case) for a high gain pair down to say 68 + 150 ohms for $H_{fe} = 40$. These lower values mean a considerable drop in loop gain, which probably accounts for much of the increased distortion, with lower gain transistors. These lower value resistors would in fact reduce the feedback by about 10 dB. If the circuit of Fig. 10 is used it will also be necessary to use lower values for R_1 and R_2 to allow for an excess current which may then be controlled. I therefore question the contention of page 74 that the circuit allows for 'precise control of the series current without affecting . . the distortion characteristics'. Any reduction of loop gain must increase the distortion.

I have just measured a very similar stage to that described and find that for an 8-ohm load and 10-watt output, a supply of around the 27 volts quoted and a standing current of 850 to 900 mA is ideal.

L. Nelson-Jones,
Bournemouth,
Hants.

The author replies:
I thank Mr. Nelson-Jones for his comments. Taking his second point first, it is evident that the current gains of the output transistors (particularly Tr_2) influence the standing current through the output chain. However, due to the flattening effect of operation at high junction currents and temperatures, the current variation from transistor to transistor with a given value of $R_1 + R_2$ is much less than the manufacturer's quote range of H_{FE} (30-200) would suggest. The tests which I made last year with limit-value devices gave a spread of ± 150 mA, when the current gains of the devices were badly matched, and rather less than this with limit-value matched pairs.

It had been in my mind at the time of writing the article that the constructor of the circuit should make adjustments to the value of R_2 (not R_1, which is part of the bootstrap circuit) to obtain the correct standing current, and made the comment (p. 72) that 'the resistor R_2 .... can be used to set the static current of the output stages'. The use of a variable resistor, in series with some suitable fixed value, would have facilitated the setting of this, and I had from time to time wished in retrospect that I had suggested this, as an alternative. Where this arrangement is adopted, however, care must be exercised in the layout of the leads to the potentiometer to avoid undesirable output-input feedback capacitances. The potentiometer should also be at the end of R_2 nearest to R_1.

With regard to Mr. Nelson-Jones's first point (about the correct standing current for a class A stage of this type) the calculations he shows are correct, and are substantially identical to those which I made myself in the initial stages of the design of this amplifier. However, in the particular case of a class A design of this type which cannot provide a load current which increases with demand, three further points must be considered.

1. The simple calculation of the ratio of peak-to-mean currents, as in the equations above, gives an answer which is valid only for symmetrical waveforms. Most of the waveforms in speech and music, for which such an amplifier will be used, are unsymmetrical and some allowance must be made for this.

2. The calculations assume that the load is resistive. In practice, loudspeakers present reactive loads, also their impedance may fall to lower than the nominal value.

3. The optimum performance of the output stage is given when the current swing does not take either transistor into current cutoff.

Because of these considerations, which were confirmed experimentally, I suggested a value of *quiescent* current which was in excess of the bare minimum 'sine wave into resistive load' value, even though this involves an increase in the thermal dissipation of the system. Safety factors can always be cut down—provided that one knows the circumstances.

J. L. Linsley Hood

New Approach to Class B Amplifier Design

*by Peter Blomley**

The class B amplifier has established itself as the most versatile and lowest cost amplifier known. This is mainly due to the excellent work in the field of semiconductor circuit design by H. C. Lin[1], R. C. Bowes[2], R. Tobey and J. Dinsdale[3], A. R. Bailey[4] and P. J. Baxandall[5]. In this article it is hoped to complement the work of these designers by putting forward a new approach† which may solve some of the problems inherent in present designs.

Conventional approach

A definition of a class B amplifier could be 'one in which the operating point of each output device is set at the lower extreme of its transfer characteristic. Hence in a push-pull design, for any symmetrical input signal, each output device conducts only one half of the output waveform'. This method of operation gives the amplifier zero (or nearly zero) quiescent power consumption, high efficiency and excellent peak current drive capability. It is unfortunate that the sacrifice paid for these virtues is the problem of ensuring a linear transfer of signal drive from one output device to the other.

So that the class B system can be analysed it is useful to approach the output circuit as two separate amplifiers (labelled X and Y in Fig. 1 for convenience), the outputs of which combine to give the complete signal. This is shown in diagrammatic form in Fig. 1, where it is assumed that the blocks representing the amplifiers form the equivalent of a complementary output stage.

The transfer characteristic for one of these 'sub-amplifiers' is shown in Fig. 2, where above the bias point A the characteristic is extremely linear and below it becomes a combination of linear and exponential relationships. The designer's task is to define this last region so accurately that when it is combined with that of the other sub-amplifier, the overall gain will remain *constant* (i.e. as the gain of one sub-amplifier decreases, the other increases equally to compensate).

The workings of a class B output circuit can be clarified by the use of g_m diagrams, these being a plot of gain—or in this case

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†The subject of patent application 53916/69.

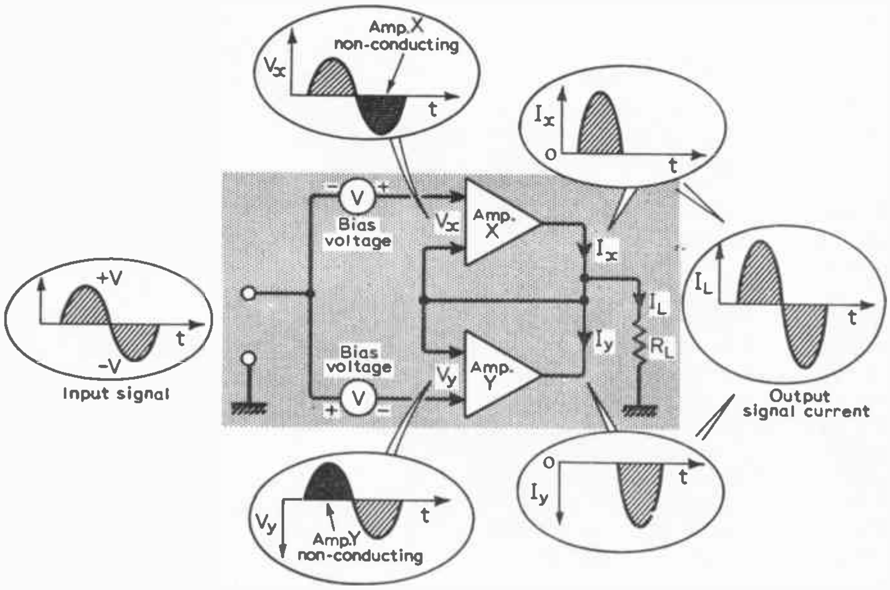

Fig.1. Block diagram of conventional class B amplifier with the two halves of a complementary output stage represented by sub-amplifiers X and Y.

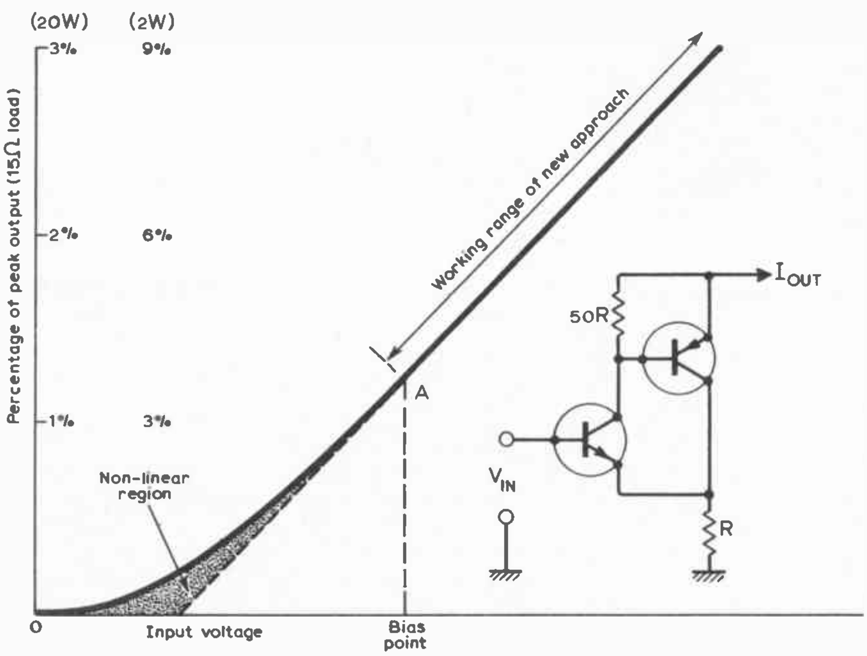

Fig.2. Transfer characteristic of sub-amplifier X which is linear above bias point A and non-linear below. In conventional class B, non-linear regions of X and Y sub-amplifiers have to be accurately matched to give overall linearity.

mutual conductance—of the complete output circuit against drive voltage. The ideal would, of course, be a straight line parallel to the input voltage axis (indicating there is no change of gain with input swing), but regrettably this is not the case with designs popular at the moment. To provide a comparison of the different types of output circuits, I have prepared gain plots showing the effects of different bias levels, these being illustrated in Figs. 3 and 4. From these it is now easy to see the characteristic change in gain which can occur during the transfer from one sub-amplifier to the other. Referring to Fig. 3 about a 10% gain change occurs during transfer, whatever the bias level is set at.

The output circuit in Fig. 4 is a quasi-complementary type giving most interesting results. The main conclusion is that it is impossible to bias this circuit for symmetrical gain change and in practice it proves very difficult to establish which biasing point would give the best results concerning the *rate* of gain change.

This method of describing a class B amplifier can give an insight into the problems involved with a conventional design. First, each sub-amplifier has to have two regions in its transfer characteristic:

—the constant gain region (above bias point A in Fig. 2)

—the non-linear region (below this point).

Second, the non-linear region of each sub-amplifier has to be complementary to its partner, otherwise the situation shown in the g_m diagrams (Figs. 3 & 4) will occur. An interesting point is that the only reason why the non-linear region of the transfer curve is important is because the input signal normally traverses this region as well as the linear portion. If this was not the case most of the design problems in class B amplifiers would be solved.

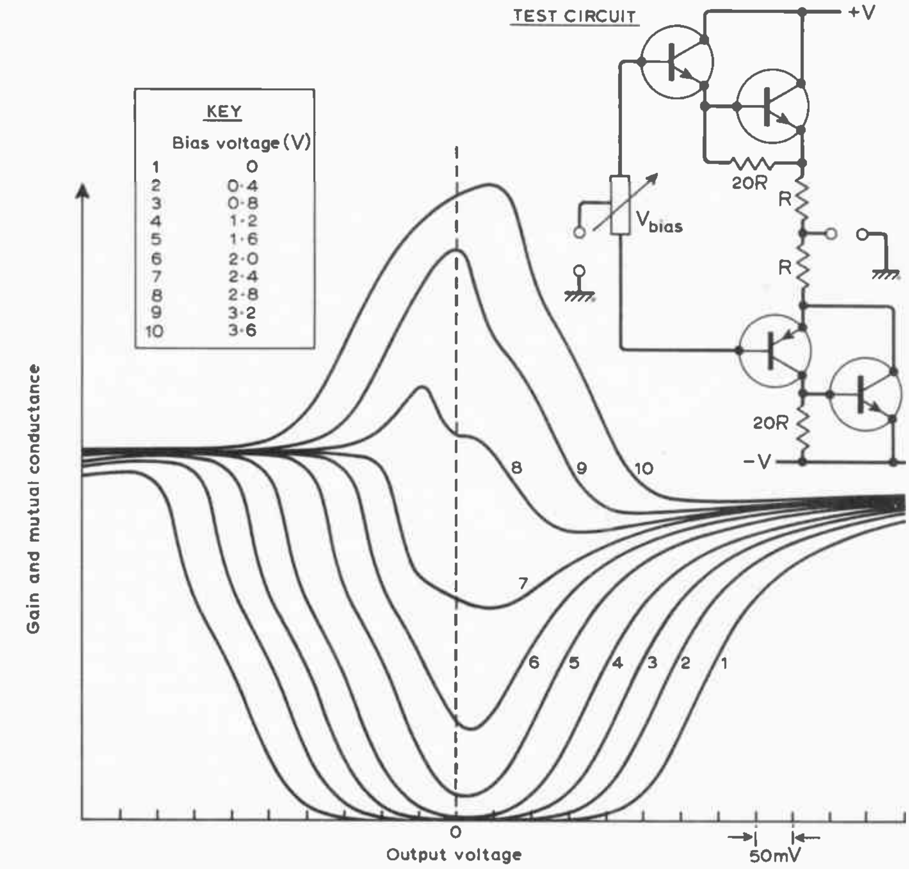

Fig.4. Curves for quasi-complementary output circuit show impossibility of biasing circuit for symmetrical gain change.

It is difficult to realize at first that a class B amplifier has to have this non-linear region in the sub-amplifier characteristic so that the two halves of the waveform can be separated. With conventional designs this is a built-in feature, but it need not be so. Assuming we define class A operation to include any amplifier where the input signal never traverses the non-linear region, the sub-amplifiers of a class B amplifier can operate in class A as long as the input signals are uni-directional. To accomplish this the required non-linear element is placed before the sub-amplifier inputs.

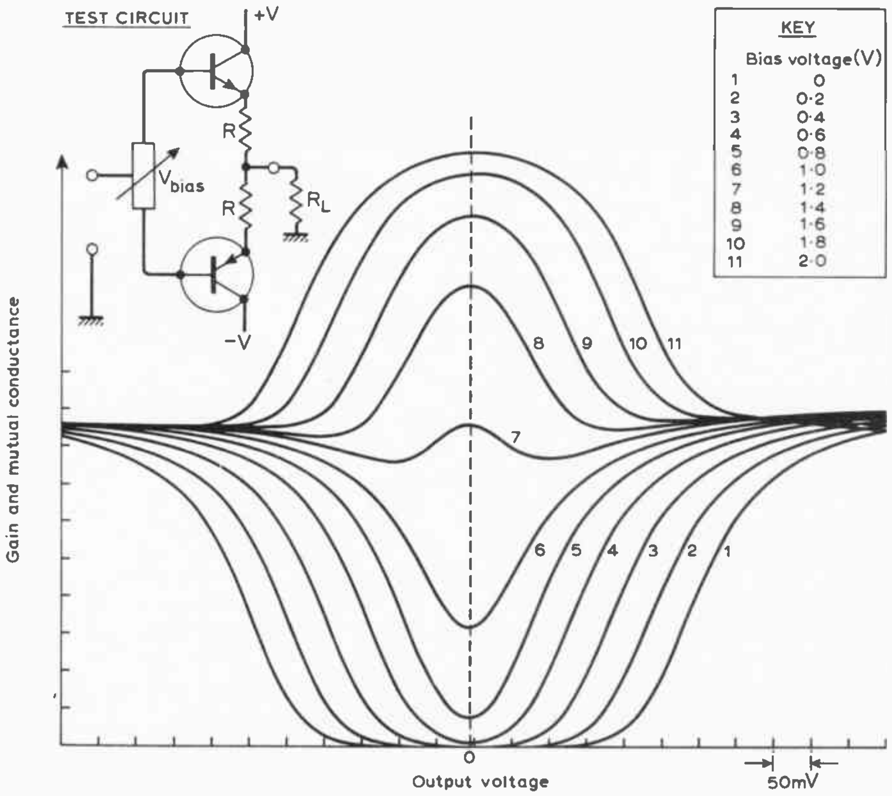

Fig.3. Gain—or mutual conductance—of simple symmetrical output circuit showing change in gain which can occur during transfer from one sub-amplifier to the other. Effect of different bias levels is shown.

New approach

Now the key to the problem is in the proposition that each sub-amplifier should be considered as a separate class A design, hence distortion generated by each of these units can be held to an extremely low level as long as the input signal can be prevented from driving the amplifier into the cut-off region. In the new approach, the output sub-amplifiers are biased *above* the non-linear region and *uni-directional* signals are fed into the input. This arrangement is illustrated in Fig. 5, where the necessary circuit changes are shown by comparison with Fig. 1. The obvious difference is the addition of the two diodes at the input which produce the uni-directional signal to drive the output sub-amplifiers. The linear transfer of signal between the two amplifiers is now dominated by this signal splitter.

Signal splitter. As the name implies the task of the signal splitter in a class B amplifier is to segregate the top and bottom halves of the signal waveform. Normally this is achieved by using the non-linear characteristics of each half of the output stage, but as this particular approach leads to problems, the new approach separates the two functions of amplifying and signal splitting completely.

To explain the problems involved with the design of a signal splitter it is usual to establish the ideal and see how this can be approached practically. As it happens there are two ideal 'half' characteristics which will give a linear cross-over when they are combined. The first, and obvious one, has a conduction path only in one direction and absolutely zero in the other. The other is more complicated and has three regions—linear region (large positive inputs), a non-linear region (transfer coefficient is proportional to signal) and a reverse region (transfer coefficient is zero).

The difficult region is the non-linear one. This will only give a linear crossover when it is combined with another conjugate characteristic. Not only this, but the relationship between the linear and non-linear portion has to be accurately defined. Normally this is achieved by altering the quiescent current in the signal splitter for minimum crossover distortion. Thus using this approach in the signal splitter means that the non-linear region has to be complementary to its partner and also that the linear and non-linear regions have to be accurately related. If additional constraints are imposed—due to device spreads and temperature changes—the situation can become very difficult unless a simple approach is used.

Returning to the first type of signal splitter, the immediate comparison which can be drawn is the simplicity of the characteristic. There are no interactions between each element and only one region has to be accurately defined. Ideally, therefore, this type of characteristic should be easy to control once a suitable device configuration is found.

Ideal element. The simple p-n diode fabricated in silicon can have a forward-to-reverse current ratio of 10^{10}; thus it approaches the ideal almost within the boundaries of measurement. This is however only considering the forward characteristic under conditions of *current drive.* If a voltage source were used the forward transfer would revert to the familiar exponential relationship between input voltage and output current (Fig. 6a). If a signal splitter is now made of two of these diodes and a *current* of changing direction fed into the common point, then from Kirchhoff's second law the current must flow either in diode D_1 or diode D_2 depending on the direction of signal current flow. The transfer coefficient for the diode must be unity, as it is only a two-terminal device, hence this type of signal splitter is extremely linear under the conditions of current drive (Fig. 6b).

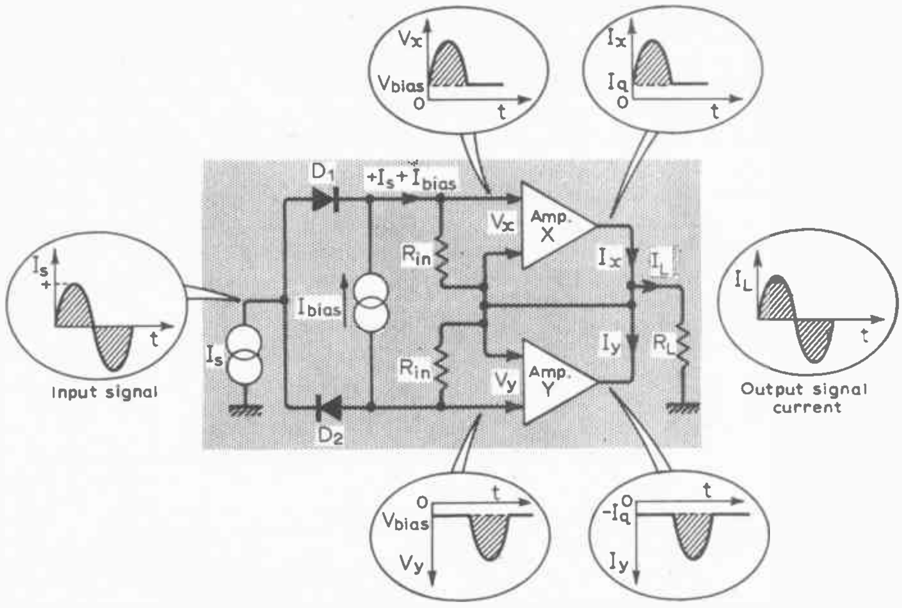

Fig.5. New approach to class B amplifier in which sub-amplifiers are biased above non-linear region and fed with uni-directional signals produced by the diodes. This effectively transfers signal splitting from the sub-amplifiers to a separate part of the circuit.

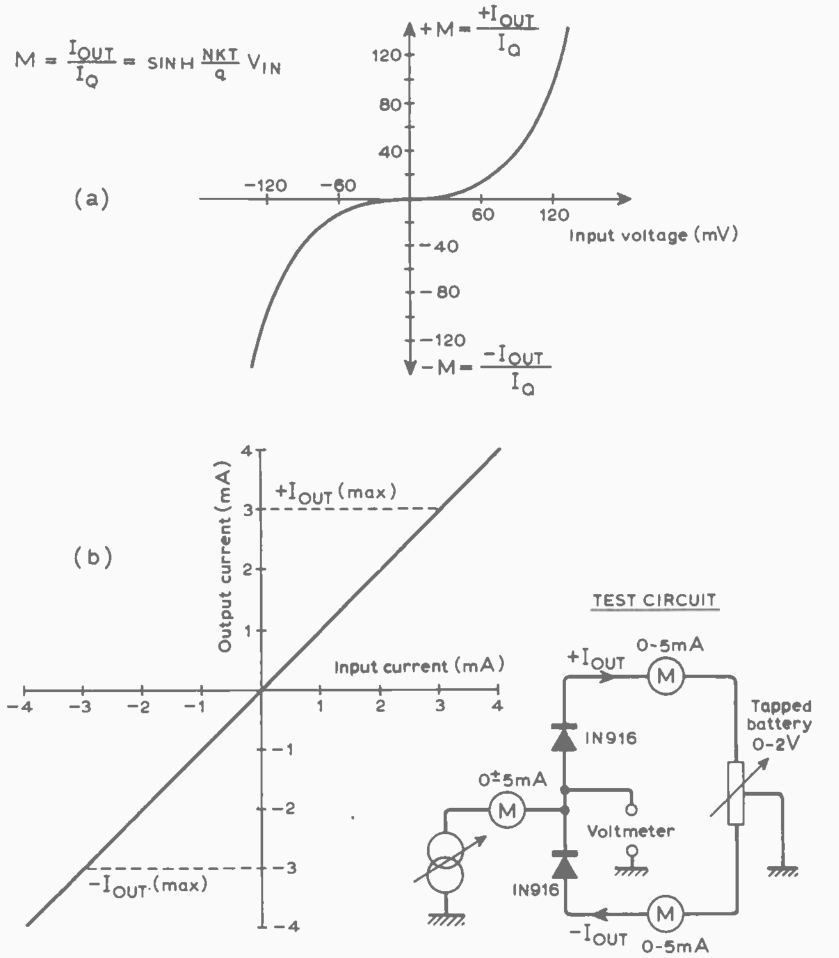

Fig.6(a). Transfer characteristic of voltage driven diode signal splitter.
Fig.6(b). Linear transfer characteristic of current driven diodes.

Transistor signal splitter. The use of a transistor as a signal splitter (Fig. 7) logically follows that of a p-n diode simply because the emitter-base junction has almost identical characteristics to that of a diode. Exactly where the transistor is superior to that of the diode depends on the design approach but in most cases the level-shifting property of a bipolar device is the main reason. This is very useful in a practical design but care has to be taken in the selection of the type of device. There is a problem with the use of transistors as signal splitters due to the emitter-base depletion capacitance. Under conditions of low injection this can add an additional phase lag during the crossover period. The problem can be overcome by using silicon planar devices with very high transition frequency (f_T) or by selecting devices in which the f_T is dominated by the diffusion capacitance as f_T remains constant down to very low emitter currents.

Synchronous signal splitter. There is a limit to the speed at which the diode or transistor signal splitter will transfer the signal path between the sub-amplifiers. If a synchronous signal splitter is used the time taken can be reduced to a few nanoseconds. This makes true class B operation possible at frequencies far higher than the audio spectrum. The system diagram is shown in Fig. 8(a). Instead of using the characteristics of the devices, as in the signal splitter which separates the two halves of the waveform, switches Tr_1 and Tr_2 are turned on and off at the required time by another amplifier labelled ST. This is a high-gain amplifier with a small amount of hysteresis, and as soon as the input exceeds a predetermined level the output from the trigger (amplifier ST) will change its polarity and turn on Tr_1 or Tr_2, depending on the signal direction, Fig. 8(b). This therefore gives almost the ideal signal splitting characteristics but the added complication might spoil its commercial possibilities.

Performance of the new design

The transistor signal splitter and the output stage circuit have a combined characteristic shown in Fig. 9 which demonstrates the excellent gain linearity. It is only when the bias of the sub-amplifiers is decreased below its optimum, allowing the output signal excursions to trace the non-linear region of the characteristic, that distortion begins to rise sharply. Further studies of these curves reveal that increasing the quiescent current through the output devices *does not* degrade, or for that matter improve, the crossover performance of the output circuit. Keeping this in mind it is therefore possible to design a class B amplifier *without any bias adjustment.* This assumes that the designer can guarantee that spreads in active devices and resistance values do not permit the quiescent current to fall below the level where the mutual conductance of the amplifier begins to decrease.

In this discussion about the performance of the design as a whole it would be fitting if the sub-amplifier design is mentioned. With conventional designs this two- or three-transistor element is fraught with compromises, one of the most serious being the decision on the inclusion of a base-emitter 'turn-off' resistance for the power transistor. Such a combination generates what can be called 'dead zone' distortion, mainly due to the change in slope of the transfer characteristic at zero crossing. One example of this is shown in Fig. 10 where, as predicted, the lower the value of resistance the more pronounced is the effect. It is very tempting to exclude this resistance altogether, especially if the current drive approach has been adopted, but the penalty would be a poor high-frequency performance coupled with overload recovery problems. This dilemma is aggravated if the designer decides to use homotaxial base powder devices (chosen for the robust nature of their construction and freedom from secondary breakdown) because the input diffusion and depletion capacitance is very high, hence the gain-bandwidth product of the device is relatively low (e.g. silicon planar $f_T \approx 90$MHz, homotaxial base $f_T \approx 1$MHz). In the latter case it is essential that the resistor is included. However, if the approach suggested in this article is adopted the sub-amplifier will never enter this non-linear region, thus the base-emitter turn off resistance can be included in the circuit to improve the performance without undue complications.

Once the decision has been taken to use the new approach the best circuit configuration has to be found and here again nature's swings and roundabouts create a difficult situation where compromise seems necessary. One of the criteria I used was that of thermal performance, following an initial consideration of the electrical properties of each configuration. The power transistor chip can change its temperature by tens of degrees centigrade during a power cycle, this being reflected by a corresponding change in the base-emitter voltage (V_{BE}) of the device. If the voltage bias to the sub-amplifier is applied directly to the power device (Fig. 11a), any change in the V_{BE} will cause a considerable change in quiescent current and in turn an

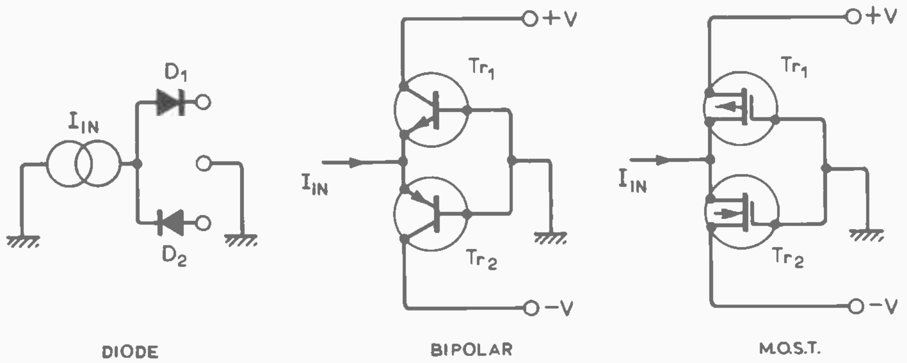

Fig.7. Types of signal splitter. Transistor type has the advantage of level shifting.

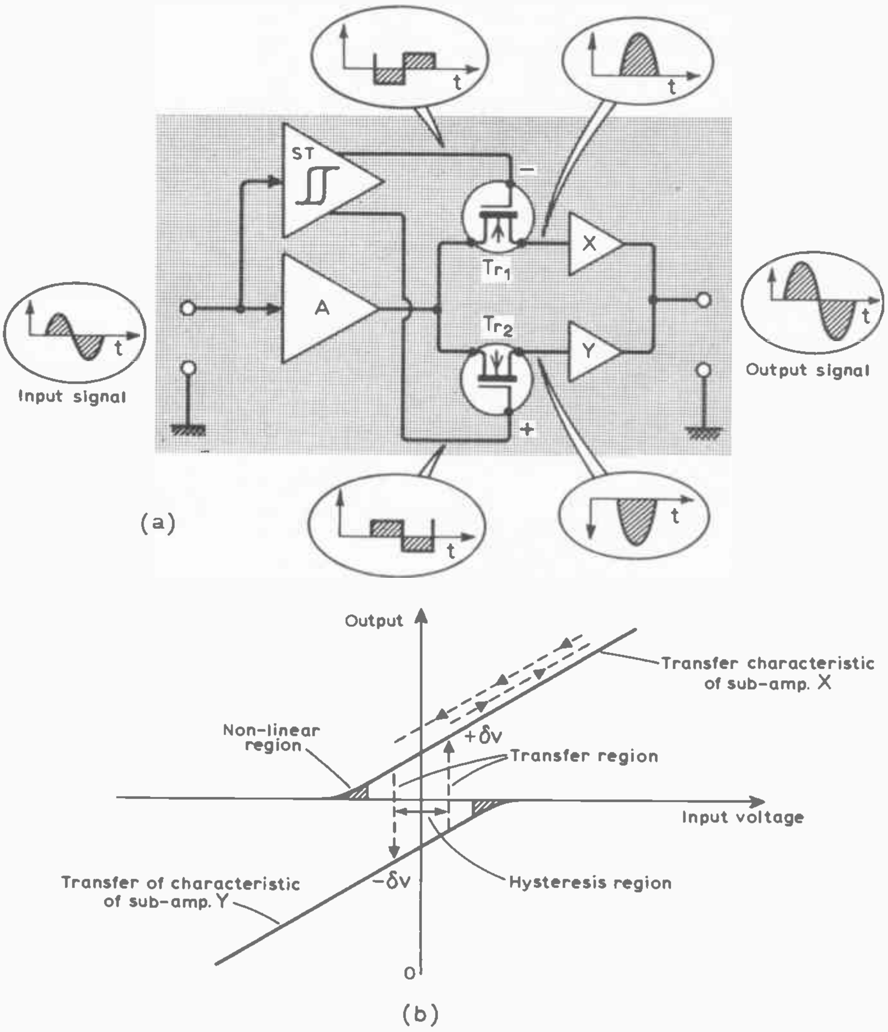

Fig.8(a). Synchronous signal splitter, with fast switching time, allows new approach to be used at frequencies well above audible range.

Fig.8(b). Operation of synchronous splitter of Fig.8(a). When input level exceeds a pre-determined level, output changes polarity and turns on Tr_1 or Tr_2.

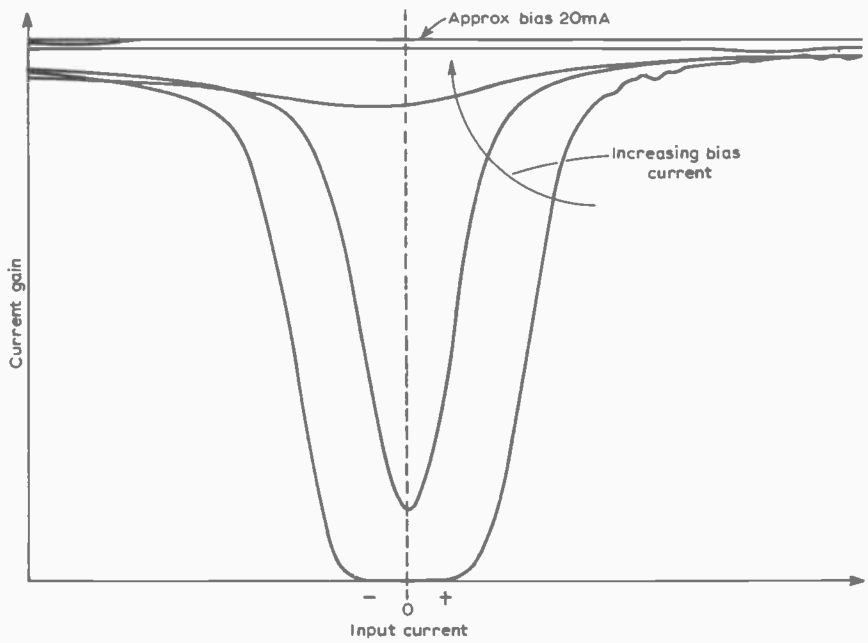

Fig.9. Combined characteristic of transistor splitter and output circuit shows excellent gain linearity.

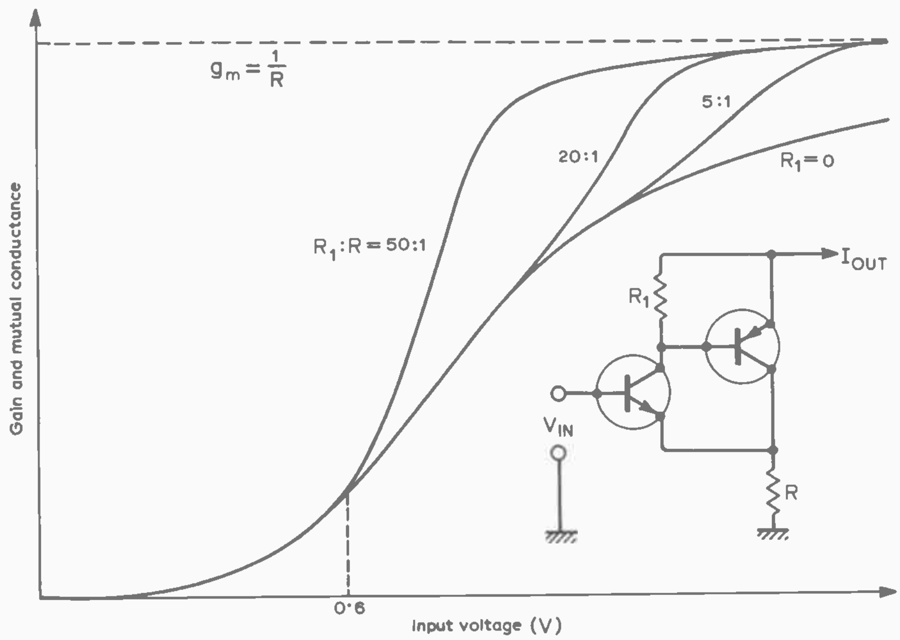

Fig.10. Transfer characteristic for conventional two-transistor sub-amplifier showing worsening effect of reducing the base-emitter 'turn-off' resistance of the power transistor. This normally generates 'dead zone' distortion due to the change in slope at zero crossing but is avoided in the new approach.

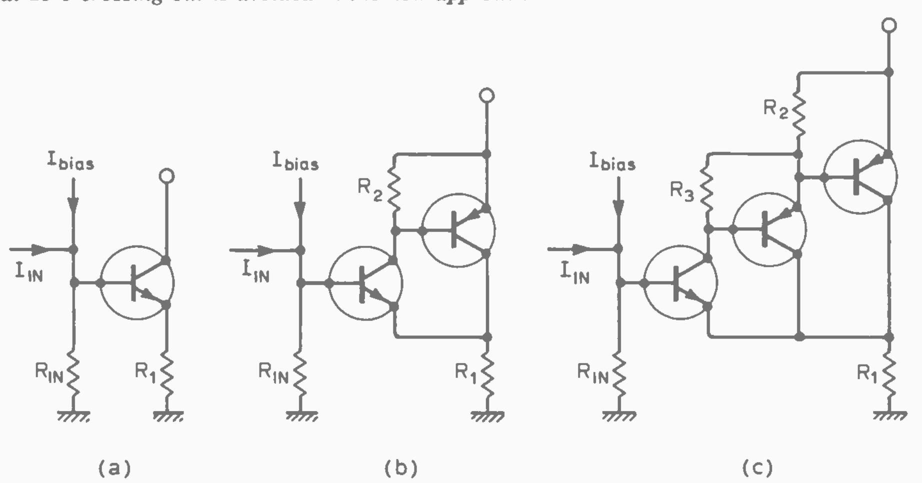

Fig.11. (a) Change in V_{BE} with temperature causes considerable change in bias current which could adversely affect intermodulation distortion. Circuits in (b) and (c) avoid this.

increase in distortion at low frequencies which could adversely affect the intermodulation performance of the amplifier as a whole. An improved design is shown in Fig. 11(b) and a more elegant version similar to that used in the Quad amplifier[6], in Fig. 11(c). It is on this latter example that I have concentrated most design effort, mainly because the performance advantages tend to outweigh the increased cost of pre-driver devices.

Returning now to an examination of the performance of the whole amplifier—the total distortion through the audio range can readily be made less than 0.1% *before applying feedback*, this performance being repeatable at almost any level of quiescent current.

Future designs

The use of class B amplifiers is not, of course, confined to the field of audio and in fact the ideas set out in this article lend themselves to applications in the high-frequency (> 1MHz) spectrum. The poor cross-modulation performance of present designs is usually due to the presence of non-linearities in the crossover region, hence substantial improvements can be expected in this direction.

Other applications where an ultra-low distortion amplifier of low stand-by power and high output capability is needed can be seen, examples of such devices being portable standard oscillators and meter calibration amplifiers. In the next article a practical design for a 30-watt audio amplifier is discussed in detail and future proposals developed in diagrammatic form.

(*To be concluded*)

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New Approach to Class B Amplifier Design

*by Peter Blomley**

This article describes a 30-watt amplifier design which embodies the author's approach to class B design, outlined last issue. Although further work on this approach is still needed, the design illustrates the kind of problems involved. The author also discusses the application of integrated components in future designs.

The general design of a complete amplifier using the new approach is relatively conventional except for the inclusion of the signal splitter (described last month). In principle, the design of each half of the output stage is made simpler as there is no cut-off, hence removing the necessity for predicting the performance in the cross-over region.

Examination of the circuit (Fig. 1) shows that the amplifier consists of three sections, the input amplifier, signal splitter and output amplifier.

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Input amplifier. This converts the input voltage into a proportional output current which drives the signal splitter. To enhance the performance of the amplifier as a whole, this section should have a reasonable mutual conductance (1A/V) and good linearity (1%). The latter does not represent a serious problem as the input amplifier is a low-level class A amplifier, but care is needed to control the maximum value of g_m otherwise frequency compensation problems arise.

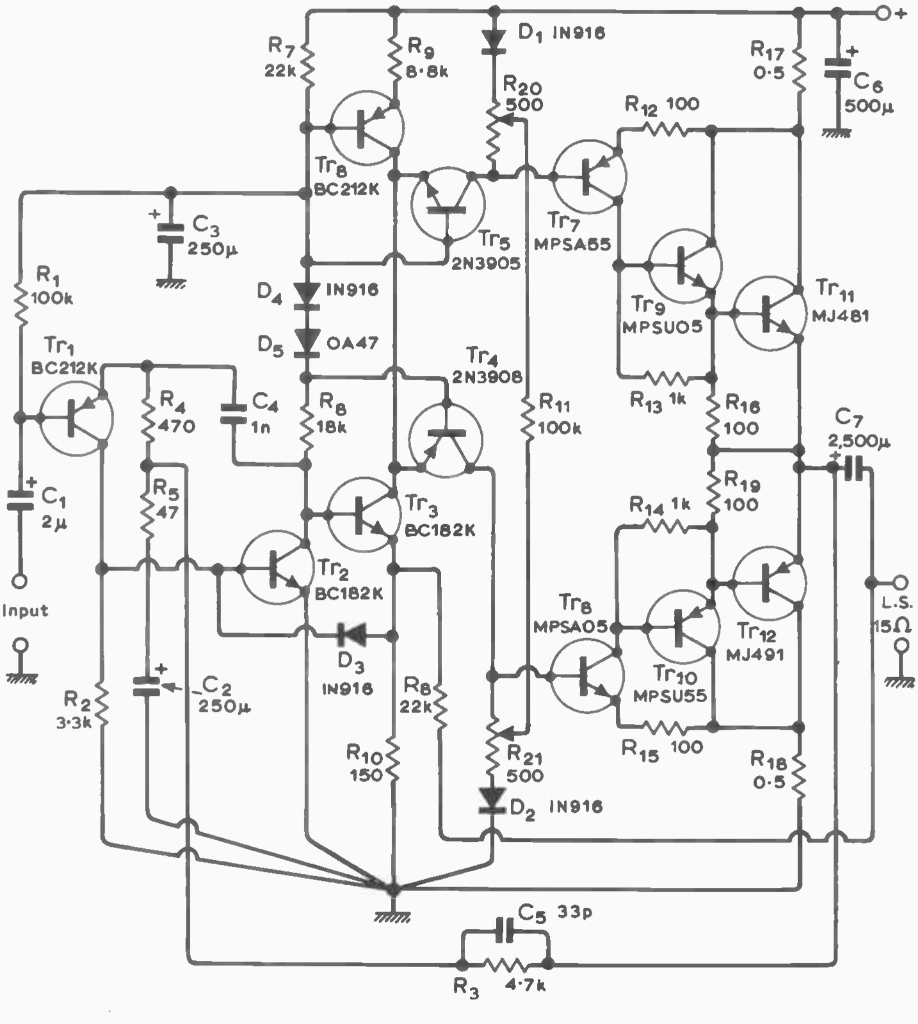


Fig. 1. Complete power amplifier circuit using new approach. Design gives harmonic distortion of 0.01% at all power levels and intermodulation distortion of 0.003%.

Signal splitter. As many fundamental details of the signal splitter were described last month, further details are confined to the biasing system. If perfect bipolar devices were available and ideal current sources existed, voltage bias across the emitter-base junction would not be needed, but such situations do not exist and distortions due to conditions falling short of the ideal can be rendered negligible by employing simple bias diodes (Fig. 2). This reduces the voltage excursion at the input to the signal splitter from 1.2V to 300mV pk-pk. The waveform with a sinusoidal output current is shown in Fig. 3.

Output stages. This now is one of the easiest to design. As long as the gain remains constant throughout the output cycle all is well. In the initial version, used to evaluate system performance, a compromise was reached between complexity, performance and cost. Thus individual adjustment potentiometers were used instead of the matched devices.

The output sub-amplifiers are similar to the Quad triples, these giving excellent linearity down to very low output currents, coupled with outstanding thermal stability. To compensate for the effect of ambient temperature changes on the quiescent current of the amplifier, diodes D_1 and D_2 cancel V_{BE} changes in transistors Tr_7 and Tr_8. It may have occurred to the reader that diodes in the forward path of the amplifier loop could generate appreciable distortion. However, in practice the maximum change in current is about 4:1 and thus almost corresponds to the change in collector current of transistors Tr_7 and Tr_8. In this way the change in voltage drop across the transistors compensates for the change in the diodes. Even if this did not occur, the resultant gain change for the output sub-amplifier is less than 4% for I_{out} values between 0 and 2A. The problem can be alleviated by increasing the current into diodes D_1 and D_2 and adding one resistor, but the advantages gained from this are negligible.

Circuit description

The function of Tr_1, Tr_2, and Tr_3 is to convert an error voltage—the difference between the input and feedback voltage—into a proportional output current. Now to produce the required mutual conductance of this stage (1A/V) without sacrificing either noise performance or linearity, the design in Fig. 1 was used. Starting at the input transistor Tr_1, this p-n-p type is used mainly as a level shifter. If we assume that the current gain of Tr_2 was extremely large (>500), then this input stage would have a maximum voltage gain of about five—not very much! If voltage gain was increased to the theoretical maximum of 30 (by decreasing the value of R_2 and R_4) problems would arise with the voltage offset at the speaker output due to increased emitter current flowing through R_3 and base current flowing through R_1.

Assuming for the moment that this first stage gain is a reasonable compromise, it now becomes obvious that the noise and distortion performance is dictated by the next stage. This stage (Tr_2, R_8) is a straightforward class A amplifier with very high gain (typically 400) and low distortion due to the limited modulation index of the collector current (0.04 max). The peak 2nd harmonic voltage generated is about 10μV and, assuming this is referred to the input of the first stage, it represents less than 0.001% 2nd harmonic distortion with feedback. Thus this second stage is the work horse of the input section, the third device Tr_3 being used both as a buffer to reduce the loading of R_{10} on R_8, and to convert the voltage changes across R_8 into an output current to drive the emitters of the signal splitter.

Resistor R_{10} performs two functions in this last stage of the input section. It defines the conversion constant *Engmen* for the stage, and it governs the maximum current which can be driven out of the collector of Tr_3. (This maximum current is defined by using the conducting voltages of D_3 and Tr_2 and the value of R_{10}.) Therefore this input section seems to have excellent performance during normal operation, but what can happen during an overload?

If the input transient was negative all would be well due to Tr_2 entering saturation. But if the transient was positive Tr_1 would turn off completely, the potential across R_{10} rising toward that at the end of R_8. (Tr_2 would also be completely cut off.) This would cause excess currents to flow in Tr_3, upsetting the bias chain R_7, D_4, D_5, R_8. After the excessive input signal is removed some time would elapse before recovery would take place, hence diode D_3 clamps the voltage and maintains Tr_2 in full conduction to reduce recovery time and improve amplifier stability.

While discussing the problem of recovery from overload, the charge across the compensation capacitor C_4 has also to be taken into account. The time for the accumulated charge to decay is a function of the amount of charge and the rate of decay. If the rate of decay is constant, the only way to reduce the recovery time is to limit the accumulated charge (in terms of voltage). Diode D_3 performs this function as well as clamping the voltage across R_{10} at 1V thus limiting drive current into the signal splitter.

The second section is the signal splitter, unique to this approach, and consists of transistors Tr_4 and Tr_5 plus a current source transistor Tr_6. The signal current into the emitter of Tr_4 or Tr_5 is derived by subtraction of two current levels, one constant and set by the voltage across R_9, and the other the output current of the input section. This signal current either appears at the collector of Tr_5—causing a voltage change across R_{20}—or at the collector of Tr_4—causing a voltage change across R_{21}. These voltage changes are converted into positive and negative output currents in the output section, which are then added together to give the final waveform. The current gain of the output sections which are conventional triples are governed by the ratio of R_{20} to R_{17} and R_{21} to R_{18}, and in this case the gain of 1000 seemed reasonable.

To keep the output triples above the minimum conduction level a bias current is provided by R_{11}. The procedure adopted for setting the standing current is to first set R_{20} and R_{21} to minimum (diode end).

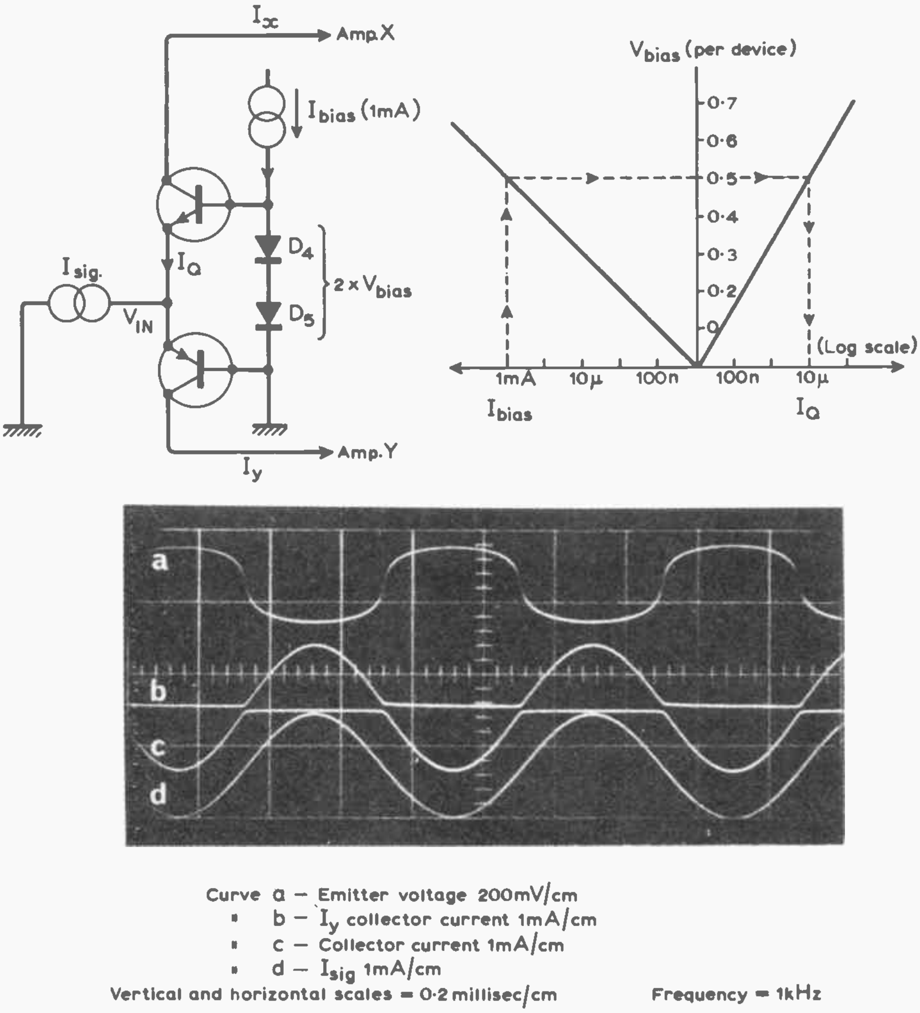


Fig. 2. Input amplifier converts signal voltage to a proportional current to feed transistor signal splitter. Bias diodes reduce voltage excursion from 1.2V to 300mV pk-pk. Bottom trace is current signal input to splitter.

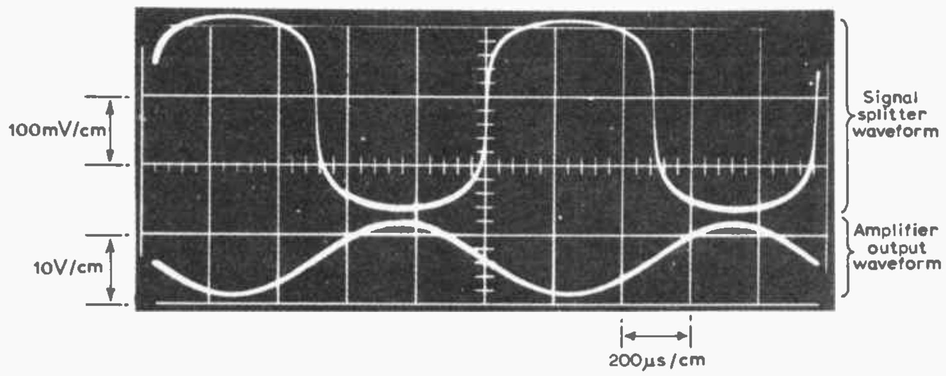


Fig. 3. Voltage excursion at signal splitter input with corresponding sinusoidal amplifier output current (R_L = 15Ω).

Set quiescent current with R_{20} and increase R_{21} until there is a small increase in current.

The only part still to be described is the biasing chain R_7-D_4-D_5-R_8-C_3. This provides the half supply voltage for the base of Tr_1 (decoupled by C_3), a load for the class A stage Tr_2, and sets devices Tr_4 and Tr_5 at the minimum conduction level required for good phase response during cross-over—by using the voltage across D_1 and D_2. By increasing the value of C_3 it is possible to reduce the rate of charging of the speaker coupling capacitor, eliminating 'thump', but capacitor size becomes very large.

Returning for a moment to the input section, Tr_2 is in a similar position to that used in many amplifiers, but instead of driving another stage (Tr_3) which only requires a limited voltage swing, it is the prime mover for the output section. To have sufficient drive capability the quiescent current in this stage may well need to be 10mA—instead of the 1mA in mine—and the voltage swing on the collector will be the full supply voltage (50 volts).

It now seems clear why the distortion of many amplifiers rises at low frequencies. The dissipation change of this device during a voltage cycle could be 500mW pk-pk in the case I have quoted giving an emitter-base voltage change at low frequencies of about 100mV. This change, even if we assumed it is basically a linear function of voltage, will cause a non-linear change in the input device and hence a considerable rise in distortion at low frequencies. In my amplifier the maximum dissipation change in Tr_2 will be less than 1mW, thus eliminating this form of distortion and improving intermodulation performance.

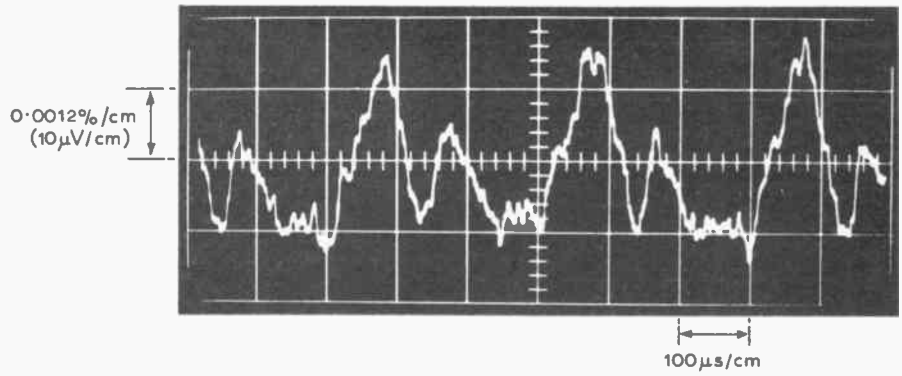


Fig. 5. Null method of assessing amplifier distortion shows distortion products to be well down in the noise. Deflection of 4cm represents 0.003% peak distortion at 10 watts (3kHz, 15Ω load).

Performance

The measurement of distortion created some difficulties especially when considering the range of frequencies over which this amplifier operates. The methods employed can be separated into two distinct techniques—spectrum analysis and nulling methods. To realize the first technique, an oscillator with a pure, single-line spectrum was needed, but the only one available at the time, approaching a reasonable performance, was the Si 451 produced by J. Sugden & Co, having a range up to 30kHz. This was found (excellent as it is) to be inadequate to permit the measurement of amplifier distortion.

So difficult in fact was the problem that it is impossible to publish distortion curves with any degree of confidence in their truth, but it can be said that using the Hewlett Packard 3590 wave analyser there was *no discernible difference* between a plot of the distortion of the oscillator and that taken after the oscillator output had been passed through the amplifier. Plots were taken over the frequency range 100Hz to 20kHz and powers of 100mW to 25W. As a matter of interest the spectrum plots of the amplifier are shown in Fig. 4 for 1kHz and 10kHz and at several power levels. The second method attempted was rather more successful but unfortunately does not present information in a usable form because it involves a comparison of output and input signals. It is also not a sequential test as in the previous method and as a result problems were encountered in successfully nulling the output against the input of the amplifier, due to the phasing of the signals and the earth loops generated by the measurement method. After considerable adjustment of the phase compensation and spurious pick-up difficulties the photograph Fig. 5 was obtained. Here the distortion generated is right in the noise (−120dB down from 20V r.m.s.) and the total deflection of 4cm represents 0.003% peak distortion at 10 watts and a frequency of 3kHz, chosen for easiest phase cancellation. The spikes usually evident in the difference waveform with this type of amplifier are completely absent, even with reactive loads, indicating that stability in the cross-over region must be excellent.

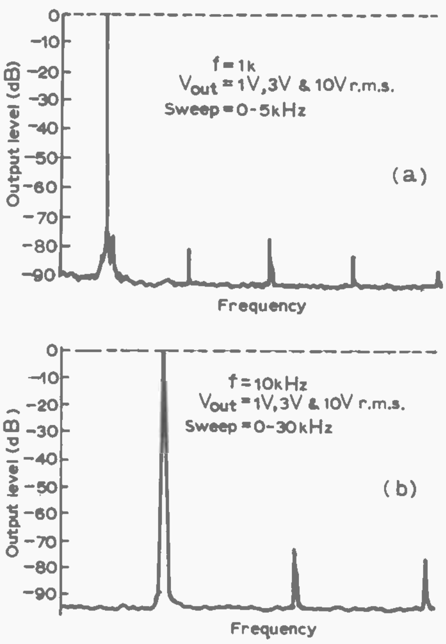


Fig. 4. Spectra made with a wave analyser showed no difference between spectra of outputs from oscillator and amplifier. Plots were made with (a) 1kHz and (b) 10kHz signals and were identical at all three power levels.

Intermodulation performance

The use of these two techniques is limited in one way or another to the evaluation of amplifier linearity. The main advantage is, of course, that a direct numerical value of distortion is obtained which can be used in comparison with other amplifiers.

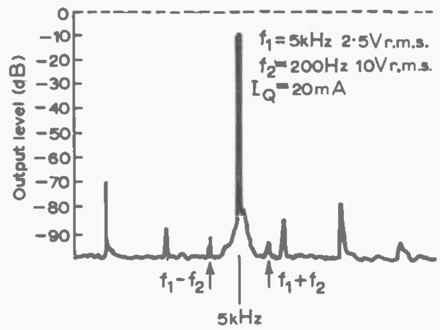


Fig. 6. Result of feeding 5kHz and 200Hz signals in a 16:1 power ratio into amplifier. Intermodulation products f_1+f_2 and f_1-f_2 are 90dB below 200 Hz signal. Other spectral lines are due to generator distortion.

The intermodulation test does not rely on low-distortion oscillators of signal cancelling techniques—in fact the only component which limits the measurement accuracy is the wave analyser itself. The real drawback is seen when an interpretation of the results is necessary! The method adopted is to "sweep" the transfer characteristic of the amplifier with a low-frequency signal of large amplitude, and to "measure" the slope of the characteristic with a low-level high-frequency signal. The two frequencies selected were 200Hz and 5kHz in a power ratio of 16:1.

The results not only ease the assessment of the amplifier performance in an absolute sense but also give some form of subjective measurement for comparison with other elements in the system. The results obtained in Fig. 6 indicate an excellent performance, the intermodulation products $f_1 + f_2$ and $f_1 - f_2$ are −90dB below the sweeping signal (200Hz) all other spectral lines being due to generator distortion.

Amplitude-frequency response

The type of frequency compensation used for this amplifier is unusual, mainly as a result of the system design. The open-loop gain begins to fall off at about 4kHz and continues on a 6dB/octave roll-off to about

500kHz where the second pole of the output section starts to contribute excess phase shift. The choice of the position of the dominant compensation was a difficult one. If it was placed in the output section, as is normally the case, the gain of the input amplifier would have to be restricted at low frequencies, affecting the distortion performance of the amplifier.

Another choice was using the dominant lag to encompass the output section as well as part of the input amplifier. This would lead to instability internal to the loop enclosed by the dominant lag and thus an internal pole would have to be introduced to remedy this condition. The final choice (shown in Fig. 7) gives the single-pole compensation needed for unconditional stability coupled with minimal high-frequency distortion. The inherent pole in the output section is subdued by the feedback resistance R_3 (so far as the main loop is concerned) but gives the required unconditional stability of the output section.

The performance with reactive loads will be spoilt if the output impedance of the amplifier is controlled by the overall feedback loop, i.e

$$Z_{out} = \left(1 + j\frac{f_1}{f_2}\right) \Big/ g_m$$

where f_1 is the signal frequency and f_2 the open-loop -3dB frequency. This expression has a simple analogy with a series inductance and resistance, where $R = 1/g_m$ and $L = 1/2\pi g_m f_2$.

A little more work† shows that if a capacitive load is used the amplifier would have a response given by

$$G = \frac{1}{p^2T^2 + a\,pT + 1}$$

This is the equation of a second-order system, where $a \int (1/g_m) \sqrt{(C/L)}$, and the natural frequency of oscillation is $w_o = 1/T = 1/\sqrt{(LC)}$. If the amplifier has an overshoot it must be due to the overall amplifier having an a-value approaching zero. If we now assume typical values and examine the worst case condition, $g_m = 10$A/V, $f_2 = 4$kHz and $a = 0.1$ (20dB peak), then $C = 4\mu$F and $w_o = 250$kHz.

If this was a perfect model for the amplifier the overshoot would be excessive, but in practice the output impedance is not only a function of frequency but also of output current. Thus a gets larger (less overshoot) as the output current increases. The basic assumption of this simple analysis is that the output impedance is controlled by the main feedback loop, but in this amplifier resistor R_6 generates another loop which effectively places a damping resistance across the apparent output inductance (Fig. 8).

The only remaining improvement to the transient performance of the amplifier is by pole-zero cancellation using the feedback element. If this term seems somewhat academic, an alternative is to study the overshoot with a second-order system with various inputs. If the input is an ideal step the amplifier will give theoretical overshoots, but if the rate of rise of the input waveform is *decreased* the overshoot will reduce and eventually disappear. The capacitor (a zero) in the feedback loop is really reducing the maximum rate of change of the voltage across the load and hence the degree of excitation given to this inherently oscillatory system. By using this type of compensation excellent performance with reactive loads has been finally achieved (Fig. 9). The overshoot with capacitative loads, such as 4μF, is about 50% with an ideal step input and far less when fed via a preamplifier, thus no difficulties should be experienced with any normal load.

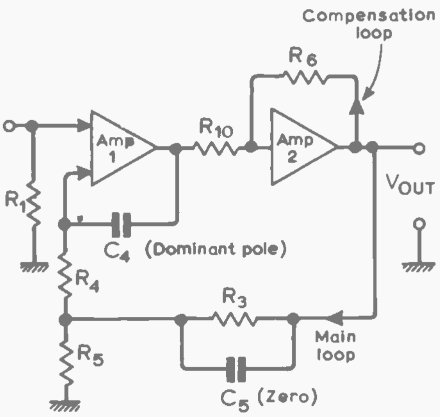
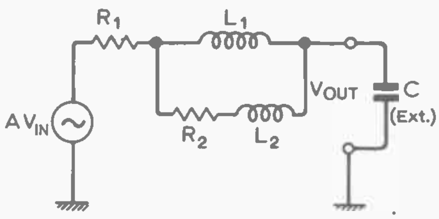


Fig. 7. Single-pole frequency compensation method used gives unconditional stability coupled with minimal h.f. distortion.



$$R_1 = \frac{1}{g_{m1}}\left(1 + \exp\left(K.\frac{I_{OUT}}{I_{MAX}}\right)\right)$$

$$L_1 = \frac{1}{2\pi f_2\, g_{m1}}$$

$$L_2 = \frac{1}{2\pi f_3\, g_{m2}}$$

$$R_2 = \frac{1}{g_{m2}}$$

Fig. 8. Power amplifier equivalent circuit. Simple analysis shows output impedance is controlled by main feedback loop, but in practice R_6 generates another loop effectively placing a damping resistance across the apparent output inductance.

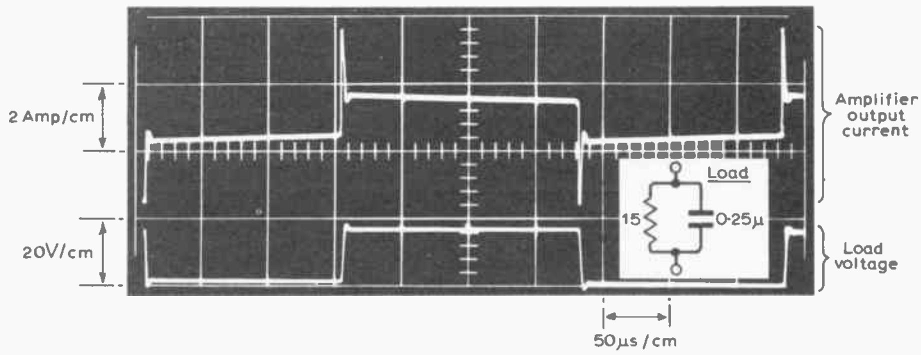


Fig. 9. Performance with a capacitative load. Capacitor in feedback loop effectively reduces maximum rate of change of voltage across load. Overshoot is much less when fed from a pre-amplifier.

Performance—with 60V regulated supply

output power	20 watts into 15 ohms 30 watts into 8 ohms
power response	30Hz to 100kHz (−3dB)
output impedance	0.1 ohm at 1kHz
total harmonic distortion	< 0.01% throughout audio band and all power levels
intermodulate distortion	< 0.003%
voltage gain	100
noise level	−120dB below full power
maximum peak output current	±3 amps, approx.

Electrostatic loads. The distortion characteristic with this type of load was still insignificant below 10kHz and gave a gradual rise up to 20kHz where it was still less than 0.05% at maximum output‡. Square-wave performance is shown in Fig. 10 at maximum ‡output. The ringing is due to the finite output impedance converting the ringing current in the inductance and capacitance of the load into ripples in the output, plus the overshoot of the amplifier itself.

Future developments

The amplifier design is hopefully only a source of ideas which may encourage further research into the whole approach to design. So that the trend may be continued, future proposals are outlined in Fig. 11. Here, the main difference is that

† See for instance "Active filters" F. E. J. Girling and E. F. Good, *Wireless World*, vol. 75, Sept. 1969, pp. 403-8.

‡ Maximum output is dictated by peak current output capability.

the output subamplifiers are oriented toward the use of integrated components. It has become obvious that past problems with class B amplifiers originated with the stabilization of the quiescent current to give zero cross-over distortion. Attempts were made to use diodes to compensate for device V_{BE} changes with fluctuations in the *ambient* temperature—the independent variations due to device dissipation could not be eliminated. Most of the time the diode did its job and the voltage defined by the combination of transistor and diode remained constant. This constant voltage was used in conjunction with low-value resistors to set the quiescent current in the output circuits.

If now an integrated component is used both the diode and the transistor are on the same chip and, apart from minor fluctuations, the combination is isothermal. As a result the quiescent current is a function *only* of the setting voltage and not ambient temperature or differential device temperatures. The accuracy with which the current can be set is largely governed by the offset voltage of the transistor pair. Typical values of ± 4mV which would represent a ± 8mA inaccuracy in the quiescent current using 0.5-ohm feedback resistors are readily obtained. With such an arrangement a reasonable quiescent current for the sub-amplifiers would be 30mA, the worst case figures would be 24mA and 38mA. Both of these values are well above the low conductance current level (5mA) which is required for good linearity of the sub-amplifiers.

The advantage of the new approach is fairly evident when it is realized that as long as the amplifiers are above the non-linear region, the spreads introduced in the sub-amplifier quiescent current will not cause the class AB situation of over-biasing (shown last month) characteristic of present designs. It is now possible to design an output stage without the normal trim potentiometers, thus giving a degree of freedom in production not possible with current amplifiers. The performance of the amplifier, once checked at the end of a production line can be guaranteed for operation in any climate and for any period of time.

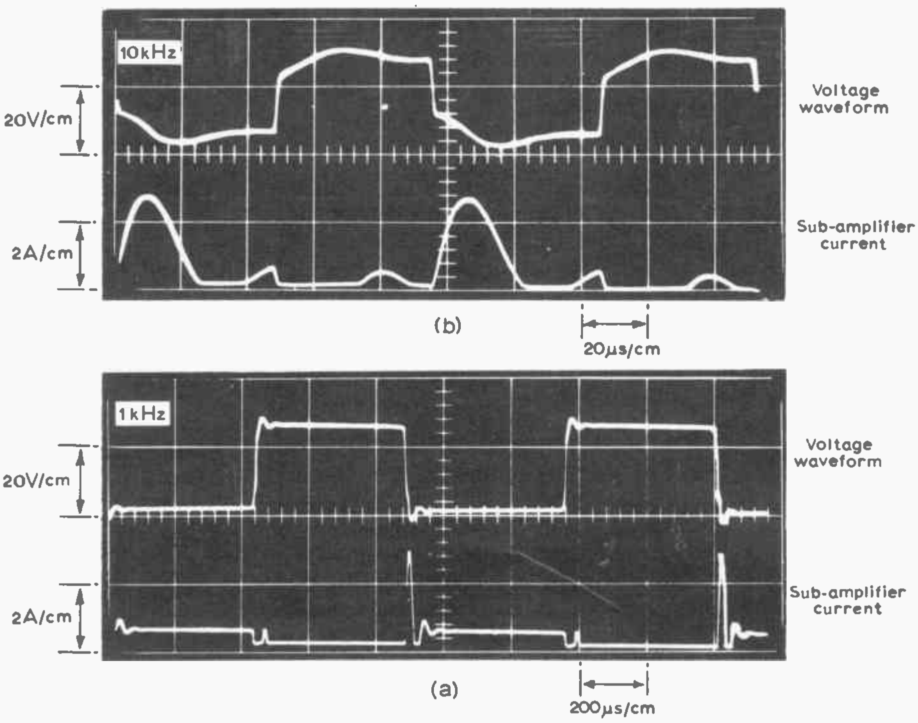


Fig. 10. Square-wave performance when driving electrostatic load at 1kHz (a) and 10kHz (b). Top traces are voltage and lower traces current out of sub-amplifier. Ringing is due to output impedance converting ringing current in L_2 and C_2 into ripples in the output.

Possible applications

The performance of an amplifier of this calibre is, in my opinion, wasted in a conventional audio set-up. In most cases, the transducers will be the weakest link.

The approach used in the design of the output sub-amplifiers does not rely on complementary matched devices—in fact, in most cases n-p-n devices are preferred for their superior secondary breakdown characteristics. This represents considerable reduction in amplifier costs especially in the 100-watt region as presently available devices boast a V_{CEO} of 120V with 100 watts dissipation at a cost of less than 75p.

The ultimate use for this amplifier would appear to lie with the high-power professional market where the performance of cascaded amplifiers in a system would have to be excellent. Use in other fields would be mainly governed by the expected gain in performance or reduction in cost. A possible application would be as a portable standard oscillator, perhaps meter calibration amplifiers, or even high-frequency low-distortion class B transmitter amplifiers. However, these are only inspired guesses which may interest those working in these relevant fields.

Thanks are due to Peter J. Baxandall for his advice and encouragement and to Hewlett Packard and the Plessey Co. for use of their facilities.

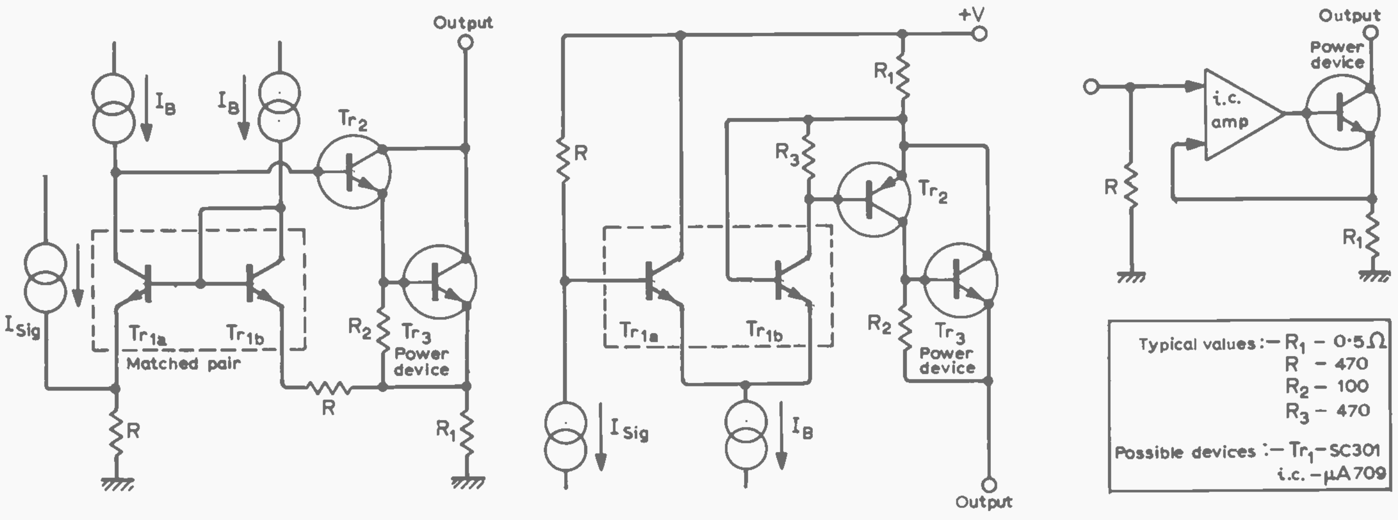


Fig. 11. Proposals for integrated components in output sub-amplifier.

A Non-resonant Loudspeaker Enclosure Design

Using acoustic transmission line with low-pass filter characteristics

By A. R. Bailey M.Sc.(Eng.), Ph.D., A.M.I.E.E.

Over the years, the design of loudspeaker units has progressed steadily until some are now available with very good performance capabilities. In particular, the advent of expanded polystyrene as a cone material has greatly reduced the distortions due to cone break-up.

Unfortunately, the design of loudspeaker cabinets has not kept pace with these developments, and there is little doubt that many enclosures now introduce more coloration than that produced by good loudspeakers. The loudspeaker enclosure to be described was developed to give as little coloration as possible, but to understand its evolution it is necessary to return to basic principles.

By far the largest number of current loudspeaker cabinets are based on the "bass reflex" cabinet design[1]. This is shown in Fig. 1 as a sectional view. This cabinet appears to have only a relatively short path-length between the back and the front of the cone and would therefore be expected to give relatively poor l.f. response. In fact the response at low frequencies can be quite large, this being due to the cabinet having a Helmholz air-column resonance at about the lowest frequency being reproduced. In the reference given previously, the theory is well explained.

Unfortunately, this resonance in the cabinet causes coloration in the bass output of the system. This "ringing" on bass transients can be very noticeable, particularly on such instruments as string bass.

In addition the cabinet itself is frequently unlagged and consequently the sound is still emerging long after the original signal has stopped. Very heavy lagging is necessary to stop this effect and in so doing the cabinet "Q" is reduced so that the bass reflex action is lost.

Sine-wave testing does not necessarily show up the defects of a speaker system. Rapid cut-off at the edges of a flat response can make it sound far worse than a slower rate of fall at the edges of a system with apparently a poorer bandwidth. This is shown in Fig. 2.

The effect of the abrupt change in slope of the amplitude/frequency characteristic is to give "ringing" at the frequency where the slope change takes place. This effect is unavoidable and is the necessary price to pay for the extension of bandwidth by the use of resonance effects. It is for this reason that loudspeaker systems can sound very "boomy" in the bass, even though the measured amplitude response shows no resonant peaks.

One method of testing that has not apparently been widely used is that of impulse-testing of loudspeaker cabinets. This method is very powerful and is described later on. For the moment it is sufficient to state that it confirmed that normal loudspeaker enclosures are not very good.

* Bradford Institute of Technology.

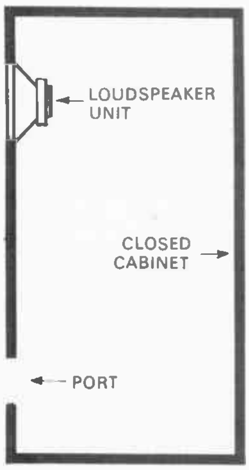


Fig. 1. Section through bass reflex cabinet.

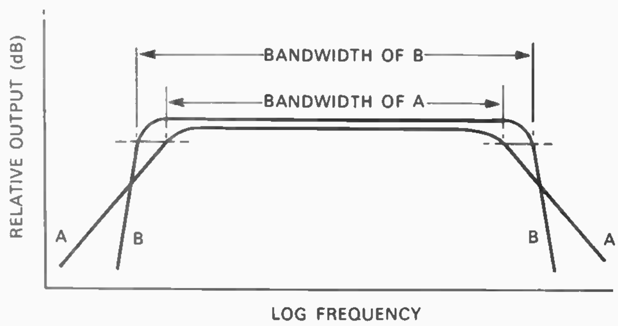


Fig. 2. Response curves having different transient responses.

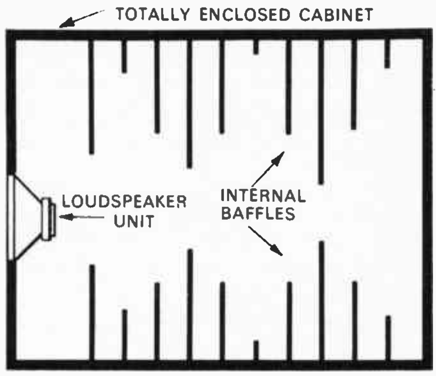


Fig. 3. Section through acoustic labyrinth cabinet.

It is now apparent that it is the sound waves produced at the rear of the cone that have to be absorbed if delayed output and resonances are to be avoided.

Acoustic labyrinths[2] have been used in the past in an attempt to "lose" the sound down multiple paths. Such an enclosure is shown in Fig. 3, but the size needed is excessive. Unless there is adequate internal lagging, then these cabinets will also possess pronounced energy storage and the consequent lack of sound clarity.

Transmission line approach

The only safe method of removing the rear cone sound energy is by transmitting it down an infinite transmission line. This is obviously impracticable so the nearest approximation was examined.

If a transmission line for acoustic waves is filled with a light acoustically-resistive medium, then the transmitted wave will be rapidly attenuated. After a certain distance the wave will be so weak that a blockage of the line will give only a minute reflected wave back to the speaker cone. The magnitude of the reflected wave can be determined by the standing-wave ratio immediately behind the speaker.

The choice of acoustic absorbing material will greatly affect the final performance so an initial investigation was made into the reflecting properties of various acoustic transmission line terminations. Many materials including glass fibre were tried, but the best results were obtained from long fibre wool. This gave a remarkably good absorption down to quite low frequencies. A typical set of the curves obtained is shown in Fig. 4.

For practical purposes a long pipe is not desirable so experiments were then made with a folded pipe of total length of about 8ft. This was built as shown in Fig. 5, the far end of the pipe having a pivoted flap so that characteristics could be taken with both open and short-circuit terminations. The results proved most interesting.

With the port closed the loudspeaker gave a very good performance with a slightly weak bass response. Sine-wave testing confirmed that the bass response fell by several dB below 60 c/s although the cone amplitude doubled for a halving of the input frequency. The trouble was finally traced to diffraction effects, the radiated wave-front changing its polar response at low frequencies. The effective bass response could be changed very markedly by positioning the cabinet away from a wall. The bass response then fell even further due to the increased diffraction at low frequencies. For test purposes a plain wall backing was used.

Opening the port had two effects. First, the bass response was improved to become approximately flat and secondly the cone excursion was greatly reduced between 30 and 50 c/s. The bass improvement was due to the line length being such that the delayed bass wave from the line was in phase with that radiated by the front of the cone. Also as the bass frequencies were radiated from two spaced sources, the diffraction effects would be reduced.

As the wool-filled line acts as a low-pass filter, the radiation from the vent cuts off before cancellation can occur at the higher frequencies. The rapid cut-off of this acoustic line is shown in Fig. 6. This shows the sound pressure at the port end of the line with the port closed.

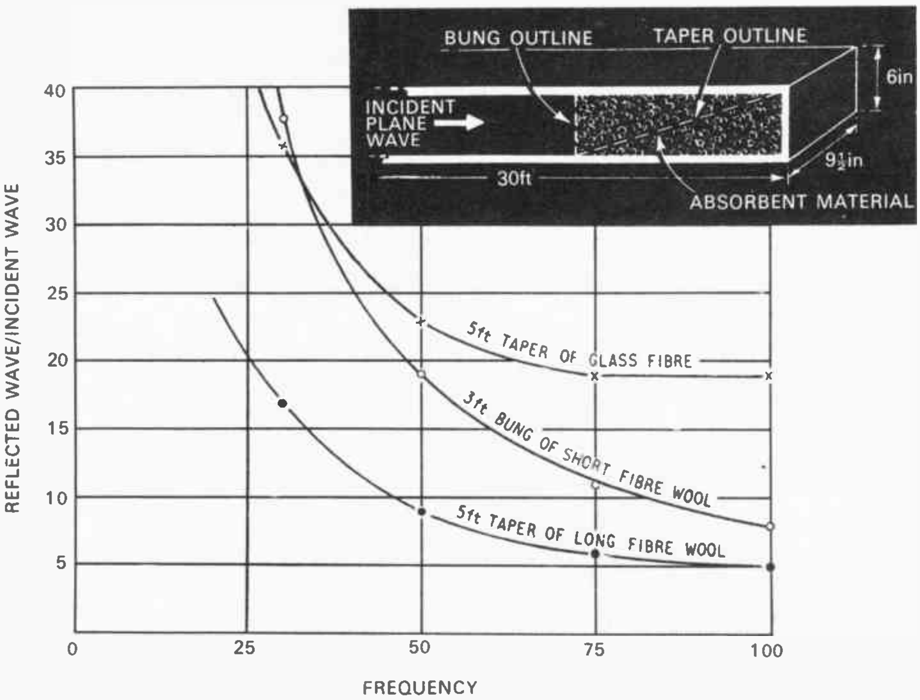


Fig. 4. Reflection characteristics of acoustic absorbents.

Impulse response

As the performance so far appeared to be satisfactory it was decided to investigate the impulse response of the loudspeaker cabinet. The square-wave testing of loudspeaker units had previously shown that it was not possible to generate a good square-wave of sound pressure, let alone an impulse. Several mechanical methods

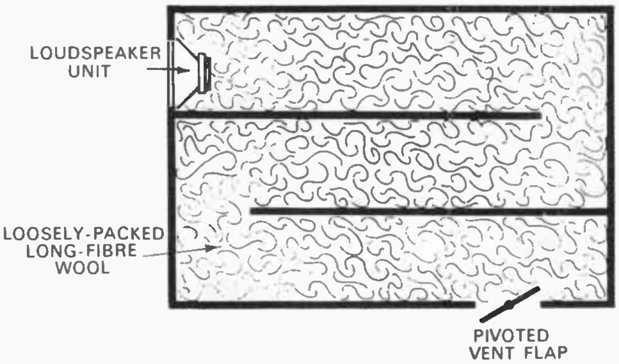


Fig. 5. Experimental acoustic transmission line cabinet.

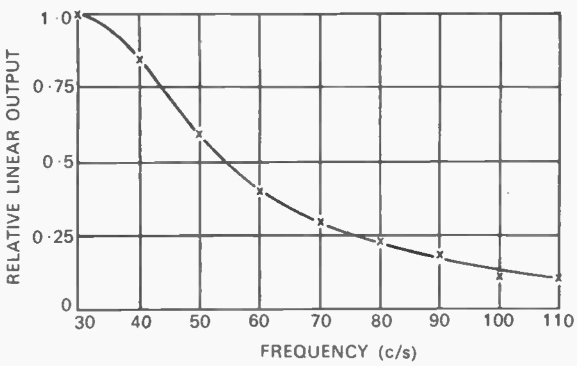


Fig. 6. Sound pressure in closed vent of experimental cabinet.

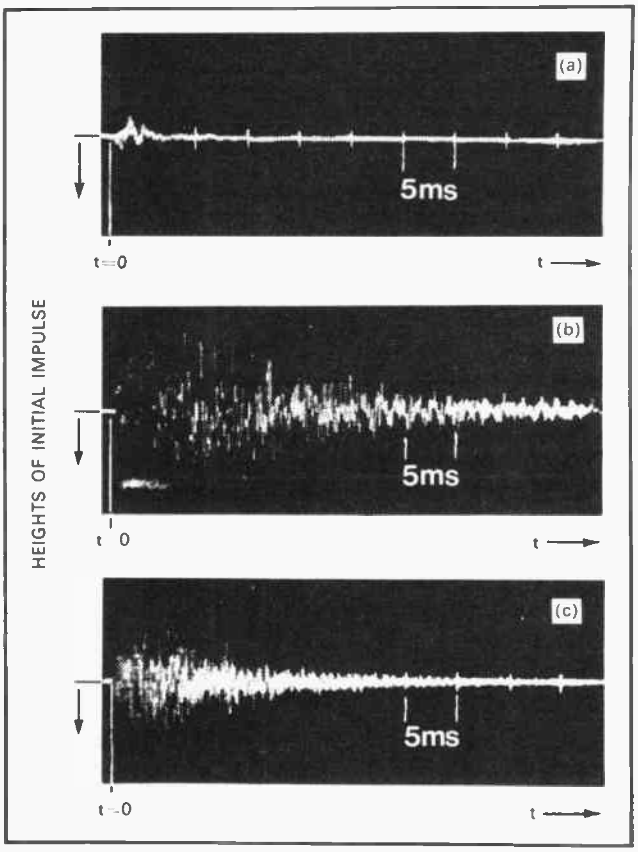


Fig. 7. (a) Impulse response of acoustic transmission line loudspeaker cabinet. (b) Impulse response of bass-reflex cabinet identical in volume to (a). (c) Impulse response of bass-reflex cabinet with internal lagging.

were then tried but none proved to be really satisfactory. The author is therefore indebted to his colleague, R. V. Leedham, for suggesting the use of exploding wires as a standard impulse source.

Exploding wires proved to be a delightfully simple and accurate method of generating an acoustic impulse. Basically the method involved discharging a low inductance capacitor of high value (1000 μF) charged to about 250 V through 1 cm of 40 s.w.g. tinned copper wire. The wire is vapourized almost instantaneously and the acoustic impulse produced had a rise time well into the supersonic region. Spark-generated impulses could have been used, but a high-voltage source is necessary of considerable stored energy if an adequate impulse is to be produced. As exploding wires were less lethal experimentally and only needed standard power supplies, the use of a spark source was not pursued.

The measuring microphone used had a working bandwidth of 30 to 10 kc/s and was used inside the cabinet at a distance of 18in from the exploding wire source. The exploding wire was operated at the position where the loudspeaker would be used, the loudspeaker opening being blanked off.

The results appear in Fig. 7a, the initial impulse being just discernible. The results were felt to be very creditable, the large damped oscillation being the flexure of the ¾in blockboard immediately behind the exploding wire. The experiment was then repeated with a bass-reflex cabinet of identical size having a port area of some 24in² and unlagged internally. The results were markedly dissimilar. Acoustically a much louder hollow explosion could be heard and the microphone pickup showed a far larger spurious output for a much longer time. This is shown in Fig. 7b, the sensitivity and time scales being identical with that of the previous test.

The cabinet was then lagged internally with sound absorbent and the test was repeated. The result is shown in Fig. 7c, the resonance obviously being better damped but still far worse than the line type of cabinet.

Listening tests proved that the cabinet had a "cleaner" sound than the bass reflex type, the effect of the line being very noticeable in its lack of coloration on speech. Transient response was definitely better on the line speaker, the sound being more "tight" and natural.

For obvious reasons it is preferable to have the long axis of the loudspeaker in the vertical plane. The cabinet ducting arrangement was therefore rearranged and one commercial form is as shown in Fig. 8. To make the most of the cabinet it is obvious that the loudspeaker units must not possess large colorations of their own. The units quoted give very good performance although other equally good units may be available. The cross-over frequency used is 1500 c/s.

The frequency-amplitude response of the complete loudspeaker system is shown in Fig. 9. The rate of fall at the low frequency end is creditably slow and far better than the majority of systems in use. It is not unknown for rates of cut-off to be as high as 18 dB per octave and to start very rapidly. This gives rise to a "heavy" bass effect that some people prefer; it is, however, not natural.

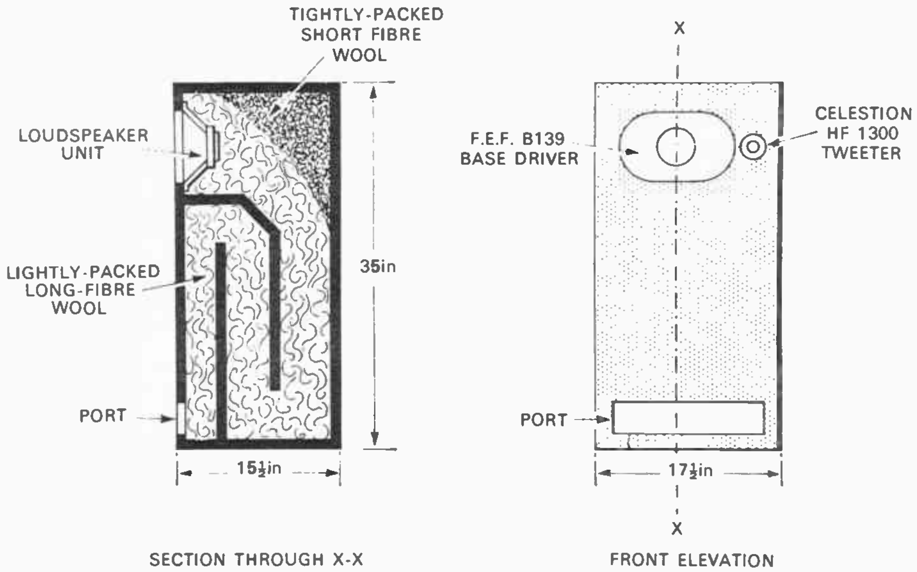


Fig. 8. Production version of the acoustic-line loudspeaker cabinet.

Cabinet material ¾in solid Afrormosia and ¾in laminated wood.
Internal partitions braced down centre line of cabinet.

The bass resonant frequency of the speaker unit is below 15 c/s in the enclosure (about 30 c/s in free air) and quite well damped, so this will have no noticeable effect on the output. As the acoustic loading of the pipe is, therefore, dominating the speaker unit, the low-frequency waveform will be better as the non-linearity in the loudspeaker unit suspension will be swamped by the linear acoustic loading.

The final subjective tests were very good. The sound quality is effortless and natural. At first hearing the bass sounds to be deficient but extended tests show that this is not so, it is merely that one has been conditioned to hearing resonant bass. The overall effect is surprisingly unexciting—only natural. In over a year's use of the system the author has noted, however, that musical listeners were very impressed with the result.

Practical Points

The cabinet design is not critical, and many variations are possible. The only cardinal point is that of keeping the pipe area above that of the cone. A rather strangled result can occur if an attempt is made to save space by restricting the pipe area much below that of the speaker cone. It must also be noted that a poor speaker does not usually sound much better in a good cabinet as the speaker deficiencies dwarf the improvement.

The application of the principle of the design is the subjective of a Patent, but there is no restriction, of course, on private individuals making cabinets for their own use. For the amateur constructor the following points may be of use:—

(1) The cabinet should be make of thick acoustically dead material, chipboard being generally better than plywood. Due to the absence of high internal pressures and the absorbing effect of the wool, the cabinet thickness and bracing are not as important as in the case of the bass-reflex.

(2) Acute bends in the pipe should be arranged to occur as far from the loudspeaker cone as possible to reduce the magnitude of standing waves due to reflections.

(3) The wool should be of long fibre length and packed fairly loosely, about one pound to every two to three cubic feet. Long-fibre wool of the correct specification can be obtained from John W.

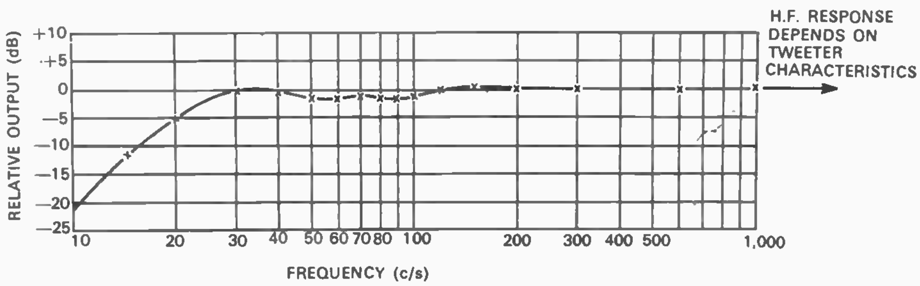


Fig. 9. Overall sine-wave response of Fig. 8.

Pennington Ltd, Midland Wool Warehouses, Briggate, Shipley, Yorkshire, at a cost of 75p plus v.a.t. per pound.

(4) Either spray the wool with mothproofer or take other suitable action or the cabinet performance may suffer from an ageing process.

The author has constructed several different cabinets of totally different sizes and geometry, and apart from narrow pipes and badly angled bends the performance has been remarkably similar. In fact a low resonance 4in unit has been used effectively and gave a good output at 35 c/s. The power handling capacity was, however, limited.

Acknowledgements.—In conclusion the author would like to thank Radford Electronics Ltd. for permission to give the details shown in Fig. 8. Also thanks are due to R. V. Leedham and other colleagues for their help and criticism.

REFERENCES

1. "Sound Reproduction" G. A. Briggs, p. 62.
2. *Ibid*, p. 56.

Letters to the Editor

Non-resonant Loudspeaker Enclosure

SINCE writing the article on the non-resonant loudspeaker, I have been examining further the performance of acoustic absorbents. Of those that are readily available, kapok is about the best but is not up to the performance of long wool. If the kapok is very well teased out then its properties are quite good. Unfortunately however it gradually compacts with use and the acoustic performance suffers accordingly. It may be possible to support it with wire netting, but this in turn can give resonance troubles.

The short fibre wool mentioned originally is very uncritical and cotton wool, kapok, or any usual cushion stuffing material is quite suitable. The purpose is only that of mid-frequency absorption and this is easily done by most textile materials.

ARTHUR R. BAILEY
Bradford

IT was particularly interesting to read Dr. A. R. Bailey's article describing a non-resonant loudspeaker enclosure, using a transmission line as a load. I would agree entirely with his contention that it is difficult to design a conventional reflex cabinet which is devoid of boom when reproducing the double bass or one which does not produce objectionable coloration of orchestral bass transients. However, I have established that it is possible to remove this defect from the conventional bass reflex cabinet by filling the interior of the cabinet with a fibrous material which provides a resistive load to the cone at low frequencies.

This system also becomes virtually non-resonant and was named a resistive reflex cabinet. The principle was used commercially early in 1962 and was the subject of a patent application on my behalf in May 1961. It is not unlikely that the subjective impression of music reproduced by means of a resistive reflex cabinet would compare favourably with Dr. Bailey's system, although their design concepts are clearly different.

Further research and development has established that reflex cabinets of only 1 ft³ can be made virtually non-resonant in the frequency range above 30 c/s. The bass quality is life-like and there is an absence of boom or chestiness in speech. As a result of further research I have established that the amplifier stability margin at the bass resonance frequency is more satisfactory with such a loudspeaker, and that the transient response of the amplifier and loudspeaker in tandem is well damped. Undoubtedly Dr. Bailey's system possesses the same virtue. It is perhaps strange that speech should sound coloured when the main system resonance is around 40 to 50 c/s. This phenomenon appears to be due to the fact that the d.c. component of the distortion produced in the amplifier appears as a pulse when there is a rapid change of signal level unless the feedback loop is d.c. coupled throughout or has a very long time constant. This internally generated pulse excites the transient response of the amplifier speaker combination and gives rise not only to boomy speech and music, but to excitation of speaker cone resonances as well.

It would seem then that those people who maintain that no two amplifier-speaker combinations should alike are probably right after all.

The transient testing procedures adopted by Dr. Bailey undoubtedly show up the spurious coloration of an enclosure very well.

Another alternative method which I have found useful is to apply a step function to the speech coil from a lead-acid battery by means of a mercury switch. The latter produces very fast rise times without contact effects. Any spurious coloration is revealed outdoors or in an anechoic chamber. It is by the same process possible to identify whether the loudspeaker or room acoustics are producing unsatisfactory bass response.

J. R. OGILVIE
Sevenoaks, Kent.

The author replies:—

I was very interested to read Mr. Ogilvie's comments with regard to loudspeaker systems. There are, however, one or two points that I would like to comment on.

Firstly, there is the perennial problem of obtaining the best possible performance from small loudspeaker enclosures. This has always been a difficult requirement due to diffraction and other effects. I would agree with Mr. Ogilvie that it is possible to make a small bass reflex cabinet virtually non-resonant, but I have always found that the small port size necessary for a low Helmholz resonance gives very little benefit unless the cabinet is very resonant. If he has indeed solved the problem, then there will be many people grateful to him.

I would be interested to know the method of damping that Mr. Ogilvie uses, as all that I have tried in small systems either put up the effective stiffness of the enclosed air to an unacceptable value, or alternatively cause distortion due to the non-linear air friction effects. These same shortcomings exist in the damping materials used in the now popular closed-box systems. Too much stuffing in a bookshelf speaker can make it sound terrible.

Regarding the effect of resonant speaker systems on their driving amplifiers; I will agree that the speaker impedance can rise steeply at resonance peaks, but this should not upset any reasonable amplifier except perhaps under overload conditions. A good amplifier should give a satisfactory transient response at the bass end even with an open-circuit as a load. Overloads on output voltage levels should also not be capable of seriously upsetting the amplifier, irrespective of the output load conditions. Any high-fidelity amplifier worthy of that name should not be upset by load conditions to an audible degree, but then I would agree that there are some amplifiers that are not as good as their title suggests.

Regarding the coloration of speech by resonant speaker systems. I feel that Mr. Ogilvie is being confused between the lowest continuous tone that can be sung and the complex components of speech. The explosive components of speech have constituents that extend below the audible spectrum, these being easily isolated by a third-octave band filter. It is these components that are subjected to the bass resonance frequency of cabinets and speakers and cause the resulting coloration.

I am rather puzzled by the reference to

d.c. components of distortion producing coloration effects. This is contrary to my own experience, where tone-burst testing an amplifier with bandwidth-limited waves gave no measurable d.c. components whatever. With a low-distortion amplifier I would not expect that any distortion products could produce audible colouring from resonances, due to their extremely low level.

When deciding on how to impulse-test loudspeaker enclosures, step waveforms were applied to loudspeakers and their acoustic outputs examined. Unfortunately no loudspeaker was found with a sufficiently good performance for the purpose. Even the best tested had far more coloration than that of the acoustic line cabinet to be tested. Certainly there is still useful work to be done before loudspeakers can be classed as giving true reproduction.

ARTHUR R. BAILEY

Loudspeaker Enclosures

DR. BAILEY'S loudspeaker design which he described in the October issue of *Wireless World*, is a resurrection of the almost forgotten labyrinth enclosure which was popular many years ago, and is still regarded in some quarters as being potentially superior in performance to bass reflex types. It bears no direct relationship to the labyrinth enclosure to which Dr. Bailey attaches the name, and in its usual form its chief disadvantage is the monstrous size required to attain the low frequency performance demanded by modern standards.

It is surprising at first sight, that the excellent bass response, shown by the curve Dr. Bailey gives us, can be produced by a phase inverting line only some 7ft. long. This will have its "free air," half-wave resonance, necessary to achieve the phase inversion required between the rear of the loudspeaker cone and the port opening, at 80 c/s. Below 40 c/s this results in the output from the port containing a component which is in antiphase with that from the loudspeaker cone, decreasing the total output and increasing the rate of fall off.

This is not the case with Dr. Bailey's enclosure, the output being well maintained to frequencies appreciably below 30 c/s, and it must be inferred that the phase inversion occurs by some means other than the free air resonance. There is a retarding effect on the waves within the enclosure, decreasing the frequency of its resonance, and thereby lowering the frequency at which phase inversion occurs. As the wool filling is the only difference between this enclosure and the simple labyrinth, it would appear that this is responsible for this effect, and there is a simple, if perhaps incomplete, explanation which indicates that this is the case.

If we consider two waves of the same frequency, but having different velocities, then:—

$$\frac{\lambda_0}{\lambda_0} = \frac{v_0}{v_1},$$ where λ_0 and λ_1 are the wavelengths corresponding to v_0 and v_1.

But $v = \frac{E}{\rho}$, where E is the elasticity of the propagating medium and ρ its density.

Hence $\frac{\lambda_0}{\lambda_1} = \sqrt{\frac{E_0}{E_1}}\sqrt{\frac{\rho_1}{\rho_0}}$, where λ_0, E_0 and ρ_0 correspond to free air conditions and E_1 and ρ_1 correspond to those in the filled enclosure.

With two assumptions, we can simplify this expression and relate it approximately to the amount of material added to the enclosure.

Firstly, it appears reasonable to assume that with a loosely packed filling, little air will be displaced. Also the fibres are themselves relatively incompressible compared with the remaining air. We can therefore say that $\frac{E_0}{E_1} = 1$ approximately since we can expect little change in the elasticity due to the filling.

Secondly, it seems quite probable that, for frequencies where there is little attenuation in the filled line, the filling, being highly compliant, will respond to the air movement, and its mass will effectively add to that of the air. Thus the density of the propagating medium will be higher than that of air, and to a fairly close approximation, can be assumed to be the density of air plus the filling rate.

The expression given above now reduces to

$$\frac{\lambda_0}{\lambda_1} \simeq \sqrt{\frac{\rho_1}{\rho_0}} \simeq \sqrt{\frac{\rho_0 \times r}{\rho_0}}$$

where r is the filling rate.

It would appear that the half wavelength resonance of Dr. Bailey's enclosure occurs at 30 c/s corresponding to a free air wavelength of 36ft. But the wavelength corresponding to the unfilled enclosure is $2 \times 7 = 14$ft.

Hence $\lambda_0 = 36$ft., $\lambda_1 = 14$ft.,

and $\rho_1 = 6{\cdot}6 \quad \rho_0 = 0{\cdot}5$ lb/ft³, taking $\rho_0 = 0{\cdot}075$ lb/ft³ at room temperature.

This means that the filling must be added at a rate of 0·425 lb/ft³ or 1 lb to every 2·3ft³ of enclosure, which is within the range recommended by Dr. Bailey.

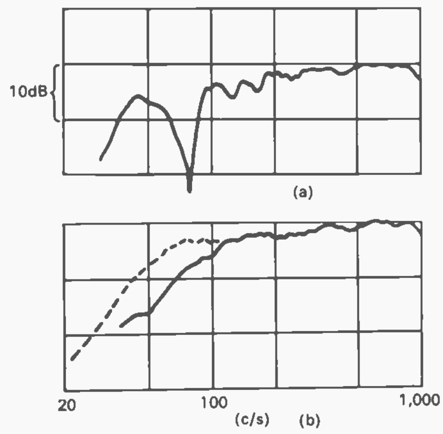
It is interesting to note that the line can be tuned to the required resonance by addition or subtraction of filling; this was always a difficulty with the simple labyrinth, since the fundamental resonance of the system changes with a change of line length, and "cut and try" could be expensive on timber. Furthermore, the use of other media is indicated since it is the weight added which is important. Provided that the low-pass characteristic can be correctly maintained, higher packing densities could be used to reduce still further the enclosure size.

So far as the reduction of spurious resonances is concerned, many of the small airtight enclosures currently available are filled with a fibrous damping medium. But it is doubtful if any of them use the velocity retardation effect at undamped frequencies other than by accident. Certainly none could use it to better effect than the labyrinth, where not only does it in this case provide a reduction of 2·6 times in the line length, but also in the other dimensions. The required volume has been shrunk from a gargantuan 100ft³ to a domesticated 5·5ft³.

This is a remarkable achievement and with its possibilities for further improvement and application is of far greater importance than the other, coincidental, properties of Dr. Bailey's enclosure.

E. A. HARMAN
Chorley, Lancs.

SIX years ago, the writer tested a labyrinth cabinet almost identical to that described as an acoustic transmission line cabinet by Dr. Bailey in the October issue. Response curves taken under free-field conditions are shown in the Figure. Variations of cabinet and absorbent gave the same result of numerous resonances, as also did a folded horn. The curve for a totally enclosed cabinet of less than half the volume is included for comparison; provided the cabinet is not long and narrow, only the fundamental is present. These results were given in a lecture to the (then) Brit.I.R.E. on January 24th, 1962. Similar results were obtained many years ago by H. J. Leak and J. Bolingbroke. The original labyrinth was essentially a



Mr. Barlow's response curves. (a) labyrinth lined with ½in thick cotton wool; impedance peaks: 87, 140, 180, 230, 330, 460, 720 and 870 c/s (fundamental 44 c/s): (b) labyrinth completely filled with cotton wool; impedance peaks: 74, 100 and 340 c/s (fundamental 23 c/s). The dotted curve is for a totally enclosed cabinet half filled with cotton wool; impedance peaks: fundamental only 65 c/s.

resonant device, in which the resonances and anti-resonances were used to equalize the speaker output. It will be noted that the rate of cut off of the totally enclosed cabinet is similar to that of the absorbent-filled labyrinth, and can be varied if need be by design. When measured standing against a wall, as is done by Dr. Bailey, the response of the labyrinth may tail off more gradually, but this would apply also to the totally enclosed cabinet. If it is desired to tail off the bass gradually from a relatively high frequency, there are simpler and less resonant devices than the labyrinth for doing this.

D. A. BARLOW
H. J. Leak & Co.,
London, W.3

The author replies:—I read Mr. Harman's letter with great interest as his theory is borne out in practice. The velocity of sound in wool is considerably slower than in free-air, and is also slower than can be accounted for by the difference between isothermal and adiabatic compression of the air. The wool mass is definitely slowing down the wave front, but as there cannot be perfect coupling between the wool and the air the effect will be somewhat less than given by Mr. Harman's calculation. On the other hand the wave will be slowed by the isothermal effects of the wool as well, so the error in assuming perfect coupling will be reduced.

As Mr. Harman surmises, the velocity of sound can be slowed down very greatly in a high packing density, but unfortunately this gives rise to high back pressure on the loudspeaker cone due to the very restricted air passages. There is therefore a maximum packing density that can be used without giving a strangled effect to the sound. The maximum density varies with speaker design and cabinet design, but is far greater than the density used in the cabinet described.

Regarding Mr. Barlow's letter, I feel that he must have misunderstood the article. This may have been my fault, but the cabinet design is based on a transmission line (which should have no reflections) having energy absorbing properties at all but the lowest frequencies. There is no desire to form a labyrinth (dictionary definition—with many turnings) at all. In fact every turning tends to cause reflections and these are contrary to what is required.

Without knowing what design of cabinet Mr. Barlow used, it is difficult to be analytical of his results. It may be of interest, however, to note that cotton wool has not proved to be a suitable material from the tests that I carried out. I would disagree that the rates of cut-off are the same in the second figure, my constructed asymptotes on the mean rate of cut-off give the labyrinth a 5 dB per actave slower rate of fall.

Incidentally my own response curve was taken with B. & K. equipment with the speaker back to the wall of a 60 ft long laboratory, the microphone being 1 ft in front of the speaker midway between the speaker and vent axes. A free-field response was not given as this is intolerably bass-heavy if a flat characteristic is obtained. A floor and a wall were felt necessary to simulate the effect of normal domestic listening conditions.

If Mr. Barlow is still convinced that a closed cabinet gives better performance, then I will be only too happy to give him a demonstration of the system's capabilities. A 25 c/s pure sine wave can be generated acoustically by the system. A very large enclosed cabinet would be needed for this as the cone resonance is increased by the enclosed air. Incidentally, the effective system resonance of the transmission line speaker is below 15 c/s for the design given. The cone resonance as such may be above or below its free-space figure depending on the sign of the reflected reactance of the transmission line. This factor, however, has little significance as line loads the cone resistively to such a degree that reactive effects are negligible within the audible range.

ARTHUR R. BAILEY

Low-cost High-quality Loudspeaker

Design for frequencies above 100Hz. 1-Construction and assembly

by P. J. Baxandall, B.Sc. (Eng.), F.I.E.E., F.I.E.R.E.

This loudspeaker may be built by the home constructor for a total expenditure in the region of £10. While it does not have the extended bass response of some much more expensive loudspeakers, it is nevertheless unusually free from the colorations and hangover effects which are unfortunately still a feature of the majority of commercial designs[1]. Consequently, on many types of programme material, it will be found to give considerably better reproduction than is obtained with many much more costly loudspeakers.

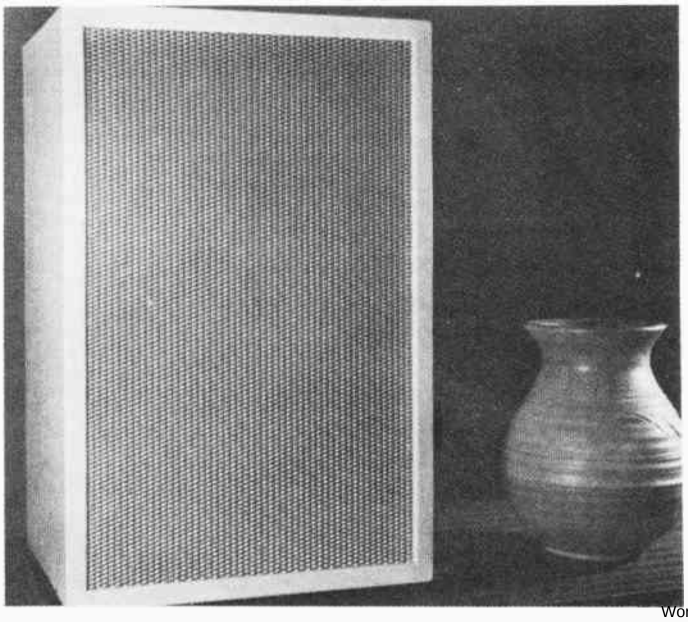
On speech, both male and female, the loudspeaker reaches a very high standard of performance. Using a high-grade capacitor microphone out of doors, an almost deceptive degree of realism can be achieved in the reproduction of familiar voices.

While many people who have heard the loudspeaker on music seem to find the bass response fairly adequate, direct comparison, particularly on organ music or large-scale orchestral music, with a good loudspeaker, such as a B.B.C. monitor[1], leaves listeners in no doubt that the reduced response below about 100 Hz constitutes the main shortcoming of the design*. Consequently it is recommended that, where space permits, the basic low-cost loudspeaker should be augmented, at frequencies below about 100 Hz, by a separate woofer. Because this has to cover a frequency range of only about one octave, in a rather uncritical part of the spectrum, there is much latitude in its choice and almost any old 12-in unit, such as can be bought second-hand for a pound or two, can be pressed into service.

In a stereo system, one such woofer can be shared very satisfactorily between the two channels, and because these very low frequencies convey almost no sense of position, the woofer can be placed in any convenient position in the room. When circumstances permit, the possibility of mounting the woofer unit in a hole in the floor, ceiling or a wall may be worth considering, as it saves the space and labour of cabinetwork.

* Of course, some small-scale music of great beauty contains almost no frequencies below 100 Hz and is therefore virtually unaffected by this shortcoming.

The complete low-cost loudspeaker. The size of the cabinet is 18in. × 12in. × 10in.(deep).



Suitable circuit arrangements for such a three-speaker stereo system will be discussed in Part 2 of this article, and it will be shown how the relative levels from the woofer and the other two loudspeakers may be adjusted to give nicely balanced reproduction in listening rooms having different acoustical properties.

A complete stereo system on the above lines can thus be built for no more than about £20, and is capable of a surprisingly high standard of reproduction. Even direct comparison with a pair of Quad electrostatic loudspeakers does not always reveal any obvious shortcomings, though careful listening over a period of time makes it evident, in particular, that the lower intermodulation and hangover distortion of the electrostatic speakers results in greater clarity and separation of instruments particularly at high volume levels. Nevertheless, the low-cost system is capable of quite impressive volume and clarity in the reproduction of orchestral and choral music in rooms of normal living room size, and in much music of a quieter nature listeners have shown no marked preference for one or the other speaker system.

Evolution of the design

The following thoughts were significant in the evolution of the present design, which aims to satisfy an evident demand for the best possible quality of reproduction at a really low price:

(a) Large loudspeaker units suitable for a wide frequency range are expensive and need augmenting by a tweeter for really first-class results.

(b) Smaller circular units, e.g. 8 in., often suffer from undesirable hangover effects in the lower-middle-frequency range[1,4,5] and the unpleasant sound of these cannot be fully removed by electrical equalization. However, it was mentioned by Dr. G. F. Dutton of E.M.I. at the discussion following Mr. Shorter's paper[1] that the use of elliptical rather than circular diaphragms gives a marked reduction in hangover distortion, which is caused by diaphragm vibration persisting in low-damped radial modes after the cessation of the signal.

(c) The surprisingly good results given by a commercial loudspeaker known as the 'CQ Reproducer', which used a cheap elliptical unit almost the same as that employed in the present design, served further to direct the author's attention to the virtues of elliptical diaphragms, and preliminary measurements on such a unit showed that it had an axial frequency response which, if its main departures from levelness were to be corrected by a cheap and simple electrical equalizer, would give a sufficiently uniform and wide-range response to meet the requirements of very high quality reproduction—except that some sacrifice of performance at very low frequencies seemed virtually unavoidable.

(d) The use of a single unit to cover the whole frequency range also simplifies matters by avoiding the problem of the unnatural changes in polar response which are liable to occur in the cross-over regions of multiple-unit systems[1].

(e) While the exploitation of cabinet panel resonances to modify the frequency response over certain ranges is a dodge which has sometimes been employed with a degree of success in cheap designs, it was felt to be such a tricky and unpredictable technique that it would probably be much better avoided.

(f) The notion that very high flux densities are essential for good transient response, while a widely propagated belief, is not in accordance with much practical experience†.

Consequently the fact that the cheap elliptical unit being considered had a rather small magnet was not regarded as of much significance in this context.

(g) Of much greater significance was felt to be the fact that only quite small diaphragm excursions, in the region of $\pm$ 1 mm, can be made without running into considerable suspension non-linearity and non-linearity caused by the rather skimpy nature of the coil and magnet geometry. Indeed it is still a source of some surprise that such substantial volume can be obtained in practice without these non-linear effects giving any obvious subjective impairment of the reproduction.

The basic recipe adopted thus involves no more than the use, in association with a simple electrical equalizer, of a particularly suitable, though quite cheap, elliptical unit having a plasticized surround, mounted in a totally enclosed box made rigid by internal bracing and containing felt damping material to reduce standing-wave effects and provide some additional damping of the main diaphragm resonance.

The size of the box is such that the stiffness of the enclosed air at low frequencies, referred to the diaphragm is about equal to the mechanical stiffness of the diaphragm suspension, resulting in a resonant frequency of about 100 Hz. This size of box is quite convenient to accommodate, and the improvement in bass performance given by even quite a large increase in volume would not be great. Moreover, the greater the overall stiffness, the less will be the intermodulation distortion when strong low-frequency signals are fed to the unit, e.g. at 40 Hz, at the same time as higher frequencies. The size of box adopted is thus thought to be a good all-round compromise.

While the use of a vented enclosure has been carefully considered, such an arrangement would either result in a considerable increase in intermodulation distortion in the presence of large inputs at very low frequencies, or, if the Helmholtz resonant frequency were made low enough to avoid this danger, the response at very low frequencies would be at a lower level than that at higher frequencies, requiring further electrical equalization. For normal circumstances, the simple totally enclosed box seemed to be the best choice, therefore.

The use of an equalizer of fixed design, not adjusted to suit individual loudspeaker units, will obviously be satisfactory only if the variations in response of the units in production is sufficiently small. All that can be said is that of several units checked, bought over a period of several years, all have had the same main features in their frequency responses, fairly closely matched by the equalizer characteristic. Thus, while it cannot be guaranteed that every loudspeaker built to the present design will have quite such a good frequency response as the prototype, it seems virtually certain that the equalizer will always effect a marked improvement in the results.

Frequency response curves

The full-line curve in Fig. 1 shows the measured axial frequency response of the loudspeaker without the equalizer‡. It will be seen that, ignoring the numerous small wiggles (which appear in virtually all loudspeaker response curves if the frequency is varied slowly enough), the main features of this curve are a region of excessive output centred broadly just below 700 Hz, and another similar region centred at about 7 kHz.

† A weak magnet may give rise to a peak in the frequency response in the region of the main resonant frequency of the diaphragm[2]. While this is not necessarily undesirable if it occurs well below 100 Hz, where some degree of ringing does not seem to give subjectively noticeable impairment of transient response, it can in any case be damped down by acoustic means, e.g. by a close-fitting felt cover over the loudspeaker unit, if this is thought desirable. At higher frequencies, many of the diaphragm resonances are so weakly coupled to the coil that little electromagnetic damping can occur even if the flux density is very high.

‡ This measurement was made out of doors using a small home-made omnidirectional capacitor microphone at a distance of 2 ft. 6 in. from the front of the loudspeaker and at a height of 4 ft above ground, on axis. The microphone was used in the r.f. bridge system described in Reference 6, and its pressure calibration was obtained by developing a constant alternating force on the diaphragm by means of an oscillator voltage applied in series with a d.c. polarizing voltage. To avoid any significant error at high frequencies due to pressure doubling, the capsule was then placed in front of the loudspeaker with its ½ in. diameter diaphragm in a horizontal plane.

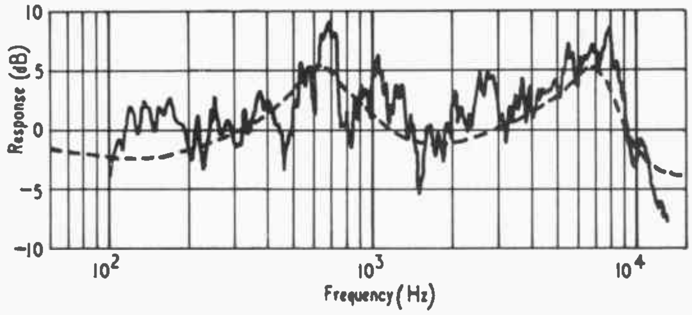


Fig. 1. Full-line: unequalized axial frequency response of loudspeaker. Broken line: inverse of equalizer frequency response.

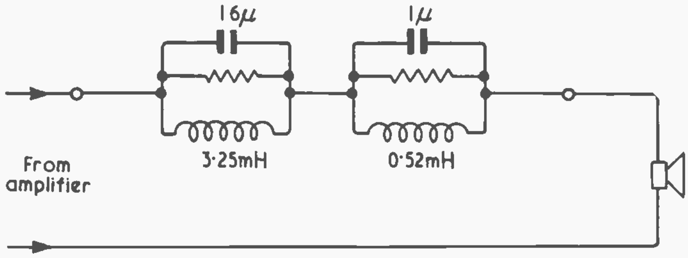


Fig. 2. Basic equalizer circuit.

The basic equalizer circuit designed to correct the Fig. 1 response is shown in Fig. 2. However, because 16 μF is an inconveniently large capacitance, the practical equalizer circuit is arranged as in Fig. 3. The full-line curve in Fig. 4 shows how the equalizer causes the voltage across the speech coil to vary with frequency for a constant amplifier output voltage. Referring to Fig. 1 again, the broken-line curve is an inverted version of the full-line curve in Fig. 4, and shows that the equalizer characteristic is quite well matched to the main features of the loudspeaker response. (The broken-line curve in Fig. 4 simply shows the effect of removing the damping resistors from the equalizer circuit.)

Fig. 5 shows the overall axial response curve of the loudspeaker with the equalizer incorporated and it will be seen that most of this lies within $\pm$3 dB limits from 100 Hz to over 10kHz, the general balance of low, medium and high-frequency response being within even finer limits.

The loudspeaker unit. The unit is an Elac 15-ohm 9 in. × 5 in. elliptical unit, Type 59 RM/109, manufactured by Electro Acoustic Industries Ltd., Stamford Works, Broad Lane, Tottenham, London N.15. The retail price is £2.47 plus v.a.t. When ordering state that 15-ohm impedance is required.

Constructing the equalizer

The equalizer circuit has already been given in Fig. 3. The inductors both employ 0.014 in. silicon iron laminations, Inter-Service No. 421. These are conveniently obtainable in kits from the Belclere Company Ltd., 385/387, Cowley Road, Oxford. Each kit consists of a stack of Silcor 107 laminations, a bobbin and a steel shroud. For each equalizer, two kits are required:—

Kit GN/Silcor (7/16 in. stack) Price 78p
Kit GX/Silcor (3/4 in. stack) Price £1.08

It is essential to specify 'T' and 'U' laminations when ordering, as the firm now normally supplies 'E' and 'I' types, but has agreed to supply 'T' and 'U' laminations for this equalizer when requested to do so.

L_1 winding. The tapped inductor L_1 uses the larger GX size core stack. First wind on 110 turns of 28 s.w.g. enamelled copper wire in four neat layers. Then wind on, *in the same direction*, 330 turns of 34 s.w.g. enam., making 440 turns for the whole bobbin. The 330 turns need not be wound in accurate layers—just wound on reasonably tidily. There is no need for any insulation between sections, but the outside of the winding should preferably be protected by empire cloth or thick paper.

L_1 core. Place all the T's through the bobbin tunnel from one side. Place all the U's in the shroud, with small pieces of cardboard, or 1/16 in. s.r.b.p. ('Paxolin') ¼ in. × ¾ in. as shown in Fig. 6, to prevent the steel shroud coming too close to the core gaps.

Each of the three gaps should be 0.025 in., which must, of course, be formed by inserting suitable insulating material. In the absence of other facilities, use may be made of the fact that the outside cover of *Wireless World* has been made of paper of thickness approximating closely to 0.005 in. for at least ten years! Thus insert five thicknesses of this paper in each gap. It will be found convenient to cut strips of widths approximately 1/8 in. and 3/8 in. for the outside and central gaps respectively and to fold these strips in zig-zag fashion to form five thicknesses. With these gap-spacers in position, the shroud should be screwed down tightly onto the wooden baseboard, 3/8 in. No. 6 roundhead woodscrews being suitable.

L_1 connections. The enamelled wires from the bobbin should be carefully bared with sandpaper or emery paper and soldered to the three tags of a tagstrip screwed down to the baseboard as shown in Fig. 6 and the photograph below. The beginning of the winding (inner end of the 28 s.w.g. section) should go to the tag nearest to the end of the baseboard, the outer end of the 28 s.w.g. section and the inner end of the 34 s.w.g. section going to the middle tag to form the tapping point. The outer end of the 34 s.w.g. section goes to the tag nearest the middle of the baseboard.

L_2 winding. The untapped inductor uses the smaller, size GN, core stack. Wind on 86 turns of 24 s.w.g. enam. in four neat layers and cover with empire cloth or paper. (For winding this and the other inductor, a simple gadget may be improvised, using bits and pieces from the junk box, Meccano, etc., for rotating the bobbin. There is no need for a turns counter—the number of turns is small enough to be counted without difficulty mentally!) If an aluminium or other non-ferrous shroud is used instead of the sheet steel one employed in the prototype (supplied by Belclere), the winding turns should be increased to 95. Also, because of the increased shunt loss resistance then obtained, the damping resistor value (Fig. 3) should be reduced from 120 ohms to 68 ohms. The use of an aluminium shroud for inductor L_1 has no significant effect on the inductance value or losses, owing to the much smaller air gap.

L_2 core. Insert all the T's through the bobbin tunnel from one side and place the shroud over the core so that the tops of the T's lie inside the top of the shroud. A piece of 3/16 in. thick soft packing material measuring 1 in. × 3/4 in. should now be obtained. This is placed between the bobbin and the wooden baseboard so that the bobbin and laminations are pressed securely up into the shroud when the latter is screwed down.

No U laminations are employed for this inductor.

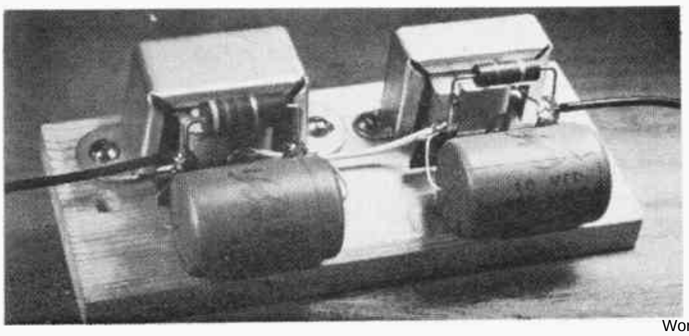
Wire supply. The three gauges of enamelled copper wire required (24, 28 and 34 s.w.g.) may be conveniently obtained, in a minimum quantity of 2 oz each, from Post Radio Supplies, 33, Bourne Gardens, London, E.4.

The three 2-oz reels contain enough wire for at least four equalizers.

Other components. The other components required for the equalizer are all readily obtainable, including tag strips, from Radiospares Ltd. through any radio dealer. Tubular 1 μF paper capacitors, 250 V d.c. wkg., $\pm$ 20% tolerance, are suitable.

Tests. Provided the above instructions have been carefully carried out, it is virtually certain that the equalizer will function correctly. However, if an oscillator is available, it is worth while to check that, with a constant voltage fed to the series combination of equalizer and loudspeaker, the voltage across the loudspeaker varies with frequency approximately as shown in Fig. 4.

The equalizer.



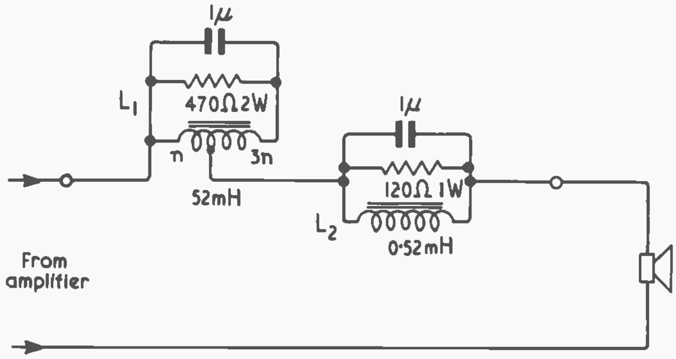


Fig. 3. Practical equalizer circuit.

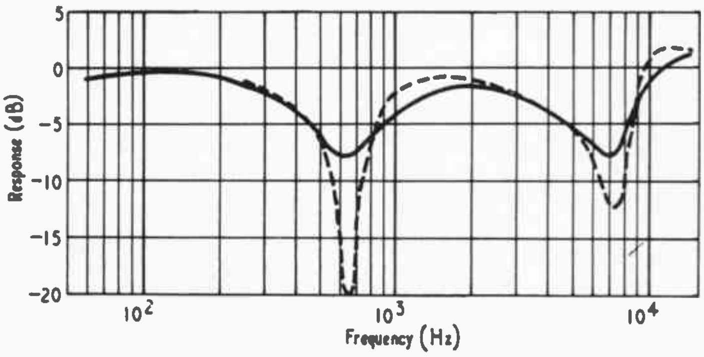


Fig. 4. Measured frequency response from amplifier output to speech-coil. Full line: with damping resistors as in Fig. 3. Broken line: without damping resistors.

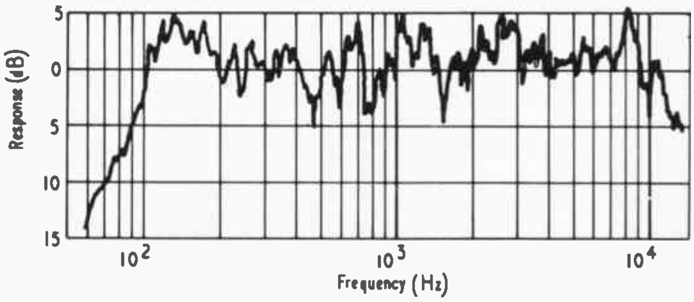


Fig. 5. Axial frequency response of loudspeaker, with equalizer.

The exact position of the lower-frequency dip is slightly dependent on the a.c. voltage at which it is determined. With a source voltage of 2 V r.m.s., the dip will occur about 20 Hz lower in frequency than with a source voltage of 0.2 V r.m.s. For voltage levels above 2 V r.m.s., the fall-off in dip frequency with increasing level is more gradual. (This effect is due to the fact that the initial a.c. permeability of silicon iron is rather low compared with its value at higher flux densities; the effect is well diluted by the presence of gaps, however, and does not seem to give rise to any subjectively noticeable distortion.)

The performance of the equalizer may be regarded as satisfactory provided the measured results fall within the following limits:

(a) Low-frequency dip, with a test voltage of about 2 V r.m.s. applied to the combination of equalizer and loudspeaker, 580 to 800 Hz.

(b) High-frequency dip (almost independent of test voltage) 6200 to 7900 Hz.

(c) Magnitude of dips (almost independent of test voltage), relative to response at 1700 Hz, −5 to −8 dB.

In the unlikely event of the performance falling outside any of the above limits, adjustments may be made as follows:

To correct (a), adjust gap of L_1. Increasing the spacer thickness by 0.005 in. will raise dip frequency about 40 Hz.

To correct (b), the value of C_2 may be modified or, alternatively, the number of turns on L_2 may be adjusted. Removing 5 turns will raise the dip frequency about 400 Hz.

To correct (c), alter the appropriate damping resistor value.

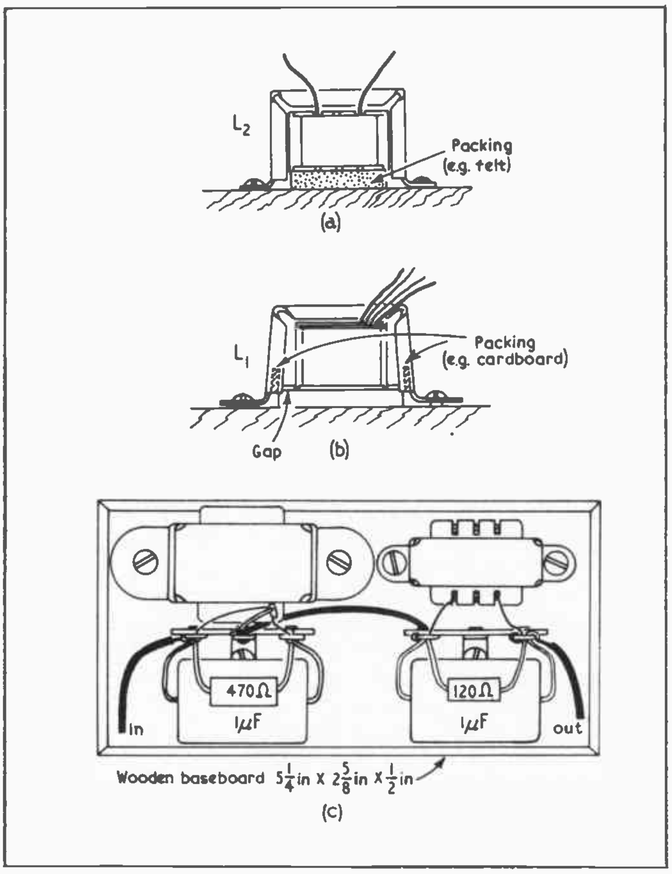


Fig. 6. Constructional details of equalizer.

Equalizer intermodulation distortion. Variation in inductance of L_1 and L_2 with the instantaneous value of large low-frequency signal currents flowing through them is a possible cause of intermodulation distortion. A test showed, however, that the inductance of L_1 dropped by less than 2% when a direct current of 0.25 A was passed through the whole winding, equivalent to a current of 1 A through that part of the winding traversed by low-frequency signal currents. The effect in L_2, because of the much larger gap, will be even smaller. It is obvious, without further calculation, that the distortion caused will be considerably smaller than that introduced by non-linearities in the loudspeaker unit itself.

Alternative equalizer design. While the above method of constructing the equalizer is attractively cheap, some readers may find it more convenient to use Mullard Vinkors. Brief winding details are:

L_1: 99 turns of 28 s.w.g. plus 297 turns of 34 s.w.g. on Mullard 35 mm, $\mu_e = 63$, Vinkor. (LA2102 core and slug, DT2180 bobbin, plus DT2151 or DT2187 casing or DT2234 mounting clip.)

L_2: 49 turns of 24 s.w.g. on Mullard 25 mm, $\mu_e = 63$, Vinkor. (LA2302 core and slug, DT2179 bobbin, plus DT2149 or DT2185 casing or DT2228 mounting clip.)

The resistor R_2 across L_2 should be 68 ohms as compared with 120 ohms in the Belclere version—this allows for the lower losses of the Vinkor. R_1 remains at 470 ohms.

Equalizer kit. A complete kit of parts for the equalizer, including ready-wound inductors on laminated cores, may be obtained from Peak Sound (Harrow) Ltd., 32 St. Judes Road, Englefield Green, Egham, Surrey. The price is £1 16s. The author understands that this company will also supply a kit of parts for the complete loudspeaker (including cabinet parts).

Constructing the cabinet

While some readers may prefer to buy a cabinet and adapt it to the present design, there must be many others who, like the author, find woodwork an enjoyable and rewarding pastime and would prefer to make their own. For this reason full details and a few constructional hints and tips are given.

The author used ½-in. gaboon plywood, but veneered chipboard ('Weyroc') is a satisfactory alternative. While minor changes to the dimensions shown in Fig. 7 may be made to suit individual requirements, the volume of the cabinet should preferably be kept about as shown.

The cabinet is held together by a combination of Cascamite glue and ⅞ in. countersunk woodscrews, screwed into the inside of the cabinet sides through the ½-in. × ½-in. wooden strips shown. The local timber yard supplied the latter, and also the 1-in. × ½-in. strip for fixing the speaker mounting board, in ramin, a beautifully straight-grained, but quite cheap, hardwood ideally suited to the purpose.

The bottom of the cabinet should be fixed with screws only (no glue) so that it can be removed easily for extracting the speaker mounting board, should this become necessary at any time for renewal of the expanded aluminium or Tygan covering —see photograph on page 97.

The use of mitred corner joints at the top corners of the cabinet allows the veneer to extend right up to the corner, and unlike some other corner joints[7] can be cut by the amateur without special rabbet-cutting tools. The author had not previously tried cutting mitred joints for box corners, but found it surprisingly easy to produce a thoroughly neat job.

The aim should be to cut the wood *very* slightly off 45°, so that when the screws are tightened into the ½-in. × ½-in. strip, the mitred joint is sure to close tightly at the outside of the box.

Either Tygan or expanded aluminium may be used for the front of the loudspeaker, according to choice[7]. While expanded aluminium causes slightly less acoustic obstruction, many people prefer the appearance of Tygan. Expanded aluminium may be obtained from the Expanded Metal Company Ltd., P.O. Box 14, Stranton Works, West Hartlepool, Co. Durham.

A suitable type is List Ref. No. 363A in plain aluminium (22 s.w.g.). It is made in standard sheets 4 ft × 2 ft, in which the 'long way of mesh' (which must be horizontal when mounted in the loudspeaker, for satisfactory appearance) runs along the 4-ft dimension. Half a sheet, 2 ft × 2 ft, is therefore a sensible quantity to order for two loudspeaker cabinets. A piece 2 ft × 6 in. will be left over, but will probably come in useful sooner or later.

Two of the four cabinets made by the author are wax polished with expanded aluminium fronts. After very thorough sandpapering, finishing with No. 1 sandpaper, one thin coat of white French polish was put on quickly with a cloth rubber, followed by wax polishing with Meltonian white shoe polish. Nothing could be much easier and quicker than this finish, which is nevertheless very pleasing. The other two cabinets, however, are painted white and have pattern U528 Tygan.

If expanded aluminium is used, it is important that it should be fixed in such a way that it cannot rattle. The author glued a ¼-in. wide strip of ⅛-in. thick felt round the periphery of the front surface of the speaker mounting board, thus separating the main area of the expanded aluminium from the board. To make doubly sure the aluminium would not vibrate against the board, ½-in. squares of ⅛-in. felt were stuck to the board at four positions round the outside of the speaker hole. This felt, the speaker mounting board and the edge of the cone-fixing cardboard on the speaker unit should be painted matt black, to prevent any of these being visible through the expanded aluminium. As previously mentioned, the 'little louvres' of the expanded aluminium should be horizontal rather than vertical: there is also a right way up to mount this material which gives minimum transparency from the usual viewing angles. Even the two sides of a piece of expanded aluminium will be found on careful inspection to be slightly different, and it is worth putting it the same way round if two cabinets are being built for stereo working.

When Tygan is used, the strips to space it from the speaker board need not be of soft material, and ⅛-in. hardboard is suitable**. The Tygan may be stuck round the edges of the board

** It might be thought that with cloth there would be no need to space it away from the board. It is easily demonstrated, however, that if the cloth is touching the board but is not stuck to it, buzzing sounds are produced at certain low frequencies. The trouble with sticking the Tygan to the front of the board is that it is liable to make the outline of the speaker hole visible from the front.

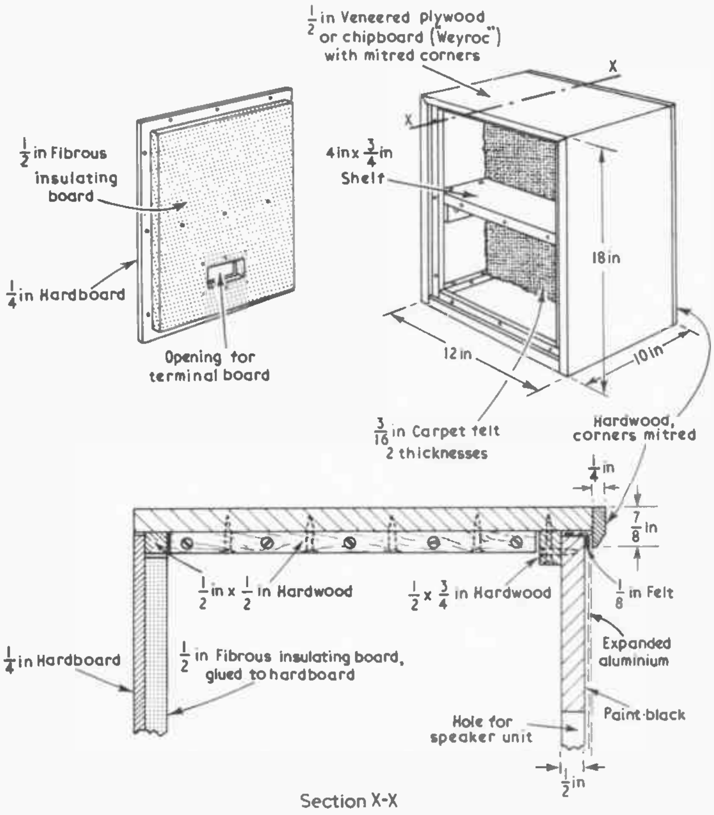


Fig. 7. Cabinet construction details.

with Evo-stik impact adhesive, care being taken to keep the warp and woof running parallel to the board edges. If this is not got quite right at first, it is possible to pull the Tygan off in the appropriate place and reposition it slightly—but every effort should be made to get it right first time nevertheless.

Finally, the heat treatment recommended by Mr. Briggs in Reference 7 should be applied—a bar-type electric radiator 'should be held about six inches away from the mesh for about five seconds, when the heat begins to contract the fibre. Remove the radiator *immediately* a slight movement is seen in the Tygan, otherwise excessive contraction will be induced'.

Before finally mounting the speaker unit, check that the coil leads are correctly positioned and in no danger of rattling against the diaphragm or the speaker chassis.

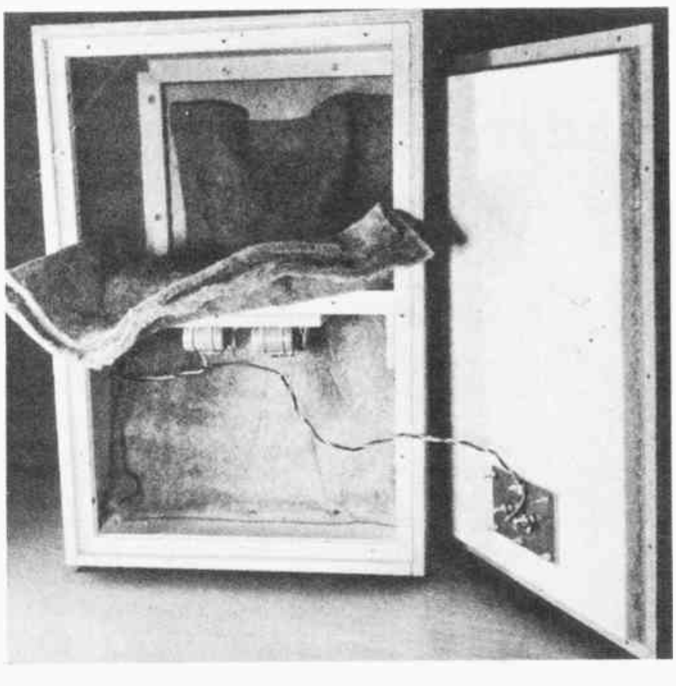
Wooden strips of cross-section 1-in. × ½-in. should be screwed and glued edgewise on, using 1¼-in. No. 8 countersunk screws, to the insides of the cabinet sides, top and bottom at a distance from the front edges sufficient to accommodate the thickness of the speaker mounting board after it has been fitted with expanded aluminium or Tygan. This distance is ⅜ in. in the author's cabinets, but it is as well to tailor it to suit the speaker boards as made. The aim is to make the latter an easy sliding fit between the 1-in. × ½-in. strips and the inside of the front mitred mouldings when later fitted. The board is secured in place by screws into it through holes drilled in the 1-in. × ½-in. strips.

The ½-in. × ½-in. strips shown may then be screwed and glued suitably in place and the carcase of the cabinet assembled. The bottom, however, should not be glued in position, so that it can be removed by undoing screws only, as previously mentioned. The strips should, of course, be spaced from the back of the cabinet by an appropriate amount, to accommodate the thickness of the back cover.

The cabinet back could be made of ½-in. plywood or Weyroc like the sides, but the author used a slight modification of a B.B.C. recipe[1], ¼-in. hardboard being glued to ½-in. builder's insulating board, as shown in the photograph on the right. This gives a composite board which is considerably lighter than wood of the same thickness and which also possesses desirable self-damping properties.

The choice of cross-section for the mitred front mouldings of the cabinet exerts a subtle effect on the appearance, and can be left to individual preference. The moulding should be glued to the carefully planed front edge of the cabinet, but a few 1-in. or 1¼-in. panel pins will make it much easier to position the mouldings nicely and ensure that they remain properly positioned while the glue sets. The panel pins should be punched down below the wood surface, and each hole filled with plastic wood made by mixing a drop or two of Durofix with plenty of wood dust obtained from sandpapering a nearby part of the same moulding. Allow to dry very thoroughly before finally sandpapering flush—the pin positions should then be almost invisible.

Components of cabinet, showing removable base, and the elliptical loudspeaker unit.



Interior of completed loudspeaker, with the two-layer felt 'curtain' partly removed to show the felt 'hat' over the drive unit. The removable back (right) is made from hardboard and insulating board.

The moulding attached to the bottom of the cabinet should not be glued near its corners, otherwise the bottom will cease to be easily removable.

The $\frac{3}{8}$ in. thick 'shelf' provides a firm anchorage for both the sides and the back of the cabinet (the back being screwed to the edge of the shelf) and thus reduces the tendency for these parts to vibrate in 'drum' fashion.

There are obviously various possibilities for the signal connections. The author made a rectangular cut-out in the back of the cabinet, $2\frac{1}{2}$ in. × $1\frac{1}{4}$ in., and fitted behind it a $\frac{1}{8}$ in. s.r.b.p. ('Paxolin') board carrying two nickel plated 2 B.A. screw terminals —all available from RS Components Ltd. through radio dealers.

It is important to keep the cabinet reasonably free from air leaks. One easily overlooked source of leak can arise when expanded aluminium is used, if it is bent right round the edges of the speaker mounting board and onto the back surface of the board. Even though the board is held tightly by screws against the 1-in. × $\frac{1}{2}$-in. strips fixed to the cabinet sides, there is nevertheless an air leak round the edges of the board through the interstices of the expanded aluminium. It was found experimentally that the diaphragm displacement at 40 Hz was reduced *several times* on sealing this leak, and there must, of course, be an accompanying reduction in intermodulation distortion. There is a good case, therefore, for cutting the expanded aluminium only $\frac{1}{2}$ in. larger than the speaker mounting board all round, and fixing it with tacks into the edge of the board, thus obviating the leak.

After finally assembling the cabinet, with the unit in place, about 30 in. of red-and-black flex should be soldered to the speaker tags, of which one is marked red by the makers. One piece of ordinary carpet felt, about $\frac{3}{16}$ in. tick and measuring 14 in. × 11 in.†† should now be tacked in place with six tacks spaced out round the unit, producing a sort of roughly fitting felt hat over the unit. This will provide considerable damping of the low-frequency resonance and will consequently reduce the acoustic output in the 100 Hz region. Without it, there may be a slight tendency towards coloration of male speech. With two thicknesses of felt tacked down more closely with a larger number of tacks, the bass response will be decidedly thin. If no woofer is to be used, some constructors may prefer the compromise of omitting this felt cover altogether—speech may then sound a little too full in the bass, but the musical reproduction may be thought better.

After dealing with the above, a 'curtain' made from two pieces of approximately $\frac{3}{16}$ in. carpet felt, each measuring about 19 in. × 13 in., should be tacked loosely in place inside the cabinet as a sort of diaphragm dividing the space into two halves, with the loudspeaker unit in one half.

The equalizer should now be screwed to the shelf and wired in series with one of the leads from the unit to the terminal board. The terminal connected to the red lead should be marked appropriately if stereo operation is envisaged.

The back should be thoroughly screwed on, using three screws along each edge plus three more along its middle to fix it to the edge of the shelf.

Finally, four rubber feet, available from most hardware shops, may be screwed to the bottom of the loudspeaker—or a piece of felt may be stuck on if preferred.

†† The pieces used by the author weighed approx. 1.6 oz.

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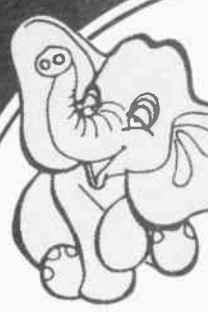


Low-cost High-quality Loudspeaker

2. Using the loudspeaker: Determining the electro-mechanical constants of the drive unit

by P. J. Baxandall, B.Sc. (Eng.), F.I.E.E., F.I.E.R.E.

An amplifier capable of giving a sine-wave mean output power* of about 10 W into a 15-Ω resistance load is quite sufficient for operating the low-cost loudspeaker. The use of higher output levels will give gross distortion and may cause permanent damage to the loudspeaker.

Measurements on the loudspeaker (described later in the article) yield an efficiency figure of about 1%, so that a 10-W amplifier is capable of producing from the loudspeaker a sine-wave mean acoustic power output of about 100 mW. From reference 8, this would be expected to give, in an average living room of, say, 1,500 cu ft, a sound intensity nearly 100 dB up on the standard reference level of 2×10^{-5} newton/sq metre r.m.s. This is about the intensity experienced in a good seat in a concert hall during very loud climaxes of orchestral music, though many musical people choose to listen at considerably lower levels at home. (An independent check with a microphone of known sensitivity, at 1 metre on axis out of doors, also gave an intensity of about 100 dB up on 2×10^{-5} N/m² for full output from a 10-W amplifier.)

Comparison in the author's living room of the measured sound intensity of a grand piano, and of the reproduction of a recording of this piano via the low-cost loudspeaker, showed that the loudspeaker, when driven by a 10-W amplifier (not overloaded), could produce an intensity very nearly, but not quite, equal to that of the piano itself when played at extreme fortissimo.

As mentioned earlier, the use of a woofer is really well worth while, adding depth and warmth to the reproduction, improving the naturalness of the balance and reducing listening fatigue.

In a mono system, the simple arrangement shown in Fig. 8 has been found very satisfactory, provided the woofer sensitivity is high enough. A suitable recipe for the choke is as follows:

Core, bobbin and shroud. Belclere Kit LX, in Silcor (1¼-in. stack of 0.014-in. laminations, Inter-Service No. 417, maximum dimension 2¼ in.) at £1.50.

Winding. 300 turns of 24 s.w.g. enam.

The choke gaps may be adjusted, using cardboard or other insulating material, to vary the output of the woofer, and Fig. 9 shows the effect of so doing on the voltage across a 15-Ω woofer mounted on a baffle board. The best setting may be determined subjectively.

* The author does not like the term 'r.m.s. output power', despite its almost universal use nowadays. The product of r.m.s. voltage and r.m.s. current is not r.m.s. power, but *mean* power.

Peter J. Baxandall held an amateur radio licence (2AZS) while at King's College School, Wimbledon, but became increasingly fascinated by the problems of low-distortion sound reproduction, electrical musical instruments, etc. He obtained his degree in electrical power engineering at Cardiff Technical College in 1942, later becoming a radio instructor in the Hankey Radio Training Course there. He moved in 1944 to T.R.E. (now R.R.E.), Malvern, and worked on microwave techniques for the first two years before joining F. C. Williams's team on electronic circuit research work. He has latterly co-operated with industry in developing transistor power amplifiers.

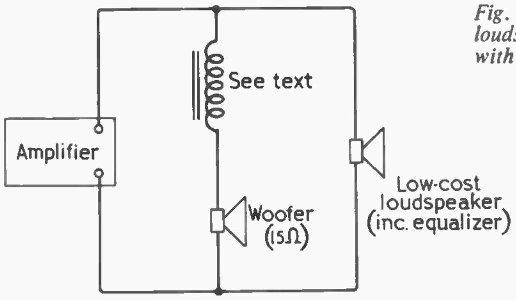


Fig. 8. Mono loudspeaker system with woofer.

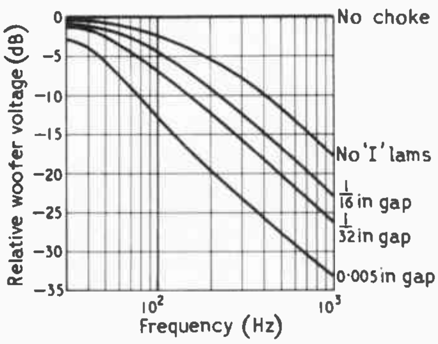


Fig. 9. Variation of woofer voltage with frequency for various choke gaps.

In a stereo system, the simplest arrangement is to feed a single woofer plus choke from the power amplifier of one channel only. This may seem very crude, because the woofer does not receive the proper sum signal, but the fact remains that it is fairly satisfactory in practice. It is conceivable, of course, that one might come across a stereo record with nearly all the low bass in the channel not feeding the woofer, but the author has yet to meet such a case among classical records.

If, however, the above simple solution does not seem attractive, there is more than one possible way of feeding the woofer with a genuine sum signal. A very satisfactory method is to connect the woofer plus choke between the live output terminals of the two power amplifiers and introduce a simple unity-gain phase-inverting stage at a suitable point in one channel, probably between the control unit and the power amplifier. The connections to one of the smaller loudspeakers must, of course, also be reversed, to restore the loudspeaker outputs to their correct phasing. (The alternative scheme of reversing the connections to one half of the stereo pickup is not such a good idea as it might at first sight appear to be—quite apart from the fact that it does not cater for the stereo radio aspect of the problem. This is because, while it will work all right under stereo conditions, the normal control unit switching arrangements connect the two halves of the pickup in parallel under mono conditions, giving, ideally, zero output if the connections to one half have been reversed.)

An incidental advantage of feeding the woofer from the two power amplifiers as just described is that the signal level is 6 dB higher than when fed from only one amplifier, assuming the two amplifiers give in-phase contributions to the woofer at these very low frequencies.† If the bass output is too powerful, it may, of course, be reduced by decreasing the choke gap, but it is probable that the extra output at these very low bass frequencies will be felt to be beneficial in most rooms.[1]

An alternative and equally effective method for obtaining the sum signal for feeding the woofer is shown in Fig. 10. The transformer can employ a Belclere Kit LX, as used for the woofer choke, each winding consisting of 150 turns (in one section) of 24 s.w.g. enamelled wire. The laminations should be interleaved, i.e., no gap.

The connection of a transformer, as in Fig. 10, directly across the output of some transistor amplifiers of the type having no output transformer, is, however, inadvisable, being likely to lead to the breakdown of one or both of the output transistors should an accidental very-low-frequency overload occur—caused, for example, by mishandling the pickup. This is because the transformer inductance presents the transistors, at very low frequencies, with a low value of almost purely reactive load, giving an instantaneous combination of high collector current and high collector voltage not met under more normal load conditions. No trouble with such breakdown effects is likely to be experienced with valve amplifiers, however.

An economical scheme, which feeds the woofer with a genuine sum signal without requiring a transformer, and which can be used quite safely with a transistor amplifier, involves connecting two resistors, of about 15 Ω each, in series between the live output terminals of the two power amplifiers, the woofer and its series choke being fed from their junction. While this arrangement draws extra power from the amplifiers, and gives reduced electromagnetic damping of the woofer, it has been found to work quite nicely in practice. The resistors should preferably be wire-wound, with a rating of at least 3 W each.

Yet another arrangement, which may be favoured by readers possessing a spare mono amplifier, is shown in Fig. 11. It is here assumed that the amplifier input impedance is at least 100 kΩ. If it is lower than this, the impedance values in the circuits should all be reduced appropriately.

The circuit shown in Fig. 12, which is believed to have been used on the Continent, has the advantage of requiring neither special iron-cored components nor an extra amplifier. It cannot be strongly recommended, however; the resonant interaction of the motional impedances of the three loudspeakers leading to peculiar dips and peaks in the frequency response which are difficult to predict or control satisfactorily.

A system employing a single woofer operated off the sum signal can give considerably less turntable rumble than a normal system, and this is a very real advantage. Rumble vibrations tend to be largely in a vertical plane, and the single woofer is non-responsive to vertical stylus movements, which normally represent the stereo difference signal. When two separate full-range loudspeakers are used, vertical stylus movements give, ideally, equal and antiphase outputs from the loudspeakers, but, because of acoustic effects in the room, the outputs do not, in general, cancel at the listener's ears—indeed, if they did, there would be no stereo effect! The loss of stereo effect inherent in the use of a single woofer does not seem to matter, provided, as in the present scheme, it is confined to very low frequencies only.

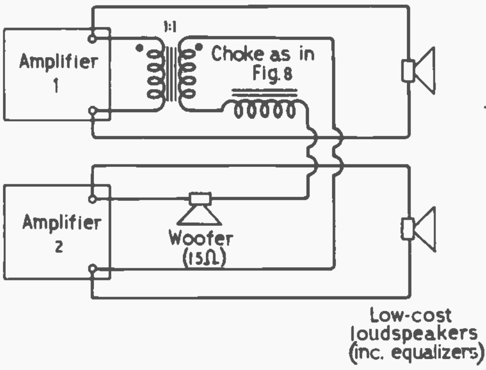


Fig. 10. Stereo system with phase-inverting transformer for feeding woofer.

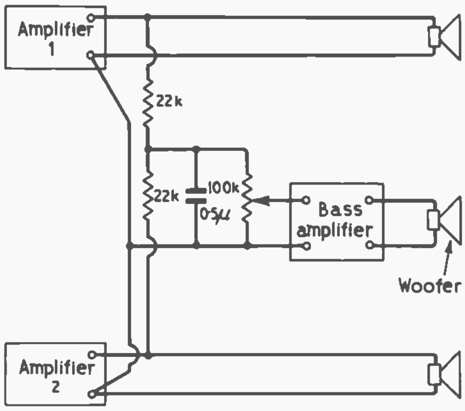


Fig. 11. Stereo system with separate amplifier for feeding woofer.

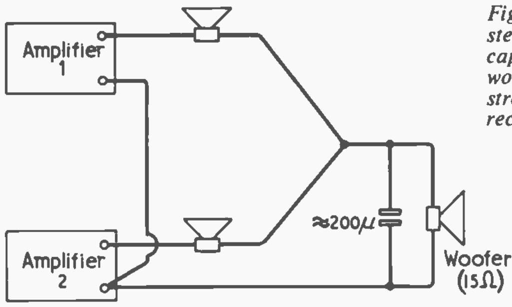


Fig. 12. Simple stereo system with capacitor across woofer—not strongly recommended.

Measurements on the drive unit

The electro-mechanical constants of the loudspeaker drive unit were determined by the following set of measurements and calculations.

The unmounted loudspeaker unit was placed face upwards on a table and fed at low level from an oscillator via a 1,000-Ω series resistor. An oscilloscope (10 mV/cm sensitivity) connected across the speech coil enabled the oscillator to be set to the diaphragm resonant frequency, f_0, as indicated by a maximum waveform amplitude. A series of ordinary brass balance weights was then carefully placed on the diaphragm near the coil, giving modified values of resonant frequency. A graph of $(1/f_0)^2$ against the mass added was then plotted. This was a good straight line, with an intercept at -5.7 gm, so the effective diaphragm mass was taken to be of this magnitude. The total mass corresponding to a particular resonant frequency could then be obtained, enabling the diaphragm suspension compliance to be determined from the relationship:

$$f_0 = \frac{1}{2\pi\sqrt{MC}} \qquad (1)$$

where: M = mass in kilogrammes, C = compliance in metres per newton, f_0 = resonant frequency in Hz.

The value obtained for the compliance was 0.67 mm/N.

A little wire tripod was then made, whose feet rested on the junction between the central dome and the paper diaphragm. This tripod carried a fine pointer arranged to indicate displacement against a scale of millimetres fixed to a wooden 'bridge' resting on the loudspeaker frame. With care, it was found possible to estimate tenths of a millimetre. Measured direct currents were then passed through the coil, in both directions, and a graph of displacement against current plotted. This was fairly straight, with a slope of 4.5 mm/amp, up to about ± 1 mm. From this figure and the compliance figure previously obtained, a force/current relationship of 6.7 newtons/amp was deducted.

The aim was to obtain all the values of the mechanical elements

† An objection to the scheme might seem to be that each amplifier will 'see' a load impedance of only half the woofer-plus-choke impedance. However, because the impedance of a nominally 15-Ω speaker is much higher than 15 Ω in the region of its resonant frequency, except, perhaps, if it is mounted in a phase-inverter cabinet, and because of the high impedance of the choke at higher frequencies, it is found in practice that there is little reduction in the apparent power-handling capability of the system.

represented by the analogous electrical circuit of Fig. 13[8,10]‡ The suffixes D and C are used for quantities associated directly with the diaphragm and cabinet respectively. F is the force which would be produced by the speech coil if it were prevented from moving. This force, as already mentioned, is 6.7 N/A. If E is the amplifier output e.m.f. in volts, and if we take the sum, R_{tot}, of the coil resistance (13.5 Ω) and the amplifier output resistance as 15 Ω, then $F = 6.7 \times E/15$, i.e., $F = \times 0.45$ newtons.

R_{EM} in Fig. 13 is the mechanical resistance introduced by electromagnetic damping, i.e., it is the ratio of the force produced by the speech coil when blocked to the velocity if would have if completely free and massless. It may be shown that:

$$R_{EM} = \left(\frac{F}{I}\right)^2 \times \frac{1}{R_{tot}} \qquad (2)$$

where: F/I is in N/A, R_{tot} is in ohms, R_{EM} is in m.k.s. mechanical ohms.

For the present loudspeaker unit, R_{EM} comes out at 3 m.k.s. mechanical ohms.

C_C is the compliance associated with the volume of air enclosed in the cabinet, referred to the diagram. It is obvious that the larger the diaphragm area, the greater will be the increase of air pressure in the cabinet for a given diaphragm movement, and that a given increase in pressure will produce a force on the diaphragm proportional to its area. Hence:

$$(\text{Air compliance})_{\substack{\text{w.r.t.}\\ \text{diaphragm}}} \propto \frac{\text{cabinet volume}}{(\text{diaphragm area})^2} \qquad (3)$$

(An alternative method of calculation involves the use of acoustical impedances rather than mechanical impedances. Acoustical impedance is pressure/volume-current rather than force/velocity. Compliance is then volume change/pressure and is a function of cabinet volume only and not diaphragm area.)

When the air in the cabinet is suddenly compressed, its temperature rises. If the increased pressure is maintained, the air will cool down again, giving a further volume reduction. Thus, for very slow changes, the compliance is higher than for faster changes. When, as normally applies for a loudspeaker cabinet, even at low audio frequencies, there is no time for the air to cool down after compression, the operation is said to be adiabatic, as compared with isothermal for very slow changes.

The effective diaphragm area for the Elac unit used, measured to the 'mid-point' of the surround, is approximately 0.0164 sq m. The effective cabinet volume is approximately 0.0252 cu m. This leads to the result that, with adiabatic operation, the compliance of the air, referred to the diaphragm, is 0.70 mm/N.

It will be noticed that the compliance due to the air is about equal to that of the unit itself, giving a rise in resonant frequency by a factor of about $\sqrt{2}$. The calculated resonant frequency in the cabinet is 114 Hz, which agrees quite reasonably with that determined experimentally. (A complicating factor is that it is found in practice that the resonant frequency of the unit depends considerably on the applied voltage at which it is measured.)

The remaining element to be determined in Fig. 13 is the mechanical resistance R, representing diaphragm suspension losses and radiation resistance. The latter varies rapidly with frequency, but the value of R at the resonant frequency is of particular interest. One of the simplest methods for determining R is to connect the contacts of a relay in series with the speech coil and a d.c. supply such as a dry cell. The relay is operated at some quite low frequency, e.g., 1 Hz, by means of a multivibrator, or in some other convenient way. When the contacts open, the electrical resistance in the speech-coil circuit becomes infinite, making R_{EM} zero. The only resistance effective in the mechanical circuit is then R, and a damped oscillatory voltage appears across the coil, as shown in Fig. 14. The Q value may be determined from the rate of decay of the oscillation, and a convenient fact is that the Q value is equal to the number of *half* cycles that occur while the oscillation amplitude is decaying from a value of unity to a value of 0.21 of

‡ The analogy used is that voltage represents force, current represents velocity. Hence voltage/current, i.e., electrical impedance, represents force/velocity, i.e., mechanical impedance. Just as the reactance of an inductance L is $2\pi fL$, so the mechanical reactance of a mass M is $2\pi fM$ mechanical ohms, etc.

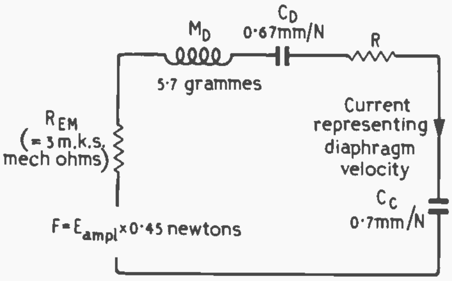
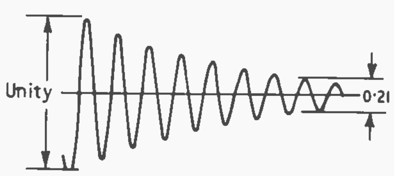


Fig. 13. Analogous electrical circuit representing mechanical system of loudspeaker unit and cabinet.



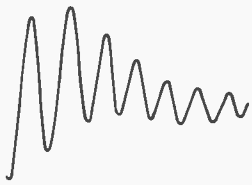


Fig. 14 (Left). Open-circuit coil voltage following interruption of current.
Fig. 15 (Right). As for Fig. 14, but with air leak in cabinet.

unity. This test performed on the present loudspeaker, with no felt in the cabinet, gave a Q value of 15 and a natural frequency of 110 Hz. The reactance of the 5.7. gm diaphragm mass at 110 Hz is 3.9 m.k.s. mechanical ohms, so that, with $Q = 15$, R is $3.9 \div 15$, i.e., 0.26 m.k.s. mechanical ohms. Thus, when the unit is fed from a *low* impedance source, the total mechanical resistance is 3.26 m.k.s. mechanical ohms, and the Q value is 3.9/3.26, i.e., 1.2.

It was interesting to observe that, while a nice simple exponentially damped sine wave was obtained with the cabinet properly sealed, quite a small leak, such as that mentioned earlier, caused by incorrectly-fitted expanded aluminium, converted the waveform to a much more complex one, somewhat as sketched in Fig. 15.

The question now arises as to how much of the above 0.26 m.k.s. mechanical ohms figure for R is caused by radiation resistance. The diaphragm area of 0.0164 sq m is the same as for a circular diaphragm of radius 7.2 cm. The radiation resistance seen by a diaphragm, which is 420 m.k.s. mechanical ohms per sq m at high frequencies, where the wavelength is small compared with the diaphragm radius, falls off inversely as the square of the frequency from a corner frequency at which radius/wavelength = 0.25. For a radius of 7.2 cm, the wavelength is thus 28.8 cm, corresponding to a frequency of 1170 Hz. Hence at any frequency, f, considerably lower than this, the radiation component of the mechanical resistance in Fig. 13 is $0.0164 \times 420 \times (f/1170)^2$, i.e., $5.0 \times 10^{-6} f^2$ m.k.s. mechanical ohms.

The acoustical power radiated is equal to the square of the diaphragm velocity times this radiation resistance—equivalent to '$P = I^2R$' in an electrical circuit. Hence the radiated power will be independent of frequency if the diaphragm velocity is proportional to $1/f$. At frequencies well above resonance in the Fig. 13 circuit, the reactance of M_D becomes the dominant mechanical impedance, giving a velocity of $F/2\pi f M_D$. As already discussed, $F = E \times 0.45$ newtons, M_D is 5.7 grammes. The radiation resistance is $5.0 \times 10^{-6} f^2$ m.k.s. mechanical ohms. Hence the power radiated is:

$$\left[\frac{E \times 0.45 \times 10^5}{2\pi f \times 5.7 \times 10^{-3}}\right]^2 \times 5.0 \times 10^{-6} f^2 \text{ W}$$

$$= E^2 \times 0.80 \text{ mW}.$$

A 10-W amplifier designed for a 15-Ω load will give an output voltage of 12.2 V r.m.s., so that, from the above, the acoustic output from the loudspeaker at 10-W level is 120 mW. Hence, for practical purposes, the efficiency may be taken as 1.2%.

In the Fig. 13 circuit, the power output is (current)^2 × (radiation resistance). But radiation resistance is proportional to frequency)². Hence power output is proportional to (current × frequency)². Now the voltage across the inductance is proportional to (current × frequency), and this leads to the useful idea, pointed out by D. E. L. Shorter in reference 2, that the output power is proportional to the square of this voltage, or the pressure produced by the loudspeaker in free space is proportional to the voltage across the inductance. In this context, the circuit may

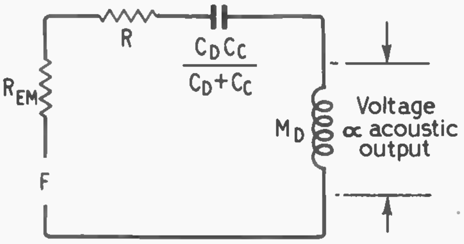


Fig. 16. Rearrangement of Fig. 13.

conveniently be redrawn as shown in Fig. 16. This is a well known circuit, whose normalized frequency response is given in reference 11. With a Q-value of 1.2, as determined above, the response would be expected to exhibit a peak of 2.4 dB just above the resonant frequency and to become asymptotic at very low frequencies to a 40 dB/decade (12 dB/octave) line going through 0 dB at the resonant frequency. The measured acoustic frequency response of the loudspeaker, Fig. 5 in Part I, will be seen to approximate fairly closely to this at low frequencies.

In conclusion, readers employing Vinkor for mounting clips for the equalizers may find it convenient to order mounting boards. These are DT 2233 for the larger core and DT 2227 for the smaller core. If, however, the whole equalizer is built on a piece of 1/16th inch s.r.b.p., holes for attaching the clips may be drilled in this and there is no need to employ Mullard boards.

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9. 'Analysis and Measurement of Programme Levels' by D. E. L. Shorter and W. I. Manson. B.B.C. Engineering Monograph No. 16, March 1958.
10. 'Dynamical Analogies' by H. F. Olson. D. Van Nostrand, New York *(1958)*.
11. 'Electric Circuits' by M.I.T. staff. John Wiley *(1943)*.

Electrostatic Headphone Design

Instructions for making a simple and inexpensive high-quality unit

by Philip D. Harvey, B.Sc.

The design described below, like that published in 1968[1], is based on the constant charge push-pull principle schematically illustrated in Fig. 1. The constant charge is derived by feeding the diaphragm from a high resistance *R*, and relying on the capacitance of the earphone to store the charge.

Basic requirements in construction are that:

1. the fixed plates be rigid, acoustically transparent, and both flat and conducting on the inner surface;
2. the spacers be flat, of uniform thickness and, above all, insulating; and that
3. the diaphragm be flexible and light.

In all, three models were constructed. In producing fixed plates for the final model the electro-mechanical analogy described in Appendix B was used.

Stroboscopic examination of an earphone had shown that the diaphragm behaves as an elliptical vibrating piston with major and minor axes set by the spacers. These dimensions were set at 75 × 45 mm to cover the ear. A short transmission "tunnel" is employed to improve low-frequency coupling with the ear. This extension is lined to reduce resonances.

The fixed plates are of single-sided copper-plated fibre-glass. Hole area is 30%—sufficient to ensure acoustical transparency without sacrificing rigidity. The holes must be deburred after drilling.

To remove the risk of charge leakage at the edges of the board and at the connecting bolt holes (due perhaps to tearing of the diaphragm and consequent shorting) about 2 mm of copper is removed from the edges of the board round the connecting bolt holes (see Figs 2 and 3) to prevent charge leakage should the diaphragm tear at the edges.

The spacers, made of polyvinyl acetate, are cut in one piece from a sheet to avoid poorly insulating joints. These are drilled, using the fixed plates as templates, and deburred.

To make a safe connection of high voltage leads, two methods can be employed for the outer plates:

(a) Alternate unrounded corners of each fixed plate are removed to allow a connection to be made to the other fixed plate. Plasticine can be used for insulating the connection. The principle is illustrated in Figs 4 and 5.

(b) A small hole may be drilled in one corner of the fixed plate, and the copper side of the board slightly countersunk. The insulation of the signal wire is then stripped off, the inner being tinned and fed through the hole, as shown in Fig. 6. The well, created by countersinking, is now filled with solder which makes good contact with both the wire and copper plating. By grinding this surface flat we have a good safe connection.

To insulate the diaphragm connection it was decided to utilize the insulating properties of both the fixed plates and the transmission tunnel. The connection was brought to the surface of one fixed plate by a brass bush as shown in Fig. 7. The connection was then made harmlessly between the tunnel and board.

The film for the diaphragm is prepared by taping it crease free over a wooden frame of inside dimensions 200 × 250 mm. The frame, with the film now flat and under tension on its upper surface, was placed over a sheet of glass 240 × 190 mm of slightly greater thickness than the frame. Under these conditions it was easier to rub Aquadag on and off the film. This should be continuous until surface resistivity is $10^8\,\Omega$. The prepared film is next mounted on one spacer using double-sided Sellotape with the resistive side exposed, and laid on to the other spacer and a fixed plate with the brass bush inserted. The brass bush

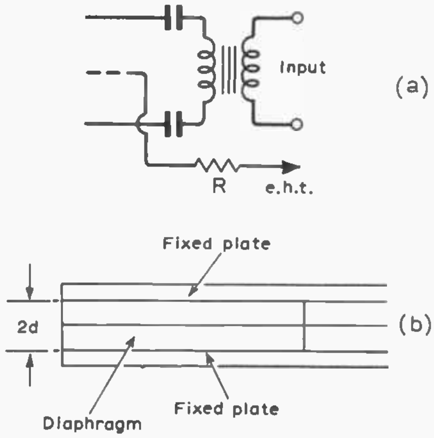


Fig. 1. Push-pull electrostatic sound generator.

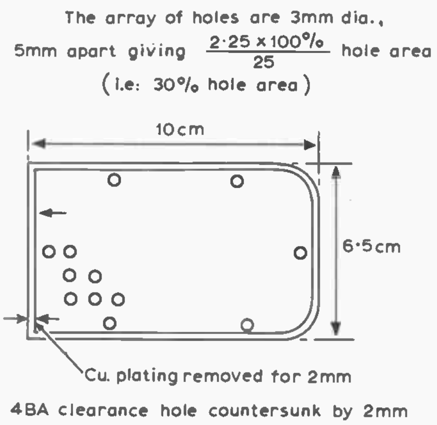


Fig. 2. Plan view of fixed plate.

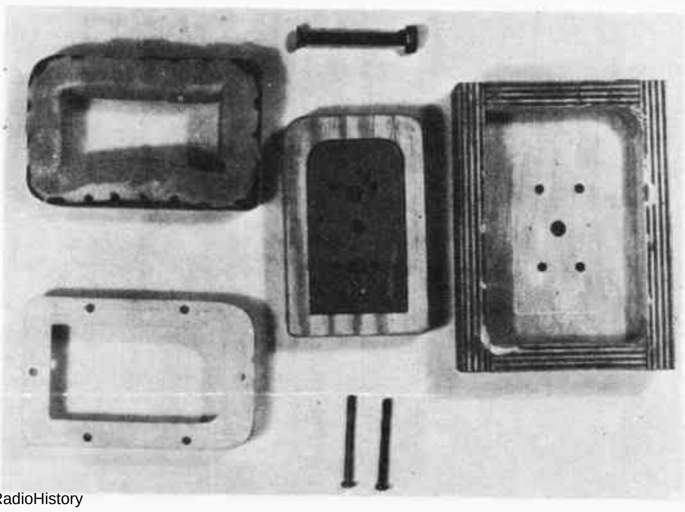


Fig. 3. Mould used for the transmission tunnel, and typical results achieved.

[1] 'High-quality Electrostatic Headphones' by J. P. Wilson, *Wireless World*, Dec. 1968.

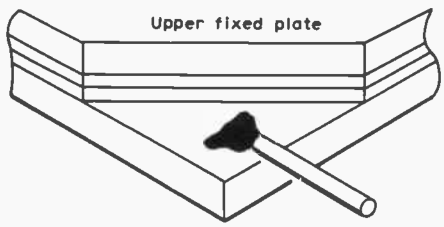


Fig. 4. One corner of final model.

now contacts the resistive coating, although it might be necessary to use some Aquadag on the contacting surfaces. The other fixed plate is laid on the assembly, followed by the transmission tunnel ready drilled, enabling the parts to be fastened together with nylon nuts and bolts. The components are shown in Fig. 5.

Before testing, the earphone is heated by warm air to tighten the diaphragm and remove any slight creases in it.

Transmission tunnel details

The transmission tunnel must be light and strong, and transmit the sound produced by the earphone to the ear. The simplest shape to do this is shown in Fig. 9. The only readily available group of materials to fulfil the above conditions is the plastics. These also have an advantage of damping incident sound, whereas metals tend to 'ring'.

The idea of casting the tunnel from polystyrene was investigated. Experiments led to the use of a wooden mould. It was found that if the mould was left overwaxed, then the excess wax was melted during the ensuing catalytic process, and this enabled the polystyrene to be removed from the mould whilst it was still pliable. Provided it was well supported whilst setting fully, the result was quite acceptable. Both the mould used (made of two parts for easier positive removal) and a typical positive are shown in Fig. 3.

Tunnels of both clear and coloured polystyrene were made, and it seems that the colouring material used gave the tunnel added strength.

It was found that latex foam rubber, used for lining the tunnel because of its excellent sound absorbing properties, was best cut on the bandsaw.

Variation of the other component elements

Under given conditions of signal and bias voltages, the two components affecting the earphone's performance are:

(a) The spacers—the thickness of which determine E and hence sound output. Spacer thicknesses of 0·18, 0·25, 0·37, 0·62 and 1 mm were tried. Decreasing the spacer thickness did not alter the frequency response but raised the sound level. Construction difficulties increased as spacer thickness decreased due to the slight and unavoidable warping of the fixed plates. This did not become too bad until ionization of the air was also a problem (see below).

Silicon resin bonded paper, paxolin, and dry paper were also tried as spacer materials. No difference was observed in the performance and it is concluded that any material having a resistivity greater than $10^{10}\,\Omega$ cm would be satisfactory.

(b) The diaphragm—through which no appreciable current should flow in less than half the time period of the lowest frequency to be reproduced. This ensures constant charge conditions. If one assumes the diaphragm to be perfectly conducting, and the earphones to have capacitance C farads, and further that the lower limit of audibility is 27 Hz, then the diaphragm must be fed via a resistance R ohms, such that;

$$RC > \frac{1}{2 \times 27} \text{(approx.)}.$$

C is calculated as 330 pF from $C = \frac{\epsilon A}{d}$

$$\text{whence } R > \frac{1 \times 10^{12}}{54 \times 330} \text{ i.e. } R > 6 \times 10^{7}\,\Omega.$$

Due to the high value of this resistance it is easier to make the diaphragm resistive than feed it through an external resistance. Experiments were made with sheets of $10^7\,\Omega$ surface resistivity and greater. As expected the bass response improves as the resistance increases. The high-frequency

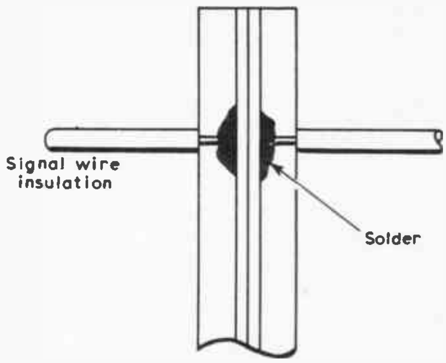


Fig. 6. Cross-section of alternative final model.

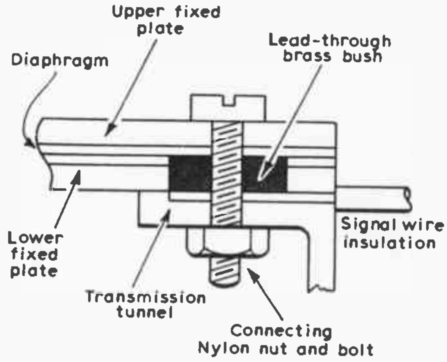


Fig. 7. Cross-section through connection to diaphragm.

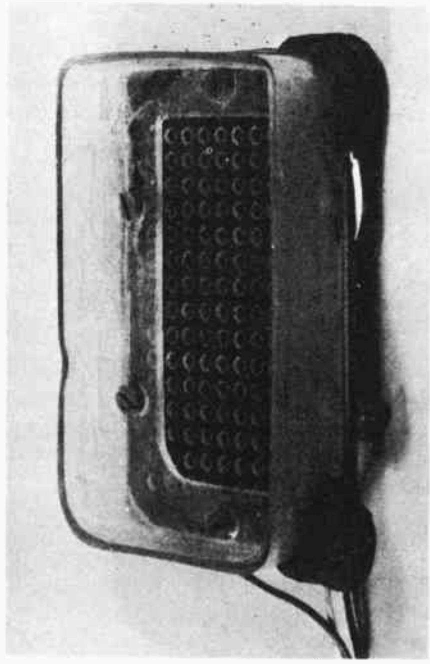


Fig. 8. The completed final earphone.

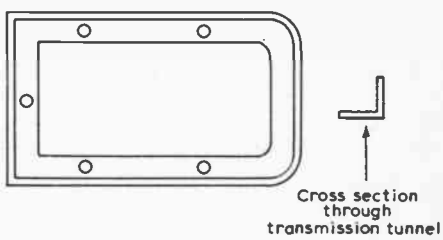


Fig. 9. Basic transmission-tunnel shape.

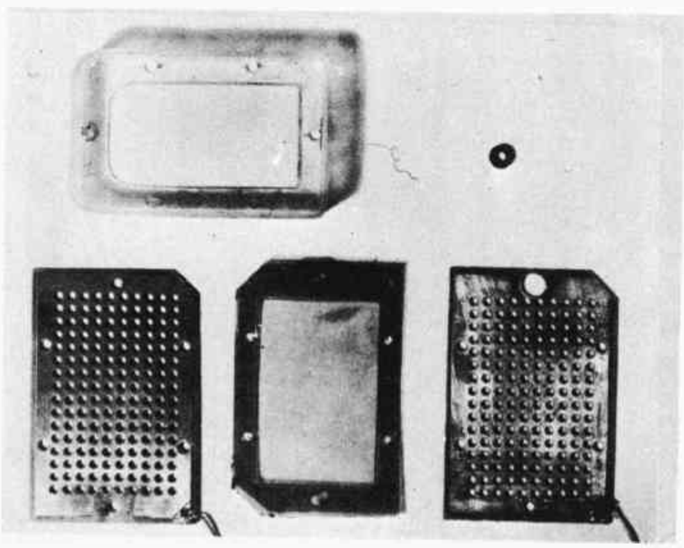


Fig. 5. Component parts of the final model.

response also improves, due presumably to the lower mass resulting from less graphite on the film. As some charging current must flow on to the diaphragm there is some limit to how high the resistance can be. Best results were obtained at the limit of measurability, i.e. a surface resistivity of approximately $10^9\ \Omega$.

Hospital anti-static polythene was tried and though it worked, the type available was thick and heavy, with a surface resistivity of only $10^5\ \Omega$. Hence both high and low frequencies suffered.

Various materials of the same type (Vitafilm) were obtained from local supermarkets. These were analysed spectroscopically and found to be the same material with the exception of that supplied by Sainsbury's. Microscopic analysis then showed that Vitafilm because of its porosity was not very suitable. The film made by The Borden Chemical Company was judged to be best closely followed by that made by Filmco in Durham.

Further tests to discover how best to apply homogenous resistive coating to the film were made on Borden's film. The use of evaporation techniques were first studied, but these posed three problems. In the conventional evaporator the film surface exposed was not large enough for an even film to be deposited over a sufficiently large area. Also at the low temperature required (not to destroy the film) oxidation of the depositing metal occurred. Finally when a film was deposited the metal permeated the plastic, altering its properties such that it became brittle and unusable.

Dry graphite powder rubbed into the surface did not alter its resistivity, presumably because the particles did not interlink and form molecule chains.

Finally a method was considered whereby a conducting medium could be sprayed as a solution in a liquid that would attack the film and hence give a permanently resistive surface. Graphite does not readily dissolve in any p.v.c. solvent, and so could not be used. A solution of silver in methyl acetate (Silver Dag) was sprayed on to a film, soaped to lower surface tension. The results were encouraging but a less active solvent would have to be used. Before pursuing this method, diluted Aquadag was substituted for Silver Dag and found to leave a completely uniform layer of graphite on the film when dry. Although this coating could be made fairly thick its resistivity remained immeasurably high until it was rubbed. Experience soon showed the amount that had to be sprayed for the required resistivity.

Drive circuits

Provisional model

The circuit shown in Fig. 10 employs the output stage of a commercial valve amplifier. The surface resistivity of the diaphragm must be greater than $10^8\ \Omega$ and hence the $10^7\ \Omega$ resistor in the feed line to the diaphragm is not necessary, but an added safety precaution.

It was found that the 0·01 μF isolating capacitors were sufficiently leaky to allow the outer plates to attain a high voltage, and the diaphragm could be earthed as an alternative form of bias. This makes the diaphragm an effective negative charge. This is not desirable because a steady high voltage on the outer conducting plates could be dangerous.

With the earphones in the circuit as shown, distortion was apparent, even at low acoustic levels. This was thought to be due to the output transformer. This amplifier was not designed to operate at maximum output continuously, and under these conditions inter-modulation distortion sets in. The earphones require a high voltage signal, but very little current. With this in mind an amplifier to deliver a distortion free signal was designed.

Designed valve amplifier

With a spacing of 0·37 mm (which changed by only 10% at full bass output) the maximum permissible voltage between the diaphragm and either fixed plate, to avoid ionization of the air between them, is 1000 V. With 300 V on the diaphragm this means that the maximum peak-to-peak voltage level on one plate can be 500 V. This leaves a large margin of safety for humid days or signal surges. The circuit of Fig. 11 was used giving only 400 V peak signal, as the valves and components were readily available. It gave no distortion observable on an oscilloscope, even without negative feedback, due presumably to the light loading on the amplifier.

Its use gave immediately discernible improvement in output level and fidelity.

Designed transistor amplifier

40 V rails are commonly available on transistor amplifiers and the circuit of Fig. 12 was built giving 32 V peak signal. Using 300 V rectified mains on the diaphragm gave a barely audible output.

The circuit of Fig. 13 was designed to give 300 V peak output. Any n-p-n silicon transistor with a $h_{fe}>50$ at 1 mA and a

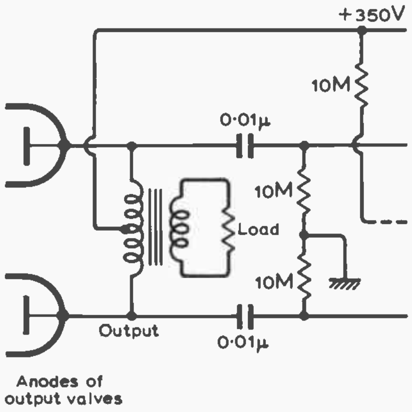


Fig. 10. Modified output of a commercial valve amplifier.

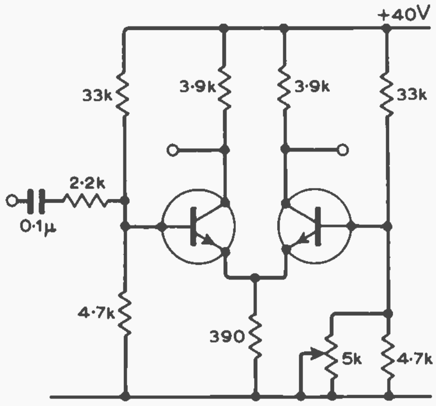


Fig. 12. Differential amplifier providing 32 V output.

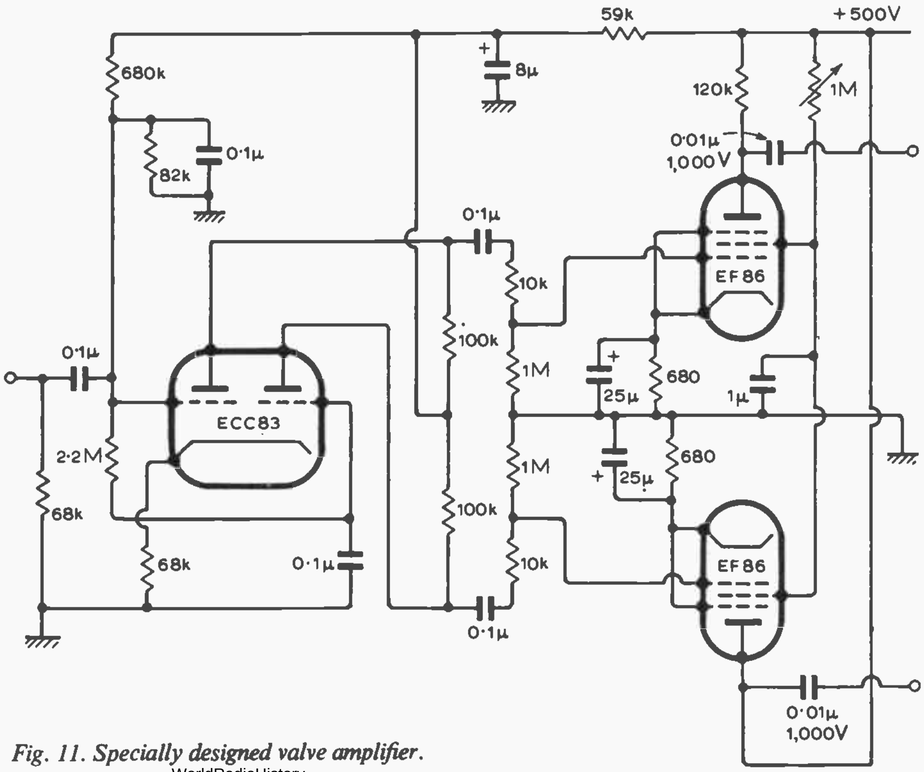


Fig. 11. Specially designed valve amplifier.

$V_{ce} > 35$ V will do for the first stage. The transistors in the differential stage should preferably be matched.

Three potentiometers are included to set up the amplifier to its optimum performance. First use R_1 to match the base voltages of Tr_1 and Tr_2; then adjust R_3 to make the average collector voltage of Tr_3 and Tr_4 115 V. Finally, using R_2, balance these collector voltages; repeat this procedure until both Tr_3 and Tr_4 collectors are at 155 V.

Measurement and analysis

From the section below and Appendix A the optimum of all the variables may be found. Although the thinner the spacers used the more the acoustic output obtained, it was found with the thinner ones (0·18 and 0·25 mm) that the air ionized on more humid days. This was apparent as a clicking noise, varying in repetition rate from one to ten hertz. It arose because constructionally the fixed plates are never equidistant from the diaphragm, and the air between the diaphragm and closest plate ionizes first. This allows attraction to the other plate increasing E, so that air here ionizes while the other reconstitutes itself. This effect is eliminated by reducing the voltage on the centre plate, but this necessarily reduces sensitivity.

The 0·37 mm spacers were therefore chosen and a plot of output versus central electrode potential revealed a levelling off at about 600 V. This is unexplained, but below this value the measured output is very near to the calculated value.

Many listeners were satisfied with volume and fidelity using 350 V on the diaphragm and the designed valve amplifier. There were many comments on the "depth" of the sound, which is due to the fact that plane waves are arriving at the ear, and these are normally associated with a distant source by the hearing mechanisms. When in use on a stereo system this effect makes it easier to identify the direction from which the sound appears to come.

Results achieved

Traces of the frequency responses are given with markings of 10 dB intervals and at the frequencies 20 Hz, 100 Hz, 200 Hz, 1 kHz, 2 kHz, 10 kHz and 20 kHz.

Fig. 14 gives the responses with different input signal voltages. The effect of increasing this voltage should be the same as decreasing spacer thickness. The relative graphs show this to be true, though the relative amplitudes differ.

Fig. 15 displays the difference made by altering the potential on the centre electrode.

Fig. 16 displays the difference in characteristic responses when plotted in the open air, and when plotted in the artificial ear.

Fig. 17 shows the best response achieved and corresponds to all the variables being optimised. The component specification for this is:

spacers—polyvinyl acetate 0·37 mm thick;

diaphragm—Borden Chemical's plasti-

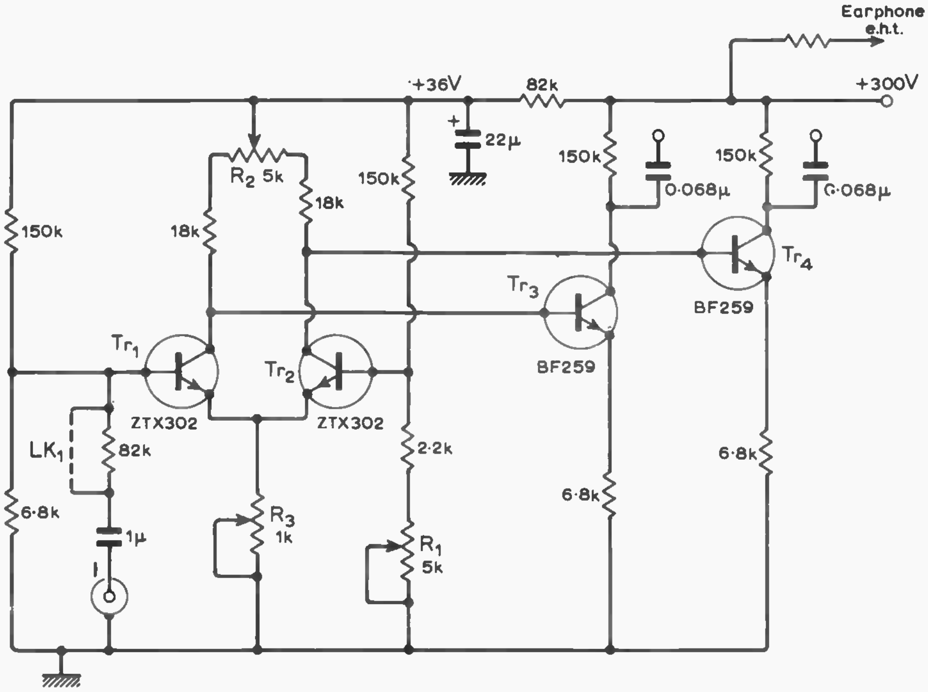


Fig. 13. Suitable transistor drive amplifier providing 300V peak output.

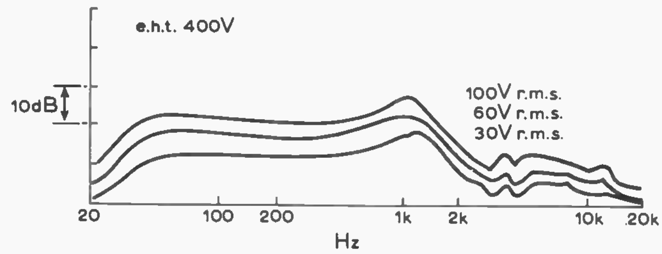


Fig. 14. Response for different signal-voltage levels.

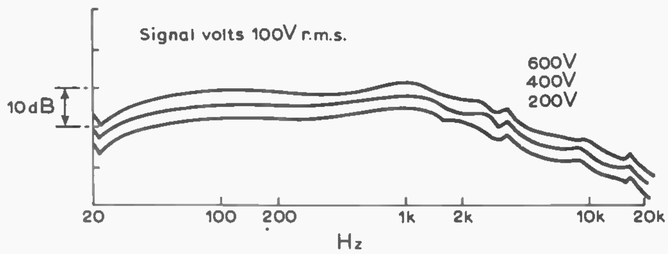


Fig. 15. Response for different diaphragm voltages.

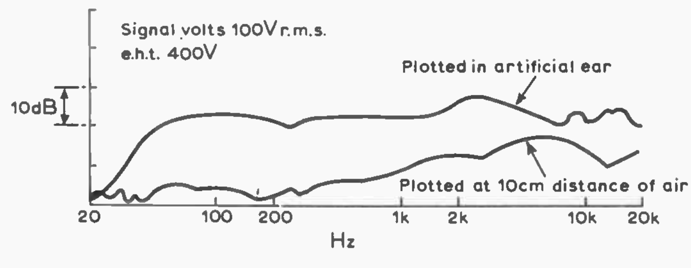


Fig. 16. Comparison of response in open air to that in artificial ear.

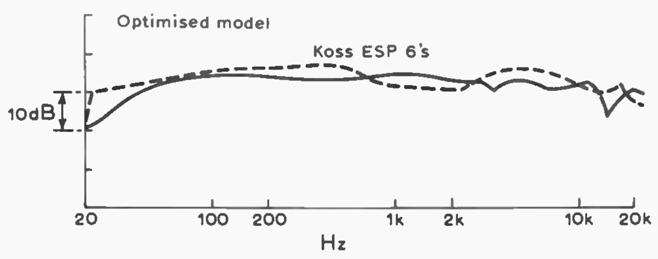


Fig. 17. Comparison of best earphone constructed with the published response of a Koss ESP6 unit.

cized p.v.c. sheeting 15 μ in thick, sheet resistivity $10^9\ \Omega$.

Safety

There are no uninsulated connections carrying high voltage near to the ear. Provided the connections at the signal generator are also well insulated, there is no danger of a fatal shock. There is always the danger of the diaphragm splitting, but even if it were to lacerate and protrude from a fixed plate, it would come up against the polyurethane foam the earphone is lined with. (This avoids cavity resonances in the sound conveyed to the ear, as well as insulating the ear.) If the diaphragm managed to touch the ear, then in the worst case at least $10^8\ \Omega$ on the film would allow only 3.5 μA to flow through the body, even assuming the body to be a dead short!

Suggested improvements

In order to achieve a broad frequency response it is essential to have slack suspension, and a low mass radiator. The first has been achieved by the use of a diaphragm which can be under quite high stress on its own plane, whilst a relatively low force can cause deflection in a transverse direction. In this design the mass of the radiator is no more than that of a layer of adjacent air a few millimetres in thickness. This could further be reduced by using a film resistive by manufacture.

The effect of resonances in this particular shape of diaphragm has not been investigated as the response curve does not indicate trouble of this kind. Three final points are worth making:

(a) The behaviour of the charge on the film is still largely unexplained as is the levelling off of the response with greater than 600 V on the diaphragm;

(b) Double-sided boards which prevent warping, along with more sophisticated construction techniques, should yield a system of adequate acoustic output using much smaller signal and bias voltages; and

(c) The quantities of different types of distortion present could be measured. Results obtained and listening tests indicate their virtual absence at low sound levels.

APPENDIX A

Measuring the response of the earphone on the ear

Without elaborate equipment, such as a probe microphone, this is difficult to do. Furthermore the earphones under test were not always safe to wear. For these reasons the ear was simulated for the tests. Artificial ears are readily available, and commonly have a volume of 6 cubic centimetres. The volume enclosed by the transmission tunnel is nearly twenty times this, and the addition of the ear's volume makes little difference to its response. The B & K microphone used for the tests was one inch in diameter, about the same as the opening to the ear. The flat wooden plate used for holding the microphone was lined with polyurethane foam, to simulate the coefficient of reflection of the skin.

The conventional B & K frequency plotting apparatus was then set up, and a constant peak-voltage sine-wave output fed to one plate with the other earthed. The inner electrode is maintained at, say 400 V by an h.t. supply. The frequency is swept continuously throughout the audio range 20–20,000 Hz, synchronized to a chart recorder into which the output of the microphone amplifier is fed.

Measuring diaphragm surface resistivity

Apply 250 V d.c. across two electrodes one inch long and one inch apart. The current flow is measured. Sufficient accuracy was obtained by quoting the result as $P \times 10^N\ \Omega$, where both P and N are integers.

APPENDIX B

The electro-mechanical analogy

This is employed to determine the output expected from the earphones, and the frequency response expected. The calculations performed assume values either already determined for the final model or values of the materials readily available.

Fig. 18 gives the equivalent mechanical circuits of the earphones, where the mass m is the mass per unit area of the diaphragm. The spring S is the suspension of the diaphragm in the transverse direction. The damping, $2R_m$ in the centre frequency band, is due to the impedance of the air. F_o is the peak force per unit area on the diaphragm.

Employing the electrical analogy of this circuit gives us Fig. 19. The mass per unit area becomes an inductance of M henries. The suspension becomes a capacitance of S^{-1} farads. The damping becomes a resistance of $2R_m\ \Omega$, and the force a voltage of $F_o \sin \omega t$ V.

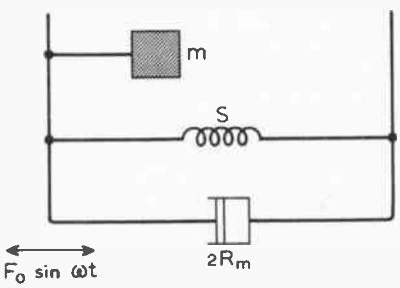


Fig. 18. Equivalent mechanical circuit of earphone.

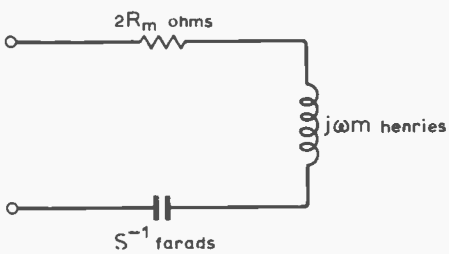


Fig. 19. Circuit given by the electro-mechanical analogy.

We know that:

$$M = 2{\cdot}4 \times 10^{-2}\ \text{kg}\ m^{-2}\ \text{(Vitafilm)}$$
$$2R_m = 2\rho c$$
$$= 820\ \text{Rayl in the mid-frequency band.}$$

S cannot be easily measured *in situ*, but a comparison with a conventional 4 inch loudspeaker indicated the same order of magnitude. It is calculated accurately knowing the free resonance to be at 55 Hz.

From Fig. 19 we know:

$$I = \frac{F_o \cos \omega t}{\left(2{\cdot}4 \times 10^{-2} j\omega + \frac{S}{j\omega} + 820\right)}\ \text{amps,}$$

and that at resonance I is real.

Hence
$$2{\cdot}4 \times 10^{-2} j\omega = \frac{S}{j\omega},$$

giving $S = 2{\cdot}4 \times 10^3$ newtons per metre. Because power $\propto$ current^2, the -6 dB points are given by

$$\frac{S}{\omega_L} = 820 \qquad \therefore\ \omega_L = 3$$

$$\omega_h m = 820 \qquad \therefore\ \omega_h = 35{,}000.$$

Therefore the -6 dB points are expected to be at ·5 Hz and 6000 Hz. In the region between these two points the movement of the plate is opposed only by the resistance of the air, so that the device is almost 100% efficient.

A light, thin material, such as that from the Borden Chemical Company considerably extends the flat response.

In order to determine the expected output, the equation $F_o = qE_o$ is utilised. The charge per unit area, q, is determined from the expression:

$$q = \frac{C \times V_{dc}}{\text{area}} = \frac{2\epsilon_o V_{dc}}{d},$$

where V_{dc} is the voltage applied to the diaphragm, and d is the thickness of the spacers:

Hence

$$F_o = \frac{2\epsilon_o V_{dc}}{\text{area}} \times \frac{v_o}{2d}$$

$$= 1{\cdot}95 \times 10^{-2}\ v_o\ \text{newtons per metre}^2$$

A loudness of 100 dBm is considered adequate, whence $F_o = 2$ newtons per metre^2.

This is achieved by signal voltages V_o of the order of 100 V in the region 6 kHz to 10 kHz. This is not a signal voltage sufficient to cause ionization of the air with 350 V on the diaphragm.

An I.C. Peak Programme Meter

by L. Nelson-Jones, F.I.E.R.E.

A design using standard i.c. operational amplifiers to achieve a transformerless design to the specification of the British Broadcasting Corporation, who pioneered this type of level indicator. Mono or stereo applications are catered for in the design, with separate or common meter indication. The circuit is stable against temperature and supply voltage variations, and is designed for use with a nominal 24V supply (16–30V). The main design aims were to obtain accuracy, stability, ease of law adjustment, and repeatability from one unit to another.

The peak programme meter dates back some 36 years when it was developed to provide a better means of measuring line levels in sound broadcasting than that provided by normal rectifier instruments such as the VU meter. In particular the instrument was given a slow decay and fast attack time to ease reading and lessen eye strain. Early designs were characterized by a very rapid response to transient peaks, but this was later modified since it was found that in practice the ear cannot easily detect the distortion produced by the clipping of very short duration peaks. The final attack time figure decided upon, and which is still standard, was 2.4 milliseconds. Such a response corresponds to a meter reading reaching 80% of peak using a square wave transient lasting 4 milliseconds. The decay time constant used is 1 second, which is a compromise between ease of reading and a response quick enough to record following peaks.

The graduations on the indicating meter were kept small in number and a black scale with white markings used to make for ease of reading. The basic scale division was chosen as 4dB, this being two steps of the standard B.B.C. fader controls. On a standard meter there are basically 7 divisions, with division 4 corresponding to 0dBm on a 600Ω line (0.775V r.m.s. sine wave, 1.095V peak).

The response of the peak programme meter (PPM) is approximately logarithmic and the divisions on the meter are approximately evenly spaced. The extreme divisions (1 to 2, and 6 to 7) represent a greater change than 4dB, namely 6dB. (Earlier meters differed in having all divisions except 1 to 2 equal to 4dB.) The present standard calibration together with the corresponding current in the meter are shown in Table 1.

The meter figures given are for B.B.C. Meter Specification ED1477, the one chosen for the design to be described.

In order to make good use of a fast charge time, the dynamic qualities of the moving-coil meter movement itself must be tightly controlled and considerably faster than that of normal movements. The meter must also be correctly damped to avoid large overshoots—two rather conflicting requirements. Whilst PPM circuits will work with standard meter movements the accuracy will be somewhat impaired unless the correct movement is used. In particular a circuit using a normal meter movement will, when set up on a standard tone level, tend to seriously underestimate short peaks on actual programme material.

Peak detection

In most previous PPMs a normal full-wave rectifier has been used, (Fig. 1), with a centre tapped signal transformer; the charge and discharge time constants being controlled by the two resistors *r* and *R*.

With the advent of integrated circuit operational amplifiers, however, one can now make an accurate peak rectifier without the need to use large voltage swings in order to overcome the forward drop of the rectifier, and the consequent non-linearity at low levels.

The basic circuit of such a peak detector is shown in Fig. 2. On a rising positive input, the output of the op-amp rises positively until the signal fed back to the inverting input of the op-amp via the diode *D* equals the level at the non-inverting input of the op-amp. When the input level falls, the diode *D* ceases to conduct as it becomes reverse biased, and the previous peak is stored on the capacitor *C* until such time as the input rises above the voltage to which the capacitor is charged, when the voltage on the capacitor again follows the input.

In practice the author has modified the basic circuit of Fig. 2 to that of Fig. 3. Apart from the two resistors *r* and *R*, to control the charge and discharge time constants, a transistor has been added to ensure adequate charging current availability. The practical values of the components are $C = 33\mu F$, $r = 75\Omega$, $R = 30k\Omega$. With such a large capacitance the peak charging current through the diode reaches approximately 100mA, which is well above the

Table 1.

PPM reading	Level dBm	Input voltage (peak)	Meter current (mA)
0	—	0	0
1	−14	0.220	0.10
2	−8	0.436	0.22
3	−4	0.690	0.35
4	0	1.095	0.51
5	+4	1.74	0.67
6	+8	2.75	0.80
7	+14	5.50	0.93
f.s.d.	—undefined		1.00

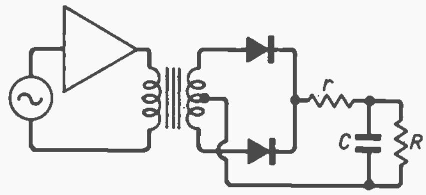


Fig. 1. Conventional PPM using centre-tapped transformer and full-wave rectifier.

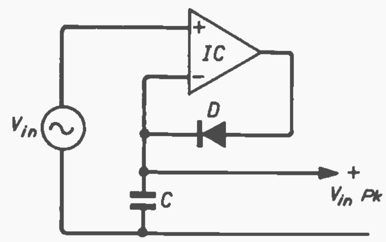


Fig. 2. Peak detecting circuit.

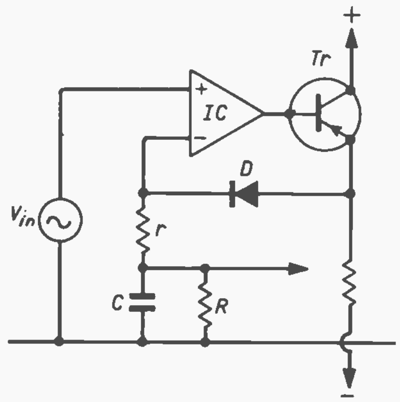


Fig. 3. Peak detecting circuit with time constants added.

capability of a normal i.c. op-amp, hence the additional transistor. (In practice r will be less than 75Ω due to the necessity of allowing for the forward impedance of the diode, and other components in the 'charging' path.)

The peak detector described operates only on positive peaks, whereas in a practical PPM it is necessary to measure positive and negative peaks equally, and to this end it is necessary to either (a) have a similar peak detector of reversed polarity to detect the negative peaks or (b) to have a second similar peak detector and precede it with a unity gain phase inverter.

It was decided to take the second course since it allowed the two positive peak detectors to be combined, sharing a common capacitor, charge and discharge resistors. In this way the highest peak from either detector will automatically be selected.

The unity gain inverter can of course be a centre tapped transformer as in previous PPMs, or else another op-amp connected as a unity gain inverter, as shown in Fig. 4.

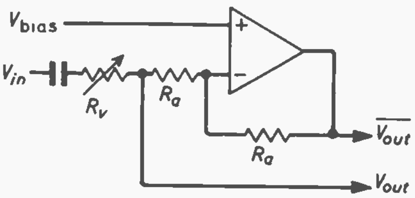


Fig. 4. Unity gain inverting circuit.

Provided that the loading on the two outputs is small, both the a.c. and d.c. levels will be equal except for the phasing. Since the input to the two peak detectors is the non-inverting input of two op-amps the loading is in fact quite low. The difference of d.c. level due to the unequal base supply resistances of the two peak detectors is approximately equal to the typical offsets of the i.cs, and is therefore fairly negligible when compared to the signal levels, i.e. they are less than 10% of the lowest division (1 on the PPM = 0.22V pk). In addition there is a zero set control on the output amplifier which can largely remove the effect from the meter deflection.

Gain adjustment is achieved by the single control R_V for both peak detectors. Whatever the value of R_V, V_{out} and $\overline{V_{out}}$ (Fig. 4) will remain equal and opposite to one another so far as signal excursions are concerned, although at the same d.c. level.

Law corrected output amplifier

The voltage across the peak storage capacitor is applied to a law corrected summing amplifier, whose input resistance (and hence the discharge time constant) will be set by an input resistor R_x to the summing point. The basic principle of this amplifier is illustrated in simplified form in Fig. 5.

The initial gain for voltages close to the bias voltage line (V_{bias}) that is from 0–3 on the PPM scale, is linear, and is set by the ratio R_y/R_x since for small output levels the transistor Tr_1 is reverse biased. When the emitter potential of Tr_1 falls below its bias voltage V_1 the additional feedback

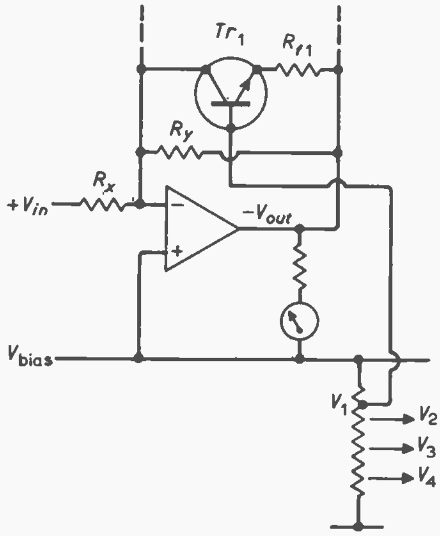


Fig. 5. Use of a transistor in the feedback path to provide law correction of transfer characteristic.

resistor R_{f1} is brought into operation in parallel with R_y so that the gain is reduced to

$$\frac{R_y \cdot R_{f1}}{R_x(R_y+R_{f1})}.$$

Further feedback resistors R_{f2}-R_{f4} together with transistors Tr_2-Tr_4 each controlled by bias voltages V_2-V_4 respectively, are similarly connected to the amplifier to successively reduce the gain with increasing negative output level. The law corrected amplifier therefore approximates the desired curve of input versus output with a five section linear gain curve as shown in Fig. 6. The choice of feedback resistors and bias voltages is made to get the best match to the actual smooth curve. In practice this was done by graphical methods together with calculation. The values were finally adjusted by trial and error to get the best result, together with the use of standard E24 values. The choice of values possible is almost infinite depending on the choice of break points.

Complete practical circuit

Fig. 7 shows the complete circuit of the peak programme meter based on the circuits described above. It is designed to work with a 1mA meter movement to B.B.C. Specification ED1477.

There are a few items in this circuit not covered in the above circuit descriptions. First, in the feedback network of the law corrected amplifier a diode has been added to prevent any appreciable positive excursion of the amplifier's output on switching on or off. Secondly, a zero set potentiometer is added to this amplifier to take out the combined zero errors of the four op-amps which although small enough to hardly affect the working accuracy is nevertheless rather annoying visually in the absence of an input level.

The zero set potentiometer is the usual value for the type 741 op-amp but is connected in a somewhat different manner. Instead of being connected between the two offset points of the 741 and the negative supply line, a resistor is connected to the slider of the potentiometer and returned instead to the 9.1V bias line. This arrangement allows a much wider range of adjustment than the usual connection, which although adequate to cope with the offset of one 741 op-amp is not sufficient to cope with the combined offset of four op-amps if these should unfortunately be additive.

The d.c. operating level of all stages is determined by the bias supply of +9.1V stabilized by the zener diode D_4 which also supplies the bias chain for the output amplifier's feedback network. This bias chain has an overall adjustment in order that the exact law correction of the completed instrument may be set up, and the tolerances of the various elements allowed for—in particular that of the zener diode stabilized voltage.

The 1mA meter to B.B.C. Specification ED1477, has a resistance of 600Ω ±5% so that with its series resistor of 4.7kΩ (R_{15}), full-scale deflection corresponds to −5.3V with respect to the +9.1-volt bias line. Maximum overdrive of the meter is limited therefore to a little less than the bias line

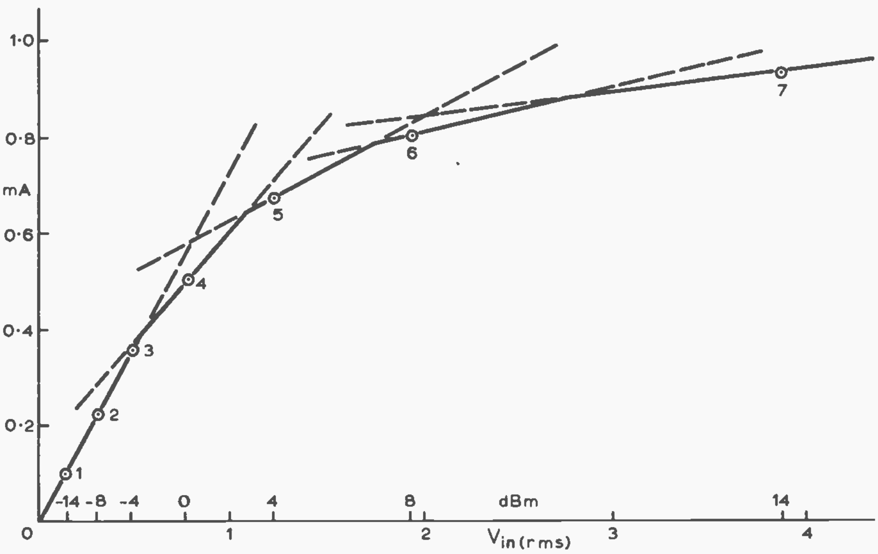


Fig. 6. Low corrected transfer curve approximated by a five section linear characteristic.

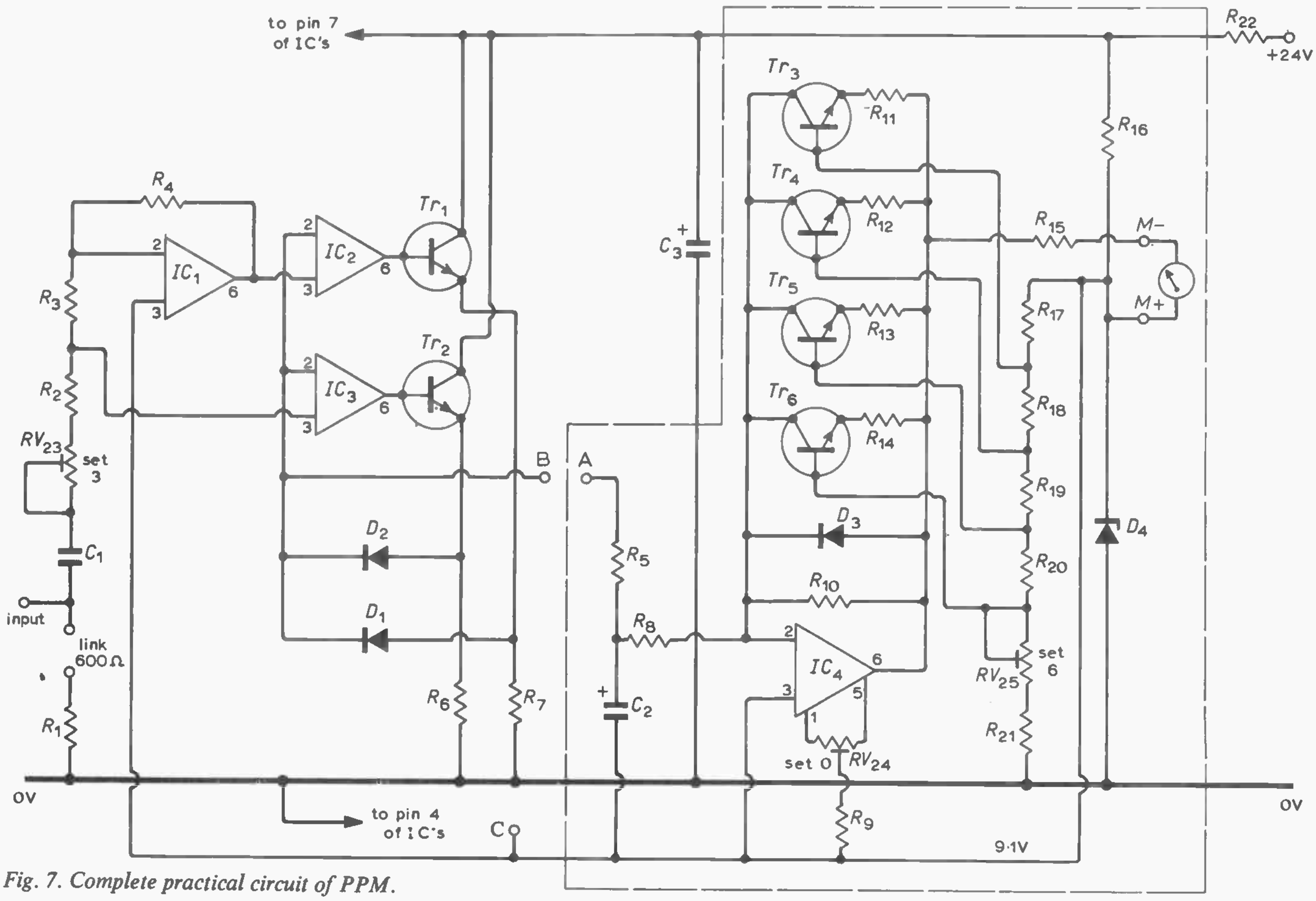


Fig. 7. Complete practical circuit of PPM.

voltage, or approximately some 8V, corresponding to roughly 150% overdrive—a reasonable value for meter protection. The general action of the circuit normally prevents reverse deflection, but in any case the diode in the feedback circuit prevents more than 0.6V being applied to the meter, corresponding to −11% deflection.

Due to the very high peak currents occurring in the peak rectifier circuit, particularly in the collector currents of Tr_1 and Tr_2, some measure of isolation from other equipment sharing the same supply line is necessary. To this end decoupling by R_{22} and C_3 is provided.

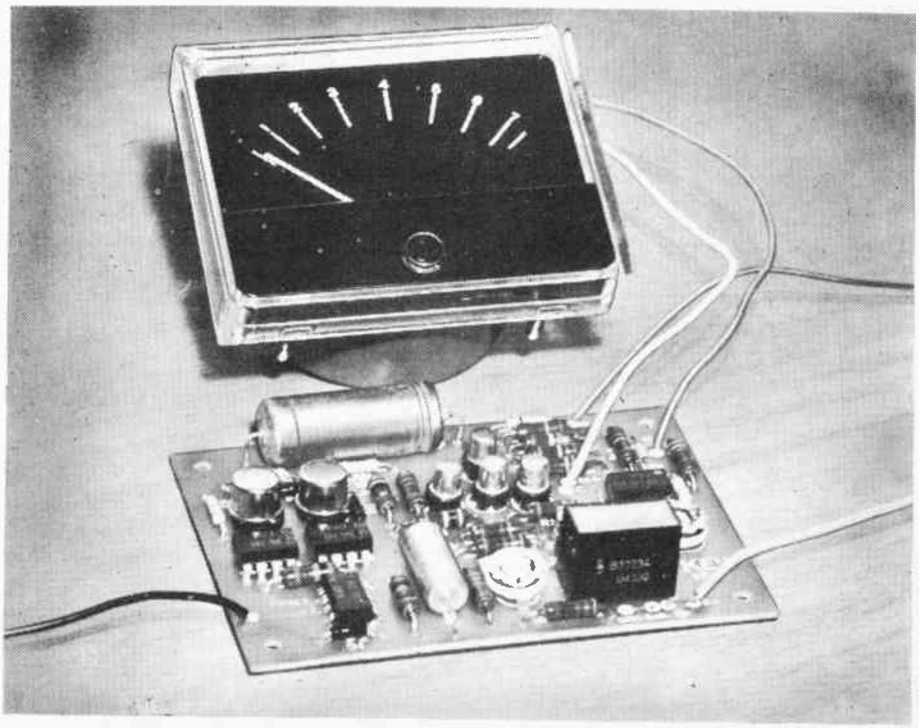
A resistor (R_1) of 620Ω is included so that it can be linked into circuit to give a line terminating impedance of 600Ω instead of the normal line bridging input impedance of around 16kΩ.

Setting-up and performance

The procedure for setting-up the PPM is a simple one. First, with zero input voltage, the zero is set ('Set 0' control) RV_{24}. Next a level corresponding to −4dBm (reading 3 on the PPM scale) or 490mV r.m.s. sine wave, 690mV peak, is applied and the 'Set 3' control (RV_{23}) is adjusted to bring the meter pointer to 3 on the scale. Finally a level of +8dBm, (reading 6 on the PPM scale) or 1.94V r.m.s. sine wave (2.75V peak), is applied to the input and the 'Set 6' control RV_{25} is adjusted to bring the meter pointer to 6 on the scale. The meter is then checked at 0, 1, 2, 3, 4, 5, 6, 7, and f.s.d. points as listed in Table 1, and any small adjustment made to the 'Set' 0, 3, and 6 controls to minimize the spread of errors. Having completed the sequence of adjustments the meter should read within 0.5dB at 1kHz at all scale marks, although f.s.d. is as stated in Table 1 undefined (it will usually correspond to around 5.3V r.m.s. sine wave).

Performance versus temperature. The PPM has very little variation with temperature. A 30°C rise in temperature (from 17°C) gave only about 10μA change in meter current at any point of the scale, i.e. about 1% of f.s.d.

Performance versus frequency. As shown in Fig. 8 there is a slight droop in the upper frequency range, and this is due to the limited slew rate capability of the 741 op-amp in the peak detectors. Amplifiers



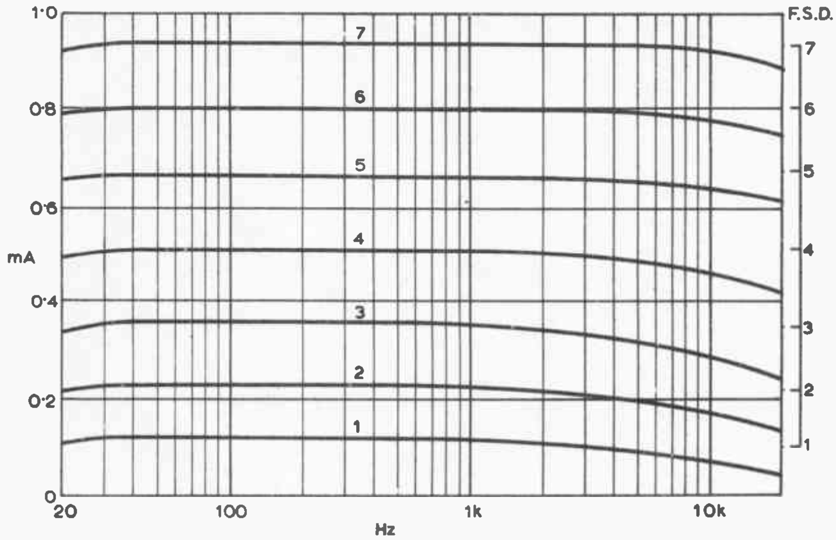
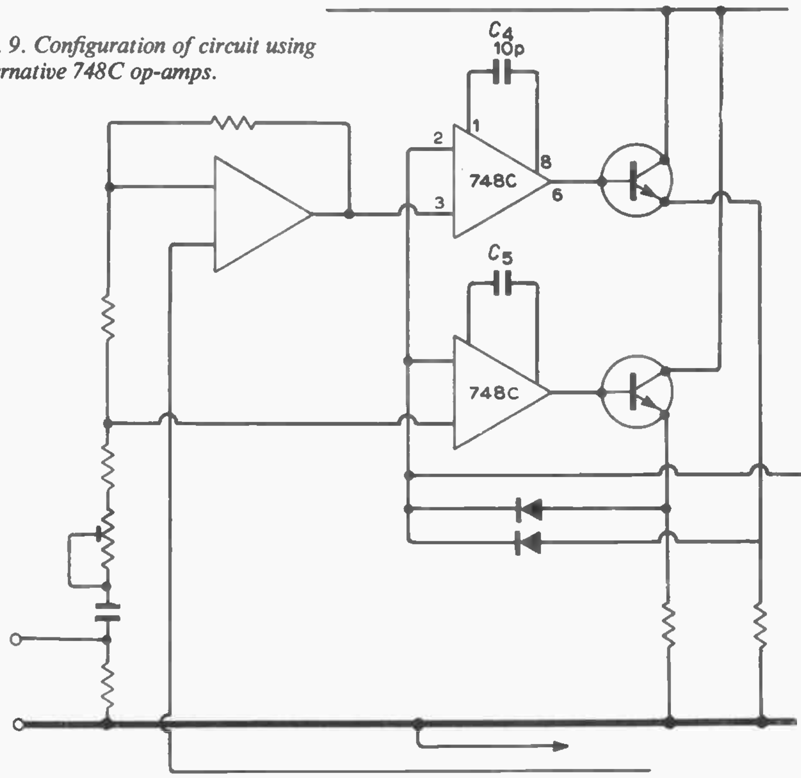


Fig. 8. Frequency response of prototype PPM.

Fig. 9. Configuration of circuit using alternative 748C op-amps.



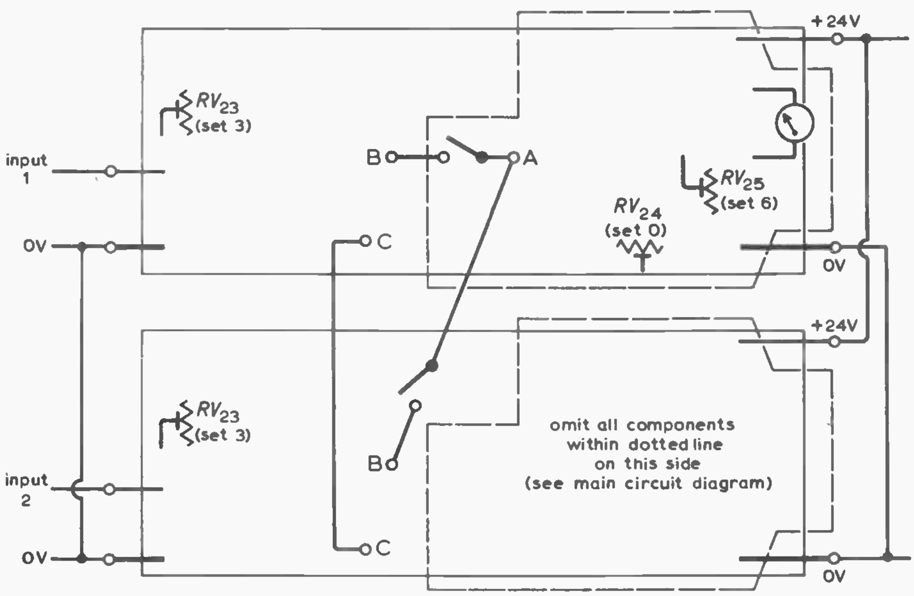


Fig. 10. Outline of printed circuit panel for mono or stereo.

having a higher slew rate have been tested and do remove this limitation in the audio range. The 748 op-amp has a higher speed performance than the 741 but uses external compensation; this allows the response to be tailored to suit any particular need. Fig. 9 shows how two 748 op-amps may be used for IC_2 and IC_3, together with appropriate extra components to obtain a flat frequency response over the whole audio band. There is still a slight fall off at 20kHz but this is greatly reduced as compared to the 741 op-amp.

In practice does this h.f. droop matter? The author would argue that for the monitoring of practical speech and music levels it does not matter to any noticeable extent. This is because of two factors. First, there is the attack response time of 2.5ms used in the circuit, meaning that a level must last for several milliseconds to register near to its true peak level, and secondly, in general, frequencies above about 5kHz do not exist at as high a level as the lower frequencies, and these lower frequencies therefore largely determine the peak amplitude at any time.

Performance versus supply voltage. Over the range of 16 to 30V there is little visible change of reading at any level of input. The circuit is designed for operation from a nominal 24-volt supply. Supply current is somewhat dependent on input level, and is typically 14mA at zero input, rises fairly rapidly as input is applied, and reaches 20mA at full scale. There will be some variation from unit to unit but at 24V the current should remain in the limits 13–22mA. The current demand is also dependent on supply voltage being lowest at 16V and highest at 30V. An absolute maximum supply voltage of 36V should never be exceeded.

Connections for stereo use with a single common meter

For economic or space reasons, it may be desired to use two PPM circuits with a common meter, and the printed circuits were designed with this in mind as an option. The method of interconnection is shown in Fig. 10 where two input circuits up to point *B* are used, with only one output circuit from point *A* onwards. The bias supply is made common to both boards by linking points *C* together.

To set up the meters in this method of connection the zero is first set at nil input level (to both inputs), 'Set 0' control (RV_2).

Next inputs of −4dBm are connected to each input in turn and the appropriate 'Set 3' control (RV_{23}) for that channel is set to give a reading of 3 on the PPM. Finally the 'Set 6' control is set to give a reading of 6 from either input at a level of +8dBm.

For the setting of the 'Set 3' and 'Set 6' controls both inputs may be connected in parallel and the switches shown in Fig. 10 operated to select the channel to be set up.

The dotted lines in Fig. 7 show the section of circuit omitted on one board and correspond to the dotted lines in Fig. 10.

Acknowledgements

The author wishes to express his grateful thanks to the B.B.C. Engineering Department and in particular to Mr. A. E. Tolladay, for considerable help and encouragement, also to Ernest Turner Electrical Instruments Ltd. for their help in the project.

Constructional appendix

The circuit is built on a printed circuit board $3\frac{1}{2} \times 3\frac{1}{2}$ inches in size with mounting centres of 3.1×2.1 inches (6BA). The board, which is suitable for either the circuit with 741s in all stages or the higher speed circuit with 748s in the peak rectification stages, is shown in prototype form in the photograph. Layout of production boards will differ slightly but all component positions are silk screened onto the component side of the board.

It is essential that the charge storage capacitor C_2 be a low leakage type, hence the specification of a solid dielectric tantalum type. An alternative is the solid dielectric aluminium capacitor such as Mullard type 121 15339 (33μF 16V) or 121 16339 (33μF 25V). However, it should be remembered that these are of 20% selection tolerance, and it may be necessary to select one to the necessary tolerance of 10%. In general normal aluminium electrolytics are not suitable, due to their high leakage (especially at elevated temperature) and very wide tolerance, even of the higher quality type (e.g. Mullard C428 is -10 $+50$%).

No special techniques are used in the construction and the only precautions needed are to ensure correct insertion of the 8-pin dual-in-line op-amp packages, and to avoid shorts on the board due to careless soldering (a miniature soldering iron is, these days, essential for printed circuit work). Mounting pads are used under the 6 transistors but are not absolutely essential. Connections are by 14–0076 p.v.c. covered leads as shown in the photograph.

All component parts in kit form together with Ernest Turner PPM meters type 642 are available from Key Electronics, P.O. Box No. 7, Bournemouth, BH7 7BS, Hants.

Components list

Resistors

R_1	620	R_{13}	56k
R_2	2.2k	R_{14}	15k
R_3, R_4	10k	R_{15}	4.7k
R_5	68	R_{17}	1.2k
R_8	30k	R_{18}	620
R_{10}	160k	R_{19}	560
R_{11}	220k	R_{20}	270
R_{12}	120k	R_{21}*	

All the above are 2% metal oxide or metal film (e.g. Welwyn MR5 or Electrosil TR5).

*Resistor R_{21} will normally be a wire link. (For use only where a higher reference line voltage than 9.1V is to be used.)

R_6, R_7	22k	R_{16}	2.2k
R_9	220k	R_{22}	47

All the above are $\frac{1}{8}$W 5% carbon film (e.g. Iskra UPM033 or Mullard CR25).

Capacitors

C_1	1μF, 100V poly (15mm mounting centres)
C_2	33μF, $\pm$10%, 20V solid dielectric tantalum
C_3	220μF, 35V, aluminium electrolytic.
C_4, C_5	10pF, ceramic tube, disc or poly.

Transistors

Tr_1, Tr_2	BFY52 or 2N2219.
Tr_3 to Tr_6	BC109.
D_1, D_2	OA200, 1S920.
D_3	1S44, 1N914, 1N916.
D_4	BZY88–C9V1 (9.1V, $\pm$5%, 400mW).
IC_1	741C (8 pin d.i.l.).
IC_2, IC_3	741C (8 pin d.i.l.) or 748C for high-speed version.
IC_4	741C (8 pin d.i.l.).

N.B. The TO–99 versions (multi-lead TO–5), of the 741 and 748 may also be used since they have the same lead layout and are easily arranged in d.i.l. lead configuration. Jermyn Industries Ltd., type MON–8L mounting pad may be used to achieve this end.

Preset potentiometers

RV_{23}, RV_{24} 10k $\pm$20% RV_{25} 5k $\pm$20%

Open cermet potentiometers (R.S. Components or A.B. Electronic Components).

Semi-sealed type Morganite 81E may also be fitted (also from R.S. Components).

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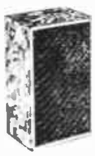

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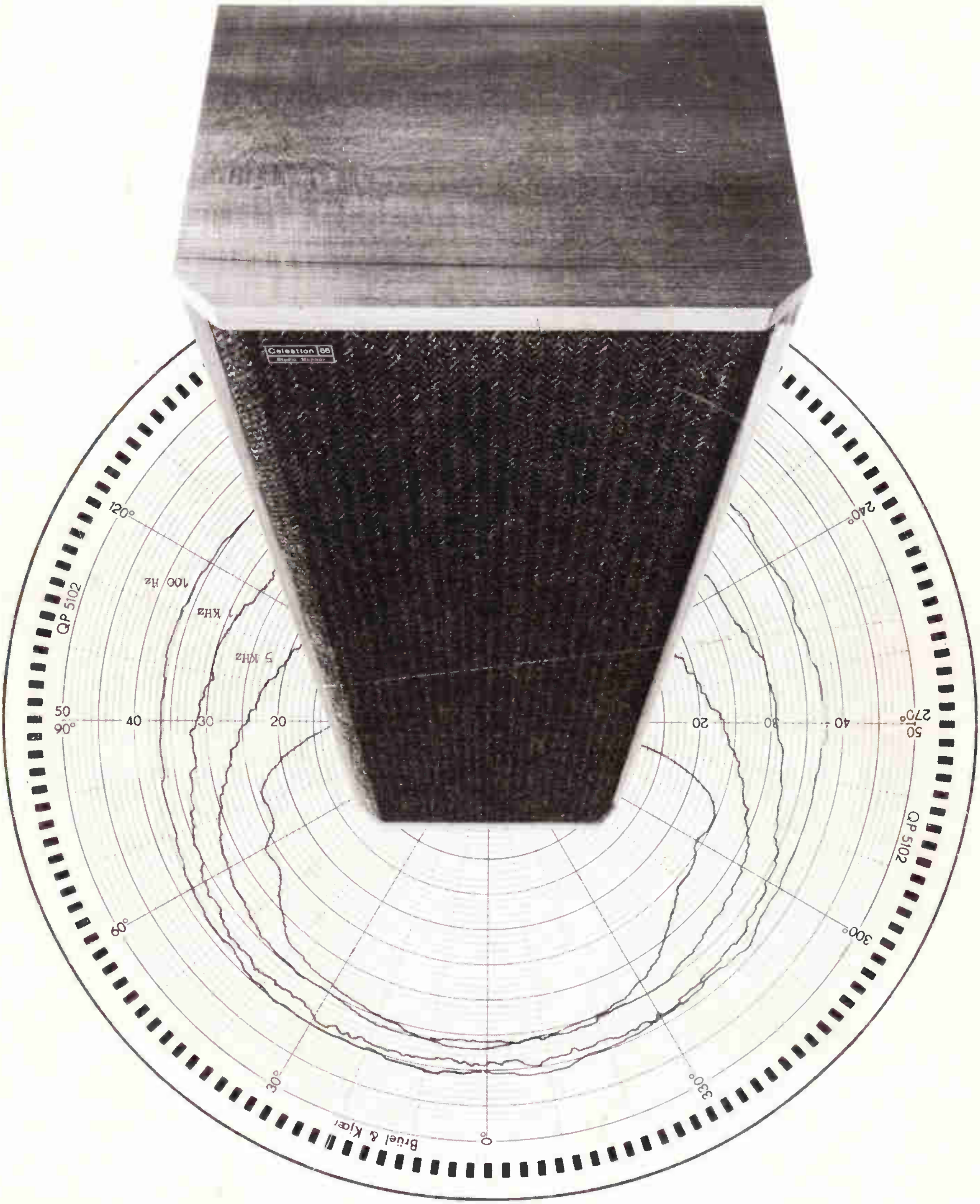

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