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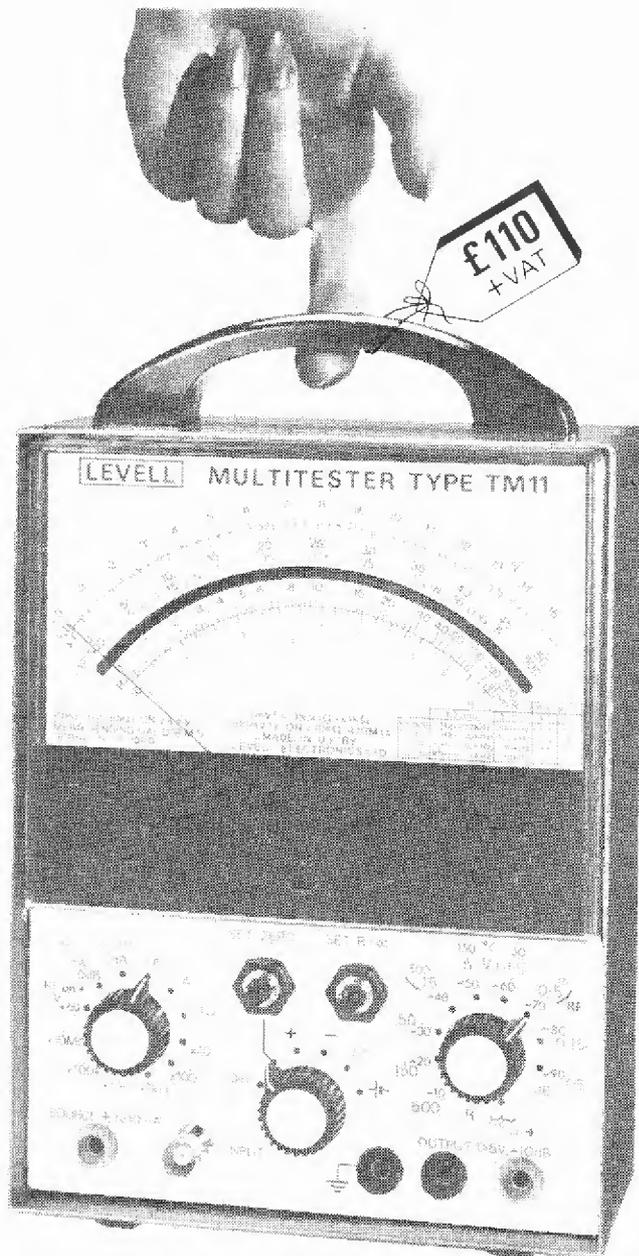
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Editorial compiled by Philip Darrington
Deputy Editor, *Wireless World*

wireless world

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Introduction

THE success of the first collection of articles clearly demonstrated that there is a need for the preservation of out-of-print information on constructing high-quality sound reproduction equipment. In this second book, we have up-dated most of the content, but have left in some of the more popular articles which are still in demand.

The inclusion of John Linsley Hood's cassette deck design prompted the succession of noise-reduction and range-expanding units, while the Walltenna was too appealing an idea to leave out. Geoff Cowie's piano is not, perhaps, strictly an audio project in the ordinary sense, but many people seem to want photocopies, so that we reproduce it here in rather better shape than the average copy. Arthur Bailey's 30W amplifier is still very much in demand and this is retained, but David Read's active crossover system is an addition. This has been very well received and appears to have started a series of copies — always a sign of quality. The transmission-line speaker is the second version, and Jack Dinsdale's horn speakers are included.

The flow of designs submitted for publication shows no sign of drying up, so that even the equipment in this collection of reprints will eventually be overtaken by improved techniques. To keep up with progress, there is no better source of information than *Wireless World* which, over 64 years, has built up a reputation for being first with new techniques, many of which are now industry standards.

FM tuner designs

Two designs with various circuit options

by D. C. Read, B.Sc.

Among the many benefits conferred on the home constructor by the increasing sophistication of available packaged components is that these allow complex circuit ideas to be not just considered as expensive ideals but realised in practice with comparative ease and economy. This is particularly true in the field of f.m. stereo reception. L. Nelson-Jones pointed the way in his articles*. Further development of these ideas has led to a more comprehensive design, two versions of which are described below. For both, apart from the usual consideration of first-class performance and stability of operation, the main aim has been to eliminate time-consuming r.f. and i.f. alignment difficulties and the possible need for expensive test equipment. In the following description, modifications and optional extra facilities are discussed, and the necessary constructional details given (see also part 2).

The first, and simplest, version makes use of the well-proven Nelson-Jones i.f. and demodulator sections (though slightly modified) and replaces the discrete-component front end with a voltage-controlled tuner module. It also incorporates on the same board a phase-locked loop stereo decoder circuit based on the Motorola chip. Pre-selected station change is by means of push-button switching with an extra switch position giving access to a manual control for tuning over a large part of Band II. Included in the design are regulator circuits for providing both the main supply rail and the constant-voltage source for tuning control.

The more advanced development uses an RCA CA3089E package containing i.f. amplifier/limiter and demodulator circuits which are sufficiently sensitive to allow all the i.f. pass-band shaping to be performed at a lower level: limiting starts with an input of about $15\mu\text{V}$ for the RCA integrated circuit. Thus, only a small amount of

Performance

Signal-to-noise ratio.

Simple version 62dB at 1mV input

27dB at $1\mu\text{V}$ input

f.e.t. version offers an improvement for low signal levels of 1-100 μV e.g. 55dB instead of 47dB at $10\mu\text{V}$.

Distortion. Simple version: 0.5% at 1kHz and 75kHz deviation (1mV), see text for method of improving to 0.1% f.e.t. version gives 0.12% at 1mV, 0.08% at $10\mu\text{V}$.

Crosstalk. 34dB 80Hz to 5kHz with C_{28} , R_{27} optimized.

RF performance. See part 2.

gain in the preceding stages is required, to make up for ceramic filter losses and provide impedance matching, and the 20dB gain i.c. (CA3053) used in the simple version is therefore not needed.

In addition to the composite signal output, the CA3089E produces:

- delayed a.g.c. voltage
- push-pull current supply for a.f.c.
- adjustable inter-station muting voltage.
- direct voltage proportional to the r.f. signal amplitude to actuate a tuning meter or show received field strength.

Of these, the first is most useful. It is used here to control the gain of an additional aerial-fed r.f. amplifier which, as well as giving the tuner increased sensitivity for reception of weak incoming signals, attenuates those of excessive strength to reduce the risk of local-oscillator pulling, an effect which can occur when the LP1186 module is over-driven. More particularly the a.g.c. circuit using this control feed can easily be tailored to suit different reception conditions according to location and requirement.

In the simple version of the tuner, an a.f.c. feed is conventionally derived from the demodulated audio and, because of other precautions taken against drift, is more than adequate for all practical purposes. The availability of a separate a.f.c. supply is not therefore particularly significant except that it does more readily offer a choice

of control sensitivity. Similarly, the other two CA3089E outputs have only limited application in the present design. For general household use muting is not required with stable push-button tuning; neither is there need to inspect the incoming signal level. These optional facilities are included only to allow for band-searching by manual tuning.

The decoder, stabilizer and tuning-voltage circuits are the same in both versions of the tuner.

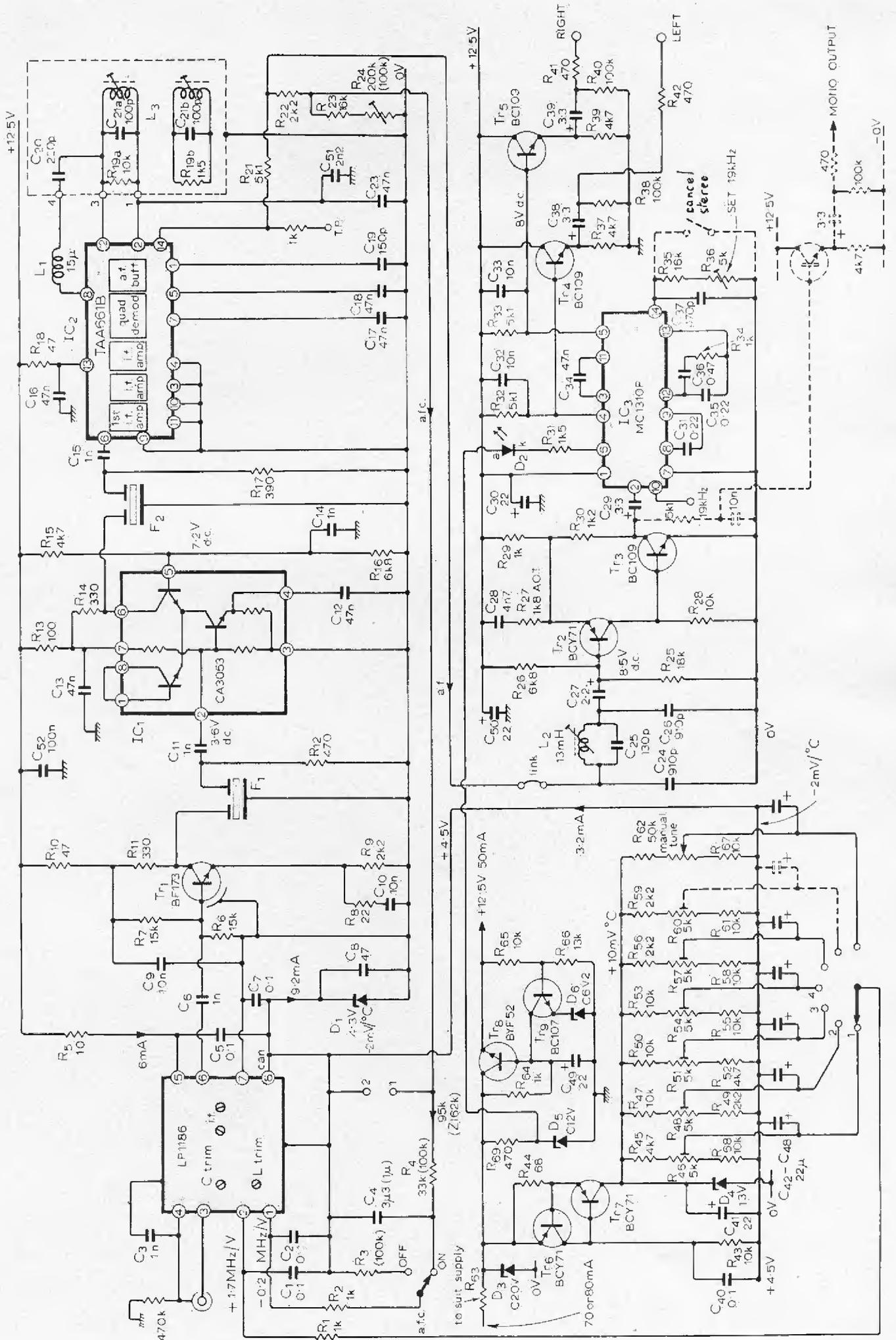
Simple version

The LP1186 module circuit is "floating", but is made unbalanced by C_3 (Fig. 1) so that it is suitable for connection to the aerial via a 75-ohm co-axial feeder. A.C. earthing is used because the module negative supply rail and the metal cover is held at about +4.5 volts with respect to the main chassis earth by zener diode D_1 . There are two reasons for this, first, the LP1186 requires an 8 volt supply, instead of 12.5 volts nominal as in the remainder of the tuner circuit. And second, the local-oscillator frequency-control circuit operates about a 0-volt zero error signal at pin 1 (ref, pin 8) but the a.f.c. voltage produced by the TAA661B has an on-tune centre value of around 5.4 volts, reduced to about 4.5 volts, in the chain R_{22} , R_{23} and R_{24} .

A minimum incoming r.f. signal of about $2\mu\text{V}$ r.m.s. for 30dB quieting is

Fig. 1. Simpler version of tuner design using LP1186 module. Stereo decoder and tuning circuits are common to the two tuner designs. See part 2 (April issue) for modification for Toko module. Station selection switch should have break-before-make contacts. Resistors can be 5% high stability or metal film types, rated at $\frac{1}{4}$ watt. Capacitors are 2% polystyrene types for signal circuits, disc ceramics for r.f. decoupling, and tantalum types for audio coupling and decoupling. Switch shown with broken lines is closed for mono reception to kill 76kHz oscillation. Values in parentheses are for lower a.f.c. sensitivity.

* F.M. Stereo Tuner, *Wireless World*, vol. 77 1971, pp.175-80, 245-9. See also p.376, vol. 78 1972, pp.179-83, pp.318/9 and vol. 79 1973 pp.271-5, p.591. All reprinted in "High Fidelity Designs", IPC Business Press Ltd, 1974.



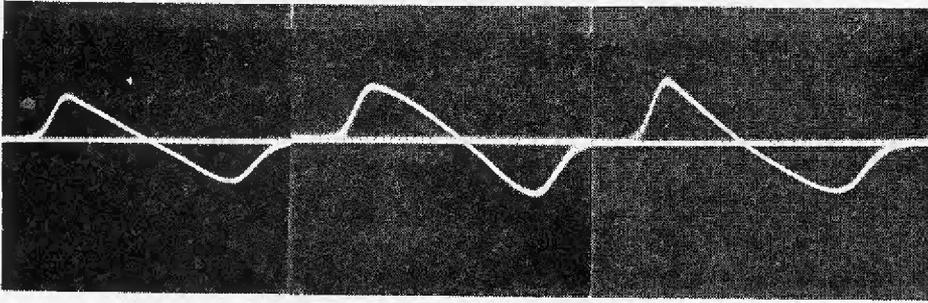


Fig. 2. Effect of changing dummy circuit resonance on transfer slope of demodulator. (a) left, optimum setting (b) middle, upper core "in" on optimum setting (c) right, upper core "out" on optimum setting. (1V/cm.)

necessary to give adequate limiting in the TAA661B amplifiers; and a maximum amplitude of between 5 and 10mV is recommended to prevent oscillator pulling.

The remaining LP1186 module at pin 2 is the tuning control voltage from the push-button selector circuit which has selected values between 4.7 volts and 6 volts; a somewhat greater range is provided in the continuously-variable manual tune position. Both this voltage feed and the a.f.c. line are decoupled (R_1/C_1 , R_2/C_2) to prevent spurious modulation effects which could be caused by hum fields or other stray signals picked up on wiring to and from switches.

AFC sensitivity

Following normal practice, of course, the a.f.c. feed passes through components, C_4 , R_4 , which filter out the audio modulation. The input resistance at pin 1 of the LP1186 is 62k Ω , if R_4 has the value 100k Ω in Fig. 1, the useful a.f.c. voltage change is about one-third of the total available. If greater sensitivity is required, therefore, the value of R_4 must be reduced and that of C_4 raised to maintain effective audio rejection. A suggested pair of values is 33 k Ω and 3.3 μ F. With these in circuit a 50% increase in sensitivity is achieved but at a greater risk of locking to an adjacent station.

The problem of incorrect selection because of excessive a.f.c. is aggravated by the interleaving allotment of transmitter frequencies and therefore mis-selection is most likely to occur at points where interleaved channels are received at comparable strengths. A typical instance might be a location midway between the Oxford and Wrotham transmitters which radiate basically the same Radio 1, 2 and 3 programmes on interleaved frequencies. In such circumstances, the station chosen by the receiver might not be the wanted one; the choice will arbitrarily depend on the direction of tuning change. As the tuning shifts up or down the band from one selection to the next, the local oscillator might be captured by an in-between transmission which creates a large enough a.f.c. voltage to make it lock to this station in error. Over-sensitivity of the a.f.c. can also result in station-jumping effects where the receiver suddenly changes tuning and switches away from one transmission to some other because of a

reduction in received signal strength; aircraft flutter, particularly, causes such mis-operation.

It is obviously good practice then to set the a.f.c. sensitivity so that it is no more than just sufficient. In the event that particular reception conditions are such that sensitivity is already too large, even with the circuit as given in Fig. 1, two pairs of diodes in series (types 1N914 and 1N916 are suitable) can be connected back-to-back across C_2 at the LP1186 a.f.c. input. With this modification, the frequency-control swing is limited to less than the 300kHz station spacing and thus station jumping or mis-selection will be prevented.

The 10.7 MHz output from across pins 6 and 7 is fed to Tr_1 which provides the correct source impedance for the first ceramic filter, and also gives some amplification. The amount of gain is set by the value for R_8 and should be such that, for a low r.f. input of 10 μ V to the tuner, a suitable signal level (say 10mV) is available to drive the first i.f. amplifier in the demodulator module; this gives 40dB quieting. A 20dB amplifier stage comprising the cascode-connected circuit in IC_1 provides the correct source impedance for F_2 which finally passes the band-shaped i.f. signal to IC_2 .

Remember that the Vernitron type FM4 components used for F_1 and F_2 must have the same colour marking. The green-coded type is recommended because these have a pass band centred on 10.7MHz and therefore match the curve normally provided by the maker's preset adjustment of LP1186 modules. If ceramic filters of another colour code are used, it may be advantageous to re-tune the two output band-pass coils in the LP1186 for optimum performance. These are accessible through holes in the module cover. The best way of making the adjustments is to use a frequency-sweep input signal displayed on an oscilloscope connected across the demodulator input (pin 6 of the TAA661B). Such ideal methods are rarely available to the home constructor, however, and the practical compromise is to select a weak incoming signal and then adjust the coils for least background noise in the sound output

from the tuner.

For reception of weak signals it may be worthwhile to carry optimization of the LP1186 one step further and adjust its input circuit to match the aerial. Two other holes in the LP1186 cover allow separate access to the aerial trimming coil and its associated capacitor. Because these components affect opposite ends of the tuning range, their adjustment is a relatively simple matter: using the manual tuning control, select a weak station radiating in the 87-89 MHz range and tune the coil for minimum noise; no more than a fraction of a turn is needed to show either that a reduction in noise is possible or that the optimum setting already exists. Change to a weak signal towards the other end of the band (96-97 MHz) and similarly adjust the trimmer capacitor. This is not an essential adjustment and will, at best, give only small improvement for weak stations.

Quadrature demodulator

The circuit surrounding IC_2 shows two main differences compared with that in the original Nelson-Jones tuner. First, the i.f. signal sample used to derive the quadrature-phase demodulating signal is taken through an inductor L_1 instead of a capacitor (note that C_{20} in Fig. 1 is now simply a d.c. blocking component). This is done so that the resulting demodulator transfer slope is in the correct sense for a.f.c. Second, the phase-shifted carrier itself can optionally be produced by two tuned circuits with twin coils L_3 having separate cores but mounted on the same former. As Fig. 1 shows, one of these circuits is a dummy, the tuning of which is adjusted so that the modifying component of current induced in the main tuned circuit is of suitable phase and amplitude to give a straighter transfer slope. The effect of changing the dummy circuit resonance is illustrated by the three sweep photographs which show: (a) the transfer slope for an optimum setting; (b) and (c) non-linearity resulting from two incorrect settings.

There is, unfortunately, a difficulty to be met in using this apparently simple and cheap modification: it is only effective if properly adjusted and although adjustment is relatively easy, it necessarily entails the use of extra test equipment. Further, the basic reduction in tuner output distortion is marginal (typically, 0.5% total harmonic content for the one-coil circuit; 0.1% with two coils) and would be hard to detect aurally. Even so, a low level of harmonics in the demodulated signal, helps to prevent intermodulation products in the overall stereo decoding process and, provided that suitable test equipment is available, the additional circuit and set-up procedure offers a worthwhile advantage.

There are two possible methods of adjustment. The first uses a distortion meter to measure the total harmonic

content in the audio output (taken via a 15kHz low-pass filter such as that described later) for an r.f. tuner input modulated by 1kHz at $\pm 75\text{kHz}$ deviation. The filter is needed here to reject the 19kHz pilot tone as well as the 23kHz transmitter switching signal remaining in the audio output. The dummy tuned circuit (upper core) is then simply adjusted to give a minimum reading on the meter. However, since the distortion figure which can be achieved is low (about 0.1%), the exact null point may be somewhat masked by noise. The alternative method overcomes this. It uses a wave analyser tuned to the 3rd harmonic of the incoming 1kHz modulation which is again at the maximum deviation of $\pm 75\text{kHz}$; in this instance there is no requirement for a low-pass filter. The adjustment of dummy circuit tuning is made for a minimum output at 3kHz.

The demodulated multiplex signal from pin 14 of IC₂ at about 0.5 volts r.m.s. for $\pm 75\text{kHz}$ incoming f.m. deviation has a d.c. component of about +5.4 volts for the in-tune condition. In fact, the value given here is a nominal one and varies between different examples of the TAA661B; since this varying direct voltage is used to operate the a.f.c. circuit and must be matched to the supply offset provided by D₁ (4.3 volt zener), variable resistor R₂₄ is included in the a.f.c. potentiometer chain to allow fine adjustment.

The method suggested for adjustment is as follows. Switch a.f.c. off. Using the manual control, set the tuner well away from any station, i.e. completely off-tune. Measure the direct voltage at the IC₂ pin 14 test point (this connects with the demodulator output via a protecting 1k Ω stand-off resistor); normally, the value obtained will be about 5.4 volts positive.

Tune through a reasonably strong incoming signal and, by observing the voltage change from maximum positive to maximum negative (a total peak-to-peak swing of, say, between 2.5 and 3.5 volts) sample the S curve to find the on-tune point, which is at a voltage nearly equal to that already established for the off-tune condition. With the tuning set at this point, transfer the meter to the tuning-indicator connection points marked 1 and 2 in Fig.1. In the on-tune condition and with R₂₄ set at a maximum, terminal 1 will be positive with respect to 2. Adjust R₂₄ to bring this potential difference to zero. Switch a.f.c. on and observe the possible slight change of meter reading. Again adjust R₂₄ to restore it to zero. Finally, check the voltage appearing at the test point and operate the a.f.c. switch to ensure that this voltage remains unchanged with and without a.f.c. applied.

Tuning voltage selector circuit

The circuit used to provide a selection of pre-set direct voltages for tuning purposes in the LP1186 module forms the

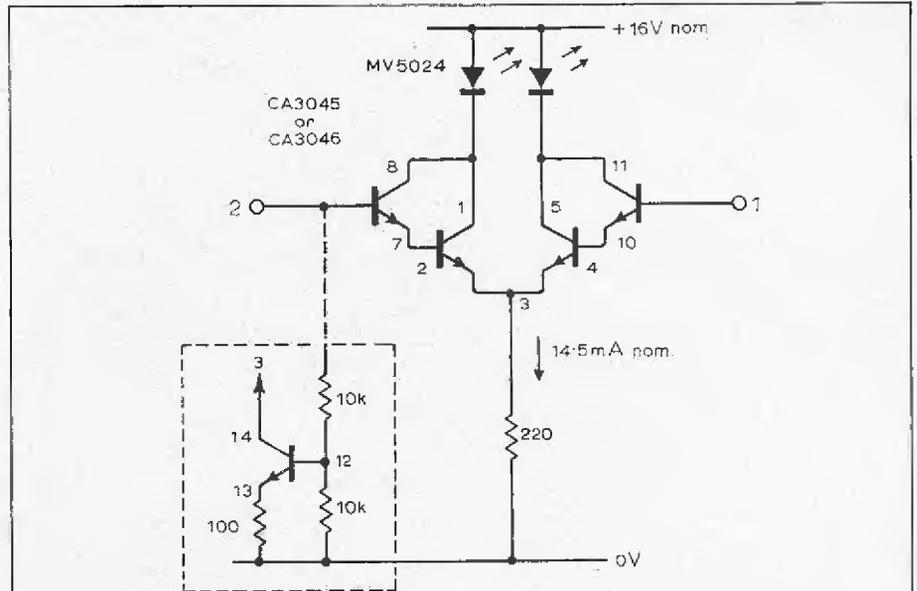
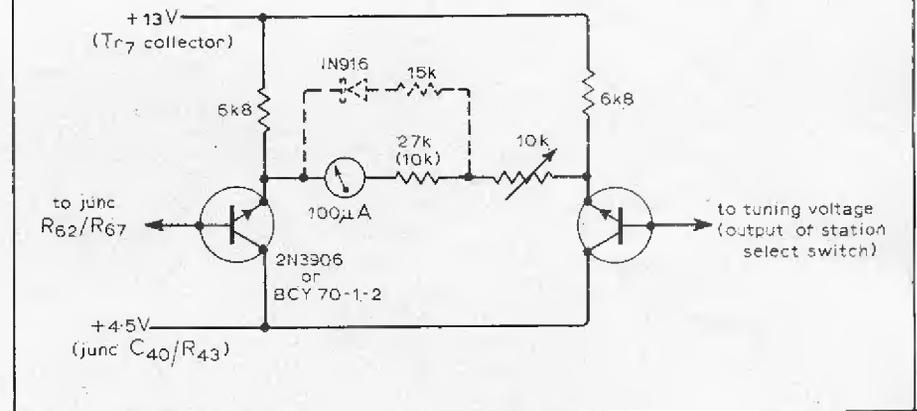


Fig. 3. In this tuning indicator option, the 220ohm resistor can be replaced with the circuit shown in broken box to ensure equal lamp brightness either side of tuning position.

Fig. 4. Optional meter circuit provides tuning scale for manual tuning. Meter is RS Component MR100.



lower left of Fig. 1; this part of the circuit also includes the main supply-voltage stabilizer.

All the tuning voltages are derived as proportions of a fixed voltage from zener diode D₄ which is supplied with a constant current by the stabilizer circuit of Tr₆ and Tr₇. D₄ is returned to 0 volts instead of the 4.5 volt rail used by the LP1186 module and the rest of the tuning voltage circuit, so of the 13-volt reference potential provided by D₃, only about 8.5 volts is used for tuning. The main reason for this arrangement is that by allowing for a greater reference voltage than necessary, the zener diode used to provide it has a positive temperature coefficient large enough to give the degree of frequency-drift correction required. The coefficient of a suitable lower-value zener, although still positive, would be too small for this purpose; the figures for comparison here are +10.5mV/ $^{\circ}\text{C}$ for D₄ and -10kHz/ $^{\circ}\text{C}$ for the oscillator.

Thus, with D₄ connected as shown, the variation of control voltage with temperature has an effect on the

tuning frequency which both opposes and, after accounting for potentiometer action and the 4.3 volt zener offset in the selector circuit as well as the tuning-voltage/frequency relationship, is about equal to the variation caused by a similar temperature change on the local oscillator itself. Obviously, for automatic compensation to become fully effective, the relevant components in the tuner must themselves have reached their normal working temperature.

If the number of pre-selected stations required is more than, say, 12 then as each chain takes about 0.4mA the total drain on D₄ might be greater than its reserve of current and it would no longer be effective in maintaining a constant voltage. In this event, it will be necessary to increase the current supply from Tr₇ by reducing R₄₄ to the next lowest preferred value (56 Ω).

Transistor Tr₈ and Tr₉ form the main parts of a conventional series regulator acting with reference to zener diode D₆ to provide the main supply rail of 12.5 volts from the nominal 16 volt d.c. input.

An incoming feed of this value is conveniently obtained by peak rectification of the output from a mains-to-12 volt transformer. It should be capable of supplying at least 250 mA r.m.s. at 12 volts to ensure the required rail voltage.

The 12 volt zener, D_5 , provides a reserve of current for supplying the stereo indicator D_2 so that the main supply rail is not affected by current changes as this l.e.d. is switched on and off.

Stereo decoder and output circuit

The right hand of Fig. 1 shows the decoder module feeding twin audio output circuits and preceded by a low-pass filter mainly comprising L_2 . This filter passes the composite multiplex signal obtained from the demodulator including the upper subcarrier sideband extending to 53 kHz but rejects frequency components outside this range. Ideally, the filter should have a flat pass-band with negligible phase distortion so that the mono and stereo information channels occupying 0 to 15 kHz and 23 to 25 kHz, respectively, can be recovered with equal fidelity. It should also cut off sharply to give the maximum possible attenuation to all signals outside this band, especially in the range 99 to 129 kHz, which includes the first odd harmonic of the stereo channel subcarrier with sidebands.

To satisfy such a requirement would entail the use of a complex network; in practice, the simple, single-section filter used in the tuner is adequate, even more so if the demodulator dummy tuned circuit, discussed earlier, is used to reduce the level of interfering harmonic components. As an added refinement, the tuning of L_2 can, if desired, be adjusted to set the first rejection frequency so that optimum separation is obtained for signals in the region of 5kHz; this is the upper end of the audio range over which good stereo separation is most important.

Further overall response adjustment is given by the feedback stage of Tr_2/Tr_3 . The C_{28} , R_{27} circuit causes a basic 6dB/octave rise which is modified by R_{29} so that the resulting slope counteracts a general slight fall in the preceding circuits. As before, equality of level for both the mono and stereo information channels at the decoder input is the criterion. The low output impedance presented by Tr_2/Tr_3 is a necessary factor in the proper operation of the MC1310P circuit; separation at the lower end of the audio band suffers if this requirement is not met.

The decoder module, IC_3 , is operated in a normal manner with surrounding circuit values much the same as in an article which introduced the MC1310P (*Wireless World* July 1972). The only addition is the optional 76 kHz oscillator-disabling switch shown in Fig. 1. If fitted, this is used to inhibit stereo operation for exceptionally weak incoming signals when the resulting 20dB improvement in signal to noise

ratio offers a worthwhile advantage.

De-emphasis of the decoded audio signals taken from open collectors at pins 4 and 5 of the MC1310P is arranged by shunting each of the load resistors, R_{32} and R_{33} with $0.01\mu F$ capacitors. The twin output signals are then available from buffer emitter followers, Tr_4 and Tr_5 . Apart from the more obvious benefits of having low-impedance outputs, these are particularly useful, with series resistors R_{41} and R_{42} suitably changed in value, for feeding the 15 kHz low-pass filters (part 2) which may be inserted between the tuner and its following amplifiers.

Extra circuits

Where stereo programmes are to be used to make mono recordings, the emitter follower circuit shown dotted in the lower right hand corner of Fig. 1 would be a useful addition. This simply provides a low-impedance output of the separately de-emphasized multiplex signal.

Another possible extra facility is the tuning indicator circuit illustrated by Fig. 3. This basically comprises two Darlington pairs in a single i.c. with a common-emitter load and light-emitting diodes in the collector feeds. When connected to a.f.c. circuit points 1 and 2, in Fig. 1, these diodes show equal illumination for equal input voltages at pins 6 and 9 to indicate the in-tune condition whereas one or other is brighter on either side of this point. An optional refinement to the basic Fig. 4 circuit (shown boxed) overcomes possible asymmetry in individual diode brightness for off-tune conditions. As the modification shows, the common-emitter resistor is replaced by a constant-current source using a spare transistor in the i.c. and two additional resistors.

As a further aid to station selection, the reader may like to include the circuit shown in Fig. 4 and thus provide a tuning-scale facility. The added circuit mainly uses a readily-available and reasonably cheap edgewise meter which in my installation is mounted together with the pre-selection buttons and other controls on a remote front panel. Fig. 4 shows how the meter is connected into the main tuning/selection system detailed in the Fig. 1 circuit which requires only two small modifications. One is the addition of a series resistor between R_{62} , the manual tune control, and the 11-volt maintained tuning-voltage supply rail. The value of the added component (typically 18k Ω) is chosen on test so that the meter full-scale deflection (indicating 98 MHz) occurs at the fully-clockwise slider setting of R_{62} . Second, the value of R_{67} will need changing to, say, 22k Ω to make the R_{62} fully-anticlockwise setting coincide with a tuning frequency of 88 MHz. The values actually required might be different because the tuning-voltage spread for the LP1186 varies by

about 1 volt at the low end of Band II and about 3 volts at the top end.

With the meter circuit as shown full line in Fig. 4 the scale (constructed by experiment) will be cramped at the low-frequency end. This is an advantage if all the pre-selected station frequencies are here, because the more open upper-end markings then give better accuracy when exploring/setting in the manual tune position. However, in some other instance it may be necessary to make the scale marking more linear and this can be done by modifying the meter circuit as shown dotted in Fig. 4 whereby the scale already provided with the meter can conveniently be used such that 0 = 88 MHz and 10 = 98 MHz.

FM tuner designs

2 — Improved performance; further facilities

by D. C. Read, B.Sc.

Changes which provide the tuner described in part 1 of this article with some additional control and monitoring facilities and a more flexible input circuit are shown in Fig. 5. The extra gain-controlled r.f. stage comprising the dual-gate m.o.s.f.e.t., Tr_1 , can be arranged to function in different ways according to local reception conditions. Two alternatives are illustrated in the circuit diagram by the indicated possible connection of a $10k\Omega$ resistor between Tr_1 source and the positive supply rail. Circuit operation is as follows.

With $10k\Omega$ resistor. The stage produces either a gain (maximum 6dB) or a loss (maximum 12dB) under the control of the a.g.c. voltage returned from the i.c. This division into two control regions makes the most efficient use of the available 18dB a.g.c. range whereby large incoming signals are reduced in level to prevent oscillator pulling but weak signals are given low-noise amplification before the LP1186 r.f. and mixer stages, so that the noise these produce is added in smaller proportion.

Without resistor. The stage gives low-noise gain with a value between zero and 12dB again depending on the a.g.c. voltage. This arrangement is suitable for tuners used in fringe areas where received signals are low; i.e. where increased sensitivity is required and high-level incoming signals are not normally encountered.

A further possibility makes even more effective use of the m.o.s.f.e.t. characteristics but at the expense of added complexity, particularly in setting up. If the Tr_{10} source is held at a fixed voltage, say by means of a low-value zener between it and the 0-volt rail with a current feed via a resistor to the positive rail, then the a.g.c. range is extended because the source-follower feedback action which modifies the effect of the control voltage on gate 2 is inhibited.

The spread of characteristics for f.e.t. devices is such that, without this stabilizing feedback, the bias on gate 1 needs preset adjustment to give maxi-

● The simpler version described in part 1 comprises tried and trusted circuits, up-dated with refinements intended to make construction, line-up and operation easy; stability and utility are the essential features. The overall design is flexible, and various special facilities can easily be added either during or subsequent to the main construction. These extras include:

- a twin tuned-circuit demodulator which reduces harmonics in the recovered multiplex signal but which needs proper adjustment using a wave analyser or distortion meter
- a stereo-inhibit switch which allows mono reception of weak stereo signals thus giving a 20dB improvement in signal-to-noise
- a buffered and de-emphasized mono feed derived before decoding and intended for tape-recording
- low-pass audio filters to remove unwanted components from the tuner outputs, useful for tape recording either stereo or mono
- a tuning-indicator circuit.

● The more advanced tuner can be provided with any or all of the additions listed above; it also shows further refinements, some optional, which give improved performance in certain respects but which increase the number of necessary adjustments both in setting up the tuner and in its normal operation. These modifications and additions are:

- an extra gain-controlled r.f. stage giving increased sensitivity and stability, and improved signal-to-noise performance. The design of this stage also allows different a.g.c. characteristics to be chosen either as a result of fixed circuit changes or subsequently by adjustment of a panel control to suit various reception conditions
- a more comprehensive a.f.c. system which, like the a.g.c. circuit, can be varied in its effect under external control (R_9 could be a front-panel variable resistor)
- a received signal-strength meter circuit with calibration curve. This meter feed could also be used for stereo-threshold switching.
- adjustable inter-station muting.

mum gain for weak signals. In practice the required bias is easily set by connecting gate 1, actually the earthy end of the input coil, to a variable tapping in a high-resistance potentiometer chain across the zener. Then, with the a.g.c. voltage on gate 2 at its most positive value, the bias is varied until the highest possible stage gain is obtained. The likely performance of such a circuit is a maximum gain of 16dB and a control range of 25 to 30dB.

The circuit which includes the LP1186 module and the impedance-matching stage, Tr_2 , is largely as in the simpler version, the only difference being an additional resistor in the a.f.c. feed. The choice of value for this component, which determines a.f.c. sensitivity, is

dictated by local reception conditions. High sensitivity is given with the value at $47k\Omega$ as shown in Fig. 5. If equal-strength neighbouring-channel signals are present, the degree of control might be too great such that the tuner could be captured by an unwanted station as the local oscillator sweeps through the relevant frequency while changing to select the wanted station. If this occurs, reduce the resistor value, possibly to as low as $5k\Omega$, which still allows a useful amount of control.

Because of the extra gain now available at the tuner front end and also in the CA3089E module, the i.f. amplifier IC₂ is not required and the correct impedance for F_2 is provided instead by a grounded-base stage, Tr_3 .

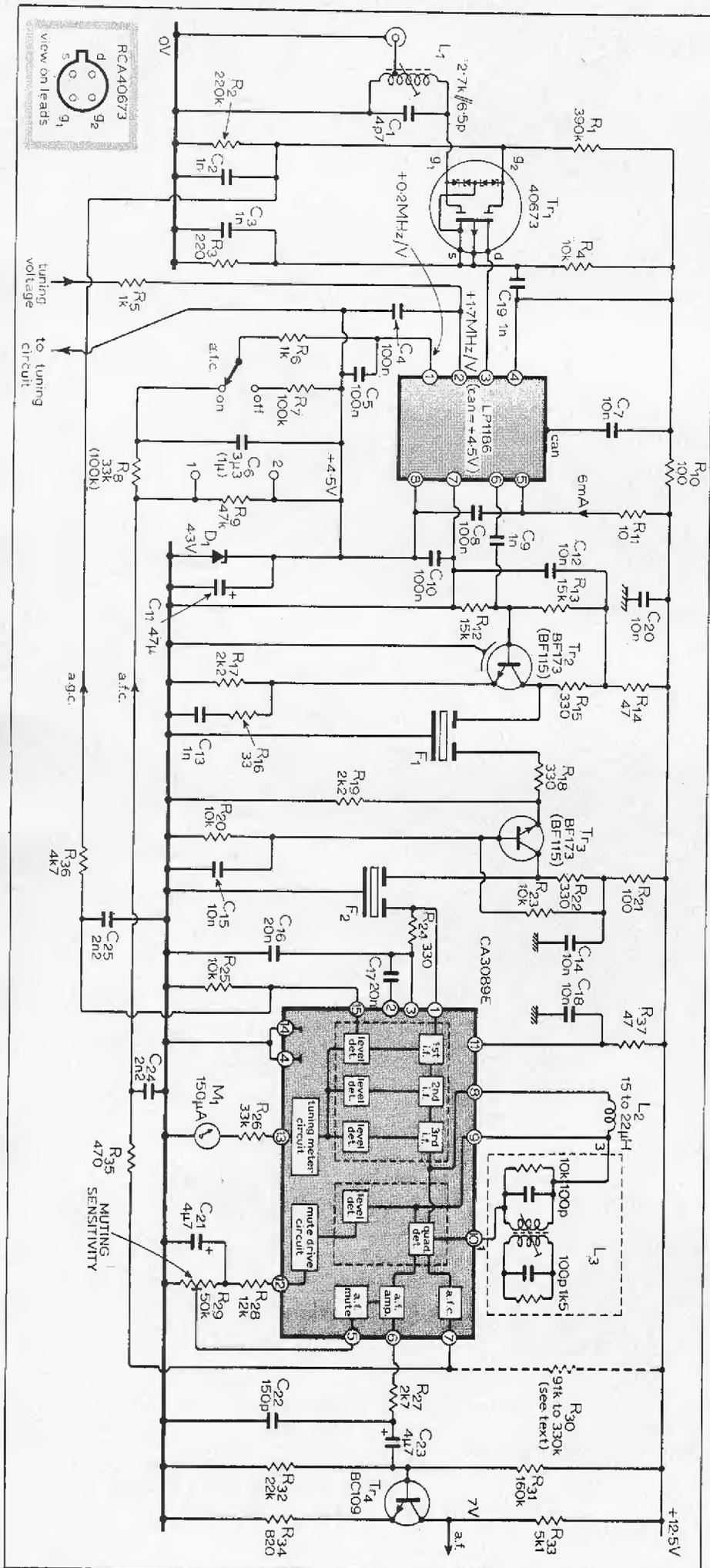


Fig. 5. Improved performance at low signal levels is obtained with this circuit, which uses the tuning and decoder circuits of Fig. 1. Sequence of connections for Tr_2 is b, e, c, screen (on-lead view). On the CA3089E pin 8 corresponds with the tag location.

Although the RCA limiter/demodulator circuit is more complex than its TAA661B counterpart it operates in a similar manner, using an inductive carrier feed to obtain the quadrature reference phase and has the optional dummy tuned circuit to improve linearity of the transfer slope. The external circuit differences mainly concern the use of additional facilities provided by the i.c. Because the a.f.c. signal is derived from a push-pull, open-collector current source in the CA3089E circuit, it is possible that the equal and opposite current condition in a given sample of the i.c. does not occur precisely at the middle of the demodulator S curve. In such a circumstance, a small correcting bias can be provided through a resistor with a value in the 91 to 330kΩ range, connected either to the positive rail, as shown, or to 0 volts, whichever is appropriate. To find the required value and the appropriate supply connection point for this resistor, a method similar to that already described for matching the a.f.c. offset voltage in the simple tuner is suggested; in this instance, however, the S curve is sampled by measuring the voltage across the 150pF capacitor in the pin 6 output circuit.

The completely off-tune condition is used to find the particular voltage value which represents the effective S curve centre and this is then established by tuning to a strong station. Now connect the meter across the a.f.c. sensitivity-controlling resistor, R_9 (points 1 and 2). With the a.f.c. switch off, vary the bias to pin 7 until the measured voltage is zero and remains so with the a.f.c. on. (Note that, as the a.f.c. drive is from a constant-current source, there is automatic compensation for the supply voltage - offset at pin 8 of the LP1186.)

The varying voltage output from pin 13, shown as the meter current in

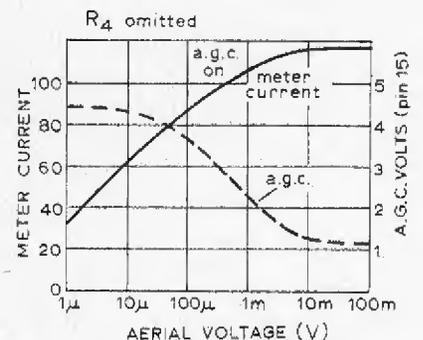


Fig. 6. Curves showing a.g.c. performance and meter current, taken with R_4 omitted. Delayed a.g.c. voltage is at pin 15 on CA3089E.

Fig. 7. Double notch output filter option. Inductors wound on 14mm Mullard Vinkor assembly, with Ferroxcore violet type LA1228. Filter, which has a 6dB loss, should have 25kΩ load.

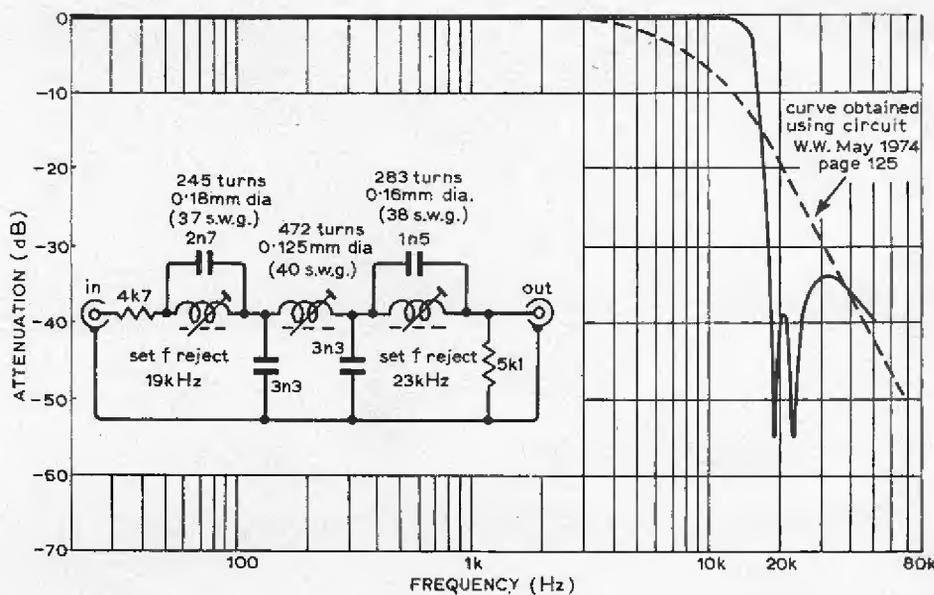
Fig. 6, is not used in this tuner for stereo threshold switching. It can, however, be fed to a suitable meter circuit to give a received-signal strength measurement by relating the indicated current to the calibration curve shown in Fig. 6.

Setting of the audio muting sensitivity control is done by tuning manually through a number of stations and increasing sensitivity until the noise between these is reduced to a minimum. The demodulated multiplex signal, at about 140mV r.m.s. for ±75kHz incoming-signal deviation, is fed via Tr₄ to the 50kHz low-pass filter, decoder and audio output circuits, already described. An extra stage around Tr₄ provides a small amount of gain to compensate for the lower output from the CA3089E demodulator and presents the correct source impedance to the filter.

Optional 15kHz low-pass filter

The output signals from the tuner contain components at the pilot-tone frequency and the switching frequency. Apart from producing noise, these

Component location and p.c. board layout for Fig. 1. Boards for Fig. 1 and Fig. 5 circuits are available from M. R. Sagin, 11 Villiers Road, London NW2, price £3 inclusive, and parts are available from Manor Supplies, 172 West End Lane, London NW6.



unwanted signals can cause difficulty when the tuner stereo output is tape-recorded. If the recording bias beats with one or other of the out-of-band components, or more probably, with their harmonics, then the product frequency could be within the audio band and the resulting signal would produce interference. Such undesirable effects can be prevented by including a low-pass filter in each of the output circuits.

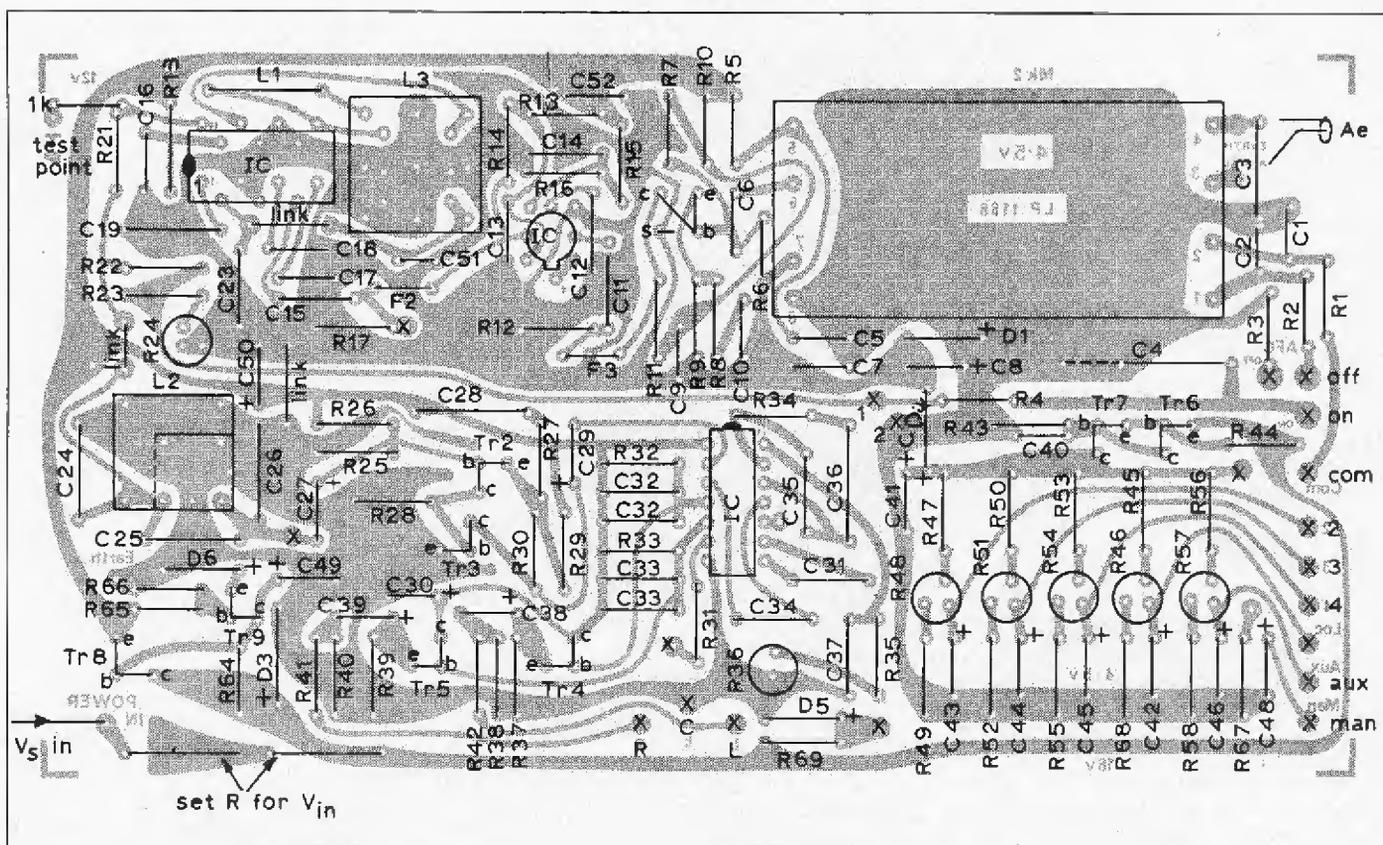
The audio band transmitted is limited to 15kHz, as a necessary factor in normal pulse-code-modulated signal distribution, so it is reasonable to use a sharp filter cut-off at a frequency just above 15kHz. The circuit of a suitable filter is given in Fig. 7 together with its

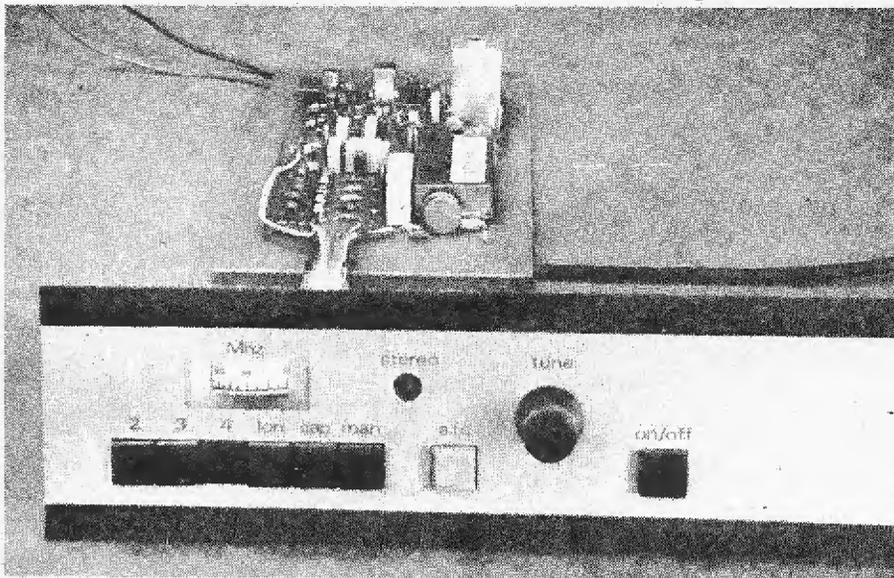
response. The second notch, at 23kHz, is at the frequency allotted to a control signal which the BBC uses for distribution-route and transmitter switching. (An active filter would have required a more extensive circuit requiring many more components to achieve the high rates of response change at cut-off and the notch sides.)

Tuner r.f. and i.f. performance

The four most important figures here are those for i.f. and image rejection, which relate to operation in the r.f. section, and for a.m. and adjacent/alternate channel rejection given by the i.f. circuits.

The first two depend on r.f. circuit





Front panel and controls can be mounted remote from the printed board.

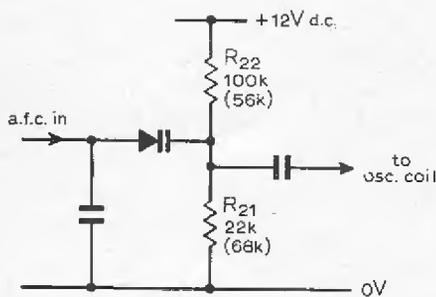
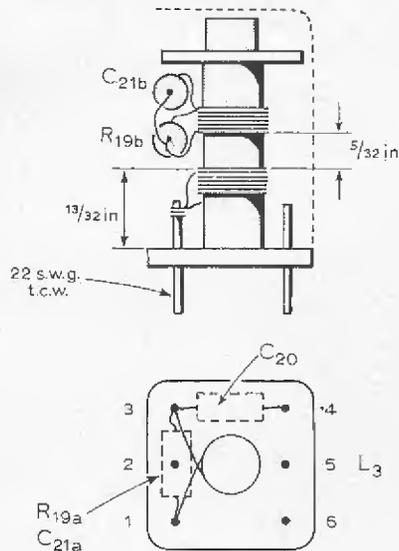


Fig. 8. If Toko EF5600U-1 module is used in place of LP1186 change values of resistors in Toko module to those shown in brackets.



- 1/4 in former, 6mm (9/32 in dia.) Neosid can 1 3/8 in. high 3/4 in. square
- 2 violet cores 6 x 1 x 12-7
- both windings 16 turns 34 s.w.g.

L₃ component numbers above as for Fig. 1. Capacitor marked C₂₀ is omitted in Fig. 5. Simpler coil uses 10 turns on Neosid E3 assembly. Phase shift coil is a Sigma SC10 screened r.f. choke, or Painton equivalent. L₂ Fig. 1 is wound on a Mullard Vinkor assembly with 14mm violet core and 172 turns of 38s.w.g. enamelled wire. L₁ in Fig. 5 is 8 3/4 turns, tapped at 1 1/4 turns, on Neosid 6mm former with 22s.w.g. wire.

The figures for a.m. rejection, quoted from the manufacturers' data for 30% a.m., are -45dB for the SGS TAA661B and -55dB for the RCA 3089E. Performance in respect of adjacent/alternate channel rejection is determined by the i.f. pass-band response characteristic which, for both tuner versions, is the resultant of two FM-4 ceramic filters in cascade. These components were also used by Nelson-Jones, and a curve showing the insertion loss for the combination appears in his original article. This gives the 3dB-down bandwidth as ±110 kHz, and off-tune loss figures of 40 dB at ±200 kHz and 60 dB at ±280 kHz. Rejection of unwanted channels is thus more than adequate.

selectivity and in both versions are determined by the performance of the Mullard LP1186 module. The specification for this quotes an i.f. rejection of 65 dB for 95MHz input and an image rejection of 40 dB.

If better r.f. performance is required, this can be easily obtained but at increased cost by replacing the LP1186 with the Toko type EF5600U-1 tuner module which contains four varicap-controlled tuned circuits and has image and i.f. rejection figures both quoted as 90 dB. A module of this type has been successfully fitted to the author's tuner with some small modifications, as below.

Fitting Toko front-end

Change of tuning voltages. For tuners operated in the London area, the necessary changes to pre-set tuning voltages for stations at the ends of the band are:

	LP 1186 (w.r.t. pin 8)	EF5600U-1 (w.r.t. 0V)
89.1 MHz (Radio 2)	2.4V	3.5V
97.3 MHz (LBC)	6.0V	7.5V

Change of d.c. offset and a.f.c. centre-voltage. The EF5600U-1 tuning voltages are referred to 0 volts instead of the +4.5-volt offset present at pin 8 of the LP1186. This difference necessitates two modifications. First, the 4.3-volt zener, marked D₁ in both diagrams, must be replaced with a shorting link. Second, in the Toko module, the maker's circuit diagram shows that the a.f.c. circuit involves a separate diode with a 2-volt bias obtained from resistors numbered R₂₁ and R₂₂ as illustrated in Fig. 8. Because this circuit is intended for operation with an incoming a.f.c. signal centred on 0 volts, it must be modified to suit the +4.5-volt centre value which obtains in the tuners. The suggested changes are marked in parentheses in Fig. 8, giving an offset of about 6.5 volts.

FM tuner designs

Further details of construction and alignment

by D. C. Read, B.Sc.

As the result of experience gained in building what might be termed production models of the f.m. tuners described recently in *Wireless World* (March and April), the following information on construction and alignment is given, together with suggestions for possible modifications and component alternatives, and some corrections to details already published.

As a board layout was arranged to take specific capacitor types, further details are:

C_1, C_6, C_{15} – polystyrene 2 per cent types were used originally but disc ceramic capacitors are also suitable;

C_{20}, C_{21a}, C_{21b} – polystyrene;

C_{24}, C_{25}, C_{26} – polystyrene (these low-pass filter components need to be as accurate in value as possible);

C_4, C_{32}, C_{33} – 10 per cent polyester, e.g. Mullard type 334.

On the p.c. board supplied for the tuner there are two positions marked for each of C_{32} and C_{33} . This provides for additional components to be installed so that the theoretical 75 μ s de-emphasis time-constant can be obtained very accurately if necessary. Unless otherwise specified, the remaining capacitors are either disc ceramic or tantalum types, the last-mentioned being marked with polarity on the circuit diagrams.

Push-button assemblies

Push-button switch assemblies may be used, but remember that as these are generally equipped with high-value adjustment resistors (100k Ω per section), the reservoir capacitors, C_{42} to C_{48} in Fig. 1, would not then be suitable because tantalum capacitors are subject to considerable variation of leakage current with change in temperature. They are unsuitable for use in a high-resistance circuit within an a.f.c. loop; given a modest change in temperature, the resulting frequency bias created by the consequent change in the tantalum characteristics could so offset the a.f.c. system as to prevent it giving satisfactory overall control.

As a compromise, disc ceramic capacitors of, say, 220nF could be used in these positions to provide a small but useful reservoir effect. Note, however, that without the decoupling action of

the 22 μ F components, the tuning-voltage feed to the LP1186 module is more liable to pick up hum and noise interference and hence allow spurious modulation of the received signal. Thus, if a push-button unit remote from the tuner is installed, it is good practice to screen this feed and/or ensure that it is kept well away from possible sources of interference, e.g. mains wiring. When loaded with the high-resistance selection circuit, the tuning voltage regulator diode, D_4 , may be fed with much more current than is needed to carry out its control function. In this event, R_{44} could be increased, say to 100 Ω .

AFC circuit

In Fig. 5 (April issue) a 3.3 μ F capacitor was specified for C_6 in the a.f.c. line from pin 7 of the CA3089E demodulator. This is an unnecessarily large value because the feed only carries an appreciable audio component under off-tune conditions. A smaller, and hence cheaper, component, say a 470nF disc ceramic or polyester capacitor, would suffice.

Too large an a.f.c. range can be a disadvantage because of station-jumping. If four diodes, arranged as two series pairs connected back-to-back in parallel, are placed across points 1 and 2 in the a.f.c. circuit, the positive and negative voltage excursions are limited each to about 1.2 volts so that the maximum possible tuning frequency change under a.f.c. action is always less than the minimum channel separation.

The other modification concerns extension of the a.f.c. sensitivity control as a front panel facility. This can be done by making R_9 a variable component, still connected between points 1 and 2, but with R_8 taken to the slider. Such an arrangement then enables reduction in sensitivity to be carried out manually whenever necessary but does not prevent the voltage changes appearing across R_9 from being used to

operate the l.e.d. indicator circuit detailed in Fig. 4 of the March issue.

Muting circuit

In at least one of the advanced tuners so far built and aligned, the CA3089E muting output from pin 12 took the form of amplitude-clipped noise instead of varying d.c. The interconnecting circuit feeding the muting input on pin 5 then produced an average of this output which was not sufficient to operate the mute when required. As the i.c. gave a satisfactory performance in all other respects, it was worthwhile making a suitable circuit change to correct for the abnormality. The circuit published in the April issue was therefore modified to give an increased output by disconnecting the existing circuit between pins 12 and 5 and connecting a 1N914 and 50k Ω potentiometer in series between pins 12 and 5 (anode to pin 12). Connect C_{21} between pin 5 and the zero-volt line.

Aerial coil

The aerial coil, L_1 , used to feed incoming signal to the tuned r.f. stage in the advanced version is constructed as shown below. Note that the total number of turns on this coil is 7, not 8 $\frac{1}{4}$ as stated on page 50 of the April issue.

Cut Neosid 6mm former to about 14mm.

Wind on 7 turns 22 s.w.g. tinned copper wire, equally spaced out to 11mm.

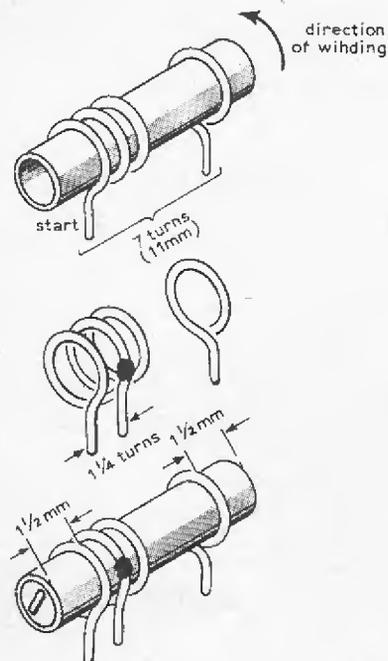
Remove former, tap at 1 $\frac{1}{4}$ turn from start, open turns adjacent to tap to avoid shorting.

Re-insert former and centralise. Screw in core together with p.t.f.e. tape. Coat with Denfix or nail varnish.

In some examples of the advanced version, C_1 may not be required because stray capacitance and the input capaci-

Filter type	3dB bandwidth	± 300 kHz rejection	stopband loss
Vernitron FM4	235kHz	30dB	40dB
Toko CFSA	300kHz	20dB	30dB

tance of Tr_1 added together are sufficient.



Ceramic filters

Difficulty has recently been found in obtaining the Vernitron filters specified for F_1 and F_2 in both versions. Fortunately, ceramic filters from the Toko type CFSA 10.7 range are readily available. These replacement components offer a performance which is not quite as good as that of the Vernitron FM4, but it is generally adequate for the tuners described. The Toko units are also cheaper – about half the cost.

If these alternative filters are used, a small change in Fig. 1 circuit values would be required because the amount of overall phase response correction given by C_{28} and R_{27} in the Tr_2/Tr_3 amplifier is no longer suitable. The values should be changed to 3.3nF and 2.4k Ω . Resistor R_{27} can be adjusted on test to obtain up to 38dB channel separation from 1 to 5kHz.

Adjustment of L_3

The quadrature-phase signal derived for both demodulators, TAA661B in the simple version and CA3089E in the advanced version, could optionally be

produced by means of a double-tuned coil at L_3 . When setting the two cores of L_3 so as to take the best possible advantage of the linearizing effect of current in the dummy coil, it is essential that the cores be kept as far apart as possible in the former, thus minimising changes in coupling between the coils when adjustment is made. The lower core is used to tune the main L_3 coil; it should initially be screwed in just enough to give as symmetrical an S-curve as possible. When this has been set satisfactorily, the upper core is added and screwed in enough to straighten the bend in the transfer slope as in the left hand trace of Fig. 2. More precise adjustment would require the use of a wave analyser or distortion meter.

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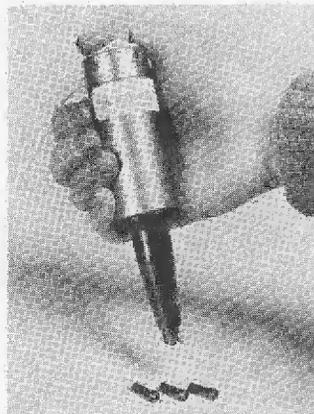
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Novel stereo f.m. tuner

1—Circuits and operation

by J. A. Skingley and N. C. Thompson

Plessey Components Ltd, Swindon

Using a ready-made front-end, integrated circuits and only one inductor, this tuner design is simple to operate, construct and set up. It includes novel circuitry to give inter-station muting, a.f.c. restricted to less than station spacing, a single-lamp tuning indicator, temperature-compensated varicap tuning allowing stations to be preset, and a linear-scale frequency meter. A simple stereo decoder circuit (part 2) uses active filters to eliminate "birdies" and remove subcarrier harmonics. Printed circuit wiring diagrams, assembly instructions and setting up procedure will appear in part 2.

The designer of technical equipment for the domestic market faces a problem. On the one hand the technical operating requirements can dictate a multiplicity of controls and demand a detailed knowledge of their use. On the other hand the operation must be simple and easily understood by people of all walks of life and professions. This design had to cater for non-technical people and children.

The first requirement then was that the system used should mask the technicalities and present the user with the simplest possible mode of operation, relying at most on traditional skills learned from more conventional a.m. receivers, without sacrificing the advantages to be had from a modern f.m. tuner.

The second requirement was to provide a tuner which was at least as good as the best commercial unit on all technical parameters. The total design objectives were therefore

requirement 1—ease of operation:

- (a) provision for push-button tuning
- (b) no undesirable outputs
- (c) powerful a.f.c.
- (d) sensitive, unambiguous tuning indication

requirement 2—good performance:

- (a) $2\mu\text{V}$ for 30-dB signal-to-noise ratio
- (b) 3-dB limiting $0.5\mu\text{V}$
- (c) image rejection 40dB
- (d) i.f. rejection 65dB
- (e) capture ratio 2dB
- (f) a.m. rejection 60dB
- (g) a.f. response $\pm 1\text{dB}$, 20Hz to 15kHz.

The combination of these objectives led to a system which to our knowledge is unique (Fig. 1).

The core of an f.m. system is its i.f. strip, and in this design it was decided to use a block filter and integrated-circuit limiting amplifier. The distribution of selectivity and gain has conflicting requirements. From the point of view of noise selectivity should

come after gain; from the view point of intermodulation effects gain should come after selectivity. The ideal choice is one where the gain and selectivity are uniformly distributed throughout the system, and this was more nearly achieved in traditional i.f. amplifiers using discrete components. The use of integrated circuits however precludes the use of this system because selectivity cannot be integrated.

There are however a number of advantages to the use of block filters over distributed systems. They can be designed as a single entity, providing a shaped response via the controlled interaction between sections to give a complex pole system, and avoiding the need for delicate stagger tuning of discrete sections. Termination conditions are easily allowed for, the filter as a whole is less sensitive to variations in transistor parameters. The filter used in this receiver is the Murata SFG-10.7 MA, which has excellent bandwidth and selectivity (Fig. 2).

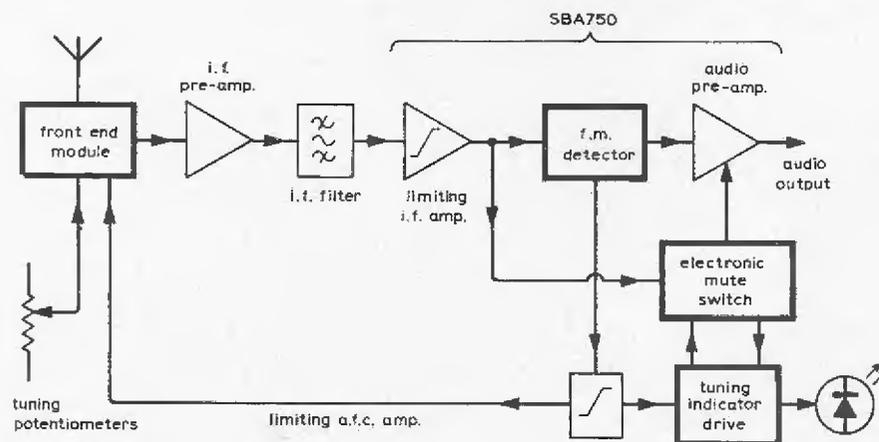


Fig. 1. In this tuner design limiting and demodulation is provided in a five-stage amplifier and a balanced demodulator, both in the SBA750A i.c., additional gain being provided by a two-stage discrete-component preamplifier. Integrated muting circuit eliminates inter-station noise and the novel one-lamp indicator makes tuning simple.

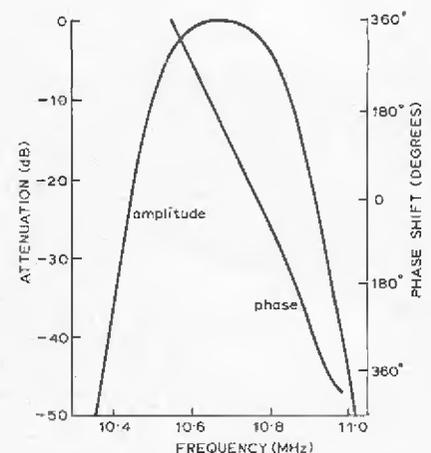


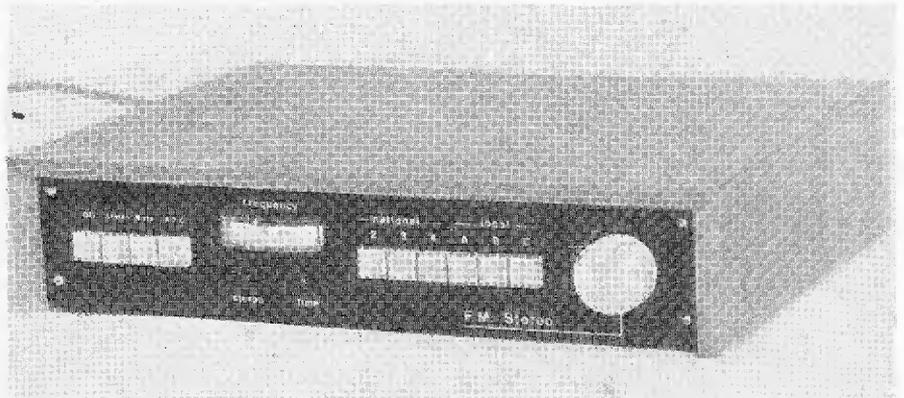
Fig. 2. Filter characteristic is maintained by making filter "see" 330 ohms at source and load.

The integrated amplifier must have excellent limiting characteristics to provide a good a.m. rejection, and the device chosen achieves this by the use of a five-stage limiting amplifier and a balanced demodulator. This is the Plessey SBA750A which has 45dB rejection at 200µV and 60dB at 2mV input. These figures correspond to 2µV volts and 20µV respectively at the input in this design. This device also features a mutable a.f. amplifier which is used in the mute circuit to be described later.

It is interesting to reflect here that, at the present stage of integrated circuit development, the system designer has a wide choice of such building blocks, and, being relieved of the detailed design of these, has far more scope for originality than he used to have. This would seem in contrast to the gloomy forecast once made that all design would be done by the i.c. manufacturer, leaving the system designer the simple task of plugging in devices. In fact it is this new freedom which brings to light the need for new building blocks, and in turn produces more advanced systems.

Objective 1(a), is met by using a varicap-tuned front end. The performance of commercially-available units, although capable of improvement, is equal to our design objectives and presents the simplest solution. The unit chosen was the Mullard LP1186, which can be conveniently mounted on a printed circuit board.

To achieve objective 2(a) more gain is required than that given by the above items. This extra gain is provided by a two-stage feedback amplifier as shown in Fig. 3. The first stage acts as a transconductance amplifier, its gain being defined by the 100-ohm emitter resistor. The second stage then functions as a transresistance amplifier or current-to-voltage converter. The combination therefore has a gain defined by the 100-Ω resistor and the 2-kΩ feedback resistor, and has a value of 26dB. The output impedance of this stage is around 90 ohms,



Features of this simple-to-operate stereo tuner include a muting circuit to prevent unwanted signals being heard (e.g. weak stations, signals affected by flutter, and mistuned stations), a single-lamp tuning indicator and a linear frequency meter.

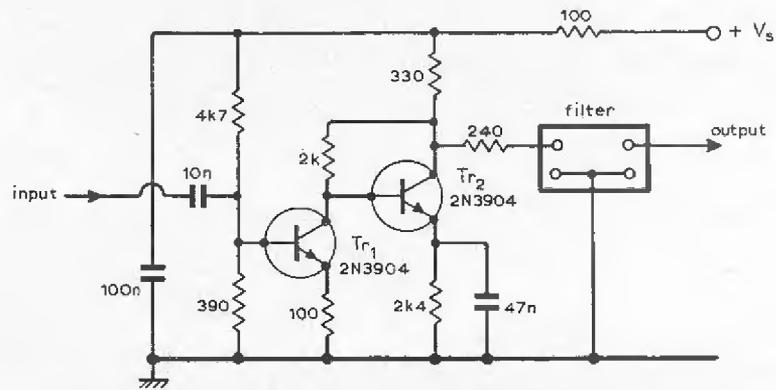


Fig. 3. Feedback amplifier consisting of voltage-to-current converter followed by a current-to-voltage converter has 26dB gain, defined by 100-ohm and 2k-ohm resistors. Output impedance of 90 ohms is made up to 330 ohms to correctly feed the filter.

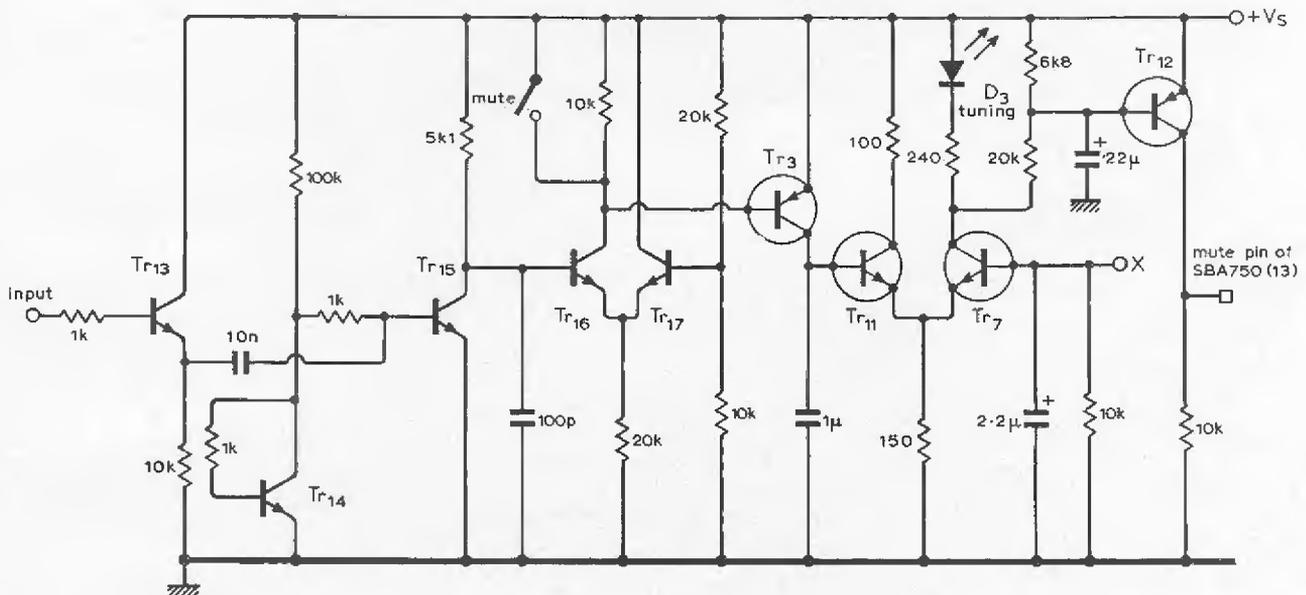


Fig. 4. Mute circuit, operated by the amount of amplitude modulation in the i.f. output, has the advantage of suppressing unwanted outputs due to detection on the non-linear regions of the S-shaped demodulation curve. Point X, fed from the demodulator via Fig. 5, feeds the tuning indicator, which is held off by Tr₁₁ in the presence of noise. An output from this circuit unmutes the stereo output from SBA750 i.c. First five transistors are contained in SL3045 i.c.

and so a buffer resistance of 240 ohms is used to present the filter with its required source impedance of 330 ohms. This introduces a slight loss of gain, but ensures that the correct filter characteristic is obtained.

Muting circuit

The system so far described provides an audio output from an aerial input and could therefore be used as a tuner as it stands. However, this is really where our story begins. If the above system is used, the first thing soon realized is that the interstation noise is highly objectionable. In fact it corresponds to a fully modulated signal over the entire audio band, since the input stage noise is sufficient to achieve limiting in the last i.f. stage, with a bandwidth equal to the i.f. filter pass-band.

In addition to this ear shattering blast, as a station is tuned a highly distorted version of the station programme is received before the correct tuning point is reached, also at a high volume. This is of course due to detection via the "S" curve of the detector, and will be produced at equal intervals either side of the correct tuning point. The net result of all this is design objective 1(b), the suppression of all unwanted outputs. Put the other way round, the only sounds heard should be correctly tuned stations.

Audio muting is achieved by using the remote gain control facility of the SBA750.

This is a pin connection usually taken to a remote potentiometer carrying d.c. only. For our use the potentiometer is replaced by a p-n-p transistor (Tr_{12} , Fig. 4) which is controlled from a number of sources.

Because the noise level is sufficient to produce a fully limited signal from the i.f. amplifier, the magnitude of this cannot be used to detect the presence of the station to un-mute the system, and this presents a problem. The solution is simple in concept.

What is required is a measure of the degree of limiting taking place within the i.f. amplifier, and this is easily monitored by detecting the amount of amplitude modulation present in the i.f. output. This has the advantage of detecting the spurious responses mentioned earlier, as these are in fact caused by the high-slope edges of the i.f. filter response converting the frequency modulation into amplitude modulation which is then detected by the quadrature detector.

The first device in Fig. 4 acts as a buffer to the i.f. amplifier output, which is very sensitive to capacitance loading. The next two devices form the amplitude detector. Transistor Tr_{14} is diode-connected and fed with a small current from the supply. As the five devices Tr_{13} to Tr_{17} are contained on one chip of silicon (SL3045), they have well matched base-emitter voltages, and this causes Tr_{15} to conduct the same current as

Tr_{14} . The 1-k Ω resistors in Tr_{14} and Tr_{15} bases provide a higher input impedance for the signal while preserving the voltage match, since the base currents are also equal.

The collector of Tr_{15} would sit at a relatively high voltage due to its smaller load resistance, but application of the i.f. signal (noise or "clean" 10.7MHz), which is about 400mV pk-pk, causes this stage to rectify bringing its collector down to a low voltage. The collector time constant is chosen so that the 10.7-MHz signal is filtered out but allowing amplitude modulation up to about 100kHz to be followed. This modulation can only be negative, as the i.f. is amplitude limited, and this produces positive-going signals, when present, at Tr_{15} collector.

The following transistor pair is biased with Tr_{17} normally on, so that this positive-going signal turns on Tr_{16} and hence Tr_3 , both of which are normally non-conducting. In the presence of noise therefore the capacitor in Tr_3 collector is charged rapidly taking the base of Tr_{11} positive. This will happen in the presence of any form of a.m. noise or spurious signals, and may be used to mute the receiver.

Single-lamp tuning indicator

There are many ways of meeting objective 1(d). However there is a need to provide an indication which is readily understood by all, without the need for instruction, which

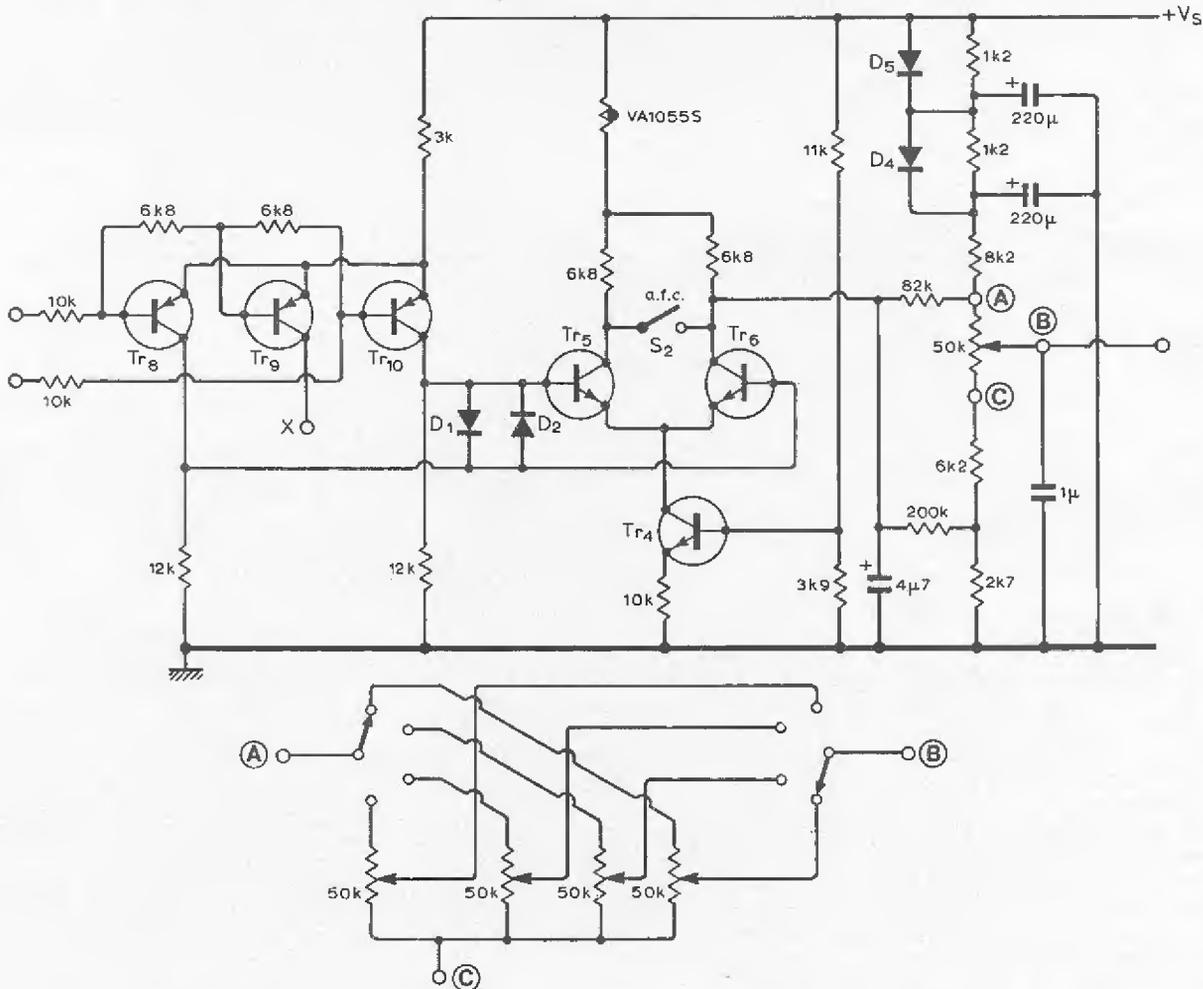


Fig. 5. A.F.C. amplifier has its voltage swing limited to restrict a.f.c. action to less than station spacing. Using a computer programme, the tuning potentiometer network was optimized to provide uniform lock-in range over the tuned band and correct temperature compensation of the front-end. Diodes in potentiometer smoothing circuit allow quick charging of capacitors and keep settling time after switch-on short.

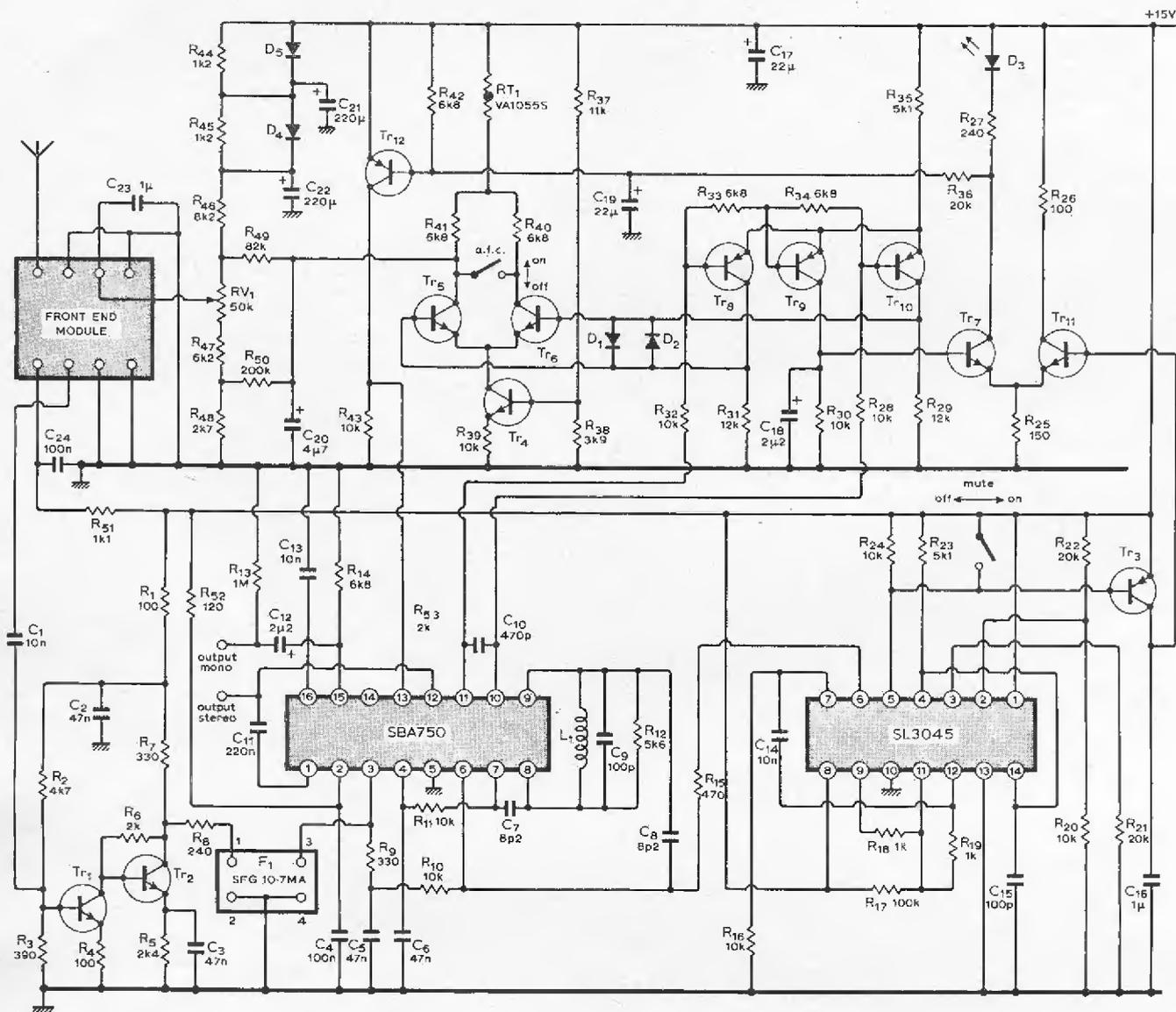


Fig. 6. Circuit of f.m. receive section of tuner with a.f.c., muting and tuning indicator circuits. Frequency is displayed on a moving-coil meter (circuit in part 2). Resistors in this circuit should be 2% tolerance.

ruled out several of the recently-used types. These include the centre-zero meter and the two-lamp system, as neither of these provide a maximum response at the correct tuning point, which is the conventional mode of adjustment. This topic is to be covered in another article and the final design is shown in Fig. 5.

Inputs are taken from the balanced demodulator output via 10kΩ resistors and fed to a long-tailed pair with the addition of a third transistor. This device is connected so that it conducts a maximum current (one third of the tail current) when the inputs are balanced. The collector of this (point X) is connected to the base of Tr_7 (Fig. 4), the d.c. conditions being arranged so that at balance, and assuming Tr_{11} is off, Tr_7 conducts 20mA into the single l.e.d. indicator, D_3 . This then has a maximum brilliance at the correct tuning point.

The Tr_{11} tuning indicator is extinguished in the presence of noise, since Tr_{11} conducts away the current supply to it when its base

is made positive by the action described earlier.

Objective 1(b) can now be easily obtained. We only require the audio signal to be unmuted when there is a correctly tuned station being received, that is, when the tuning indicator is fully lit. The voltage-current characteristics of a l.e.d. are similar to those of a zener diode; therefore a 240-ohm resistor is placed in series to give a voltage-varying point, and this is used to operate the mute transistor Tr_{12} (Fig. 4). The system will only pass a correctly tuned signal of significant signal-to-noise ratio. Very weak stations, stations suffering from gross interference or aircraft flutter, spurious responses, and even strong stations which are mis-tuned by more than a few tens of kHz will be completely muted.

A.F.C. circuit

There now only remains the objective 1(c), the provision of a.f.c. This function received a lot of thought and discussion before the

system described was finalized. A high degree of loop gain was required to reduce the tuning error to negligible proportions, but when this was tried, however, several disadvantages came to light.

Firstly it was found impossible to tune the receiver with a.f.c. applied. One station would be captured and the tuning control rotated past several others on the dial before the original station was lost, and then it was not known which station had been recaptured. The tuning was completely ambiguous, in opposition to the main requirements and objective 1(d) in particular. There are several expensive commercial tuners having this fault.

The solution to this problem is simple in hindsight. What is needed is indeed high gain, but the frequency range should be restricted to less than the typical station spacing. This is easily achieved by limiting the swing voltage available from the a.f.c. amplifier. One requirement for the above system to be successful is that the tuner

Components (main board)

Filter Murata SFG 10.7MA	C_{16}, C_{23}	1 μ F (stack foil polycarbonate)
Front-end module Mullard LP1186	C_{17}, C_{19}	22 μ F (25V electrolytic)
Mains transformer	C_{20}	4.7 μ F (63V electrolytic)
RS Components (634 Trans)	C_{21}, C_{22}	220 μ F (16V electrolytic)
Push buttons		
RS Components (Press 2P)		
Pre-set pots		
RS Components (Lin Cerm 50k)		
Multi-turn pots		
RS Components (Multi pot 50k)		
Regulator		
RS Components (MVR 15)		
Rectifier		
RS Components (REC 70)		
Meter RS Components (MR 42A)		
Semiconductor devices		
IC_1	SBA750A	
IC_2	SL3045	
Tr_1, Tr_2	2N3904	
Tr_3, Tr_8 to Tr_{10}, Tr_{12}	BC179	
Tr_4 to Tr_7, Tr_{11}	BC109	
D_1, D_2	1N4148	
D_3	5082-4403 (l.e.d.)	
D_4, D_5	1N4001	
Capacitors		
C_1, C_{14}	10nF (disc ceramic)	
C_3, C_5, C_6	47nF (disc ceramic)	
C_4, C_{24}, C_2	0.1 μ F (disc ceramic)	
C_7, C_8	8.2pF (min. plate ceramic)	
C_9, C_{15}	100pF (min. plate ceramic)	
C_{10}	470pF (disc ceramic)	
C_{11}	0.22 μ F (stack foil polycarbonate)	
C_{12}, C_{18}	2.2 μ F (63V electrolytic)	
C_{13}	10nF (stack foil polycarbonate)	
	$R_{10}, R_{11}, R_{16}, R_{24}, R_{28}, R_{30}, R_{32}, R_{39}, R_{43}, R_{20}$	10k Ω
	R_{12}	5.6k Ω
	R_{13}	1M Ω
	$R_{14}, R_{33}, R_{34}, R_{40}$ to R_{42}	6.8k Ω
	R_{18}, R_{19}	1k Ω
	R_{17}	100k Ω
	R_{21}, R_{22}, R_{36}	20k Ω
	R_{23}	5.1k Ω
	R_{25}	150 Ω
	R_{29}, R_{31}	12k Ω
	R_{35}	5.1k Ω
	R_{37}	11k Ω
	R_{44}, R_{45}	1.2k Ω
	R_{46}	8.2k Ω
	R_{47}	6.2k Ω
	R_{48}	2.7k Ω
	R_{49}	82k Ω
	R_{50}	200k Ω
	R_{51}	1.1k Ω
	R_{52}	120 Ω
	RT_1	VA1055S thermistor
	R_{15}	470 Ω
	R_{38}	3.9k Ω

signal. Therefore, the resistance is split and two 220 μ F capacitors inserted as shown. This has the desired effect, but produces a slow sweep across the tuning range as these capacitors charge up after switch-on. This is annoying as several stations are captured by the a.f.c. in passing, before the wanted station is reached.

To overcome this defect, it is arranged that D_4 and D_5 diodes rapidly charge the capacitors, by-passing the 1.2-k Ω resistors.

When fully charged, the voltage drop across these is insufficient to keep the diodes forward biased and they cease to conduct, allowing the full time constant to become effective. By this means the switch-on settling time is greatly reduced. These diodes can be seen in Fig. 6 which shows the discrete components and their inter-connection with the integrated circuit and front-end module.

The tuner circuit described provides the optimum performance with regard to the needs of all users. Operation is simplicity itself. The set is silent while tuning between stations and the tuning indicator is un-lit. When a station is encountered the tuning indicator lights fully and the set un-mutes. The tuning control is now left untouched in the knowledge that no further adjustment is required. If an adjacent station exists, the control may be turned further and the tuning lamp will blink off then on again and the new station will be heard; again no further adjustment will be required. This action is similar to turning a rotary switch. If this new station is further away, the set will remain silent until it is reached; mis-tuned stations cannot be received. If the a.f.c. is disabled, for instance when hunting up-band for foreign stations, the tuning indicator allows a sensitive fine adjustment to be made. The mute facility may also be removed under these conditions to allow weak stations of less than 5 to 10 μ V to be received.

The design has been sufficiently thorough to ensure that the performance can be guaranteed, provided that the components specified are used and the layout of the p.c. board is as recommended. Setting up can be done with nothing more than a trimming tool, for the single tuned circuit, and a pair of eyes.

should not have a temperature drift greater than the hold-in range of the a.f.c. This would anyway render the tuning calibration unreliable. In this design the a.f.c. amplifier is used to provide temperature compensation independently of the hold-in range. The circuitry used is shown in Fig. 5.

The differential output from the tuning indicator triple is further amplified by Tr_5 and Tr_6 . Due to the action of the triple there is a common-mode signal present in its output, and this is rejected by Tr_5 and Tr_6 by providing them with a constant-current tail from Tr_4 . Also under extreme mistuning it would be possible for Tr_5 or Tr_6 to bottom, causing a spurious output voltage and incorrect a.f.c. action. Diodes D_1 and D_2 are included to prevent this from happening. They limit the swing available by clamping Tr_8 and Tr_{10} collectors together at a maximum of 0.6 volts difference.

The a.f.c. can easily be cancelled by closing switch S_2 , thus removing differential gain. The output voltage is now determined by the current from Tr_4 , which is fixed, and the total load resistance in the common mode. This includes the thermistor common to both collectors which is selected to provide the correct compensation of the front end, via the following network.

This network was optimized by a computer programme to provide a shift of the end voltages of the tuning potentiometer by

amounts representing equal frequency shifts. This results in a uniform lock-in range over the tuned band, and also correct temperature compensation. A possible method of switching and tuning potentiometers is shown; alternatively these may be connected in parallel provided that the total resistance remains at 50k Ω .

The positive end of the tuning potentiometer, point A, is connected via a total of 10.6k Ω to the positive supply, and any residual mains ripple at this point will modulate the tuning and produce a hum at the output. Less than 0.1mV ripple is required to produce a hum level at least 60dB down on full

Performance

Sensitivity	1.8 μ V for 30dB s/n (mono, see graph in part 2 for stereo)
3-dB limiting	0.7 μ V
image rejection	42dB
i.f. rejection	60dB
capture ratio	1.8dB
a.m. rejection	60dB (see graph)
mute level	3 to 5 μ V
output level	180mV stereo 1.7V mono
harmonic distortion	0.5% for 75kHz dev.

Novel stereo f.m. tuner

2—Stereo decoder, assembly and setting up

by J. A. Skingley and N. C. Thompson

Plessey Company Ltd, Swindon

Fig. 7 shows the internal circuit of the SBA750. Pins 3 and 4 receive the input from the i.f. which is passed along the chain of five limiting amplifiers to pins 6 and 7. From here the signal is passed internally direct to the quadrature detector, and also externally via the quadrature coil and pins 8 and 9. The demodulated signal is then available at pins 10 and 11, and it is from here that the drive for the a.f.c. and tuning indicator is taken. This audio signal is taken internally to an amplifier giving a single-ended output on pin 12.

The stereo signal is taken from this pin and fed to the stereo decode board. It is also coupled via C_{11} to pin 1, and after further amplification is available de-emphasized as a mono signal on pin 15. De-emphasis is accomplished by C_{13} on pin 16. Both stereo

and high-level mono outputs are therefore available.

The amplifier output to pin 12 can be attenuated by varying the current fed into pin 13. In this design pin 13 is open circuited to fully mute the amplifier while preserving the a.f.c. drive from pins 10 and 11. This ensures that the receiver captures a station from the muted condition.

Improved stereo decoder

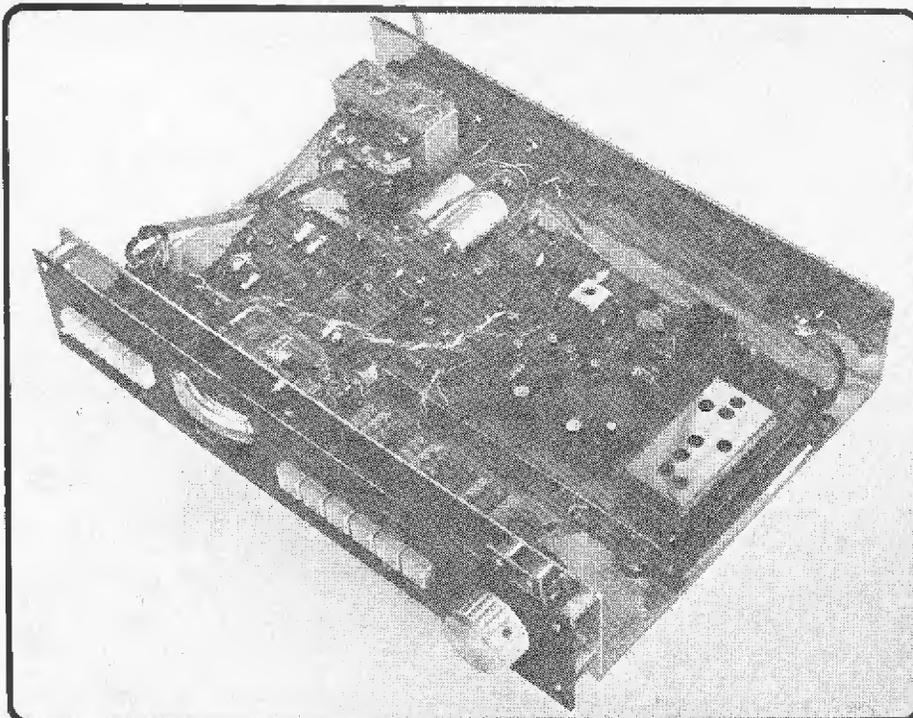
The stereo decoder is shown in Fig. 8. When this decoder integrated circuit was first used "birdy"-type interference was experienced under certain conditions. The causes of this have been reported elsewhere, but it is worthy of further explanation judging by the lack of effort to remove it in expensive receivers.

The nature of a frequency modulated signal is such that a bandwidth many times that of the deviation is needed for accurate transmission and reception, around 300kHz often being used. The spacing of broadcast programmes is however only 100kHz and this inevitably results in frequencies from one station arriving at the detector of a tuner receiving an adjacent station. The products from the detection are however normally supersonic and therefore inaudible.

This is fine until we introduce stereo reception which involves demodulation of the stereo channel at 38kHz usually using square-wave switching. This process also demodulates signals around the odd harmonics of 38kHz, i.e. 114, 190, 266kHz etc. The first two of these will produce audible signals from the adjacent channels at 100 and 200kHz away from the wanted station, giving interference centred on 14 and 10kHz respectively. These sound like high-pitched twittering sounds commonly called birdies.

Knowing the cause, the effect can be greatly reduced, if not completely eliminated. The wanted stereo information extends up to 53kHz, so by filtering the signal above this frequency before the stereo decoder, the unwanted adjacent channel signals can be attenuated. This also brings about an improvement in signal-to-noise ratio during stereo reception, as noise above 53kHz is also reduced. Such noise can be demodulated down to the audio band by the harmonics of the 38kHz switching frequency in a similar fashion to the adjacent channel signals if it is not removed.

The filtering required is carried out by Tr_1 in Fig. 8, which also shows the complete stereo decoder. (This circuit has been built on a separate board, and for this reason the components have been numbered independently). Transistor Tr_1 forms an active filter of the Sallen and Key type and provides a second-order response. There is an addi-



tional pole supplied by C_{10} between pins 10 and 11 of the SBA750 and this, together with the two poles of the active filter, combine to give a three-pole optimally flat response up to 53kHz, followed by a sharp roll off of 18dB per octave. This transistor is directly coupled and biased from pin 12 of the SBA750, and its output fed to the input of the decoder integrated circuit.

This decoder is of the phase-locked loop type, and in appreciating the advantages of this form of decoder, it is worth looking briefly at the conventional type. These decoders work by generating the necessary 38kHz switching frequency either by converting up the 19kHz pilot tone, or by phase locking a local oscillator to this pilot tone. The 38kHz signal may then be used to switch the multiplex signal into its two separate paths. Obviously, this switching must not only be at the correct frequency but it must also be in synchronism with the original coding. In other words the phase of the coding and decoding signals must be identical. If a tuned circuit is used as in the conventional decoder to separate the 19kHz from the audio, then a high-Q tuned circuit must be used to avoid phase jitter from the residual audio. A high-Q tuned circuit however has a high phase shift for a small amount of mis-tuning. For this reason a low Q would be required. Hence there is a compromise.

The phase-locked decoder operates by generating a 38kHz signal (in this case a 76kHz signal divided by two). This is divided by two to give 19kHz which is compared in phase with the pilot tone. The difference between the two is used to provide

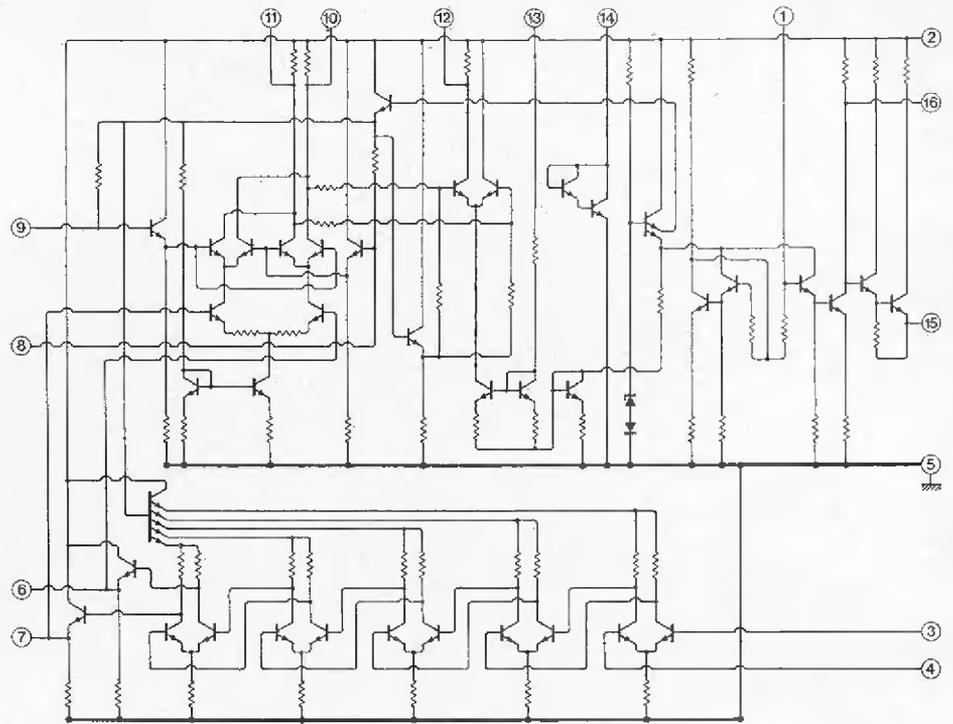


Fig. 7. Internal circuit of SBA750A i.c. includes circuitry i.f. amplifier (below) balanced quadrature detector and a.f. preamplifier (used on mono only). Signal to unmute output is applied from Fig. 4 to pin 13. Drive for a.f.c. and indicator circuit is taken from pins 10 and 11 via circuit of Fig. 5 and drive for mute circuit is taken from pin 6.

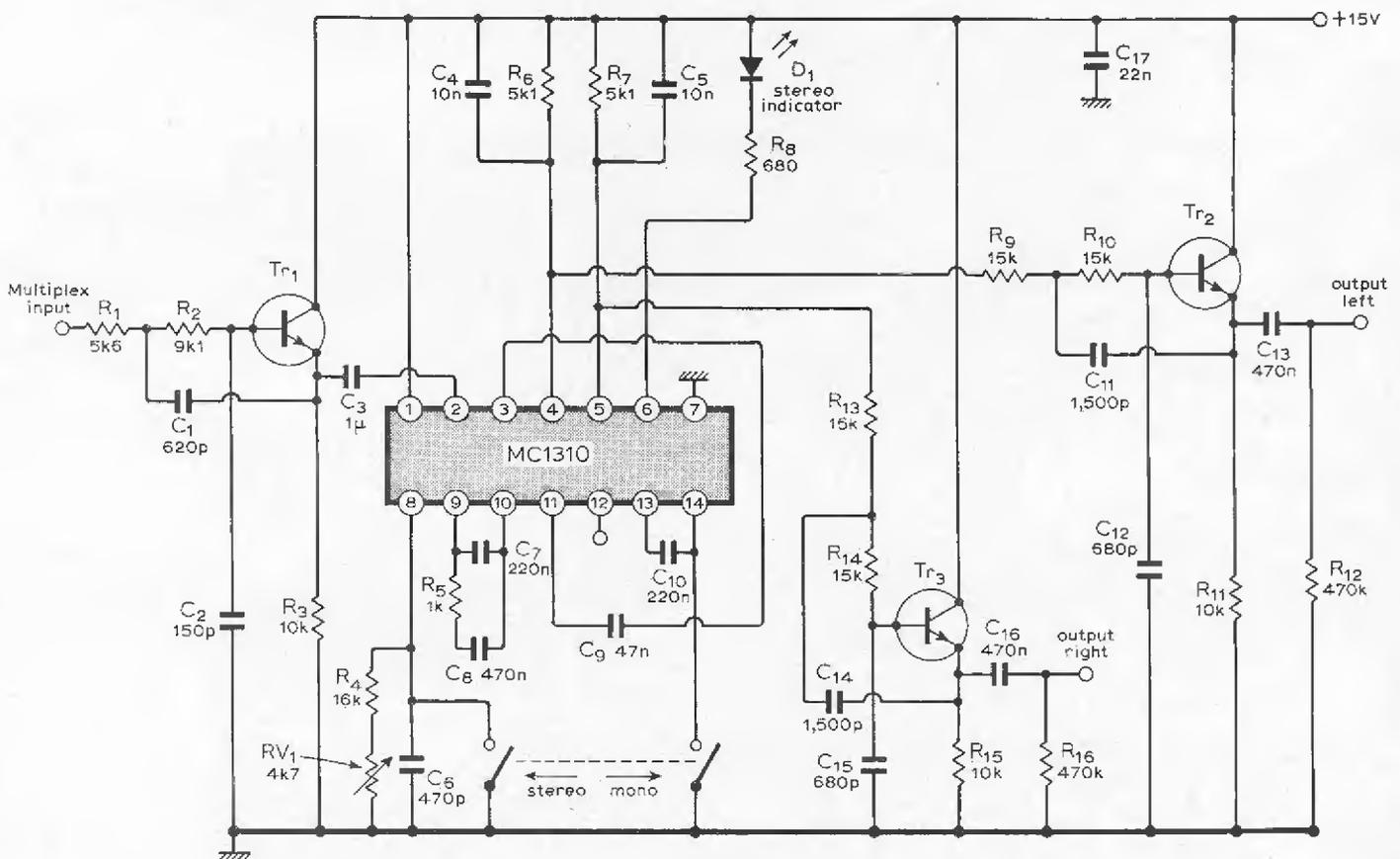


Fig. 8. Decoder circuit includes an active filter to roll-off response at 18dB per octave above 53kHz to prevent "birdies" that result from interference between odd harmonics of 38kHz and adjacent carriers. Two further active filters remove the 38-kHz harmonics from the outputs.

feedback, and hence phase lock the local oscillator to the pilot tone. The time constant of the feedback path may be made long so as to reduce the phase jitter to a negligible amount. This is equivalent to producing a high-Q tuned circuit, but one which cannot drift in phase provided the loop gain of the feedback path is high. This system also has the advantage that no coils are required. The complexity required dictates the use of integrated circuits on economic grounds, which has the added advantage that fewer corners need be cut in the design stage, so that the full potential of the system may be realized.

Two more active filters have been added, one per channel, formed by Tr_2 and Tr_3 (Fig. 8) and these are directly biased from the integrated circuit. Their function is to remove unwanted signals from the outputs, such as the 38kHz sub-carrier and its harmonics, which could otherwise cause trouble when tape recording.

The MC1310 has a direct output for a stereo indicator lamp and the facility for disabling the decoding process if desired. In the tuner described this is implemented by a second pole to the switch which also stops the oscillator to prevent any possible interference. The stereo decoding may need to be stopped if the signal is weak and a poor signal-to-noise ratio is obtained. Reverting to mono reception will provide an improvement. You may prefer to use the mono output provided from the receiver board, but this will need attenuation to give a compatible level when switching to mono. The tuner will, of course, automatically give a mono output in the absence of a pilot signal.

Construction

Layout and general presentation of the tuner is largely a matter of personal choice, and in this connection the mechanics de-

scribed represent only our solution, offered as a suggestion. The layout of the printed boards are critical, and it is strongly advised that the board design offered here is used. The system employs a generous amount of gain at high frequencies and even small deviations from the layout given could prove troublesome. This layout, given in Figs 9 and 10, follows good engineering practice and ensures stable performance.

When assembling a board of this com-

plexity it is a good idea to insert a few components at a time, solder these and clip their leads before inserting a few more. Start with the passive components and finally the transistors, integrated circuits, filter and front end. There are four wire links and these are made using discarded resistor ends. The single coil is 15 turns of 33 s.w.g. cotton cord wire, close-wound on a Neosid type A screened assembly using an F16 screw core. Before soldering the wire ends of this, insert the capacitor C_9 (100pF) into the same pins

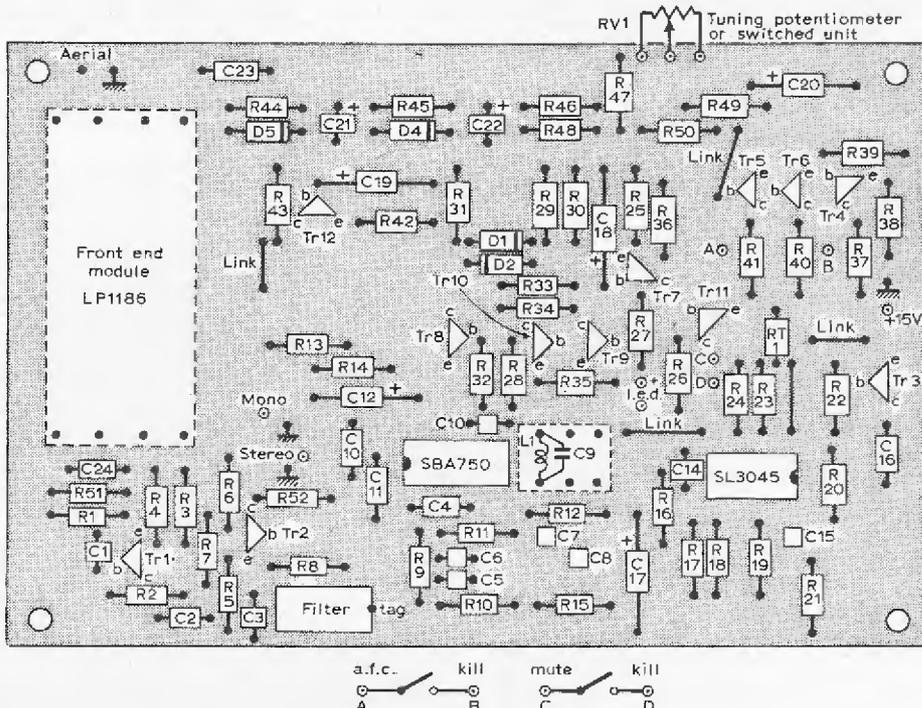


Fig. 9. Component layout for Fig. 6 is critical and p.c. board shown in Fig. 10 should be used. Points A and B connect to the a.f.c. switch and points C and D to the mute switch.

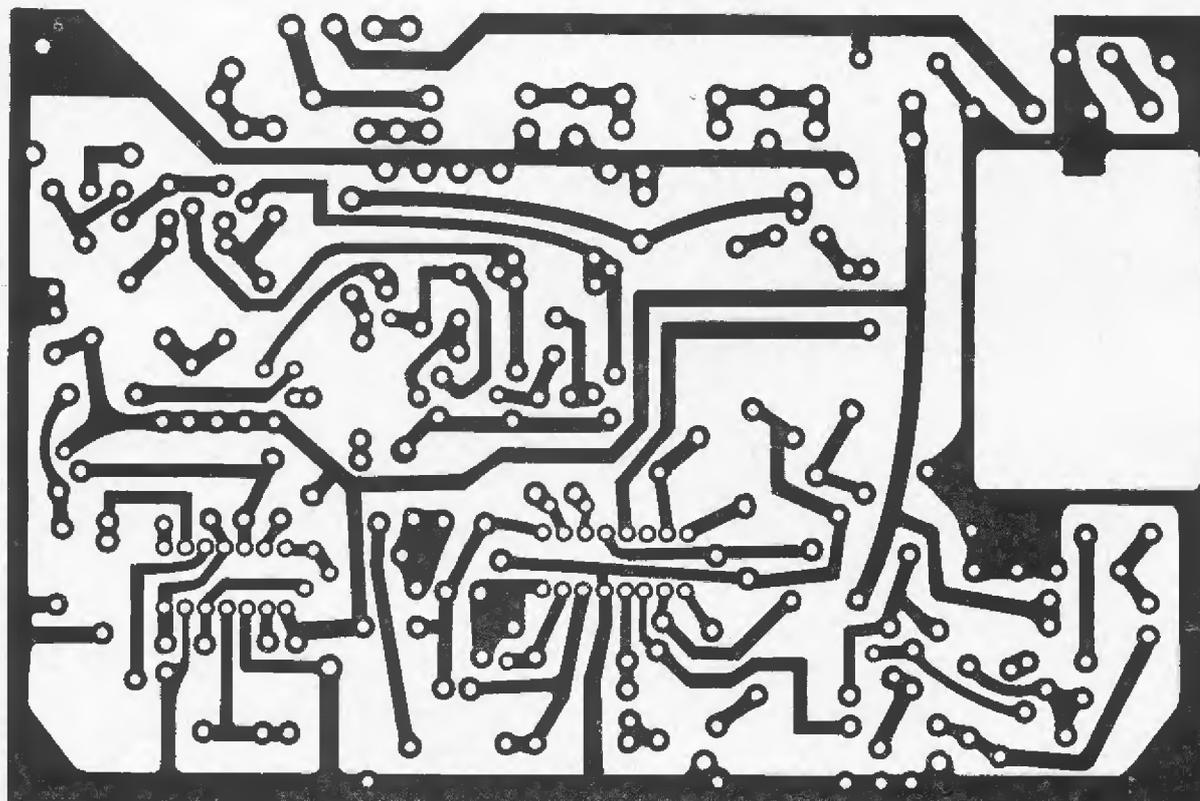


Fig. 10. Copper side of printed board, actual size. Component side shown above.

as the coil. This capacitor is slim and is easily accommodated within the can.

The emitters of the transistors are adjacent to the tag on the can. Transistors Tr_1 and Tr_2 are an exception to this. Insert so that the flat face of the plastic package is on the opposite side to the base of the triangle. If the centre (case) lead is bent forward towards the flat all these leads will fall into place.

Finally, solder twisted-pair wires at points A and B for the a.f.c. switch and C and D for the mute switch, together with pairs for the tuning indicator lamp (observe polarity), audio outputs and 15-volt power supplies. Mount the two boards on $\frac{1}{2}$ in brass pillars at the four corners, preferably on a metal chassis to ensure good earthing and screening.

The chassis system (Fig. 12) was constructed from 16 s.w.g. aluminium sheet, a piece 11in square being required. After drilling the chassis is bent into a U-shape where shown, and the front panel fitted on $\frac{1}{2}$ in brass pillars. This can be made of aluminium, sprayed and marked with Letra-set. Alternatively, perspex may be used, marked in mirror image and sprayed on the reverse side. There are many ways open to home constructors these days, and the production of a professional finish is largely a matter of ingenuity and personal taste.

The front panel is made $\frac{1}{4}$ in deeper than the front edge of the chassis to allow the use of rubber feet on the chassis, yet leaving a smaller clearance beneath the front panel. A false front panel has the advantage of hiding most of the screws and allowing the push buttons and meter to protrude the correct distance. It also allows space for a hidden pilot lamp to illuminate the meter. A cover made of polished wood or sprayed metal may easily be made to fit over this chassis, consisting of inverted U-shape forming the sides and top. This should be about 8 or 8 $\frac{1}{2}$ in from front to back, depending on the front and back overhang desired, and the sides the same drop as the front panel.

The only difficulty is in mounting the ten-turn potentiometer which has a short spindle and must be brought forward $\frac{1}{4}$ in from the chassis. This is achieved by cutting a clearance hole in the chassis 1-in square and mounting the potentiometer on a strip 1 x 2in held on $\frac{1}{2}$ in pillars, or long screws and double nuts. The spindle will then protrude from the front panel by the correct distance.

The prototype used six push buttons for pre-selected stations mounted on a printed board with the pre-set potentiometers between them. This resulted in a compact switch unit, but the method necessitated a minor trimming of the width of the potentiometer by rubbing their faces with emery paper until a sliding fit was obtained. This board also held a drive circuit for the meter used to display the frequency. This is shown in Fig. 13, and the p.c.b. in Fig. 14. This unit may be mounted on 0.1in pitch Letrokit board and hand wired if preferred, using the p.c.b. layout as a guide.

The meter drive circuit is basically an emitter follower driven from the tuning voltage, but the addition of a Plessey SL301 matched pair as shown results in a tempera-

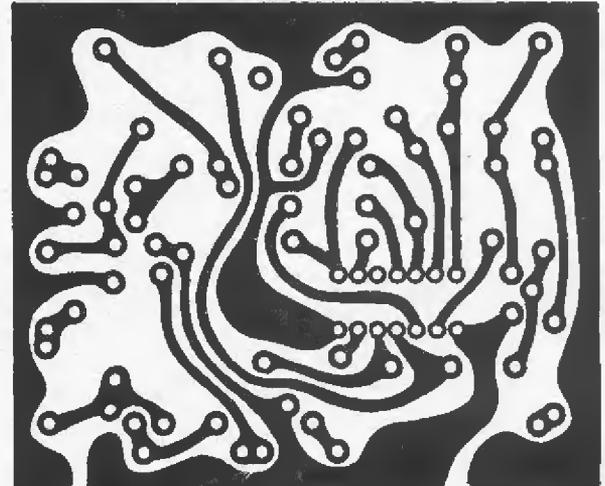
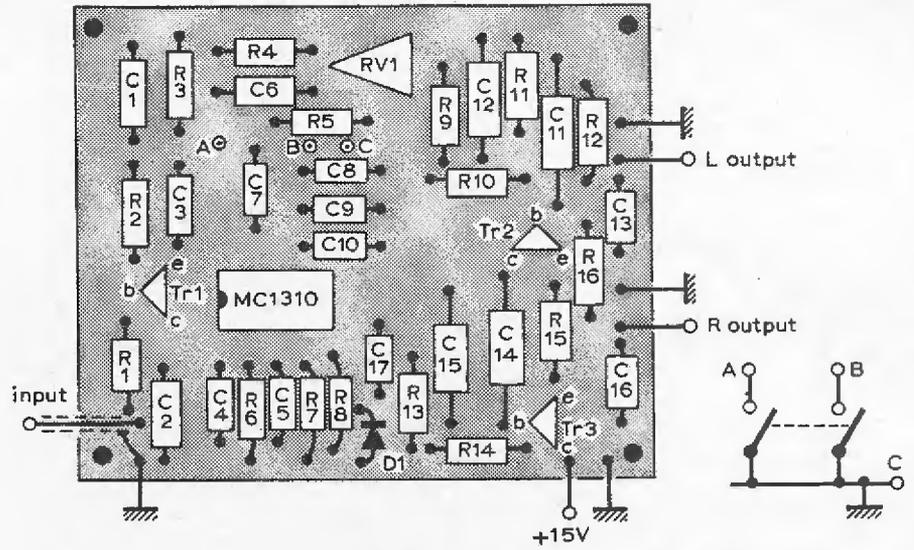
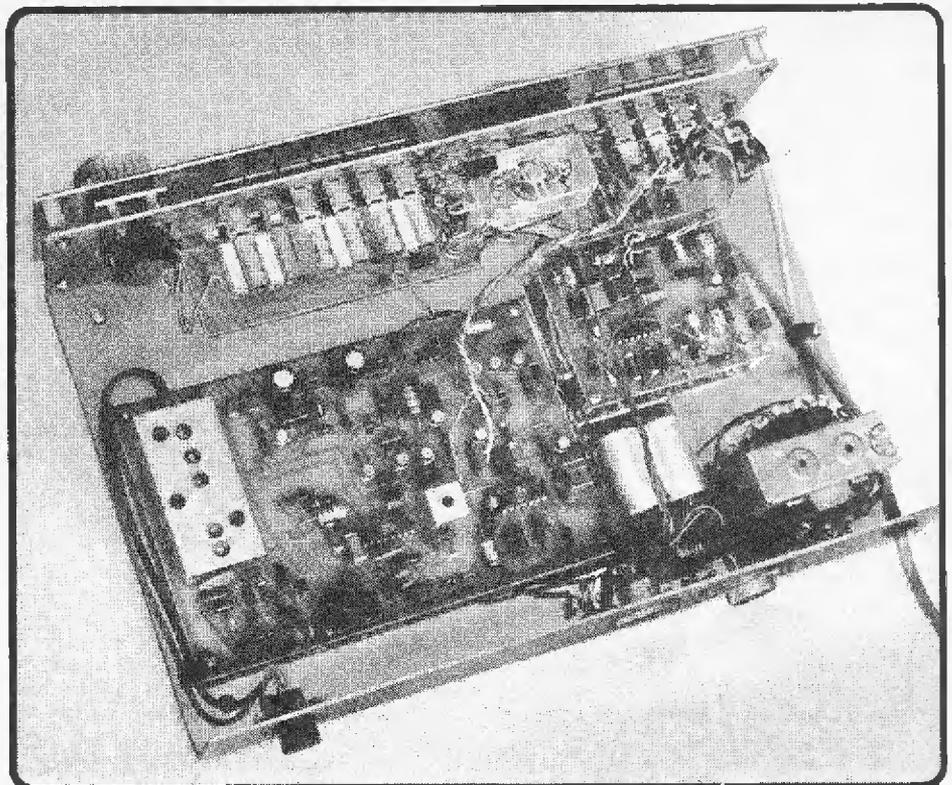
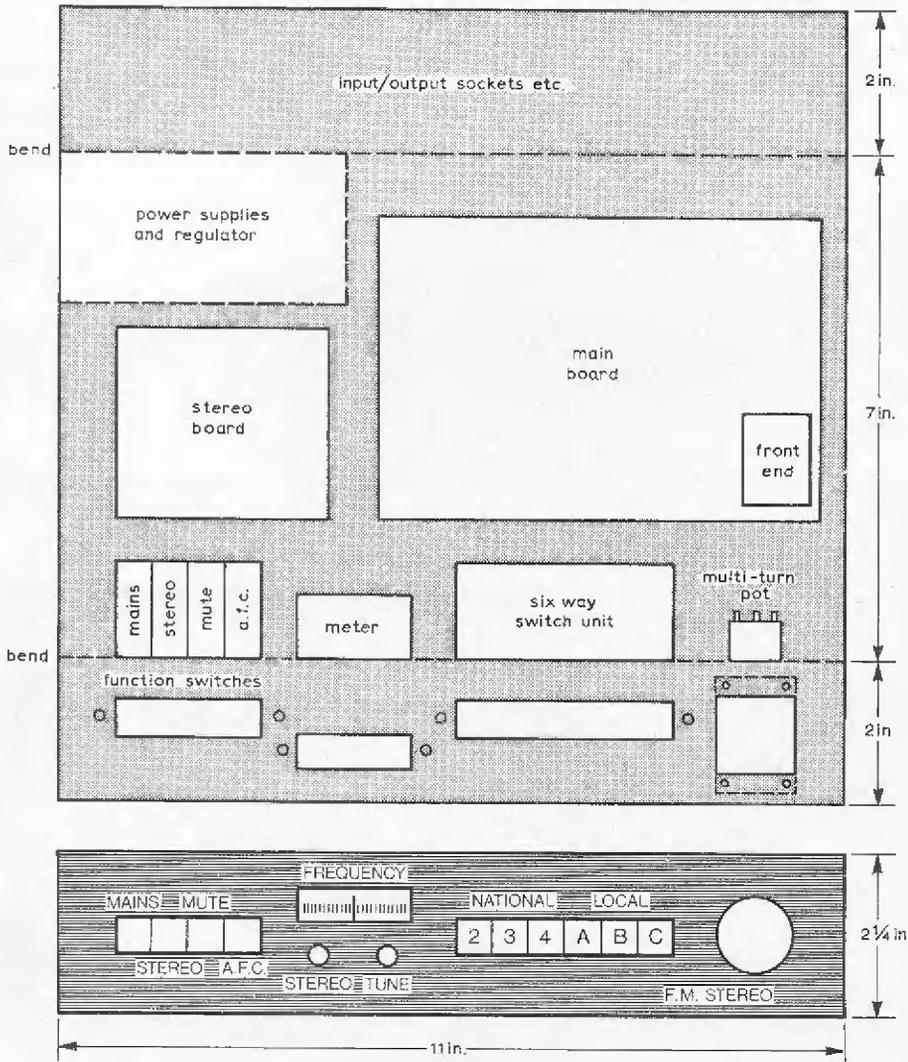


Fig. 11. Stereo decoder layout. Points A, B and C connect to mono/stereo switch shown actual size.





Chassis for complete receiver can be made from an 11in square 16-s.w.g. sheet of aluminium, bent into a U-shape (Fig. 12, above). Separate front panel and cover improve appearance. Drive circuit for frequency meter gives linear frequency scale, varicap non-linearity being matched by matched-pair integrated circuit (Fig. 13, below). Layout on right shows pre-set potentiometers and frequency meter components (Fig. 14).

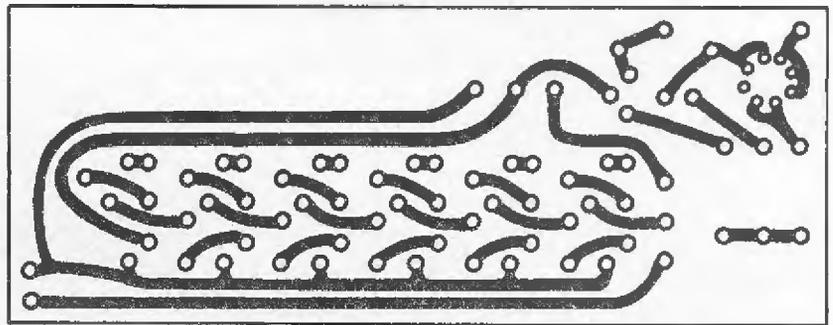
ture-stable law-bending circuit producing a linear frequency scale from the non-linear law of the front-end varicaps. The meter is and RS Components miniature edge meter MR42A, 25-0-25 μ A. By removing the two screws the case may be removed and the scale lifted from its mounts. A wipe with acetone or nail varnish remover removes the lettering which may then be replaced with a suitable frequency scale using Letra-set—or free-hand if you have a steady one! Take care to avoid damage to the pointer, and the case should be replaced to exclude dust while the scale is being redrawn. Final calibration is done by mechanical adjustment of the zero adjuster while receiving a known station.

It is important that the power supply should be free of ripple and temperature stable, and this is achieved by a regulator integrated circuit RS Components MVR15V. This, together with bridge rectifier REC70, 1000 μ F capacitor, and transformer (634) from the same supplier complete a stable power supply for little effort.

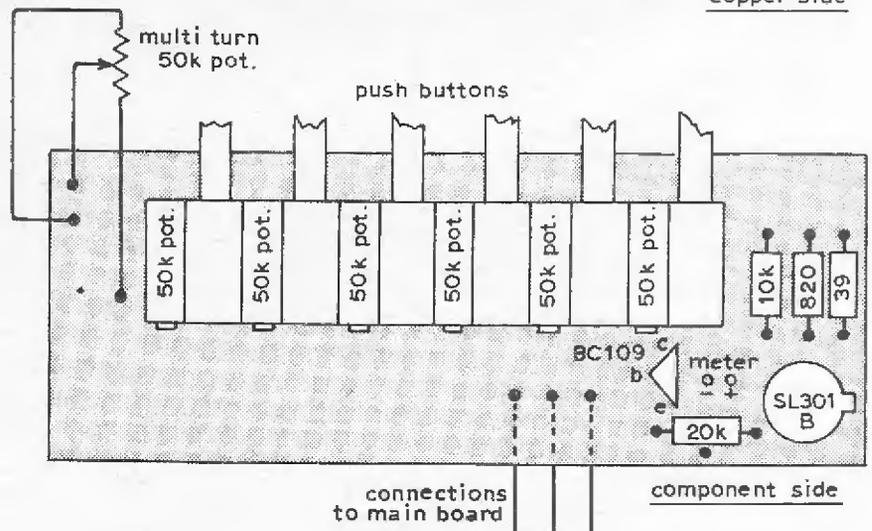
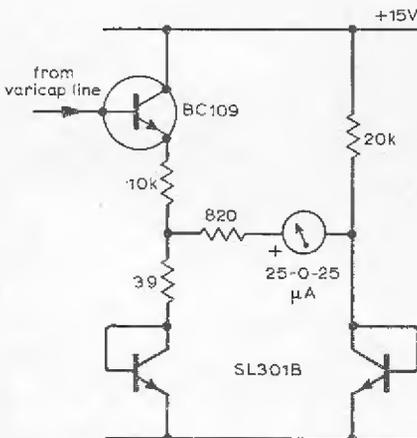
Top-grade components were used to ensure reliability and consistency of performance. It is strongly recommended that the components specified are used. There are parts of the circuit which require 2% resistors, for example, to ensure correct biasing and balance of the tuning point and a.f.c. circuitry.

Setting up and testing

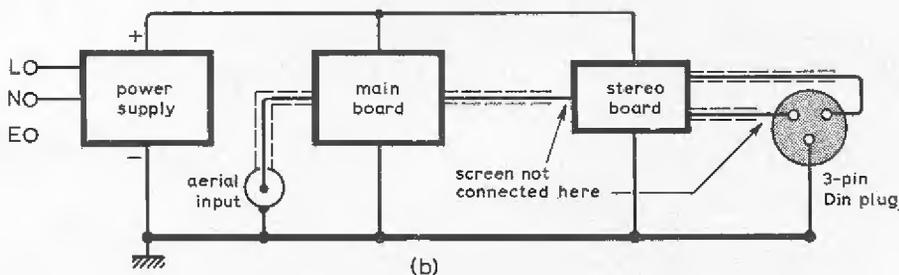
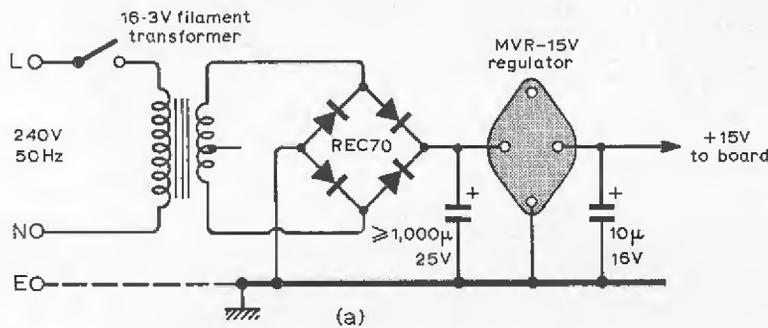
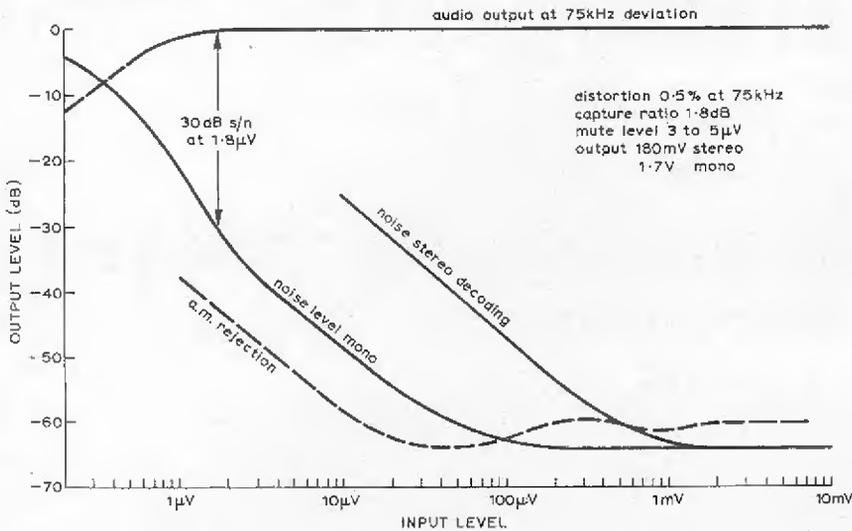
When the boards have been wired and mounted make the appropriate interconnections. Connect l.e.d.s and switches and check everything before switching on. Put switches in the a.f.c.-off and mute disabled



copper side



Tuner performance



Components for suitable power circuit, above, were listed in part 1.

positions, and leave the aerial unconnected. Do not adjust the front-end module which is pre-aligned and should not be touched. Under these conditions there will be a band of noise defined by the i.f. filter passing through the limiting amplifier to the detector. If the core of the single coil is adjusted until the tuning lamp responds and be trimmed for maximum brilliance, the detector will be correctly adjusted to the centre of the i.f. pass band.

Now connect an audio amplifier and speaker. A loud smooth hiss should be heard. If the mute switch is operated this hiss should be silenced and the tuning lamp extinguished. Connect the aerial—a short length of wire may receive several stations—and, with all push-buttons out, adjust the ten-turn potentiometer to find stations. Observe a.f.c. action by mis-tuning a station until the tuning indicator is just extinguished and the output muted. Switching in the a.f.c. should recover the station.

A good aerial will now reveal anything around a dozen stations. The three least noisy should prove to be the national stations, and the frequencies of these are given in the local Radio Times. Trim the meter using the internal adjuster and obtain pre-set stations by pressing the appropriate button and adjusting the adjacent pre-set potentiometer. If this adjustment procedure cannot be achieved switch off and check all components and interconnections. Particularly check for incorrect polarity i.e.ds and capacitors, misplaced n-p-n an p-n-p transistors, capacitors omitted in coil assembly or wrong number of turns.

If the alignment of the main board is achieved set the stereo decoder. This involves adjusting the oscillator using the single pre-set potentiometer. First find a transmission known to be stereo; check with the *Radio Times*. Ensure that decoding is not disabled by the switch provided and adjust until the stereo indicator lights. Now

Components (stereo decoder)

Transistors BC109
IC MC1310
Diode 5082-4403 (i.e.d.)

Resistors

R_1	5.6kΩ
R_2	9.1kΩ
R_3, R_{11}, R_{15}	10kΩ
R_4	16kΩ
R_5	1kΩ
R_6, R_7	5.1kΩ
R_8	680Ω
$R_9, R_{10}, R_{13}, R_{14}$	15kΩ
R_{12}, R_{16}	470kΩ
RV_1	4.7kΩ

Capacitors

C_1	620pF (polystyrene)
C_2	150pF (polystyrene)
C_3	1µF (polycarbonate)
C_4, C_5	10nF (polycarbonate)
C_6	470pF (polystyrene)
C_7, C_{10}	0.22µF (polycarbonate)
C_8, C_{13}, C_{16}	0.47µF (polycarbonate)
C_9	47nF (polycarbonate)
C_{11}, C_{14}	1.5nF (polystyrene)
C_{12}, C_{15}	680pF (polystyrene)
C_{17}	20µF (polycarbonate)

release and pre-press the stereo switch and adjust the potentiometer until the i.e.d. lights in the shortest possible time after the switch is pressed. This is the correct setting.

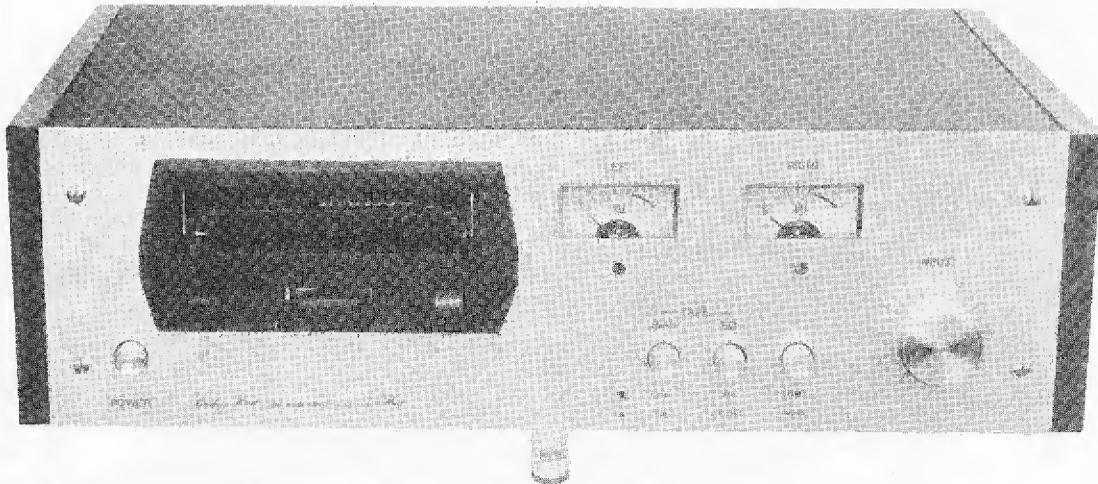
Finally, a word about aerials. This tuner will receive stations on a few feet of wire, but if the full potential and maximum signal-to-noise ratio is to be obtained, a good aerial is essential. Any system can only be as good as its signal source, be it a pick-up cartridge, tape head or radio aerial, and this can easily be the weakest link in the chain.

Modification. A resistor (R_{53}) of 2kΩ should be added in place of the wire link shown just below R_{43} in Fig.9. Because Motorola have altered the specification for the MC1310, a 150-Ω resistor should be added in the supply line to the decoder board, returning the i.e.d. current directly to the +15V rail. Increase C_{17} to 20µF to decouple this.

HART ELECTRONICS

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J. L. Linsley-Hood High Quality Cassette Recorder



As these circuits are capable of such an excellent performance we feel that it is not sensible to sacrifice this potential by designing a kit down to a price. We have therefore spent a little more on professional hardware allowing us to design a very advanced modular system. This enables a more satisfactory electrical layout to be achieved, particularly around the very critical input areas of the replay preamps. These are totally stable with this layout and require no extra stabilising components. Many other advantages also come from this system which has separate record and replay amps for each channel plugging in to a master board with gold plated sockets. The most obvious is the reduction of crosstalk and interaction which could cause trouble on a single plane board, with our modular system the layout is compact but there is no component crowding. Testing is very easy with separate identical modules and building with the aid of our component-by-component instructions is childishly simple, but the finished result is a unit designed not to normal domestic standards but to the best professional practice.

All printed circuits are of glassfibre material, fully drilled with a tinned finish for easy and reliable soldering. Component locations are printed on the reverse side of the board and are arranged so that all identification numbers are still visible after assembly.

- 71x Complete set of parts for Master Board, includes bias oscillator, relay, controls, etc. £9.83 + £1.23 VAT.
- 72x Parts for Motor Speed and Solenoid Control for Lenco CRV deck. This is the proper board layout as given in the articles. £3.52 + 44p VAT.
- 73x Complete set of parts for stereo Replay Amps. and VU Meter drive £8.12 + £1.02 VAT.
- 74x Complete set for stereo Record Amps. £6.74 + 84p VAT.
- 75x Complete set of parts for Stabilised Power Supply to circuit given in Article. This uses a special low hum field transformer with better characteristics than the commonly used toroid. £8.79 + £1.10 VAT.
- 700M2 Individual High Quality VU Meters with excellent ballistics. £8.48 + £1.06 VAT. Per Pair.
- 700C2 High Quality Custom built steel Case. Complete with Brushed aluminium front plate, mains switch, record microswitch, turned record level knob, plastic cabinet feet, all bolts nuts and mounting hardware. All necessary holes are punched and all surfaces are electroplated. Complete step-by-step assembly instructions are included. The cover is finished in an attractive black crackle surface. £16.50 + £2.06 VAT.

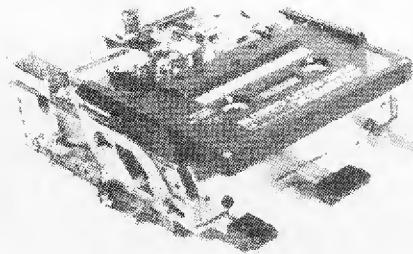
TOTAL COST OF COMPLETE KIT £83.08 + £10.39 VAT.

ALL PARTS ARE POST FREE

Please send 9 x 4 SAE for lists giving fuller details and Price breakdowns.

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Personal callers are always welcome, but please note we are closed all day Saturday

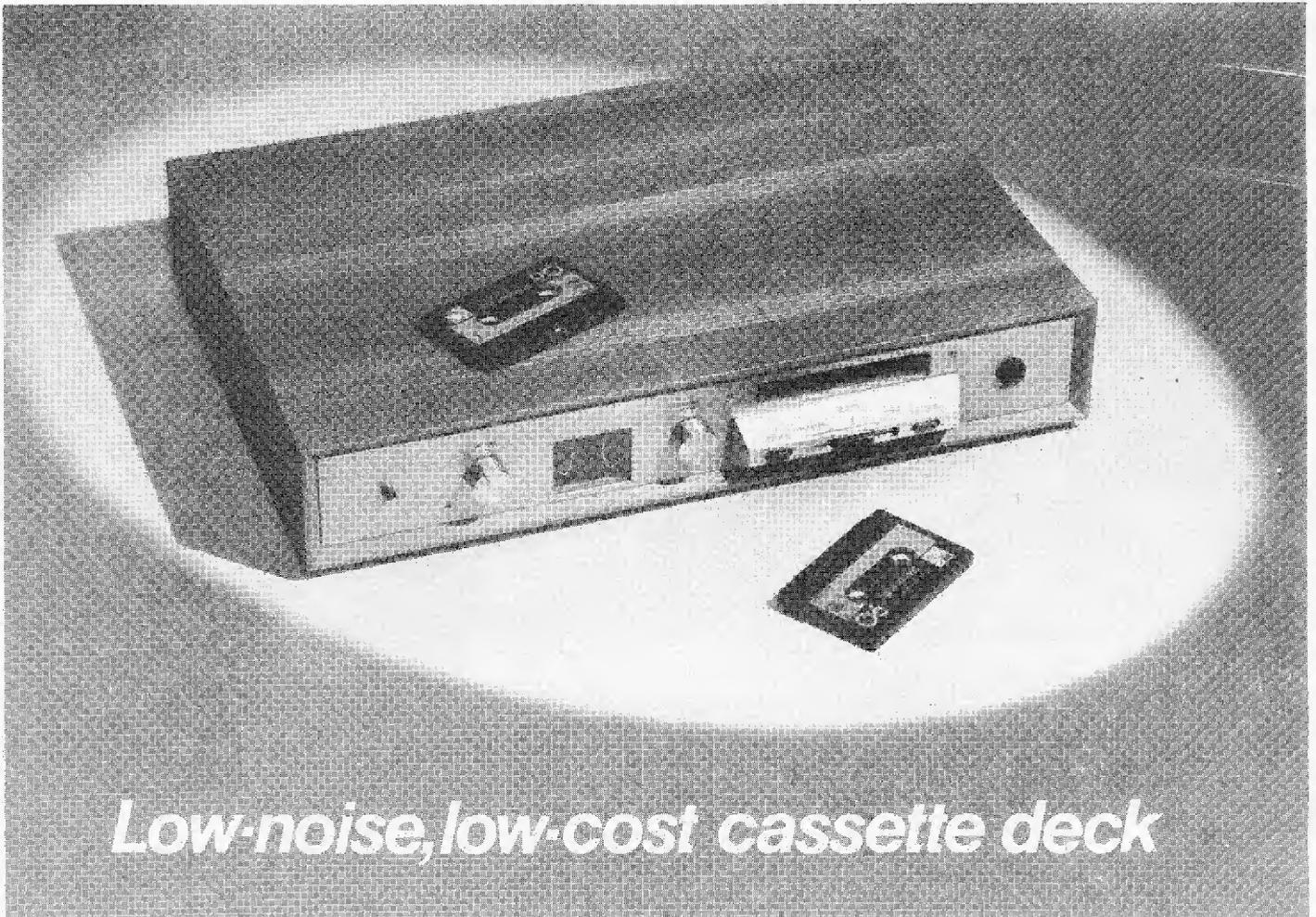


LENCO CRV CASSETTE MECHANISM

High Quality, robust cassette transport for Linsley-Hood recorder. Features fast forward, fast rewind, record, pause and full auto stop and cassette ejection facilities. Fitted with Record/play and erase heads and supplied complete with Data and extra cassette ejection spring for above horizontal use. Price £21.80 + £2.70 VAT.

We also supply complete kits to make a fully integrated 30 watt stereo amplifier using the Bailey Power Amplifier circuit and the Bailey/Burrows Pre-amplifier with the Quilter Tone control modification.

Printed circuits and components are available for the Stuart tape circuits. These articles described a high quality tape link circuit for use with a reel-to-reel deck. Reprints of the three articles are available from us price 40p. Post Free (No VAT).



Low-noise, low-cost cassette deck

For some years the author had contemplated the possibilities for the provision of music of reasonable technical quality, by way of headphones, while away from home on camping holidays – which were normally taken in scenically attractive but physically remote parts of the countryside. Of the available alternatives, the use of previously recorded tape cassettes seemed the most satisfactory, but it is unlikely that further action would have been taken on this matter but for the current availability at an attractive price of good-quality cassette mechanisms made under Star patents by Garrard and Goldring-Lenco.

It must be explained, at the outset, that the intention was not to provide an instrument which would equal or exceed that of expensive and carefully engineered "transcription" cassette recorders, but rather to evolve a straightforward and relatively inexpensive circuit arrangement which would nevertheless provide a standard of performance which would be acceptable in the context of existing, high quality, audio equipment. In the event, the performance of the prototype has substantially exceeded expectations, and has led to a major revision of the author's opinion of the performance obtainable from this medium.

In particular, it would appear that, with good system design and appropriate attention paid to recording and bias levels in a direct recording made

High-quality design for mains/battery use

by J. L. Linsley Hood

from a good quality l.p. disc onto a reasonable quality ferric-oxide cassette tape, the major component of noise on replay is likely to be the surface noise on the original disc. Also, the differences between the source material and the cassette transcript can be sufficiently small that they are not readily apparent, even on A-B comparison.

Basic circuit

The general layout of the system adopted is shown in Fig. 1. The d.c. power supply unit has two outputs – one of about 12-14V at 200-400mA to feed the d.c. drive motor which operates the cassette feed, and which has its own speed control system incorporated by the manufacturers, in the case of the Garrard CT4 used in the prototype – and one having a well-smoothed and electronically stabilized output preset to a nominal 13.5V, which feeds either the replay or record amplifiers. Between

these two lines there are two change-over switches, to the centre point of which can be connected a 12-14V d.c. supply, so that the system can also be operated from batteries.

The changeover switch in the amplifier supply line is a small microswitch, not supplied with the cassette mechanism but operated by a protruding tag on the side of the record push button on the mechanism. To make a recording, this is depressed before the cassette is inserted, when a mechanical interlock retains the button in the inward position. When the d.c. supply is connected to the record amplifier panel, it also energises a 12V, three-pole change-over relay connected in parallel with it. This relay transfers the connections from the combined replay-record heads from the input to the replay amplifier to the output of the record amplifier. Under normal replay conditions, neither the relay nor the record amplifier panel are energized. The bias/erase oscillator is mounted at the output end of the record amplifier and is supplied with power when this panel is energized. By using separate record and replay amplifiers some additional component cost is incurred, but the internal switching is greatly simplified.

Replay amplifier

The use of the extremely low tape speed of the Philips cassette design, coupled with the small head gaps necessary for good high frequency response, and the

better surface recombination noise characteristics of the n-type base material, at an appropriately low collector-to-emitter voltage, say 3 to 4V, with as low a collector current as is permissible and a base circuit impedance giving a suitable compromise between Johnson noise and device noise figure requirements. In the case of the Texas Instruments BC214LC, the optimum collector current and base circuit impedances are $10\mu\text{A}$ and about 800 ohms. This gave, on the prototypes of this amplifier, a measured noise referred to the input of some $0.2\mu\text{V}$ which is only slightly above the predicted Johnson noise value for the known input impedance and equalized bandwidth. In practice, the input noise introduced by this stage is sufficiently less than that of the tape background for it to be unimportant as a contribution to the overall system noise figure.

In the second stage of this amplifier, where the replay equalization (frequency/amplitude response shaping) is performed, a good-quality integrated operational amplifier "gain block" is employed, as in all the other gain stages of the system. The unit chosen is the Motorola MC1741CG, which is a fairly standard 741 but in an 8-pin TO39 metal-can encapsulation, and is, in the authors experience with these devices, much to be preferred on grounds of reliability. Two equalizing characteristics are provided, having $70\mu\text{s}$ and $120\mu\text{s}$ upper time-constants. Of these, the former is the internationally agreed standard for chrome tape, and the latter is the normal standard for ferric types.

The output from this amplifier, about 0.4volts r.m.s., at 0VU and 660Hz, is taken to the output socket, and the VU meter through an isolating silicon diode. A similar isolating diode on the output of the record amplifier circuit allows the VU meters to be used both on record and replay settings, which is

useful for assessing tape output characteristics, and the recording levels of recorded cassettes.

The two replay characteristics are shown in Fig. 3, and are determined by the switched values of R_{10} , R_{12} and C_6 , C_8 . Some additional treble lift to compensate for head limitations is given by R_9 , C_5 and gives rise to the part of the curve indicated in Fig. 3.

Although the author has some personal reservations about the use of series feedback configurations in the case of magnetic pick-up input equalization arrangements, where at the upper end of the recorded frequency range it is possible to generate relatively large pickup output voltages with consequent risk of distortion due to common-mode failure, in the case of cassette replay heads the likely output voltages are so small in relation to the input device C_{be} voltage that this is a negligible problem. Also, to design for the lowest practicable noise level, series feedback configurations remain the simplest form to implement, although in higher-speed, higher-output recorder systems it could be worthwhile to introduce feedback, around an inverting amplifier, at a low impedance at the earthy end of the playback coils.

To avoid replay head magnetization problems due to switch-on current surges through the replay coil windings on the charging of an input series capacitor, the replay coil is connected between the input reference voltage source and the base of the input transistor, so that the total current flow through this is limited to the base current of this device — about $0.1\mu\text{A}$. (Head magnetization is less of a problem on record due to the demagnetizing effect of the fairly large bias voltage applied to it during recording. It is, however, important that the time constant of the record output circuit should be shorter than that of the decay of bias voltage, which is ensured by the use of fairly substantial capacitor values on the record amplifier positive supply line.)

The measured total harmonic distortion of the replay amplifier, input to output, at up to 1V r.m.s. output, is less than 0.01%, and a very high degree of

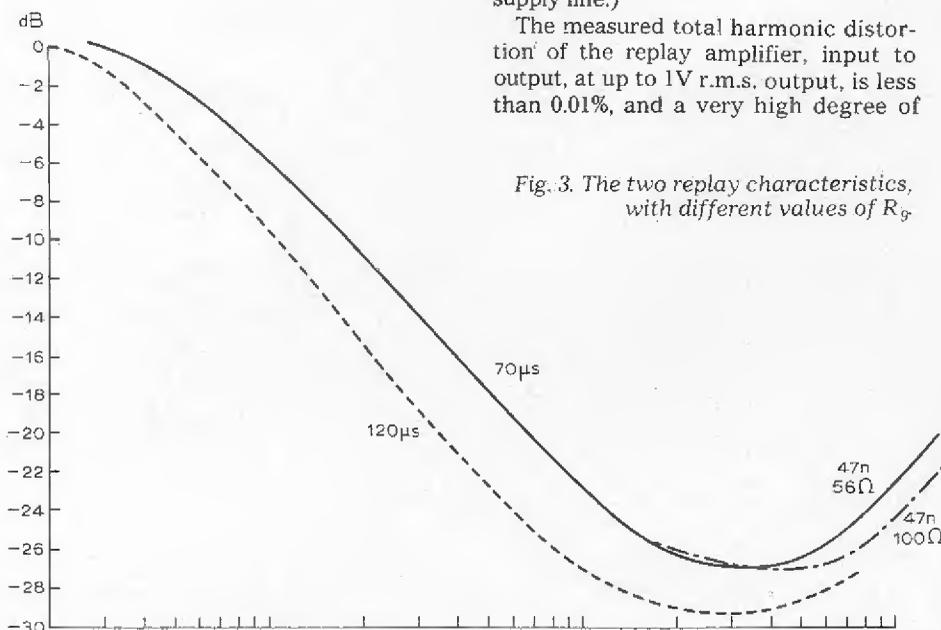


Fig. 3. The two replay characteristics, with different values of R_9 .

Measured performance figures of prototype (Garrard CT4 mechanism)

Frequency response $\pm 1\text{dB}$ 35Hz
 —12kHz (BASF LH Super C90)
 Channel separation 45dB at 1kHz
 Erasure better than 50dB
 T.h.d. at 0VU (660Hz) 0.75%*
 Replay amplifier background noise, CCIR weighted, —56dB.
 Zero recorded level background noise, CCIR weighted, —52dB.
 Bulk erased tape background level, CCIR weighted, —54dB.
 The above figures refer to a 1kHz tone recorded at 0VU on BASF LH Super C90. Both channels are identical to within 1dB.
 Record amplifier t.h.d. at +3VU less than 0.02%
 Replay amplifier t.h.d. at +3VU 0.01% (Residual distortion less than background noise at —6VU.)

*This figure should be considered in the context of typical disc replay figures (e.g. 1.2% and 0.6% harmonic distortion for 20cm/s at 1kHz, vertical and lateral modulation respectively) for a good-quality pick-up cartridge in a good-quality arm, rather than in comparison with the less than 0.1% t.h.d. typical of a good-quality audio amplifier.

h.t.-line noise and ripple rejection is given by the use of a constant-current-source load (Tr_2) in the first stage.

Record amplifier

Since the design value of input sensitivity for this amplifier is not very high — 50mV r.m.s. input at 1kHz for a 0VU record level — great care to obtain a high signal-to-noise ratio is unnecessary (the difference in recorded noise obtained by replacing the input MC1741CG with a very low noise circuit such as that used in the replay amplifier is only of the order of 0.75dB). A simple amplifier design based on a pair of these operational amplifiers is therefore entirely adequate, and confers a number of minor advantages in addition to those of simplicity and economy of component cost.

To avoid the necessity for winding coils for the generation of the required peaky record characteristic (desirable to offset shortcomings in the head performance, tape and recording characteristics at the upper end of the recording range) an active RC circuit arrangement is employed. This is shown in the circuit diagram of Fig. 4, and consists of the network R_{16} , R_{17} , C_{12} , C_{13} in conjunction with R_{10}/VR_2 and C_{15} . The recording characteristics obtainable from this are shown in Fig. 5, for various component values, which may be of use if it is desired to use different record heads to those supplied with the Garrard CT4. The magnitude of the pre-emphasis hump in the 13-15kHz region is determined by the setting of VR_2 (a preset component on the circuit board), while the basic recording treble

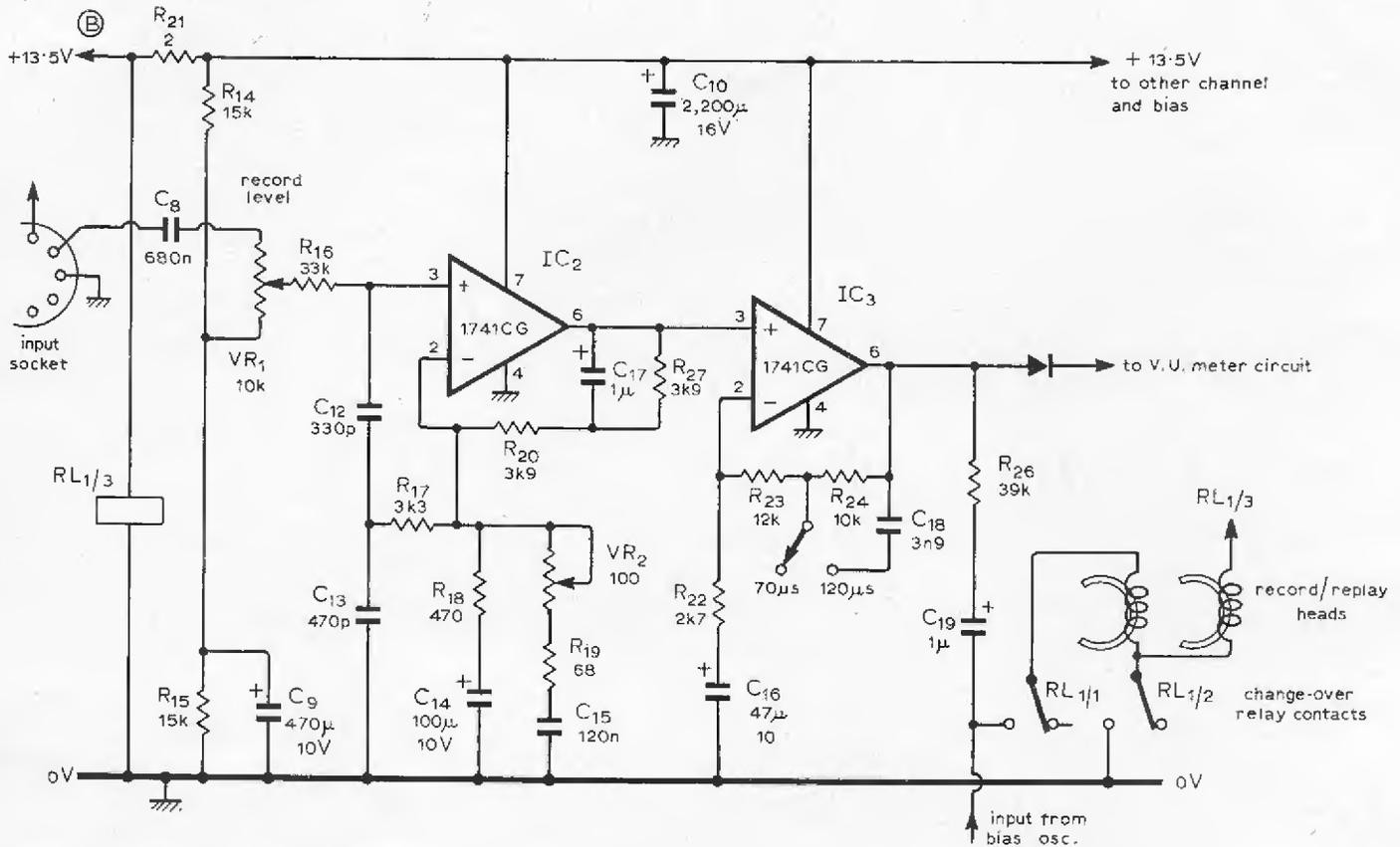


Fig. 4. Recording amplifier.

lift time constants are determined by C_{12} and C_{15} .

Changeover from the basic $70\mu s$ recording characteristic to the $120\mu s$ one is by switching C_{18} into circuit. The new cassette-standard bass pre-emphasis at $3180\mu s$ is provided by C_{17} , R_{27} . A $39k\Omega$ swamping resistor is interposed between the output of the record amplifier and the head, to approximate to a constant-current recording condition. Since the impedance of the head at the upper end of the frequency range of the recorder is less than $10k\Omega$, the loss of h.f. due to this is small, and readily compensated for in the equalizing circuitry. With this value of output swamp resistor, attenuation of the bias voltage by the low output impedance of the 1741 is sufficient to eliminate the need for any additional bias-trap circuit, while allowing record amplifier circuit outputs of up to $+3VU$ with less than 0.02% t.h.d. at $1kHz$.

With the recording heads used in the prototype, a $0VU$ record level at $660Hz$, chosen to avoid regions in which pre-emphasis characteristics would influence the result, corresponded to $2.25Vr.m.s.$ at the output of the recording amplifier. Since the output magnetic flux characteristics of the heads were not specified, this level was chosen arbitrarily as the one at which a third-harmonic distortion level of approximately 1% was given at $660Hz$ on a good quality (BASF Super LH C90) ferric tape. This gives a $+3VU$ setting of $3.1V r.m.s.$, which is below the amplifier clipping level on $13V$ supply line voltage.

The output of the record amplifier is taken to the VU meter circuit through a silicon diode, but since the record

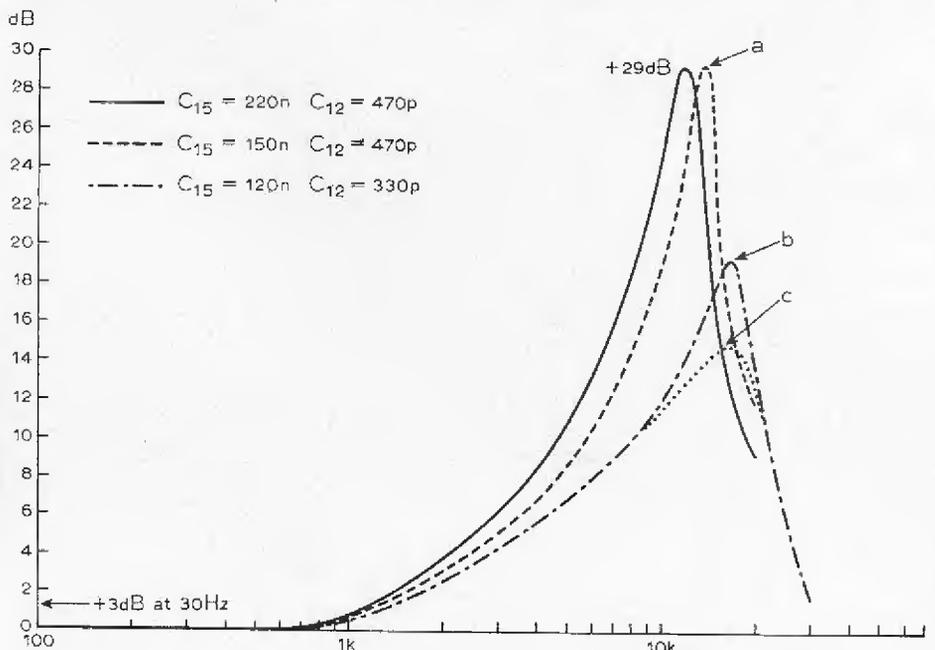
output is higher than that of the replay, an attenuator is included in this circuit to bring the two outputs to equality. The $47k\Omega$ resistor to the zero-volt line serves to provide a forward current to bias the diodes into conduction. Switching between record and replay in the VU meter circuit is automatic since only the circuit in use has an output

above the zero-volt level, the other one being disconnected from the supply line. Unwanted signal transfer through this diode feed network is of a very low order magnitude.

VU meter

This is a straightforward precision millivoltmeter of conventional type, in which the meter rectifier bridge is connected in the feedback loop of an operational amplifier as shown in Fig. 6. Although this is a more elaborate arrangement than most conventional VU meter systems, the cost of the operational amplifiers and the associated germanium diode rectifiers is small in comparison with even a modest twin

Fig. 5. Recording characteristics with variations in C_{15} and C_{12} . The peak heights are adjustable by VR_2 , (b) being the compromise adjustment and (c) the setting for optimum square wave reproduction.



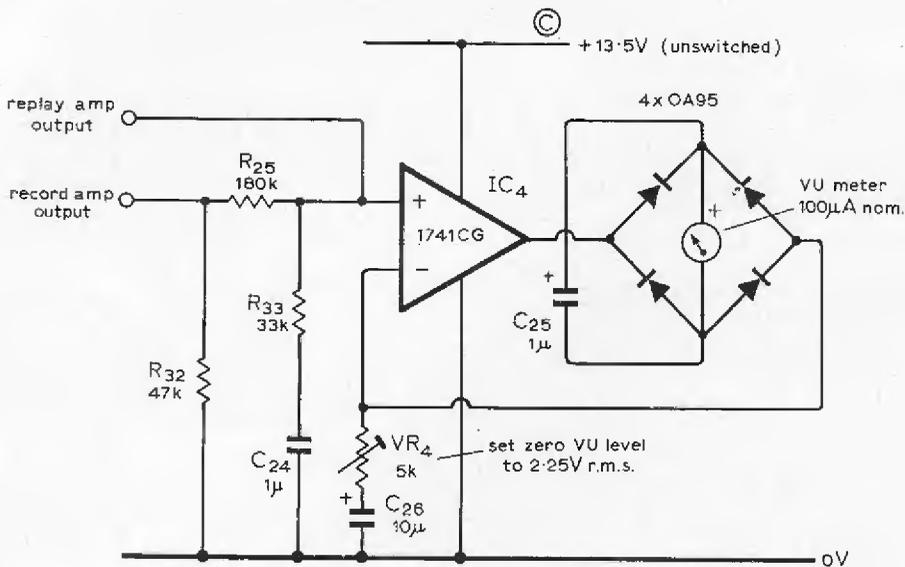


Fig. 6. VU meter circuit.

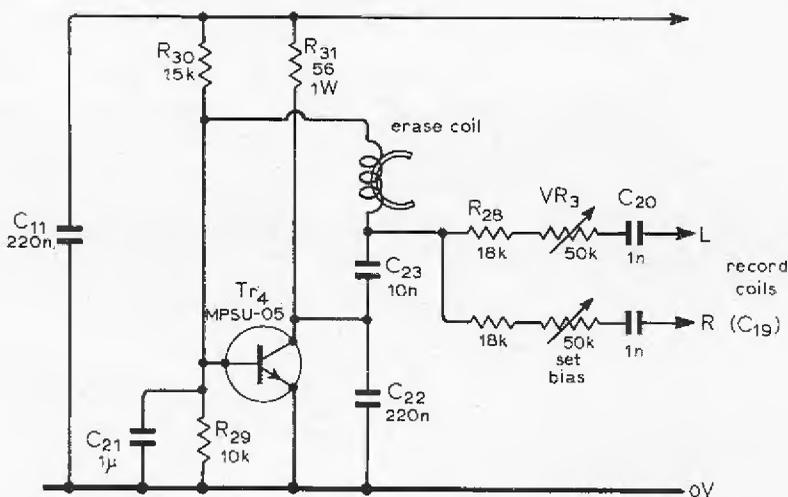


Fig. 7. Erase and bias oscillator, with continuous variation of bias level. Provision is made on the p.c.b. for the level to be switched.

VU meter, and the arrangement has much in its favour in a very linear a.c.-to-d.c. conversion, flat frequency/amplitude response, high input impedance, and short output voltage rise time due to the low output impedance of the amplifier. This latter feature is of particular value in tape recording, where the signal level meter should ideally have zero inertia so that it can follow the modulation of the signal without missing short-duration peak levels.

Bias and erase oscillator

A fairly common and irritating feature of inexpensive cassette recorders is their inability to erase fully an existing programme on a tape, when a further recording is being made on top of this. For satisfactory erasure of ferric and ferrichrome tapes, at least 20V r.m.s. should be supplied to the erase coil, and for chrome tapes a value as high as 25V may be required with typical cassette

erase heads. To obtain voltages as high as this with low-voltage lines, it is customary to use a push-pull oscillator driving a step-up transformer, but some care is necessary to avoid harmonic distortion which can impair the recorded signal quality and s/n ratio.

A simpler method, which avoids many complications, is to use the erase head as the coil in a self-oscillating circuit, and employ the Q-multiplication of the tuned circuit around the erase coil both to provide the necessary voltage swing and also to improve the purity of the waveform. The circuit shown in Fig. 7 is a modified Colpitts, and provides an output of 25-33V r.m.s. at the required erase frequency (50kHz), with supply voltages in the range 12-14 volts and with a waveform distortion of less than 1%, even when loaded with the bias circuitry. The current consumption is, however, of the order of 100mA, giving a transistor dissipation of about 0.7W. The Motorola MPS-U05 is particularly suitable, but other high-

gain, high-transition-frequency 1W devices are quite suitable since the circuit is not particularly critical of component values or types, except in so far that these may modify the operating frequency, which should be within the range 50kHz \pm 5%.

The h.f. bias waveform is also derived from the erase coil, by way of a resistor-capacitor chain, VR₃, R₂₈, C₂₀, to each record head output (VR₃ is twin-gang). Since the purity of the bias waveform at the recording head is the design requirement, it is tempting to use a value of series capacitor (C₂₀) which will be series resonant with the record coil at the bias frequency, as is fairly standard commercial practice. However, on reflection, confirmed by measurement, it is better to use a larger value of C₂₀, and take advantage of the integrating characteristics of the series network to attenuate higher order distortion components in the bias waveform, as seen at the head.

The bias voltage required across the record coil is dependent on the tape used but, as a guide, should be in the region 5-7V r.m.s., with the CT4 heads. The signal level, for reference, at this point, is only about 50mV.

Garrard Engineering Ltd now tell us that production of the CT4 mechanism is to stop in June. As mentioned in the article, however, Goldring Ltd also market a unit made under the Staar patents and this will continue to be available for some years. The type number is CRV and one difference between the two is that the CRV does not incorporate motor speed stabilization.

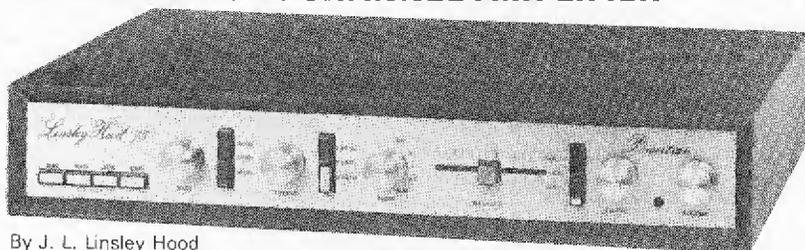
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HI-FI NEWS 75W/CHANNEL AMPLIFIER



By J. L. Linsley Hood

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 Linciac which is employed in the two most critical points of the system, namely the equalization 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 equalized and two linear, each having independently adjustable signal level. The attractive slimline unit pictured has been made practical by highly compact PCBs and a specially designed Toroidal transformer.

- | Pack | Price |
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| 1. Fibreglass printed-circuit board for power amp | £1.15 |
| 2. Set of resistors, capacitors, pre-sets for power amp | £2.15 |
| 3. Set of semiconductors for power amp | £6.50 |
| 4. Pair of 2 drilled, finned heat sinks | £1.10 |
| 5. Fibreglass printed-circuit board for pre-amp | £1.75 |
| 6. Set of low noise resistors, capacitors, pre-sets for pre-amp | £3.40 |
| 7. Set of low noise, high gain semiconductors for pre-amp | £2.40 |
| 8. Set of potentiometers (including mains switch) | £3.15 |
| 9. Set of 4 push-button switches, rotary mode switch | £4.50 |
| 10. Toroidal transformer complete with magnetic screen/housing primary: 0 117-234 V; secondaries: 33-0-33 V, 25-0-25 V | £10.95 |

- | Pack | Price |
|--|-------|
| 11. Fibreglass printed-circuit board for power supply | £0.85 |
| 12. Set of resistors, capacitors, secondary fuses, semi-conductors for power supply | £4.60 |
| 13. Set of miscellaneous parts including DIN skts, mains input skt, fuse holder, inter-connecting cable, control knobs | £5.35 |
| 14. Set of metalwork parts including silk screen printed fascia panel and all brackets, fixing parts, etc. | £7.30 |
| 15. Handbook | £0.30 |
| 16. Teak cabinet 18.3" x 12.7" x 3.1" | £9.85 |
- 2 each of packs 1-7 inclusive are required for complete stereo system. Total cost of individually purchased packs £83.75

Designed in response to demand for a tuner to complement the world-wide acclaimed Linsley Hood 75W Amplifier, this kit provides the perfect match. The Wireless World published original circuit has been developed further for inclusion into this outstanding slimline unit and features a pre-aligned front end module, excellent a.m. rejection and temperature compensated varicap tuning, which may be controlled either continuously or by push button pre-selection. Frequencies are indicated by a frequency meter and sliding LED indicators, attached to each channel selector pre-set. The PLL stereo decoder incorporates active filters for "birdy" suppression and power is supplied via a toroidal transformer and integrated regulator. For long term stability metal oxide resistors are used throughout.

- | Pack | Price |
|--|-------|
| 1. Fibreglass printed board for front end IF strip, demodulator, AFC and meter circuits | £2.15 |
| 2. Set of metal oxide resistors, thermistor, capacitors, ceramic preset for mounting on pack 1 | £4.30 |
| 3. Set of transistors, diodes, LED, integrated circuits for mounting on pack 1 | £5.25 |
| 4. Pre-aligned front end module, coil assembly, three section ceramic filter | £8.50 |
| 5. Fibreglass printed circuit board for stereo decoder | £1.10 |
| 6. Set of metal oxide resistors, capacitors, ceramic preset for decoder | £2.60 |
| 7. Set of transistors LED, integrated circuit for decoder | £2.90 |
| 8. Set of components for channel selector switch module including fibreglass printed circuit board, push-button switches, knobs, LEDs, preset adjusters etc. | £8.80 |
| 9. Function switch, 10 turn tuning potentiometer, knobs | £5.30 |
| 10. Frequency meter, meter drive components, fibreglass printed circuit board | £9.45 |

- | Pack | Price |
|--|-------|
| 11. Toroidal transformer with electrostatic screen, Primary: 0-117V-234V | £4.45 |
| 12. Set of capacitors, rectifiers, voltage regulator for power supply | £2.95 |
| 13. Set of miscellaneous parts, including sockets, fuse holder, fuses, inter-connecting wire, etc. | £1.50 |
| 14. Set of metal work parts including silk screen printed fascia panel, acrylic silk screen printed tuning indicator panel insert, internal screen, fixing parts, etc. | £7.50 |
| 15. Construction notes (free with complete kit) | £0.25 |
| 16. Teak cabinet 18.3" x 12.7" x 3.1" | £9.85 |
- One each of packs 1-16 inclusive are required for complete stereo FM tuner. Total cost of individually purchased packs £76.85

Published in Wireless World (May, June, August 1976) by Mr. Linsley-Hood, this design, although straightforward and relatively low cost nevertheless provides a very high standard of performance. To permit circuit optimization separate record and replay amplifiers are used, the latter using a discrete component front-end designed such that the noise level is below that of the tape background. Push button switches are used to provide a choice of equalization time constants, a choice of bias levels and also an option of using an additional pre-amplifier for microphone use. The mechanism used is the Goldring-Lenco CRV, a unit distinguished in its robustness and ease of operation. Speed control and automatic cassette ejection are both implemented by electronic circuitry. This unit which is powered by a toroidal transformer and uses metal oxide resistors throughout offers an excellent match for the Wireless World Tuner and the Linsley-Hood 75 Watt Amplifier.

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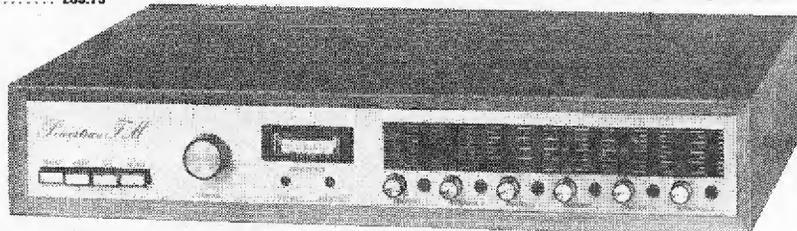
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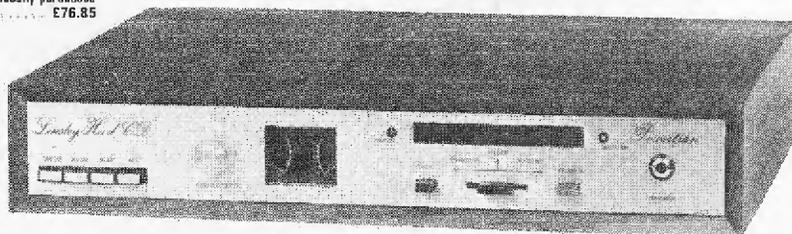
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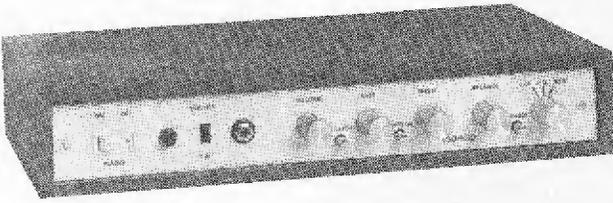
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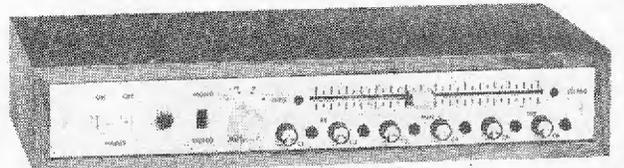
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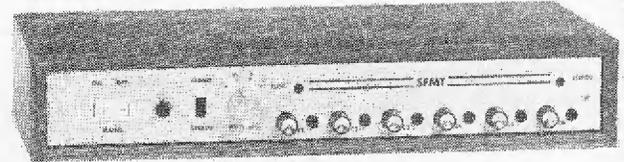
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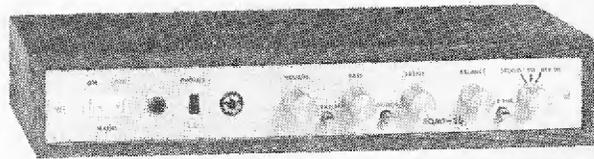
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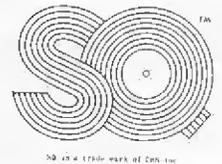


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Low-noise, low-cost cassette deck — 2

by J. L. Linsley Hood

Any convenient power supply circuit may be used for a.c. mains operation, provided that it can be set to give a stable, ripple-free output of 13.5-14V at output currents up to 250mA. A suitable design is shown in Fig. 8. The low-value resistor in the record amplifier supply line is to limit the supply line current surge through the changeover micro-switch when the large capacitor on this line is connected in parallel with the capacitor on the output of the power supply. To avoid noise originating from the pulsating current demand from the d.c. cassette-drive motor-control circuitry, the recorder supply is taken directly from the power supply reservoir capacitor through a 20 Ω , 10W resistor, with the negative return line being also directly connected to the reservoir capacitor, rather than to a chassis return.

The chassis itself is only connected to the zero-volt line at the input to the replay amplifier.

Recording and bias levels

One of the most obscure areas in the field of tape recording, in the eyes of the layman, is the interaction between tape types and biasing levels. While much

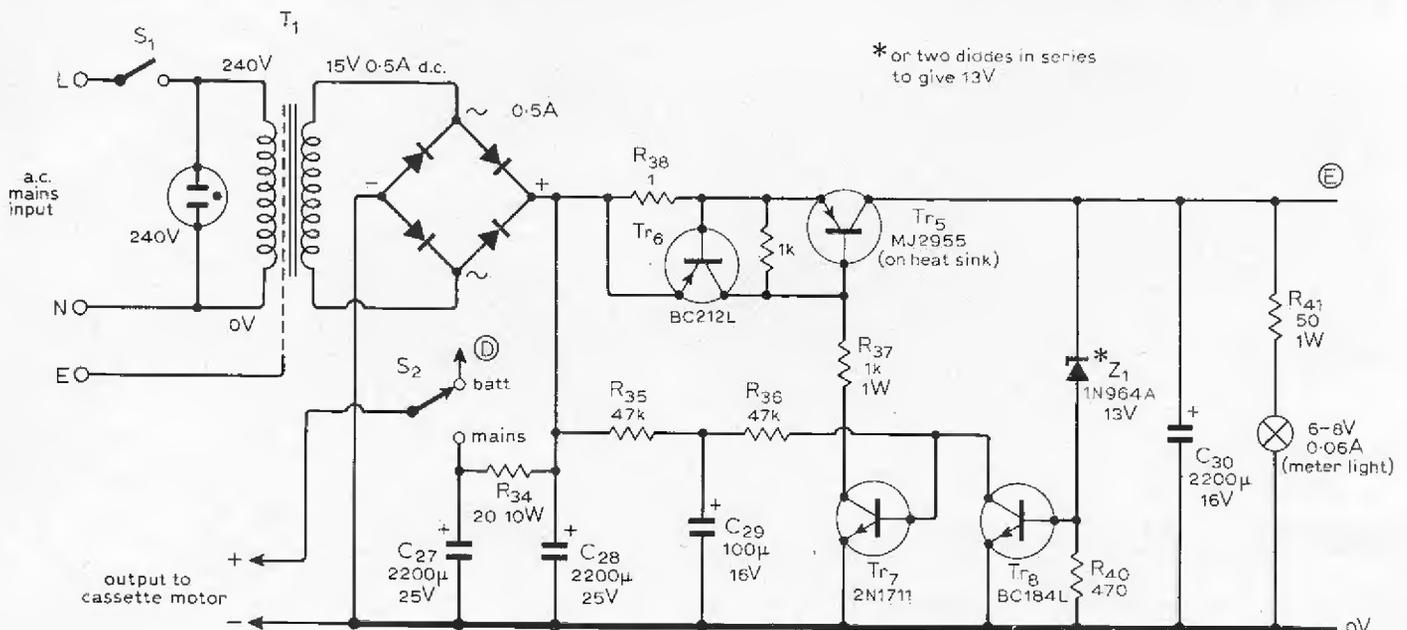
information has been published on this subject, it is often discussed in obscure terms which make the argument difficult to comprehend. Since it is possible that the construction of a cassette recorder of this type may be of interest to those with no previous experience in this medium, an attempt has been made to provide a simple introduction to this topic.

In general, it is not practicable to obtain an adequate remanent magnetic flux in a magnetic tape, for the reproduction of signal waveforms of low harmonic distortion and good output level, unless a high frequency "bias" of suitable magnitude is superimposed on the signal at the time of recording. The effect upon the various signal parameters of variations in the bias level is shown in schematic form in Fig. 9. From this it will be seen that there is not a single biasing level which is optimum for all recorded frequencies, and that the optimum level for 1kHz is in excess

of that which gives the highest output for, say, 10kHz. Also, the level which gives the lowest recorded noise level is less than that which gives the lowest t.h.d.

It is apparent from this that the setting of this parameter is one which demands some compromise, and the one which is chosen will depend upon the preferences of the user. In general, for cassette recorders, the chosen bias is that which gives the maximum output at 330Hz, or a slight excess of that optimum for 1kHz, and the reduction in output at higher frequencies is compensated by modifications to the record pre-emphasis curve. However, the required bias and compensation characteristics will be different from one make of tape to another, and from one design of record/replay head to another. In this design, the decision had been taken, partly in the interests of running costs, and partly in the interests of minimizing wear in the Permalloy heads, to optimize the design for "ferric" tapes rather than chrome types, and the Philips standard low noise C90 was taken as the reference. It was found, however, that the settings derived for this was also optimum for

Fig. 8. Power supply. The motor is fed with 3.5V by the stabilizer, but the solenoid is provided with the full 12-14V.



"super" tapes of the types exemplified by Memorex MRX₂, and BASF Super LH, although these gave an improved performance.

In general, there is very little difference in the background "bulk-erased" noise levels of most good-quality commercial cassette tapes, although there will be larger differences between the noise outputs of tapes passed through a recorder set to record at zero signal level. The greater the degree of homogeneity of the tape oxide layer, the lower the zero-recorded-level noise will be, down to a minimum which depends on the fundamental granularity of the oxide medium. The recently introduced "super" series tapes have a more uniform oxide coating, which can give a 1-2dB zero-signal background level improvement.

However, there are also improvements which have been made in the output level and harmonic distortion for a given output level, due to a more careful balance of grain shapes and sizes. The extra 2-3dB in output can lead, in total, to a small though audible improvement in overall signal-to-noise ratio (some typical results are shown in Fig. 11, and Table I.) Useful though this is, a far greater capacity for improvement in overall performance lies in the hands of the user in a careful choice of the recording level setting. Ideally, one should record at as high a level as possible, so long as few signal peaks significantly exceed the DVU level. A 1-2dB excess is unlikely to be noticed in replay, especially with good tape, provided that the duration of overrun is brief, and the difference in s/n ratio from "correct" to over-cautious choice of record levels can readily exceed the difference between a cheap tape and an expensive one.

Probably the best way of choosing recording levels is to set the mechanism to Record but with the pause button pushed in, and in this condition to experiment with the gain settings until the optimum setting is found, when the recording can be started. In the case of a live performance, assuming one has the co-operation of the performers, it is usually possible to persuade them to execute a known fortissimo and set the record levels appropriately for this.

Noise characteristics of the system

During the development of this circuit, the sources of noise in the system were investigated at some length, since, although it was envisaged that some form of external noise-reduction system would be incorporated for use during replay, it seemed advisable to try to minimize noise in the design before taking additional palliative action. The reduction of noise in the replay circuit has already been described; if the unweighted noise level at this stage, without tape, is taken as the unit, the bulk-erased-tape noise level corresponds to about one and a half or two units, and the zero-record-level noise

(with the circuit parameters arranged in accordance with normal practice) was then equivalent to some five or six units.

Obviously there was little that could be done about the noise level of the tape, as received, but it seemed that quite a lot could or should be done about the 14dB worsening of this during no-signal recording. Indeed, if it were possible to get down to the level of the original tape, the overall performance would have been beyond reproach, in

spite of the low tape speed and narrow tape width. An additional piece of evidence which seemed of interest was that the replayed noise was "whiter" than would have been expected from an original white-noise source (i.e., the tape) when replayed through the type of equalizing characteristic employed in the replay amplifier. This seemed to be a common characteristic of commercial cassette recorders, which all sounded "whiter" on the replay noise tone than

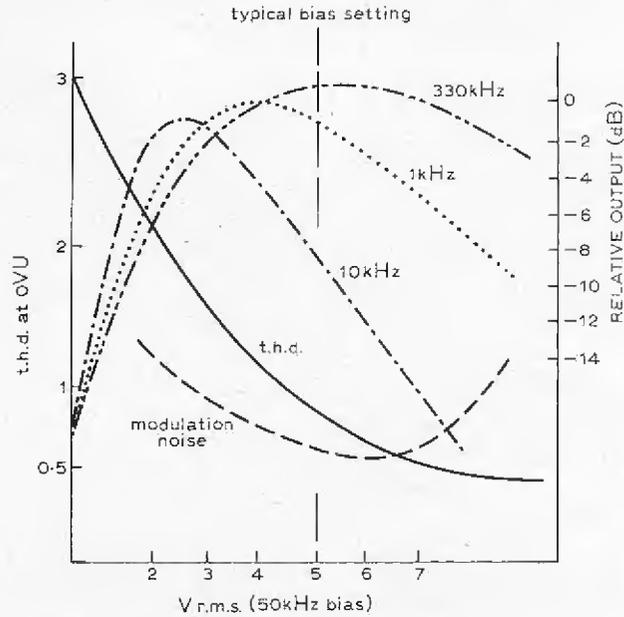


Fig. 9. The effects of changes in the h.f. bias level on output, t.h.d. and modulation noise.

Fig. 10. Record/replay frequency/amplitude curves for a variety of tapes, optimized for maximum flatness and for best square-wave reproduction

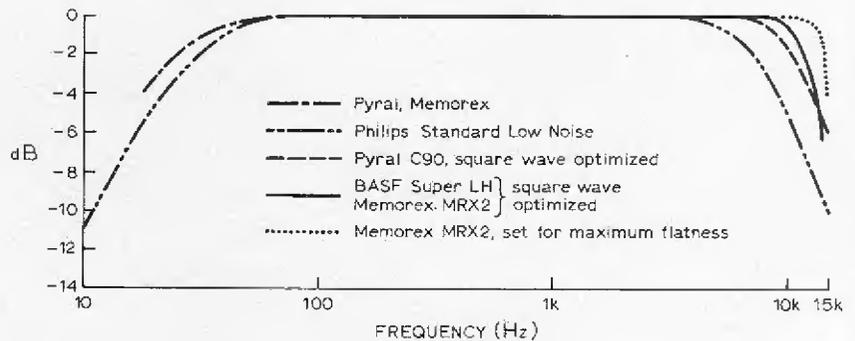
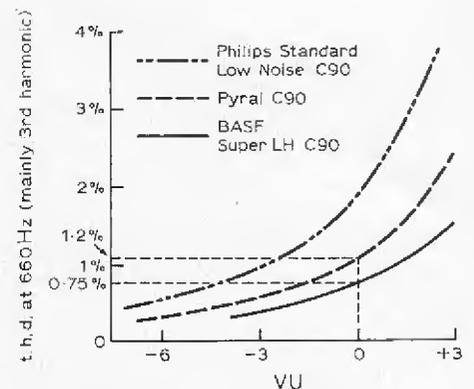


Table I.

Relative output at 660Hz (0VU recording level)	
Philips Standard C90	0VU (+0dB)
Pyral C90	+2dB
Memorex MRX ₂ C90	+2dB
BASF Super LH C90	+3.5dB

Fig. 11. Harmonic distortion (including noise) as a function of recording level for three types of tape.



0VU = 2.25V r.m.s. @ 660Hz at output of record amp

seemed reasonable to expect. It was also observed that changes in the level, frequency or shape of the bias waveform made no difference to the result. The conclusion began to grow that the problem was due to generally-distributed noise, in the source and record amplifier, being selectively amplified in the 10-15 kHz band and applied to the tape at high levels through excessive signal pre-emphasis.

Having become convinced on this point, the attempt was made to determine the optimum compromise between flatness of frequency response and signal-to-noise ratio for a given record level. At this stage it was found that optimization of a recorded 1kHz square wave to reduce the initial overshoot and ringing found when more conventional magnitudes of pre-

emphasis were applied gave also the best performance compromise on bandwidth against s/n ratio. The bias levels found during this exercise were compared to those obtained by more conventional setting-up procedures, and found to be substantially identical. However, the zero-signal-level recorded noise was found to have been reduced by about 10dB by the process of square-wave optimizing, as compared to that given by frequency response optimizing. The two curves are shown in Fig. 10, and it will be seen that the h.f. loss amounts to only 1dB at 10kHz and some 4dB at 14kHz, which can be remedied by the use of amplifier tone controls on replay with very little detriment to performance.

The final conclusion is that in general far more h.f. pre-emphasis is employed

on recording, in the interests of maximum flatness of the published response curves, than is sensible in the light of the overall performance, and that with more prudence exercised in this respect, noticeable improvements could well be made. Interestingly, programme recordings made before and after optimizing of the square-wave performance of the recorder did not show the expected small loss of higher frequencies, with the upper register seeming both cleaner and more extended than before, possibly due to the lessening of the incidence of h.f. tape or head saturation.

One final recommendation in this respect is that for one's own use, even on ferric tapes, the 70µs characteristic should be used both on record and replay. However, this is a choice which can be assessed readily by individual experiment. Certainly, in the case of the author's prototype, the use of this equalizing time-constant, in association with an optimized square-wave characteristic, has given a system in which the tape noise, on a good quality tape, is sufficiently unobtrusive to render further noise reducing circuitry unnecessary.

Bias and equalization settings

It will be apparent from Fig. 9 that adjustment of the h.f. bias level of the recorder will have the effect of altering the whole response curve, by altering the effective recorded levels of the h.f. and l.f. components relative to one another, and it may therefore appear difficult to optimize either the bias or the equalization. The suggested method is therefore as follows:

- Set VR₂ so that the response curve on record is as curve A on Fig. 5., (approx. 85Ω) when measured at the output of IC₃.
- Set the bias level so that there is approximately a 1-2dB drop in output at 1kHz.
- Record a square wave at 660Hz, and make small adjustments to VR₂ until the cleanest leading edge is obtained on the replayed square wave, with only a small single overshoot.
- Finally, leaving VR₂ set at the chosen value, record a 660Hz square-wave at various bias levels and adopt the one which gives the best overall square-wave shape for the tape which it is desired to employ.

The technique of square-wave optimization is well known in the audio field as a means for setting tone-controls and filters to optimum flatness, because of the facility which it offers for a simultaneous examination of a wide range of frequencies. On exactly the same score it would appear to be an excellent method of optimizing bias levels.

While it is hoped that the performance given by the prototype will prove to be fairly typical of the results given by other models built to this design, it is appreciated that in a system in which

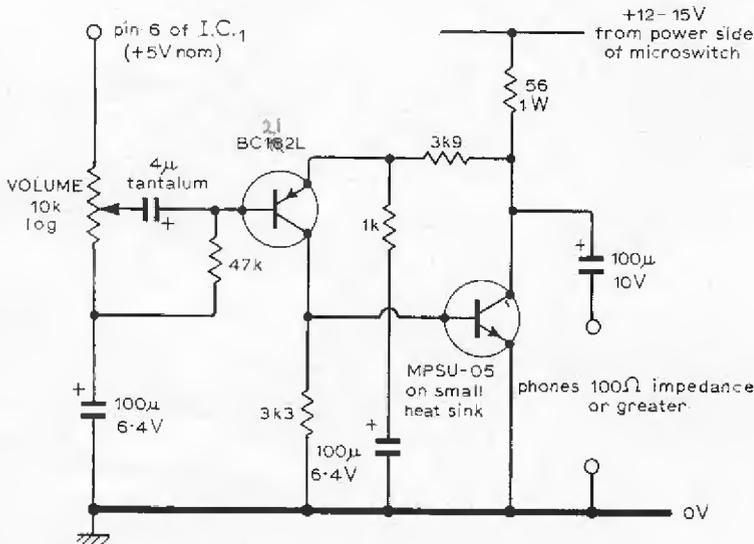
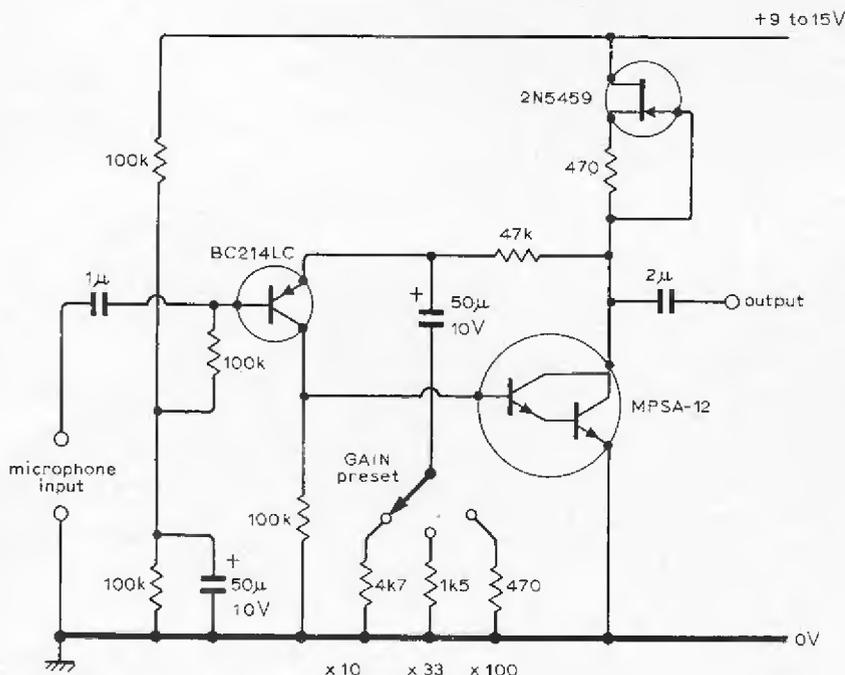


Fig. 12. Class A headphone amplifier, with a gain of 5.

Fig. 13. Low-noise microphone pre-amplifier.



not only will components vary in types and tolerances, but also the tape transport mechanism and heads (which may be changed during the manufacturer's production run for reasons of commercial availability) may differ from those used by the author, the scope for variability is considerable. Also, from personal experience, and measurements, there is a considerable variation in performance from one tape type to another, although the consistency of performance of the better-grade tapes from the better manufacturers appears to be fairly good. On the credit side may be set the fact that one does not have quality control problems in a unit that is one-off, and that one can optimize one's channel balance and h.f. performance for the tapes one prefers and the heads one happens to possess.

Headphone operation

Both the output level and the drive capability of the final 741s of the replay amplifier are adequate to give a satisfactory signal strength and quality into the 2kΩ load impedance of the author's headphones (Sennheiser HD414), so, for simplicity, this was the course adopted. However, for those with lower-impedance or less-sensitive headphones, a suitable circuit is shown in Fig. 12. This operates in class A, and is suitable for load impedances down to 100 ohms.

Direct microphone recording

The sensitivity of the record input is only intended to be sufficient for recording from an existing audio amplifier or radio tuner capable of delivering some 50-100mV output at a fairly low impedance, and it would not be suitable for microphone inputs. For this purpose a pre-amplifier can be used, of which a suitable circuit is shown in Fig. 13. Three preset gain positions are given, of 10, 33 and 100×, which should cope with the bulk of microphones likely to be found in practice. A typical gain suitable for a low-output cardioid capacitor electret microphone is of the order of 33, for a normal recording level at half-gain setting on the recorder.

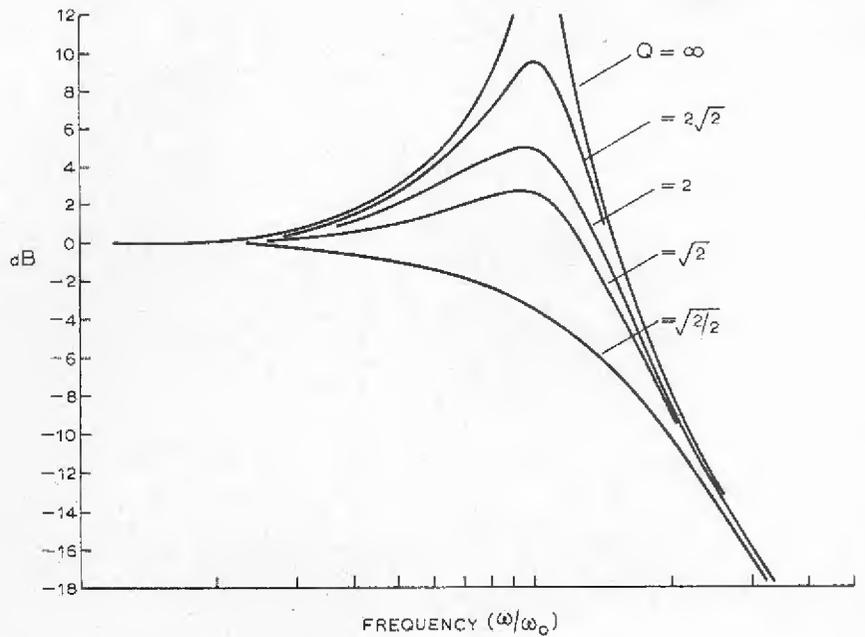


Fig. 15. Frequency/gain response of filter in Fig. 14.

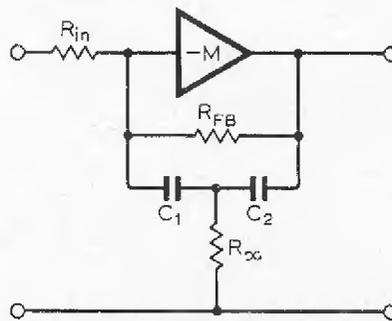


Fig. 14. Basic second-order active low-pass filter (as Reference 1).

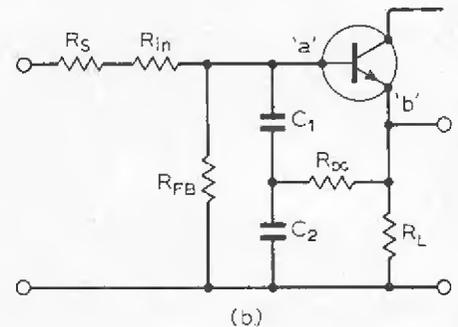
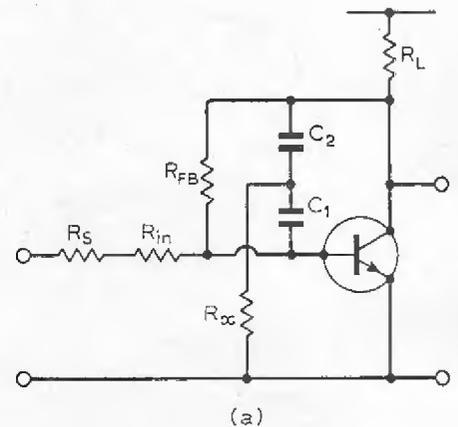


Fig. 16. Common-emitter (a) and common-collector (b) arrangements of basic circuit shown in Fig. 14.

since it was desired, in the interests of simplicity, to avoid the use of inductors, and it was also required to avoid possible trouble due to the intrusion of 38kHz signals from multiplex stereo decoders, the decision was made to use the gain/frequency characteristics of an under-damped second-order active low-pass filter, such as that shown in Fig. 14.

This is one of the classic forms of active element, and was analysed by Girling and Good¹ in the first part of their survey of active filters in *Wireless World*. It has a gain/frequency response of the type shown in equation (1) and

illustrated in Fig. 15 for various values of Q (1/α)

$$\frac{V_{out}}{V_{in}} = \frac{1 + j\alpha \frac{\omega}{\omega_c}}{1 + j\alpha \frac{\omega}{\omega_c} - \left(\frac{\omega}{\omega_c}\right)^2} \quad (1)$$

If this is redrawn, ignoring biasing, in the form in which the amplifying element is a single, common-emitter-connected transistor, as shown in Fig. 16(a), it will clearly be seen that this can be rearranged into the common-collector form of Fig. 16(b) without significant alteration of the expression for the

Appendix

Derivation of record equalisation characteristics.

The generation of a recording pre-emphasis characteristic of the general form shown in Fig. 5 is normally done by incorporating a damped LC parallel resonant circuit in the feedback loop of an inverting amplifier stage. However,

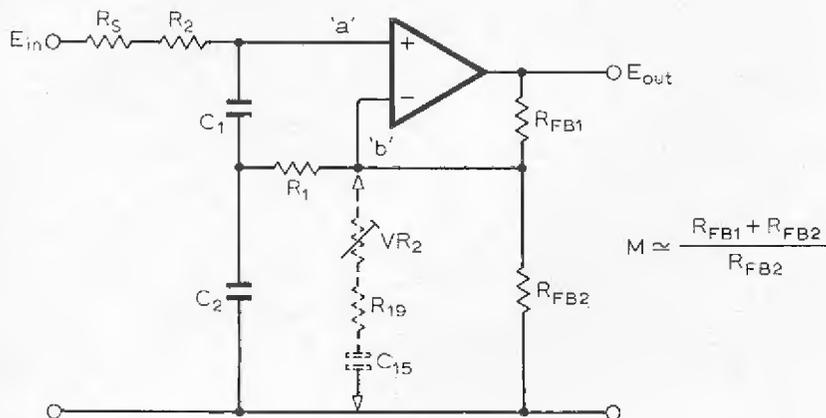


Fig. 17. Use of the active filter in a stage possessing gain, as in Fig. 4.

frequency response. (It was pointed out by Girling and Good that a similar rearrangement leads to the evolution of the filter circuit due to Sallen and Key² from the type based on a feedback loop containing an integrator and a lag, described earlier by Baxandall³.)

Filters of the type shown in Fig. 14, and 16(b) were used in the authors "Modular preamplifier" design (*Wireless World* July 1969) and the subsequent postscript to this and the earlier class A amplifier, published in December 1970. The performance of the circuit shown in Fig. 16(b) referred to as a bootstrap or "H" filter, was analysed by the author in another place⁴ and, in a

different context, by Hemingway⁵.

The operation of the filter circuit of Fig. 16(b), like that of the Sallen and Key configuration, can be realised by any active element in which there is approximately unity gain between points "a" and "b". This allows the use, for example, of a voltage-follower operational amplifier as the active circuit element, or even a non-inverting operational amplifier having more than unity gain, as shown in Fig. 17, provided that an adequate fed-back gain margin is available at the frequency of interest. The gain and frequency of maximum response of such a circuit is given by equations (3) and (4) where M is the

normal gain of the stage with feedback.

$$f_c \approx \frac{1}{2\pi \sqrt{[C_3 C_4 R_1 (R_2 + R_S)]}} \quad (3)$$

$$\left| \frac{E_{out}}{E_{in}} \right| = M \left(1 + \frac{R_2 + R_S}{R_1} \frac{C_1 C_2}{(C_1 + C_2)^2} \right) \quad (4)$$

While, basically, the Q of the system is determined by the ratios of $R_1 : R_2 + R_S$ in a stage with a shunt network, such as that of VR_2 , R_{19} and C_{15} in Fig. 4, the gain will change in the frequency region of the filter attenuation band. The associated phase shift due to this network also modifies the Q and allows adjustment of this by VR_2 , which is a feature of some practical convenience.

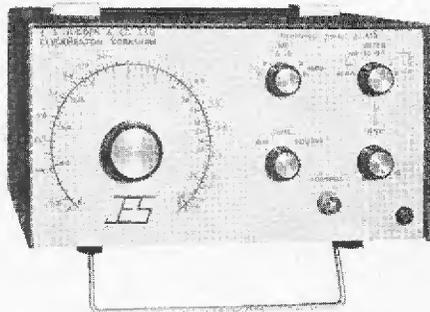
References

1. Girling, F. E. J., and Good, E. F., *Wireless World*, vol. 75, No. 1406, pp 348-352.
2. Sallen, R. P., and Key, E. L., *Trans. IRE.*, vol. 2, No. 1, pp. 74-85.
3. Baxandall, P. J., *Wireless World*, vol. 61., No. 1, pp. 8-14.
4. Linsley Hood, J. L., *Hi-Fi News and Record Review*, vol. 18, No. 2, pp. 293-294.
5. Hemingway, T. K., *Electronic Designers Handbook*, pp. 283-288.

Component supplies

Goldring-Lenco CRV cassette mechanisms can be obtained from Goldring Ltd, 10 Bayford St, Hackney, London E8 3SE, or Hart Electronics, Penylan Mill, Oswestry, Salop. VU meters are available from J. E. T. Electronics, 90a Mawney Road, Romford, Essex.

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Motor control and further notes

by J. L. Linsley Hood

For further updates see W.W. Feb 1978 pages 35-40.

In response to one or two queries, the following notes are offered. Several cassette decks have now been completed, using alternative designs of printed board, and have proved very successful.

Motor control

Circuitry for the control of the drive motor and solenoid is shown in Fig. 20. It is required to supply or withhold current from the cassette-retaining solenoid and to supply a constant drive to the motor in the presence of supply variations.

Solenoid control. Tr_3 normally conducts and energizes the solenoid. As the motor turns, the pulse-generating switch in the mechanism (yellow and green leads in the Goldring deck) keeps Tr_1 conducting, which cuts off Tr_2 and allows current to flow through the solenoid and Tr_3 . When the motor stops, so does the switch: Tr_1 ceases to conduct and, after 3 seconds (C_2R_3) Tr_2 conducts, cutting off Tr_3 and de-energizing the solenoid. The cassette is thereby released. If the "pause" contacts are made, the motor stops, but the cassette is retained in position.

Speed control. The motor is supplied with constant current via Tr_5 . Tr_4 is conducting. Back e.m.f. developed by

the motor beginning to turn is applied to Tr_4 emitter, reducing its forward bias. This reduces the current into Tr_5 base and tends to reduce the motor speed — the effect is to stabilize the motor. Tr_5 behaves as a constant-current source by virtue of the feedback from its collector to Tr_4 base.

Record input impedance

There are, unfortunately, two conventions on the impedance levels employed

for signal handling prior to tape recording. Of these, the older, and I think the more sensible, is the "600 ohms, 0 VU" (+0 to -60dB, ref. 0.77 V r.m.s.), system which seems to be used by many recording studios, and gives a signal level which can be handled comfortably without problems of degradation due to noise. The other, and the one which is being used increasingly in commercial amplifier "recorder" outputs, is the DIN standard, which implies basically a constant-current source, developing a nominal 1mV r.m.s. for each 1k Ω of recorder input impedance. Predictably, this leads to a degradation of signal

Fig. 19. Buffer amplifier to match a DIN source to the recording amplifier.

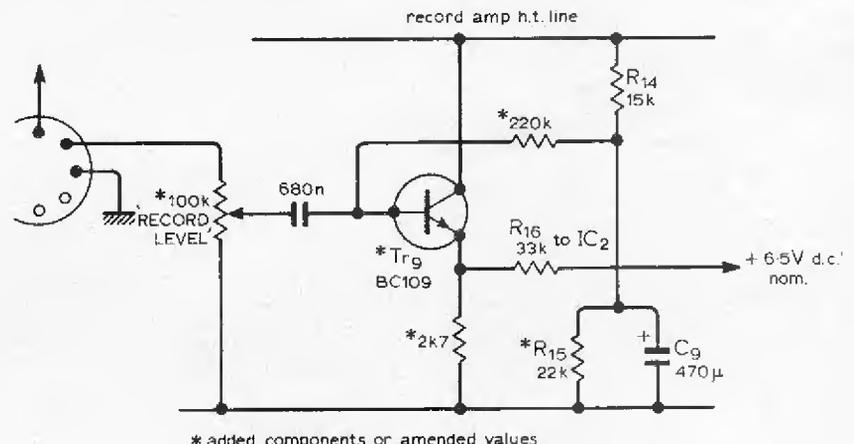
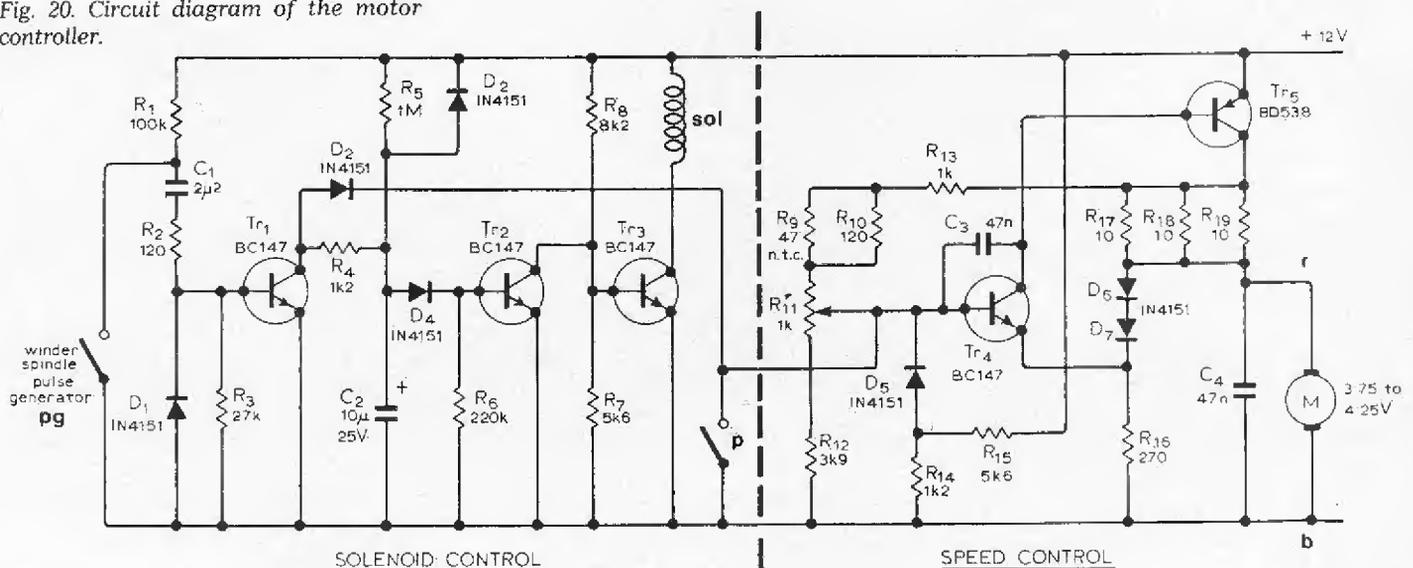


Fig. 20. Circuit diagram of the motor controller.



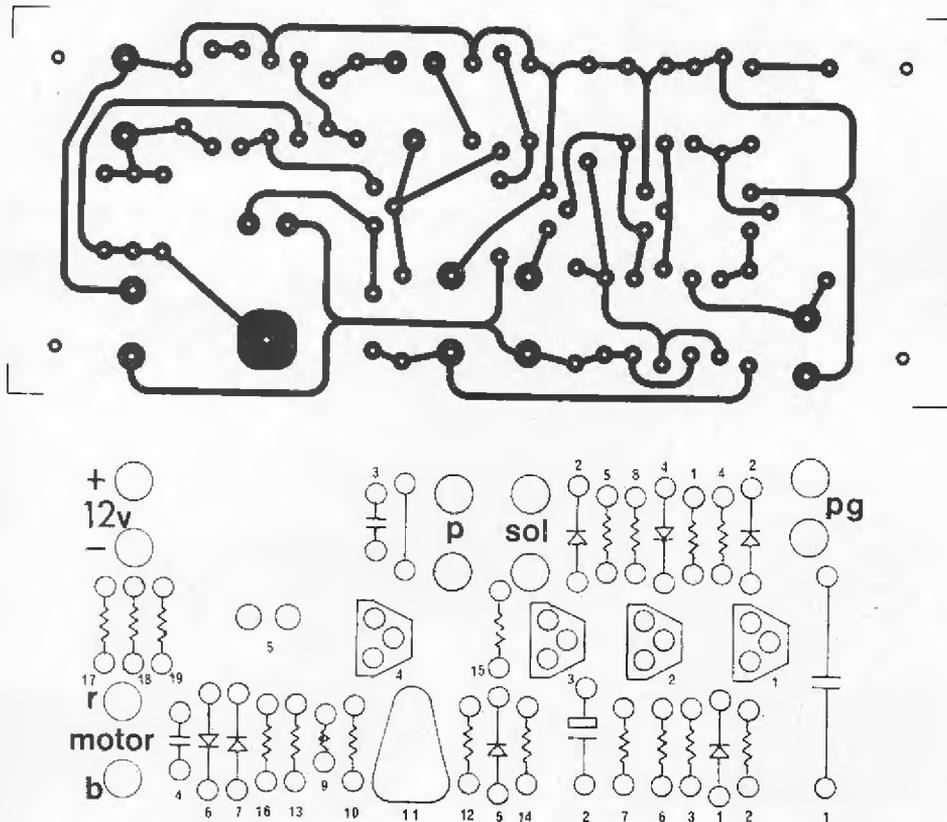


Fig. 21. A suggested, actual-size layout for the controller. The layout and modifications to the speed control circuit are due to Mr A. H. Milligan.

cable simply to increase the input record level potentiometer to 50k Ω or 100k Ω since the source impedance of IC₂ influences the Q of the h.f. pre-emphasis system (see Appendix). While the effect of the existing 10k Ω potentiometer, when driven from a fairly low source impedance, is negligible, this would not be true for a higher value DIN input.

If, therefore, this is to be used with a commercial unit having this convention (as distinct from a home-constructed item, in which it is probably most convenient to take the recorder feed at the pre-amp output, in parallel with the power amplifier input), it is recommended that a small buffer circuit should be attached to the output of the record level potentiometer, as shown in Fig. 19.

Replay h.f. stability

Proximity of output and input leads may cause instability in the replay amplifier. If this cannot be avoided due to layout constraints, a small capacitor (330pF or so) can be connected across the replay output relay terminals (RL₁/1, RL₁/2) – across the replay coil output in the replay position – without any adverse effect on the h.f. performance.

quality due to thermal noise unless fairly high value input impedance circuits are employed.

The convention for which the recorder described above was designed was the 600-ohm source impedance one

although, taken in general terms, this means any range of source impedances in the range zero to a few kilohms, and the system as it stands would probably have inadequate gain if operated from a DIN source. It is, however, not practi-

Letters to the Editor

LOW NOISE CASSETTE DECK

We should like to take the eminent Mr J. Linsley Hood to task for advising the use of the 70 μ s equalisation characteristic for use with normal low-noise ferric cassettes. This is most misleading because one of the most serious problems with these cassettes is their lack of response to high-level, high-frequency signals; the 120 μ s post-emphasis was adopted to try to alleviate this. Even this results in a fully-saturated recorded level of about 10dB below Dolby level at 10kHz. Adopting the 70 μ s equalisation characteristic reduces the h.f. overload figure by almost another 5dB which makes an already bad situation intolerable. This would produce severe h.f. intermodulation distortion when recording typical musical material at "normal" mid-band modulation levels.

The reason that the 70 μ s equalisation is adopted for chrome cassettes is simply that they are much less susceptible to h.f. overload because of the smaller particle size and higher coercivity of the oxide formulation.

We would, however, endorse Mr

Linsley Hood's suggestions for optimising bias and equalization settings. In our opinion too many manufacturers align their machines to attain the ultimate in frequency response (or "specmanship") to the detriment of other aspects of the reproduced quality. One notable exception to this is the British manufacturer NEAL, who quite deliberately align their machines to be -2dB down at 12kHz on ferric tapes. This compromise produces a similar result to that obtained using Mr Linsley Hood's square wave technique.

C. J. Evans, J. Dawson,
A&R, Cambridge.

The author replies:

The original Philips recommendation for the equalisation of the "Musicassette" was for time constants of 1590 and 120 μ s, of which the first was to compensate for anticipated inadequacies in the l.f. response and hum pick-up problems with the circuitry, and the second was to remedy the known shortcomings of recording head and tape characteristics. Improvements in system design have led to the universal adoption of the 3180 μ s low-frequency equalisation time constant, to bring the

cassette system into line with other reel to reel, recording systems, and the advent of chromium dioxide cassette tapes has prompted the adoption of a 70 μ s h.f. time-constant in order to secure some of the advantages which these improved tape types can offer.

As a consequence of this, most modern cassette recorders will offer a choice of h.f. time constants, whose use is at the discretion of the user, when he is making recordings for himself.

However, the design of recording heads and cassette tape materials has not stood still in the intervening years, and it is my belief that there are many ferric tapes which will give an improved signal to noise ratio, without any significant penalty in terms of h.f. overload on typical programme material when used with the "chrome" (70 μ s) equalising time constant, and it was this belief, based on a quite substantial number of tests, which led me to make the recommendation to which Mr Evans and Mr Dawson object.

A shrewd friend once observed to me that rules were made for the guidance of the wise, and the blind obedience of fools, so, in this context I would urge, even in the light of hind-sight, which is said to be an exact science, that users try out the available options, and judge the issue for themselves.

J. L. Linsley Hood

Wireless World Dolby noise reducer

1—An introduction to the Dolby noise reduction system

by Geoffrey Shorter

This noise-reducer design is intended mainly for hiss reduction in magnetic-tape recording machines. The unit can be switched to decode commercially available Dolby B-encoded cassette tapes, Dolby B-encoded f.m. radio transmissions (as in the USA), or to encode blank tapes from any source. As an alternative, it can be used in trading some of the noise improvement for reduced distortion at peak recorded levels. The *Wireless World* processor can be aligned without any additional test instruments, the circuit board being arranged to provide the necessary alignment and calibration tones. This article gives background to the B system and to the functioning of the noise reducer and subsequent articles describe construction, alignment and calibration of the unit.

In audio systems dynamic range can be defined as the ratio of the largest to the smallest programme signal. Dynamic range is typically limited at the high-level end by tape saturation or amplifier signal handling problems: there is usually a fairly well-defined level beyond which compression occurs and distortion rises at a rapid rate. At the other extreme there is a limit on the lowest signal that can be handled, set typically by the noise level of electronic circuits, tape noise, surface noise on discs, or granularity on optical soundtracks.

In concerts, dynamic range can be as high as 90 to 100dB, but once such programme material has been recorded, dynamic range is reduced to 60 or 70dB. (When broadcast the range can be as low as 20 to 40dB.) In this situation there are three options—lose that part of the programme below noise level, distort the peaks, or distort the range by compression either manually or automatically. None of these options is altogether acceptable in itself, all distort the original in some way. What is needed is a way of getting round this limitation of dynamic range without the distortion of overmodulation, without losing programme in noise and without distortion of range. Before discussing various techniques that have been proposed and tried, we will be more specific about what is required.

As well as not introducing any perceptible non-linear or dynamic range distortion of both steady-state and transient signals, any proposed technique for high quality use should not perceptibly alter the signal in respect of frequency response and transient response. Any signal processors must be able to operate to the normal constraints of audio

channels, i.e. operation should not depend on freedom from phase and amplitude versus frequency errors or changes, nor on a linear phase-frequency response; channel overload characteristics should not be worsened. In addition to compatibility with transmission channels, there must be compatibility between processors to the extent that recordings can be interchanged. In reducing perceptibility of noise, there should be no noticeable noise modulation effect and ideally all noises should be reduced by a similar amount, otherwise reducing one kind might unmask another.

Noise-reducing techniques

“Static” methods. The most well-established methods of avoiding the constraints imposed by high noise levels are “static” ones. Examples are the high-frequency pre-emphasis, and subsequent de-emphasis, applied to f.m. broadcasts and gramophone records and the low-frequency pre-emphasis used in tapes. They are static because the amount of emphasis given is fixed and does not take account of the signal in any way. At some frequencies, there is thus an intrusion into the possible range of levels that signals can occupy which may mean that some lower than normal limit must be placed on the programme level.

Single-ended methods. An alternative approach is the dynamic one of altering the level of a signal by an amount that depends on the signal level, at either the sending/recording end or at the listening end. In examining such dynamic techniques it is expedient to look at the possibilities from a steady-state signal level point of view, with the thinking that frequency and time-dependent variations

can be seen as special categories within a level classification. In practice, however, the success of each kind will undoubtedly depend on how well complicated time-varying multi-frequency signal patterns are responded to by the processing circuitry; and to whatever psychoacoustic, or perceptual, effects such as auditory masking, can be discovered and made use of.

The simplest kind of device, within our terms of reference, is the low-level noise gate, depicted graphically in Fig. 1(a), which eliminates signals below a certain threshold level. More useful is a stepped noise gate, where signals and noise below a certain threshold are attenuated by a finite amount rather than an infinite amount—Fig. 1(b). There are a host of variants on this theme, Fig. 1(d) showing another possibility.

A number of commercially-available expanders have used the general approach of Fig. 1(b), including H. H. Scott's “dynamic noise suppressor” and R. Burwen's “dynamic noise filter”, operating only at low and high frequencies and with a passband that varies according to signal level. The Philips “dynamic noise limiter” is another example, though its operation is restricted to high frequencies. With these devices, the bandwidth restriction at low signal levels must inevitably cause some loss of programme. Further, any reduction of noise level that can be achieved is likely to be modulated by intermittent mid-frequency signal components, giving rise to what is called breathing. Because they are “single-ended” these techniques must result in a distortion of dynamic range. Thus you can either have the original dynamic range plus non-reduced noise, or a distorted dynamic range and loss of some

low-level information with a reduced noise level—but not both at the same time.

Besides altering the level of low-amplitude signals, a similar expansion can be achieved by expanding high-amplitude signals, Fig. 2(c), but as well as exhibiting the two major disadvantages already mentioned, this would suffer a third. By having a variable-gain element operating at a high level there are obviously greater risks of generating intrusive unwanted signals as a result of overshooting, high non-linear distortion and a high circuit noise level.

Dynamic processing is often carried out prior to recording or transmission. The low-level compression characteristics of Figs. 1(c) and (e) and the high-level characteristic of Figs. 2(a) and (b) both enable average signal level to be increased relative to the noise level. But in themselves they suffer from the same disadvantage as do the expanders. Clearly, single-ended methods are inappropriate to normal high quality reproducing systems.

Complementary methods. The only way of avoiding the difficulty of alteration to dynamic range is by the complementary method—the dynamic equivalent of static “equalization”. In complementary systems, signal processing before transmission and recording, normally compression, is followed by an equal degree of complementary processing, normally expansion, prior to audition so that the original dynamic range is restored. Noise added by the medium after compression is reduced by the degree of expansion used. In the expander of Fig. 1(b) the complementary compressor characteristic would be (c) and the complement of (d) would be (e). Likewise, the transfer characteristics of Figs. 2(b) and (c) form another compander system.

Fig. 1. Low-level noise gate (a) simply loses both noise and signal below a certain threshold level. Finite attenuation of low-level signals is achieved with the expansion transfer characteristics of (b) and (d). Such “single-ended” expanders reduce noise at the expense of distorting dynamic range. Compressors at the signal source end can raise low level signal above noise levels, but similarly distort range (c) and (e).

Fig. 2. High-level limiter and compressors (a) and (b) and expanders (c) suffer an additional disadvantage because of processing at a level where distortions would be more obvious.

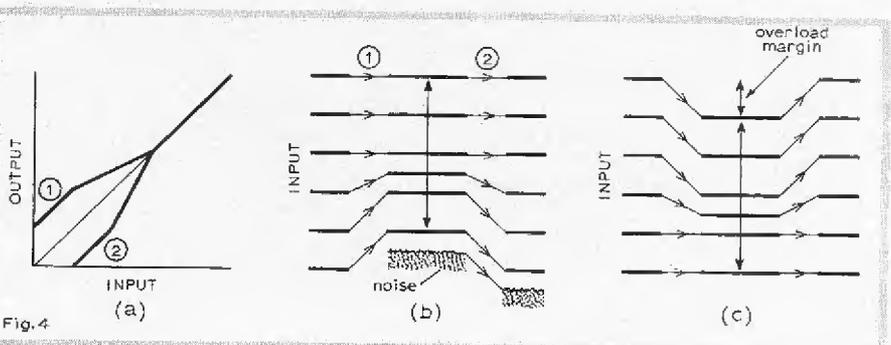
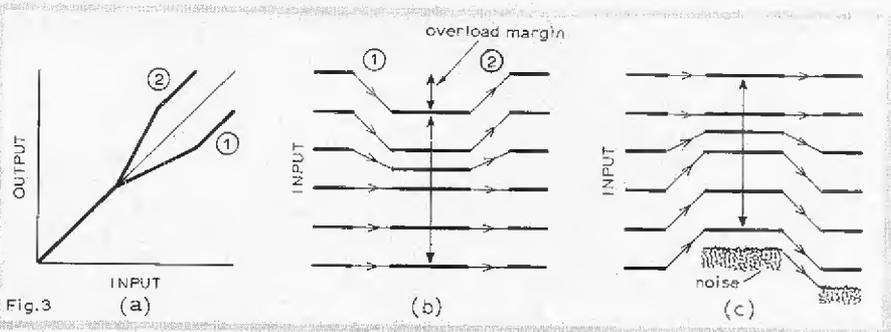
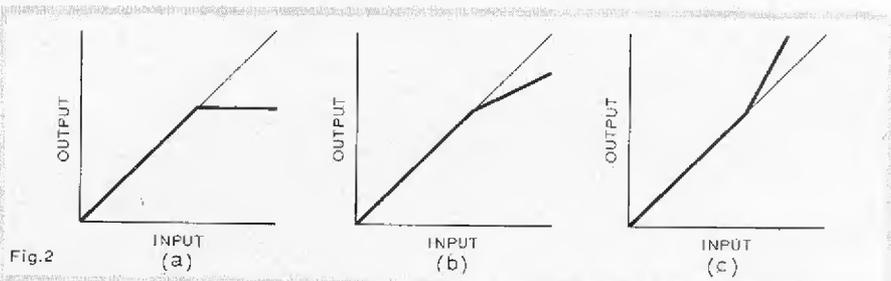
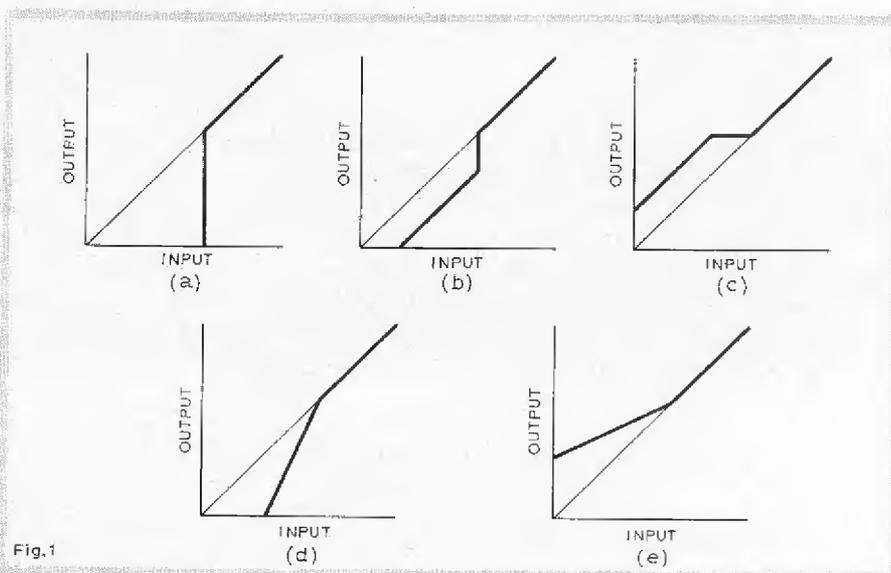
Fig. 3. Complementary high-level system (a) is able to reproduce original dynamic range while either reducing maximum level to give more overload margin (b), reducing noise (c), or giving a combination of both.

Fig. 4. Low-level complementary system (a) has the advantage that any distortion products are at a low level where they are less likely to be audible.

Another kind of diagram makes it easier to visualize what happens so far as levels are concerned. Fig. 3(a) is a typified high-level compander characteristic, showing both the compression and expansion curves. Its equivalent level diagram of Fig. 3(b) shows the reduced dynamic range (indicated by arrows) where the maximum level to be handled by the interposing medium is assumed to be the same—the region marked “overload margin” giving an increased margin against overload and thus lower distortion. Fig. 3(a) shows the same reduced dynamic range

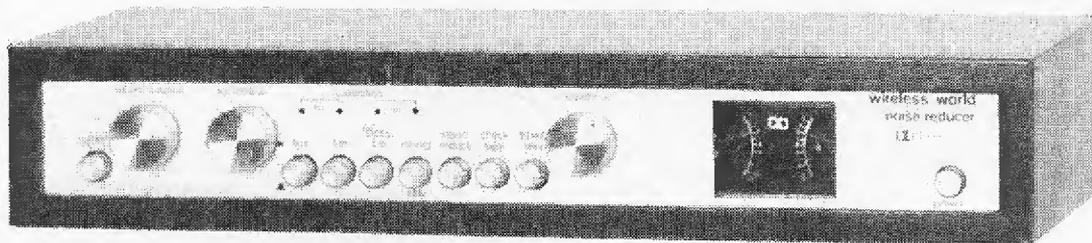
produced by the characteristic of Fig. 3(a), but with the intermediate gain shifted so that the low signal levels can be increased in relation to the noise level.

Fig. 4(a) shows low-level compander characteristics, with the level diagram of Fig. 4(b) illustrating the use of the compressed dynamic range to bring up the low-level signals relative to the noise. Fig. 4(c) shows how, by reducing the levels by a constant amount, increased overload margin can be obtained. (Notice the similarity between Figs. 3(b) and 4(c) and between Figs. 3(c) and 4(b), the





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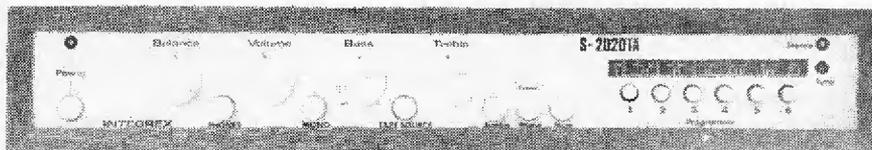
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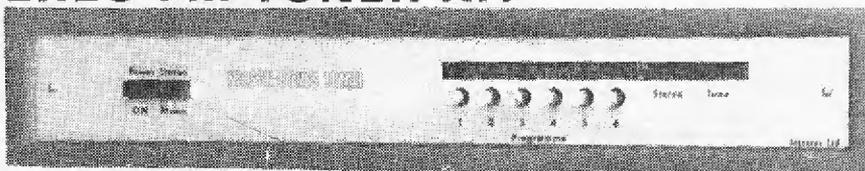
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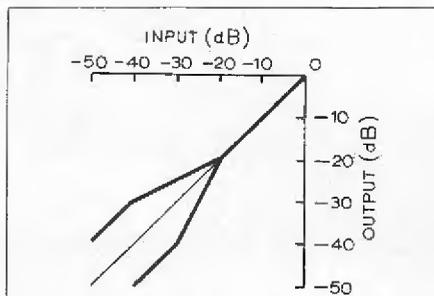
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difference being the siting of the region of "linear" operation at either a high level or a low level. Despite the immediate visual contrast between Figs. 3(a) and 4(a) there is clearly a close resemblance between curves 1.)

In practice the characteristic curves do not have the discontinuities shown, corners being rounded to prevent objectionable noise modulation. The curves should be capable of easy realization, be readily reproducible and the two complementary curves must be matched to within the required tolerance.

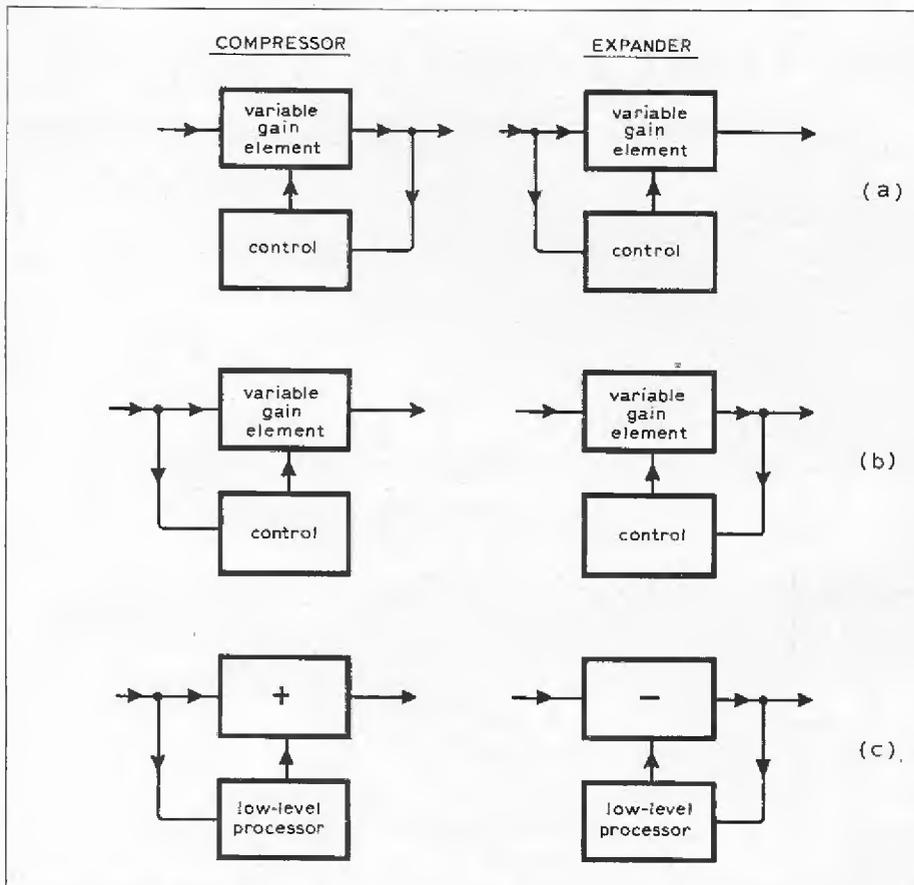
Two recently-introduced studio companders use the general approach of Fig. 2(b) and (c), but with a threshold that is much lower than indicated. The dbx Inc. compander uses a square-law curve above a certain threshold (-60dBm), which in logarithmic terms is a 2:1 compression ratio. The Burwen "noise eliminator" uses a cubic law (logarithmically, a 3:1 compression ratio) above a certain threshold. (A fixed h.f. pre-emphasis and a level-independent bandwidth are also features of these systems.)

In general, such high level companding techniques suffer from a number of drawbacks: poor tracking between the two processors, high sensitivity to errors in gain in-between processors, overshooting and a risk of overmodulation, both of which could lead to compression in the transmission medium that would go uncorrected on expansion, noise modulation by signals, modulation-product formation as a result of rapid gain changes, all of which are undesirable in a high quality link. High level companders can be very useful however in telephone circuits for example and the Post Office's Lincom-



In most electronic signal processing systems there is usually some maximum level beyond which the signal must not be allowed to go and to which levels are frequently referred. Transfer characteristics are therefore usually given in the quadrant shown in which the point of reference is made to be some arbitrary maximum level, rather than the zero signal level of Figs. 1 to 4. (A zero at the axes intersection would represent 0dB and not an origin as in cartesian coordinates.) In practice such curves are not discontinuous but are smoothly connected to prevent unwanted modulation and to permit easy realization and matching.

Fig. 5. Conventional companders use the equivalent complementary systems of (a) or (b) whereas the Dolby system (A and B) uses an additive to method (c) enabling processing circuitry to be separated from the main signal path.



pex scheme is an example of a compander in which dynamic range is reduced to zero. (Subsequent expansion would not be possible were it not for the fact that information on signal amplitudes is contained in separate pilot or control channel.)

The low-level method (Fig. 4) has a high tolerance of channel gain errors, produces modulation distortion at low signal levels rather than high levels, and there is less risk of overloading the medium. It seems a good idea anyway because one might expect the ear to be less sensitive to low-amplitude effects than to the same effects at high level. This then is the basic companding technique chosen for the Dolby system.

Dolby low-level compander

In conventional companding systems there are two equivalent ways of achieving compression and expansion. One is to derive a control signal, after subjecting the input signal to a variable-gain element (compressor); expansion or "decoding" would then be achieved by the converse process—the control signal being derived prior to a variable-gain element (expander), Fig. 5(a). The equivalent, alternative, way is to derive the control in the compressor part before the variable-gain element and to subsequently expand by using a control obtained after the variable-gain device, Fig. 5(b). (The first-mentioned method is used in the dbx and Burwen high-level companders.)

The Dolby technique makes use of a different approach—with an important difference; compression is achieved by deriving a special low-level signal that is added to the main signal, and expansion is obtained by subtracting a low-level signal from the main one, Fig. 5(c). (Within the low-level processor block, compression is achieved with method (a).)

Of course, the required compander characteristics could have been derived in the normal way, i.e. by direct action of a compressing circuit on the main signal path Figs. 5(a) and (b); but in the low-level approach the whole range need not be subjected to processing. It is obviously in the interests of quality that low-level signals be processed separately, leaving the main signal to a linear path whose quality is not restricted by that of the variable-gain path.

Tracking at high levels becomes easier using this low-level approach, and a tracking error due to channel gain variation would occur at an unobtrusively low level. Additionally with this technique, it is found that sufficiently accurate tracking can be maintained using a control derived from peak and average signal values. Thus the elaboration of an r.m.s.-derived control, which would strictly be necessary for channels having a non-linear phase-frequency response, is avoided.

Notice that in the subtractive part of Fig. 5(c), a negative feedback loop is effectively formed in the low-level "contribution" to the main path. Advantage of

this is taken in the Dolby system in that an identical network to that used to produce the additive low-level signal at the encoder, can be used in forming the subtractive component at the decoder, merely by inserting the network in the negative feedback loop of a main path amplifier. Among other things this means a single processor can be used for both encode and decode functions by a suitable switching arrangement.

In a wideband compander of this kind having the kind of characteristic at Fig. 4, a low-amplitude signal below the operating threshold would result in the maximum amount of low-level boost being applied, and on decoding the noise level will be appropriately reduced; a high-amplitude signal would result in no noise reduction. Thus an intermittent high-amplitude signal could modulate the noise level, producing breathing (unless high-level signals were present in the same frequency band as the noise. This breathing can occur in any kind of wideband compander, of course).

In the Dolby A system this effect is overcome by splitting the audio band into sections in the additive signal path, each section having its own compression and control circuitry. A high-amplitude signal in one band will not then prevent noise reduction being obtained in bands above and below. Within each band, the presence of a high-amplitude signal is relied on to mask, that is reduce the perceptibility of, noise components close to that signal. Studies of auditory masking show a shift in the hearing threshold in the presence of a (masking) tone, which effect can extend upward in frequency to a considerable extent; downward to a much lesser extent, the amount depending on the level of the masking tone.

When the economics of band splitting are judged against the extent of this masking effect, the amount of noise reduction required, and the value of threshold level in relation to the benefits of the additive technique, it turns out that four bands give a satisfactory compromise of cost versus performance. Splitting the band with 12dB per octave filters in the ranges 80Hz low pass, 80Hz to 3kHz band pass, 3 to 9kHz band pass, and 9kHz high pass would give a uniform 10dB boost (and hence noise reduction) to low-level signals, as determined by setting compression threshold at 40dB below peak operating level. By making the 3 to 9kHz bandpass filter into a high-pass filter, an additional boost is obtained, gradually increasing from about 5kHz to a maximum at 15kHz. The lowest band provides reduction in the hum and rumble range, the second reduces mainly broadband noise, tape print-through and cross-talk, while the upper bands reduce hiss.

Dolby B-type system

The cost and complexity of the A system is not really appropriate to consumer products. Moreover, in slow-speed tape machines in particular the noise spectrum has a different distribution to that occurring in the studio situation, on account of the slower tape speed and thin

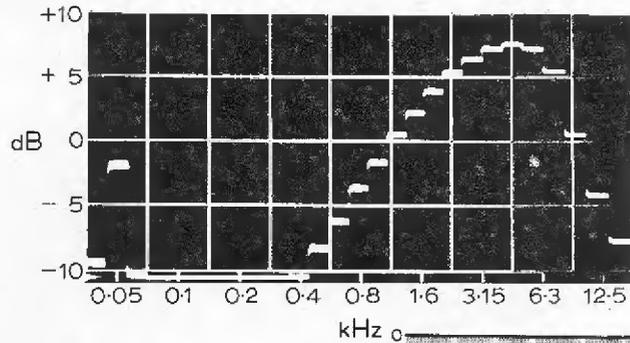


Fig. 6. Noise spectrum of low-noise ferric oxide cassette tape shows problem is a mid- to high-frequency one, rather than a broadband one for which the Dolby A system was developed.

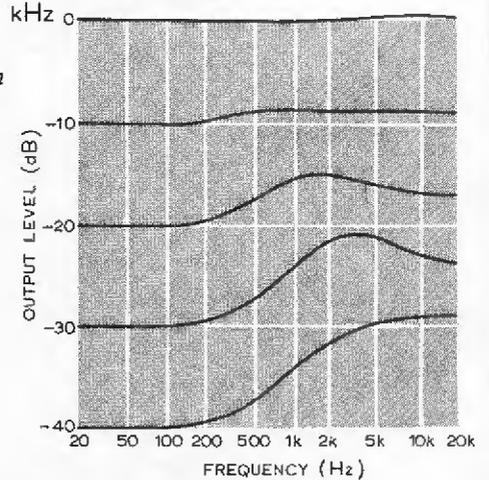
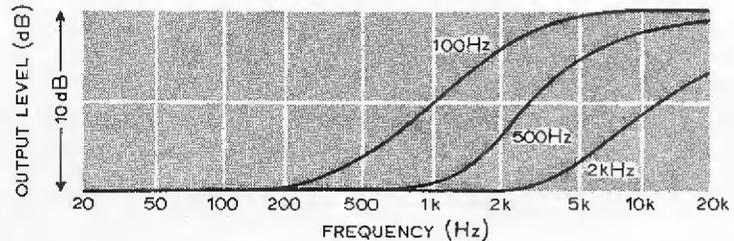


Fig. 7. In the Dolby B and JVC a.n.r.s. systems, low-level high-frequency signals are boosted during encoding by 10dB at low levels, the amount of boost decreasing as input level increases. Characteristics shown are amplitude-frequency response curves, with input level as a parameter, for the encoding process.

Fig. 8. Because the compressor circuit is made frequency sensitive in B-type processors, frequency at which boost, and hence noise reduction, occurs rises with increasing input level. Thus noise reduction is preserved in the presence of mid-frequency signals at high amplitude, which would otherwise reduce or prevent noise reduction. Curves show response below threshold level in presence of 0dB tones.

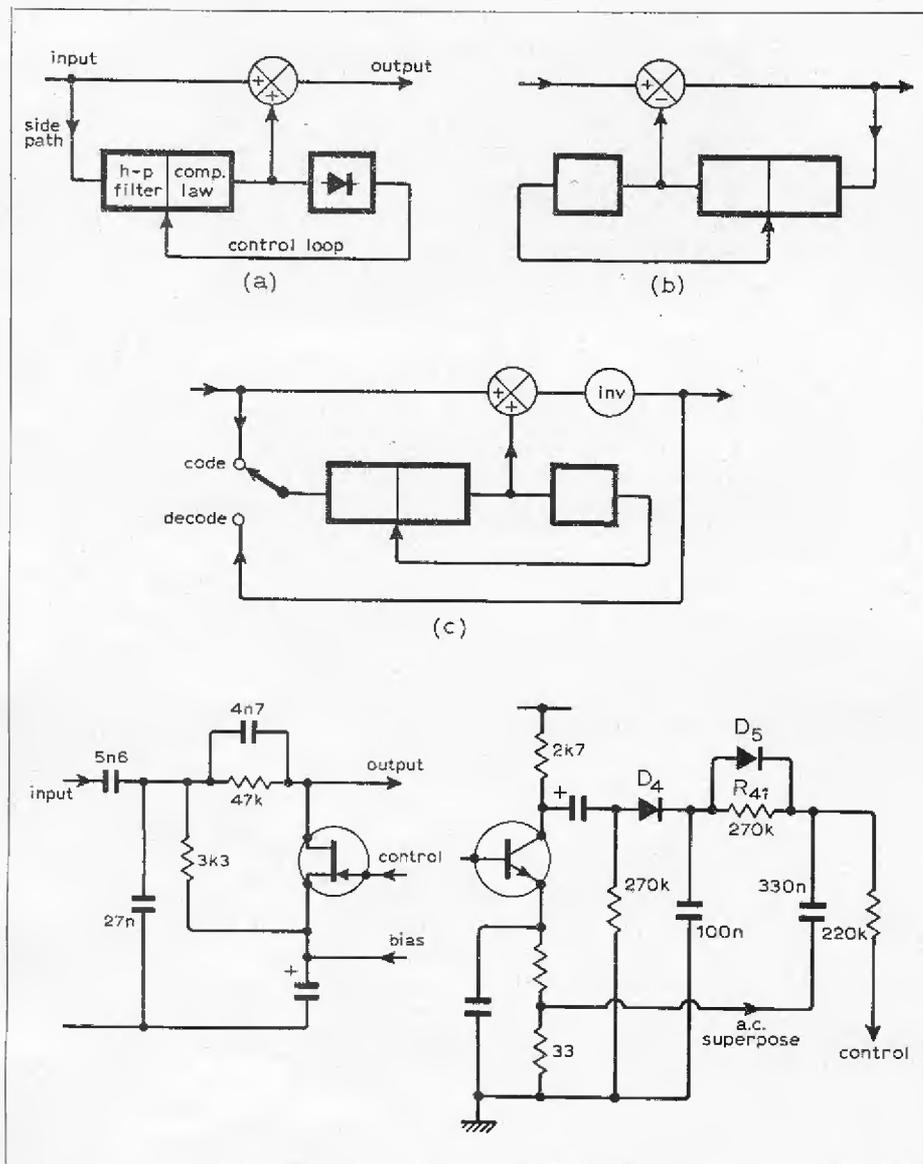


oxide layers used in tape cassettes. Fig. 6 gives a typical DIN-weighted noise spectrum taken from a low-noise ferric oxide tape cassette, showing the noise problem is mainly a mid- to high-frequency one. Noise reduction in the B-type system is therefore limited to this frequency range and Fig. 7 shows the amount of boost (hence noise reduction) applied at various input levels; a fixed high-pass filter placed in the subsidiary signal path would achieve this end. What, then, about noise modulation which in the A system was reduced to imperceptible amounts by the multiband feature?

In the B system, such a filter prevents high-level low-frequency tones from activating the compression circuit, so there is no noise modulation by l.f. components. But there could still be modulation by high-level signals close in frequency to the filter cut-off. The trick

to avoid this, unique to the Dolby B circuit, is to move the filter passband higher in frequency, so that the high-level signal would then be below the filter passband. The curves of Fig. 8 show the effect of the variable-frequency filter under the influence of a high-level tone at three different frequencies; the lowest-frequency curve representing the lower limit of the combined filter's translation in frequency. As the figure shows, with a high-amplitude tone of 500Hz applied, there is some 8 or 9dB of noise reduction at 10kHz; even with a tone at 2kHz there is still some noise reduction obtained. Had the filter passband remained fixed, these high-level tones would have caused the variable-gain element to operate, resulting in reduced or zero contribution from the subsidiary path, and hence little or no noise reduction.

Fig. 9 shows a simplified block diagram



of B-type processors, the encoder at (a), and the decoder at (b) with the same filter and compressor circuitry now in a negative feedback loop. In (b) a phase inversion is clearly required, which in (a) it is not. A simple dodge, that leads to a simplified encode/decode switching arrangement, is to re-site this phase inverter in the main signal path after the summing amplifier. The inverter can now remain in-circuit permanently, forming part of the feedback loop only during decode, Fig. 8(c).

Circuit operation. The way in which the voltage-variable filter and compressor operates is interesting. A fixed high-pass filter, formed by the parallel combination of the 5.6 and 27-nF capacitors (fed from a low impedance source, they are effectively in parallel) and the 3.3k Ω resistor determines a turnover frequency of 1.5kHz (Fig. 10). Imagine that a simple compressor then follows, i.e. a variable attenuator formed by a fixed resistor and the f.e.t. voltage-variable resistor (ignoring the 4.7nF capacitor). The f.e.t. is to be controlled by a direct voltage obtained after rectification of the signal passed by the filter/f.e.t. combination. Without any direct voltage applied to the f.e.t. gate, as would be the case for inputs of any level below the filter passband and

for low-level inputs within the passband, the f.e.t. resistance is nominally infinite. The filter circuit would thus give minimum attenuation of h.f. signals and pass them to the main path, allowing h.f. noise reduction to be obtained. When an h.f. input is of sufficiently high level for the control signal to overcome the f.e.t. bias (this determining the compression threshold), the direct voltage to the gate would cause the f.e.t. resistance to fall, attenuating the signal, and reducing the amount passed to the main path. As the h.f. signal increased, a progressively smaller amount would be returned to the main path. Operation of this principle is shown by the curves in Fig. 7(a), which in fact apply to the Dolby B and a.n.r.s. circuits.

By replacing the fixed resistor with a capacitor (4.7nF) in series with the f.e.t. resistance a second, variable, high-pass filter is formed. With increasing f.e.t. gate voltage, actioned by an increasing signal frequency and/or level, the filter characteristic rises in frequency, "overtaking" the fixed filter curve to largely determine a new, higher, passband (after equilibrium between signal level control and filter is reached). Thus the frequency at which a significant signal is returned to the main path is raised, as depicted in

Fig. 9. Characteristics of Fig. 8 are realized by a voltage-controlled filter and compressor which adds up to 10dB of subsidiary signal to the main path during encoding (a). In decoding, a similar network is used to subtract from the main path (b), the network forming part of a negative feedback loop. This loop means that identical networks can be used for encoding and decoding. By placing the phase inversion in the main signal path, as shown (c), it can be left permanently in-circuit, simplifying encode/decode switching.

Fig. 10. Output of high-pass filter decreases after the compression threshold, set by gate bias, has been exceeded by the control signal. Response curve of combined fixed and variable filter sharpens when the two turnover frequencies coincide.

Fig. 11. Control-loop integrator has variable attack and decay times depending on speed and amplitude of signal changes. Large transients cause D_5 to conduct, shortening loop response time. Superposition of a.c. signal on control loop is to allow f.e.t. to operate symmetrically, thus keeping second harmonic distortion to a low level.

Fig. 8, preserving some h.f. noise reduction in the presence of mid-frequency signals. In the region where the two filter curves are close, the combined filter shape is sharpened to around 10dB/octave, so the effect of the filter action is heightened in this region, and the immunity of the circuit to noise modulation therefore improved.

Dynamic operation

To avoid modulation products being generated by rapid changes of gain in the compressor, which may or may not be cancelled in the complementary expansion process, a long attack time is desirable in the rectifier circuit providing the f.e.t. control voltage. On the other hand, a short attack time is needed to minimize the effect of overshoots, which could have an amplitude equal to the amount of compression.

The extremely elegant solution chosen is to use a time constant that depends on the rate of change of signal. Referring to Fig. 11, the 2.7-k Ω collector resistor and the 100-nF capacitor allow rapid following of a slowly changing input signal. But the time constant of the 270k Ω (R_{41}) and 330-nF component gives an attack time for the control signal of 100ms—long enough to prevent audible modulation products being formed. Diode D_5 is not brought into conduction because the voltage drop across it is never large enough (the discharge time of the 100-nF path being shorter than through the 330-nF capacitor). For large transient changes of input signal the potential across the 100-nF rises faster than that at the 330-nF capacitor so D_5 conducts, reducing

attack time to around 1ms or less. Between these two extremes charging of the 330-nF capacitor is shared by D_5 and R_{41} , as determined by the p.d. across them.

While the effects of transients are limited by the variable attack time, high amplitude transients require more rigorous treatment. Overshoots, as a result of the control loop not operating quickly enough, are limited to a maximum amplitude of 2dB by two silicon clipper diodes. When added back to the main path the clipped subsidiary signal can result in a momentary distortion of a few percent for around 1 or 2 ms, but this occurs at a time when, because of the causal transients in the main path, the ear is least susceptible to it.

As with attack time, recovery time is as much a problem—it must be so short that noise reduction immediately following a high amplitude signal is restored within the time the ear takes to recover its normal hearing threshold, but not so short that low-frequency or modulation distortion results. The circuitry ensures a 100-ms decay time.

In Fig. 11 there is a proportion of a.c. signal from the emitter resistors superimposed on to the direct control voltage. This is to maintain symmetry of operation in the f.e.t. and thus keep second harmonic distortion to a low level by ensuring that $v_{gd} = v_{gs}$. Therefore an a.c. signal is applied to the gate that is half the value of that at the drain. By this means, and by keeping the signal voltage at the f.e.t. low by the capacitance divider prior to the f.e.t., distortion is reduced from a peak of 0.5% to 0.05% (at 1.5kHz and -15dB).

This simplified introduction to noise-reducing systems should help in understanding operation of the B-type circuit, to be given in next month's issue in full.

Acknowledgement. We wish to thank Dolby Laboratories Inc. for their co-operation in developing this *Wireless World* design and particularly Ian Hardcastle for his valuable assistance.

Wireless World

Dolby noise reducer

2 — Construction

by Geoffrey Shorter

This Dolby B noise reduction unit can be used with both open-reel and cassette tape machines. It is intended for decoding Dolby B-encoded tapes and f.m. transmissions, and for encoding and decoding your own tapes.

The circuit diagram is split into three parts: the main signal path, Fig. 12 (top), the subsidiary or side path, Fig. 12 (bottom), and the circuitry used in setting up the unit.

The input signal to be processed from the auxiliary, tuner or tape inputs passes via the switching arrangement of Fig. 13 to point D in Fig. 12 (top). In addition to providing 12dB of gain, Tr_1 ensures a proper source impedance for the low-pass filter. Filter components L_1 and C_5 provide a gradual attenuation (-3dB at 28kHz), while the 19kHz filter switch brings in additional components to give a response ± 1 dB at 15kHz, -31dB at 19kHz and -22dB at 38kHz.

With high-quality open-reel machines whose response is flat up to 19kHz, the additional filter may be out of circuit when the source is free from spurious signals. But because the bandwidth of signals into the record processor should be the same as that for signals entering the playback processor for proper matching, it is usually advisable to have the filter in, especially with cassette machines having a fast-falling response. If there is any risk of unwanted signals above audibility, for example from a stereo decoder or tape bias oscillator, the filter must be switched in. If such signals are above the compression threshold the noise reduction will not operate correctly.

The direct-coupled pair Tr_2 and Tr_3 have a low output impedance for driving the voltage-controlled filter and it is at this point that the signal path is split during encoding. The main signal path continues via the summing junction following R_{14} .

The final directly-coupled amplifier pair Tr_4 and Tr_5 must be inverting because on decoding the subsidiary or side signal path is arranged to form a feedback path from its output to input via R_{15} (See Fig. 9c May issue).

For encoding, the signal at point A

passes via a series of switches to point B in the side-path section, Fig. 12 (bottom), and is returned to point E after processing. Point G feeds the meter amplifiers. The processed output is available at C, passing through the switching arrangement of Fig. 13 to the record output socket, Skt_1 , pin 4.

In decoding, the signal is taken from a recorder via pin 5 of Skt_1 to point D. The output from Tr_4 , Tr_5 at point C is passed to the side path at B, through switch Sw_{1b} in Fig. 13. Decoded output appears at Skt_2 (pin 5) via Sw_{1c} .

From the side-path dynamic filter, whose operation was described in the May issue, the signal is amplified by 26dB by Tr_6 and Tr_7 , and extracted at the overshoot suppression diodes, D_2 and D_3 . When combined with the main path signal via R_{15} this results in either a boost of up to 10dB during encoding or a loss of up to 10dB during decoding. (Diode D_1 forms part of a temperature compensation network for the f.e.t. bias.) The variable time-constant control-voltage circuit, following Tr_8 and described last month, also provides an a.c. signal of half the f.e.t. drain voltage. This signal, obtained by attenuating the 26-dB amplified signal with R_{37} and R_{44} , is passed through C_{20} to linearize f.e.t. operation.

Setting-up circuitry (in kit version)

Because this noise reduction unit can be used with a variety of tape recorders, the side-path includes its own 400Hz oscillator so that a standard-level tone can be recorded, played back and the processor calibrated for the particular tape used. The 400Hz tone is obtained by switching the side-path circuit (Sw_{3c}), to form a Wien-bridge oscillator with R_{46} , C_{27} , C_{26} and R_{45} around Tr_{106} , 107 & Tr_{108} . Oscillator output is taken from point E and applied via point F to the processor input, D, by Sw_{3c} and Sw_{3a} . Switch Sw_{3f} alters the control time constant to prevent oscillator instability. Potentiometers RV_3 and RV_{103} are used to set the level of the 400Hz tone for both left and right channels respectively, but only the left-channel side-path circuit is wired to oscillate. This

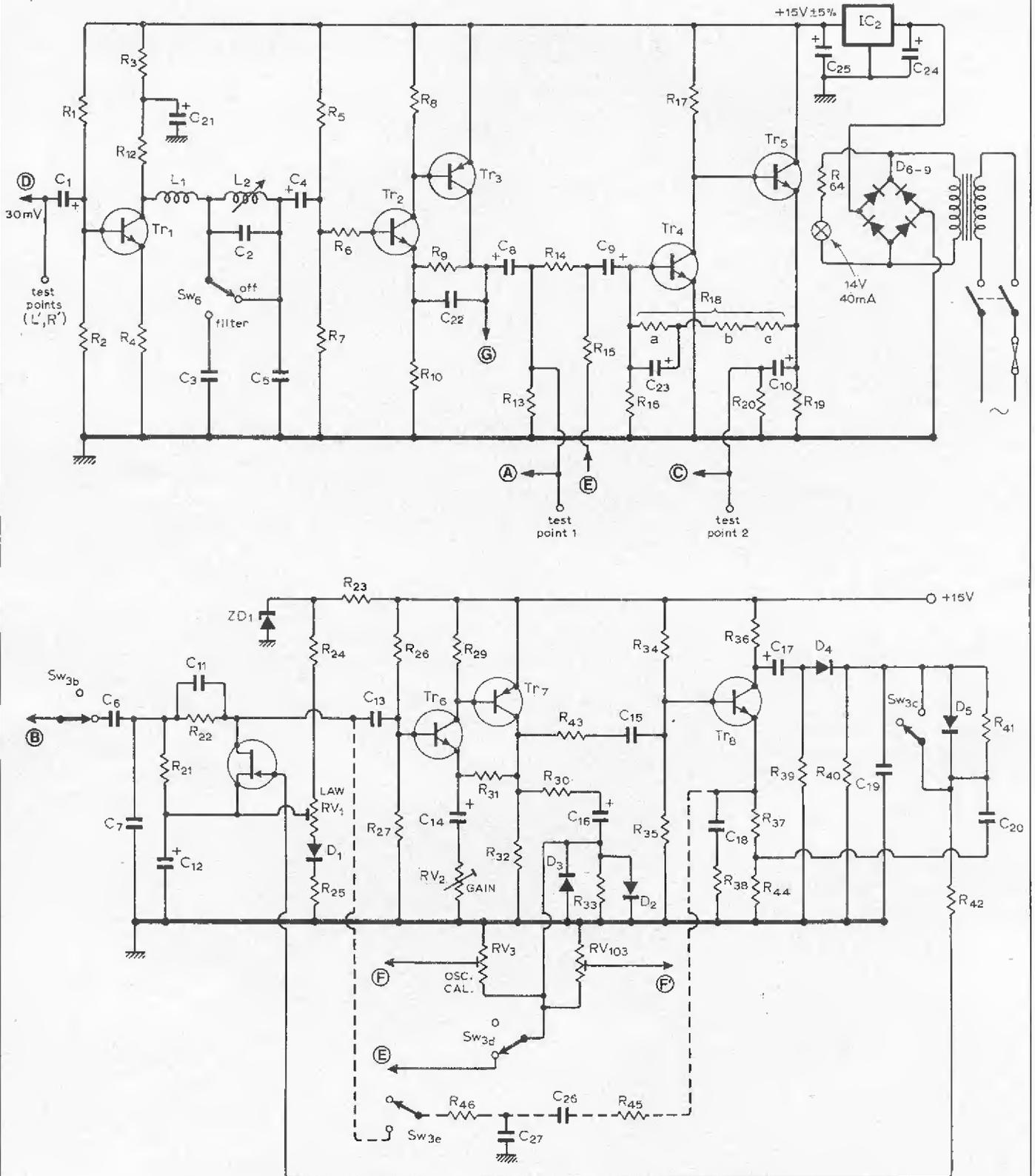


Fig. 12. Circuit of one channel of the stereo Dolby B noise reduction unit. Upper circuit is of main signal path, input at D, output at C. Point G feeds meter circuits of Fig. 15, while point A or point C feeds the side-path input B (bottom), according to whether encode

or decode is switched by the interface circuit of Fig. 13. Side-path output from E is combined with main signal via R₁₅. Connection shown with broken line forms a Wien bridge oscillator to provide a 400-Hz calibration tone. Output is via oscillator level controls

RV₁₀₃ and feeds point F in Fig. 13. This additional circuitry, including potentiometers and Sw_{3b-e} is used on one channel only (the left channel in the kit design). Resistor R₃₃ is omitted in left channel if used as an oscillator.

adjustment, and that of RV_{1,101} and RV_{2,102}, are made with the aid of the right channel meter, calibrated in the kit design by a further oscillator (Fig. 14). This oscillator provides a well-defined output of 580mV, whose accuracy is determined by the supply line regulation of 5%.

For the kit design the oscillator of Fig. 14, including the components shown by the broken lines and with its output feeding the attenuator of R₆₀ and R₆₁ (a), provides the 580mV signal to calibrate the meter. After this calibration, R₄₈ and R₅₇ are removed and the second network, (b), of Fig. 14 wired in to provide a 5kHz sinewave source for aligning the circuit. The input filter coil L₂ is used temporarily in this oscillator.

The two meter circuits of the kit design use two parts of an LM3900 Norton or current-differencing amplifier in a "perfect diode" arrangement, Fig. 15. Because the circuit is set-up at low levels, R₅₅ is temporarily reduced in value to increase sensitivity for these measurements. Additional current gain is provided by Tr₉. Only the right-channel

meter is used to measure the low levels.

Circuit options

The unit can of course be constructed without using the kit. Provided that normal good practice is followed in circuit construction, assembly on Lektrokit or Vero circuit boards should be no problem. But for those constructors unfamiliar with normal practice, we recommend using either the full kit or a smaller p.c. board. This smaller board is for a single-channel processor without the switching and setting-up circuitry of the full stereo board, and is available separately.

If similar functions to those of the kit are required the same switching arrangements of Fig. 13 can be used. Selected field-effect transistors are available separately through *Wireless World* (see panel).

The simplest possible circuit option is for playback of B-encoded cassettes. Designed for use as a noise reduction unit, the circuits have many more facilities than required for a playback-

only processor; nevertheless, Fig. 12 can be used in this application with an enormous simplification of the switching. The circuit can be permanently wired in the decode mode, and needs only the switch Sw₄ in Fig. 13. Point C is permanently wired to point B via Sw₄ and the signal from the head amplifier wired to point D via the play cal. control. The filter components can be omitted if use is to be always limited to playback of recorded cassettes.

Inclusion of the facility for decoding B-type f.m. transmissions can be added to this basic design simply by retaining Sw_{2a} and Sw_{1a} and associated input circuitry. More simply, the two switches can be combined into one.

Maximum cost-effectiveness is clearly obtained with the encode/decode version, as almost all of the circuitry is common to both modes — see Fig. 9, May issue, page 204. The first basic simplification possible of this switchable family is omission of the f.m. facility. Switch Sw₂ is eliminated, being permanently wired in the position shown in Fig. 13.

Components

Electrolytic capacitors are 16-volt working (except C₂₄, C₁₄, C₁₁₄, C₂₀₉, and C₃₁, C₁₃₁). Polystyrene capacitors may be marked with a "k" multiplier instead of "n". (Polyester capacitors are colour coded.)

- C_{1,101} 10μ electrolytic
- C_{2,102} 3n 5% polystyrene
- C_{3,103} 3.9n 5% polystyrene
- C_{4,104} 10μ electrolytic
- C_{5,105} 2.2n 5% polystyrene
- C_{6,106} 5.6n 1% polystyrene
- C_{7,107} 27n 1% polystyrene
- C_{8,108} 10μ electrolytic
- C_{9,109} 10μ electrolytic
- C_{10,110} 10μ electrolytic
- C_{11,111} 4.7n 1% polystyrene
- C_{12,112} 10μ electrolytic
- C_{13,113} 100n metallized polyester
- C_{14,114} 47 or 50μ 6-volt electrolytic
- C_{15,115} 100n metallized polyester
- C_{16,116} 10μ electrolytic
- C_{17,117} 10μ electrolytic
- C_{18,118} 100n metallized polyester
- C_{19,119} 100n metallized polyester
- C_{20,120} 330n metallized polyester
- C_{21,121} 10μ electrolytic
- C_{22,122} 22p polystyrene
- C_{23,123} 10μ electrolytic
- C₂₄ 1000μ 25-volt electrolytic
- C₂₅ 10μ electrolytic
- C₂₆ 10n metallized polyester
- C₂₇ 47n metallized polyester
- C₂₈ 100n metallized polyester
- C₂₉ 10μ 10-volt electrolytic
- C₃₀ 33n metallized polyester
- C_{31,131} 10μ 10-volt electrolytic
- C_{32,132} 330n metallized polyester
- C₃₃ 1.5n disc ceramic
- C_x(two) 2.7n* polystyrene

Resistors 1/4-watt, 5% tolerance unless otherwise stated.

- | | |
|-------------------------------|----------------------------------|
| R _{1,101} 470k | R ₃₃ [†] 22k |
| R _{2,102} 47k | R _{34,134} 120k |
| R _{3,103} 1k | R _{35,135} 47k |
| R _{4,104} 470 | R _{36,136} 2.7k |
| R _{5,105} 43k | R _{37,137} 1k 2% |
| R _{6,106} 100 | R _{38,138} 47 |
| R _{7,107} 6.8k | R _{39,139} 15k |
| R _{8,108} 2.2k | R _{40,140} 270k |
| R _{9,109} 820 | R _{41,141} 270k |
| R _{10,110} 180 | R _{42,142} 220k |
| R _{11,111} 270k | R _{43,143} 8.2k |
| R _{12,112} 3.3k | R _{44,144} 33 |
| R _{13,113} 33k | R ₄₅ 27k |
| R _{14,114} 150k 2% | R ₄₆ 6.8k |
| R _{15,115} 180k 2% | R ₄₇ 1M |
| R _{16,116} 27k | R ₄₈ 1M |
| R _{17,117} 22k | R ₄₉ 4.7k |
| R _{18a,118a} 150k | R ₅₀ 2.2M |
| R _{18b,118b} 150k 2% | R ₅₁ 3.9M |
| R _{18c,118c} 10k | R _{52,152} 560 |
| R _{19,119} 1k | R _{53,153} 150k |
| R _{20,120} 33k | R _{54,154} 150k |
| R _{21,121} 3.3k 1% | R ₅₅ 330k 2% |
| R _{22,122} 47k | R ₅₆ 330k |
| R _{23,123} 2.2k | R _{56,156} 330k |
| R _{24,124} 6.8k | R _{57,157} 1k |
| R _{25,125} 2.7k | R ₅₈ 10k |
| R _{26,126} 1M | R ₅₉ 3.9M |
| R _{27,127} 1.8M | R ₆₀ 110k 2% |
| R _{28,128} 1k | R ₆₁ 10k 2% |
| R _{29,129} 15k | R ₆₂ 15k 2% |
| R _{30,130} 6.2k | R ₆₃ 6.8k |
| R _{31,131} 8.2k | R ₆₄ 82 |
| R _{32,132} 10k | R _x (two) 18k* |

† Two needed if cal. osc. not used.

Transistors

- Tr_{1,101}, Tr_{5,105} ZTX109C, BC109C or equivalent
- Tr_{2,102}, Tr_{4,104}, Tr_{6,106}, Tr_{8,108}, Tr₉ ZTXA11, or ZTX109, BC109, etc.
- Tr_{3,103}, Tr_{7,107} ZTXA21, 2N4058 or equivalent
- f.e.t.s (two) 2N5458, MPF104, 2SK30D or GR specially selected.

Diodes

- D_{1,101}, D_{4,104}, D_{10,11} OA91
- D_{2,102}, D_{3,103}, D_{5,105}, D_{110,111} 1N914
- D_{6,9} 1N4001 or 1N4002
- ZD_{1,101} BZV19C8V2 (8.2V zener E-line package)
- IC₁ LM 3900, MC3401 or MC3311
- IC₂ L131 or TDA1415

Potentiometers

- RV_{1,101} 5k or 4.7k lin. preset (law)
- RV_{2,102} 470 lin. preset (gain)
- RV_{3,103} 50k or 47k lin. preset (400-Hz osc. level)
- RV_{4,104} 50k or 47k log. preset (play cal.)
- RV₅ 5k or 4.7 lin. preset (5kHz osc. level)
- RV_{6,106} 20k log. preset (record cal.)
- RV_{7,107} 5k or 4.7k log. preset (f.m. cal.)
- RV_{8,108} 1k lin. preset (meter cal.)
- RV₉ 50k dual log/reverse log. (record balance)
- RV₁₀ 50k dual log. (record level)
- RV₁₁ 5k dual log. (output level)

Inductors

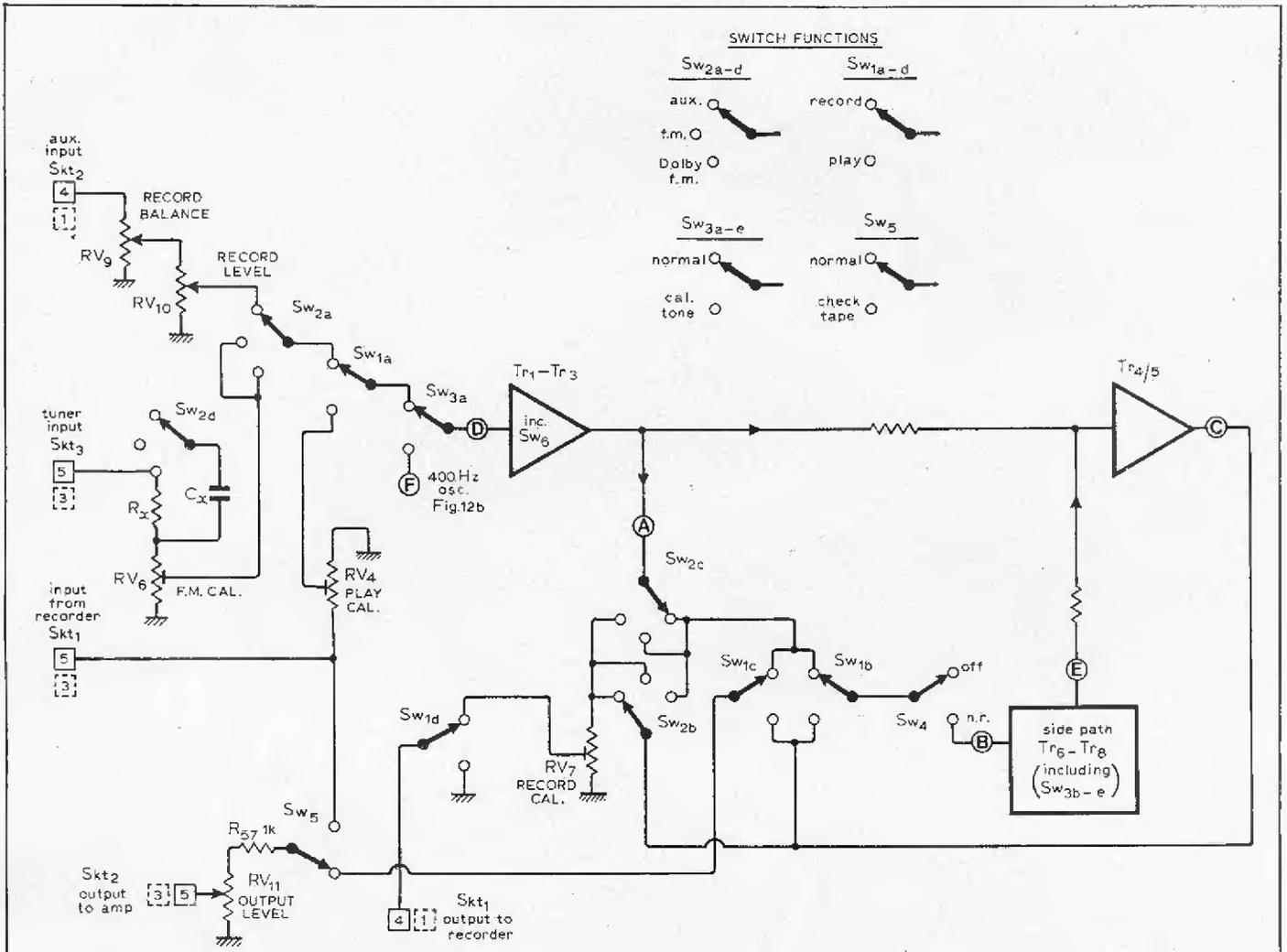
- L_{1,101} 36mH ± 5% (Toko 30569 in kit)
- L_{2,102} 23mH, Q ≥ 60 (Toko 30568 in kit)
- Transformer 240/17V nominal

Other parts (all supplied in kit)

- Dual 200-μA meter, plastic foam, wire-ended 14-V 40-mA lamp ● fuse and holder ● 7-button switch unit, 6-pole switch (Sw₂), mains switch ● two printed boards ● three knobs ● three DIN sockets ● chassis, front panel, screws, tag strip, meter bracket ● labels, connecting wire, mains lead, strain-relief bush ● cabinet.

*Values for 50 to 25μs change in time constant. For 75 to 25μs change, as in USA, use 1.8nF and 39kΩ.

not on main board



If a separate audio oscillator is available, the circuit of Fig. 14 version (b), need not be used. If the unit is to be built into a tape machine you may wish to omit the meter circuits, and adopt a simpler switching scheme. But you would then need an a.c. millivoltmeter for setting up. The 400Hz oscillator wiring, shown by the broken line in Fig. 12, could also be omitted if the same tape is always used. We recommend retention of this feature to take account of tapes with different sensitivities (see part three).

Setting-up procedure

For proper operation, the encoding and decoding signal processors and the intervening signal channel must be matched at all frequencies of interest and all levels. Any errors in channel gain, on a wideband or frequency-selective basis, can produce a mismatch, or error, in overall response. But first, the circuit must be adjusted to provide the correct degree of low-level h.f. emphasis and de-emphasis (10dB at 5kHz), and the correct threshold level. Matching between encode and decode modes must be checked. Then the processor must be level-matched to the equipment and media (tape of f.m. radio) it is to be used with; to be covered in part three.

If the circuit of Fig. 12 is constructed

Fig. 13. Switching interface for one channel of Dolby B processor allows decoding and encoding of tapes, recording and simultaneous decoding Dolby f.m. transmissions (current in the USA), encoding of normal f.m. transmissions, and a normal signal for monitoring during recording. This arrangement is used in the kit design, but could be simplified in other constructions, for instance by omitting Dolby f.m. provision given by Sw₂. Switch Sw_{3a} appears in both channels, but remainder of Sw₃ is used in one channel only. Pin numbers on kit DIN sockets are indicated for both channels (dashed boxes for left).

without using the kit, apply the following setting-up procedure (see part 3 for kit). You will need an a.c. millivoltmeter and an oscillator, unless you adopt the technique using the circuits of Fig. 14 & 15, as in the kit design.

Before starting, make sure that the f.e.t. gates are shorted to earth. Start in the record mode with the noise reduction switched out (also the cal. tone off and the filter out, if used).

—Set law control RV₁ to produce maximum positive voltage on the f.e.t. source.

- Feed in 5kHz signal at a level to give 17.5mV at test point 1 and note signal level at test point 2.
- Switch in noise reduction and adjust gain control RV₂ to give a 10 ± 0.25 dB rise at test point 2. Note signal level*.
- Remote f.e.t. gate short and adjust law control RV₁ for a 2 ± 0.25 dB drop at test point 2.
- Replace gate short and check that level returns to that identified by*. Finally, remove gate short.

Encode/decode matching check. Without altering the control settings, switch to play mode.

- Switch out noise reduction and short f.e.t. gate.
- Feed in 5kHz signal at a level to give 44mV at test point 2.
- Check that signal drops by 10 ± 0.5 dB when noise reduction is switched in.
- Remove gate short and switch in noise reduction. Check that signal at test point 2 is $17.5\text{mV} \pm 0.5\text{dB}$.

Decode-only processor. As with the switchable encode/decode version, ensure that f.e.t. gates are shorted to earth, and switch noise reduction off.

- Set law control RV_1 to pinch-off f.e.t. i.e. maximum positive voltage on source.
- Feed in 5kHz signal to give a level of 44mV at test point 2.
- Switch in noise reduction and adjust gain control RV_2 to give a fall of $10 \pm 0.25\text{dB}$ at test point 2. Note signal level*.
- Remove gate short and adjust law control RV_1 to give a rise of $2 \pm 0.25\text{dB}$ at test point 2 (should be 17.5mV).
- Replace f.e.t. gate and check that level returns to that indicated by*
- Remove gate short.

Meter and oscillator calibration. If the meter circuits are to be fitted, calibrate them by applying a 580mV tone and adjusting for a 0dB reading. One of the meters can then be used to calibrate the 400Hz oscillator level, if used. (The circuit of Fig. 14 from the kit design could be used if fed from a sufficiently well-regulated supply line; 5% in the circuit of Fig. 12.)

- Apply input signal to point D to give 580mV at point G.
- Adjust RV_8 for 0dB meter reading.
- Operate cal. tone switch (if oscillator fitted).
- Adjust RV_3 to give 0dB meter reading.

The unit is now ready for use. But to ensure compatibility with commercially-available Dolby tapes, and to ensure interchangeability of tapes from machine to machine, it must be calibrated using a level-setting tape, to be detailed in part three of this article.

Kit construction

Successful operation of the unit depends on a number of factors. As well as proper matching of the unit, strict adherence to component tolerances and alignment procedure, use of selected f.e.t.s, and a low ripple in the supply line are all essential to correct operation. For these reasons the parts for the unit are available as a complete kit.

The printed board of the kit is designed to keep wiring to an absolute minimum; it is for this reason that switches, calibration controls, and DIN sockets are board-mounted types. First

Fig. 14. Oscillator circuit used in kit for generating a 1-kHz tone (a) for calibrating the meters. Though a square wave, the magnitude is chosen to give the same reading as a 580-mV sine wave. Circuit is subsequently used to provide a 5-kHz circuit alignment tone at (b) by temporarily using L_2

thoughts indicated a double-sided board would be needed together with plated-through holes, but this would make an expensive board. The same effect could be achieved with a larger single-sided board but would result in a large number of links. The relatively large number of controls finally decided the format. To keep board length down, some controls had to be mounted above others, and as there was to be a minimum of wiring, the top controls are mounted on to a separate board. The advantage of this sandwich board technique is a saving of about 24 links.

In the instructions, component numbers for the left-channel have 100 added to the number for the right channel: thus R_{121} is the left channel component corresponding to R_{21} in the right channel.

Kit assembly instructions

A number of pins are supplied with each kit; in fitting them insert from the track side of the board, tap down lightly with a hammer and solder into place. Insert pins as follows

- two pins for the transformer input, marked V_{in} close to IC_2

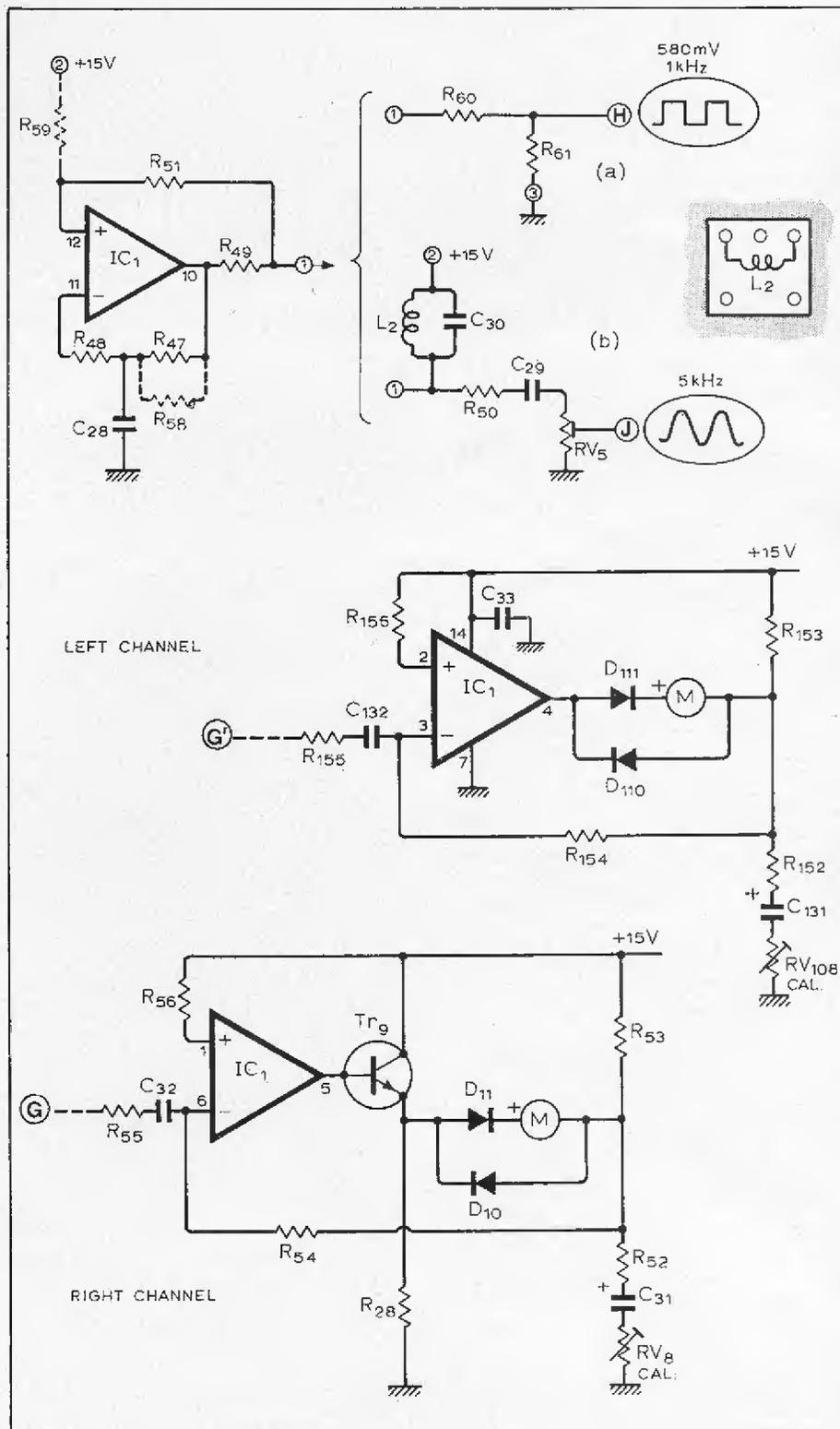
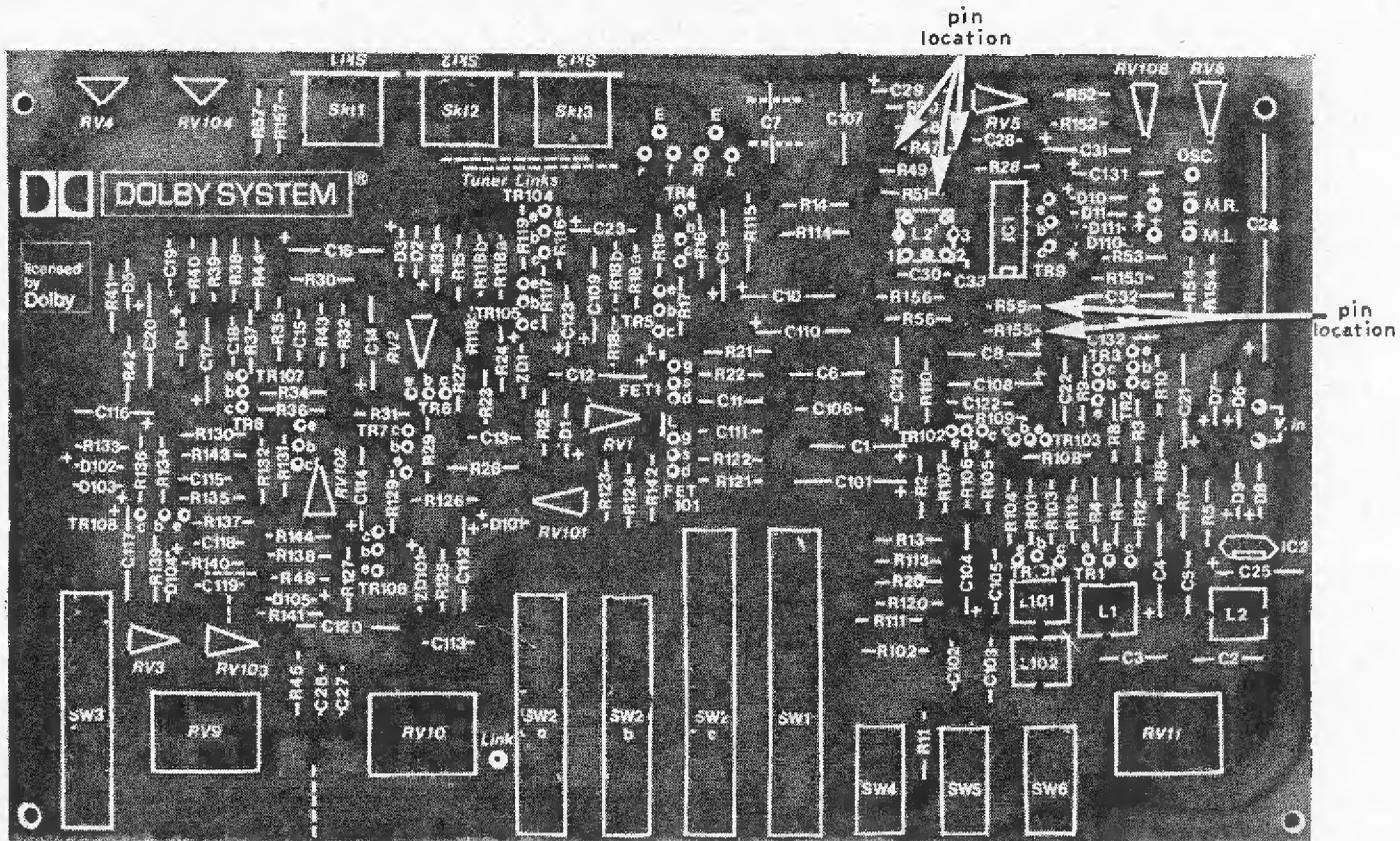


Fig. 15. Meter circuits using "perfect" diode arrangement. Right-channel meter circuit at bottom includes extra gain to allow measurement of low signal levels during alignment.



- four pins for right and left meter outputs, marked \pm M.R., \pm M.L.
- two pins in resistor R₄₇ position
- one pin at the end of R₅₁ close to IC₁ (see Fig. 16)
- one pin each at the end of R₅₅, 155 close to C₃₂ (see Fig. 16)
- three pins in the L₂ position, marked with broken lines, next to IC₁
- one pin at the 5kHz oscillator output point, marked "osc"
- six pins in the holes marked E, R, L; E, r and l between socket Skt₃ and C₇.

There are seven links to be inserted on the main board; two further links are used if a tuner is to be connected to the auxiliary input socket, rather than the tuner input. The two f.e.t.-gate links should be looped, to allow easy breaking and making of the gate during alignment. Close-tolerance components, i.e. resistors of 2% tolerance or better and capacitors of 5% tolerance or better, are separately packed.

- Insert seven or nine links, as appropriate.
- Mount close-tolerance resistors R_{21,121} - R_{14,114} - R_{15,115} - R_{18b,118b} - R_{37,137}.
- Follow with close-tolerance capacitors C_{2,102} - C_{3,103} - C_{6,106} - C_{7,107} - C_{11,111}.
- Mount the remaining fixed resistors and capacitors identified on board, excepting C₃₀, R₄₇, R_{55,155}.

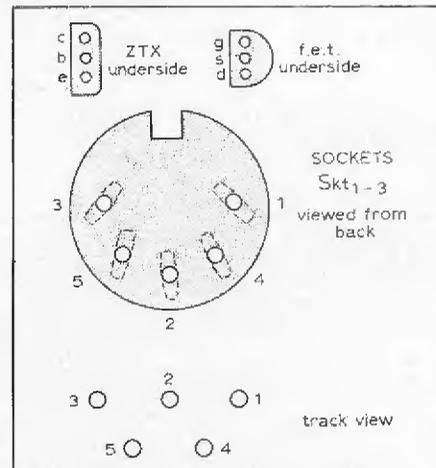
Make sure electrolytic capacitors are inserted the correct way round, that is, indented end to the hole marked +. Note that R₃₈ to R₆₄, R_x and C_x will be left over, in addition to the four components already mentioned.

Fig. 16. Main board markings show seven essential links plus two optional links, for use if a tuner is to be applied to auxiliary socket. Some pin locations are shown. (Boards in kit have a slightly different track arrangement.)

Fig. 17. Socket connections, viewed from "holes". If E-line zener diode is used, as supplied in kit, the + sign on the board should correspond with the position of the collector lead in the E-line package shown right.

- Add pre-set potentiometers RV_{1,101} - RV_{2,102} - RV_{3,103} - RV_{4,104} - RV₅ - RV_{8,108}.

There are four types of diodes, easily identified by the quantities supplied. Zener diodes have the connections of the E-line package, the + lead corresponding to the collector position in Fig. 17. Of the others, the OA91 germanium diodes will be the largest and glass-encapsulated; the rectifier diodes will be the four plastics-encapsulated ones; and the 1N914s should be the smallest, of either glass or plastics. The band-end is to correspond with + on the board. Base connections for the transistors are shown in Fig. 17. The field-effect transistors may have various markings but nevertheless will have been specially selected. Transistors Tr_{1,101} and Tr_{5,105} must be type ZTX109C, but the remain-



ing n-p-n type may be supplied as either ZTXA11 or 109C. IC₁ is located so that the end having the indent or other marking corresponds with the board marking. Solder next in place

- diodes ZD_{1,101} - D_{1,101} to D_{5,105}, D₆ to D₉, D_{10,110} and D_{11,111}
- transistors Tr_{1,101} Tr_{5,105} (ZTX109C), Tr_{3,103} and Tr_{7,107} (ZTXA21), field-effect types, followed by remainder
- integrated circuits IC₁ IC₂.

When positioning the three DIN sockets make sure they are vertical and in line with each other, for appearance's sake. Check functioning of the push-button switches as they are difficult to remove once soldered. As the switch board markings will be covered by the switches, identify them before assembly. Take care to push them fully into the board and ensure that they fit

squarely: any skew will result in misalignment with the front panel. Fit and solder

- three DIN sockets
- switches Sw₁ to Sw₆
- inductors L₁, L₁₀₁, L₁₀₂, but not L₂.

Sub-printed board Components are fitted on to the track side of the subsidiary printed board.

- Solder components C_x, R_x.
- Solder potentiometers RV_{6,106}, RV_{7,107}.
- Attach plastics adjuster inserts into RV₆, RV₇.
- Cut off potentiometer legs flush with the board.

The sub-board should be spaced about 0.09in away from the top of the main switches to ensure potentiometer centres line up with the front panel holes. Matchsticks form convenient spacers.

- Lay matchsticks on Sw_{2a} and Sw₁
- Position sub-board, check alignment and solder
- Join areas on sub-board marked R, L, r, l to corresponding points on main board using twin-screened cable. Earth at one end only to points marked E.

- Connect link point on sub-board to link point on main board almost underneath.
- Insert links marked "Mpx" for use with 25- μ s B-Type f.m. transmissions.

Returning to the main board, be careful to align potentiometer spindles horizontally.

- Solder dual potentiometers RV₉ (log/reverse log), RV₁₀, RV₁₁.
- Check underside of board for solder shorting and dry joints.
- Crop leads to avoid touching chassis.
- Insert thin sheet of card between board and chassis.
- Fix board in position with 6BA screws.

Off-board assembly. Fix in position

- transformer
- fuseholder
- mains switch to meter/switch bracket
- bracket with tag strip under one screw.

At this point you can tape the meter to the bracket temporarily with the piece

of foam plastic material between; normally the meter will be held in position by the front panel. Continue with off-board wiring

- transformer secondary to two points of tag strip (not earth tag)
- the two tags to V_{in} terminals on board
- meter illumination lamp, in series with R₆₄, to the two tags, the junction to a third tag (not earth tag)
- meter terminals to \pm M.R. and \pm M.L. on board (note + terminals on meter)
- mains cable brown lead to transformer primary via fuseholder and switch
- mains cable blue lead, via switch to transformer primary
- mains cable earth lead to earthed tag on strip
- insert strain-relief bush in hole and pass cable through
- stick on labels: one to identify sockets and play calibration potentiometers, the Dolby Laboratories label on the rear close to socket Skt₃, and the third inside chassis close to transformer.

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Wireless World Dolby noise reducer

3 — Kit alignment and calibration

by Geoffrey Shorter

Constructors who build a Dolby-B processor without using the full WW kit have the option of using the power supply included in the circuit of Fig.12 or of using an alternative one, for instance one built into existing equipment. Component values for the circuit of Fig. 12 have been optimized to provide an overload margin of 16dB (equivalent to 1200nWb/m on open-reel) for a 15-volt supply, but voltages between 15 and 24 volts could be used provided component voltage ratings are chosen appropriately. The main requirement is that supply ripple be less than 200 μ V r.m.s. Current consumption at 15 volts is 20mA per processor; with IC₁ and IC₂ it is 30mA. The voltage regulator IC₂, whose output is 15 volts \pm 5%, is essential if the meter calibration oscillator of Fig. 14 is used. Input to the regulator should be not greater than 25V and not less than 18.25V.

Kit setting-up procedure

The procedure for setting up the kit design is a little more elaborate than the basic alignment instructions because it is designed to eliminate necessity for additional equipment i.e. a.c. millivoltmeter and variable-frequency a.f. oscillator. It therefore includes a facility for generating a 5kHz circuit alignment tone, as well as a 400Hz calibration tone. Two meter amplifiers, and a 580mV source (1kHz oscillator) to calibrate the meters, are included to obviate the need for an a.c. millivoltmeter.

In using the in-built meter scale in setting up, it is better to use close-tolerance resistors in an attenuator so that all measurements can be made at one meter reading (0dB). Errors in meter reading are minimized by this technique, and errors due to an inaccurate scale eliminated.

Right-channel meter calibration

The unit is aligned using part of IC₁ as a meter calibration oscillator. The amplifier section of IC₁ based on pins 10, 11 and 12 is first used as shown in Fig. 14. In this mode the amplifier is wired as an astable multivibrator switching between the 15V supply rail and 0V,

Typical performance

Noise reduction: better than 9dB weighted

Clipping level: 16.5dB above Dolby level (measured at 1% third harmonic content)

Harmonic distortion: 0.1% at Dolby level (typically 0.05% over most of band, rising to a maximum of 0.12%)

Signal-to-noise ratio: 66dB (20Hz to 20kHz, signal at Dolby level)

Approximate voltage readings (AVO 8)

	collector	emitter
Tr ₁	9.0	0.6
Tr ₂	14.3	1.5
Tr ₃	7.6	rail
Tr ₅	rail	7.6
Tr ₆	---	8.8
Tr ₇	8.4	rail
Tr ₈	8.0	2.6
IC ₁	pin 4	6.8V
	pin 5	7.7V

with a mark-to-space ratio of about 1:1 and a frequency of around 1kHz. The real voltage swing is a little less due to saturation voltages, but is highly repeatable from one sample to another.

- Connect resistor R₅₉ (3.9M Ω) from the pin at R₅₁ to pin 2 or the L₂' position.
- Wire R₅₈ (10k Ω) in parallel with R₄₇ (1M Ω) across the pins at R₄₇ position.
- Form an attenuator with R₆₀ (110k Ω 2%) and R₆₁ (10k Ω 2%) in series, Fig. 14, earthing the end of R₆₁ by connecting to pin 3 of L₂' and connecting R₆₀ to pin 1.
- Solder one end each of R₅₅ (330k Ω 2%) and R₁₅₅ (330k Ω) to their pins. Take the other end of R₅₅ to the junction of R₆₀, R₆₁ (R₁₅₅ remaining floating). Switch on.
- Adjust RV₈ (Fig. 15) until the r.h. meter reads 0dB. Switch off.
- Remove R₅₅, R₅₈, R₅₉, R₆₀, R₆₁ and do not alter the setting of RV₈.

Circuit alignment

The now-calibrated r.h. meter is used to set the gain and f.e.t. bias controls of both left and right processors with the help of a 5-kHz oscillator, Fig. 14, adapted from the 1-kHz oscillator circuit by using arrangement (b).

- Solder C₃₀ in position, removing and replacing the p.c.b.
- Solder L₂ on to pins 1 and 2 of the L₂ position. Gently screw in the core.

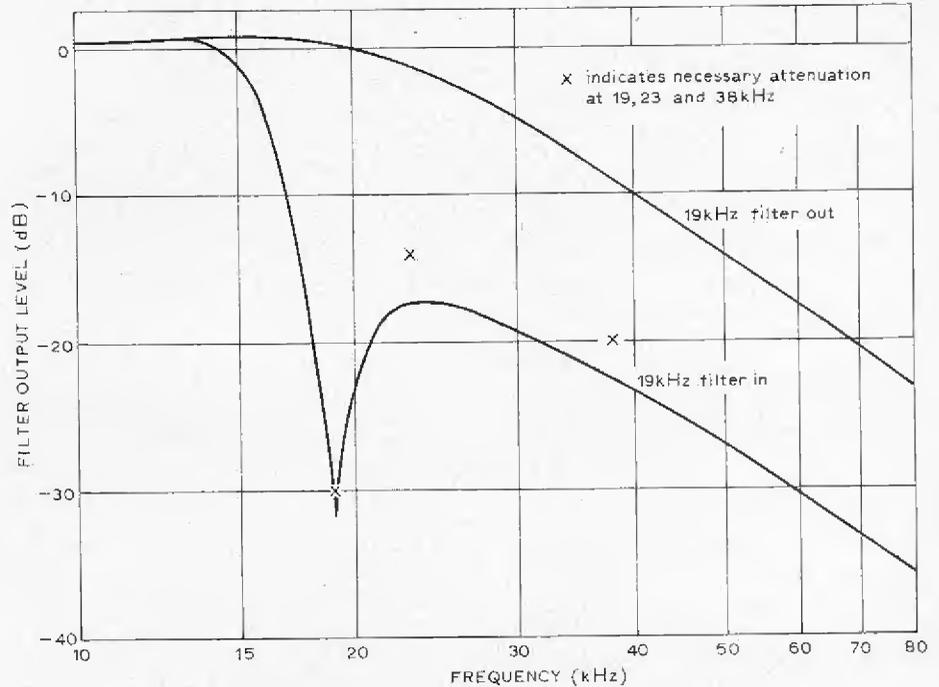
Right-channel circuit alignment.

- Connect R₆₁ (10k Ω 2%) between the R₅₅ pin and test point 1 (TP1) on the sub-board.
- wire the oscillator pin, marked "osc." to the sub-board pin marked R' (input to processor).
- Set RV₅ (oscillator level) fully anticlockwise. Check that no plugs are connected into the sockets. Set RV_{2,102} fully anticlockwise. Switch on.
- Select the auxiliary position for Sw₂. Set the balance control RV₉ to

mid-position and the input level control RV_{10} fully clockwise.

- Ensure the calibration tone switch Sw_3 , the noise reduce switch Sw_4 , and the 19-kHz filter switch Sw_6 are in the off position (out), and the check tape switch Sw_5 is in the normal position (out).
- Check that the f.e.t. gates have previously been shorted to ground by two looped links.
- Turn the law control RV_1 fully clockwise to pinch-off f.e.t.
- Switch Sw_1 to record and adjust RV_5 until the meter reads 0dB (equivalent to 17.5mV at TP1). Switch off.
- Transfer the end of R_{61} from TP1 on the sub-board to TP2 and switch on. Meter should read within ± 1 dB of the previous, 0-dB reading. Note actual reading *. Switch off.
- Solder R_{62} (15k Ω 2%) and R_{63} (6.8k Ω) in series with R_{61} (i.e. between the R_{55} pin and TP2), decreasing meter sensitivity by 10dB. Switch on and check meter reading reduces by roughly two thirds.
- Switch on noise reduction, Sw_4 and adjust RV_2 (gain) to bring back meter reading to that noted above at *. Switch off.
- Cut the f.e.t. gate short for the

Fig. 18. First part in setting-up procedure for kit version (left) shows arrangement used in calibrating the right-channel meter. For aligning the noise reduction circuit the meter calibration oscillator is changed to a 5kHz oscillator, using L_2 temporarily in the L_2' position (centre). Its output, via the "osc" pin, is taken to the processor input (R' for the right channel). To calibrate the 400Hz oscillator, L_2 is put in its normal position, the i.c. oscillator disabled, and the oscillator output taken from TP1 or TP101 (right).



Amplitude response with and without 19kHz filter.

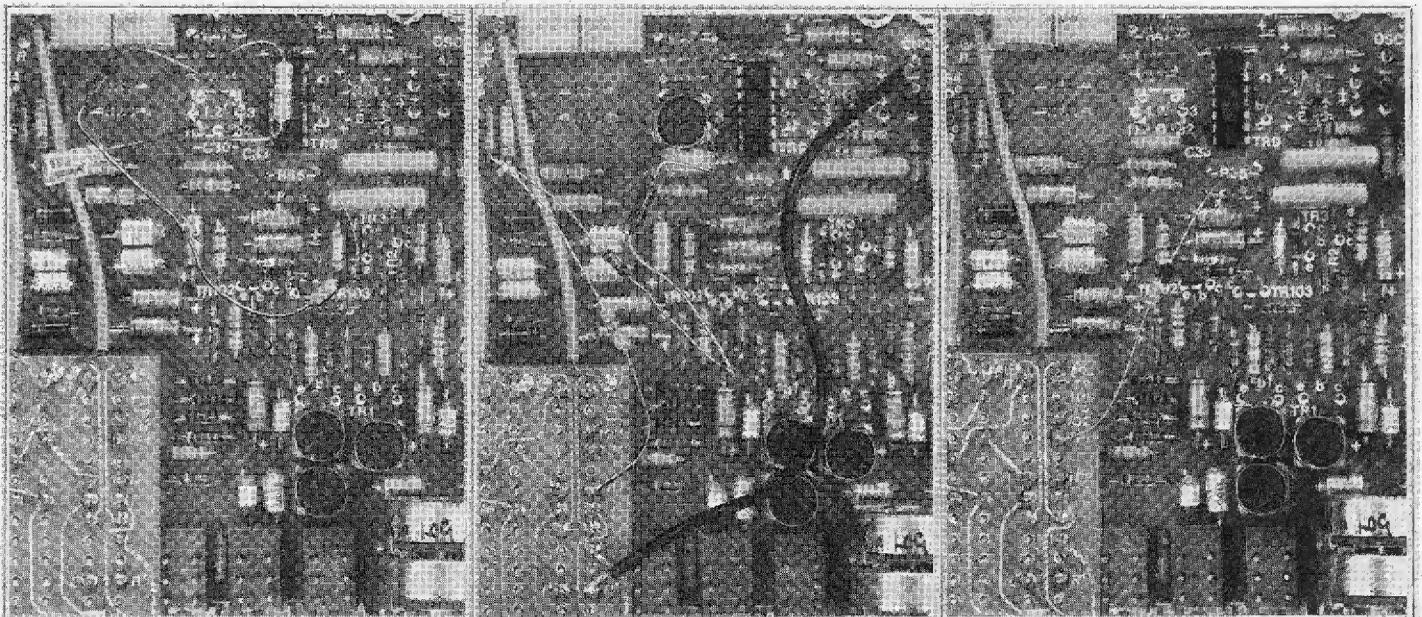
- right-hand channel with wire cutters and short-circuit R_{63} increasing meter sensitivity by 2dB. Switch on.
- Adjust RV_1 (law) until meter reads as noted above, at *. Switch off.
- Re-apply f.e.t. gate short and replace R_{63} . Switch on and check meter still reads as above, at *. Switch off. Remove gate short.

Encode/decode matching check.

- Switch Sw_1 to play and switch noise reduction off, Sw_4 .
- Short-circuit R_{63} , leaving R_{61} and R_{62} connected. Set $RV_{4,104}$ fully clockwise. Switch on.
- Adjust 5-kHz oscillator output level control RV_5 until meter reads 0dB (equivalent to 44mV at TP2). Switch off.

- Switch noise reduction on, Sw_4 . Short-circuit R_{62} and R_{63} so that only R_{61} is in circuit. Switch on. Meter should read 0dB to within ± 1 dB. Switch off.

Left-channel circuit alignment. Now repeat this process for the left channel, starting from the point of connecting R_{61} between the R_{55} pin, (not R_{155}) and the test point — now to be TP101 — on the sub-board. Note that the right channel meter, being calibrated, is still used in setting up the left channel, and that TP101 is to be read for TP1, TP102 for TP2, RV_{101} for RV_1 , RV_{102} for RV_2 , and that the left-channel f.e.t. gate-shortening loop is now implied. The "osc" pin is to be connected to the point L' on the sub-board at the appropriate time.



After repeating for the left channel, switch off. The gain and law adjustments are now complete.

- Remove the f.e.t. gate shorts, R_{61} , R_{62} and L_2 , inserting L_2 into its normal (final) location.

400Hz oscillator calibration

- Solder one end of R_{55} to its pin and connect the other end to TP1.
- Short pins 1 and 3† at the L_2' position, and remove the wire from osc pin to point L'. Switch on.
- Switch Sw_1 to record, press the noise reduce switch Sw_4 off and switch on the 400-Hz calibration tone oscillator, Sw_3 .
- Adjust RV_3 (oscillator level) until the right-channel meter reads 0dB. Switch off.
- Transfer the end of R_{55} from TP1 to TP101 and switch on. Adjust RV_{103} until the r.h. meter reads 0dB.
- Repeat this procedure because of a slight interaction between RV_3 and RV_{103} . Switch off.

Left-channel meter calibration

- Disconnect R_{55} from TP101 and connect the free end of R_{155} to TP101 and switch on.
- Adjust RV_{108} to obtain 0dB at the left-channel meter, being careful not to disturb RV_8 . Switch off the cal. tone oscillator. Switch off.

†Experience has shown that a better method of disabling the 5kHz oscillator is to remove R_{47} .

19kHz filter adjustment

- Wire R_{155} permanently onto the main board, replace R_{55} with R_{61} and connect free end to TP1.
- Connect an f.m. stereo tuner to the auxiliary input and with the aux-tuner links wired in, switch on and tune to a BBC stereo test transmission.*

Alternatively, if a high accuracy (± 50 Hz) 19kHz oscillator is available, connect its output to point R' on the sub-board.

- With zero a.f. modulation,* adjust the record level control RV_{10} to give a 0dB meter reading. Switch the 19kHz filter on, Sw_6 .
- Adjust L_2 for minimum reading on the right-channel meter. Do not adjust L_1 or L_{101} . Increase record level for sharper null near tuning point.

Repeat for the left channel starting by transferring end of R_{61} from TP1 to TP101, and adjusting L_{102} for minimum reading. (In using a 19kHz oscillator, connect to point L' on the sub-board and transfer R_{61} lead from TP1 to TP101 before adjusting L_{102} .)

Calibration

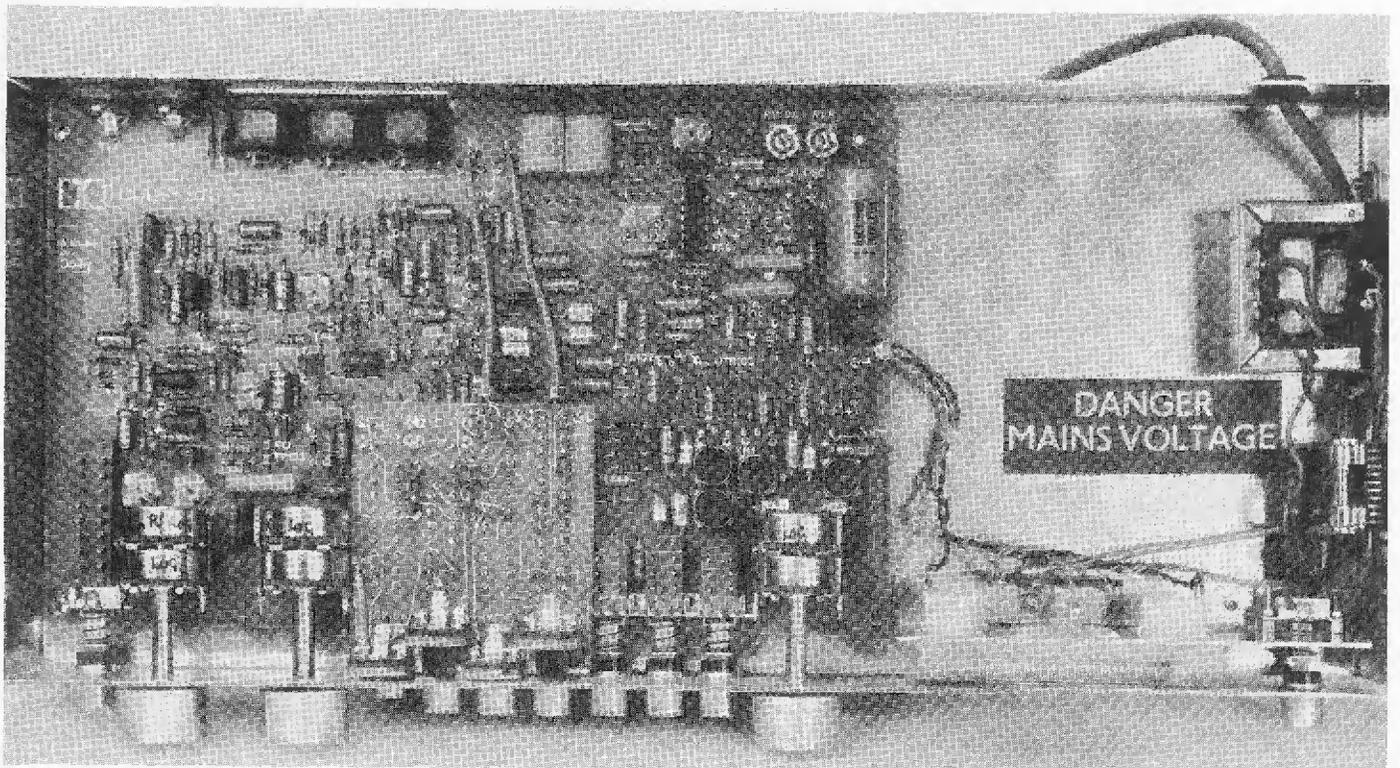
To ensure interchangeability of all Dolby B-encoded tapes and of Dolby

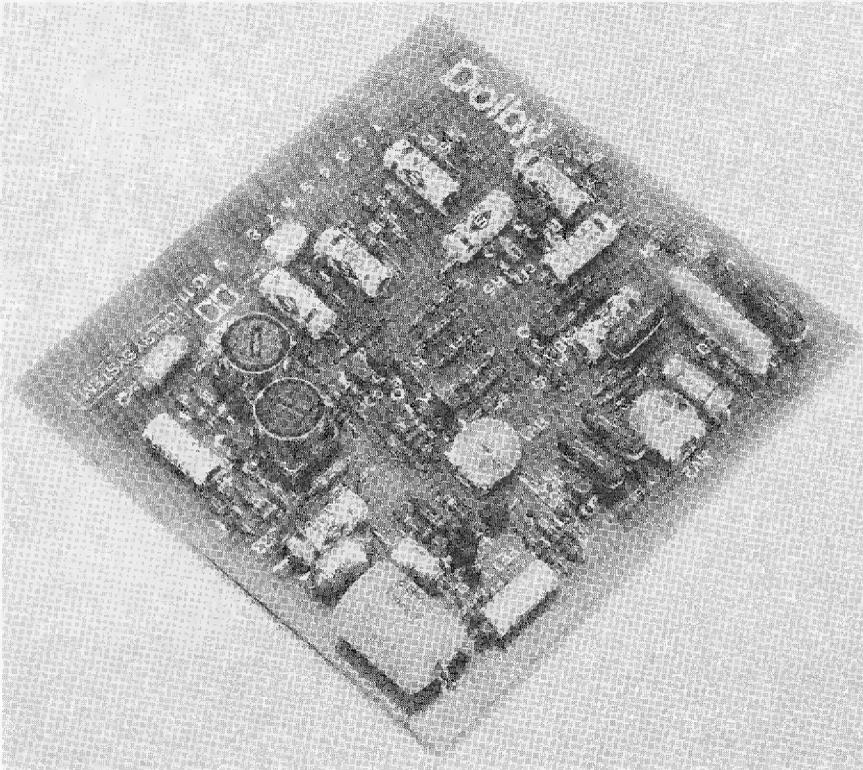
B-equipped machines, the voltage levels in the processors must be related to flux levels on the tape. A certain amplitude level is used that bears a fixed relationship to the noise reduction parameters and to conditions between encoder and decoder. The level chosen corresponds with a flux on open-reel tapes and cartridges of 185nWb/m, with 200nWb/m for cassette tapes, with a deviation of 37.5kHz on f.m. transmissions, and with a voltage level at the processor output of 580mV r.m.s.

This level, often called Dolby level, should not be taken to imply an operating level. If the level-setting meters in the unit are to be used as modulation-depth meters, a mark may be made on the meter to indicate the reference level. Whilst setting this level equal to 0VU on meters can often lead to reasonable modulation depths, this is not always the case: for cassette recorders it is best set at +3VU.

The 400Hz oscillator and tapes recorded with a 400Hz tone to the above level are used in calibrating units, once the circuitry has been set up. When playing or recording the standard flux level, the 580mV level is set by adjusting the play calibration potentiometers during play, and the record calibration potentiometers during recording.

Playback-only decks and units. As the signal levels on encoded tape cassettes are to be related to those in the decoder during playback only, the 400Hz oscillator is not required and calibration is achieved with a calibration cassette, containing the reference flux.





An alternative to the kit design is this single-channel processor, using the circuit of Fig. 12 but excluding power supply, alignment and calibration circuitry. (Track diagram will be given in a subsequent issue.)

- Switch noise reduction off.
- Play calibration tape. Set play gain control on tape deck to 0VU on deck meter, if possible, or to mid-position otherwise.
- Adjust play cal. control for 580mV on meter or Dolby level indication, depending on meter used.

Playback gain controls on the recorder in the signal path en route to the processor input should not now be disturbed.

Switchable encode/decode processors.

Playback calibration

- Switch to play and switch off noise reduction. Connect millivoltmeter to point G if meters not built-in.
- Play calibration tape. Set play gain control on tape deck if fitted to 0VU on deck meter, if possible, otherwise to mid-position.
- Adjust play cal. control for 580mV indication.

This completes playback calibration and the play gain controls on the tape deck should not be altered. Adjust listening level with the output level control following the decoder output (as in Fig.13).

Record calibration

Start by setting record gain control on tape deck to mid-position, if fitted. (If

combined with playback gain, do not adjust.)

- Switch to record.
- Fit blank tape (as recommended by maker or for which bias is correctly adjusted) and feed in 400Hz at points from external or internal oscillator. (If unit has been built into cassette machine and 400Hz input is via line input socket, adjust record level control so that meter reads 580mV, or Dolby level.)
- Record on tape for a few seconds, rewind and playback, switching to play on the noise reduction circuit as well as on the deck. Note whether meter shows about or below 580mV, or Dolby level.
- Make small adjustment to record cal. controls in appropriate direction and record 400Hz tone again, observing meter reading on playback. Repeat this procedure as many times as necessary to obtain correct reading.

This completes record calibration for tapes. If the circuit of Fig.13 or similar has been adopted, recording level is adjusted with record balance and level controls on the noise reduction unit, the level being judged by the tape deck's normal meters.

When the noise reduction unit is connected to a three-head machine with a simultaneous monitoring facility the tape signal may be monitored in its

encoded form by operating the check tape switch.

Simultaneous encode/decode circuits. Constructors with three-head machines having a simultaneous monitoring facility can use single-processor boards permanently wired in the encode and decode modes. If provision for encoded f.m. transmissions is required switching must be arranged so that encoding does not take place during recording. A monitor switch can be provided at the input to the decoder, to switch from tape, via a play cal. potentiometer, to source i.e. a connection to the encoder output via a 580-30mV attenuator, Fig 19.

Playback calibration procedure is as above, but record calibration is simplified.

- Set record level controls on tape recorder to mid-position. Set monitor switch to tape.
- Record on blank tape, operating the calibration tone switch or injecting a 400Hz tone from an external oscillator.
- Adjust record cal. control so that meter reads 580mV, or Dolby level.

FM calibration. If you wish to set the controls for encoded f.m. transmissions, currently being transmitted by stations in the USA, an approximate calibration can be achieved by tuning to a local station, switching to f.m. or Dolby f.m. and setting the f.m. cal. control to give meter readings similar to those obtained when playing pre-recorded tapes. More

Change of time-constant for encoded f.m. transmissions

There are two commonly used pre-emphasis time constants, 50 μ s and 75 μ s. Under certain conditions, these values can lead to reduced modulation at low and medium frequencies or severe amplitude distortion at high frequencies. In the USA the FCC has approved Dolby Laboratories' proposal of using 25 μ s for encoded transmissions, and to receive such broadcasts it is necessary to alter the de-emphasis time constant. In the circuit of Fig.13 this is achieved with components R_X and C_X, values being given in the components list on page 259 (June) for the change from 75 to 25 μ s and for a change from 50 to 25 μ s (not yet authorized in 50 μ s countries). When recording such broadcasts the encoding function of the noise reduction unit is clearly not required and the "Dolby f.m." switch position automatically switches off the encoding function. Application of the Dolby B system to f.m. broadcasting is discussed in two articles in the *Journal of the Audio Engineering Society*, June 1973, pp.351-62, and briefly in the July 1974 issue of *Wireless World*, page 237.

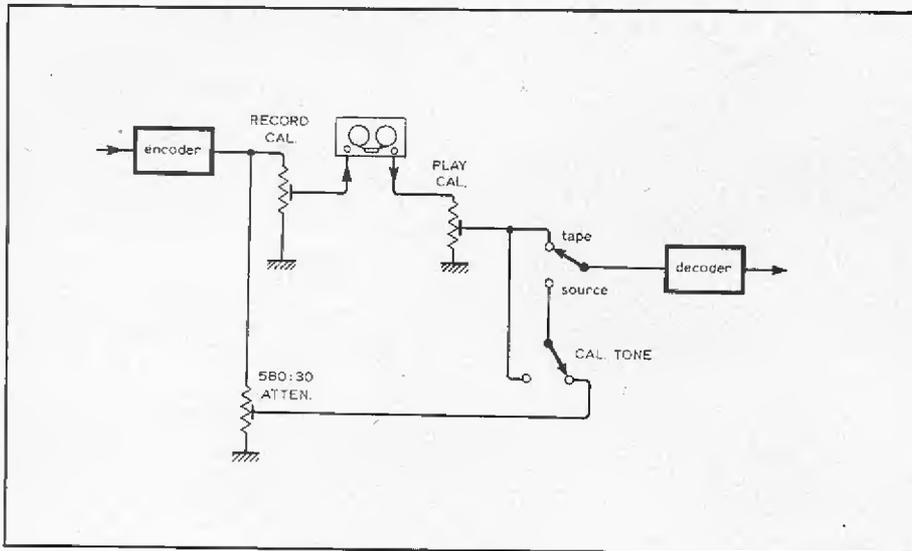


Fig. 19. Monitor switch arrangement for permanently-wired processors in three-head machines.

accurate adjustment can be obtained if a station can be received which transmits the 400Hz calibration tone, identified by a characteristic warbling, or alternatively by using an f.m. generator. In this last-mentioned case, modulation frequency should be set to about 400Hz with a peak deviation of 37.5kHz (not including pilot tone).

- Tune in to whichever of these signals is available.
- Switch to record, and to either f.m. or Dolby f.m.
- Adjust f.m. cal. control so that meter reads 580mV, or Dolby level.

Using the unit

The calibration procedures described theoretically apply to the one tape speed used during calibration. Whether the calibration will hold for different tape speeds depends on the design of the deck, so check calibration when speed is changed. The calibration tape available can be used at 4.75 and 19cm/s, as well as 9.5cm/s. (For 38cm/s tape speed, where the noise spectrum is wideband, applying the B-type system may result in the remaining mid and low-frequency noise becoming more apparent). When the brand of tape is changed it is usually necessary to readjust the record cal. controls, the play cal. setting remaining unchanged. The characteristics of cassette tapes are more critical, and changing brand will normally require adjustment of bias (and equalization when using CrO₂ tapes).

When the unit is connected to the normal input and output points of a tape

recorder, the recorder's own input and output controls from part of the calibrated system. The settings used during calibration should not be disturbed, input and output level controls being provided on the noise reduction unit, and it is a good idea to mark the tape recorder control settings.

The amplitude response of the tape recorder must be flat and its gain unity, measured between point G of the processor in record and play, to ensure correct operation, so that the signal voltage in the decoder is the same as that at the encoder (to within 2dB). If there is a bandwidth restriction between encoder and decoder, e.g. if the response of the recorder does not extend up to at least 10kHz, a non-complementary situation arises, unless of course the encoder input bandwidth is similarly limited.

In using the unit don't forget that it will only reduce noise generated after the encoder and before the decoder. If the input signal is noisy in itself or is made noisy by poor circuitry prior to encoding, this noise will be reproduced unaltered along with the signal. In some cassette decks, the line inputs are attenuated prior to amplification by a sometimes noisy microphone pre-amplifier.

As the sensitivity of the processor is of the order of 30mV, a line input amplifier is not required when the circuits are built into a tape recorder, and the input signal should be taken directly to the input gain control via a switch, or socket with switch, to disconnect the microphone pre-amplifier. It's a good idea too to make sure any automatic level limiter operates only in the microphone input and not in the line input.

Letters to the Editor

DOLBY KIT FILTER ADJUSTMENT

The use of the BBC test transmissions seemed to me to be a little hit-and-miss for setting-up the 19kHz filter of the Dolby noise reducer (July issue) since the vital zero modulation part only lasts for about two minutes. I also did not have a suitable signal generator available.

However, a little thought showed that a precise 19kHz signal was available from pin 10 of my MC1310 stereo decoder when receiving a stereo signal. Since this is the signal that the filter is required to attenuate it seemed logical to use this for alignment purposes. The signal was applied with a 2MΩ potentiometer in series and alignment was easily completed using the signal generator instructions.

There was possibly some modulation of the signal as the meter flickered slightly, but in spite of this the null was very precise.

Your readers may find this of interest to enable them to set up their kits without having to wait for Radio 3 to close down.

M. S. Maisey,
Coulsdon,
Surrey.

Wideband compander design

Simple square-law circuit gives 100dB dynamic range

by John Vanderkooy

University of Waterloo, Ontario

The wideband compander described can preserve the dynamic range of virtually any input signal when recorded by a normal tape recorder. Operational amplifiers and matched photocells allow accurate compansion with no necessity for calibration or care in recording levels. The unit can be used in compression mode for recording or playback in noisy environments, and for speech signals.

The dynamic range of tape recorders has never been adequate for high quality reproduction. If a high input level is used in an attempt to decrease the effects of tape noise, distortion results on loud passages and transients are severely distorted. Reducing the level to allow even moderate transients to be captured with little distortion means that small signals will be lost in noise. A good quality half-track reel-to-reel machine can expect a signal-to-noise ratio of about 60dB and because normal audio signals vary more than this, most recording is caught in the compromise between noise and signal distortion.

Several commercial devices are available to solve these noise problems. Dolby A systems are compressor-expanders that work in a number of frequency bands in the audio spectrum. They are very virtuous but are beyond the stage where home construction can be contemplated. The Burwen compressor-expander¹ is a device that works over the whole audio spectrum, using cube-root compression and cubic expansion, along with some equalization. Circuits for this compander were not available to the author, but the appeal of the basic system was such that a design suitable for home construction was sought and finally achieved. A recent advertisement from DBX indicates their companders may be similar to the one described here. Recently, Self² has detailed a circuit for compression only. Stuart³ has described several other active systems but not the power law compander so its basic merits will be given below.

If a compressor is designed to take an input audio signal I of large dynamic range and compress it to an output θ in accordance with the law

$$\theta = k_1 I^{1/n}$$

where k_1 is a constant and n is a positive integer (it could be fractional, but the electronics is more complicated), then the dynamic range of θ can be such that a tape recorder can faithfully record this signal. We assume that the playback signal P is equal to θ for a unity gain recorder. (There will be some error and its effect will be discussed later). Then if an expander is made which gives a final output signal S given by

$$\begin{aligned} S &= k_2 P^n, \\ \text{then } S &= k_2 \theta^n = k_2 (k_1 I^{1/n})^n \\ &= k_2 k_1^n I \end{aligned}$$

and hence except for the constants, which will vary if level controls are altered, the signal S is a scaled version of the original input I . For domestic use it is argued that $n=2$ is a good choice. An input signal of 100dB variation will be compressed to 50dB at the tape, thereby achieving a good performance with modest recorders. The $n=3$ system used by Burwen¹ shows deficiencies when used with domestic recorders having considerable variations in response with frequency. If the recorder has a response error of X dB then the final expanded signal will have an error of nX dB. A wide spectrum signal such as normal audio will relieve these difficulties, but if a 6dB variation in response exists, the cubic system is impractical.

Recent articles by Shorter⁴ on the Wireless World Dolby B noise reducer give much useful information on compansion in general and prompt a comparison between wideband compansion and Dolby B. I have always preferred the transmission of programme material by simple 2:1 logarithmic compression, rather than Dolby B methods, because the frequency response is then not altered by receivers not equipped with standard decoders. I fear the extra top will become so enticing to people that the Dolby decoders will hardly find use. In essence it boils down to a preference for distortion in level as opposed to distortion in frequency response. An interesting view of Dolby methods from the BBC recently appeared in a letter to the editor⁵.

A real advantage of the Dolby B approach is that only high frequencies are altered, and gain changes can be made so quickly that no noticeable noise modulation and breathing exist. Present-day wideband companders can partially solve these problems as well. Firstly, the attack time can be very short, so that extra pre-emphasis can be used with a consequent reduction in noise. This however, creates more incompatibility with existing components and pre-emphasis is not used in the present design. Secondly, by using special filters to eliminate self-modulation distortion, but still retaining a rapid decay-time, the effects of noise modulation and breathing are subjectively reduced. This concept is used in this design. A definite advantage of wideband compansion is the much greater degree of noise reduction for low-level signals, as will be evident later. Professional

assessments of companders and Dolby systems are given in recent reviews⁶. For the moment it is to be appreciated that wideband compansion prevents overloading the recorder, reduces the effects of noise at low signal levels, and virtually makes recording level controls unnecessary. In addition an accurate power law device will reproduce faithfully irrespective of the settings of the level controls. No reference levels are necessary as in Dolby systems or other non-linear companders.

Requirements

The heart of a compander is a gain-controlled amplifier which can divide or multiply the gain by means of a control voltage. It must be capable of 50 or 60dB gain variation with an accurate characteristic. A good audio bandwidth must be maintained over the whole variation, and the distortion should not exceed 1%. The gain variation must be rapidly programmable as well. A servo system driving a potentiometer would be accurate but too slow. A good figure to shoot for in response time is several milliseconds. This allows even transients to be respectably dealt with.

A well-built transconductance multiplier will satisfy the above characteristics, but it has too much wideband noise. This is due to the necessity of small signal levels at the bases of the multiplying transistors to prevent distortion.

As well as an accurate multiplier-divider, the circuits which caliper the audio level and produce a smooth rectified signal proportional to the amplitude must be accurate and have an attack time less than a few milliseconds. The release time should be rapid to prevent pumping but not rapid enough to cause distortion by "self modulation" of a low-frequency signal.

Experiments

Early experimental attempts at making the multiplier-divider centred on f.e.t.s and their source-drain characteristics near the origin. Distortion is high if the f.e.t. is used in a straightforward way. It can be greatly reduced if the gate is driven not only as a control voltage but as an alternating voltage which is midway between that of the source and the drain as in Fig. 1(a). This gives the device a drain characteristic of odd symmetry. Thus all even harmonics are entirely

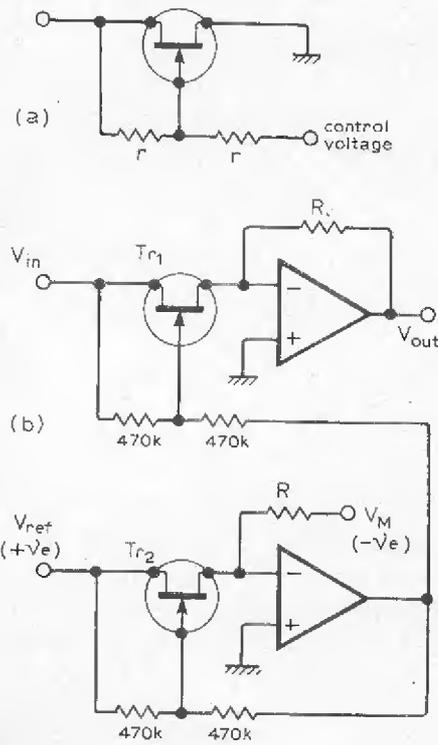


Fig. 1 (a) The distortion of the f.e.t. as a voltage-controlled attenuator can be reduced by driving the gate with an alternating voltage midway between the source and drain voltages. (b) by using a second matched f.e.t. the variation of the gain can be made proportional to V_m/V_{ref} .

the multiplier-divider. Initial experiments indicated that when a light-emitting diode was suddenly turned on the coupled photocell would respond with approximately two time constants, one a fast but rather small relative behaviour, the other a slower rise of about 10ms to a final conductance level. This is not suitable for a fast-acting gain control circuit. Also the final conductance value was not properly proportional to the i.e.d. current. Fig. 2 shows the characteristic of resistance versus current for a CL904N photocell coupled to an i.e.d. A straight line of slope -1 would represent ideal behaviour.

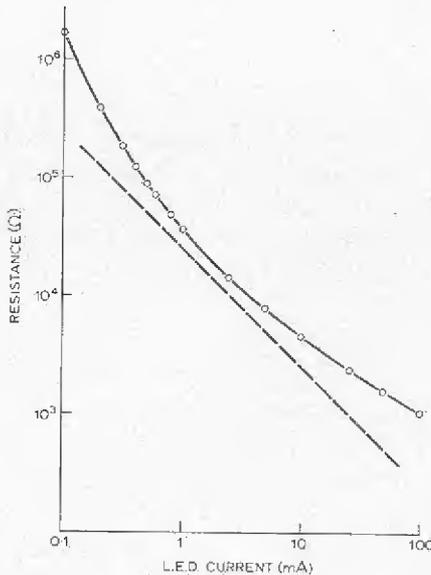


Fig. 2 Curve of photocell resistance versus i.e.d. current for a Clairex CL904N photocell when illuminated by a red i.e.d. The unbroken line is drawn through the experimental points. The dashed line represents ideal behaviour.

removed by this "push-pull" technique, and only odd harmonics, mainly third, occur at higher signal levels. The problem still remains that a large gain variation of greater than 30dB is difficult to achieve, and the gain is not a simple function of the control voltage. The last-mentioned problem can be alleviated by using one of a matched pair of f.e.t.s (a dual) to generate a specific resistance using an operational amplifier.

In Fig. 1(b), Tr_2 has a resistance which is determined by setting the input current of the op-amp equal to zero with due respect for the virtual earth, i.e. $V_{ref}/R_{f.e.t.} = V_m/R$. Transistor Tr_1 will have the same resistance and the upper circuit, which handles both polarities for audio, will obey a law $V_{out}/R = V_{in}/R_{f.e.t.} = V_m V_{in}/V_{ref} R$. Thus $V_{out} = V_{in} V_m / V_{ref}$. Division and multiplication have both been accomplished! This circuit technique must be remembered for the multiplier to be described later. It is not suitable in the present form since not enough gain variation is available.

Another method attempted was to make a transconductance multiplier using f.e.t.s as the input active elements. They would not be as linear as transistors on a relative basis but since the voltage scale on which they turn on is about a volt as opposed to the 25mV for a transistor, much less attenuation of the input signal is necessary and this together with lower f.e.t. noise would reduce the noise to small values. Disadvantages of the design are the difficulty of obtaining division and the requirement of four well-matched f.e.t.s.

Photoconductive cells were also considered as possible gain control elements for

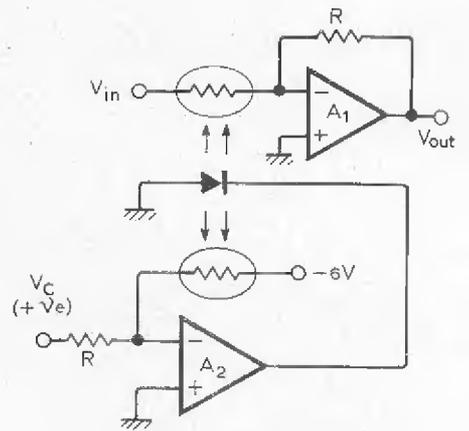


Fig. 3 Photocells and a controlling i.e.d. are used here as a multiplier using a concept similar to that shown in Fig. 1 (b). For this circuit $V_{out} = V_c V_m / 6$.

showed that a 60dB range was possible and the attack time for a large step increase in the control voltage was about one millisecond. The feedback has thus considerably reduced the sluggishness of the cell. At low light levels the cell seems to have a longer time constant and a nonlinear network placed in series with the i.e.d. maintains good control stability over all voltage levels.

Fig. 4 shows the rectifier circuit that was adopted to produce a direct control voltage for the multiplier-divider. Amplifier A_2 has circuitry which creates an absolute value circuit with a gain of 2/3. Diode D_2 is used to create a virtual earth at the inverting input for positive input signals so that the upper 5k and 10k resistors can form a simple attenuator. This diode also prevents op-amp saturation and hence allows accurate response up to the highest audio frequencies, a feature which many precision rectifier circuits do not have.

Amplifier A_3 is a peak detector in which D_4 prevents saturation of the op-amp when the input voltage from the absolute value rectifier is lower than the voltage on C_1 . Another advantage of this diode is more subtle. If a rectified sine wave of constant amplitude is fed into the peak detector, the

It was decided to employ op-amps to linearize the cells using the technique mentioned earlier. Fig. 3 shows the basic idea for a multiplier. A divider can be constructed by interchanging the resistor and the photocell as gain-determining elements for the amplifier A_1 . The i.e.d. shines equally onto both photocells. Tracking of the photocells is essential for an accurate power law comparison, but an error does not significantly affect the overall characteristic, see later. From five photocells at least two would track well over factors of 100 change in the resistance. Experiments with this multiplier-divider

Fig. 4 Schematic of the circuit used to obtain an absolute value of the audio signal and produce a control voltage proportional to the peak of the waveform.

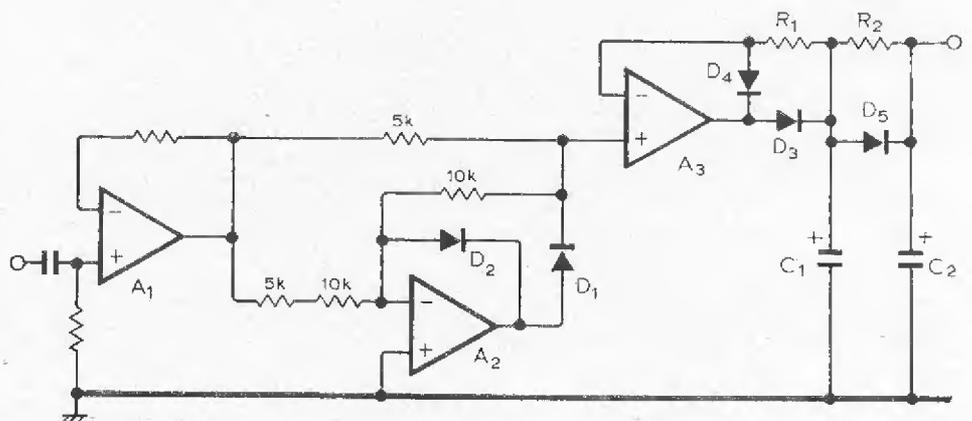


Fig. 5 Complete schematic of the compander. Op-amps are assumed to have $\pm 15V$ power supplies. For best performance amplifier A_1 should have separately decoupled supplies. The $10k\Omega$ resistor in the compensation should be referred to the negative supply.

droop at C_1 is much less than in a circuit in which R_1 is returned to ground rather than the inverting input terminal. This is so because the negative input terminal follows faithfully the input signal on the non-inverting input terminal. However, if the audio signal disappears, then R_1 is effectively returned to ground and the decay time constant is short. For audio frequencies $> 1/2\pi R_1 C_1$, the droop is only $1 - (2/\pi) \approx 0.36$ as large in this circuit as when R_1 is returned to ground. Components R_2 and C_2 provide extra filtering and D_3 allows the control voltage to rise quickly in the presence of audio transients. The input follower A_3 is necessary because the input impedance of the absolute value circuit changes with signal polarity.

Circuit description

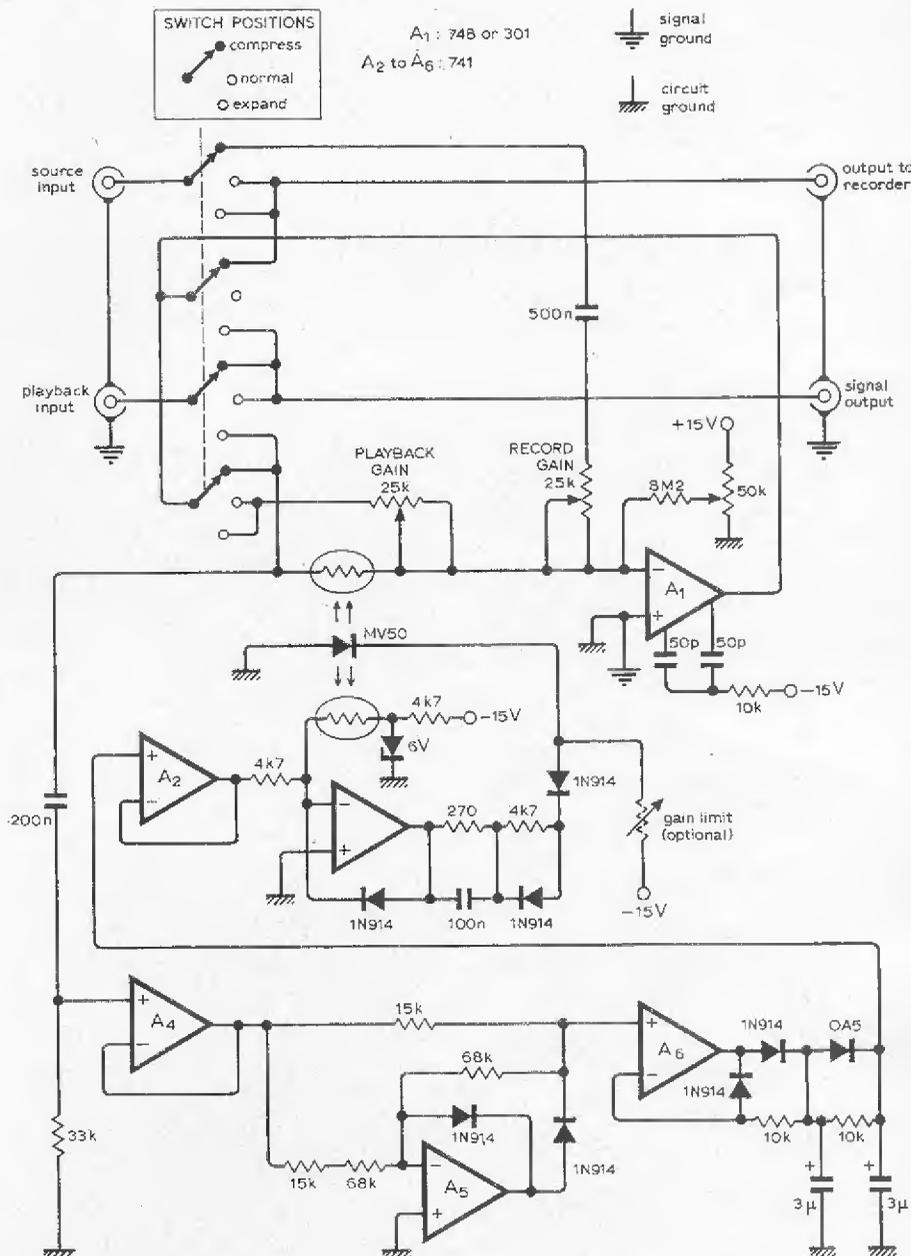
The complete compander circuit diagram is shown in Fig. 5. Switching allows the circuit to be used as a compressor during recording, and as an expander during playback. In the compression mode, the control voltage acts to decrease the audio gain as a divider. Hence the output θ will be related to the input by the relations $\theta \propto I/V_c$. But V_c is derived from the amplitude of the output signal, hence $V_c \propto \theta$. Thus $\theta \propto I/\theta$ or $\theta \propto \sqrt{I}$, a square-root compressor. When in the expansion mode, the audio gain is precisely proportional to the control voltage. Hence the final output signal S is related to the playback signal by the relation $S \propto PV_c$ but V_c is derived now from the audio signal P , hence $V_c \propto P$. Thus $S \propto P^2$ and a square-law expander results.

If the photocells do not track well, the division or multiplication factor must still be the same function of control voltage, say $f(V_c)$, because of the way the photocell is switched in the compress and expand modes. Hence $\theta = I/f(V_c)$, and if we assume again that $P = \theta$ (for a good recorder) then

$$S = Pf(V_c) = \theta f(V_c) = I.$$

Thus a perfectly complementary system still results. Careful analysis shows that this is true only if the recorder has unity gain, because otherwise the playback signal would produce a different control voltage than that used during recording. Only a power-law behaviour of the function $f(V_c)$ will preserve the relative level differences. In the present circuit $f(x) = x$, a simple function indeed.

The major circuit blocks in Fig. 5 can easily be recognised from the earlier discussion, but several features warrant special consideration. The operational amplifier A_1 used in the multiplier is used in a circuit in which the gain is varied by up to 60dB. At unity inverting gain, a compensation capacitor of about 15pF between pins 1 and 8 (half of that for a voltage follower) is necessary for stability. But this is detrimental to the frequency response when the gain is high (a small amount of feedback), as for example during small signal levels in compression. Pin 8 comes from the output circuit of the op-amp. Pin 1 is a high impedance point which has a signal referred to the negative supply line. The difficulty occurs when the



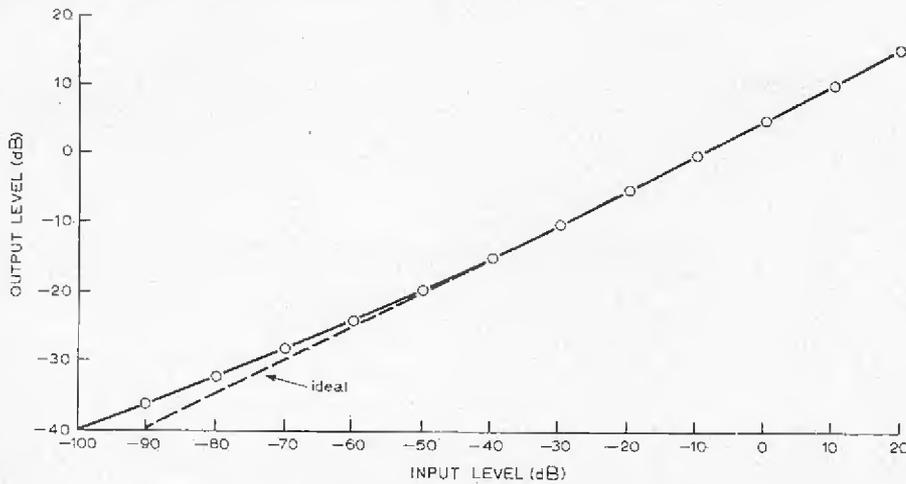
high level signal from pin 8 is injected into pin 1 through the normal compensation capacitor. The gain is drastically reduced at high audio frequencies. But there is no problem with op-amp stability at these frequencies; instability only occurs near 1MHz. The $10k\Omega$ resistor between the two $50pF$ capacitors shunts the gain reducing signal to the negative supply line thus restoring the gain at audio frequencies while not materially affecting stability considerations at megahertz frequencies.

An important point is the selection of photoconductive cells. Impedances of $1k\Omega$ are ideal for op-amp gain determining resistors. Lower values might tend to cause current limiting at high signal levels. It is therefore recommended that the photocells have resistances of not much less than $1k\Omega$ when illuminated by an i.e.d. carrying a current of 10mA. The i.e.d. can be glued to the two matched photocells (or dual photocell) with clear epoxy.

In a stereo system one has a choice of building a compander for each channel (the best solution) or of combining functions together. In a combined system it will be necessary to control three photocells with one i.e.d. If matched i.e.d.s are available then

two double units can be used. But they are not easily matched. Some require a threshold current before they start to emit light. In any event the right and left channels should be summed before peak detection. The voltage follower in the rectifier circuit can easily be rewired to act as an inverting summer. Of course two separate op-amps will be necessary with compensation as described earlier.

Due to the switching in the rectifier and peak detecting circuits, it is recommended that separate decoupling be used for the supply lines to the signal op-amp. All input and output connectors should have their signal ground connected to the non-inverting input of A_1 . You may wonder why a low-noise audio op-amp such as the Fairchild 739 was not used in the signal circuits. This is because these do not have adequate reserve gain for the multiplier - divider action necessary here. They also draw more input current, causing greater offsets when the gain is high. It is wise to include the input offset current adjustments for the signal op-amps as shown in Fig. 5. The offset is adjusted to provide zero direct output voltage with a very low signal input in the compress mode. If this is not done a low frequency thump will occur when the gain



changes quickly.

Another point is that for no signal level in the compress mode, the gain is very large and is limited mainly by the rectified output noise. This is not usually a problem since most sources for recording in the home such as discs and microphone arrangements have enough background noise. An easy way to eliminate such problems is the inclusion of a resistor shown dotted in Fig. 5 which limits the maximum gain by preventing the l.e.d. current from becoming zero.

As shown, the compander responds to very low frequency signals and has low phase shift. Sometimes a turntable can have a large low frequency rumble which can modulate the compressor gain. In such cases a filter should be used to remove such low frequencies. A simple solution is to decrease the value of the $0.5\mu\text{F}$ record input coupling capacitor to give an appropriate cut off frequency. If a recorder with restricted bandwidth is used, it is wise to restrict the input to the compander to the same extent. This ensures that the rectifier circuits will see similar signals on compression and expansion.

The power supplies should be well regulated for optimum performance, but unregulated supplies with good ripple filtering are acceptable. The transformer should supply 11V a.c. on open circuit and allow 70mA of current drain.

All diodes can be silicon signal diodes, such as 1N914, 1N4148, 1S44; only D_3 in Fig. 4 should be a germanium signal diode as this will help reduce overshoots in compression.

Input impedances are simply given by the values of the record and playback preset potentiometers. The outputs are low impedance, and perhaps 560Ω resistors should be added in series with these outputs to prevent damage if high signal levels are inadvertently applied to these outputs.

Performance

The most important characteristics are the compression and the accuracy of the whole process. Fig. 6 shows the graph of output level versus input level in the compression mode of operation. The deviation of the curve at very low input levels is due mainly to photocell tracking error, and partially from the amplified noise of the 748 op-amp. The input voltage is not measured at these low levels; it is inferred from the settings of an accurate low impedance attenuator. Deviation from a square root behaviour is never more than 1dB for well over 80dB of dynamic range. All levels are in dBm ($0\text{dBm} = 0.775\text{V r.m.s.}$).

Fig. 6 Curve indicates output level versus input level when the compander is in the compression mode. Note that a 60dB input variation is compressed to 30dB of output variation to be recorded on the tape.

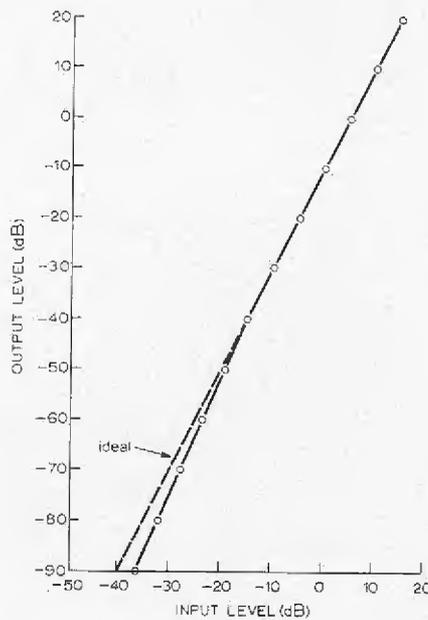


Fig. 7 Output level versus input level in decibels for the expansion mode. A 30dB input variation from the recorder on playback would be expanded to an output of 60dB to be fed to a power amplifier.

Fig. 7 shows the output level versus input level in the expansion mode. The curve again shows almost no discernable deviation from a square law. Output levels are difficult to measure with standard a.c. voltmeters below about -90dB . The complete characteristic from recording input to final signal output is linear to much better than 1dB because of the exact complementarity discussed earlier.

Even the dynamic characteristics are precisely complementary, because the audio signal used to produce the control voltage is derived from the output in compression

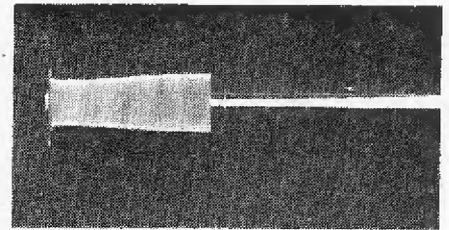


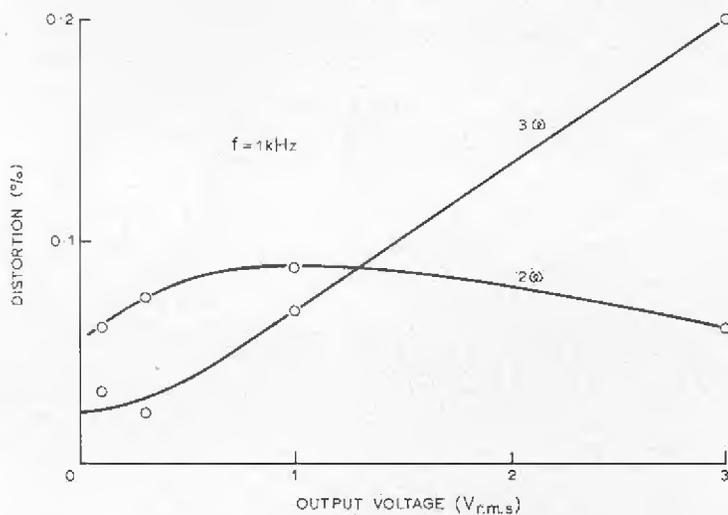
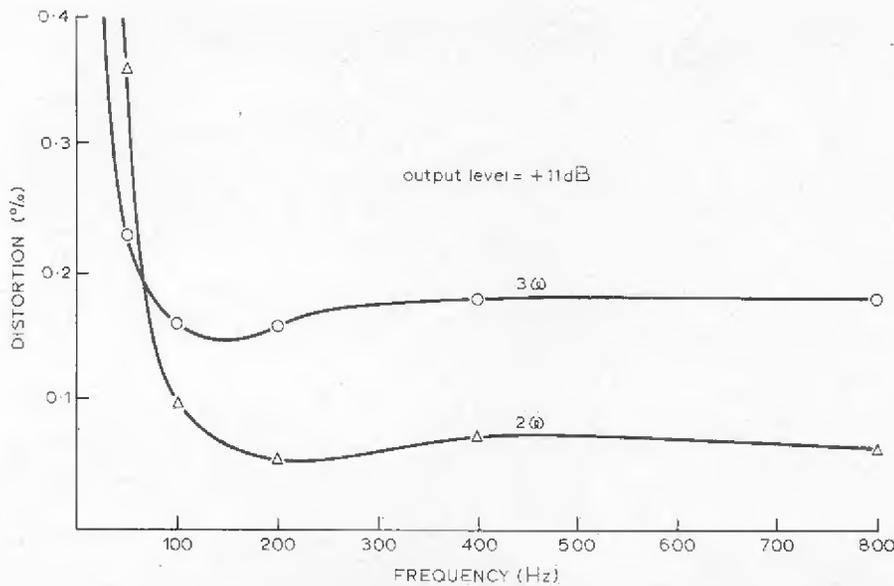
Fig. 8 Oscillograph of the compressor output when the 1kHz input signal is suddenly increased by about 30dB by a mercury wetted relay. (20ms/div horizontally, 2V/div vertically.)

mode, and from the playback input signal in expansion mode. These two signals should be the same for a good tape recorder. An overshoot in the compression mode, which is very difficult to suppress completely because of the time constants of the photocells and the peak rectifier, will not be exactly undone in the expansion mode. Only the leading edge of a transient sound will perhaps not be faithfully recorded, but the ear will forgive severe distortion for periods of several milliseconds. Fig. 8 shows the output signal to the recorder in compression mode when the input signal is suddenly increased by about 20dB. The signal frequency is 1kHz. Notice that there is a slight overshoot in the compression that lasts about 10ms. The transient edge dies away with a time constant of about a millisecond. There is some dependence in Fig. 8 on the phase of the input signal at the moment of switching in the higher level. One would expect this in fast-acting circuits. A real audio transient is likely to be less severe than the instantaneous switching used here as a test signal.

The release time constant is less than a tenth of a second, giving a fast enough action that even on a rapid reduction in signal level, no noise is noticeable on replay. The rapid release time is also advantageous if the compressor is used on the output of an automobile radio. The normally large variations in signal level will be reduced so that low levels are not masked by the ambient noise. (I have often wanted something akin to an engine-speed dependent volume adjustment on my automobile radio.)

For high fidelity purposes the compander must have low distortion. Fig. 9 shows the measured second and third harmonic distortion versus frequency in compression mode for an input level of $+10\text{dB}$. The rise at low frequencies is due to the ripple from the peak rectifier. The wideband distortion is due to the photocell characteristic and is mainly third harmonic.

Fig. 10 shows the second and third harmonic distortion versus the output signal level (the voltage across the photocell) at a frequency of 1kHz. Except at high output level, the distortion quickly falls near the noise limit of the wave analyser. There is approximately 0.05% of residual third harmonic distortion in the oscillator which may slightly raise or lower the measured third harmonic, depending on phase relationships. Distortion level is low enough because it does not represent a crossover distortion, only a gently curving transfer characteristic. However, it would be unwise to be too defensive



about distortion levels in photocell circuits. Some cells have much larger distortion than others. A number of different types were tried. I admit I could not hear the difference, but measured distortions of up to 2% at high level were occurring for some cells. The circuit whose measured low distortion is shown in Fig. 9 and 10 uses quite inexpensive cells, type VT-833, manufactured by Vactec Incorporated.*

Any reasonably fast CdSe photocell could be used with resistance characteristics as described earlier. A quick check on distortion can be made by applying 10V r.m.s. at 1kHz to a divider made up of 10kΩ resistor and the cell, illuminated to a resistance of about 10kΩ. If no appreciable curvature exists on an X-Y oscilloscope display or cell voltage versus oscillator voltage, the cell has suitable characteristics.

In Fig. 11 the clipping characteristics of the compander are shown in compression and expansion modes. The break point is due to current limiting of the amplifier A_3 in Fig. 5 which drives the i.e.d.

The final test of performance of an audio circuit must be the human ear. In microphone arrangements using the compander there is dead quiet at no signal level. This is far from true without the compander. Replay sounds natural and the settings of the level controls on either recording or playback are

Fig. 10 Distortion of the compressor versus output voltage level at a signal frequency of 1kHz. There is a residual distortion of about 0.05% third harmonic in the oscillator which may alter the third harmonic results somewhat at low levels.

unimportant as long as overload is prevented. Using good discs as a source there is no noticeable difference in the dynamics even on piano music when the compander is used. This is impressive performance for such a simple circuit.

How can the audio enthusiast use the compander? If his tape recorder has a signal-to-noise ratio not much worse than the sources at his disposal, then it is hardly worthwhile using it to preserve dynamic range. However, modern stereo cassette recorders have signal-to-noise ratios of around 50dB, whereas a live f.m. broadcast can have 70dB. Then 20dB of increased dynamic range will result. In a live microphone setup with a low noise preamplifier the increase in dynamic range is greater than 40dB, and here the compander allows almost complete disregard of the level controls.

If master tapes and discs were made with a square root compressor and radio stations would broadcast these directly, then an expander in the receiver could bring back the full dynamic range of the original signal. Another use for the compander occurs whenever there is a high background noise

Fig. 9 Distortion of the compressor versus frequency for an input level of +10dB. Curves are substantially constant beyond 800Hz with a small increase beginning beyond 10kHz.

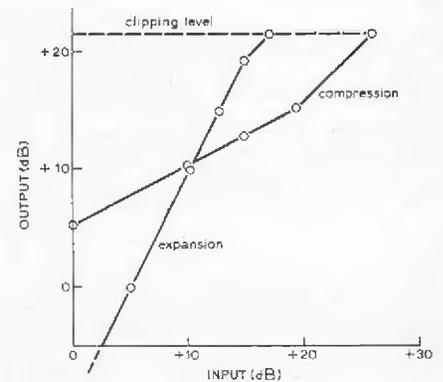


Fig. 11 Characteristics of the compander for high level signals. Clipping level in the compression or the expansion mode is determined by the supply voltage of the operational amplifiers. Deviation from square-root compression and square-law expansion results from the current limiting of the amplifier A_3 driving the i.e.d. These levels were obtained by setting the record and playback present potentiometers to 10kΩ. Altering these values will alter the point at which the behaviour saturates.

level, such as in an automobile, workshop, or a home with children. The unit can be used to process a signal using compress mode, so that the dynamic range of the signal stays sensibly above the noise level. It is wise to include the dotted resistor of Fig. 5 in such setups to reduce the noise output when the signal level is very low.

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*U.K. agents Teknis Ltd, Teknis House, Meadow, Godalming, Surrey GU7 3HQ. The cells cost around £1.

High-quality compressor/limiter

A variable law, low distortion attenuator incorporating second harmonic cancellation circuitry

by D. R. G. Self, B.A.

University of Sussex

Compression and limiting play an increasingly important role in the resources of a modern sound studio. The conventional function of signal level control is to avoid overload, but it can be used in the realm of special effects. To date, however, relatively few designs for high-fidelity compressor/limiters have been published.

The main design problem is the voltage-controlled attenuator, v.c.a., which increases attenuation of the input signal in response to a voltage from a control loop as shown in Fig. 1. In limiting, this circuit block continuously monitors the peak output level from the v.c.a. and acts to maintain an almost constant level if it exceeds a threshold value, or, in compression, allows it to increase more slowly than the v.c.a. input signal. This is illustrated in Fig. 2., which shows the input-amplitude/output-amplitude characteristic for both compression and limiting. Note that limiting makes use of a much tighter slope to ensure that the output voltage cannot exceed the chosen limit, and that the threshold (point of onset of attenuation) takes place at a higher level than for compression.

Traditionally, studio-quality compressor/limiters (as the two functions are so similar it is logical to produce a system that can be used for either compression or limiting) used one of two types of v.c.a. Either the audio signal was chopped at an ultrasonic frequency by a variable mark/space square wave — which requires complex circuitry and careful filtering of the audio output to avoid beats with tape-recorder bias frequencies — or it was attenuated by an electronic potential divider, one arm of which was a photoresistor, the control signal being applied via a small filament bulb. The last-mentioned has disadvantages because photoresistors are non-linear devices, therefore noticeable distortion is introduced into the audio signal, and the thermal inertia of the bulb filament limits the speed of attenuation onset.

Most modern compression systems use field-effect transistor operated below pinch-off as a voltage-variable resistance in a potential divider. This

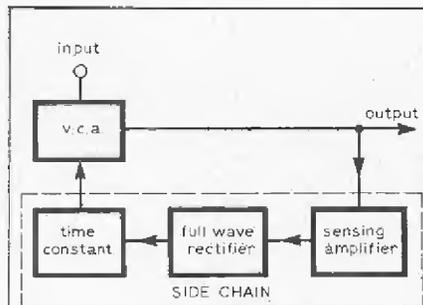


Fig. 1. Voltage-controlled attenuator with d.c. control loop.

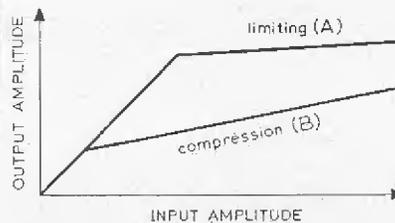


Fig. 2. Amplitude characteristics for compression and limiting—the last-mentioned uses an almost zero slope to prevent the output exceeding a preset level.

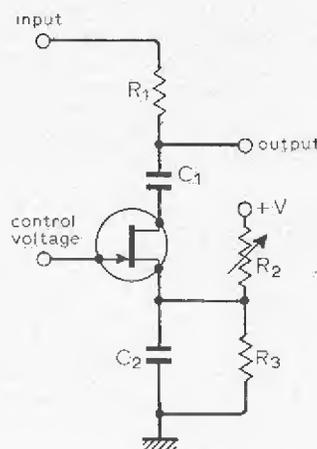


Fig. 3. Basic v.c.a. circuit providing up to 45dB of attenuation. This configuration introduces second-harmonic distortion which is greatest at 6dB of attenuation.

technique has many advantages; it is a simple, cheap, and fast-acting configuration that can provide an attenuation variable between 0 and 45dB. The only problem is that an f.e.t. is a square-law device, and tends to generate a level of second-harmonic distortion that increases rapidly with signal amplitude. A typical arrangement is shown in Fig. 3 — R_2 , R_3 and C_2 allow the source of the f.e.t. to be set at a d.c. level above ground, so that a control-voltage that moves positive with respect to ground can be used, to avoid level-shifting problems in the control loop. This d.c. level is isolated from the input and output by C_1 .

The distortion introduced by this circuit is at its worst for the 6dB attenuation condition, because at this point the drain-source resistance equals R_1 , and the maximum power level exists in the f.e.t. Table 1 shows the level of second-harmonic distortion introduced into a sine-wave signal of 100mV r.m.s. amplitude, under the 6dB attenuation condition, for three different f.e.t. types. Measurements were made with a Marconi TF2330 wave analyser, higher-orders of harmonic distortion proved to be negligible amplitude in all cases. These measurements were made on one sample of each type of f.e.t. and, because production spreads are large, the results should be treated with some caution. However, it is clear that these levels of distortion are unacceptable for high-quality applications.

Fortunately, a technique* exists for reducing f.e.t. distortion to manageable levels, if the control-voltage is applied to the f.e.t. gate and summed with a signal consisting of one-half the voltage, from drain to source, then the distortion level is dramatically lowered. The configuration in Fig. 4 shows a simple way of realising this; the signal fraction fed back is not critical and 10% resistors can be used for R_4 and R_5 . Surprisingly, this distortion cancellation procedure leaves the attenuation/control-voltage characteristic almost unchanged. Table 1 shows the new maximum distortion values for 100mV r.m.s. input. (Note that the maximum no longer occurs at 6dB attenuation, but at a point that

varies with the f.e.t. type, where cancellation is least effective.) From these results the 2N5457 and 2N5459 are superior, the 2N5459 was used in the final version of the v.c.a.

To determine appropriate signal levels in the v.c.a., measurements were made of maximum distortion generated, ie the v.c.a. was set to 2dB attenuation, against r.m.s. input voltage; results are shown in Table 2. The question now arises as to whether this distortion performance is adequate for a high-quality compressor/limiter. There is no general agreement as to the amount of second harmonic distortion that can be introduced into a program signal before it becomes aurally detectable, but 0.1% is a figure that is quoted. This means that the permissible input voltage to the v.c.a. would be restricted to below 100mV r.m.s. In practice, however, the attenuation level will be constantly changing, and because distortion level peaks fairly sharply with attenuation change, this level of distortion will only be present for a very small percentage of the time. In any case, second harmonic distortion alone has a relatively low "objectionability factor". The proof of the pudding is in listening to the compressor output signal; inputs of music around 200mV r.m.s. produced no trace of audible distortion. (Good class A power amplifiers and headphones were used for monitoring).

The control loop consists of an amplifier which senses the v.c.a. output level. A full-wave rectification system is normal practice because program waveforms have positive and negative peaks that can vary by as much as 8dB, and an 8dB uncertainty in the output level is usually unacceptable. A time-constant arrangement is used with the rectification circuit to control the attack and decay rates.

The output sensing amplifier in the system is a non-inverting op-amp which allows a high input impedance because the output impedance of the v.c.a. stage reaches a maximum of about 39k Ω at zero attenuation. The full-wave rectification system consists of a transistor phase-splitter driving two op-amp precision-rectifier stages in antiphase. The principle of a precision rectifier is illustrated in Fig. 5. The rectifying element is placed in the feedback loop of an op-amp, so that the effect of the forward voltage drop on the output voltage is divided by the open-loop gain. During positive half-cycles, if the input voltage exceeds the d.c. level stored on the capacitor C, the op-amp output swings positive and C is charged through diode D until its stored voltage is equal to the input voltage. Thus C takes up a voltage across it equal to that of the positive peak of the input signal. During negative half-cycles, and while the input is less than the voltage on C during positive half-cycles, the op-amp saturates negatively and D remains firmly reverse-biased. Obviously this is only a half-wave rectification circuit, the

Table 1. Second-harmonic distortion level introduced into a sine-wave of 100mV r.m.s.

Device	2N3819	2N5457	2N5459
2nd harmonic at -6dB	13%	10%	8.9%
2nd harmonic with cancellation attenuation shown	0.39% 2dB	0.12% 10dB	0.12% 2dB

Table 2. Maximum distortion generated by various input voltages at 2dB attenuation.

Input (mV, r.m.s.)	2nd harmonic (%)
20	0.005
50	0.10
100	0.12
200	0.19
500	0.34
1,000	0.56

Table 3. Prototype calibration data and compression ratios

VC ₂ (V)	Threshold (mV, pk)	Compression ratio
2.9	10	2.3
3.5	20	5.1
5.0	50	10
6.7	100	20
8.5	200	35
9.8	500	50

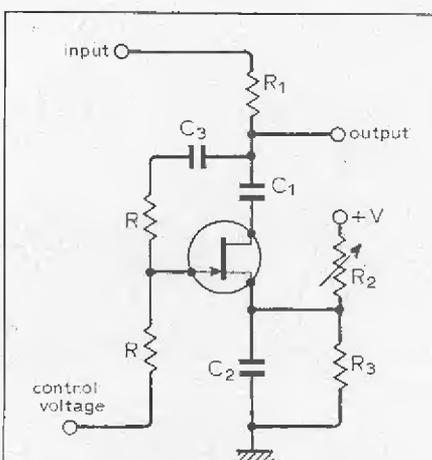


Fig. 4. Standard circuit technique for reducing f.e.t. distortion by summing half of the drain/source voltage with the control voltage.

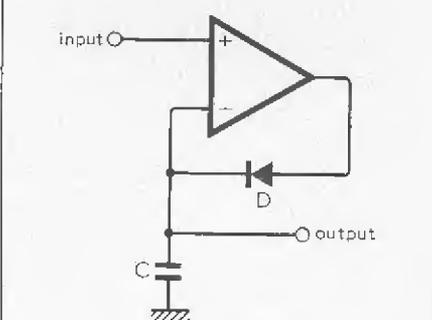
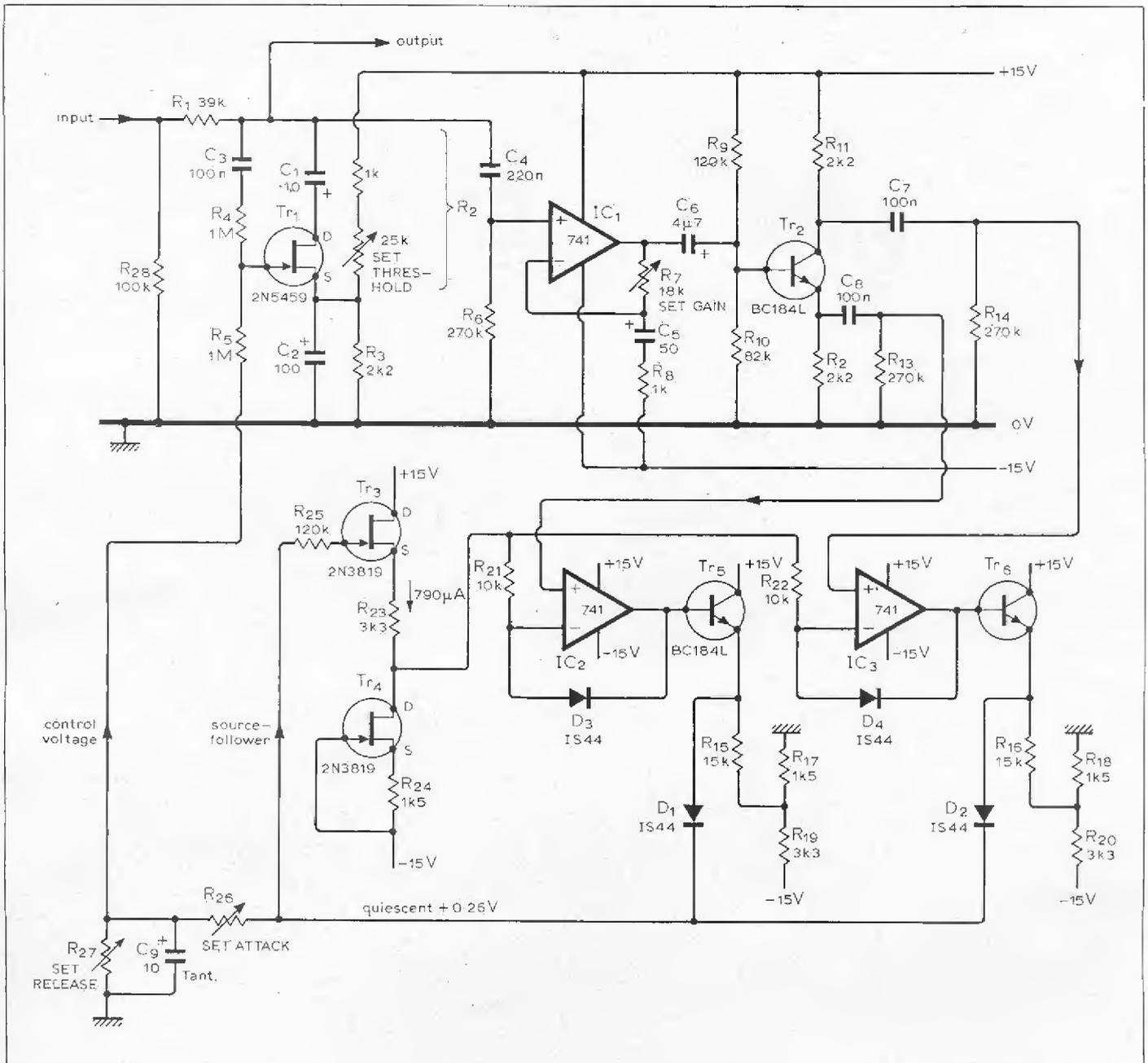


Fig. 5. Basic precision rectifier circuit where the rectifying element is in the feedback loop of an op-amp.

full-wave version uses two of these driven in antiphase, and charging a common capacitor. A resistance through which the charging currents flow determines the attack time, and another in parallel with C defines the decay time-constant.

The complete circuit is shown in Fig. 6. The v.c.a. is essentially as described above and the attenuation threshold is set by the variable resistance R_2 . As the resistance is increased the level of control voltage required for attenuation to begin is reduced, and the system's input/output characteristic moves smoothly from A to B on Fig. 2. The threshold decreases and the compression slope becomes less flat as the system turns slowly from a limiter into a compressor by the manipulation of a single control. The output sensing amplifier consists of IC₁ and has a gain of 19 over the audio band. This is rolled off to unity at d.c. by C₅. Transistor Tr₂ and its associated components form a conventional phase-splitter driving IC₂ and IC₃, the precision rectifiers. The rectifier circuitry is more complex than implied above, three modifications have been made to improve the performance. Firstly, IC₂ and IC₃ charge C₉ via current amplifier stages Tr₅ and Tr₆ otherwise the current-limited 741 outputs would be unable to provide enough current for the faster attack times (less than 1mS). Secondly, the feedback loop from C₉ to the inverting inputs of IC₂ and IC₃ is completed via a f.e.t. source-follower. Without this, C₉ would be loaded by the two 741 inputs, and this would severely limit the maximum decay times available. Incorporating the source-follower allows decay times of several minutes by using large resistance values for R₂₇. The conventional source-follower has a large negative offset voltage and is unusable in this application because due to their rectifying action IC₂ and IC₃ are unable to provide a voltage on C₉ that is negative of ground. This would be required to allow the source-follower output to be at ground when there is no input to the rectifiers. However, if a modified source-follower is used, with a constant-current source and resistance combination in the source circuit, the offset voltage can be varied on either side of zero by manipulation of R₂₄ which varies the driving current. The offset voltage is arranged to be plus 0.3V, to allow a large safety margin for thermal variations, component ageing, etc. This means that under no-signal conditions C₉ takes up a standing quiescent voltage of plus 0.3V. The effect of this is taken up in the calibration of R₂.

The third modification is the addition of R₂₁, D₃ and R₂₂, D₄. These two networks prevent IC₂ and IC₃ from saturating negatively, during negative half-cycles of their input voltage, by allowing local negative feedback through D₃ and D₄. This limits the negative excursion of the IC outputs to



about two Volts. The prevention of saturation is necessary because the recovery time of the 741s causes the frequency response of the precision rectifier circuit to drop off at about 1kHz. The addition of the anti-saturation networks provides a frequency response that starts to fall off significantly above about 12kHz which is ample for our purposes as program signals have very little energy content above this frequency.

The final part of the circuit defines the attenuation time constants. Resistor R_{26} sets the attack time constant and R_{27} the decay time constant; these can range between zero and $1\text{M}\Omega$ ($220\mu\text{s}$ and 10s) for R_{26} , and $1\text{k}\Omega$ and infinity (10ms and 20min) for R_{27} . They can be either switched or variable resistances, depending on the range of variation required.

The circuit in Fig. 6 shows the compressor output being taken directly from the v.c.a. This is only suitable if the minimum load to the output is greater

Fig. 6. Complete circuit where the output is taken directly from the v.c.a.—this may be buffered for loads greater than $100\text{k}\Omega$.

than $100\text{k}\Omega$, otherwise the v.c.a. attenuation characteristic will be distorted by excessive loading. If lower resistance loads are to be driven a buffer amplifier stage must be interposed. The IC_1 amplifier stage is suitable for most applications, and its gain is $(R_7 + R_8)/R_8$. For the unity gain case R_8 & C_5 can be eliminated and R_7 replaced by a direct connexion.

The compressor should be driven from a reasonably low impedance output (less than $5\text{k}\Omega$).

Construction is straightforward; the layout is not critical and the prototype was assembled on 0.1in matrix Veroboard. To set up the circuit R_{24} is adjusted so that the voltage across C_9 is about plus 0.3V with no signal input.

The value required will vary due to production spreads in the f.e.t.s. To calibrate R_2 it is necessary to relate the level of input signal at which attenuation commences, with the voltage across C_2 . This can be done with an oscilloscope, or preferably an a.f. millivoltmeter. As a guide the calibration data for the prototype is shown in Table 3, along with the values of the compression ratio (number of dBs the input must increase by to increase the output by 1dB). This data must be regarded as only a guide. It is worth noting that as the controlling factor setting the compression/limiting function is the voltage across C_2 R_2 could be replaced by a $1\text{k}\Omega$ resistor connected to a remote voltage source.

The compressor/limiter is quite straightforward in use, provided a few points are kept in mind. Firstly, if it is being used in the limiting mode to prevent overload of a subsequent device, the fastest possible attack time should be used, to catch fast transients, and a

fast decay time (say 100ms; $R_{27} = 10k\Omega$), to allow the system to recover rapidly when the transient has passed. Secondly, if a noisy programme signal is being compressed a long decay time should be employed, otherwise the noisy background will be faded up during quiet passages, and the familiar compressor "breathing noises" will be heard. Finally, signals with a large v.l.f. content should be avoided or filtered, otherwise v.l.f. modulation of the signal will result, if a fast decay time is in use.

If a stereo compressor/limiter is constructed from two of the systems described above it is necessary to gang together R_{26} , R_{27} and R_{28} between the two channels. A direct connexion between the non-grounded sides of the two C_9 s is also needed. It might be necessary to select matched f.e.t.s to avoid stereo image shift during compression, due to differing attenuation characteristics in the two v.c.as. A well-smoothed p.s.u. providing $\pm 15V$ should be used to power the compressor/limiter.

Components list

IC _{1, 2, 3}	741	
Tr _{2, 5, 6}	BC184L or equivalent	
Tr ₁	2N5459	
Tr _{3, 4}	2N3819	
D _{1, 2, 3, 4}	1S44 or low-leakage equivalent	
R ₁	39k	
R ₂	25k variable, with 1k in series	
R ₃	2.2k	
R _{4, 5}	1M	
R ₆	270k	
R ₇	18k	
R ₈	1k	
R ₉	120k	
R ₁₀	82k	All resistors
R _{11, 12}	2.2k	(except R ₂) 1/4W
R _{13, 14}	270k	
R _{15, 16}	15k	
R _{17, 18}	1.5k	
R _{19, 20}	3.3k	
R _{21, 22}	10k	
R ₂₃	3.3k	
R ₂₄	see text	
R ₂₅	120k	
R _{26, 27}	see text	
R ₂₈	100k	
C ₁	10 μ F 25V electrolytic	
C ₂	100 μ F 25V electrolytic	
C ₃	100nF 250V polyester	
C ₄	220nF 250V polyester	
C ₅	50 μ F 40V electrolytic	
C ₆	4.7 μ F 40V electrolytic	
C _{7, 8}	100nF 250V polyester	
C ₉	10 μ F 16V tantalum bead	

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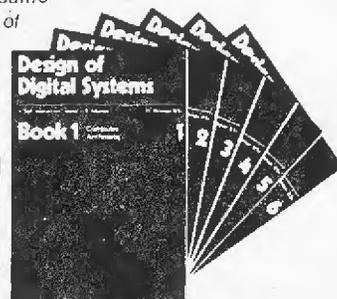
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An automatic noise-limiter

A simple muting circuit for use with f.m. tuners

by P. Hinch, B.Tech.

In recent years the automatic noise limiter has become an increasingly common addition to high quality f.m. receivers. Such a circuit greatly simplifies tuning of the receiver by selecting a minimum signal level below which the audio output is muted. Apart from the removal of inter-station noise, a squelch circuit can also ensure that only the local transmitters of the national stations are received. With high sensitivity tuners (such as the Nelson-Jones design¹), it is not always immediately apparent when the "wrong" transmitter is being received, until the poorer signal-to-noise-ratio becomes evident. A further bonus is the removal of tuning ambiguities in the absence of a.f.c., caused by the shape of the discriminator response curve; a high level, distorted signal is received on either side of the true signal due to the i.f. falling on the wrong slope of the discriminator response.

The usual method of achieving the a.n.l. function is to detect amplitude modulation of the i.f. after limiting. If noise is being received, the i.f. amplitude occasionally drops to zero due to noise cancellation. These gaps in the i.f. waveform can be

detected, and used to operate the muting circuit. However, in a circuit designed to be an add-on unit for existing tuners, it was considered undesirable to make connections into the i.f. strip of the receiver. The circuit described requires no modifications to the tuner, except, in the case of monaural reception, removal of the de-emphasis capacitor.

The circuit relies on the fact that, while the signal bandwidth is restricted to a maximum of 53kHz (for stereo signals), the noise bandwidth extends to over 100kHz. A third order high pass filter is used to reject the signal and yet allow noise to pass through. The resultant signal is amplified and detected, so producing a d.c. output related to the amount of noise being received. This is used to operate an f.e.t. switch, which mutes the output of the receiver. For mono reception, provision is made for adding a de-emphasis capacitor at the output.

Circuit description

The first stage is an emitter follower designed to provide a high input impedance which is substantially constant with

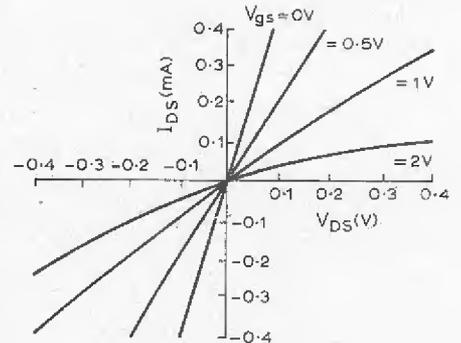


Fig. 2. A sketch graph of the f.e.t. $V_{DS} = I_{DS}$ characteristics.

frequency. This is important in order to avoid amplitude and phase distortion of the stereo multiplex waveform when fed from a receiver having an appreciable output impedance. The input capacitor to the emitter follower has a value of 68pF, giving a first order high-pass characteristic with a cut-off frequency of 100kHz. The variation in amplitude at the input when fed from a source impedance of 2.2k Ω

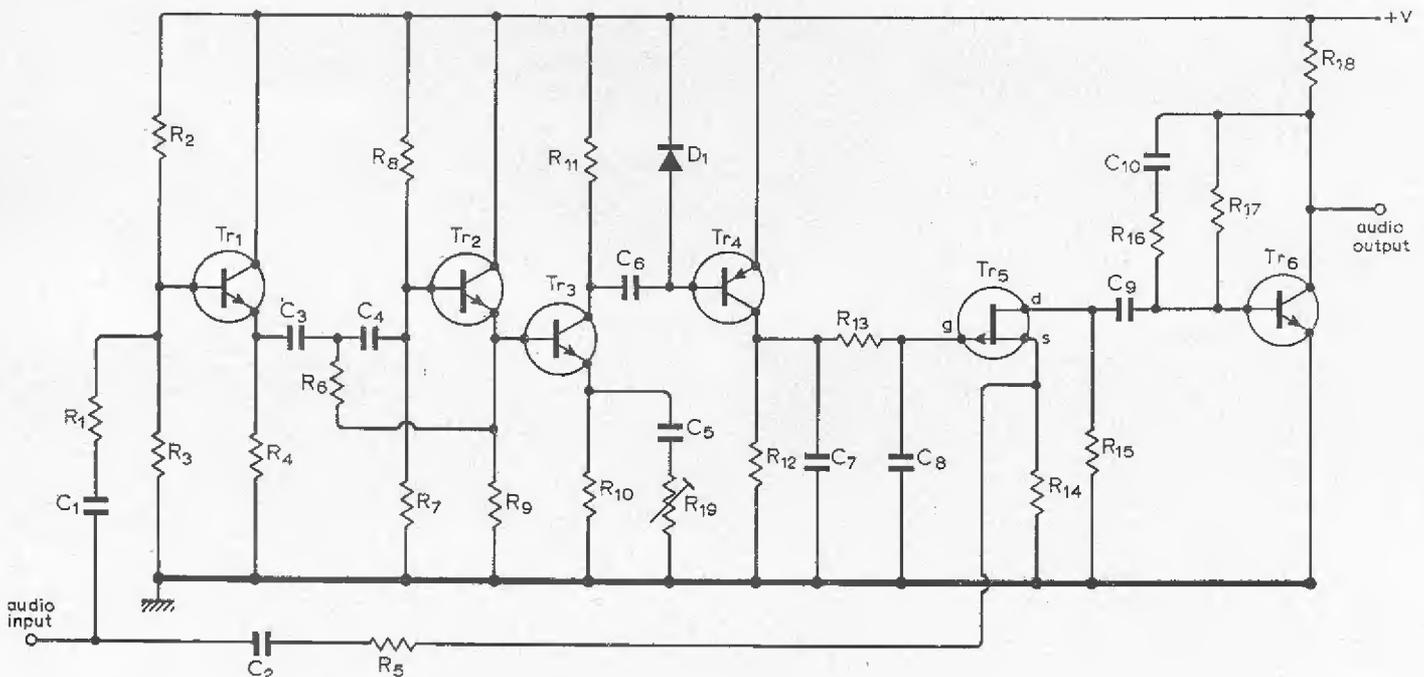


Fig. 1. Complete circuit of automatic noise limiter.

(as in the Nelson-Jones design) is then only 0.3dB from 1 to 53kHz.

The second stage is a Sallen-Key type second order high pass filter with a cut-off of 100kHz, presenting a low impedance drive to the voltage amplifier stage (Tr_3 in Fig. 1). The detector Tr_4 switches when the amplifier output reaches about 1.4 volts peak-to-peak. The detector output passes through a low pass filter (R_{13} , C_4) which prevents accidental muting caused by brief noise spikes on an otherwise low noise signal (for example, those caused by badly suppressed car ignition systems). The muting action is performed by a p-channel junction f.e.t. used as a switch.

Design of the f.e.t. switch

If an f.e.t. is operated under conditions of low gate-source voltage and low drain-source voltage, it acts as a linear resistance, the value of which is controlled by the gate-source voltage (see Fig. 2). For the 2N3820 device used in this design, the minimum "on" resistance is typically around 400Ω . In order to avoid distortion it is clear that, in the "on" state, the drain-source signal voltage must be kept to a minimum, as also must the gate-source signal voltage. If either of these is allowed to rise, the drain-source resistance will vary over the cycle, and distortion will be generated. Thus an f.e.t. switch as shown in Fig. 3 was found to generate 0.5% distortion at 0.5V r.m.s. input. For higher input levels the distortion increased drastically. This was considered unacceptable for high quality reproduction.

The solution to this is to connect the f.e.t. (Tr_5) to the virtual earth point of a feedback amplifier, as shown in Fig. 1. At this point, signal levels are very low.

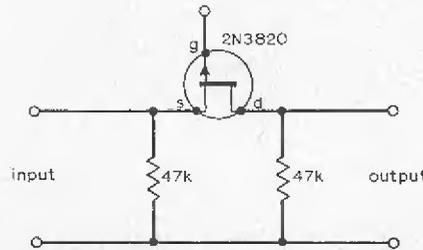


Fig.3. Elements of the f.e.t. switch used to control the receiver output.

In this circuit, distortion was found to be 0.03% at 53kHz, and 0.5V r.m.s. input. The distortion was almost entirely second harmonic, and at low frequencies the level was reduced still further. The attenuation in the "off" state was found to be -60db relative to 0.5V r.m.s.

This design has the added advantage that de-emphasis can be added for mono reception, by connecting a 2.2nF capacitor across the base and collector of the transistor.

Constructional Details The layout is not particularly critical, but long leads should be avoided, especially to the base of Tr_1 . It is, of course, important to remember to remove the receiver de-emphasis capacitor if one was fitted for mono reception. In the case of the Nelson-Jones tuner the designer recommends replacing this component with 150pF.

Performance The circuit has been in use for some time in the author's Nelson-Jones tuner. It has proved to be highly immune to transient interference, and greatly simplifies tuning of the main national and local

Components list

Resistors:

R_1 1k Ω	R_{10} 2.2k Ω
R_2 47k Ω	R_{11} 4.7k Ω
R_3 47k Ω	R_{12} 47k Ω
R_4 2.2k Ω	R_{13} 220k Ω
R_5 22k Ω	R_{14} 47k Ω
R_6 10k Ω	R_{15} 47k Ω
R_7 15k Ω	R_{16} 22k Ω
R_8 39k Ω	R_{17} 560k Ω
R_9 2.2k Ω	R_{18} 2.2k Ω

All 5% carbon.

Potentiometer:

R_{19} 1k Ω lin. preset

Capacitors:

C_1 68pF silver mica	C_6 10nF
C_2 1 μ F	C_7 10nF
C_3 150pF silver mica	C_8 100nF
C_4 150pF silver mica	C_9 1 μ F
C_5 100nF	C_{10} 470nF

All capacitors except C_1 , C_3 , C_4 , may be 20%. C_1 , C_3 , C_4 should be 5%.

Transistors:

Tr_1 to Tr_3	2N930
Tr_4	BC214L
Tr_5	2N3820
Tr_6	2N930

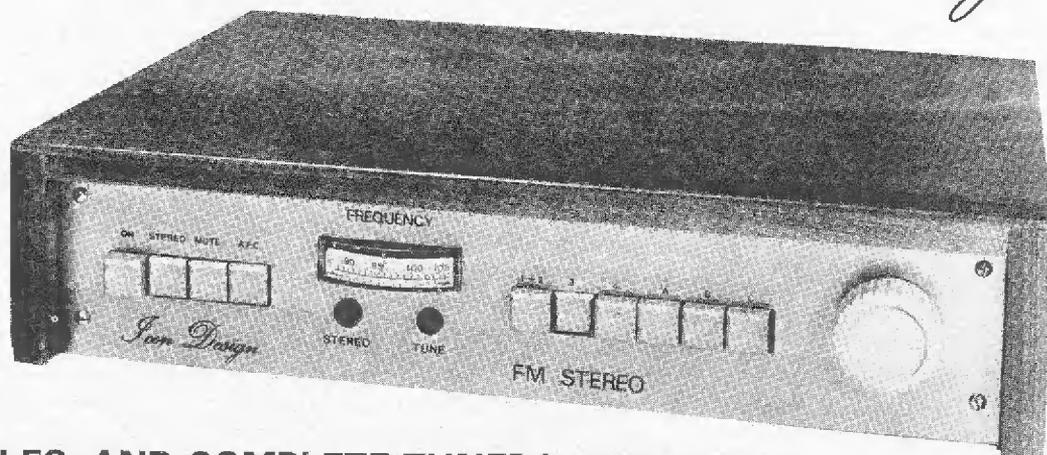
Diode:

D_1 1N914

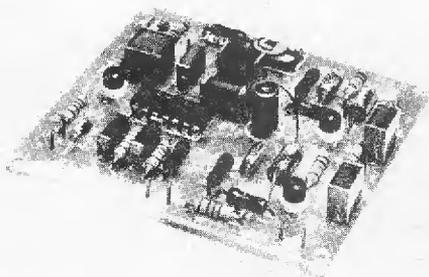
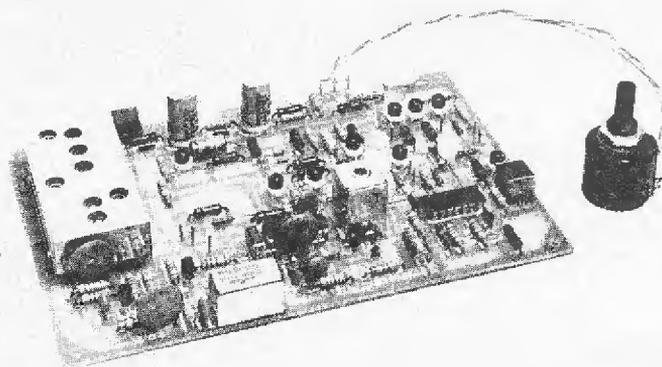
stations. To enable reception of distant signals a switch has been included to short the gate of Tr_5 to ground and defeat the muting operation.

Reference

1. "F.M. Stereo Tuner" by L. Nelson-Jones, *Wireless World*, April-May 1971.

F. M. STEREO*Icon Design***KITS, MODULES, AND COMPLETE TUNERS AVAILABLE.****MAIN RECEIVER BOARD**

Kits 1 to 4 form the main receiver board shown here, providing all functions including a.f.c., mute, and tuning indication. Features include a pre-aligned front end, integrated circuit i.f. amplifier, 3 section ceramic i.f. filter, integrated circuit mute system, varicap tuning via a 10 turn pot., plus 17 other semiconductors.

**STEREO DECODER**

Kits 5 to 7 form the stereo decoder board as shown, including an integrated circuit 'phase locked loop' decoder, anti-birdy active filters, and output pilot tone filters for trouble free recording.

M6 DIGITAL FREQUENCY METER

- ★ CMOS CIRCUITRY
- ★ FULL POWER SUPPLY
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- ★ MONEY BACK GUARANTEE

We are very pleased to be able to announce this new module which has been in development for over 12 months. It has been designed to suit the specific requirements of F.M. tuners, to provide accurate station identification at a glance, at a sensible price and a sensible size. The complete outfit, which is NOT a kit, consists of the transformer, logic and display module and local oscillator take-off board. This latter fits directly to our main receiver board, and requires NO direct connections or modifications to the LP1186 front end module.

If you are interested drop us a line for full details and prices of all our modules.



All kits are guaranteed to work when correctly assembled, and will be repaired free of charge.

Icon Design

33 Restrop View,
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SN5 9DG.

Modular integrated circuit audio mixer

by J. H. Evans,* Dip.Tech.(Eng.) and P. Williams,* B.Sc.

An audio mixer is always a complex constructional project and many must have fallen by the wayside in the face of the difficulties of layout and wiring complexity. A modular approach has been adopted which considerably reduces these problems and at the same time takes advantage of the simplicity of using low cost, readily available integrated circuits. As an additional aid, very complete details of the wiring and front panel design have also been given, which if accurately reproduced should negate hum and instability often associated with home constructed mixers.

The system to be described was designed at the request of the audio-visual aids department of a College of Technology. The requirements were rather unusual in several respects, and as a result the system, though following standard audio practice, has many unusual features. The first, and perhaps the most attractive to many users, is that it was of necessity designed with minimum cost in mind. To that end the active devices used throughout (with one exception) were the lowest cost operational amplifiers available at that time. Since designing and building the equipment the choice seems to have been justified on this ground at least, in that the cost of these operational

amplifiers (type 741) has fallen still further. At the time of writing the total cost for the amplifiers used in this system would be about £4.00 costed at the 100-up rate. However, the price of the amplifiers (of the order of 20-25p each) is such that they can now be thrown around with the gay abandon previously allotted to surplus transistors.

From a circuit standpoint the main difference between this and more usual mixer systems is that a larger number of channels have to be accommodated and, most important, that each channel was required to have separate bass and treble tone controls. This was a user requirement stemming from experience with multiple microphone inputs in locations where little control was possible over the acoustics. The

results had previously been that a single tone control on the mixed channels was inadequate to compensate for the variation in the individual speakers, the microphones and their placements. Since it appeared, from a study of the literature, that passive tone controls were unlikely to cover the wide range of bass and treble lift and cut that might be required in such circumstances, then it became even more important that the active element used in the tone control should have the lowest possible cost. It is clearly arguable that, since a single transistor still costs less than a single operational amplifier, the choice is weighted in favour of the transistor. However, advantages of circuits based on the operational amplifier are that separate biasing networks are not required and the input and output are at almost zero d.c. voltage. By eliminating both coupling and decoupling capacitors in the tone control circuit together with the bias resistors, a saving of fourteen components proved to be possible over a recent transistor tone control circuit of otherwise similar type. For the private user it is not easy to quantify the saving in component cost and the saving in time, although this must be considerable. As with all decisions of an engineering nature other compromises are introduced, and the requirement of both positive and negative supply rails for the amplifier would seem to be a more serious disadvantage. If it is noted that in this circuit, for example, there are eight separate tone control stages together with nine other amplifying and filtering circuits using operational amplifiers, then the additional cost of such a supply, spread over the circuits, is a relatively minor factor.

*Paisley College of Technology.

Parameters	Typ N5741	Supply voltage rejection ratio	10
		Large signal voltage gain	200,000
Parameters	Typ N5741	Output voltage swing	±14 ±13
Input offset voltage	2.0	Output resistance	75
Input offset current	20	Output short circuit current	25
Input bias current	80	Supply current	1.4
Input resistance	2.0	Power consumption	50
Input capacitance	1.4	Transient response (unity gain)	
Offset voltage adjustment range	±15	Risetime	0.3
Input voltage range	±13	Overshoot	5.0
Common mode rejection ratio	90	Slew rate	0.5

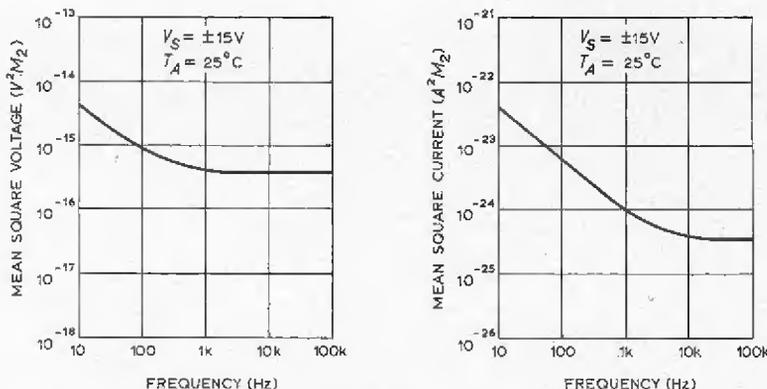


Fig. 1. Typical characteristics for a type 741 op-amp. (Courtesy Signetics Corporation)

Noise considerations

Once the decision has been taken on the grounds of cost and simplicity to use operational amplifiers of a particular type, then the user no longer has the very wide range of choices available to him in other designs. Thus, where the absolute minimum of input noise is the most important factor in the design, it would be necessary to revert to one of the excellent designs published in this field by other authors. Alternatively, the same system could be adopted but using an operational amplifier of higher cost specifically designed for low-noise input performance. That is not to say that the present circuit has poor noise performance,

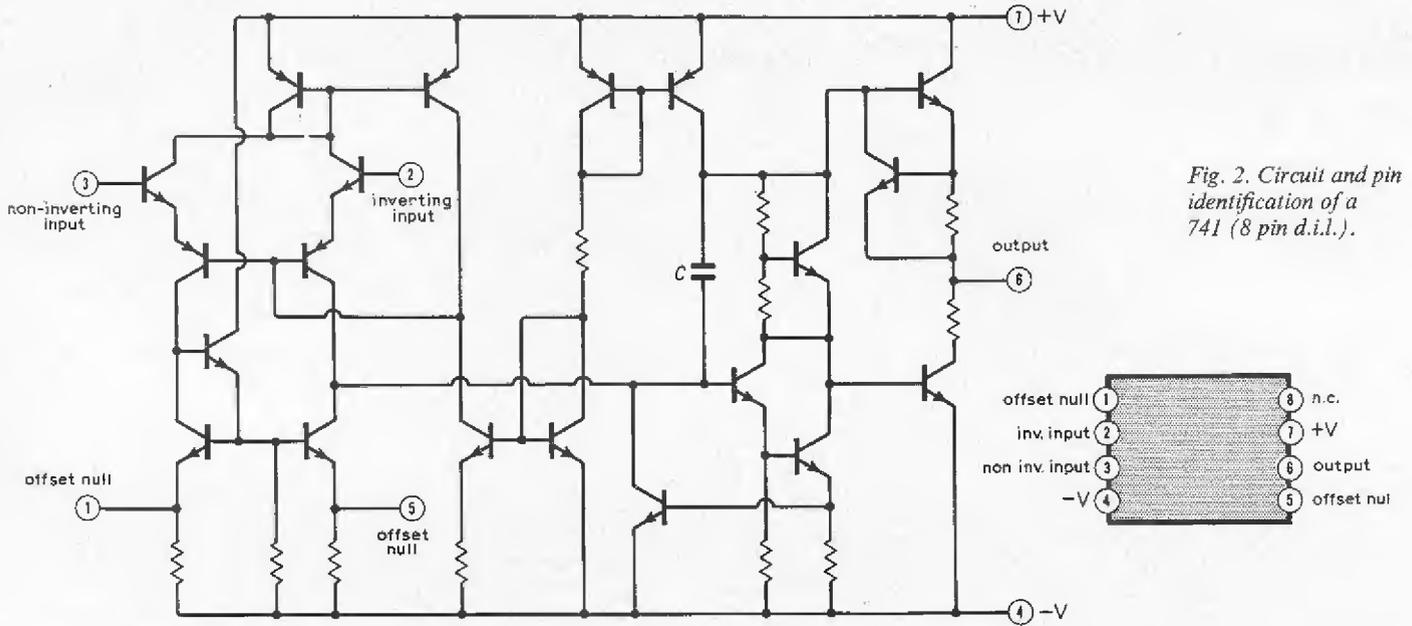


Fig. 2. Circuit and pin identification of a 741 (8 pin d.i.l.).

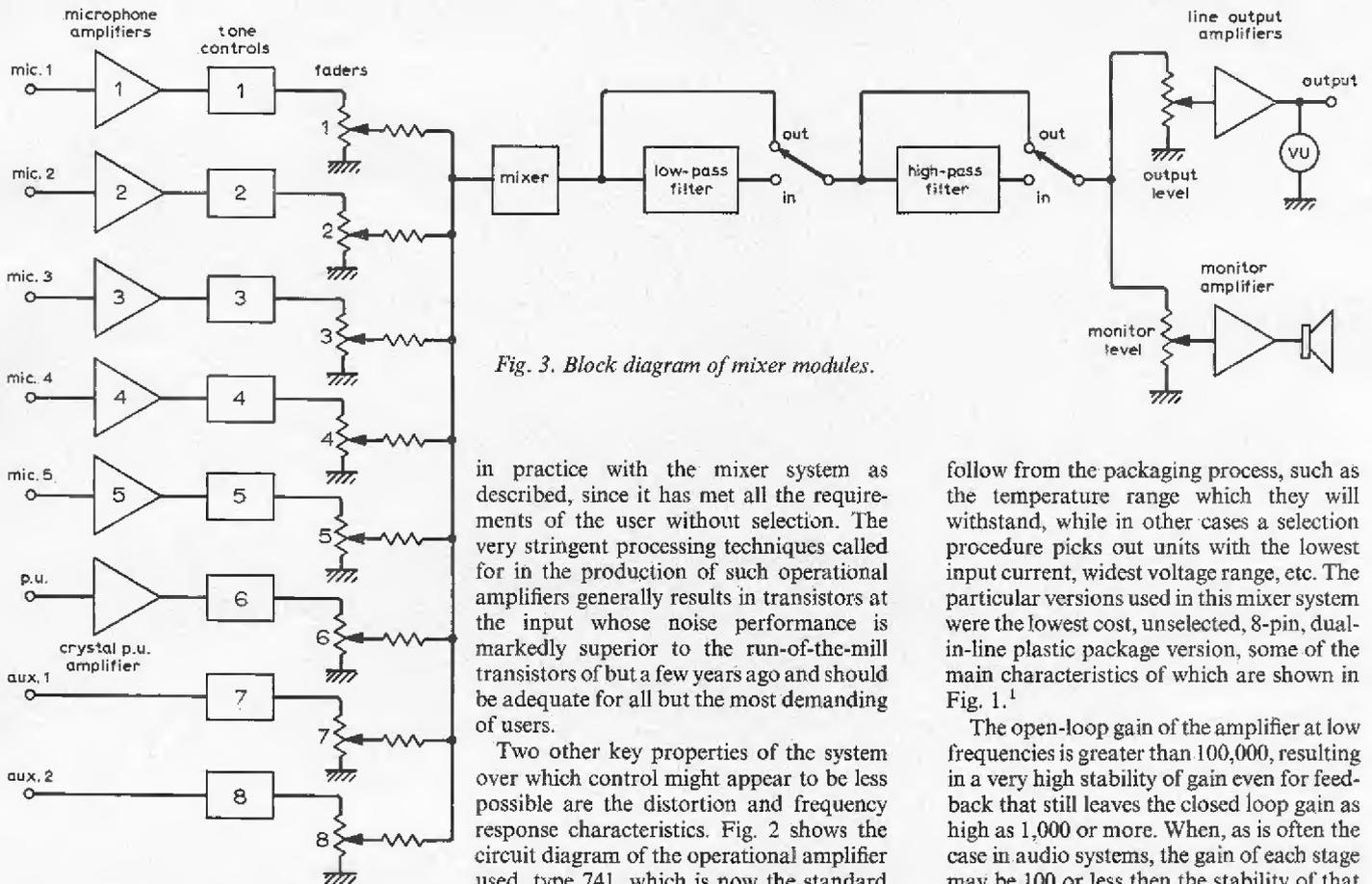


Fig. 3. Block diagram of mixer modules.

in practice with the mixer system as described, since it has met all the requirements of the user without selection. The very stringent processing techniques called for in the production of such operational amplifiers generally results in transistors at the input whose noise performance is markedly superior to the run-of-the-mill transistors of but a few years ago and should be adequate for all but the most demanding of users.

Two other key properties of the system over which control might appear to be less possible are the distortion and frequency response characteristics. Fig. 2 shows the circuit diagram of the operational amplifier used, type 741, which is now the standard operational amplifier for the majority of routine industrial circuits, though of course the integrated circuit manufacturers are continually producing newer i.c.s of ever-increasing performance. This particular operational amplifier is offered by nearly every manufacturer under different code numbers although usually the numbers 741 appear somewhere in the code. The same circuit is produced in a variety of packages: flat pack, TO-5, 14 pin dual-in-line, 8 pin dual-in-line, etc. The circuit is also produced to various specifications which sometimes

follow from the packaging process, such as the temperature range which they will withstand, while in other cases a selection procedure picks out units with the lowest input current, widest voltage range, etc. The particular versions used in this mixer system were the lowest cost, unselected, 8-pin, dual-in-line plastic package version, some of the main characteristics of which are shown in Fig. 1.¹

The open-loop gain of the amplifier at low frequencies is greater than 100,000, resulting in a very high stability of gain even for feedback that still leaves the closed loop gain as high as 1,000 or more. When, as is often the case in audio systems, the gain of each stage may be 100 or less then the stability of that gain depends only on the stability of passive components. At higher frequencies the picture is different. Because the amplifier is designed to be stable and free of oscillations with feedback of up to 100% then great care has to be taken in the control of the frequency response of each stage within the amplifier. If at any time the phase shift around the loop reaches 180° then feedback intended to be negative becomes positive and if the loop gain is still above unity self-sustaining oscillation will result. This explains the presence of capacitor C in the

as is seen from Fig. 1 which gives the characteristics of the type 741 operational amplifier. The input noise voltage and currents given are only typical figures, and where optimum performance is required at the input, a further advantage of using the same active device throughout becomes apparent: the user may simply interchange the operational amplifiers at the input until the unit with the minimum noise performance is found. This has not proved to be necessary

circuit of Fig. 2. This capacitor is connected at a critical point in the circuit where the impedance is relatively high, and ensures that the gain of the amplifier has fallen to a very low level before the frequency is reached at which the phase shift in the other stages becomes significant. Thus, at no time does the total phase shift of the amplifier approach 180° before the gain has fallen below unity. To achieve this effect the reduction in gain begins at a very low frequency so that the gain is reduced by a factor of 100,000 before the critical frequency is reached; i.e. the cut-off frequency for the open loop condition is of the order of 10Hz. In a system in which there is only a single dominant high frequency time constant the gain-bandwidth product remains constant for all values of the feedback. A resistive feedback network that reduces a low frequency gain from 100,000 to 10 will roughly increase the upper cut-off frequency from 10Hz to 100kHz. Broadly speaking, with amplifiers of this type gains of up to 100 in each stage may be achieved with no serious attenuation in the audio frequency band. In the particular circuits used in this mixer the gain per stage is typically $\ll 100$ and a bigger problem is that of restricting the high frequency performance from a noise standpoint.

The output stages of a 741 operational amplifier can be seen from Fig. 2 to be a

form of Class-B push-pull. Without feedback the resulting cross-over distortion would be serious, but, as explained above, the feedback is usually heavy and the feedback is correspondingly reduced. This is not the full story; at higher frequencies the reduction in open loop gain makes the feedback network less effective in reducing distortion and it would not be an ideal circuit for obtaining large output swings at frequencies above 100kHz, certainly not with light feedback. Direct measurements on the amplifier are discussed later and it will be seen that it meets most audio requirements with ease, though once again it must be stressed that the very highest quality amplifiers specifically designed for this application should be capable of better performance. The very low cost of the 741 operational amplifier means that competing systems are likely to suffer a heavy cost penalty for any significant improvement in performance.

System requirements

To meet the needs of the audio-visual aids department concerned, the system as shown in block diagram form in Fig. 3 was devised. The specifications it was intended to meet are shown in Table 1. Each of the eight channels was required to have separate bass and treble, lift and cut tone controls, five of the channels being microphone amplifiers,

one from a ceramic pickup and two line inputs receiving signals from tape recorders and the like. A line output stage was required capable of operating into 600Ω although generally operating into a higher impedance. A monitor amplifier of 1-3W output was also included. To cope with signals having an excessively high or low frequency content, it was decided to have both a low-pass and a high-pass filter and to make these with widely-variable cut-off frequencies. This solution was adopted because the system, of which this mixer was to form a part, was still in the process of considerable expansion and the precise requirements were not known. All of the circuits described so far were based on the 741 operational amplifier but, at a later stage, a VU meter was added and this uses the only separate transistor in the system. Figs. 4-13 show the circuit diagrams of the various functional blocks included in the system. The only components not shown in these figures are the coupling capacitors C_{4a} and C_{4b} which connect the two line input channels to their respective tone control circuits. A fuller discussion of the component values used in the mixer and the reasons for their choice are given in the following sections.

Input amplifiers

The low output impedance and low voltage generated by high quality microphones is in conflict with the input characteristics of the operational amplifier. If the noise properties are to be optimized then the effect of both noise current and noise voltage must be investigated. The noise voltage of a typical 741 amplifier over the audio band is about 3 microvolts. Since the voltage generated by a microphone might be as low as a few tens of microvolts, the signal-to-noise ratio for direct connection would clearly be poor. If we note that the noise current, however, is of the order of 20 pico-amp, then we see that its contribution to the total noise, depending as it does on the low resistance of the microphone input combination, is small. It is this which forces the use of a microphone transformer upon us. From a purely electronic standpoint, there would be much to be said for using a microphone with a high resistance obtained by winding a larger number of turns. A microphone with an impedance in the range 50-500 Ω may be used with a transformer of step-up ratio 5-50 such that the impedance presented to the input of the amplifier is of the order of thousands of ohms, or higher. The step-up ratio ensures that the voltage presented to the amplifier input terminals is now much greater than would have been the case for direct connection. Hence, the input noise voltage is less significant, but the input current is now flowing through a greater source impedance and is more significant. For any given type of amplifier there will be an optimum turns ratio which ensures that the contribution of noise due to voltage and current is comparable. For any other turns ratio, or variation in microphone impedance, the reduction in equivalent noise due to either voltage or current will be less than the increase due to the other. It does not follow that the ratio which

Table 1
Specification for c.c.t.v. audio mixer

- Frequency response**
Flat response from 30Hz to 20kHz.
- Hum-noise**
70dB below rated output.
- Input noise**
200 Ω source, 120dB below 1 volt.
- Channels**
Eight inputs as follows:
 - (a) Five mic. input
 - (i) Source impedance 50 to 500 Ω .
 - (ii) Source e.m.f. 0.125 to 0.28mV.
 - (b) One tele-cine
 - (i) Source impedance 8 Ω (or 16 Ω).
 - (ii) Source e.m.f. 500mV.
 - (c) One record player
 - (i) Source impedance high.
 - (ii) Source e.m.f. 100mV.
 - (d) One tape-recorder
 - (i) Source impedance 15 Ω .
 - (ii) Source e.m.f. 100mV.
- Tone control**
To be provided on all input channels.
Bass ± 10 dB; treble ± 10 dB.
- Outputs**
 - (a) (i) to input of video recorder; unbalance impedance bridging for 600 Ω line.
or (ii) 0.5mV into 200 Ω .
 - (b) To supply 3-W loudspeaker.
- VU meter**
To be provided on output (a).
- Type of controls**
 - (i) Slider-type potentiometers for volume control.
 - (ii) Rotary-type potentiometers for bass and treble.
- Console dimensions**
Dependent on the case or cabinet available, but the operational position of control surface to be 30° to the horizontal.
- Power supply**
240V, 50Hz.

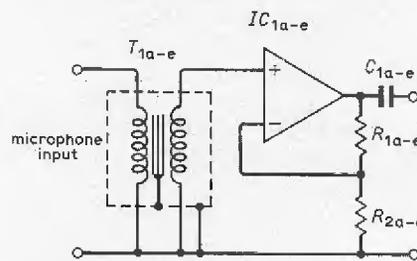


Fig. 4. Practical microphone pre-amplifier.

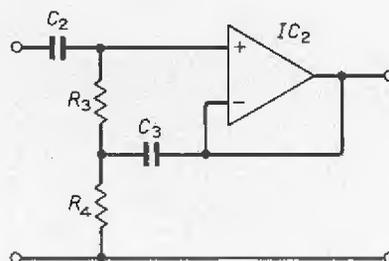


Fig. 5: Pre-amp for ceramic cartridges.

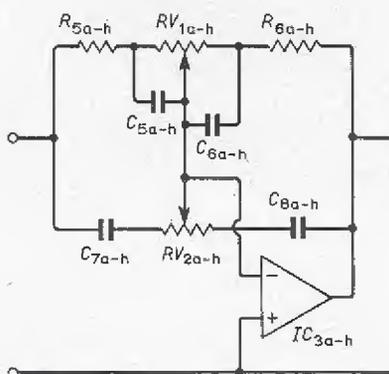


Fig. 6. Tone control circuit.

is optimum from a noise standpoint will also be the optimum ratio for extended frequency response. For example, if the turns ratio is too high the effective source impedance of the microphone may be such that capacitive effects cause a fall in the upper cut-off frequency while it may be difficult to maintain a sufficiently high inductance to avoid losses at low frequencies.

The practical circuit used in this mixer is shown in Fig. 4. For a typical microphone resistance of 200Ω the 15:1 turns ratio of the transformer results in an equivalent source resistance of $45k\Omega$. It is tempting to select feedback resistors that give gain in this stage to eliminate the risk of noise pick-up in subsequent stages, but this is inadvisable for two reasons. Firstly, the smaller the feedback the less is the input impedance of the amplifier raised, while it is important that the input of the amplifier should not place an excessive load on the secondary of the transformer. Secondly, it must be remembered that operational amplifiers of this kind are compensated internally to make them stable for all feedback ratios and hence the compensation reduces the gain-bandwidth product sharply. A low-frequency gain of 1000 is easily obtained, but the upper cut-off frequency might then fall well below 10kHz. With the component values indicated, the overall voltage gain from the microphone to the output of this stage is well in excess of 100. The output voltage thus obtained is of the order of a few tens of millivolts and is comparable to that expected at the line inputs.

For gramophone use the mixer was intended to operate from a ceramic cartridge and a high input impedance is required. The circuit of Fig. 5 provides this with a voltage gain of unity since the output voltage of the cartridge is already equal to or greater than that of the preceding cases. As such a cartridge does not provide a d.c. path for the amplifier input current, it is necessary to provide such a path as with resistors R_3 and R_4 . These would provide a load resistance for the cartridge which would result in a poor low-frequency response, since the cartridge impedance is capacitive. The centre tap of the resistors is therefore bootstrapped from the output of the amplifier by capacitor C_3 . This ensures that the p.d. across R_3 is very small and hence that the current through it is negligible, corresponding to a very high dynamic input impedance. Much has been written recently about alternative input stages for ceramic pickups but there is some advantage to retaining at least one input of this form which may accommodate any voltage source including those of high internal impedance.

Tone controls. The Baxandall tone control is standard within high quality amplifier circuits² and meets all the requirements of the present unit (Fig. 6). Since there was to be additional voltage gain in the following mixer stage and in the line output unit, the resistor values in this stage were chosen for a mid-band gain of unity. The intended input voltages are of the order of a few tens of millivolts, i.e. well above noise level, while leaving a very large overload margin. In an attempt to reduce the component count as

low as possible, advantage has been taken of the almost zero d.c. output voltage of the amplifier by omitting the electrolytic capacitor that would normally be used between stages. Since the overall voltage gain from the input of the tone control circuit to the line output is little more than 30 and the equivalent input offset voltage is but a few millivolts, the line output offset is very much less than 1V. This must be allowed for if d.c. coupled into the load but is not of sufficient magnitude to contribute anything to the non-linearity of the output stage. It is in such modifications as this that the inherent balance at the input of an operational amplifier combined with its large dynamic output range allows for such simplifications. Only experience will show whether the small amount of d.c. voltage that remains across the gain controls will result in noisy operation.

Mixer stage

This is a conventional summing amplifier (Fig. 7) with an output voltage equal to about five times the sum of the input voltages. It would be perfectly possible to use this stage for line output functions, if necessary increasing the gain of the stage as appropriate. As the equipment was at the experimental stage, it was decided to include high-pass and low-pass filters of variable cut-off frequency to be switched in or out as required. It then became advisable to follow these with a final line output stage which also contributed voltage gain. Since no individual stage was called on to provide a high voltage gain, the bandwidth easily exceeded the specification of 20kHz.

Low pass filter. The circuits used for both this and the following filter (Fig. 8) are standard in industrial practice, making use of the operational amplifier in its unity gain mode, i.e. as a voltage follower. The filter has a second order response with cut-off frequency controlled by the ganged variable resistors in the range 500Hz to 20kHz. This range is clearly wider than would be required for audio applications alone, but it was felt that the availability of the filter for experimental demonstrations might be worth while. For this, as with the preceding

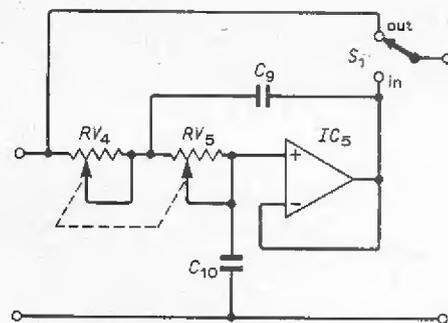


Fig. 8. Low-pass filter.

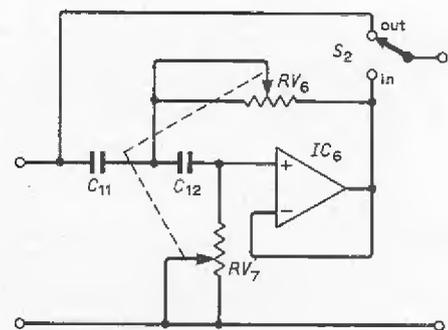


Fig. 9. High-pass filter.

circuits, the output impedance is low, partly because of the low output impedance of the amplifier itself, and partly because of the heavy negative feedback.

High pass filter. This function is simply obtained from the previous circuit by interchanging resistors and capacitors (Fig. 9). Once again the cut-off frequency, in this case a lower cut-off frequency, is variable over a wide range and was chosen to be from about 10Hz to 5kHz.

Monitor amplifier. The increasing availability of low cost integrated circuit power amplifiers greatly simplifies the job of the audio designer. Advantage was taken of the amplifier module type EA1000 (Fig. 10) since this fulfilled all the requirements for this unit. The supply voltage needed was a

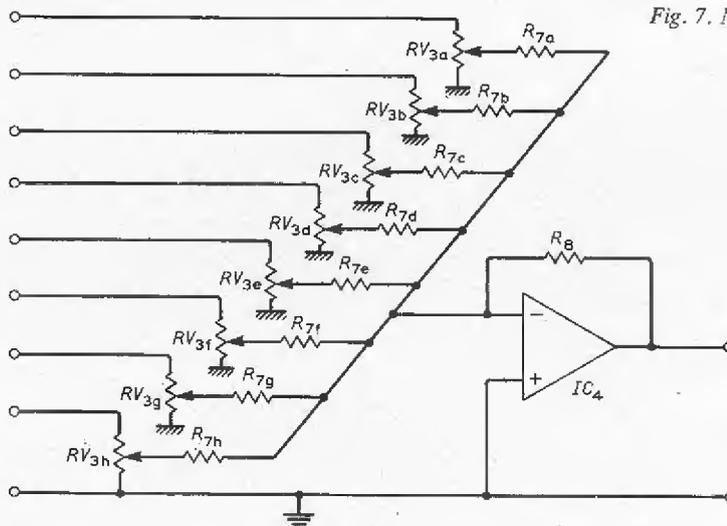


Fig. 7. Mixer amplifier.

single-ended positive supply of some 20V and the circuit operates from the smoothed, but unregulated, supply from which the +15V regulated supply is also derived. On load this supply falls from an initial value of 24V to 21V with a peak output power of 2-3W. At lower powers the distortion of the amplifier is more than adequate and the frequency response, particularly at high frequencies, is very good. The sensitivity of

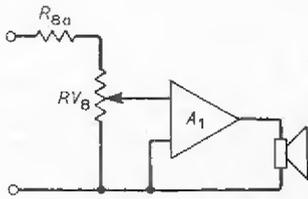


Fig. 10. Monitor amplifier.

the amplifier is such that it can be used directly from the ceramic pickup and the excess gain is, in this application, almost an embarrassment and explains the need for the additional 1MΩ resistor in series with the monitor gain control. For the given value of supply voltage the load resistance should be 15Ω.

Line output stage. This is basically similar to the mixer (Fig. 11), consisting of a virtual earth amplifier preceded by a gain control. Since the integrated circuit used has current limiting, no damage should occur either to the amplifier or to external circuits under any short term fault conditions. Alternative

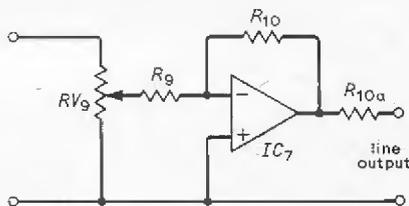


Fig. 11. Line output stage.

values of R_{10} may be used if the sensitivity range provided by RV_9 is not suitable for particular applications. The 600-Ω resistor R_{10a} was included to provide matched operation with a video tape recorder but may be omitted if a lower output resistance is required.

VU meter. (Fig. 12). Lack of experience with the meter circuits of this type led the authors to adopt a circuit described in a recent article.³ It is the only circuit in this unit that uses a transistor and performs admirably. Again, replacement by an operational amplifier might lead to some economy in bias components but this is presumably one circuit where a high open-loop gain is less necessary. The VU meter directly monitors the line output which, because of the relatively high supply voltage, might cause overloading of the meter in some cases.

Power supply. This is entirely conventional (Fig. 13) providing a nominal ±15V for the operational amplifiers and +15V for the VU meter circuit. The positive unregulated

Edge connector wiring details

P.C.B.1: Mic. chans. 1 and 2

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	13	RV _{2b}
2	RV _{2a}	14	0V (ground)
3	RV _{1a} and RV _{2a} *	15	n.c.
4	n.c.	16	RV _{1b}
5	n.c.	17	n.c.
6	RV _{3a}	18	RV _{1b}
7	RV _{1a}	19	RV _{3b}
8	n.c.	20	T _{1b}
9	RV _{1a}	21	n.c.
10	T _{1a}	22	RV _{1b} and RV _{2b} *
11	n.c.	23	RV _{2b}
12	RV _{2a}	24	-15V

*Wiping contacts.

P.C.B.2: Mic. chans. 3 and 4

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	13	RV _{2d}
2	RV _{2c}	14	0V (ground)
3	RV _{1c} and RV _{2c} *	15	n.c.
4	n.c.	16	RV _{1d}
5	n.c.	17	n.c.
6	RV _{3c}	18	RV _{1d}
7	RV _{1c}	19	RV _{3d}
8	n.c.	20	T _{1d}
9	RV _{1c}	21	n.c.
10	T _{1c}	22	RV _{1d} and RV _{2d} *
11	n.c.	23	RV _{2d}
12	RV _{2c}	24	-15V

P.C.B.3: Mic. chan. 5 and crystal p.u.

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	14	0V (ground)
2	RV _{2e}	15	n.c.
3	RV _{1e} and RV _{2e} *	16	RV _{1f}
4	n.c.	17	n.c.
5	n.c.	18	RV _{1f}
6	RV _{3e}	19	RV _{3f}
7	RV _{1e}	20	n.c.
8	n.c.	21	Crystal p.u.
9	RV _{1e}		i/p socket
10	T _{1e}	22	RV _{1f} and RV _{2f} *
11	n.c.	23	RV _{2f}
12	RV _{2e}	24	-15V
13	RV _{2f}		

*Wiping contacts.

P.C.B.4: Aux. chans. 1 and 2

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	13	RV _{2a}
2	RV _{2a}	14	0V (ground)
3	RV _{1a} and RV _{2a} *	15	n.c.
4	n.c.	16	RV _{1a}
5	n.c.	17	n.c.
6	RV _{3a}	18	RV _{1a}
7	RV _{1a}	19	RV _{3a}
8	n.c.	20	Aux. chan. 2 i/p socket
9	RV _{1a}	21	n.c.
10	Aux. chan. 1 i/p socket	22	RV _{1a} and RV _{2a} *
11	n.c.	23	RV _{2a}
12	RV _{2a}	24	-15V

P.C.B.5: Mixer

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	14	0V (ground)
2	n.c.	15	RV _{3f} *
3	RV _{3a} *	16	n.c.
4	n.c.	17	RV _{3a} *
5	RV _{3b} *	18	n.c.
6	n.c.	19	RV _{3a} *
7	RV _{3c} *	20	n.c.
8	n.c.	21	n.c.
9	RV _{3d} *	22	n.c.
10	n.c.	23	RV ₄ and S ₁
11	n.c.	24	'Out' contact
12	n.c.		
13	RV _{3e} *		

P.C.B.6: L.P. and h.p. filters and line output amplifier

Pin No.	Externally wired to:	Pin No.	Externally wired to:
1	+15V	13	n.c.
2	n.c.	14	0V (ground)
3	RV ₇	15	Junction RV ₄ -RV ₅
4	n.c.	16	n.c.
5	RV ₆ and S ₂	17	n.c.
6	n.c.	18	n.c.
7	RV ₅ *	19	n.c.
8	n.c.	20	RV ₅ *
9	RV ₉ *	21	S ₁ 'In' contact
10	n.c.	22	n.c.
11	C ₁₃ and Line o/p socket	23	n.c.
12	S ₁ 'Lever' contact and S ₂ 'Out' contact	24	-15V

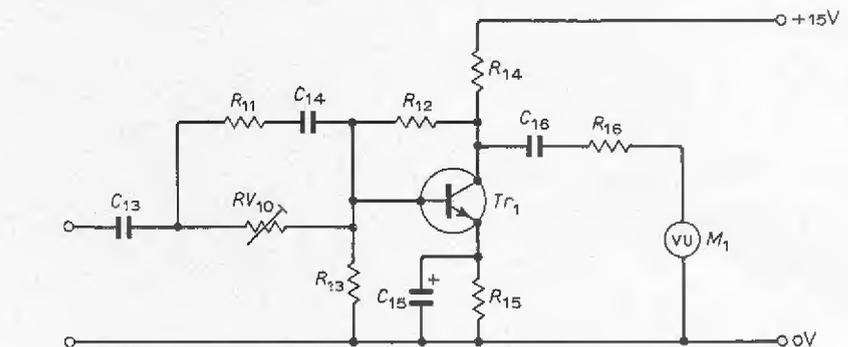


Fig. 12. V.U. meter circuit.

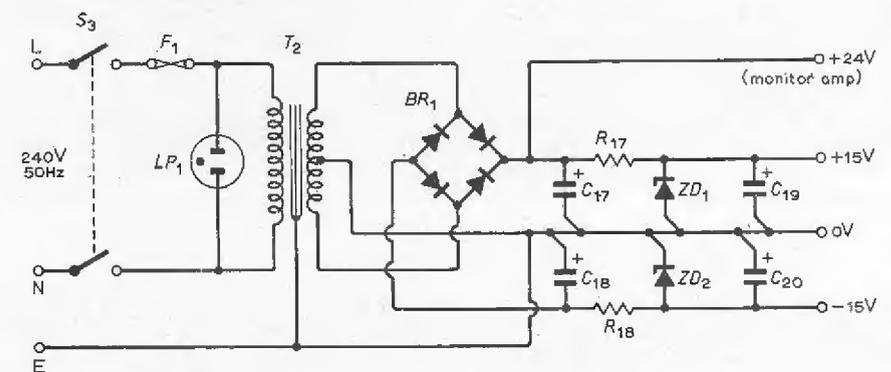


Fig. 13. Power supply.

output is used for the monitor power amplifier, the open-circuit voltage being 24V with the components used. With a continuous sinewave output, just short of clipping, in the power amplifier this unregulated output falls to about 21V, leaving an adequate margin for the operation of the resistor-zener diode stabilizer.

Since each operational amplifier typically consumes less than 2mA—the exception is the line amplifier when driving into a low load resistance—the standing current in the zener need not be high. A simple regulator of this kind seems adequate in view of the excellent rejection of line voltage variation offered by the operational amplifiers. Even on the most sensitive inputs, a total hum and noise of 60dB below the signal level was achieved.

Construction

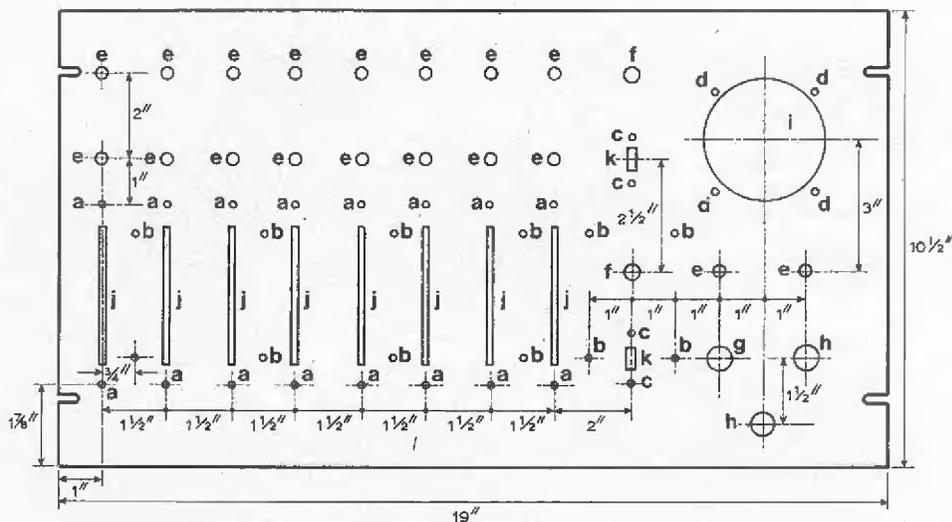
The mixer was to be incorporated in a combined audio/video control-room console, which led to a requirement that the overall width of the mixer should not exceed about twenty inches. A standard 19in panel carrying all components was used to meet this requirement.

The controls to be accommodated at the front of this panel consisted of eight linear-motion faders, twenty rotary-motion potentiometers, two slide switches and a mains on/off switch. The VU meter, a mains fuse and a neon indicator were also required to be mounted on the front of the panel. The rear of the panel carried a sub-chassis for the power supply, brackets for the input and output sockets, monitor amplifier and input transformers, six edge connectors for the printed circuit boards and a printed circuit board for the VU meter circuit.

Such a large number of front panel components required that a compromise be made between the separation of controls, for ease of use, and the length of the wired interconnections, to reduce noise and crosstalk. The resulting front panel layout (Fig. 14), shows that the tone controls and fader for each channel are in vertical alignment and that this approach lends itself to modular construction. The common apparatus in the system is all mounted to the right of the panel, allowing the tone control and fader components for the desired number of channels to be added to the left.

The tone control and input circuit components were mounted on commercially-available printed circuit boards carrying the integrated circuits. Each board carries all the components for two channels and is mounted in a 24-way edge connector located between the faders. The high-pass and low-pass filter components were mounted on a single board, and the mixer components were accommodated on a separate board. These two boards were of the same type as those used for the tone control circuits and were mounted in the same manner. A printed circuit board was designed to carry all the components in the VU meter circuit so that it could be mounted directly on the pins at the rear of the meter.

The power supply and mains plug were supported on a small sub-chassis raised above the output volume controls and



a	fixing holes (8 pairs) 6BA clear on $4\frac{3}{16}$ " centres	g	component hole $\frac{9}{16}$ " dia.
b	" " (6 pairs) " " $2\frac{7}{8}$ " "	h	" " $\frac{1}{2}$ " "
c	" " (2 pairs) " " $1\frac{1}{8}$ " "	i	" " $2\frac{3}{4}$ " "
d	four fixing holes 6BA clear on $3\frac{1}{4}$ " dia.	j	" slots $3\frac{5}{16} \times \frac{1}{8}$ " "
e	component holes $\frac{5}{16}$ " dia.	k	" " $\frac{1}{2} \times \frac{1}{4}$ " "
f	" " $\frac{3}{8}$ " "		

Fig. 14. Front panel layout.

mains switch by means of pillars. The line output socket and monitor loudspeaker socket were fixed to a single bracket supported by the bolts holding the mixer and filter edge connectors. The monitor amplifier printed circuit board was fixed to the other mixer-board pillar by means of a simple bracket. The eight input sockets and five input transformers were mounted on a single bracket running along the front panel below the faders.

Most of the wired interconnections were made by means of 8-way, unscreened, twisted cable running underneath the bracket carrying the input sockets and transformers. Screened pairs were found to be unnecessary, the only screened wire being a 3-in length between the line output volume control and the high-pass filter output. The edge connectors need to be raised from the front panel by small pillars to allow a clearance for the wiring pins. The wired interconnections can easily be made by fixing the edge connectors in the inverted position during construction and then restoring them to the correct position when the work is completed.

Results

Appendix 1 gives a comprehensive picture of the behaviour of the system and should allow the individual to use whatever combination of the sub-sections is most convenient. Other amplifier/mixer circuits have been described having performance that exceeds that of the present unit. In most cases, however, the component count has been considerably higher with a corresponding increase in constructional costs. It is recognized that the component cost may not show the same advantage; this will clearly depend on the cost of the integrated circuits to the individual user. As in industrial engineering, the shortening of design time, the increased flexibility and the reduc-

tion of auxiliary components makes the case for integrated circuit operational amplifiers in audio frequency designs almost unanswerable.

Appendix 1

Performance data

Line output

NOTE: All measurements made with $R_{10} = 1\text{k}\Omega$ and $R_{10a} = 0$.

(1) Line input auxiliary channel 1:

Tone controls: 3 o'clock. Filters: out.
Freq.: 1 kHz. Reference level: 10mW into 600 Ω , i.e. +10dB.

Frequency response relative to above:

+0dB to -1dB at 5Hz and 20kHz.
Response sensibly flat to 40kHz -5dB
Attenuation at 100kHz -23dB
Rise time 15 μ s.

Maximum power into 600 Ω (undistorted)

> 50mW at 100Hz.
> 50mW at 1kHz.
~ 40mW at 10kHz.
Output resistance at 1kHz $\ll 1\Omega$.

Tone controls

Bass: +13.5dB } 50Hz
-15dB }
Treble: +13dB } 12kHz
-15dB }

Filters

For 30dB attenuation at 10Hz and 50kHz.
3dB bandwidth is 50Hz-14kHz.

Filter cut-off frequency range

Low-pass (scratch) filter ~ 500Hz-20kHz.
High-pass (rumble) filter ~ 10Hz-5kHz.

Distortion

Output +10dBm into 600 Ω .

(a) Frequency 1kHz: 2nd harmonic < -65dB
3rd harmonic < -64dB
4th and higher harmonics
-not measurable.

Total harmonic distortion ~ 0.04%
(including test oscillator distortion).

(b) Frequency 100Hz:

Total harmonic distortion ~ 0.04%.

- (c) Frequency 10kHz:
Total harmonic distortion ~0.18%.
- Noise At maximum sensitivity:
(a) Input 30mV r.m.s. for 0dBm into 600Ω:
Total noise -58dB (~0.13%).
- (b) Input >300mV r.m.s.:
Total noise <0.04%.
- (c) With all gain controls at minimum, residual noise more than 70dB below rated output.
- (2) Microphone input channel 1
Tone controls: 3 o'clock. Filters: out.
Reference level: 10mW into 600Ω, i.e. +10dBm.
Frequency response: for source resistance of 200Ω:
+0dB to -3dB at 45Hz and > 20kHz.
With bass boost ±1.5dB 20Hz-20kHz.
- Sensitivity
500μV for +10dBm into 600Ω.
Equivalent voltage gain ~5000.
- Noise
Equivalent input noise for 20kHz bandwidth ~0.5μV r.m.s.
Noise -60dB w.r.t. 500μV signal.
Further reduction of 2dB by restricting bandwidth to 10kHz.
- NOTE: Theoretical noise performance for industrial grade operational amplifier in audio range 20Hz-20kHz.
Noise voltage ~5 × 10⁻¹⁶ V² Hz⁻¹.
Noise current ~2 × 10⁻²⁶ A² Hz⁻¹.
For 20kHz bandwidth this gives:
Noise voltage ~3.1μV r.m.s.
Noise current ~20pA r.m.s.

Optimum noise performance will be achieved when the noise contribution from current and voltage sources are equal. This occurs when the effective noise source resistance is:

$$\frac{\sim 3.1 \times 10^{-6}}{78 \times 10^{-12}} \Omega$$

$$\approx 40k\Omega.$$

With a 15:1 step-up transformer this requires a source resistance of:

$$\frac{\sim 40k\Omega}{225} \approx 178\Omega.$$

Hence, a microphone having a resistance of about 200Ω will give optimum noise performance.

Monitor output

Line input auxiliary channel 1:

Tone controls: 3 o'clock. Filters: out.
3dB bandwidth 45Hz-25kHz at 300mW into 15Ω.

n.b. For low distortion, low-frequency cut-off should be kept to ~80Hz.

Distortion

- (a) Frequency 1kHz, 0.5W into 15Ω.
Second harmonic ~0.1%
Third harmonic ~0.2%
Fourth harmonic ~0.2%
Fifth and higher harmonics <0.1%
Total harmonic distortion ~0.3%
- (b) At 100Hz and 10kHz.
Total harmonic distortion under above conditions <3%.

Appendix 2

Components list

Resistors

R _{1a} -R _{1e}	1kΩ	0.3W	5%	Mullard
R _{2a} -R _{2c}	100Ω	0.3W	5%	Mullard
R ₃	47kΩ	0.3W	5%	Mullard
R ₄	47kΩ	0.3W	5%	Mullard
R _{5a} -R _{5h}	10kΩ	0.5W	5%	Piher
R _{6a} -R _{6h}	10kΩ	0.5W	5%	Piher
R _{7a} -R _{7h}	22kΩ	0.5W	5%	Piher

R ₈	100kΩ	0.3W	5%	Mullard
R _{8a}	1MΩ	0.3W	5%	Mullard
R ₉	10kΩ	0.3W	5%	Mullard
R ₁₀	220kΩ	0.3W	5%	Mullard
R _{10a}	600Ω	0.3W	5%	Mullard
	(1.2kΩ/1.2kΩ)			
R ₁₁	47kΩ	0.5W	5%	Piher
R ₁₂	39kΩ	0.5W	5%	Piher
R ₁₃	5.6kΩ	0.5W	5%	Piher
R ₁₄	4.7kΩ	0.5W	5%	Piher
R ₁₅	270Ω	0.5W	5%	Piher
R ₁₆	3.6kΩ	0.5W	5%	Piher
	(2 × 1.8kΩ)			
R ₁₇	270Ω	1W	10%	Morganite
R ₁₈	270Ω	1W	10%	Morganite

Capacitors

C _{1a} -C _{1e}	10μF elect.	15V		Radiospares
C ₂	0.1μF	10%		Mullard
C ₃	1μF non-pol.	20%		Radiospares
		63V		
C _{4a} -C _{4b}	4.7μF non-pol.	20%		Radiospares
		63V		
C _{5a} -C _{5h}	47nF	10%		Mullard
C _{6a} -C _{6h}	47nF	10%		Mullard
C _{7a} -C _{7h}	10nF	10%		Mullard
C _{8a} -C _{8h}	10nF	10%		Mullard
C ₉	1nF	2.5%		Radiospares
C ₁₀	1nF	2.5%		Radiospares
C ₁₁	0.1μF	10%		Mullard
C ₁₂	0.1μF	10%		Mullard
C ₁₃	1μF non-pol.	20%		Radiospares
		63V		
C ₁₄	150pF	2.5%		Radiospares
C ₁₅	50μF elect.	15V		Radiospares
C ₁₆	2.2μF non-pol.	10%		Mullard
C ₁₇	2000μF elect.	25V		T.C.C.
C ₁₈	2000μF elect.	25V		T.C.C.
C ₁₉	500μF elect.	25V		Radiospares
C ₂₀	500μF elect.	25V		Radiospares

Variable resistors

RV _{1a} -				
RV _{1h}	100kΩ log.	50mW		Radiospares
RV _{2a} -				
RV _{2h}	22kΩ log.	50mW		Radiospares
RV _{3a} -				
RV _{3h}	10kΩ log.	0.2W		Preh
RV ₄	250kΩ log.	ganged 0.25W		Radiospares
RV ₅	250kΩ log.			
RV ₆	250kΩ log.	ganged 0.25W		Radiospares
RV ₇	250kΩ log.			
RV ₈	100kΩ log.	50mW		Radiospares
RV ₉	100kΩ log.	50mW		Radiospares
RV ₁₀	22kΩ 1in. preset	0.25W		Radiospares

Integrated circuits

IC_{1a}-IC_{1e}; IC₂; IC_{3a}-IC_{3h}; IC₄; IC₅; IC₆; IC₇;
All Signetics Type N5741V.

Transistor

Tr₁ ME4103 Micro-Electronics

Transformers

T_{1a}-T_{1c} U204 A.K.G., G.m.b.H.
T₂ MT112C Douglas

Zener diodes

D₁ and D₂ ZL15 15V, 1.5W

Amplifier

A₁ EA1000 S.G.S.

VU meter

M₁ MR65P Shinohara

Bridge rectifier

Br₁ BS1 International Rectifier

Switches

S₁ 250V, 1A Slide Radiospares
S₂ 250V, 1A Slide Radiospares
S₃ 250V, 1A Radiospares
d.p.d.t.

Fuse

F₁ 100mA, 20mm, Radiospares
antisurge

Neon indicator

LP₁ 220-250V Radiospares

Hardware

1 off: Front panel 19" × 10³/₄" Lektrokit
6 off: 24-way edge connectors, type LK2271 McMurdo
6 off: 24-way printed circuit boards, type LK3111 Lektrokit
20 off: Black knobs plus red pointers Radiospares
1 off: 20mm fuse holder Radiospares
1 off: VU meter printed circuit board P.C.T.
1 off: 3-way miniature mains plug Bulgin
1 off: 3-way miniature mains socket Bulgin

References

1. Signetics Corporation, 'High Performance Operational Amplifier Linear Integrated Circuit, Type N5741V', August 1970.
2. Quilter, P. M., 'Low Distortion Tone-Control Circuit', *Wireless World*, April 1971, pp. 199-200.
3. Walker, H. P., 'Stereo Mixer', *Wireless World*, June 1971, p. 298.

The "walltenna"

Foil antenna array hidden by wallpaper

by Ray Schemel and Dennis Brown

For f.m. stereo, long-distance TV, or improving reception in marginal areas, one's first temptation is to install the largest and most elaborate fishbones, wire mesh reflectors, and other pieces of aluminum-mongery at the highest and most eye-offending part of a house. It is odd, but more time, effort, and money is expended on extracting the last decibel of gain from an antenna than almost any other part of the receiver chain. After six months' exposure to the weather, those precious decibels may well have been lost and the extra signal level might better have been obtained, for example, by paying attention to the antenna matching. In fact an excellent location for an antenna could well be in the same room as the receiver — it goes without saying that the room should have walls of paper and be in the attic! The antenna structure is not exposed to the elements, water cannot enter the feeder, feeder losses are minimized, and the appearance of our towns and villages would be much improved.

The "walltenna" was conceived as a method of receiving the Wrotham and Norwich f.m. transmitters at a location near the coast of Holland at distances of 180 and 135 miles respectively. At this range the transmitters are well below the normal horizon, but a weak signal is almost always present. Signals from the "walltenna" fluctuate over a wide range, and in good weather conditions, presumably when sufficient refraction or ducting takes place, stereo reception is perfectly possible. Estimates of quality tend to be very subjective, but the signal is well above a significant degree of quieting for a high proportion of the time, although not necessarily of entertainment value because of the rapid fading. The estimates are quoted, not because this article is about f.m. reception, but to give some idea about the pick-up properties of the finished antenna.

The basic principle behind the antenna is simple. Why not make a large array essentially two dimensional, and then hide this behind wallpaper or some other decorative medium? The large size of the array would compensate for the losses of the walls, and the bother of

mounting antennas on the roof would be avoided. Various alternative schemes were considered, including multi-element Yagi and rhombic arrays on the ceilings and floors, but the arrangement described here was found to be the most suitable because it was not overly directional in the horizontal plane.

It consists of a vertical array of dipoles mounted broadside to the direction of propagation; the greater the number of dipoles, the greater the signal pick up. It is often overlooked that the signal power extracted from a given field strength is broadly independent of the wavelength, being only a function of the area of the antenna. An elaborate 20-element beam used for u.h.f. TV probably picks up less power than a piece of wire connected to a medium-wave radio. To increase the power available one has no real alternative but

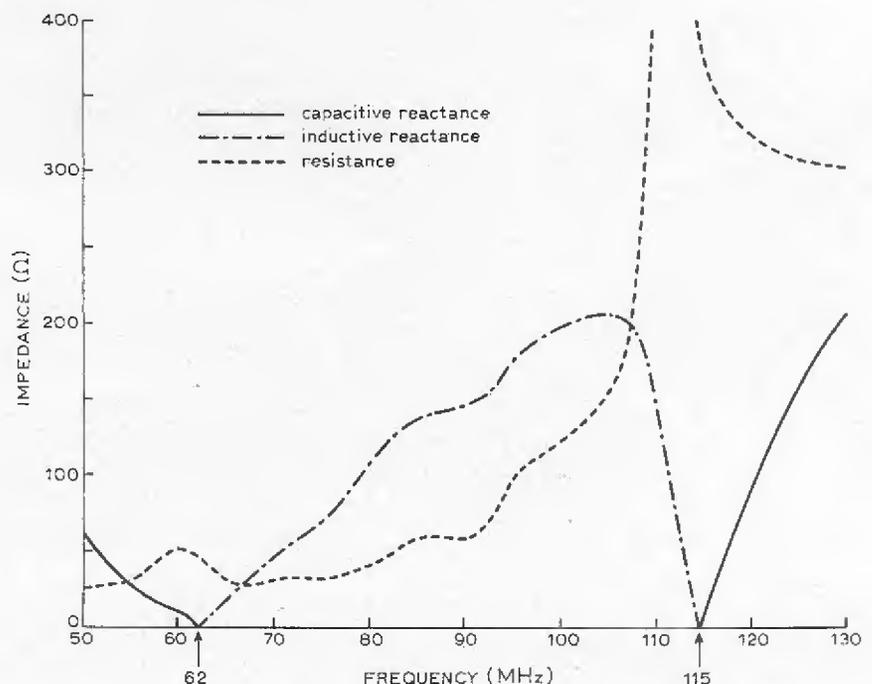
to use the largest capture area practicable.

A dipole in free space is resonant when it is almost one half wavelength long. It then has a gain of about 2.2dB relative to isotropic. As the length is increased, the familiar figure-of-8 polar diagram narrows, and when the dipole is a full wavelength long and is centre fed, the gain increases by a further 1.8dB. Of more importance is that the antenna is inherently a good radiator (or receiver) as the length increases beyond a half wavelength.

Four such dipoles, suitably phased, could give a gain of up to 10dB, and if one were fortunate enough to be able to place a plane wire mesh reflector behind the dipoles, the gain could go as high as 16dB, not allowing for mutual coupling losses. A reflector is scarcely possible in a living room, but even so, a gain of up to 10dB is quite promising. Four dipoles fit nicely into an average height of living room, and so this particular design was adopted. In practice the gain will depend on the wall characteristics.

The situation when dipoles are placed adjacent to a lossy wall is complicated in that the impedance and the resonant frequencies are markedly altered. The resonant frequency is always decreased, and the radiation resistance is also altered for a given electrical length. In addition, radio waves must penetrate what amounts to a lossy dielectric. Reflection and refraction occur at each wall-air interface and attenuation occurs in passing through the brickwork. Complicated conduction and displacement currents are set up in the wall. In spite of this, the signal pick up properties of the antenna are not necessarily degraded provided that full account is taken of the changed impedance of the dipole. Fig. 1 shows the impedance of a 1½-metre dipole mounted on an 18-in thick breeze block and

Fig. 1. Series impedance of 1.5 metre dipole mounted on a wall (measured values).



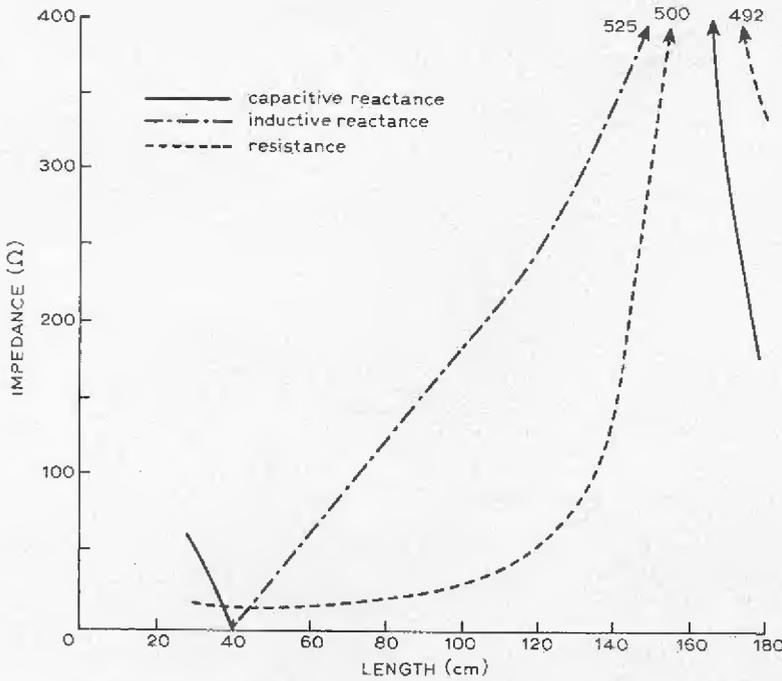


Fig. 2. Series impedance of a dipole mounted on a wall at 92MHz (measured values).

brick wall. The antenna was made of 1½-in wide aluminium foil taken from a capacitor, and would normally be resonant at about 97MHz in free space. Notice that the half-wave resonance has shifted to 62MHz and the full-wave resonance occurs at 115MHz. An alternative measurement is shown in Fig. 2, where the frequency is held constant at 92MHz and the length of the dipole is varied. In this case the dipole was placed along an 18-in thick reinforced concrete beam; notice again the length at which the half and full-wave resonances occur.

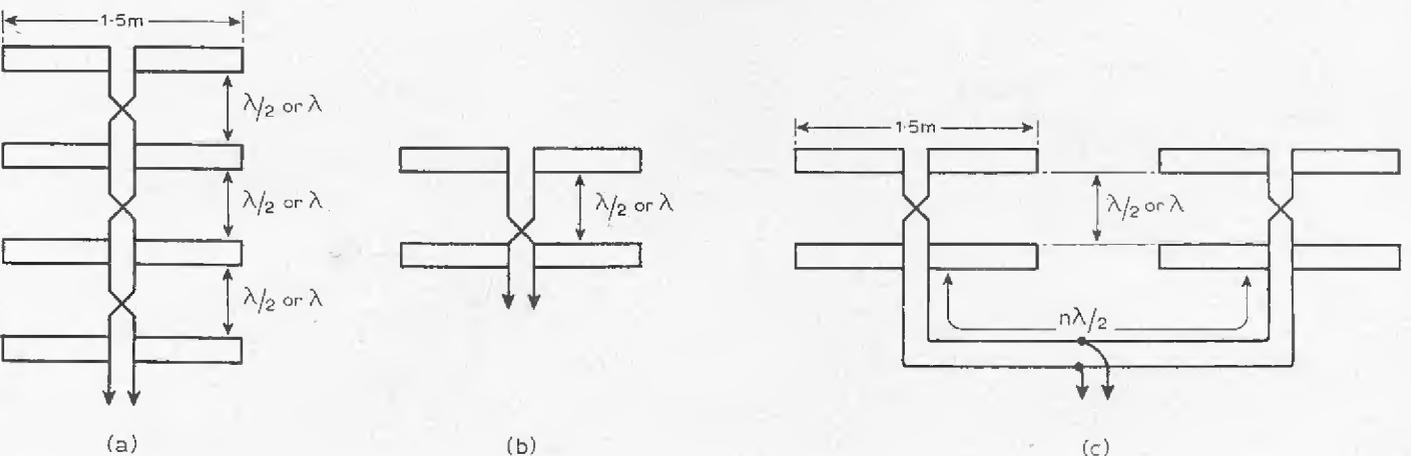
Most *Wireless World* readers will not have the facilities for measuring antenna impedances, which in any case vary in a complex manner with frequency and with the type of wall. Fortunately, it is not necessary. An antenna does not have to be resonant, and neither does it have to be any particular length. Provided that it is long electrically, i.e., greater than, say, one half wavelength, and that some suitable impedance matching to the receiver is performed, it will exhibit reasonable pick up properties. A good rule of thumb is to make an antenna a full wave-length long. It

then yields useful gain and is not too directional or large in size. For average brick walls this corresponds to it being about one half wavelength long in free space.

Constructional details

The configuration of four vertically-stacked dipoles referred to earlier will be found the most convenient for an average eight-foot living room. Details of this are shown in Fig. 3(a), and its construction is largely self-explanatory except for the feeders. Figs. 3(b) & (c) show other arrangements which may be found more suitable, and in these cases the general rules for combining any number of elements are as follows:

Fig. 3. Four vertically-stacked dipoles (a) most convenient for average living rooms; alternatives are (b) and (c). Electrical lengths of feeders shown.



- When $n/2$ wavelengths of feeder are terminated in an impedance Z_r , the input impedance of the feeder is also Z_r . The signal is shifted in phase by $180n$ degrees. Usually $n = 1$.
- Balanced feeders can always be reversed to give a fixed phase shift of 180 degrees.
- When $(n/2 + 1/4)$ wavelengths of feeder are terminated in an impedance Z_r , the input impedance of the feeder is Z_o^2/Z_r , where Z_o is the characteristic impedance. Usually $n = 0$.
- When dipole elements are not mounted above one another, there is a relative phase shift caused by the difference in time of arrival of the received signal. The phase difference is $360D\sin\theta$, where D is the antenna spacing in wavelengths and θ is the angle of arrival relative to the line joining the elements.
- The phase difference of two groups of dipole elements may be compensated for by using unequal feeder lengths, provided that the sum length of both feeders is $n/2$ wavelengths.
- The electrical length of feeders is longer than their physical length. Because of this, the phase shift is $360/v$ degrees per wavelength of feeder, where v is the velocity factor.
- Dipoles should be spaced no closer than a quarter of a free-space wavelength apart, otherwise mutual coupling effects reduce the potential signal pick up.

An excellent material for the antenna is the foil from large paper or electrolytic condensers. Aluminium baking foil or one of the proprietary aluminium-backed wallpapers can also be used. The width is not critical, but to make the antenna reasonably wide band, it should be 2in or more. After cutting the foil it may be pasted directly onto the wall, leaving a small amount free at the centre for making connection to the feeders.

Feeders present a small problem if they are to be invisible. A surprisingly

low-loss feeder suitable for the purpose of v.h.f. can be made from aluminium foil and ordinary newspaper. Lay a long strip of aluminium foil on a flat surface and tape down the ends. Then paste a similar width of newspaper onto the foil. When it is dry, lay two lengths of thin wire, say 36 to 40 s.w.g., along the middle of the newspaper about 0.25in apart, and tape down the ends. Finally, paste a second strip of newspaper on top to hold the wires in position. Alternatively, those who are a little more adventurous could try dispensing with the top layer of newspaper and instead building up the feeder directly on the wall.

A feeder made in this way had a velocity factor of 0.5 and an impedance of 170 ohms, but these figures will be dependent on the dielectric constant of the paper used. For short runs of feeder, say up to one wavelength, the approximations involved are not likely to give significant effects, but to remove the uncertainty connected with the velocity factor it is hoped that a manufacturer will be found to produce a standard type of strip line. There are other alternatives: ordinary screened balanced 75-ohm feeder can be channelled into the wall, or unscreened twin feeder can be placed in plastic conduit.

Connect feeders to the antennas by bending over the aluminium foil, pressing down firmly over the join and then folding a second time at 45° to the first, and finally placing adhesive tape over the join to hold the foil in position.

The special feeder described above is only required down to the skirting board, and for convenience it should be made an integral number of quarter wavelengths long if possible. At the skirting board it should be matched into ordinary coaxial cable or a balanced feeder.

The impedance matching makes use of a pi network and a balun as shown in Fig. 4 (the balun is not required if the receiver has a balanced input). To obtain the maximum range of adjustment, inductance L_1 should be wound on a former that accepts brass and ferrite slugs. Also, some obstinate antennas may still need additional encouragement to match, in which case an extra quarter wavelength of feeder can be inserted on the receiver side to give a further impedance transformation. The pi network could, with advantage, be placed on the unbalanced side of the balun as it makes the two adjustable capacitors earthy on one side. However, 3-30pF trimmers are easy to obtain whereas ones four times this value are not, and the balun always works at the correct impedance level.

Baluns are available but you may wish to make your own. A satisfactory balun can be made by winding 10 turns of bifilar wire onto a coil former, about 1/4in diameter with a ferrite slug. The two windings are then placed in series as shown in Fig. 4.

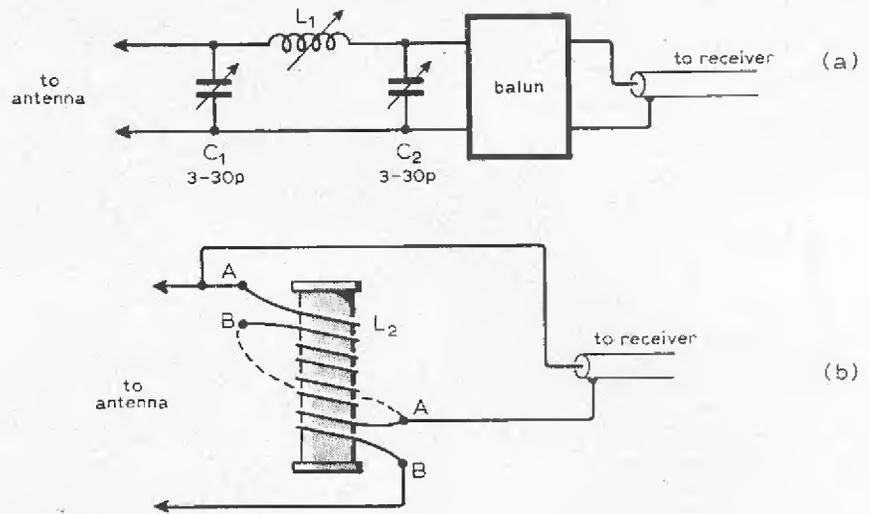


Fig. 4. Impedance matching unit (a) requires a balun (b) for receivers with unbalanced input circuits. L_1 is 5 to 10 turns on a 0.25in dia. former about 0.3in long. L_2 is 10 turns of close-wound bifilar wire on 0.25in dia. former wire gauge can be around 28-30 s.w.g.

Measurements and setting up

Before constructing the "walltenna", make a folded dipole as shown in Fig. 5, remembering to include a balun. This is most important, as unscreened twin feeders can pick up appreciable amounts of signal which may upset measurements. If the receiver normally operates with a balanced input it is well worth using a twin screened feeder to overcome the pick up problem, or even considering the use of two baluns and ordinary coaxial cable. The temporary antenna should be moved to different points on the wall to find the best positions, and these should be used for the finished dipole elements.

After the antenna is finished, the pi network will require adjusting for maximum signal transfer. To do this a meter is really required, perhaps placed on the a.g.c. line of the receiver, but an acceptable alternative is to use a weaker station and to tune for minimum noise on stereo. First adjust C_2 ,

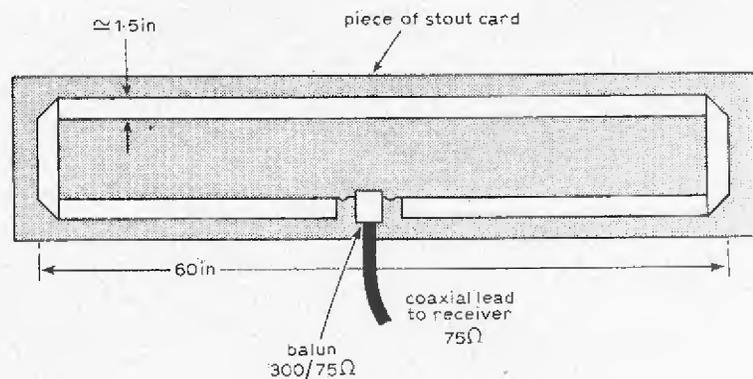
then L_1 , and finally C_1 . Remember that C_2 and C_1 are live as far as the signal is concerned and must not be touched, as opposed to adjusting, when measuring.

If the antenna of Fig. 3(c) is being used, the tapping point along the feeder must first be optimized. This can be done by calculation; the displacement from the centre point away from the dipole nearest the transmitter is $1/2 D \sin \theta$ as per the points made earlier. Alternatively, a temporary sliding connection can be made along the feeder and then this is adjusted together with the pi network.

Follow the above procedure before wallpapering, because, in spite of its simplicity, difficulties which cannot be rectified afterwards may occur. In principle, every element added to those existing should increase the total signal, after an impedance adjustment has been carried out. In practice it may be found that adding further elements gives no increase or even a decrease in signal, and better results are obtained by reversing the feeder connections. Such effects need not be unexpected. They result from differences in the pick up capabilities of each array element and the variable impedance of the wall. In these cases it is simply a question of cut and try methods, and if all else fails, of experimenting with fresh positions.

The "walltenna" principle is applicable to a wide variety of antennas and it is hoped to be able to give constructional details for u.h.f. TV at a later date

Fig. 5. Temporary antenna is used to find best position on wall.



Electronic piano design

Simple touch-sensitive piano using ready-made keyboard — 1

by G. Cowie, B.Sc.

The instrument described is a simple touch-sensitive electronic piano which is small and portable. The circuitry is designed on a modular basis using i.c.s extensively and is not difficult to construct. It generates tones by an oscillator-divider system, the tones being keyed by individual touch-sensitive key circuits. Costing around £70 to build, the design is believed to be the most cost-effective available, in terms of what it is intended to do, and a commercial instrument with this touch-sensitive feature would seem to cost at least £300.

This design was built to fill a real need; if there had been an acceptable instrument on the market I would have bought it instead of spending three months and £50 in making one. I was learning to play the piano and wanted an instrument of my own for practice. As I live in furnished flats, moving frequently, a full-sized upright just was not practicable. The alternatives were a "mini" piano or an electronic piano. I ruled out the first on finding that second-hand instruments were surprisingly expensive because of the demand; moreover they were not portable enough.

This left electronic pianos. I looked at several but they cost a lot of money and I did not like them. The trouble is that they all have the same artificial keying action in which pressing the key beyond a certain point suddenly generates a note of fixed loudness. On the cheaper instru-

ments the note cannot be held after the key is released.

I wanted an instrument which behaved just like a string piano, even if it did not sound much like one. The prototype is quite true to my original intentions: if you play loud the sound comes out loud, if you play very softly the sound hardly comes out at all. In playing a chord, one can make some notes loud and others soft at the same time. There is a "loud" pedal which can be used to sustain notes after the key is released, and the sound dies away just as in a real piano.

The low notes have long decay time constants, and the high notes have very short time constants. The tone is a bit like that of an electric piano, a harpsichord, and an electric guitar. Most listeners find the tone pleasant; it is much less harsh than that of a real piano. In any case to imitate a grand piano perfectly would

be very difficult. Most important of all, the instrument is about the size of a large suitcase and light enough to be carried by one person.

The essential feature of the design is that the volume of sound generated depends on how fast the key is pushed down, a feature not then available to my knowledge on any other electronic piano. This feature is what makes a pianoforte what it is, and its importance was impressed on me by a musician friend with whom I discussed the project. In simple terms, the effect is to add a new dimension, that of loudness, to the sound. Although I made no attempt to vary the tone along with the volume, the characteristics of the human ear and of the power amplifier cause such an effect with my instrument. I find the instrument sufficiently interesting to play that I have not found it necessary to add tone-shaping circuits though this could be done quite easily. The only extra is a swell pedal which is useful for increasing the dynamic range. Functionally, the instrument is much the same as a string piano, except that there is no "soft" pedal. Purists may argue that the key action is not the same as that of a real piano. Strictly speaking this is so, but the art of playing the electronic piano is so similar to the art of playing a real one that there is no difficulty in changing from one to the other.

To produce an electronic imitation of a real piano would be an ambitious undertaking, and potentially an uneconomic one. If the imitation is to be useful by reason of its small bulk and competitive cost then compromises are necessary. The sound that a piano makes has a complex harmonic content. This is not an insuperable difficulty in itself, but the harmonic content varies according to the loudness with which the note is struck, and with time; it is also different for notes of different pitches. Loudness of course varies with time, with a fast attack and slow decay. As if this weren't enough



to contend with, most of the notes employ two or three strings which do not vibrate in perfect unison.

The keys of a piano have a characteristic feel when pressed down by the finger: there is a constant resisting force caused by the weight of parts on the inner end of the key, a reactive component caused by the inertia of the fast-moving hammer, and a small amount of friction. The harder one tries to play, the greater the reactive part of the opposing force becomes. From the musician's point of view this characteristic of the piano is most desirable, as it makes it easier to exploit the touch-sensitive loudness which is inherent in the piano. The faster the key goes down, the faster the hammer moves, the harder the hammer hits the strings, the louder the sound. Music tutors exhort one to think of the key speed rather than the pressure. Technically this makes sense as only the final speed of the hammer matters and it is easier to accelerate it by a smooth pressure than by jabbing the key (and the music sounds better). In this piano design, each key is linked to its own timing circuit so that sound output depends on the average velocity with which a key is pushed down.

Various electronic pianos are on the market. Those priced competitively with respect to conventional uprights seem to adopt similar solutions to the problems outlined above. Keyboards are similar to those used in electronic organs, with the same touch, and most of the instruments are not touch-sensitive at all. One infers that the acoustic waveforms are square waves treated by low-pass filters, and by high-pass filters for special effects. All have the right sort of fast attack, slow decay as this is very easy to effect for square waves. Again by inference the frequency generation is done by oscillators and dividers as in electronic organs.

Design considerations

After preliminary thought and discussion, I decided that my electronic piano would be touch-sensitive, have a sustain pedal, use square-waves as the working waveform, and have twelve master oscillators, using t.t.l. 7493 integrated-circuit frequency dividers to generate the lower pitches. On seeing ready-made organ keyboards and keyswitches I decided to use these and, as five octaves is a standard size, that the little-used top and bottom octaves of the conventional 88-note piano keyboard could be dispensed with. This simplifies the work considerably and makes the finished instrument significantly smaller and lighter.

The essential features of electronic key circuit are shown in Fig. 1. The key position must be determined electrically, to control an envelope shaper whose output is used to modulate the amplitude of a continuous-pitch waveform. Tone is determined by the pitch waveform, and attack and decay are determined by the envelope. As the piano is to be touch sensitive then the initial height of the envelope must be variable.

A number of envelope waveforms are shown in the Fig. 2. These show a note played and released, a note played and

Table 1. Fundamental frequencies for C-C keyboard

Octave section	C	B	A	A	G	G	F	F	E	D	D	C
1 (osc. freq.)	2093	1975	1865	1760	1661	1568	1480	1397	1318	1244.4	1174	1108
2 (1st div.)	1046.4	987.7	923.3	880.0	830.6	783.8	739.8	698.4	659.2	622.2	587.2	554.2
3 (2nd div.)	523.2	493.8	466.2	440.0	415.3	391.9	369.9	349.2	329.6	311.1	293.6	277.1
4 (3rd div.)	261.6	246.9	233.1	220.0	207.6	196.0	185.0	174.6	164.8	155.6	146.8	138.6
5 (4th div.)	130.8	123.4	116.5	110.0	103.8	98.0	92.5	87.3	82.4	77.8	73.4	69.3
6 (5th div.)	65.4	61.7	58.2	55.0	51.9	49.0	46.2	43.6				

N.B. For modified C-C keyboard (see text) or an F-F keyboard, range is 43.6Hz to 1397Hz.

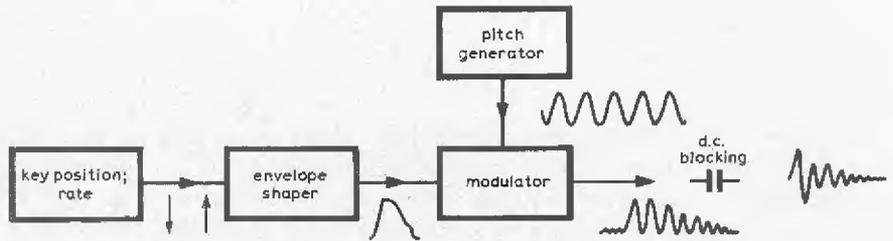


Fig. 1. Key switching and rate information provides an envelope that modulates signals from a pitch generator (c.w. oscillator).

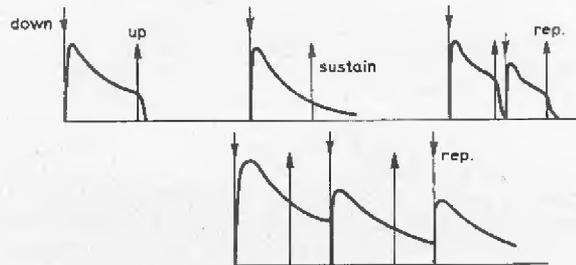


Fig. 2. These envelope waveforms are determined by switches actuated by keys and the sustain pedal. Diagrams show single and repeated notes with and without sustain action.

sustained, and notes repeated without and with sustain. Generally, electrical key contacts are used to signal the states "up", "down" and "moving" and this can be done by a single contact moving between two busbars. But I found that the simplest and cheapest system must use more contacts to simplify the electronics. An electronic piano must have an electronic switch to block the pitch signal when it is not required; in my circuit a diode is used.

Twelve oscillators generate the twelve pitches for an octave, and the pitches for lower octaves are obtained by dividing by 2, 4, 8, 16 (Table 1).

The oscillators use operational amplifiers instead of LC circuits — it is cheaper to buy op-amps than to buy special coils. Also, the op-amp circuit is easier to design and can be tuned by a cheap pre-set potentiometer. A detailed discussion of this type of oscillator is given in *Electronic Engineering*, Nov. 1971, page 54. Complex m.o.s. microcircuits are now available which will produce the twelve top-octave frequencies when driven by a radio-frequency master oscillator. Thus all the key pitches are synchronized with the master oscillator and the organ or piano never needs retuning. Such a device would add about £3 to the cost of the project

and a suitable (optional) module will be described in part 3 of this article.

Regulated supplies of +5 and -5 volts are provided as a regulated 5-volt supply was needed in any case for the t.t.l. divider circuits. The advantages of integrated-circuit frequency dividers over discrete dividers in cost, time, space etc are such as to make them the only choice. About half of the piano circuitry is inside the divider i.c. packages.

I devoted much thought to making the key circuits as simple as possible. As there are sixty-one key circuits, elimination of even one component could save hours of work and pounds of hard cash. Wood was the obvious choice of material for the case. The case is styled after my own conception of how an electronic piano should look, and has no lid as this wasn't essential and would not fit into the design. There is more room at the rear of the case than is strictly necessary; this was deliberate in that making the case too big would cause nothing like as much trouble as making it too small.

Circuit Description

Fig. 3 is a schematic diagram of the complete circuitry which is too complex to be drawn in full. Under the keyboard

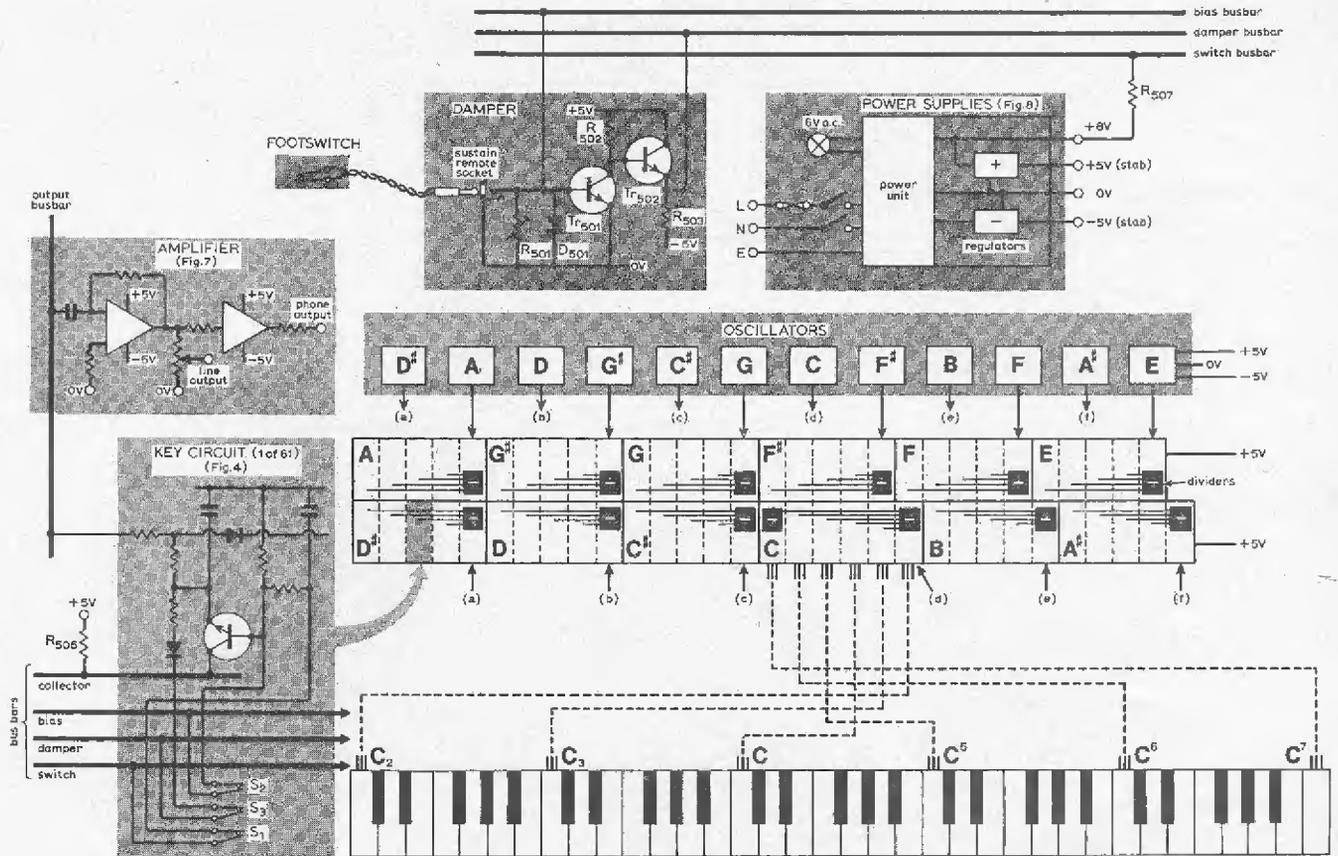


Fig. 3. Time between S_2 and S_1 , S_3 opening as the key is pressed determines loudness of sound, achieved with the key circuit. Twelve RC oscillators feed 12 dividers (shown above keyboard in diagram), giving 60 of the 61 tones for modulation by the key circuits. For a five-octave keyboard 61 tones are actually required, the additional tone (C^7) being provided by a 13th divider, as the i.c.s will only divide by a maximum of 16. Outputs from the key circuits are mixed in the amplifier via the output busbar. Three of the busbars run alongside the keyboard so that three keyswitch leads can be directly soldered to them. R_{502} is $1k\Omega$.

are 61 sets of three normally-closed gold-plated wire switches, one set being shown. One pole of each switch is connected to one of three busbars which run under the keyboard. The other poles are wired to a key circuit, which is one of a letter-group of five as there are five octaves; and there are twelve letter-groups. There is a sixth note C for which an extra key circuit and an extra divider to give + 32 provided.

Each letter-group of key circuits is fed with signals from an oscillator and frequency divider. The key circuit for one note is drawn in full to show the interconnections.

To each of the 61 keyswitches six connections are made. Three of these are common busbars (bias, damper, switch), and the other three are bias, damper, and switch signal lines and all go to one key circuit, linking the key to the electronics.

The power supply feeds +5 volts to the oscillators, frequency dividers, summing preamplifier and output amplifier and, via a resistor, to the collector bus which feeds all 61 key circuits. It also feeds an unregulated +8-volt supply to the switch bus, and -5 volts to the oscillators and amplifiers.

The output bus is a virtual earth line

fed from all 61 key circuits. The bias and damper buses are controlled by the sustain pedal.

Key circuits

Though the key circuits appear simple (Fig. 4), each has three sections which are more or less analogous to parts of a string piano. Components C_1 , R_1 , R_2 , S_1 , S_2 , form a velocity-measuring circuit which gives the piano its touch-sensitive property. The charge in C_1 , when the key is depressed, represents hammer velocity. Transistor Tr_1 provides isolation between the input and output sections of the circuit — the equivalent in a real piano is a device allowing the hammer to fall back. Capacitor C_2 has a charge representing the vibrational energy of a string. Components D_1 , S_3 , R_3 form the damper circuit, which may be disabled to give a sustain action. Diode D_1 blocks the pitch signal when the circuit is on standby and, when the circuit is active, forms a chopper and output circuit with R_4 and R_5 . The discharge times of capacitors C_2 vary to imitate the peculiarity of the string piano whereby bass notes die away more slowly than the treble.

Velocity section. Standby operation is as follows: current flows from the switch busbar, which is at about +8V, through S_1 , R_1 , S_2 , to the bias-busbar at about +0.7V, so that Tr_1 is just cut off. Capacitor C_2 is charged to about +0.4 V. When the key is partway depressed, contacts S_1 and S_3 open, and C_1 discharges toward +0.7V, with a time constant of 18ms. (This is a critical time constant that influences the playing properties of the instrument.) When the key is almost fully depressed, S_2 opens and the remaining charge in C_1 passes through R_1 into the base of Tr_1 , which conducts heavily, causing a corresponding charge to appear in C_2 . If R_2 were not included in the circuit, then S_2 having opened, a capacitance of C_1 times the gain of Tr_1 would be added to C_2 ; R_2 ensures that C_1 always discharges faster than C_2 .

When the key is released, S_2 closes first, S_3 closes discharging C_2 , and S_1 closes, recharging C_1 from the other 60 capacitors in parallel. The resulting current surge does not damage the contacts as they will handle up to 2A at low voltages, and the power factor of the capacitors is very poor. All the contacts have so far survived 21 months of use. The action of the circuit is such that when

the key is pressed very swiftly, C_1 loses potential by only a volt or so, and a potential of nearly five volts appears on C_2 . When the key is pressed very lightly, C_1 discharges almost to the bias voltage, so that a very small charge is delivered to C_2 .

Envelope section. Capacitor C_2 , having been charged, begins to discharge in pulses through R_4 , D_1 and the 7493 i.c. (Fig 5) with a time constant $2R_4 C_2$. A square chopped signal appears across R_4 and is taken out via R_5 ; the amplitude of the signal being C_2 voltage minus D_1 volt drop. The t.t.l. divider outputs have two transistors in a sort of class B arrangement so that in the high state it acts as a current source; in the piano circuit this is a nuisance and is blocked by D_1 . In the low state it acts as a current sink to ground. The voltage applied to the output must not exceed 5V.

This diode chopper was chosen because it is the simplest modulating circuit with a precisely definable output and a low feed-through of pitch signal in the off state. The impedance of this section is low to reduce the effect of D_1 leakage and capacitance in its off state. Hence C_2 must be relatively large. The impedance of the velocity section is relatively high to minimize standing currents, hence a current amplifier Tr_1 is necessary.

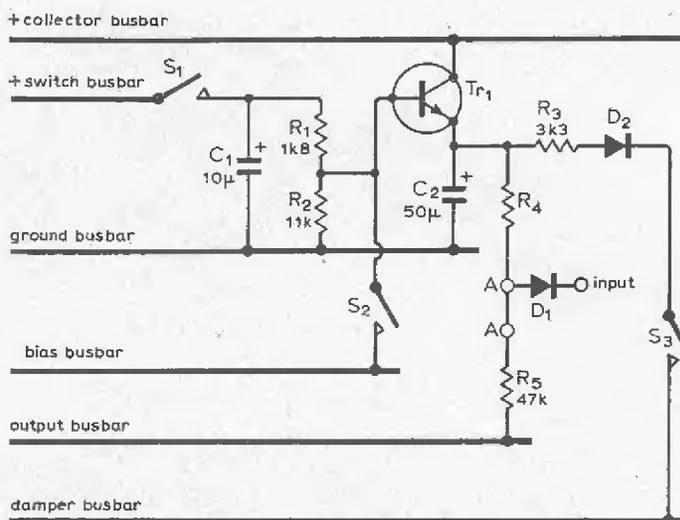
Damper section: When the key is released, S_3 closes, discharging C_2 through R_3 and D_2 . The value of R_3 is made large enough to avoid a key click. A sustain action is effected by raising the potential of the damper bus so that no current can flow into it through D_2 , and so C_2 discharges through the chopper, irrespective of the key position. Normally, R_3 , D_2 drain some leakage current from C_2 . The sustain allows capacitors C_2 in circuits on standby to charge up slightly via Tr_1 until D_1 begins to conduct. To suppress this chorus effect, the bias busbar potential is reduced.

Resistor R_4 varies from $1k\Omega$ (top C) to $15k\Omega$ (low C) — Table 3 lists values.

Oscillator circuits

The 12 oscillators use operational amplifiers in a precision relaxation oscillator circuit (Fig. 5). The output of the op-amp switches between nearly the positive and negative supply voltages. When the output has just gone positive, the negative input voltage starts to change positively as charges. When it reaches the positive input voltage, $V_0R_1/(R_1 + R_2)$, the inverting action causes the output to fall negatively, almost instantaneously. The circuit has a bridge configuration which almost eliminates the effect of load and supply voltage. At the instant of switching a differential voltage of $2V_0R_1/(R_1 + R_2)$ exists at the amplifier inputs, which limits the ratio R_1/R_2 that can be used at the chosen supply voltage with 709 series op-amps (but not 741 types). Sources of drift in the circuit include offset voltage and bias current changes.

Supplies of +5V, -5V were chosen to simplify the power supply and buffer circuits. This has rendered the oscillators



$$C = \frac{6}{f} \text{ to } C = \frac{60}{f} \pm 20\%$$

f = frequency in Hz; C in μF

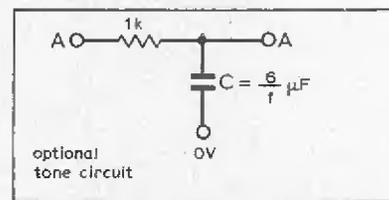


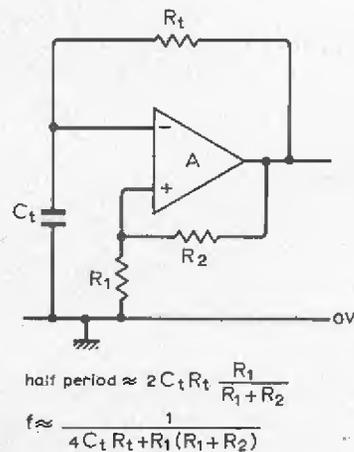
Fig. 4. Key circuit (one of 61) acts as envelope shaper and modulates input from the dividers to feed amplifier via busbar. The three busbars wired to the three switches run alongside keyboard switches. Optional tone circuit was used on the bottom twelve keys of the author's design to make low tones less harsh.

Table 2. Oscillator resistance values

Pitch	Frequency (Hz)	Nom. tuning R (kΩ)	Use (kΩ)
C ¹	1108	36.5	33 + 4.7
D	1174	34	33 + 4.7
D ¹	1244.4	32	30 + 4.7
E	1318	30	27 + 4.7
F	1397	28	27 + 4.7
F ¹	1480	27	24 + 4.7
G	1568	25	24 + 3.3
G ¹	1661	24	22 + 3.3
A	1760	23	22 + 3.3
A ¹	1865	21	20 + 3.3
B	1975	20	18 + 3.3
C	2093	19	18 + 3.3

more sensitive to ripple and one-sided supply voltage changes. Frequency depends on the values of R_{201} , R_{202} , R_{203} , R_{204} , and C_{201} , and is set by R_{203} (Fig. 6). High-stability components are required.

A buffer circuit R_{205} , Tr_{201} is incorporated as the load of the divider input and the discharge current from the top-octave key circuits may be as much as 6.6mA, which is more than the op-amp can be guaranteed to drive. The 7493 i.c. divides the oscillator frequency by 16 and has outputs for 2, 4, 8, divisors also. It changes state for an input transition from high to low, and has two reset inputs, one of which must be grounded for operation as a divider or counter. The outputs will sink more than 16mA ample for driving the key circuits. Eleven of the oscillator and divider circuits are as shown, but for C an overall divisor of 32 is required and so the output $\div 16$ is wired to the input of another divider stage whose output then feeds low C.



$$\text{half period} \approx 2C_t R_t \frac{R_1}{R_1 + R_2}$$

$$f \approx \frac{1}{4C_t R_t + R_1(R_1 + R_2)}$$

Fig. 5. Relaxation oscillator uses bridge configuration to keep effects of load and supply voltage changes to a minimum.

Summing preamplifier

The summing circuit is a standard op-amp arrangement. Capacitor C_{401} (Fig. 7) blocks the d.c. component of the output, i.e. the C_2 voltages which would be summed to about 25V. Resistor R_{401} is not mounted on the board but at a suitable point among the key circuits, thereby using the differential inputs to minimize pickup of unwanted signals. The headphone amplifier likewise is a fairly standard discrete-component op-amp with complementary emitter-follower output. It can be readily altered to drive various loads.

Table 3. Decay time constant resistor values (R_x in Fig. 4). (61 resistors needed.)

Note	Octave section							
	1	2	3	4	5	6		
C	}	1k	1.8k	3.3k	5.6k	10k	18k	(low C)
B								
A [♯]								
A								
G [♯]	}	1.2k	2.2k	3.9k	8.8k	12k	18k	(top F)
G [♯]								
F [♯]								
F								
E	}	1.5k	2.7k	4.7k	8.2k	15k	—	(low F)
D [♯]								
D								
C [♯]								

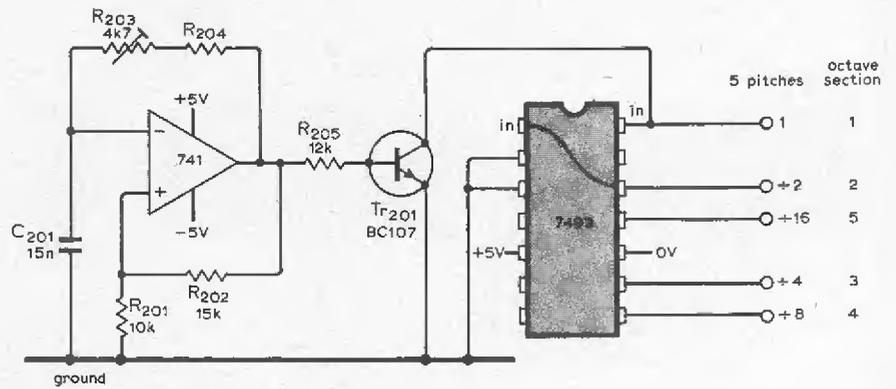


Fig. 6. One of 12 RC oscillators which feed the 12 dividers, giving five octave-related tones for each of 12 notes. Circuit shown in Letters is alternative.

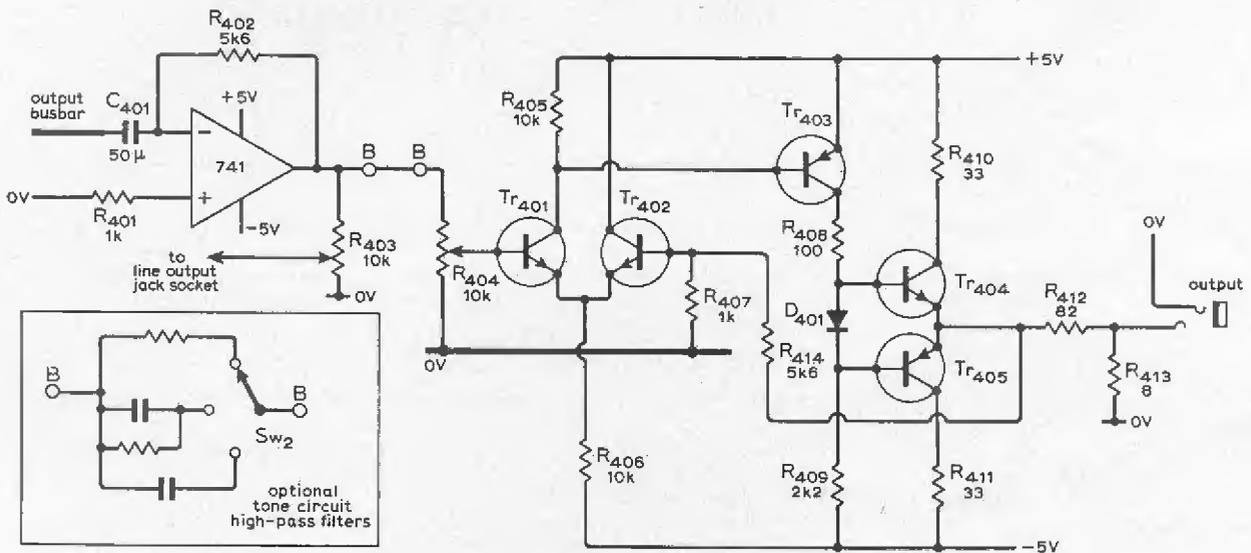


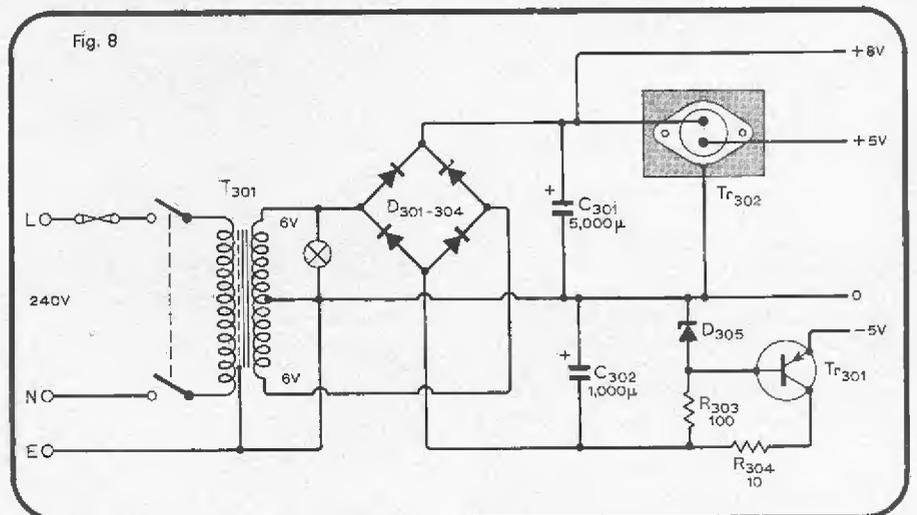
Fig. 7. Amplifier accepts inputs from 61 key circuits via output busbar. Discrete-component amplifier can be omitted if headphone operation is not needed.

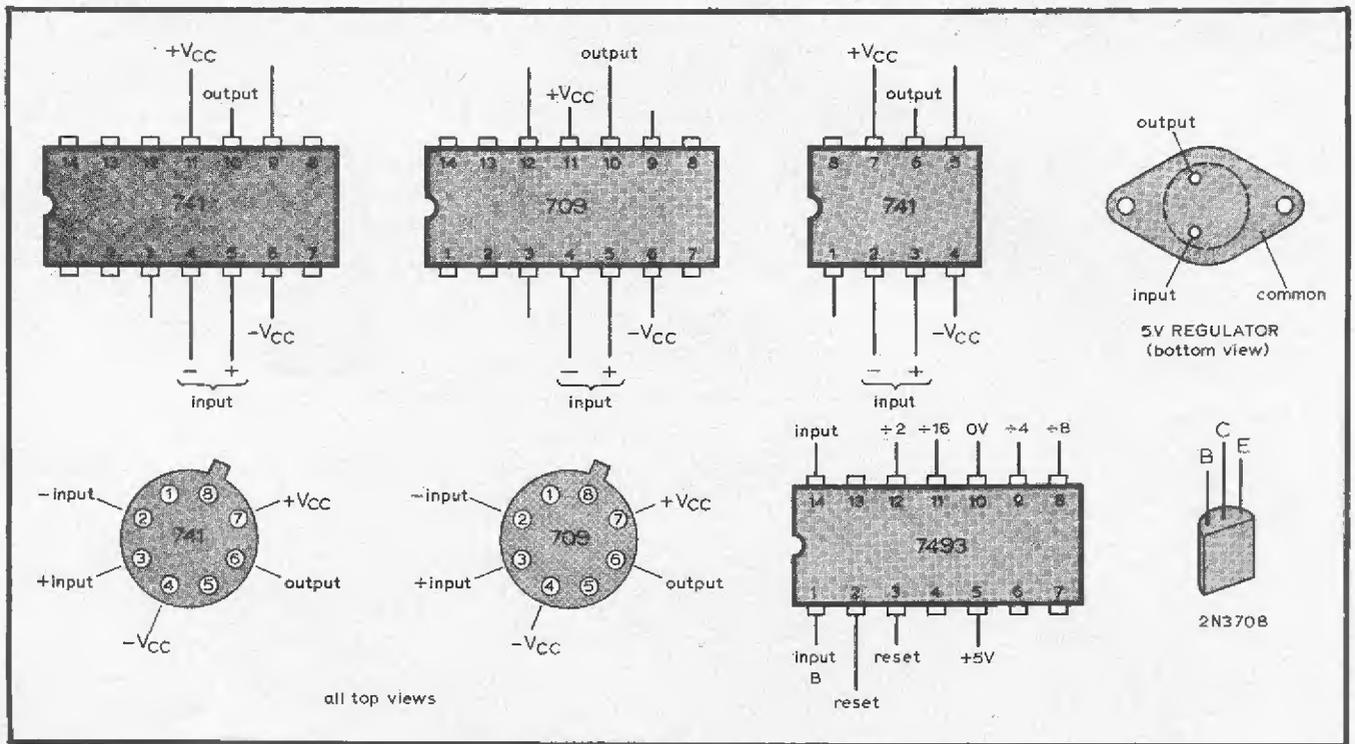
Power supply

The power supply unit (this page) provides regulated voltages of +5V at 600mA, -5V at 50mA, an unregulated +8V at 250mA, and a low-voltage a.c. to light an indicator lamp. The twin full-wave rectifier arrangement produces two "raw" d.c. supplies with a minimal voltage loss across diodes. A simple discrete-component regulator is used for the negative line, and an integrated regulator is used for the positive supply to maximize reliability. The l.t.l. divider circuits may be ruined if a regulator failure causes their supply voltage to exceed 7V. Regulated supplies assist in maintaining the frequency stability of the oscillators.

Busbar lines

A number of wires feed signals and power to all parts of the piano circuitry (Fig. 3.) These wires are for the +5V regulated power line, the -5V regulated line, ground potential, the switch bus fed via a resistor from the unregulated





+ supply, the output bus, the bias bus and the damper bus.

The bias bus is returned to ground via a rectifier diode and trimming potentiometer in parallel, so that the bus voltage may be finely adjusted. When the sustain pedal is pressed, thus closing a remote switch, the bias voltage is reduced, cutting off Tr_{501} whose base is connected to the bias bus. An emitter-follower Tr_{502} is used to control the damper bus, whose potential is 0.7V below that of Tr_{501} collector. Therefore when the sustain pedal is depressed, the damper bus potential rises by 5V to +4.3V.

Keyboard

The keyboard used in the prototype is a five-octave plastics-keyed C-C electronic organ keyboard of Italian origin, supplied by Elvins Electronics. Other keyboards may be obtained, e.g. from Harmonics Ltd and from Kimber-Allen, and if these are used some of the cabinet dimensions may have to be altered.

For musical reasons it is most preferable that the keyboard be F-F with low F at 43Hz. I believe that such keyboards may now be obtained to order. With a C-C keyboard, one has the option of modifying it to F-F by cutting away the bottom five notes and re-attaching them at the top of the keyboard.

Converting the C-C keyboard. Unhook the springs from the bottom ends and poke out the pivot wire. Put the keys aside in order. Cut the frame flush with low F and cut off the top C mountings also. Rearrange the C-E portion at the top of the keyboard with the top C (now top F) above it, and repair the saw cuts with sheet metal and self-tapping screws, being sure to get the key spacing

exactly right. Replace the keys on the board, threading the pivot wire from the bottom end. A similar operation can be performed on the Kimber-Allen keyboard. If you are at all doubtful about cutting up the keyboard, it would

be better to order the F-F version. As the pitch of the mountings varies over an octave, it is not possible to simply shunt the keys down the board.

Parts summary

Power supply (Fig. 8)

Panel fuseholder and 0.5A fuse
Illuminated push-on, push-off switch (0.5A, 250V)
6.3 plus 6.3-V, 1.5-A transformer, or 9-0-9V, 1A (Osmabet)
Bridge rectifier for 30V, 2A
Capacitors — 1000 and 500 μ F 16V

100 μ F 6V
Resistors — 10 and 100 Ω , $\frac{1}{2}$ -watt
Regulator 5V, 600mA (RS Components)
BZY88-C5V6 zener diode
BCY38 transistor
Three-way mains panel socket (Bulgin)
Oscillators and dividers (Fig. 6)
709 or 741 op-amps (twelve)
2N3708 or BC107 transistors (twelve)
Capacitors — 15nF 5% polystyrene (twelve)
Resistors— 18k, 30k Ω , 2%, 20k, 22k, 24k, 27k, 33k Ω , 2% (two each)
10k Ω , 15k Ω , 12k Ω 20% (twelve)
preset pots (pref. w.w.) 3.3k, 4.7k Ω (six each)

7493 i.c. dividers (thirteen)

Preamplifier (Fig. 7)

741 op-amp
Capacitor 50 μ F 6V
Resistors — 1k, 5.6k Ω , 10%
— 10k preset carbon pot

Headphone amplifier (Fig. 7)

2N3708 or BC107 — Tr_{401} , Tr_{402}
OC203 — Tr_{403}
2N3703 — Tr_{405}
2N3705 — Tr_{404}
OA200 silicon diode
Resistors — 8.2, 33 (two), 82, 100 Ω , 10%
 $\frac{1}{2}$ -watt
1k, 2.2k, 5.6k, 10k Ω (two), 10%
10k preset carbon pot

Busbar terminations (Fig. 3)

Resistors — 3 Ω , 5W (R_{506})
3 or 18 Ω , 5W (R_{507})
1k Ω (R_{503}), (R_{502})
22 Ω variable (R_{501})

1A silicon diode (D_{501})

2N3703 or 2N3903 (Tr_{501} , Tr_{502})

Key circuits (Fig. 4)

Capacitors — 10 μ F 10V (sixty one)
50 μ F 6V (sixty one)
Resistors— 1.8k (R_1), 10 or 11k (R_2), 3.3k Ω (R_3)
 $\frac{1}{2}$ -watt, 5% (sixty one each)
see Table 3 for R_4 values (sixty eight needed)
47k Ω (R_5), $\frac{1}{2}$ -watt 5% (sixty one)
BFY50, BFY51, BFY52 or 2N697 (sixty one)
OA200 (D_1 , D_2) silicon diode

Hardware

Keyswitches gold-plated type (sixty two) and keyboard from Elvins Electronics at 8 Putney Bridge Road, London SW18 1HU. Alternative keyboards from Harmonics Ltd, PO Box 32, Chiselhurst, Kent BR7 5RU. Alan Douglas, 4 Lees Barn Road, Radcliffe-on-Trent, Notts.
Veroboard 17 x 5 x 0.1in (three)
Vero pins (200)
Vero track cutter and pin inserter
Edge connector, 16-way 0.1-in pitch (two)
Flex — 10/0.1mm 50 metres p.v.c. ins. (two)
16/0.2mm 5m } p.v.c. ins.
24/0.2mm 3m } assorted colours
Tinned copper wire — 22 s.w.g. 3m
20 s.w.g. 1m
Jack sockets and plugs
Tagstrip — 6-way (two) and 12-way
Pushbutton switch (to stand 100lb load)
 $\frac{1}{4}$ -in grommets
Cable clamps

Electronic piano design

Part 2—assembling circuits and case

by G. Cowie, B.Sc.

The piano is designed so that the 33-in keyboard will just fit inside its case. There is a metal projection at the low end of the keyboard that must be sawn off flush with low C for the keyboard to fit. The keyboard base plate is used as a chassis for the keying circuitry, and the remaining circuitry is mounted in the upper rear of the case.

Before starting work, you should consider the finish for the case. As described, the case is suitable for finishing with paint, leatherette, or, like the prototype, with Formica. If a Formica-type finish is wanted then a lot of time can be saved by using a plastics-surfaced board for parts 2, 3, 5, 7 and possibly 11 (Fig. 9). Some dimensional changes will be necessary, and parts 4, 5 and 6 must be shortened to 33.1in. Exposed wood edges may be painted. Wood parts are fastened together by gluing and dowelling or nailing.

Start by cutting the wood parts to shape, using a rasp if necessary to trim them to exact size. Parts 2 have notches cut in them to fit parts 4 and 6. Parts 2 should be identical, and this can be checked by clamping them together. It is important that the internal width from front to back is a fraction over 33in (say 33.1in) otherwise the keyboard will not fit. This dimension is identical to the length of part 3.

The case is deeper from front to back than strictly necessary, and if you do not intend to put anything other than the circuitry described inside the case, you can reduce the front-back dimension by up to three inches. This involves shortening parts 1, 2 and 7.

Case assembly

Begin assembly by gluing and panel-pinning part 1 to part 4. Add parts 2, and dowel part 4 to parts 2. Next attach part 6 with glue and dowels. (Dowelling should be done by drilling one of the parts to be joined and then, holding the parts in their final position, passing the drill through the hole already drilled to drill the mating part to about an inch deep. Spread glue on the dowel and the mating surfaces, place the parts together and knock the dowel fully in.)

Part 3 rests on top of part 1 and is fixed to parts 2 by dowelling. Tack parts 3 and 1 together with a few panel pins and fill up the crack with glue and strips of wood. Part 3 has one edge planed off at 45°, and the op-

posing edge is also planed at 45° to fit against part 1.

Part 5, the back panel, should be left slightly oversize until fixed in place. It is more convenient to drill the holes, with a bit and brace, before fixing. Three holes of $\frac{3}{8}$ -in dia. and two holes for a mains connector and fuse are required, their positions not being critical. Apply glue to mating surfaces and secure the part 5 with panel pins.

At each step in the assembly check that the parts are squared up. If a rear panel of plastics-coated board is used, cut out a neat aperture about 3×4in before assembly. Subsequently the aperture can be closed by a connector panel fixed to the inner surface of part 5. When the lower case assembly is completed set it aside for the glue to harden.

The keyboard is bolted to two metal

channel sections 13, which in turn are screwed to the hinged upper half of the case. In their normal position, parts 13 rest on part 1, touching parts 2, so that they fix the height of the keyboard and its position; ideally 0.5in clear of parts 2.

Parts 13 are fixed to the vertical front part of the keyboard chassis by two brackets of $\frac{3}{4}$ -in angle, 1½-in long. For each fixing, drill two holes in the keyboard chassis and two in parts 13, the last-mentioned being countersunk. A section of flange of each part 13 must be cut away over a length of three inches to clear the top and bottom C actuator slots and the keyswitch mounting area behind. (To do this with an ordinary hacksaw cut and file away a portion of the flange so that the hacksaw blade can be got into position to cut away the remaining

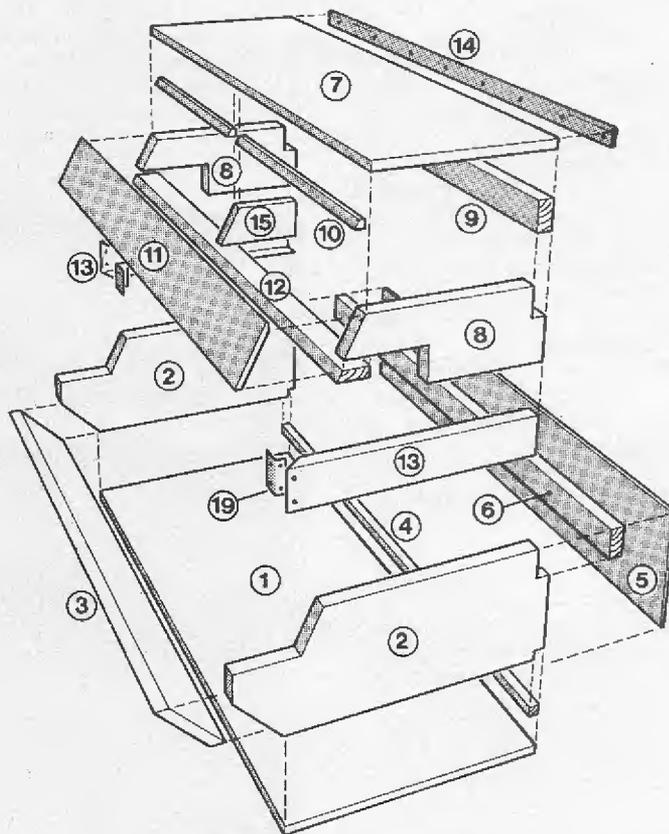


Fig. 9. Top part of the case, comprising nine wood parts and a front panel, carries the keyboard—bolted to parts 13 with angle brackets—and lifts to expose wiring about hinge (14). Parts 16, 17 & 18 are shown in Fig. 10. Dimensions given in parts list.

Notes on components

The pieces of Veroboard for the key circuits boards are 17×5in, as advertised by various suppliers, but which does not appear in Vero's current lists. If the board specified cannot be obtained, a slightly longer and narrower board could be used instead.

The resistors used on the keying boards are $\frac{1}{2}$ -watt because larger resistors would be more difficult to fit in the space available. Similarly the diodes should not be bigger than the Do-7 size, which is about 0.3-in long excluding leads. "Untested" silicon signal diodes were used in the prototype, and proved to be satisfactory. They are very cheap, but it is necessary to test them with an ohmmeter, both to weed out a few faulty diodes from the batch and also to determine the polarity of each diode. The ohmmeter will give a reading characteristic of a silicon junction when its red lead is applied to the cathode (+) end and the diode and its black lead to the other end.

"Untested" transistors may be used to save money but in my experience they are less satisfactory than cheap diodes. It is to be expected that nearly all the batch will be usable for something, but that half will be leaky with a tendency to deteriorate. They should be tested as thoroughly as possible to ensure that the transistors used have a gain of greater than 40 at 100mA and a leakage of less than a microamp or so. The transistors specified are a medium-power type, and low-power transistors should not be substituted in the key circuits. The dividers and op-amps must be good devices, not rejects. They can be bought nowadays for little more than 55p and 25p each, respectively. In the remainder of the circuitry the choice of semiconductor devices is not critical.

Capacitance values for the key circuit capacitors C_1 , C_2 should be adhered to as closely as possible. Voltage ratings are low to minimize bulk and cost, but are not critical provided that C_1 is 10V or more and C_2 is 6V or more. Tantalum capacitors are preferred as the tolerance on capacitance values is closer. On the other hand it is not known what effect the better power factor of the tantalum capacitors (without series resistors) will have on the life of contacts S_1 and the tantalum capacitors are more expensive. Aluminium electrolytic capacitors performed adequately in practice. Essentially the question is one of initial cost versus reliability.

The operational amplifier used for the summing preamplifier should be a 741 as this type is internally compensated. As no compensation is needed for the oscillator op-amps the slightly cheaper 709 type can be used instead. Lead connections for several packages were given in part 1.

scrap section with a longitudinal cut.) Bolt the parts 13 in place, with the nuts inside, and place the whole assembly on a flat surface. Make sure that parts 13 are square to the keyboard, and drill at each end of the keyboard, near the springs, a 6BA-clearance hole so that the chassis plate can be bolted to part 13 flange.

Place the assembly inside the lower case to check that it fits without jamming, and decide on the precise position of the keyboard with respect to part 3. If the keyboard is to be well forward, as in the prototype, then the edge of part 3 must be relieved so that the white keys do not strike it. (It is a good idea to put small packing pieces underneath parts 13 as then it will not matter so much if swarf and shavings collect there.)

On looking at the top of the keyboard, there is an obvious transition between the seen and unseen parts of the black eyes. It is important to know the exact horizontal distance (about .11in) between this point and part 6. This dimension, less 0.05-in clearance, less the hinge thickness, less the thickness of any laminate finish on part 11, is the front-to-back depth of the upper case assembly. If a mistake is made then either the black keys will strike on part 11 or there will be an unsightly gap.

Lay the parts of the upper case, 7-12, together to check that they form a structure of the right size when fitted together. Parts 8 rest on parts 13, and part 7 placed on top should be flush with the top of the lower case. Make sure the gap between pieces 12 and 13 is sufficient to clear the keyboard.

Glue and nail parts 7 and 9 together, and add parts 8, dowelling it to part 9 and dowelling or nailing them to part 7. Plane the front edge of part 12 to a 30° angle and fix it to parts 8 with dowels. Add part 15 with dowels to give some central support to piece 12. Leave this assembly to dry. Check that the upper case assembly fits into the

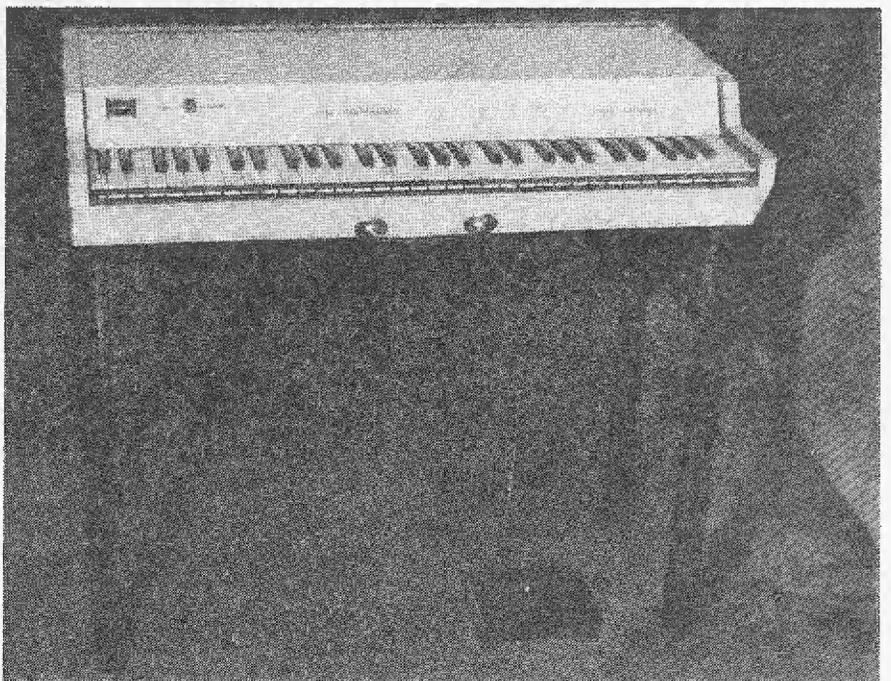
lower case without jamming or leaving an excessive gap at the sides.

Part 11 is thin so that the mains on/off switch and any extra controls can be mounted on it. Fit it by relieving the front of part 7 at 30° or by relieving part 11, leaving some gap to be filled or concealed; the lower edge of part 11 should be relieved at 30° on the inner side. Use glasspaper or a sanding disc to smooth down the mating surfaces for part 11. Attach part 11 with glue and a few panel pins. Use the parts 10, suitably shaped, and some glue to fill and strengthen the join between parts 7 and 11. Put the assembly aside to dry upside-down so that parts 10 will stay in place. (At this stage the upper and lower case halves may be painted on the inner surfaces.)

Attach two lengths of $\frac{1}{2}$ -in or $\frac{3}{8}$ -in square wood (parts 16 and 17, Fig. 10) to the keyboard chassis plate to act as bearers for the key circuit boards, which are to be $\frac{7}{8}$ to 1in below the main plate. Fix part 16 by woodscrews to the front vertical plate, which must be drilled for four fixing holes. Secure part 17 by woodscrews at its ends to parts 13; a countersunk hole is required in each part 13. Drill the main plate in two places by the springs, about one-third and two-thirds of the length of the board, so that packing pieces can be fixed in place to stiffen part 17. The keys scratch easily so protect them during drilling.

The desired surface finish is most conveniently applied to the outside of the upper and lower case at this stage. In the prototype all visible surfaces are white Formica except for the inner surfaces of parts 2 and the top edge of part 3, which are black.

Various accessories are next fitted to the case. Fit the three jack sockets, the mains connector socket, and the mains fuseholder to the rear panel. Next fit the mains switch and indicator lamp to the front panel. Attach the piano hinge to the upper case, with the axis rod above the top surface.



Footswitch for sustain action is simply a push-button switch embedded in a wood block.

Arrange the upper case in the fully open position, supported on a block, and screw the hinge to the lower case. This is trickier than it seems and it is best to put in two small screws first so that the assembly can be closed to check whether it is all lined up and fits properly. If all is well then the hinge may be securely fastened by $\frac{1}{2}$ -in counter-sunk-head brass screws.

Screw parts 13 to parts 8, thereby fixing the keyboard into the upper part of the case. Screw an eyelet into each side of the case, inside, next to part 3, and provide a similar fixing on the keyboard chassis. Permanently attach two restraining cords to hold the upper case open in a suitable position for working inside. Check that the assembly still closes. Fit the case carrying handle and add metal ties between parts 3 and 1 at the handle location. Two large self-tapping screws are used to hold the case closed. Turn the case upside-down and drill two holes near the front of the case in such a position that they pass through the flange of part 13. These holes must be of the root diameter of the self-tapping screw. Next drill the wood to the clearance diameter, grease the screws and screw them into place by repeatedly making one turn in and half a turn out.

Circuit assembly

In one sense the assembly work is very simple as there are only five major circuits involved, none of which has more than a dozen components. There is, on the other hand, a great deal of work to be done, and very boring and repetitious work at that. To give some idea of the time involved, fitting 60 components on boards will take $1\frac{1}{4}$ to $1\frac{1}{2}$ hours from unpacking to checking the soldered joints, and putting in the 183 key wires will take at least a day. Clean soldering and general neatness are very important, greatly increasing the chances that the circuitry will work as intended.

Make up the power supply, oscillators, and amplifiers first, so they can then be used to test the key circuits.

No detailed assembly is shown for the power supply as it is a simple wired unit and the original was made from junk. For safety it is best to make the power supply on a metal baseplate and completely enclose it in an earthed metal case. Mount the +5-V regulator on a small heatsink of $\frac{1}{16}$ -in thick aluminium forming the lid of the case. As the case of the regulator forms its 0-V connection it should be isolated from the heatsink with the standard kit of mica washer etc, so that the 0-V line can be floated or grounded as necessary. Make sure that no live terminal is within $\frac{1}{4}$ -in of anything on the low-voltage side. Earth any metal parts of the on/off switch showing on the front panel, as well as the metal case of the power supply. Mount the unit in the lower left-hand part of the case.

Keyswitch modification

The keyswitches as bought are of the triple normally-open type, not normally-closed as required. This is of no consequence as it is a simple matter to hook the bent wire of each pair around the straight wire, so that the switch becomes normally-closed (Fig. 11).

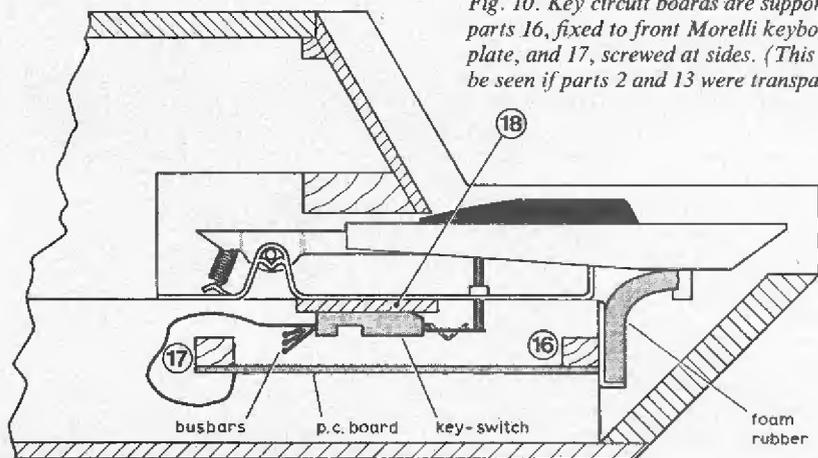


Fig. 10. Key circuit boards are supported by parts 16, fixed to front Morelli keyboard plate, and 17, screwed at sides. (This would be seen if parts 2 and 13 were transparent.)

Fig. 11. Keyswitch as bought is normally-open (top) and is simply modified to normally-closed (bottom).

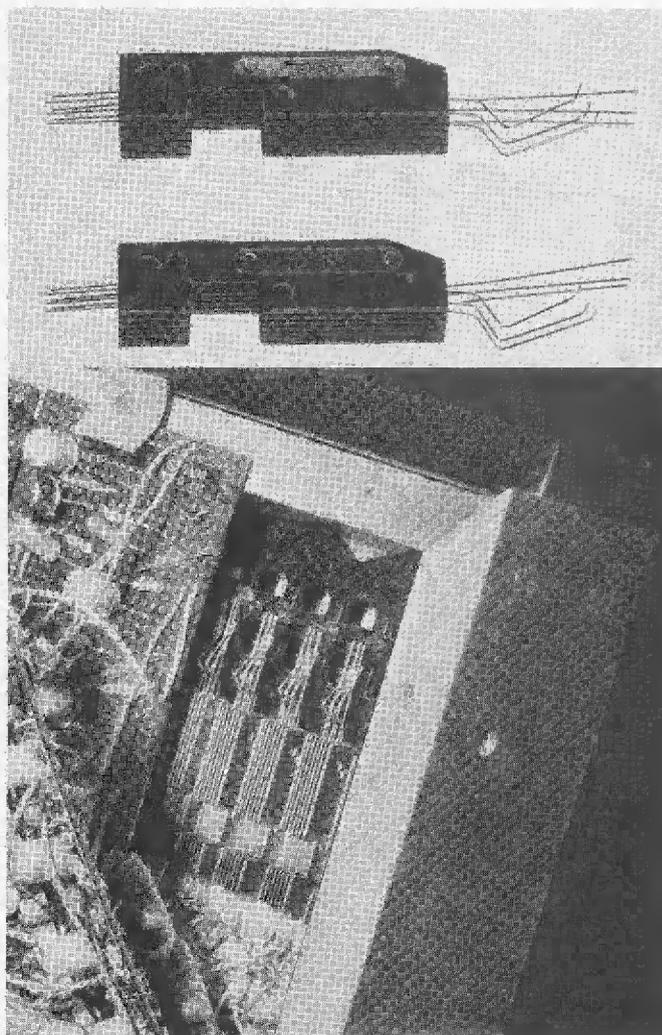


Fig. 12. View of underside of keyboard (with lid raised) shows key circuit boards with dividers and, beneath, keyswitches and actuators.

It is less easy to arrange for the contacts to open in the correct sequence. In the prototype this was done by mounting the keyswitches and then bending both wires of all three pairs until the desired result was achieved. As the differential between S_1 and S_2 opening is fairly critical, and there are 366 wires in all, this was tedious. I therefore devised a procedure which is not only less laborious but should give more satisfactory results.

Under each key is a plastics actuator rod

capped by a rubber boot which is designed to push the straight key switch wires (see Figs 12 & 13). A notch 0.1-in deep must be effectively produced at one side of each pusher, and the simplest way of achieving this seems to be to stick a tiny square of plastics sheet, not to the rubber boot, but to the two wires that must move first (Fig. 13). This should be 0.10-in (2.5-mm) thick, and about 0.15-in (4-mm) square. One side of the keyswitches has a large rectangular notch which exposes the wires. This is the visible side, and the opposite side is the

mounting side. The plastics squares of course go on the mounting side of the wires.

Simplest way to attach them is to apply a little glue with a matchstick to the ends of the two wires, touch them on the square to pick it up, adjust it carefully for position, and leave the assembly upside-down to set. Contact adhesive should stick most materials; alternatively a solvent adhesive or Araldite could be used.

The keyswitches must be mounted on a spacing piece (part 18, Fig. 10) stuck to the main plate of the keyboard chassis. The thickness of the spacer must be determined before the keyswitches are mounted. Find a piece of packing which will support one of the keyswitches in an operating position so that the rubber actuator is almost touching the plastics pad. This should ensure that when the key is pressed, the actuator opens first two contacts almost together, and the third before reaching the end of the travel (the differential is 0.10in). A piece or pieces of packing, preferably of plastics, two inches wide and in all 33-in long are stuck down with contact adhesive, and the keyswitches are stuck to this with contact adhesive. They should be as far back as possible consistent with correct operation; and they can be placed with sufficient accuracy by hand.

Oscillator assembly

The twelve oscillators (Fig. 14) are identical except for the tuning resistors (see Table 3 in part 1). The frequencies used depend on whether a C-C or F-F keyboard is chosen: 2093 to 1108Hz and 1397 to 739.8Hz respectively.

The layout shown in Fig. 15 is designed to enable six oscillators and six buffer transistors to be assembled on a 2½ × 5 × 0.1in pitch Veroboard. In the prototype two of these were mounted on plug-in carriers so that the oscillators could be removed easily for tuning and repairs, but to make things as simple as possible I recommend that all twelve oscillators and the buffer transistors be mounted on a piece of Veroboard of at least 5 × 5½in which is wired for plugging into a suitable edge connector with a minimum of 16 ways (that is twelve outputs, three power lines, and a space for a polarizing/locating key).

Use the layout given, placing the oscillators in four rows of three. The order is not important, but it is helpful if they are arranged on the board in strict alphabetical order. Break the copper tracks where shown with a Vero spot face cutter or drill. All tracks should be broken between oscillators except the three power lines, and the four tracks under the i.c. must also be broken. Wire resistor R₂₀₅ to the buffer transistor.

The integrated circuits may be either the 709 or the 741; see part 1 for pin connections. Note that the twelfth or eighth pin is not connected in the 741 but is internally connected in the 709 and so no connection must be made to it. Either skeleton or button-type preset pots may be used, the skeleton type being easier to mount. Resistors R₂₀₁ to R₂₀₄ should be 2%, ½-watt types, and the capacitors polystyrene or polyester. The component leads may be left a little long to stand them off the board. The presets form a small part only of the tuning

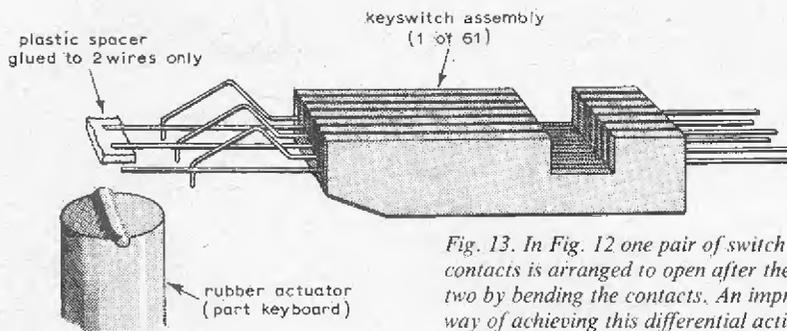


Fig. 13. In Fig. 12 one pair of switch contacts is arranged to open after the other two by bending the contacts. An improved way of achieving this differential action is to glue pads to contacts as shown.

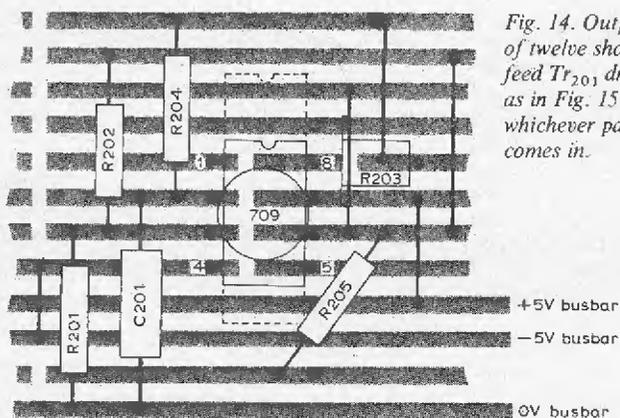


Fig. 14. Output from the oscillators (one of twelve shown) is taken from R₂₀₅ to feed Tr₂₀₁ drivers, which can be arranged as in Fig. 15. Board wiring is the same whichever package the 709 or 741 i.c. comes in.

resistance so their stability is less important.

Mount the edge-connector of the oscillators in the upper right part of the case, and arrange a clip to hold the free end of the oscillator board.

Amplifier assembly

The summing preamplifier and headphone amplifier (Figs 16 & 17) can both be assembled on a 2½ × 5in piece of Veroboard, of 0.1-in pitch, and arranged to plug into a suitable edge-connector. An internal connection must be made between R₄₀₄ and the input of the headphone amplifier, and the external connections are the three power lines, two inputs, low level output, and headphone output. A 741 op-amp must be used in this circuit.

Mount the edge-connector for this board in the upper right part of the case, together with a 12-way tagstrip.

Wire the three power lines to the tagstrip, using fairly thick stranded wire (24/0.2mm). Connect a lead for the +8-V line from the power supply to the tagstrip, and check that none of the wires can be trapped when the case is closed. Mount the dropper resistors R₅₀₆ and R₅₀₇ on the tagstrip. Mount the components R₅₀₁₋₃, D₅₀₁, Tr_{501/2} on the tagstrip, making connections to the power lines. Two of the tags are also terminals for the bias and damper lines. Wire the amplifier edge-connector to the two output sockets. The low-level socket should be wired conventionally, but the headphone socket should be a three-way socket with the 0-V line wired to the middle contact and the signal to the inner contact, so that normal connection is made to a two-way jack, but the coils of stereo headphones are con-

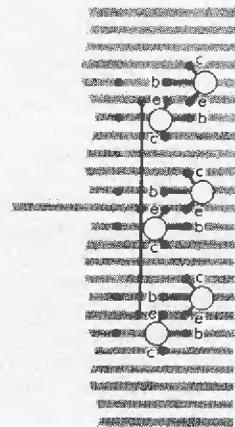


Fig. 15. Twelve top-octave driver transistors (Tr₂₀₁, six shown) can be arranged to enable six oscillators and drivers to be accommodated on 2½ × 5in Veroboard or, preferably, all twelve on a 5 × 5in board.

nected in series. Wire the sustain pedal connection from the tagstrip to the sustain pedal socket. Connect power lines to the oscillator and amplifier edge-connectors from the tagstrip. If the power supply uses a 9-volt transformer then R₅₀₇ must be about 18 ohms to drop the raw direct voltage to the 8V required for the switch busbar, and for a 6.3-volt transformer it must be about 3 ohms.

Key circuit assembly

The key circuits are assembled on two large pieces of 0.1-in pitch Veroboard, one 5 × 15in, the other, containing the odd key circuit, 5 × 16½in. The overall length is calculated to be just less than the distance

between parts 13 flanges. The 12 groups of key circuits are identical, and the 61 key circuits are themselves identical save for the connections of diode D_1 . The assembly details are given in Fig. 18 which shows the divider and the top key circuit. The whole group should be exactly five inches long, the pitch of the key circuits being 0.9in.

Mark each circuit board with two rows of three $2\frac{1}{2} \times 5$ in rectangles, using a scribe. Mark within these the areas occupied by the key circuits. Each occupies a width of eight holes, the boundaries between key circuits falling on the ninth rows of holes where several tracks must be cut to electrically isolate the circuits. Fit the divider i.c.s in their correct positions, at the top of each rectangle and one hole from the left-hand edge. Use Veroboard pins or some cheaper equivalent to attach the key wires, three to each key circuit.

At the assembly is quite a large operation, much time will be saved if it is approached methodically, for instance by taking 61 components (e.g. R_3), bending all the leads, fitting to the board, bending the leads flush with the tracking, clipping the leads to 0.1in from hole, and then soldering.

Each group of key circuits must be assigned to a letter, C, A etc. and the resistors R_4 fitted, following Table 3. The lowest values go with the highest pitch notes, and so the key circuit nearest the divider will have R_4 of 1k, 1.2k or 1.5k Ω . The positions of diodes D_1 vary; in the no. 1 key circuit the cathode goes to track 1 (pin 14 of i.c.); in no. 2 to track 3 (pin 12); in no. 3 to track 6 (pin 9); in no. 4 to track 7 (pin 8) and in no. 5 to track 4 (pin 11).

The lowest pitch key circuit takes the form of a small 13th letter-group, in physical arrangement. It fits on the extra $1\frac{1}{2}$ -in of board next to the C letter-group. The divider could be almost any t.t.l. flip-flop but I recommend a 7493, mounted on a dual in-line socket so that it can be changed easily. The connections are easily deduced: pin 14 is wired to track 4 of preceding stage, D_1 cathode goes to track 3.

Lines for +5-V, 0-V, output, and collector busbars run the length of each board. Three tracks should be commoned for the ground busbar to reduce the resistance.

Key wiring

Six wires about $\frac{1}{2}$ -in long extend from the back of each keyswitch (Fig. 19). Bend down the extreme right-hand wire of each switch (looking from the rear), and solder to them a bare tinned-copper wire of about 22 s.w.g. Inspect this work and cover the wire (switch busbar) with a length of adhesive tape. Do the same for the third wire from the right (left, as shown), and cover the damper busbar with tape. The third, bias, busbar should be of slightly thicker wire, and is soldered to the gold wires as shown. The bias contacts are the pair that open last. Bend the remaining gold wires upwards and outwards to make it easier to connect the flex wires.

To mount the two key circuit boards, use six small nails, bent over at 45°, in part 17, on which the boards rest. To hold the boards while the case is closed, drill four holes in

Fig. 16. Summing preamplifier and headphone amplifier (Fig. 17) can be included on the same board.

Fig. 17. Headphone amplifier, together with R_{404} , could be omitted if an external amplifier is used.

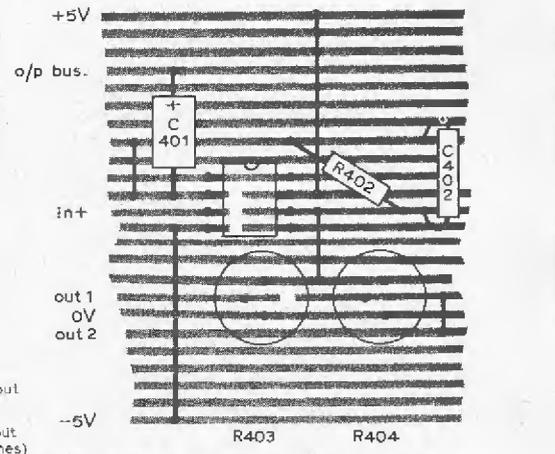
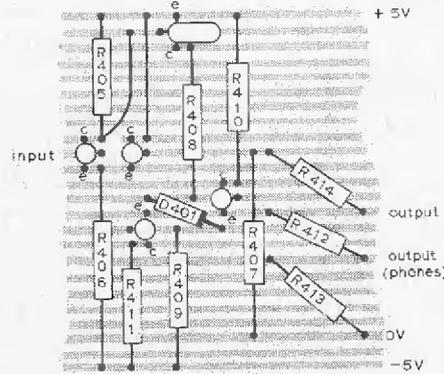


Fig. 19. Three busbars alongside the keyswitches are insulated by tape, the "switch" busbar being soldered and insulated first.

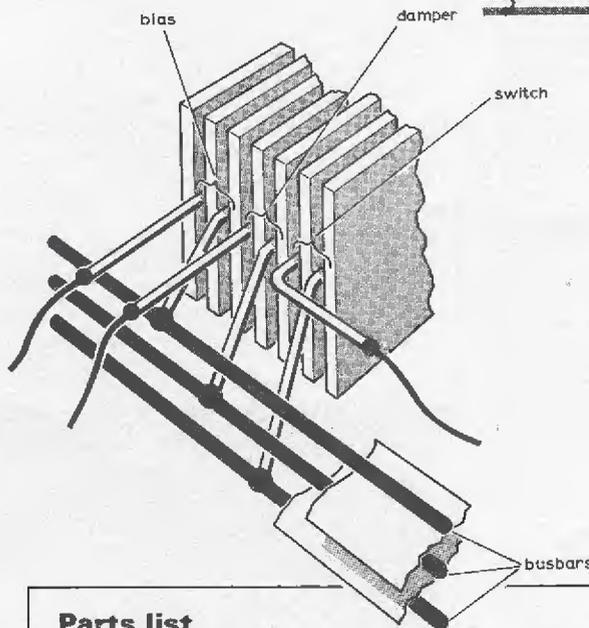


Fig. 18. Layout of one divider and key circuit. Outputs from divider feed four similar (except for D_1 position) key circuits alongside that shown. (Thirteenth divider feeds only one key circuit). Two large boards bear 61 key circuits and 13 dividers. In the five key circuits related to any one divider, uppermost wire of D_1 connects to different i.c. terminals or tracks (see text). A minor modification to this board is suggested in part 3.

Parts list

- 1 Plywood $34\frac{1}{2} \times 16 \times \frac{1}{4}$ in
- 2 Plywood* $18 \times 6 \times \frac{3}{4}$ in (two)
- 3 Ply or hardwood* $33 \times 3\frac{1}{4} \times \frac{3}{4}$ in
- 4 Deal $34\frac{1}{2} \times \frac{1}{2} \times \frac{1}{2}$ in
- 5 Ply or hardboard* $34\frac{1}{2} \times 6\frac{1}{2} \times \frac{3}{16}$ in
- 6 Ply or hardwood $34\frac{1}{2} \times 1\frac{1}{4} \times \frac{3}{4}$ in
- 7 Plywood* $33 \times 9\frac{1}{2} \times \frac{3}{8}$ in
- 8 Plywood or deal $10 \times 3\frac{1}{2} \times \frac{3}{4}$ in (two)
- 9 Deal $33 \times 1\frac{1}{2} \times \frac{3}{4}$ in
- 10 Deal $15 \times \frac{3}{8} \times \frac{3}{8}$ in (two)
- 11 Ply or hardboard $33 \times 3 \times \frac{3}{16}$ in
- 12 Deal $33 \times 1\frac{1}{2} \times 1\frac{1}{2}$ in
- 13 Aluminium or steel channel $13 \times 2 \times \frac{3}{8}$ in (two)
- 14 Piano hinge 33in
- 15 Plywood or deal $4 \times 1\frac{3}{4} \times \frac{1}{2}$ in
- 16 Deal $31 \times \frac{1}{2} \times \frac{1}{2}$ in (or $\frac{3}{8} \times \frac{3}{8}$ in)
- 17 Deal $33 \times \frac{1}{2} \times \frac{1}{2}$ in
- 18 Plastics spacer 33×2 in (see "keyswitch modification")
- Dowel rod $2m \times \frac{1}{4}$ in
- Aluminium angle-bracket $2\frac{1}{2} \times \frac{3}{4}$ in
- Deal block to house sustain switch $6\frac{1}{2} \times 5\frac{1}{2} \times 1\frac{1}{4}$ in
- Perspex 4 sq in $\times 0.1$ in thick
- Carrying handle and bolts
- Aluminium plate $\frac{1}{16}$ -in for heatsink TO-3 insulator kit
- Metal box for power supply
- Deal for stand: $26 \times 1\frac{1}{2} \times 1\frac{1}{2}$ in (four)
- $16 \times 6 \times \frac{1}{2}$ in (two)
- $34 \times 5 \times \frac{1}{2}$ in

*Parts 2, 3, 5 & 7 could alternatively be Contiboard

part 16 to take retaining screws with large washers, which overlap the boards.

Key wiring must be done with miniature stranded flex wire. Connect a wire to the switch contact, top C. Pass the wire through the gap between part 17 and the keyboard plate and lead it along part 17, then up through a right angle towards the connect-

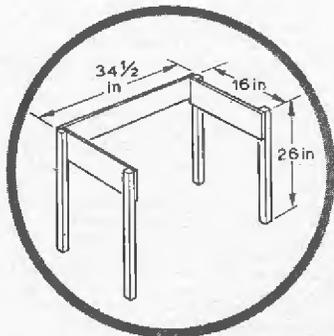


Fig. 20. Simple stand for the piano can be made from deal.

ing point—the pin on track 11 of the top C key circuit. Similarly, connect the damper wire to the pin on track 2, and the bias wire to the pin on track 18. Place cable ties at intervals along part 17 to hold the wires and tie wires together to the board in groups of six where they pass up over the lower edge of the boards.

When several keys have been wired up check whether the system works. Wire the bias, switch, and damper busbars to the tag-strip, and connect the ground 0-V line to each key circuit board. Check that the power supply delivers the correct voltages and connect the +5-V and collector busbar lines to both boards. Connect oscillator signal lines for the letter-groups in use, plug in headphones, connect the output busbar to the amplifiers, temporarily ground the positive input of the summing preamplifier, plug in the oscillators and amplifiers, and switch on. If the wired keys do not work look for faults as detailed later (part 3).

The remaining key wires may be put in when you are satisfied that the system is more or less in working order. The wires should be cut a little long, leaving about two inches of slack.

Electronic piano design

Part 3—tuning the touch-sensitive design, an m.o.s. master oscillator and other optional circuits.

by G. Cowie, B.Sc.

When you have completed the wiring of the piano the remaining work is to commission and tune it. In a project of this complexity it is inevitable that a number of faults and wiring errors will occur and the purpose of the commissioning stage is to weed these out and make various adjustment.

Start by applying power and making the following voltage checks

- supply busbars, 0V, +5V, -5V
- collector busbar, about 0.5V
- switch busbar (after R_{507}), 7 to 8V
- damper busbar, about -0.4V
- bias busbar, about +0.8V
- output busbar, +0.4V.

Turn both potentiometers on the amplifier board to mid-position, and check that the pre-amp and amplifier outputs are about 0V before proceeding further.

Connect headphones or an audio amplifier to one of the output sockets. A hiss and probably some whistles will be audible but ignore this for the moment. If nothing is heard, check the output amplifier and preamplifier. Check that a note is heard

when each key is struck, and that the pitches are more or less in the right order, bearing in mind that the oscillators are not tuned.

The system is organized in such a way that with a little thought any fault evident can be localized to one or two connecting lines, or a few components either in a letter-group or in a key circuit. Absence of signal from a key circuit will most likely be caused by absence of input signal, resulting from a reversed diode, or a misconnection at a divider or oscillator. Whistles, traceable by their pitch, will be caused by leaky transistors or bias connection faults, or damper faults. Faint notes will be caused by collector open-circuits. Open or bent key-switches will show up as faults. The sort of defects to look for are missing wires, missing links near i.c.s, bad soldered joints, assembly mistakes, solder bridges, parts touching where they should not touch, tracks not cut. It may possibly be necessary to add power supply decoupling capacitors on the various circuit boards.

If the power supply uses a 9-volt trans-

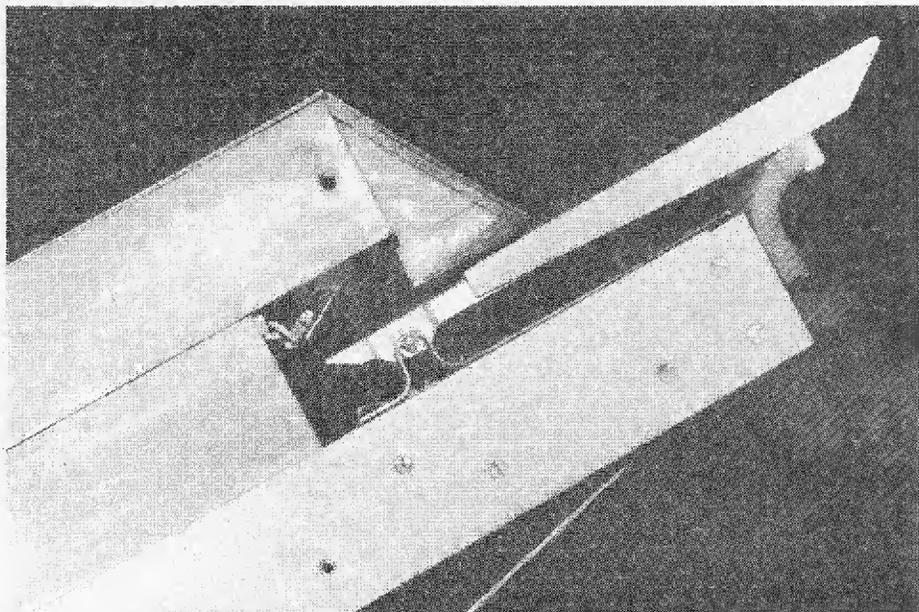


Fig. 21. To simulate the inertial load of the keying action, a piece of foam plastics material is inserted between the key undersides and the lower flange on the keyboard chassis.

former, R_{507} must be increased to about 18 ohms to drop the raw direct voltage to the 8V required for the switch busbar.

Check that each note ceases when its key is released, then short-circuit the sustain pedal socket and check that the notes now are sustained after the keys are released.

When all the circuits have been made to work, several finer adjustments can be made. Disconnect the positive-signal input of the summing preamplifier from 0V and extend it from the edge-connector with about 30in of flex wire. Solder the wire to the end of resistor R_{401} and then attach R_{401} to the 0-V busbar on the key circuit boards, in a position that gives minimum pick-up of unwanted signals. Run the positive signal lead alongside the output busbar as far as possible. It is important to arrange the 0-V lines carefully, to minimize stray signals. Connect a lead directly between the key circuits 0-V busbar and the preamplifier ground. Hiss should be inaudible while the instrument is being played.

If there is still too much background hiss,* then reduce the value of R_{501} , until the hiss is reduced to a low level. If the resistance is reduced too much there will be an undesirable effect whereby the keys refuse to work at all when pressed down very slowly.

If the damper action is felt to be too abrupt then the potential of the damper bus can be raised by connecting a resistance of a few hundred ohms in series with R_{502} , and connecting the base of Tr_{502} to the junction of the resistors.

If the action of any of the keys is noticeably different from that of the majority, the cause should be looked for and put right, otherwise it will be a permanent irritation when playing the instrument.

It is important to alter the light spring-loaded feel of the keys if the touch-sensitive action is to be satisfactory. The ideal arrangement would compromise 61 viscous dampers simulating inertial loads, but in the prototype a much simpler device was adopted which proved to give an acceptable result.

Cut a piece of flexible foam plastics material to dimensions of $2 \times \frac{3}{8} \times 33$ in. If a piece 33-in long cannot be found then use shorter lengths. Push this into place between the white key undersides and the lower flange on the key chassis (see Figs 10 & 21). In this position the black keys will rub against the foam and the white keys will deform it so that a drag is felt on the keys. Key action will not be satisfactory unless this is done.

Tuning

The easiest and most satisfactory way of tuning the oscillators is to remove the oscillator board from the instrument and take it to a laboratory which has a suitable digital timer/counter and d.c. power supplies. The oscillators should be connected to a ± 5 -V supply. They should be moni-

tored at the op-amp output where a clean square wave is present. It is advisable to use an oscilloscope to look at the waveforms and see that all is well.

The oscillators can then be adjusted to the frequencies in Table 1 (part 1) by the tuning pots to an accuracy of ± 2 parts in 1000. If an adjustment is outside the span of the potentiometers it will be necessary to add padding resistors. Label the oscillators after they are set so that they can be wired up without confusion.

If you do not have access to the above equipment, either through work or friends, the oscillators must be tuned *in situ*, by ear. The person doing the tuning must have a musical ear and be equipped with an "A" tuning fork. The piano case must be open so that the tuning pots are accessible. Tune the A group of notes against the fork. Notes F are next tuned against the A; the interval should be a major third. Next tune C by means of the major chord F-A-C, and adjust F again if necessary. Use C to tune E, the interval being a major third, then tune G in the major chord C-E-G. Use G to tune B (major third) then tune D using the major chord G-B-D. This completes the white notes.

Tune E_b using the chord F-A-C- E_b and B_b using the major chord E_b -G- B_b . Tune A_b using the chord B_b -D-F- A_b or the major third C- A_b . Tune $F^\#$ using the major

third D-F $^\#$. Tune C $^\#$ using the major chord A-C $^\#$ -E. You will probably need to go over the process several times. Fortunately the electronic piano is much easier to tune than a string piano, which has over two hundred adjustment points.

If you use a m.o.s. top-octave synthesizer i.c., only one tuning adjustment will be necessary, and will take only a few minutes (see later).

Circuit options

On the key circuit boards (Fig. 18), remove the links between track 24, to which resistors R_2 are grounded, and track 23. Link the tracks by a silicon signal diode on each board, connecting the cathodes to track 23. This modification is based on work done after the prototype was completed, and if carried out should slightly improve the keying action at the "soft" end of the dynamic range.

The prototype is a basic instrument in that no attempt is made to modify the inherent tone of its working waveform, except by a single low-pass filter. It is purely by coincidence and the laws of physics that it sounds as much like a harpsichord as anything. If this tone is not acceptable then additional wave-shaping can be added fairly easily. There is room on the keying boards to fit a low-pass filter between D_1

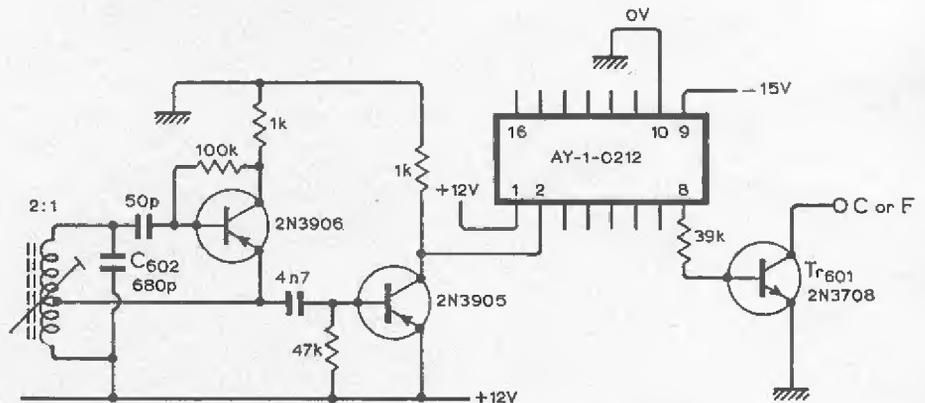


Fig. 22. This alternative tone generator, which can replace the twelve RC oscillators used in the original design, avoids conventional tuning procedures.

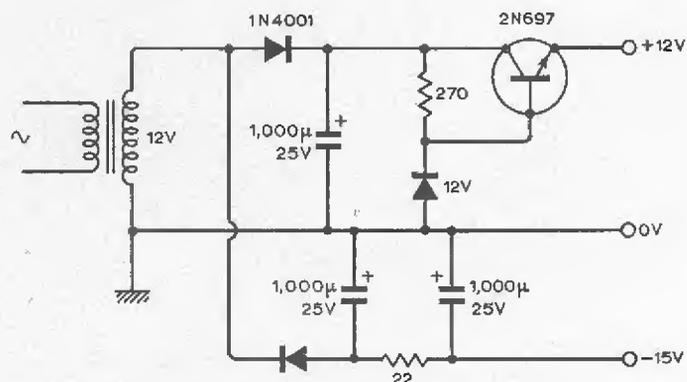


Fig. 23. Power unit to provide appropriate supplies for the m.o.s. tone generator also feeds the preamplifier and headphone amplifier. If the m.o.s. circuit is used, the -5V supply in Fig. 8 (part 1) can be omitted.

*If a hiss present while the piano is not being played is annoying, it can be eliminated with a "squelch" circuit, given later.

and R_5 , I fitted filters on the 16 lowest notes to take harshness out, with resistors of $10k\Omega$ and capacitors of $20nF$. More complex circuits could be built on a separate board, linked by a 61-way cableform.

A power amplifier could be fitted inside the case, but as this would need a separate power supply it does not seem worthwhile. Similarly, loudspeakers could be fitted in the case, but as fairly large speakers will be needed to reproduce the bass notes properly, and as the case is not designed as a loud-speaker enclosure, this is not recommended.

A tremolo circuit could easily be fitted after the preamplifier. It is not known whether vibrato can be applied to the oscillators successfully, and anyway this seems inappropriate.

The headphone amplifier could be replaced by an integrated circuit. The MFC 4000 is a cheap i.c. designed for the output stage of transistor portable radios.

A board containing twelve $L-C$ oscillators could be made as a plug-in replacement, but as discussed in part 1, relaxation oscillators are preferred for several reasons.

Amplification

For low-power amplification, as for practising, almost any amplifier and speaker can be used provided that the speaker is of adequate size. A speaker of 8in diameter should give good results, bearing in mind

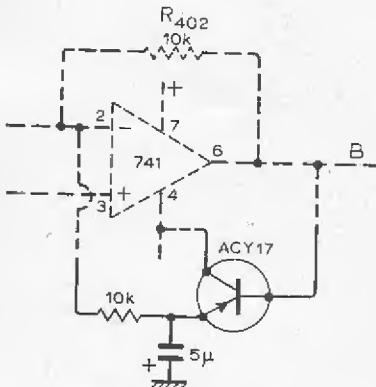


Fig. 24. Optional "anti-thump" circuit comprises three additional components to the circuit of Fig. 7.

that the bass notes have more fundamental than those of an acoustic piano. If high powers are wanted then it is essential to use heavy-duty speakers of the type used for electric guitars. The speakers should have a higher nominal rating than the amplifier otherwise the percussive piano waveform will probably damage them.

A standard volume control pedal may be bought and fitted between the piano and amplifier, and R_{403} preset to a suitable level.

M.O.S. master tone generator

The r.f. oscillator and AY-1-0212 master tone generator replace twelve oscillators (Fig. 22). The frequencies generated are within 0.1% of an equal-temperament scale, so the piano will work perfectly well without being tuned at all! It can easily be tuned against another instrument or a frequency counter if desired. It would be possible to arrange the circuit for rapid transposing; this is impracticable with twelve oscillator generators. It is still necessary to have twelve interface circuits to the t.t.l. dividers.

An extra power supply is necessary also. The circuit shown (Fig. 23) provides regulated +12V and smoothed unregulated -15V for the AY-1-0212. The -5V supply of the original piano design is omitted. The preamplifier and headphone amplifier should be powered from the +12V, -15V power supply.

A coil of the 470kHz, i.f. type, with a 2:1 turns ratio is needed. It may be necessary to change C_{602} to get the frequency into adjustment range of the ferrite screw core. The circuit will oscillate with any coil of about the right turns ratio.

The AY-1-0212 will stand a certain amount of abuse but should nevertheless be treated with care. It should be mounted in a 16-pin dual in-line socket. The one-off price is about £5.90. A Veroboard of 3×4 in should give ample room for the oscillator, i.c. and the transistor/resistor array. It is advisable to bring the connections out at the edge of the board in scale order; the i.c. outputs are in a random order.

If high-frequency noise causes trouble on the lower octaves, additional power line filter capacitors should be fitted on the cir-

**Parts for
m.o.s. tone generator**

AY-1-0212 i.c. from General Instrument Microcircuit Sales, 57 Mortimer Street, London WIN 7TD.
470-kHz i.f. transformer, 2:1 turns ratio, with ferrite core
2N3708— T_{r601} (twelve)
2N3906
2N3905
Capacitors—50pF, 680pF (C_{602}), 4.7nF
Resistors—1k (two), 100k, 39k Ω (twelve)
16-pin d.i.l. socket

Power unit (Fig. 23)
12-V, 50-mA transformer
Capacitors—1000 μ F, 25V (three)
Resistors—270, 22 Ω
1N47001 diodes (two)
BZY88—C12V zener diode
2N697 or BFY52 transistor

AY-1-0212 connections

1 $V_{SS} + 12V$	9 $V_{GG} - 15V$
2 in 500 or 334kHz	10 $V_{DD} 0V$
3 C^* or F^*	11 A^* or D^*
4 E or A	12 A or D
5 G or C	13 F^* or B
6 G^* or C^*	14 F or A^*
7 B or E	15 D^* or G^*
8 C or F	16 D or G

cuit board, i.f. capacitors can be fitted at the T_{r601} collectors or bases.

Squelch and anti-thump circuits

Some degree of "thump" will always be present where waveforms are not generated symmetrically about ground potential. Fig. 24 shows the anti-thump circuit now fitted in the prototype.

Hiss due to leakage though the reversed-biased diodes D_1 is very low, but can be completely eliminated by the "squelch" circuit of Fig. 25. Connect between points B-B in Fig. 7.

Letters to the Editor

Tuning electronic pianos

Readers who are tempted to construct the novel design of electronic piano by G. Cowie, and who try to use the tuning method outlined in the third part of his article (May), will find themselves faced with a very difficult task. Tuning any instrument to equal temperament is not easy, but the job is straightforward, if a little arduous, if the method used by professional piano- and organ-tuners is employed.

The main difficulty arises because the major third is such an imprecise interval.

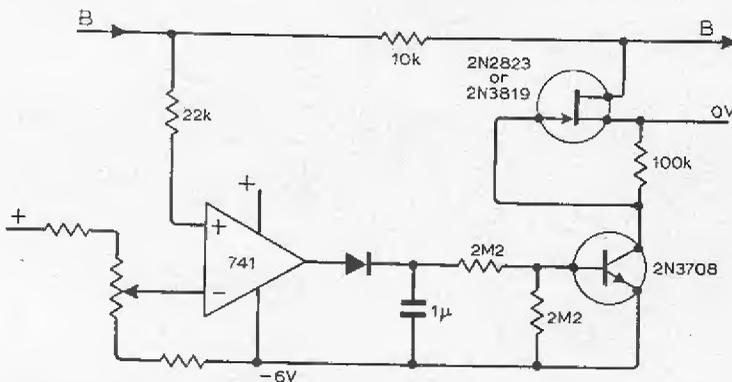


Fig. 25. Use this squelch circuit to eliminate "hiss" due to leakage currents that may be evident during standby. Suitable values for potential divider are $10k\Omega$ and $5k\Omega$ for the preset resistor.

An instrumentalist who can vary the pitch of his instrument (a violinist, for example) will say that he varies the interval of his major thirds (and minor ones, for that matter) according to the mood and tension of the music; yet the interval still sounds in tune. The same cannot be said of the mathematically defined intervals of an octave, fifth or fourth. I put these intervals in that order deliberately, because we can tolerate mis-tuning least in the octave, more so in the fifth, and most in intervals like the third, sixth, second and seventh.

This suggests that we should tune a scale by means of the intervals to which the ear is most acutely sensitive, the octave (I am neglecting the unison, since it is not relevant to laying a scale) and fifths and fourths. This sounds simple enough, but this is where the arduous part comes in. Because in the Western world we are used to music in dodecaphonic scales (containing twelve intervals) we like to be able to use these intervals freely to construct scales starting on any key-note. It is also convenient to be able to modulate from key to key at will without losing proper tuning, and herein lies the problem. A scale of C major, for example, constructed mathematically by using the exact ratios of the octave (1:2) and the fifth (2:3) has twelve intervals which are not all equal, and when these are used in other keys, the different sized intervals occur in the wrong places, making the scale out of tune. In equal temperament, all intervals are theoretically made equal. Like George Orwell's animals, some turn out, in practice, to be more equal than others, but as an ideal to aim for, it is satisfactory.

In order to make the intervals equal, the intervals are "bent" slightly, the fifths being made slightly flat and the fourths slightly sharp. Octaves are of course tuned exactly. Around middle C, the amount of this flattening or sharpening is a little less than one beat per second on average, and takes a careful ear to hear and quantify it. It is easiest to notice if the interval is first tuned exactly (no beats apparent) and then narrowed or widened slightly accordingly.

The method, then, is as follows: tune A (440Hz) exactly with a fork; take the A below middle C (which the divider circuits have tuned for you) tune up to E, up to B, down to F sharp, down to C sharp, up to G sharp, down to D sharp, up to A sharp (B flat), down to F, down to C, up to G, down to D, and if you are lucky, up to A. The first time round, you will probably find you have overdone the interval "bending", but work back in the opposite direction and adjust slightly until you have laid a perfect, equally tempered scale, then celebrate by playing the entire 48 Preludes and Fugues by J. S. Bach, written for just such a keyboard instrument.

Interested readers can find an excellent article on tuning and temperament in Percy Scholes' *Oxford Companion to Music*, which gives more details than I can.

David K. Taylor, Charlbury, Oxford.

Tuning the electronic piano

I would like to suggest the following method of tuning any locked-divider electronic polyphonic keyboard instrument.

1. Ensure vibrators or tremulants are "off". Tune middle A to 440Hz, using a fork or BBC test tone transmission.
2. Tune the D below so that it sounds a perfect fifth (zero beat) with the A440. (This should be easily recognized as the commonly heard violin tuning procedure.)
3. Sharpen the D (increase frequency) until it beats with the A at approximately 1 beat per second.
4. Now play the D and the G below it and tune the latter (first to a perfect fifth, then sharpened to 1 beat per sec.).
5. Continue sounding the G *one octave above* (392) and tune the C below in like manner.
6. Confine the sequences to the middle octaves (by jumping up one octave as required) and carry on tuning in fifths (adjusting the lower note each time) until the final interval is reached, which is E (659) sounded with the fifth below (your original A440). This should sound like the other intervals (a perfect fifth slightly diminished).

The musical reason for diminishing each interval is simply that all modern keyboard instruments are "equal temperament" tuned so that they can be played in any key.

To prove the point; if the instrument is tuned in fifths as described without diminishing each interval the final check interval (E to A) will be so diminished as to sound appalling!

For greater precision, if the instrument can be made to sustain, the beats can be counted over 10 seconds, as per the table given in the interesting book by Richard H. Dorf, *Electronic Musical Instruments*. Kenneth Palmer, Kensal Rise, London.

Electronic piano design

I would like to reassure actual or potential constructors who may have been disturbed by Mr Mitchell's letter in the August issue.

The reliability and objectivity of Mr Mitchell's remarks leave something to be desired. He refers, without being specific, to "considerable circuit duplication". Now it should be clearly understood that while the piano *does* contain many duplicated circuits, *none* of these is redundant. Electronic pianos and organs can be designed along very much the same lines; the main differences being in the key circuits. Now in a polyphonic instrument (and any worth-while instrument must be polyphonic) each key must have an entirely separate piece of circuitry associated with it. In an organ these circuits are quite simple, but in a piano they are not, neither do they lend themselves to total integration.

On the subject of cost, it should be pointed out that the electronics represent

only half of the total cost of the project. It does not seem to be possible to significantly cut the cost of the electronics even by a major redesign; they are already very simple and use cheap components.

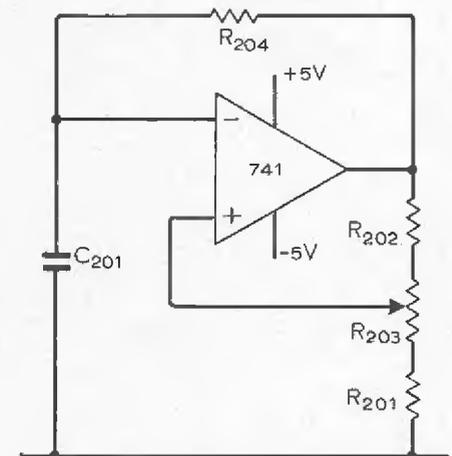
There are only about three possible realizations of the oscillator section that are at all likely to be satisfactory in terms of frequency stability; these are LC oscillators, RC oscillators using high-gain op-amps, and full-octave synthesizers driven by a single oscillator. See the May 1974 *Wireless World* pp. 143-5 for details of the latter. Special i.c.s of the "555" type probably are not stable enough. The most costly solution, the full-octave synthesizer i.c., is probably the best. The necessary buffers cost little.

I hope that those readers who ordered demonstration cassettes found them helpful; they were of course intended to demonstrate the characteristic "electronic-piano" timbre which differs somewhat from acoustic piano sound. My apologies are extended to anyone who was expecting anything musical; nothing of the sort was promised!

Geoff Cowie,
London, N10.

Electronic piano design

May I suggest an improvement to G. Cowie's electronic piano (March-May issues)? My experience of his RC relaxation oscillators has been that they are potentially extremely stable, but suffer from variations in contact resistance of the pre-set.



The problem can be minimized by a simple rearrangement of components as shown in the diagram. Any change in contact resistance is to be compared with the input resistance of the 741, not R_{204} . Since the input resistance of a 741 is typically $1M\Omega$ an increase in stability greater than an order of magnitude can be expected.

The formula for frequency of oscillation should be

$$f \approx \frac{R_1 + R_2}{4 C_1 R_1 R_2} \text{ and not as published.}$$

M. Walne,
Brighouse, Yorks.

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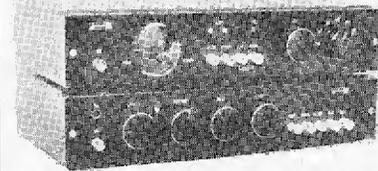
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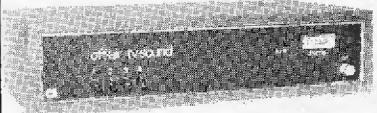
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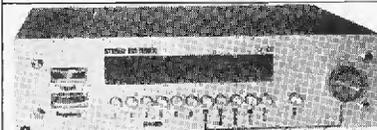
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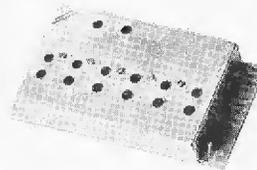
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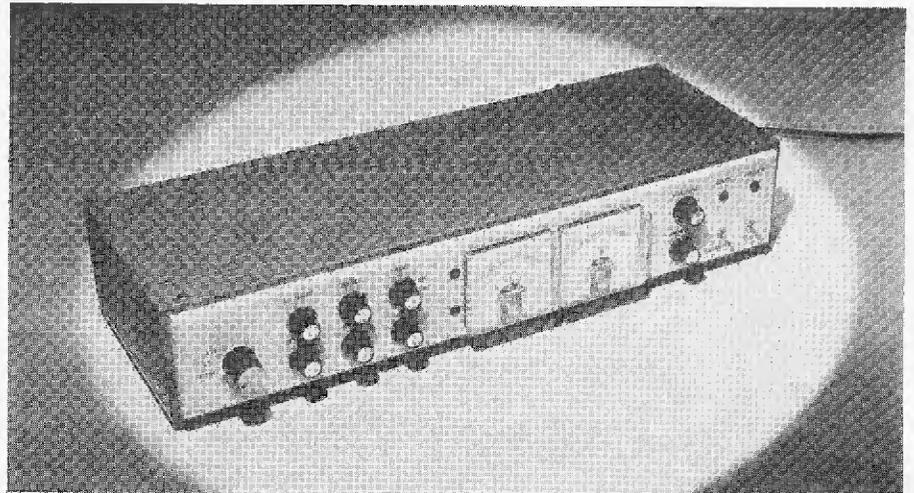
by D. Self, B.A., *Electrosonic Ltd*

This preamplifier design offers a distortion figure of below 0.002%, an overload margin of around 47dB, and a signal-to-noise ratio of about 71dB for the disc amplifier. A novel noise gate mutes the output when no signal is presented to the disc input and conversely, by using the subsonic information present on record pressings, eliminates the problem of muting low level signals.

This article describes a stereo pre-amplifier that equals or exceeds the performance of many of those available. The circuit incorporates a novel method of muting the signal path, when the disc input is quiescent, by using a noise gate that never mutes a wanted low-level signal.

Many of the important performance factors, such as signal-to-noise ratio, overload margin, and accuracy of the RIAA equalization, are essentially defined by the design of the disc input circuitry. This therefore merits close attention. The best attainable s/n ratio for a magnetic cartridge feeding a bipolar transistor stage with series feedback is about 71dB with respect to a 2mV r.m.s. input at 1kHz, after RIAA equalization. This has been clearly demonstrated by Walker¹. The equivalent amplifier stage with shunt feedback gives an inferior noise performance over most of the audio band due to the rise in cartridge source impedance with frequency. This limits the maximum s/n ratio after equalization to about 58dB. These facts represent a limit to what the most advanced disc input stage can achieve.

Overload margin appears to be receiving little attention. The maximum velocities recorded on disc seem to be steadily increasing and this, coupled with improved cartridges, means that very high peak voltages are reaching disc inputs. Several writers have shown that short-term voltages of around 60 to 80mV r.m.s. are possible from modern discs and cartridges, and higher values are to be expected. This implies that to cater for signal maxima, a minimum overload margin of 32dB with respect to



2mV r.m.s. at 1kHz is essential. Obviously a safety factor on top of this is desirable. However, most pre-amplifiers at the top end of the market provide around 35-40dB only. There are certain honourable exceptions such as the Technics SU9600 control amplifier which achieves an overload margin of 54dB, mainly by the use of a staggeringly high supply of 136V in the disc input amplifier. The Cambridge P50/110 series offers a margin in excess of 60dB by the artifice of providing unity-gain buffering, for correct cartridge loading, but no amplification before the main gain control. This allows the use of an 18V supply rail, but does limit the maximum s/n ratio.

The overload margin of a pre-amplifier is determined by the supply voltage which sets the maximum voltage swing available, and by the amount of amplification that can be backed-off to prevent overload of subsequent stages. Most pre-amplifiers use a relatively high-gain disc input amplifier that raises the signal from cartridge level to the nominal operating level in one jump. Low supply voltages are normally used which reduce static dissipation and allow the use of inexpensive semiconductors. The gain control is usually placed late in the signal path to ensure low-noise output at low volume settings. Given these constraints, the overload performance is bound to be

mediocre, and in medium-priced equipment the margin rarely exceeds 30dB. If these constraints are rejected, the overload margin of the system can be improved.

Two separate gain controls remove the most difficult compromise, which is the placement of the volume control. This approach is exemplified in the Radford ZD22 and the Cambridge P60 circuitry. One gain control is placed early in the signal path, preceded by a modest amount of gain. Cartridges of high output can be accommodated by the use of this first control. The second is placed late in the pre-amplifier and is used as a conventional volume control, see Fig. 1.

The other performance criterion which is largely defined by the disc input circuitry is frequency response, as defined by the accuracy of the RIAA equalization. Assuming that the relevant amplifying stage has sufficient open-loop gain to cope with the bass boost required, the accuracy of the equalization depends entirely on the time constants within the feedback loop. Careful design, and the use of close-tolerance components can assure an accurate response to within ± 0.2 dB from 30Hz to 20kHz.

Pre-amplifier distortion seems to have received little attention compared with that generated by power amplifiers, perhaps because the former has

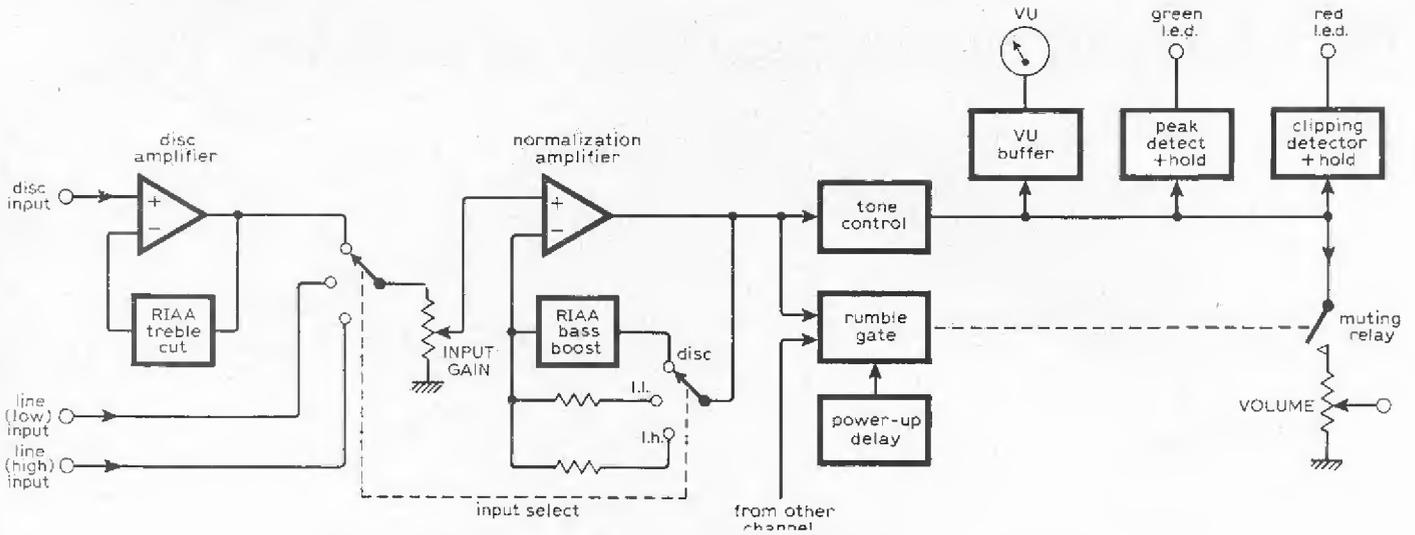


Fig. 1. Block diagram of the complete circuit. Two gain controls are used in the signal path to allow a substantial increase in overload margin.

traditionally been much lower. However, power amplifiers, with such low t.h.d. that the residual harmonics can no longer be extracted from the noise at normal listening levels, are now commonplace, particularly with the advent of techniques such as current dumping. This desirable state of affairs unfortunately does not extend to pre-amps, which in general produce detectable distortion at nominal operating levels, usually between 0.02% and 0.2%. In this design the t.h.d. at 1kHz is less than 0.002% even at 25dB above the nominal operating level of 0dBm. A Sound Technology 1700A distortion measurement system was used during development.

At this point it is convenient to consider the noise gate principle. When the pre-amplifier is being used for disc reproduction the output from each channel is continuously sampled to determine if a signal is present; if nothing is detected within a specified time interval, dependent on the previous signal levels received, the pre-amplifier is muted by the opening of a reed relay in series with the output signal path. This allows only power amplifier noise to reach the loudspeakers and considerably reduces the perceived noise generated by a quiescent sound system. Noise in the quiescent state is particularly noticeable when headphones are in use. The reed relay is also used to prevent switch-on transients from reaching an external power amplifier. So far this circuit appears to be a fairly conventional noise gate. The crucial difference is that signals from disc that have not been subjected to rumble filtering are always accompanied by very low frequency signals generated by record ripples and small-scale warps. Even disc pressings of the highest quality produce this subsonic information, at a surprisingly high level, partly due to the RIAA bass boosting. The l.f. component is often less than 20dB below the total programme level but this is quite sufficient to keep the pre-amplifier unmuted for the duration of a l.p. side.

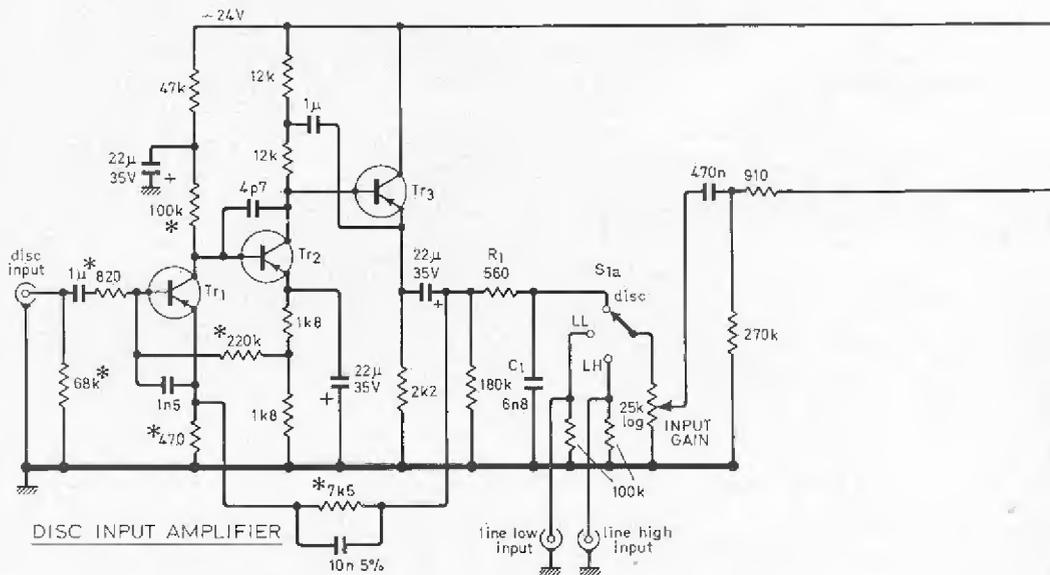
The pre-amplifier is unmuted as soon as the stylus touches the disc, and muted about a second after it has been raised from the run-out groove. This delay can be made short because the relative quiet at the start of the run-out groove is sensed and stored. The rumble performance of the record deck is largely irrelevant because virtually all of the subsonic information is generated by disc irregularities.

Audio circuitry

A detailed block diagram of the pre-amplifier is shown in Fig. 1, and Fig. 2 shows the main signal path. The disc input amplifier uses a configuration made popular by Walker, but the

Fig. 2. Circuit diagram of the signal path. Constant-current sources are biased from a l.e.d./resistor chain for improved thermal stability.

collector load of the second transistor is bootstrapped. This increases the open-loop gain and hence improves the closed-loop distortion performance by a factor of about three to produce less than 0.002% at an output of 6.5V r.m.s. (1kHz). This stage gives a s/n ratio (ref 2mV) of about 70dB and a gain of 15 at 1kHz. This is sufficient to ensure that the noise performance is not degraded by subsequent stages of amplification. The maximum output of this stage before clipping is about 6.5V r.m.s. and the nominal output is 30mV r.m.s. Because this is the only stage before the input gain control, these two figures set the overload margin at 47dB. To ensure that this overload margin is maintained at high frequencies, the treble-cut RIAA time-constant is incorporated in the feedback loop. This leads to slightly insufficient cut at frequencies above 10kHz because the gain of the stage cannot fall below unity, and hence fails to maintain the required 6dB/octave fall at the top of the audio spectrum. This is exactly compensated for outside the feedback loop by the low-pass filter R₁ C₁, which also helps to reject high frequencies above the audio band.



For convenience I have referred to the next stage of the circuit as the normalization amplifier because signals leaving this should be at the nominal operating level of 0dBm by manipulation of the input gain controls. Separate controls are provided for each channel to allow stereo balance. A later ganged control is used for volume setting and causes no operational inconvenience. In the disc replay mode, the normalization amplifier provides the RIAA bass boost, by the feedback components $R_{2,3}$ and C_2 . Two line inputs are also provided; line low requiring 30mV and line high 100mV to give 0dBm from the normalization stage with the input gain control fully advanced. When these inputs are selected, the feedback networks are altered to adjust the gain and give a flat frequency response. Ultrasonic filters are incorporated to ensure stability and aid r.f. rejection. Capacitor C_3 in the feedback arm reduces the gain to unity at d.c. for good d.c. stability. If a fault causes the amplifier output to saturate positively the capacitor is protected by a diode which has no effect on the distortion performance.

The circuitry of the normalization amplifier is complicated because its performance is required to be extremely high. The harmonic distortion is far below 0.002% at the maximum output of

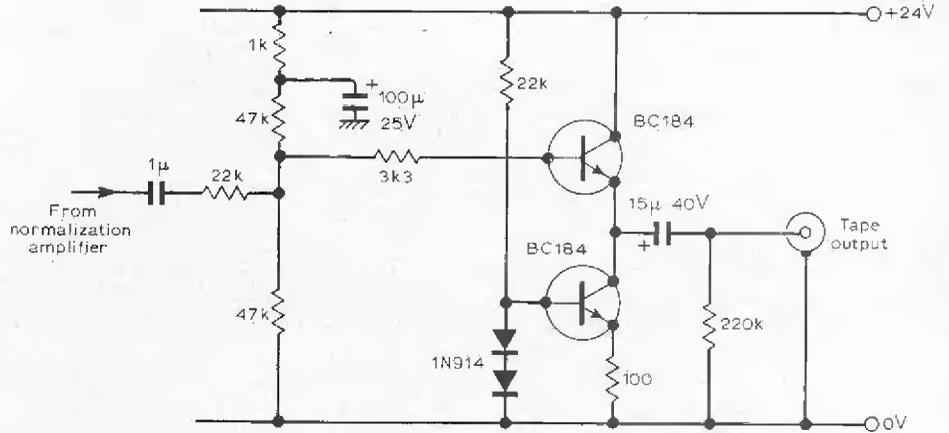


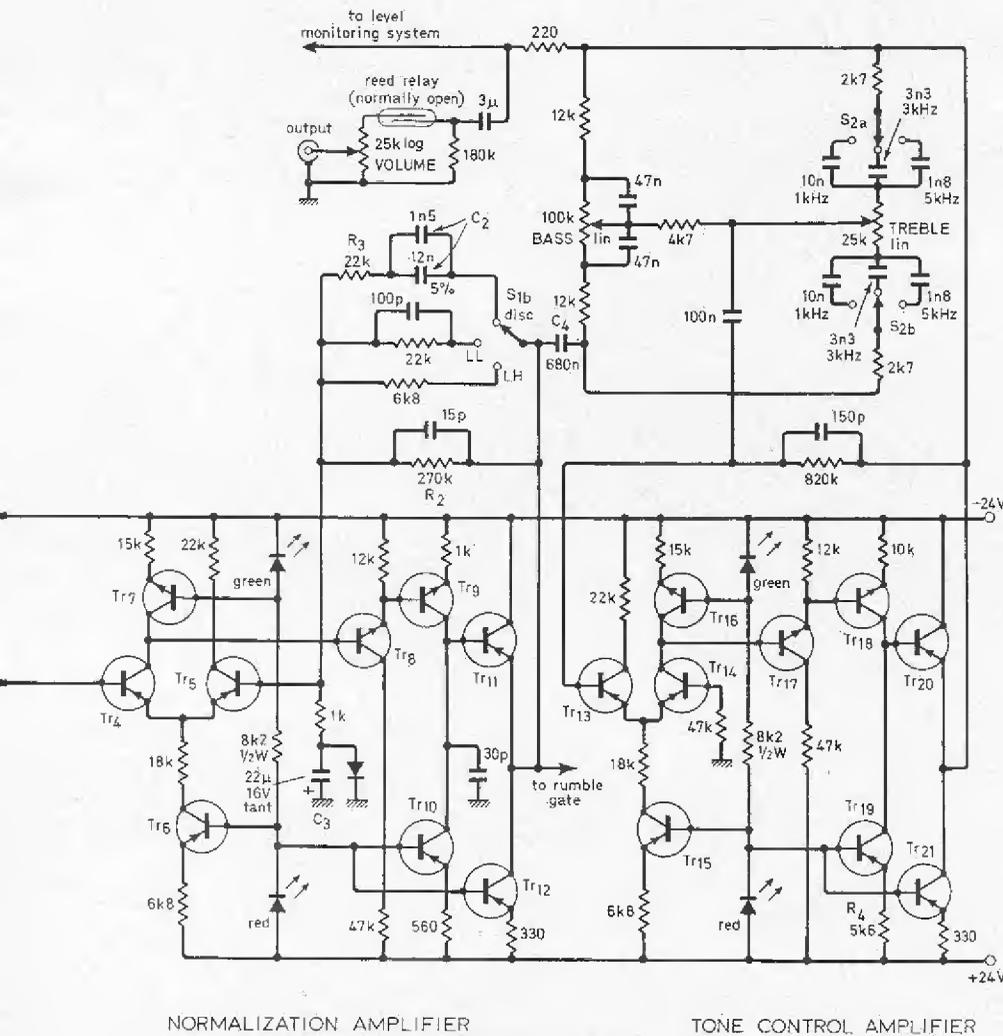
Fig. 3. Tape output circuit. The smallest allowable load impedance for an undistorted output is about 2.2kΩ. Line inputs of the pre amplifier are suitable for playback purposes.

14.5V r.m.s. which is 25dB above nominal operating level. This large amount of preamplifier headroom allows gross preamplifier overload before clipping. The input stage of the amplifier is a differential pair with a constant-current source for good common-mode rejection. The operating currents are optimized for good noise

performance, and the output is buffered by an emitter-follower. The main voltage amplifier, Tr_9 has a constant-current collector load so that high voltage gain at low distortion can be obtained. This performance is only possible if the stage has very little loading so it is buffered by the active-load emitter-follower. The various current sources are biased by a l.e.d.-resistor chain because the forward voltage drop of an l.e.d. has a negative temperature coefficient that approximates closely to that of a silicon transistor V_{be} drop. Hence, this method provides exceptionally stable d.c. conditions over a very wide temperature range.

After the normalization stage the signal is applied to a tone-control circuit based on the Baxandall network. The main limitation of the Baxandall system is that the turnover frequency of the treble control is fixed. In contrast, the bass control has a turnover frequency that decreases as the control nears the flat position. This allows a small amount of boost at the low end of the audio spectrum to correct for transducer shortcomings. The equivalent adjustment at the high end of the treble spectrum is not possible because boost occurs fairly uniformly above the turnover frequency for treble control settings close to flat. In this circuit the treble turnover frequency has been given three switched values which have proved useful in practice. Switch 2 selects the capacitors that determine the turnover point. The maximum boost/cut curves are arranged to shelve gently, in line with current commercial practice, rather than to continue rising or falling outside the audio range. In addition, the coupling capacitor C_4 has a significant impedance at 10Hz so that the maximum bass boost curve not only shelves but begins to fall. Full boost gives +15dB at 30Hz but only +8dB at 10Hz. The tone control system has a maximum effect of ±14dB at 50Hz and ±12.5dB at 10kHz.

The tone-control amplifier uses the same low distortion configuration as the normalization stage, but it is used in a virtual-earth mode. The main difference is that the open-loop gain has been traded for open-loop linearity by increasing the emitter resistor of the



NORMALIZATION AMPLIFIER

TONE CONTROL AMPLIFIER

main voltage amplifier from 1kΩ to 10kΩ thus increasing local feedback. Resistor R₄ has been increased to 5.6k to maintain appropriate d.c. conditions. This modification makes it much easier to compensate for stability in the unity-gain condition that occurs when treble-cut is applied.

Level detection circuitry

From the tone-control section the signal is fed to the final volume control via the muting reed-relay. Note that this arrangement allows the volume control to load the input of the external power amplifier even when the relay contacts are open, thus minimising noise. The signal level leaving the tone-control stage is comprehensively monitored by the circuitry shown in Fig.4. Each channel is provided with two peak-detection systems, one lights a green l.e.d. for a pre-determined period if the signal level exceeds 1V peak, and the other lights a red l.e.d. if the tone-control stage is on the verge of clipping. Each channel is also provided with a VU meter driver circuit. Transistor Tr₂₂ forms a simple amplifying stage which also acts as a buffer. Voltage feedback is used to ensure a low-impedance drive for the meter circuitry. The first peak detector is formed by IC₁ and its associated components. When the voltage at pin 2 goes negative of its quiescent level by one volt, the timer is triggered and the l.e.d. turns on for a defined time. The relatively heavy l.e.d. current is drawn from an unstabilized supply to avoid inducing transients into any of the stabilized supplies.

The clipping detector continuously monitors the difference in voltage between the tone-control amplifier

output and both supply rails. If the instantaneous voltage approaches either rail, this information is held in a peak-storage system. Normally Tr₂₄ and Tr₂₅ conduct continuously but if the junction of D₁ and R₅ approaches the +24V rail then Tr₂₄ and hence Tr₂₅ turn off. This allows C₅ to charge and turn on Tr₂₆, and Tr₂₇ and hence the l.e.d. until the charge on C₅ has been drained off through emitter-follower Tr₂₆. If the measured voltage nears the -24V rail, then D₁ conducts to pull up the junction of R₆ and R₇, which once again turns off Tr₂₅. In this way both positive and negative approaches to clipping are indicated. This comprehensive level indication does of course add significantly to the task of building and testing the preamplifier. If desired, any or all of the three sections may be omitted.

Noise gate

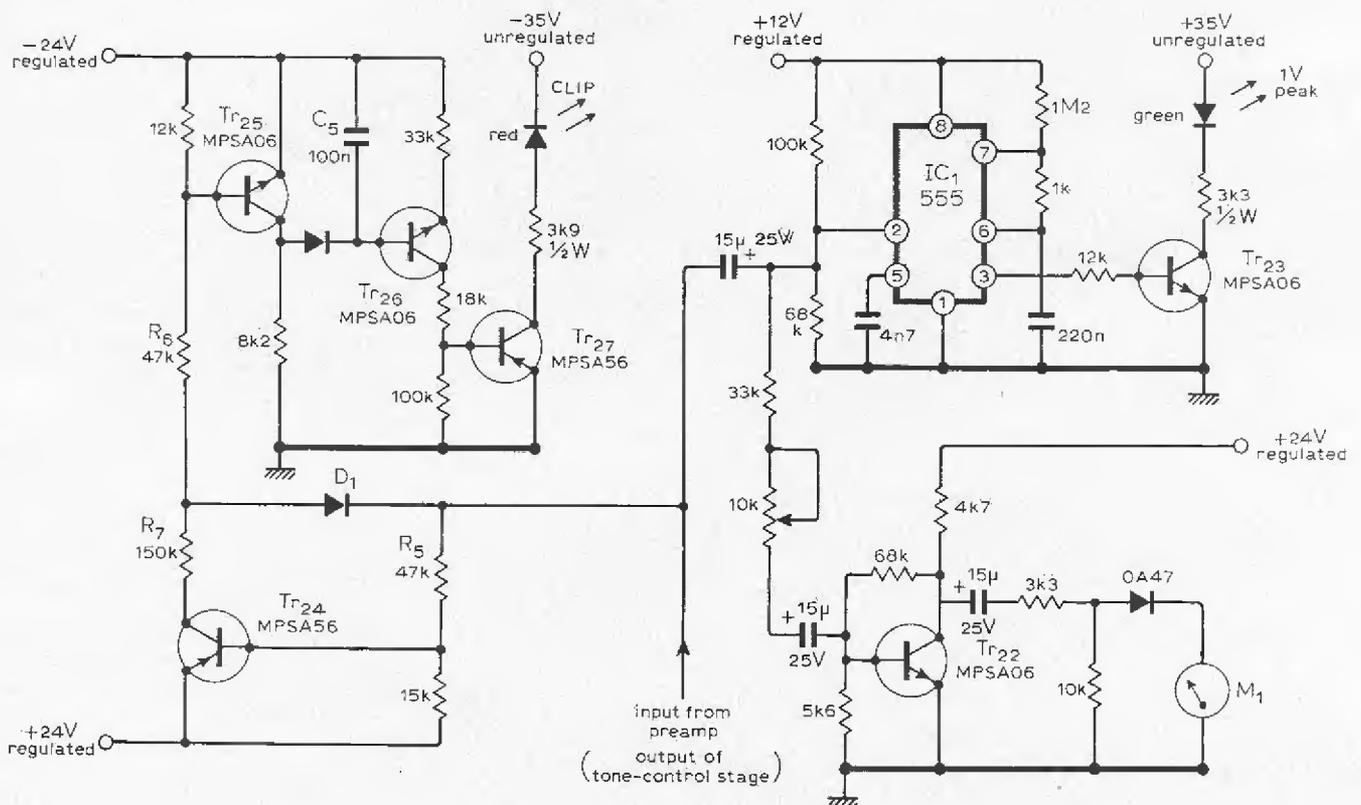
The final section controls the muting reed-relay. At switch-on, the +12V rail rises rapidly until stabilized by the zener diode. Pin 2 on IC₄ is, however, briefly held low by C₆, and the 555 is therefore immediately triggered to send pin 3 high. This saturates Tr₂₈ which prevents Tr₂₉ from turning on. At the end of the time delay, pin 3 goes low and relay driver Tr₂₉ is no longer disabled.

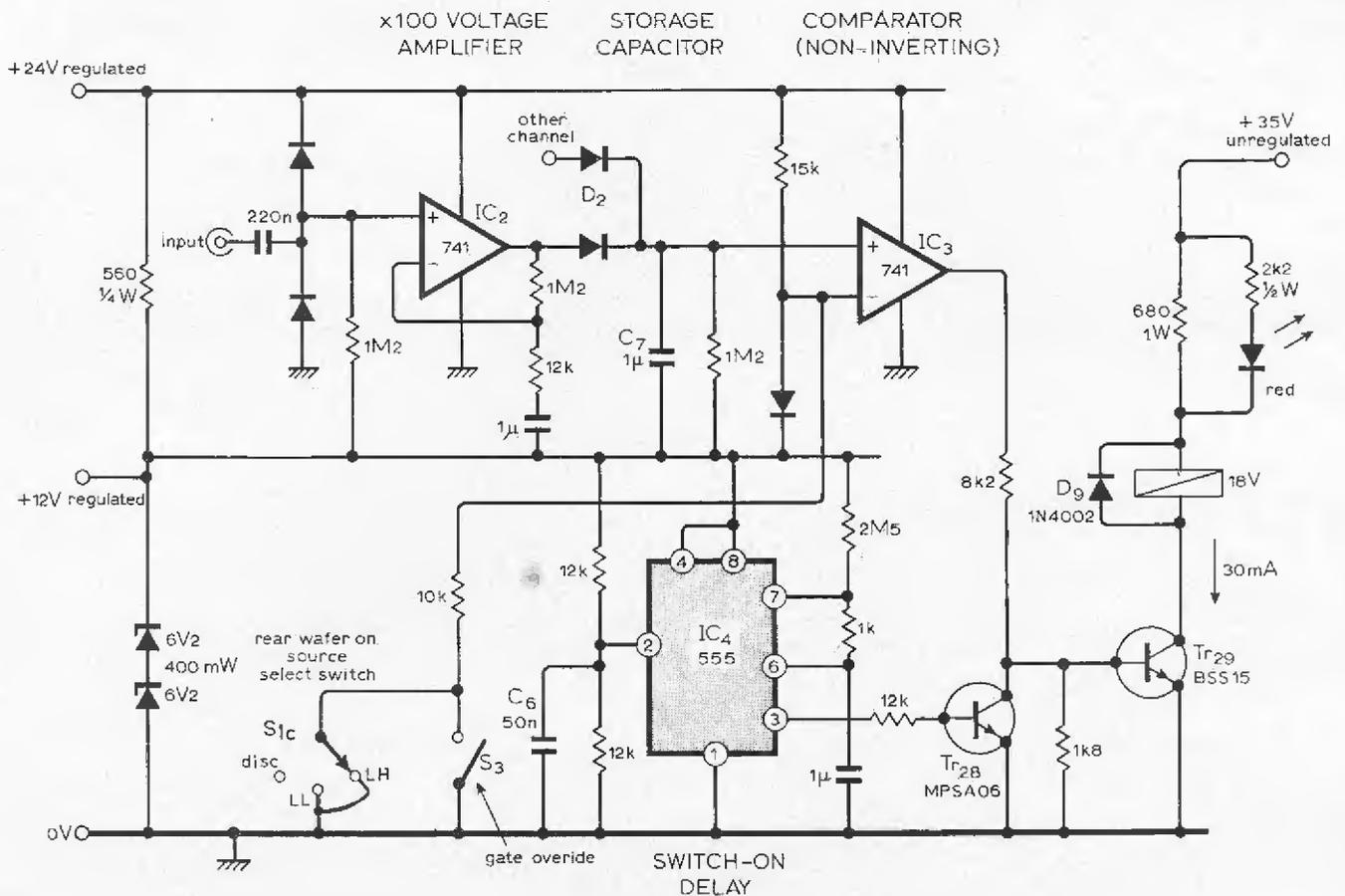
The noise gate uses two amplifiers with gains of about 100. These sample both channels at the output of the normalization stage and the inputs are clamped with diodes so that the norma-

lization amplifiers may use their full voltage swing capability without damaging the 741s. Due to their high gain, under normal signal conditions the op-amp outputs move continuously between positive and negative saturation which keeps the storage capacitor C₇ fully charged. In the silent passages between l.p. tracks the l.f. signal is not normally of sufficient amplitude to cause saturation but will usually produce at least +3 to +4 volts across C₇ which gives a large margin of safety against unwanted muting. To facilitate this the response of the amplifiers is deliberately extended below the audio band. When the stylus leaves the record surface and the l.f. signals cease, C₇ slowly discharges until the non-inverting input of comparator IC₃ falls below the voltage set on the inverting input. At this point the 741 switches and its output goes low to cut off the base drive to Tr₂₉, and switch off the relay. When the stylus is replaced on a record, the process takes place in reverse, the main difference being that C₇ charges at once due to the low forward impedance of D₂. To prevent the relay sporadically operating when the preamplifier is handling signals presented through the line inputs, an extra wafer on the source-select switch is arranged to override the rumble-sensing circuit, and provide permanent unmute. This is achieved by pulling the inverting input of comparator IC₃ negative of the +15V rail by the 10kΩ resistor so that even when C₇ is fully discharged, IC₃ will not switch. In addition, S₃ provides a manual override for testing and comparison purposes.

The power supply is shown in Fig.6. Regulators are used to provide stabi-

Fig. 4. Level monitoring circuitry. Although three separate circuits are shown, these may be omitted as required.





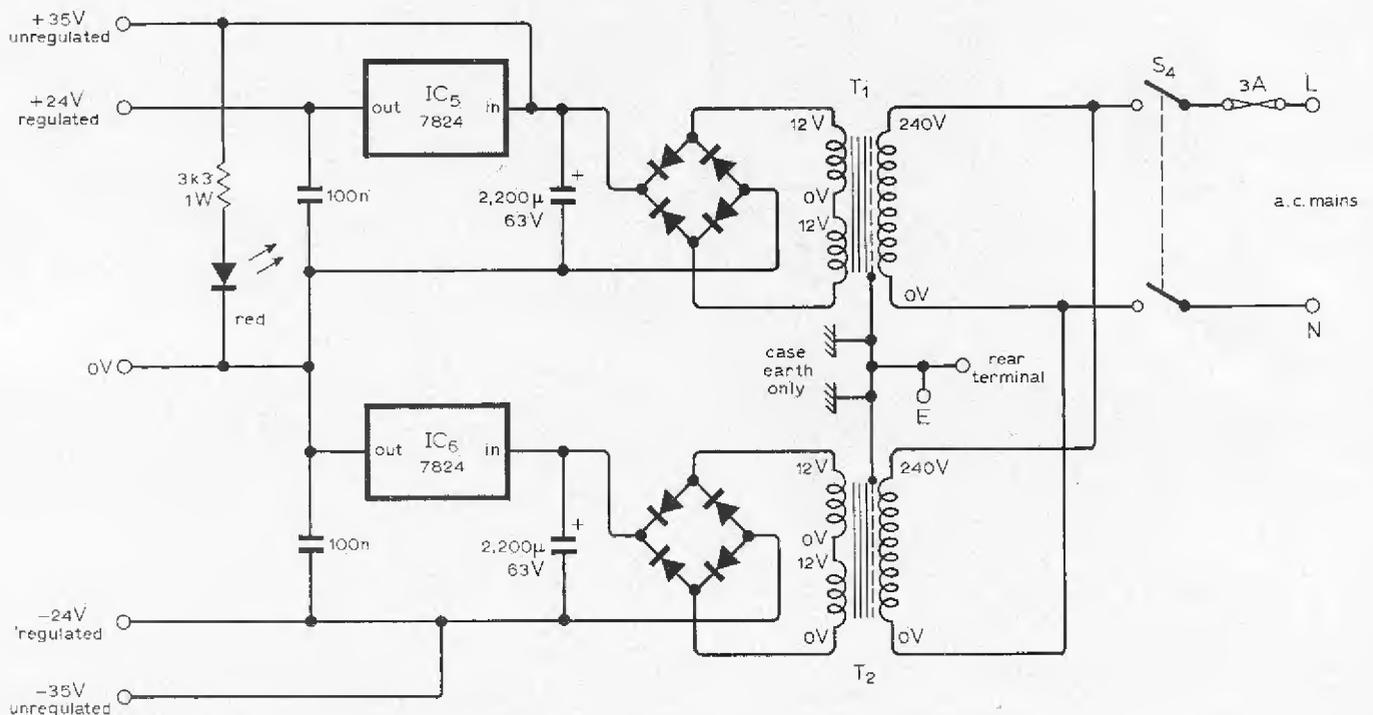
lized $\pm 24V$ rails. The unregulated supply rests at about $\pm 35V$. The signal circuitry has been designed to withstand $\pm 35V$ appearing on the supply rails, so that even in the unlikely event of both regulators failing, no further destruction will arise. Each regulator i.c. requires about $7cm^2$ of heat sink area.

Physical layout of the preamplifier is no more critical than that of any other piece of audio equipment. In general it is wise to use a layout that places the disc

input amplifier as close as possible to its input socket, and as far as possible from the mains transformer. Screened cable should be used between the disc input stage and its input socket, and between the final volume control and the output socket. The earthing requirements are straightforward and the circuit common 0V rail is led from the input sockets through the signal path to the output volume control, and finally to the 0V terminal of the power supply. This arrangement minimises the possi-

Fig. 5. Noise gate and delay switch on circuitry. The noise gate is provided with an override switch for use with line input signals. The delay switch-on overrides all of the circuitry. Amplifier IC₂ is repeated for a stereo system.

Fig. 6. Power supply. Two regulator i.c.s are used which should be mounted on heat sinks.



Component notes

All unmarked diodes are 1N914 or equivalent.

Red bias l.e.ds are TIL209 or equivalent.

Green bias l.e.ds are TIL211 or equivalent.

Resistors marked with an asterisk should be metal oxide types

Tr₁ to Tr₆ and Tr₁₃ to Tr₁₅ are BCY71
Tr₇, 8, 9, 16, 17, 18, 22, 23, 25, 26, 28 are
MPS A06.

Tr₁₀, 11, 12, 19, 20, 21, 24, 27 are MPS
A56

Tr₆ is BFX85 or equivalent.

The muting reed relay should be a 2 pole make type with an 18V coil. If a different coil voltage is used, the value of the dropper resistor should be adjusted.

The VU meter should have a 1mA movement.

If an internal diode and series resistor are fitted, the external components should be omitted.

Switch 1 (source select) is a 5 pole 3 way

Switch 2 (treble frequency) is a 4 pole 3 way

bility of spurious e.m.f.s arising between stages. The only problem likely to be encountered is the formation of an earth loop when the preamplifier is connected to a power amplifier. Therefore, it may be satisfactory in a permanent installation to have the preamplifier circuitry connected to mains earth only through the signal lead to the power amplifier. The preamplifier case must of course be connected to the mains earth for safety reasons. It is preferable to define the potential of the preamplifier even if the power amplifier is disconnected. In the prototype the 0V rail was connected to the mains earth via a 22Ω resistor which stops the formation of an earth loop and prevents the signal circuitry from taking up a potential above earth due to leakage currents etc.

Testing is relatively straightforward, providing the preamplifier is constructed and checked stage by stage. Dynamic parameters such as t.h.d. are not accurately measurable without expensive test gear, but it has been found in the course of experimentation that if the d.c. conditions are correct then the various signal stages almost always show the desired a.c. performance. The non-signal circuitry should be relatively

simple to fault-find. No problems should be encountered with the noise gate section which has proved to be very reliable throughout a protracted period of testing. The only preamplifier adjustment is for the VU meter calibration. This should be set to 1V r.m.s. = 0VU, which is completely non-standard but very useful in terms of the dynamic range of the signal path. For normal operation the input gain controls should be set so that the meter indications do not exceed 0VU, to preserve a safety margin in the later stages. This completes the preamplifier design.

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2. King, Gordon J. "The Audio Handbook", Newnes-Butterworth, 1975.
3. Heidenstrom, P.N. "Amplifier Overload", *Hi-Fi News*, December 1974.

It's the cat's whiskers!



A fascinating excursion into the past. The author has unearthed some 400 trade names from the crystal set days, along with nearly 200 manufacturers—giving the name of the set, technical description and original price. He also reviews the first days of broadcasting and looks at the difficulties experienced by crystal set users. Concise information and over 40 illustrations make this book a valuable work of reference as well as a rare piece of nostalgia for collectors.

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High quality tone control

A low distortion design

by J. N. Ellis

It is recognized^{1,3} that to obtain low noise the usual one-transistor configuration² gives generally poor results, and has a distortion level approaching 1% at about 1V r.m.s. output. The signal-to-noise ratio can be greatly improved by using two transistors directly coupled, with the first device operating in common emitter and the second in common collector mode. The first stage current can now be $100\mu\text{A}$, giving us a much better signal-to-noise ratio. This two-transistor design is often used, but suffers from latch up on overdrive.

The author's design raises the signal level from 100mV to 1V r.m.s. to drive a power amplifier and uses a cascode circuit to provide a more stable operating point and lower distortion. This is because the instantaneous collector voltage of the common-base transistor does not appreciably affect the current flowing in it. With a similar transistor cascode pair, the bias resistors may be low enough to inject a noise current into the lower device. Use of a complementary cascode configuration allows the selection of reasonable values of bias resistance.

To make full use of the advantages of the design (Fig.1) the tone network is fed from an impedance equal to that presented at the output, essentially R_{16} and R_{17} in

parallel. This allows a flat response when the potentiometers R_{19} and R_{20} are mechanically central⁴. The buffer stage (Tr_1) allows the impedance to remain constant, independent of the volume control setting.

Component values of the tone network have been selected so that maximum bass boost or cut occurs at 50Hz, and the treble boost or cut maximum at 10kHz, as it has been found that the full 20dB (theoretical) at 20kHz is unnecessary, as the sensitivity of the ear is reducing rapidly at that point⁴. Making R_7 and R_8 equal to $1k\Omega$ allows the greater range to be obtained for the impressionist. The frequency response is shown in Fig.2.

Without C_9 the square-wave response showed slight ringing, eliminated by making $C_9 = 4.7\text{pF}$. By increasing C_9 to 10pF the response is made 3dB down at 175kHz and the low frequency 3dB point is 5Hz.

The design has an overall gain of 10 (20dB), and for 1 volt output with $R_L = 10k\Omega$ and $R_S = 100\Omega$, the total harmonic distortion (measured) was less than 0.1% at 1kHz. The signal to noise ratio could not be accurately measured on the equipment available at the time, but is

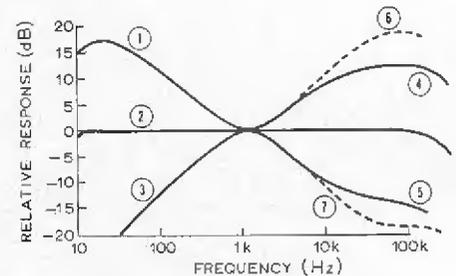


Fig.2. Amplitude/frequency response curves of tone control circuit. 1. Bass boost max. 2. Flat response 3. Bass cut max. 4. Treble boost max. 5. Treble cut max. 6, 7. Treble boost and cut with $R_7, R_8 = 1k\Omega$.

estimated to be -110dB and certainly greater than -100dB using low noise transistors — an improvement of 10 to 20dB over other designs.

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1. "Low Distortion Tone Controls", *Wireless World*, April 1971.
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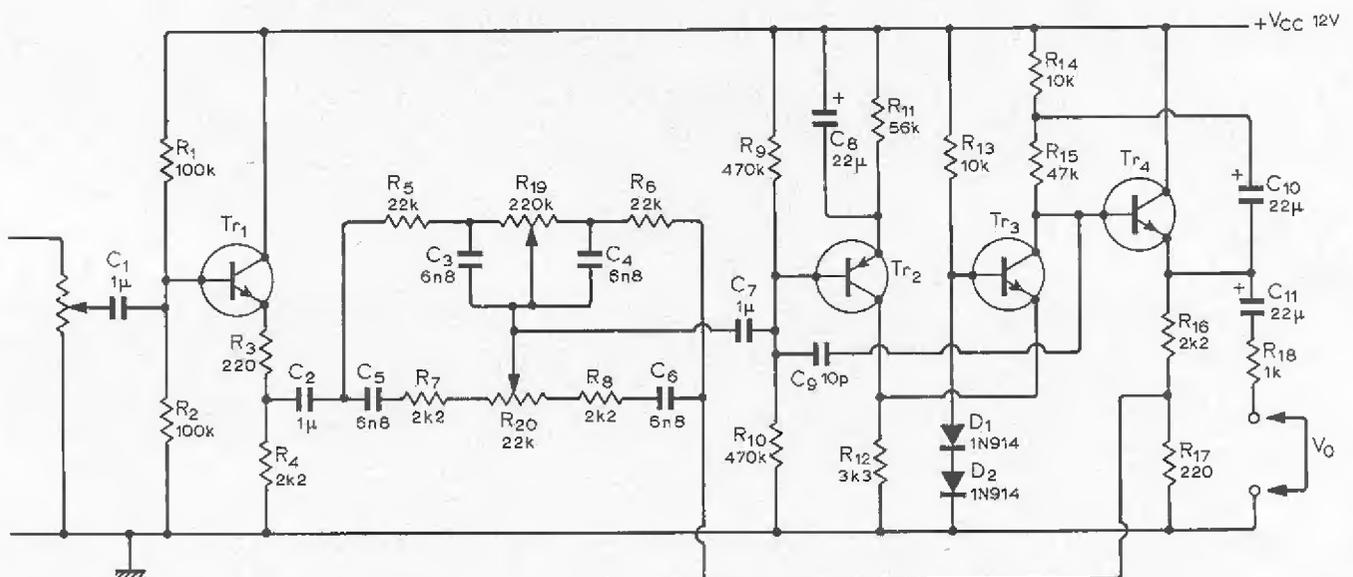


Fig. 1. Circuit diagram of tone control. Transistors $Tr_1, Tr_3, Tr_4 = \text{BC109, BC114, BC184}$. $Tr_2 = \text{BC15, BC214, BC309}$ etc.

Multi-channel tone control

A five-channel unit using active band-pass filters

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

New ideas in tone controls have appeared in the last few years^{2,3} which improve or supplement bass and treble controls, which are usually of the Baxandall type.¹ If the function of these controls is clearly understood, a fairly flexible amplitude frequency response curve can be built up. However, an easier and more versatile method of building a desired response is to use a number of overlapping, variable gain, frequency channels, spaced evenly throughout the audio spectrum. By using linear slider potentiometers, the control panel presentation can be made to resemble a response against frequency graph. In this form the system is sometimes called a "graphic equalizer". The unit which will be described for construction has five channels with centre frequencies of 50Hz, 200Hz, 800Hz, 3.2kHz and 12.8kHz. The linear slider potentiometers used give a control range of ± 12 dB. The designed signal level for normal use is around 0dBm (approximately 800mV), a level matching most modern amplifier and recording equipment inputs.

Negative feedback system

The usual method of obtaining a multi-channel response is by using an LC band-pass filter and gain control for each channel. The range of inductance needed to cover the ten octaves or so of the audio band makes the filters expensive and unattractive. In addition, much trimming is normally needed to obtain a basically flat response, due to crossover interactions between channels.

Negative feedback methods offer many advantages, such as reduced noise and distortion and accurate gain setting using potentiometers of linear law. In the case of the multichannel system, the filter specification may be greatly relaxed, and trimming can normally be eliminated.

A block diagram of this type of system is given in Fig. 1. At the centre frequency of a channel, and assuming no interaction between channels

$$|V_{out}|/|V_{in}| = R_a/R_b$$

If a simple tuned circuit bandpass filter is used, in order to cross over evenly between channels the Q value should be $\sqrt{2^x/(2^x-1)}$, where x is the channel spacing in octaves. In a negative feedback system this is not

critical and the value of Q can be raised to reduce the interaction, the penalty being a rise in noise at the crossover points. Error in the centre frequency of a filter has a similar effect; as a guide, approximately 1.5dB increase in noise is produced by 50% Q error, or 15% error in centre frequency.

At a first glance it would seem desirable to make the first and last filters low and high pass respectively, but in actual fact this response is produced in a negative feedback loop using only bandpass filters. The penalty is increased noise again, but this time well outside the audio range, assuming a reasonable choice of channel frequencies has been made. Using only bandpass filters simplifies the system since only one basic filter type is required. The design of this network will now be considered.

Active bandpass filter

In some systems there could be thirty or more channels, and since there is one filter circuit per channel, this must be designed as

Performance of the five-channel tone control

-3dB bandwidth at 1V r.m.s. signal level, controls flat: 8Hz-30kHz
distortion at 1V r.m.s. signal level, controls flat: 100Hz-0.05%
1kHz-0.05%
10kHz-0.05%
clipping level, controls flat: 2.9V r.m.s.

economically as possible within the limits of the specification. Assuming no inductors are to be used for the reasons mentioned before, the number of active devices should be minimized, unless component tolerances can be relaxed by using more. Close tolerance capacitors can prove especially expensive.

A simple circuit that meets these requirements is the Wien bridge derivative⁴ shown

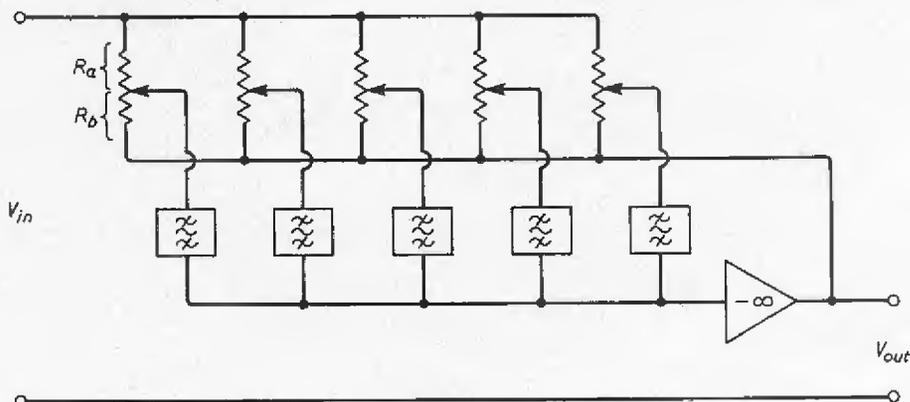


Fig. 1. Simple block diagram of the multi-channel negative feedback system.

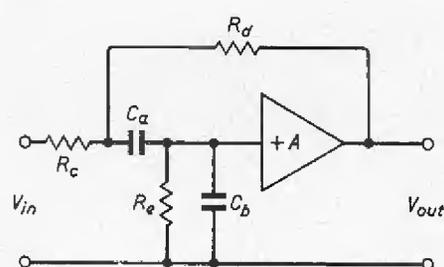


Fig. 2. Example of a band-pass active filter.

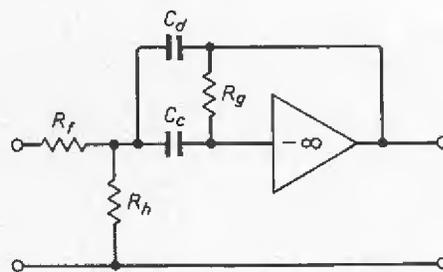


Fig. 3. Band-pass filter using multiple feedback.

*University of Durham

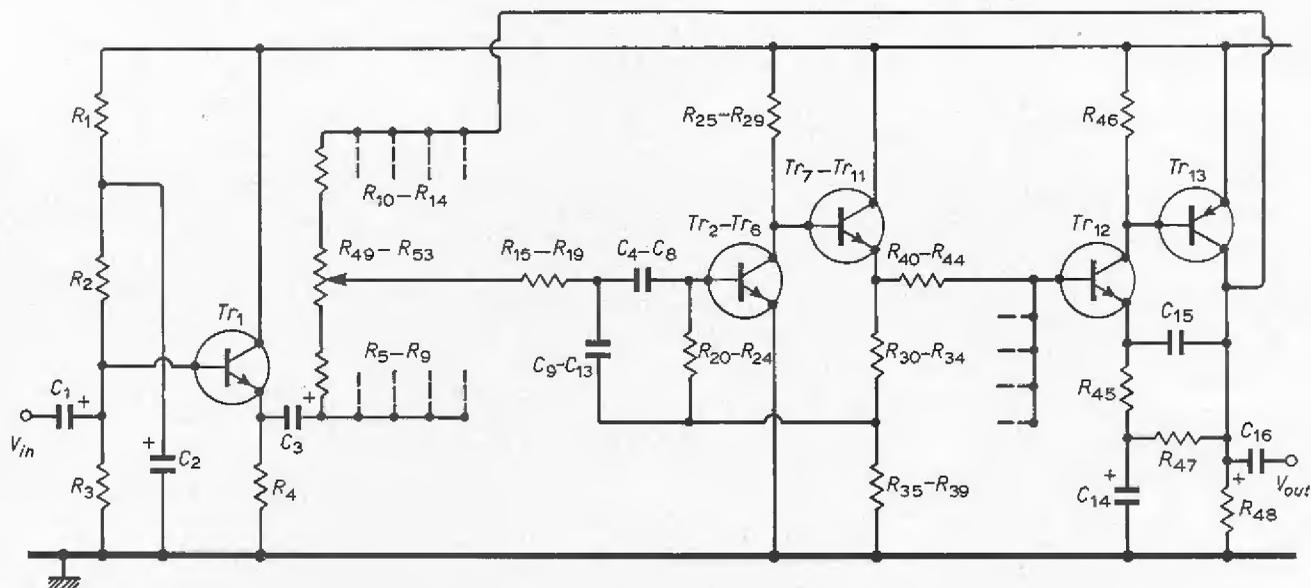


Fig. 4. Circuit of the five-channel tone control unit. Only one of the filters and linear potentiometer is shown.

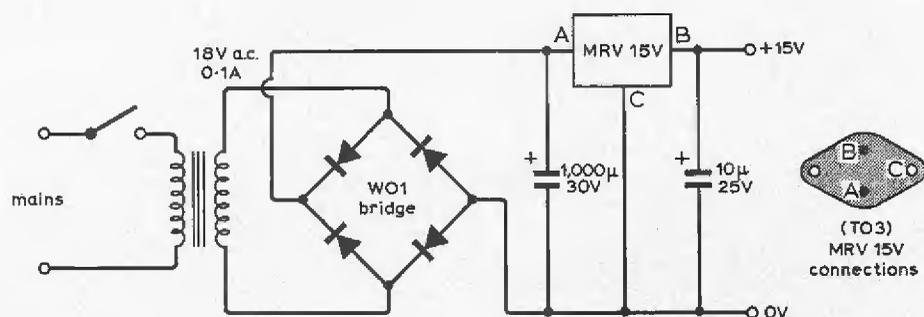


Fig. 5. Suggested power supply circuit using an integrated regulator.

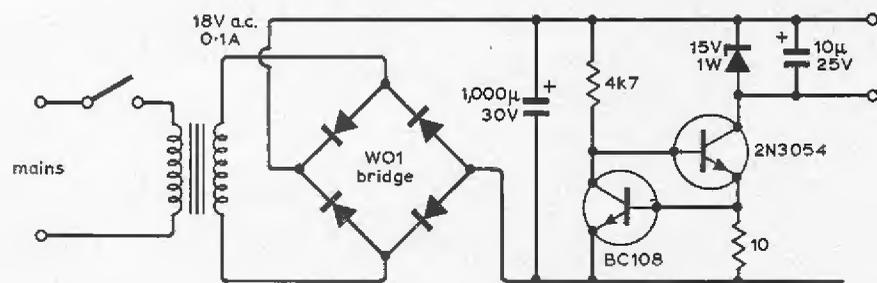


Fig. 6. Alternative discrete component power supply.

in Fig. 2. The response follows the tuned circuit law

$$V_{out}/V_{in} = -H\omega_0(j\omega)/\omega_0^2 - \omega^2 + a\omega_0(j\omega)$$

where $a = 1/Q$, ω_0 is the resonant frequency and the gain of the voltage amplifier is

$$H = 1/3(6.5 - a)$$

The component values are $C_b = C_a/2$, $R_c = 2/\omega_0 C_a$, $R_d = R_c/3$ and $R_e = 2R_c$. The most desirable feature of this circuit is that R and C values are independent of Q , and ω_0 is independent of amplifier gain H . Unfortunately, the Q value becomes extremely sensitive to the value of H for $Q \geq 1$, and the margin of stability in such cases is narrow.

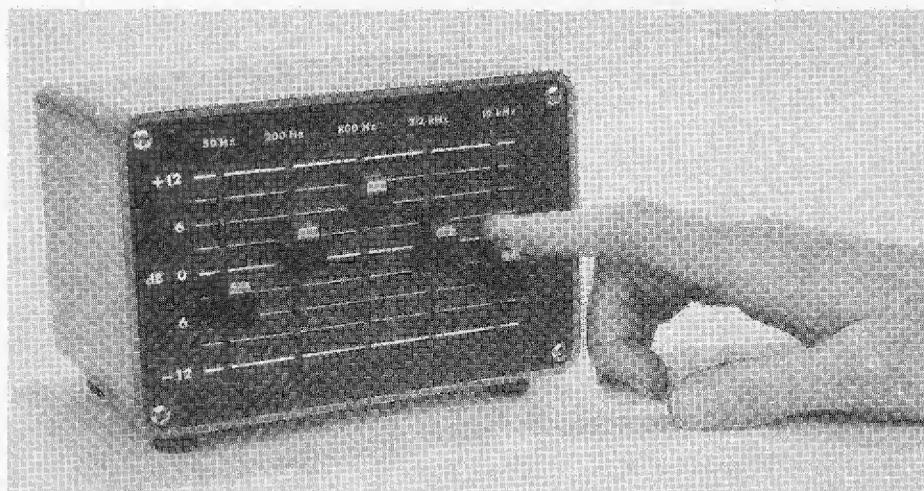
A circuit much more suited to this application⁵ consists of an inverting op-amp with multiple feedback loops (Fig. 3).

The response of this circuit is the same as before, but the component values are $C_c = C_d$, $R_f = 1/H\omega_0 C_c$, $R_h = 1/(2Q - H)\omega_0 C_c$ and $R_g = 2Q/\omega_0 C_c$. If $H = 2Q$, R_h can be left out and V_{out}/V_{in} at resonance becomes

$$V_{out}/V_{in} = -2Q^2 = R_g/2R_f.$$

The minus sign in this expression indicates a phase reversal, so the main amplifier in Fig. 1 will now need an in-phase gain to obtain an overall negative feedback relationship.

The above equations for this active filter circuit show that Q , passband gain and resonant frequency are determined solely by two identical capacitors and two resistors. The demands on the op-amp are not great, and a single bipolar transistor suffices for low Q values. The transistors are used in the circuits to follow to provide a high stage gain in addition to filtering. This reduces the noise contribution of the main amplifier, and so long as the gain is greater than the number of channels used, the dominant noise contribution is that due to the first transistor stage of the active filter. This means a signal to noise ratio of about 70dB,



Five slider potentiometer controls on the front panel can be set to represent the required amplitude frequency response.

Table 1

Channel frequency (Hz)	C_{4-13} ($\mu\text{F} \pm 10\%$)
50	0.22
200	0.056
800	0.015
3.2k	3,900p
12.8k	1,000p

which is better than most preamplifier inputs.

Practical circuit

The circuit is shown in Fig. 4. Only one filter is shown as the other four are identical, except for the values of C_x which are given in Table 1. An emitter follower Tr_1 provides a low impedance input source for the bank of control potentiometers, R_{49-53} . The fixed resistors R_{5-14} restrict the range of boost and cut to $\pm 12\text{dB}$. The active filter consists of a common emitter amplifier with a d.c. coupled emitter follower output. The Q value employed is unity and the passband gain at resonance should be approximately 28dB. The filter outputs are fed to the main amplifier through R_{40-44} which also provide the d.c. bias for this stage. Overall h.f. stability is obtained from the 6dB/octave roll-off provided by C_{15} and R_{45} .

The unit complete with power supply fits into a $16 \times 11 \times 11\text{cm}$ case without difficulty.

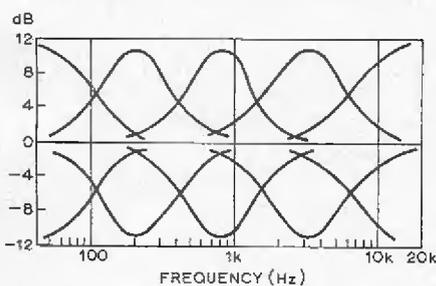
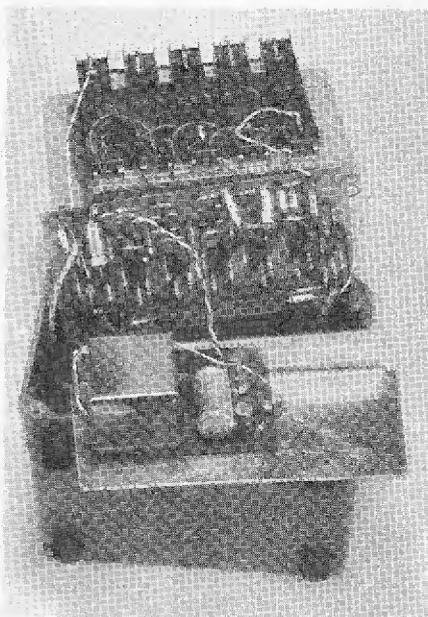


Fig. 7. Typical response curves of the five-channel unit.



Slider pots mounted on the front panel are shown at the top with the main p.c. board (centre), which carries the filter circuitry and input/output devices. The power supply is on a separate board (bottom).

Table 2

Modified filter component values for narrower channel spacing. For worst case component tolerance, channel spacing error is 30%.

Channel spacing (octaves)	Q	R_{15-19} (k Ω)	R_{20-24} (k Ω)	R_{30-34} (k Ω)	Tolerance C_{4-13}, R_{15-24} (%)
1	1.7	3.9	47	6.8	10
$\frac{1}{3}$	4.5	1.5	120	0.68	3

A suitable power supply circuit using a monolithic regulator is given in Fig. 5. An alternative discrete component version, with a "ring of two" constant current generator feeding a zener diode in parallel with the load is given in Fig. 6. Either unit offers a ripple level which is inaudible with the unit in operation.

A small point worth noting is that calibration of the potentiometers in linear dB steps does not strictly follow the mathematical law, although only 10% error is produced over a 12dB range.

More channels

The performance of the five channel unit is encouraging enough to consider expanding the controls in range and number. As for range of control, about $\pm 25\text{dB}$ is the most that would normally be required, even for special effects. One can obtain this range by reducing the values of R_{5-14} down to 2.7k Ω . To make sure that the overload margin is maintained under all conditions, it may be necessary to increase the standing current in Tr_1 and Tr_{13} to provide sufficient drive voltage for the potentiometer bank. Transistors Tr_1 and Tr_{13} may then have to be types of a higher dissipation rating. The number and spacing of controls will depend chiefly on application, thirty controls of one third octave spacing being a reasonable limit. The filter circuits can remain the same as the five channel unit; modified resistor values for two higher values of Q are given in Table 2. Using these modified circuits does not significantly alter the capacitor values for a given frequency, which are given by the equation

$$C(\mu\text{F}) = 1 (\omega_0 \sqrt{R_f \cdot R_g}) \approx 11.8/f_0(\text{Hz})$$

Components

Values for the power supply components are shown in Fig. 5 or Fig. 6.

Transistors

Tr_{13} - BC212 All others BC109

Capacitors

C_1 - 1 $\mu/10\text{V}$ C_{14} - 50 $\mu/10\text{V}$
 C_2 - 10 $\mu/15\text{V}$ C_{15} - 0.005 μ
 C_3 - 10 $\mu/15\text{V}$ C_{16} - 10 $\mu/15\text{V}$
 C_{4-8} } - C_x (see table 1)
 C_{9-13} }

Resistors

R_1 - 47k R_{30-34} - 10k
 R_2 - 47k R_{35-39} - 680
 R_3 - 100k R_{40-44} - 22k
 R_4 - 1k R_{45} - 100
 R_{5-9} } - 3.3k R_{46} - 10k
 R_{10-14} } - 3.3k R_{47} - 4.7k
 R_{15-19} - 6.8k R_{48} - 1k
 R_{20-24} - 27k * R_{49-53} - 10k lin
 R_{25-29} - 100k

*Slider pots are "Alp" type LV60 (mono) or LG60 (stereo). The cheaper Radiohm types are also suitable.

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Bailey-Burrows preamplifier

Flexible design for use with Bailey 30w amplifier

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 Ω), 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 Ω 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 Ω (not 1–M Ω as commonly provided). If 18M Ω 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.

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.

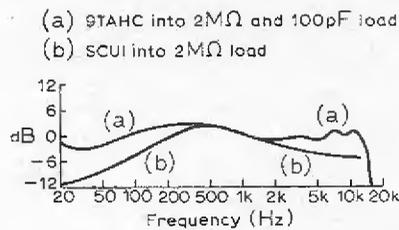
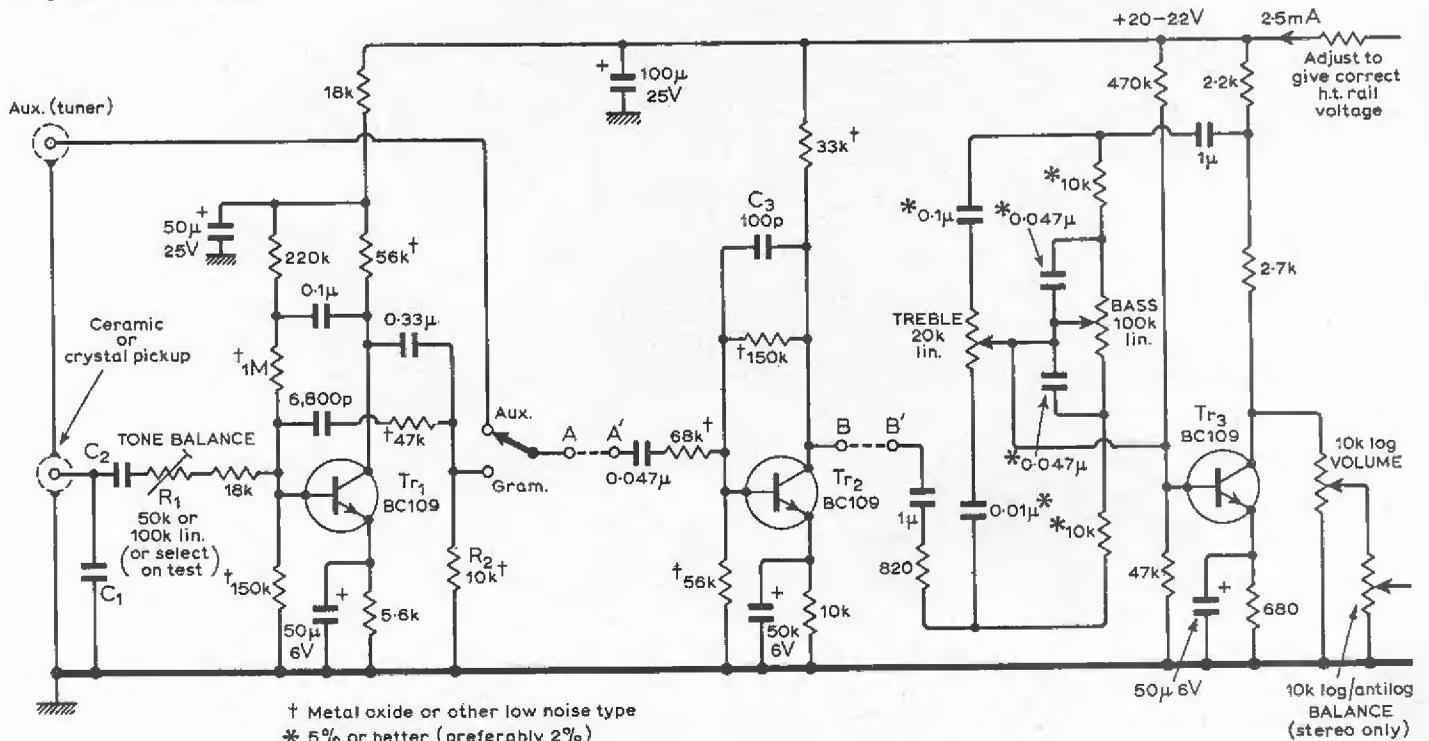


Fig. 1. Voltage/frequency curves of two well-known ceramic cartridges when used with a conventionally-designed pre-amp with $R_{in}=2M\Omega$, and a flat frequency response.

†See Appendix II.



† Metal oxide or other low noise type
* 5% or better (preferably 2%)

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	0.1 μ F	18-27k Ω	low output
Goldring CS90 } Sonotone 9TAHC }			56k Ω	
Connoisseur SCU1 } B.S.R. SC5M }	3.3nF	0.1 μ F	22k Ω	medium output
Acos GP94/1 } Garrard KS40A }			0	
	10nF	8.8nF	22-56k Ω	high output

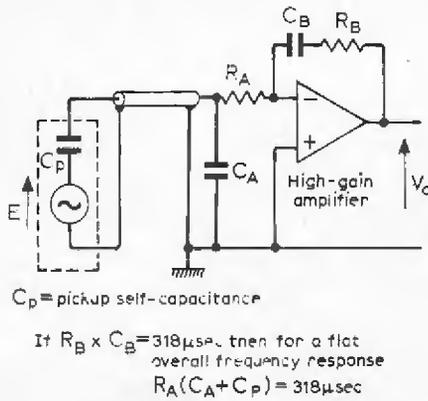


Fig. 3. First-stage design of equalization circuit.

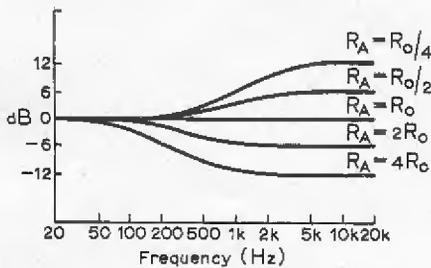
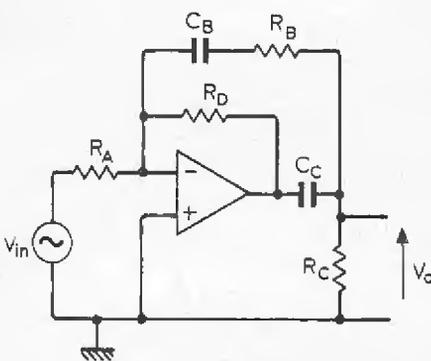


Fig. 4. Operation of tone-balance control, R_A in Fig. 3.



Design formulae for Q=1

- Choose R_C
- Make R_B several times R_C
- $C_B = \frac{1}{2\pi f_1 R_B}$
- $C_C = \frac{1}{2\pi R_C} \left(\frac{1}{f_0} - \frac{1}{f_1} \right)$
- $R_D = R_B \left(\frac{C_C R_C + C_B R_B}{C_C R_C C_B R_B} - 1 \right)$

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 75k Ω then gives an input time constant of 318 μ s (equivalent to 500Hz); to match this, the feedback circuit has a time constant of 318 μ s 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'. 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 SCU1 needs $R_A = R_0/4$.

With an infinite gain amplifier 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. Baxandall². 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 $C_1 C_2 / (C_1 + C_2) \approx 4000\text{pF}$	
input impedance	not applicable (68k Ω for aux input connected as shown)
disc equalization	In conjunction with the better ceramic pickups can be adjusted to flat $\pm 1.5\text{dB}$ 30Hz-10KHz. Low-frequency performance independent of pick-up capacitance.
rumble filter	18dB/oct, $f_0 = 50\text{Hz}$ independent of pick-up capacitance
low-pass filter	fixed, $C_3 = 100\text{pF}$ gives $f_{3dB} = 12\text{KHz}$ Scale C_3 up in proportion for low f_{3dB}
tone controls	h.f. about $\pm 14\text{dB}$ l.f. about $\pm 14\text{dB}$.
current consumption	$\approx 2.5\text{mA}$.

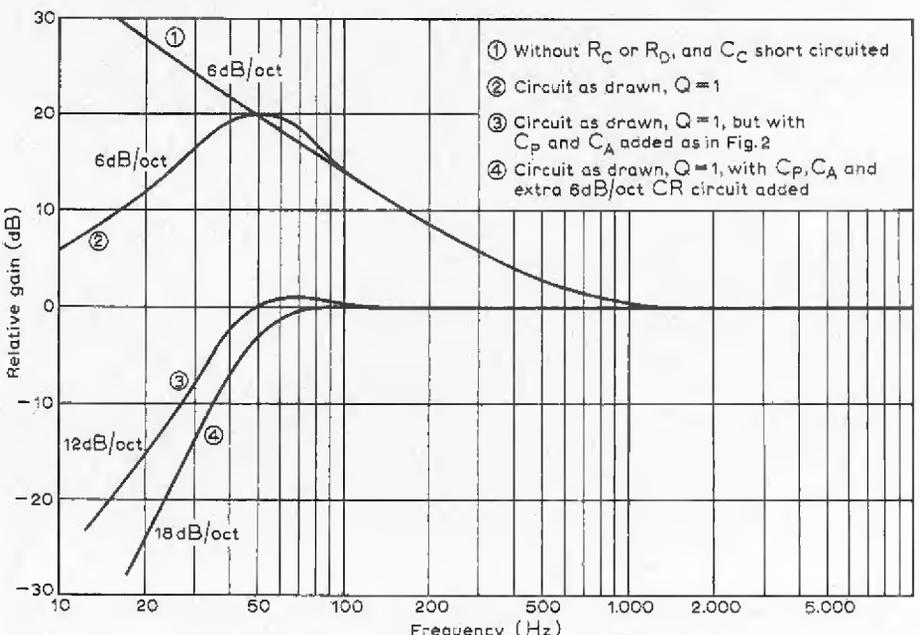


Fig. 6. Performance of circuit of Fig. 5 with $f_0 = 50\text{Hz}$ and $f_1 = 500\text{Hz}$.

30-watt high fidelity amplifier

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.

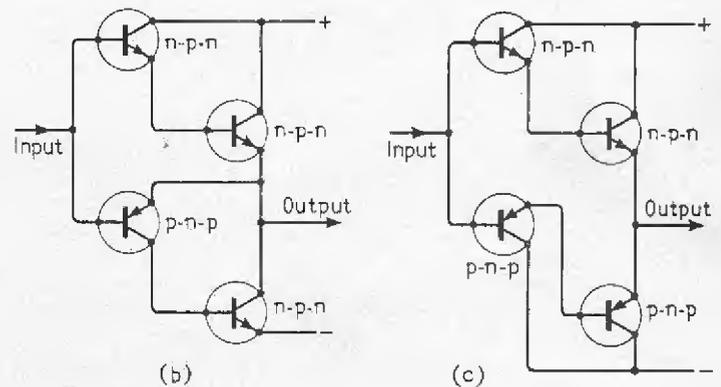
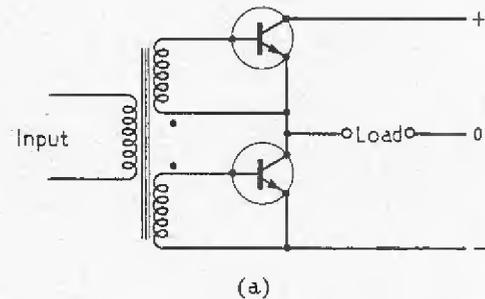


Fig. 1. Direct-coupled output stages: (a) with driver transformer; (b) quasi-complementary; (c) fully complementary.

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

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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', 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.

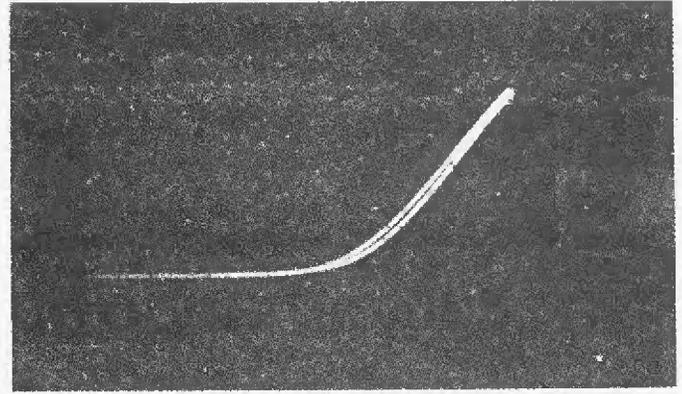


Fig. 2. Transfer characteristic of upper half of Fig. 1(b).

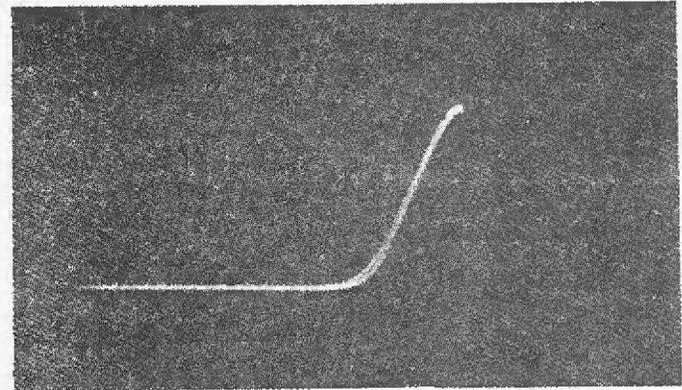
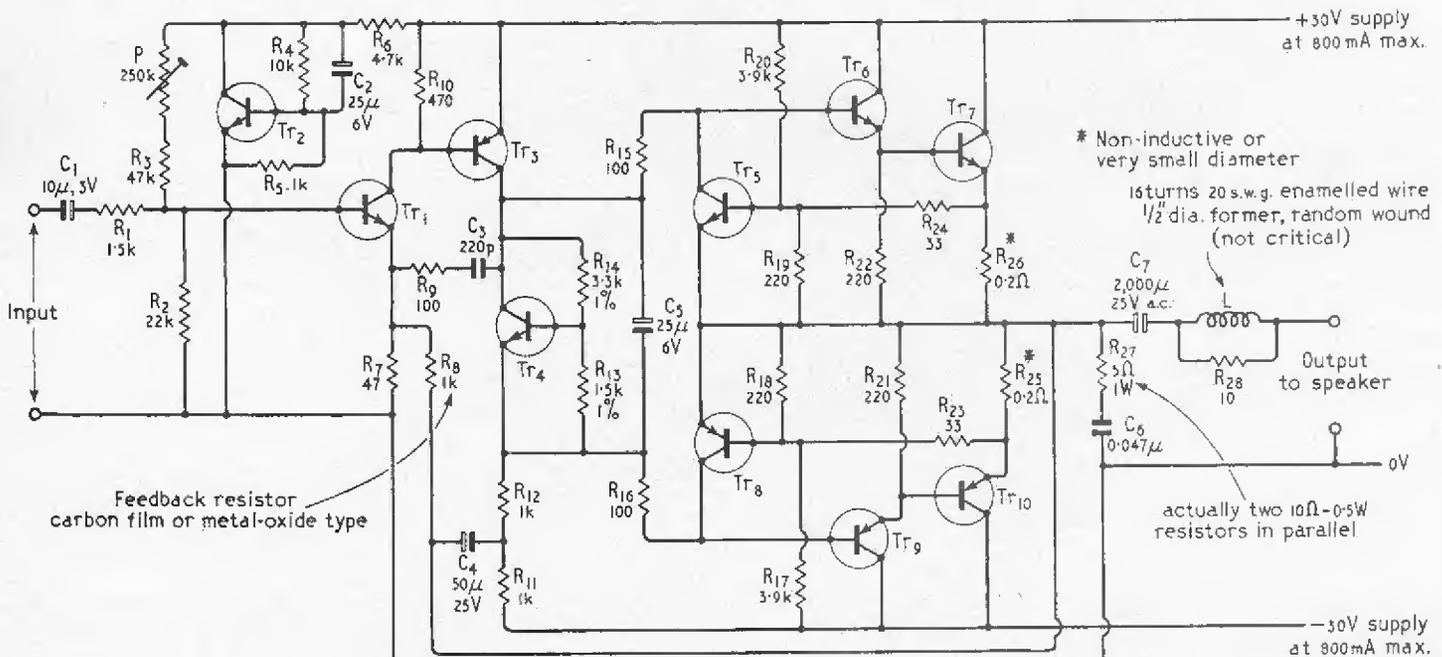


Fig. 3. Transfer characteristic of lower half of Fig. 1(b).

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 kHz 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- μ F polarized electrolytics connected 'back-to-back'.



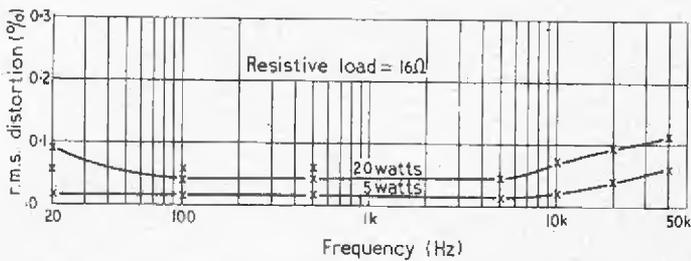


Fig. 5. Distortion characteristics of amplifier with 16-ohm load.

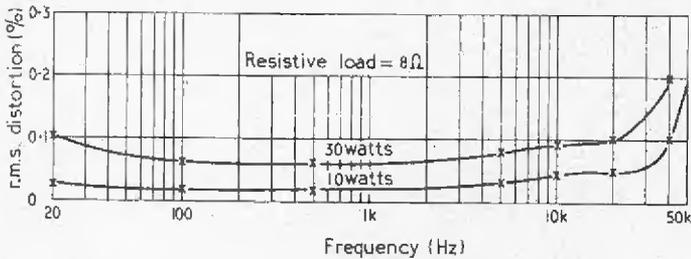


Fig. 6. Distortion characteristics of amplifier with 8-ohm load.

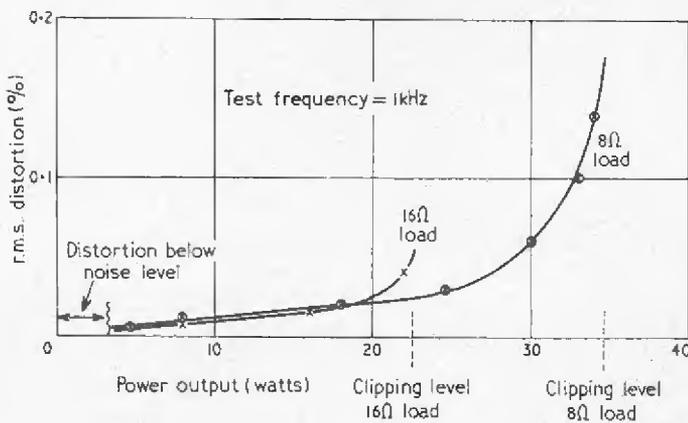


Fig. 7. Variation of distortion with output power level.

current. This transistor, T_{r2} , 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, T_{r1} , 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 T_{r1} . 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

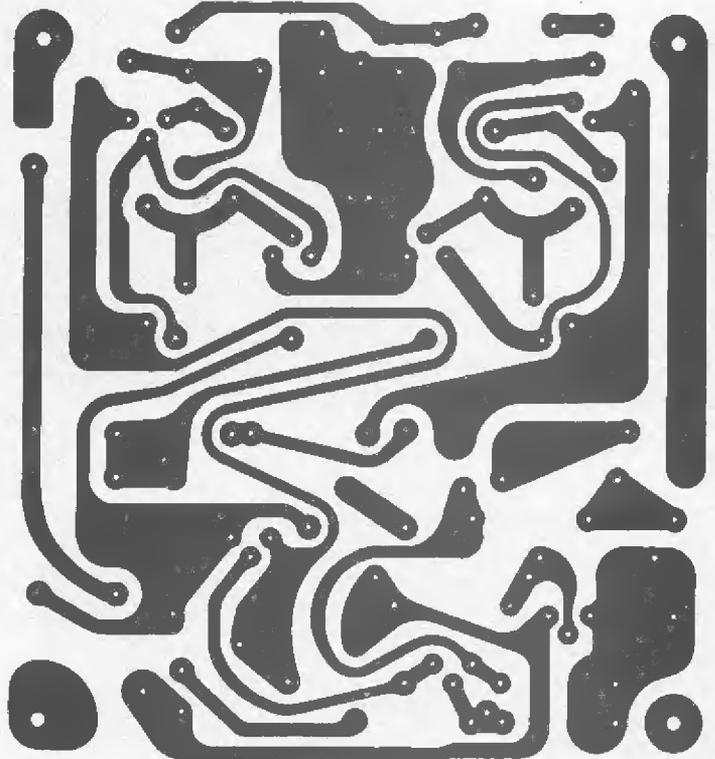
the ultrasonic region. Equally the inductor in series with the output lead, which improves the stability with capacitive loads, need have only a very small inductance. This wide bandwidth gives exceptional high-frequency performance as can be seen from the distortion figures in Figs. 5, 6 and 7. Unfortunately, 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.

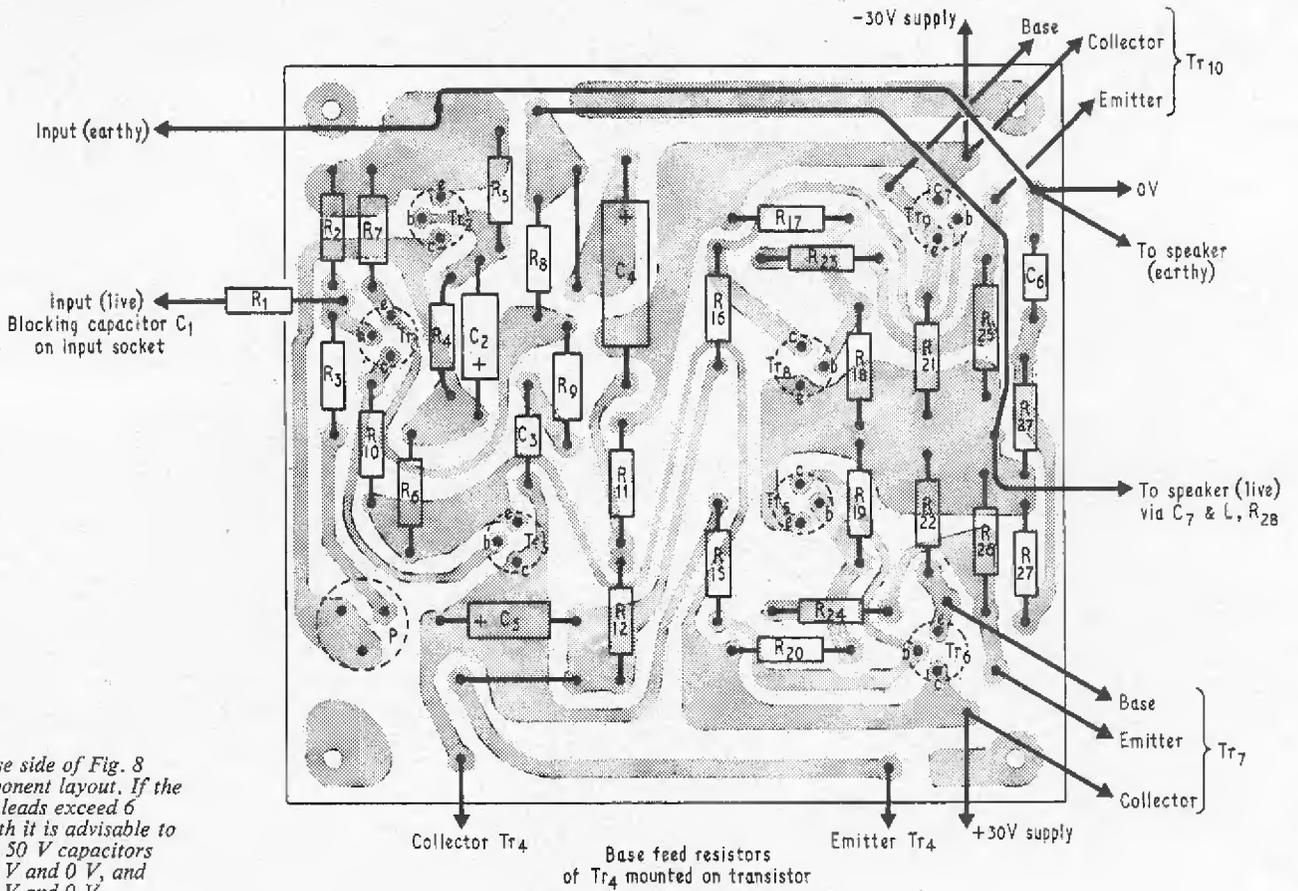
The overall performance of the amplifier is very good, considerably better in fact (on paper) than the best valve amplifiers. Unfortunately, listening tests have shown that the performance of the amplifier is only slightly, if any, better than the best valve amplifiers. Extensive listening tests indicate 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.)





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



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

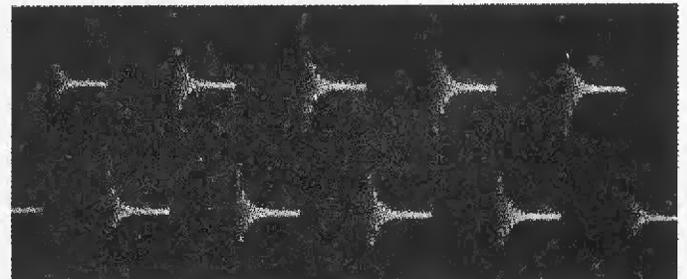
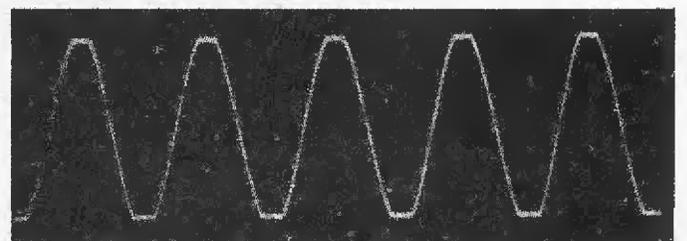


Fig. 13. Overdrive with sine-wave input, showing clean limiting 1 kHz and 8-ohm load.)



30-watt amplifier modification

Operation from a single supply rail

by Arthur R. Bailey*, M.Sc., 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 2000 μ F 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-k Ω 1-W type, and its associated resistor R_7 is increased to 68 ohms.
4. To maintain d.c. balance in the amplifier, R_8 is increased to 10 k Ω and R_4 to 22 k Ω .
5. Owing to the changes produced in 4 above, C_2 is now 10 μ F 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

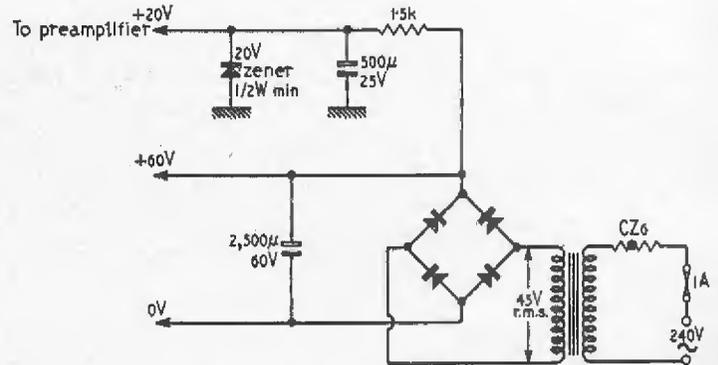
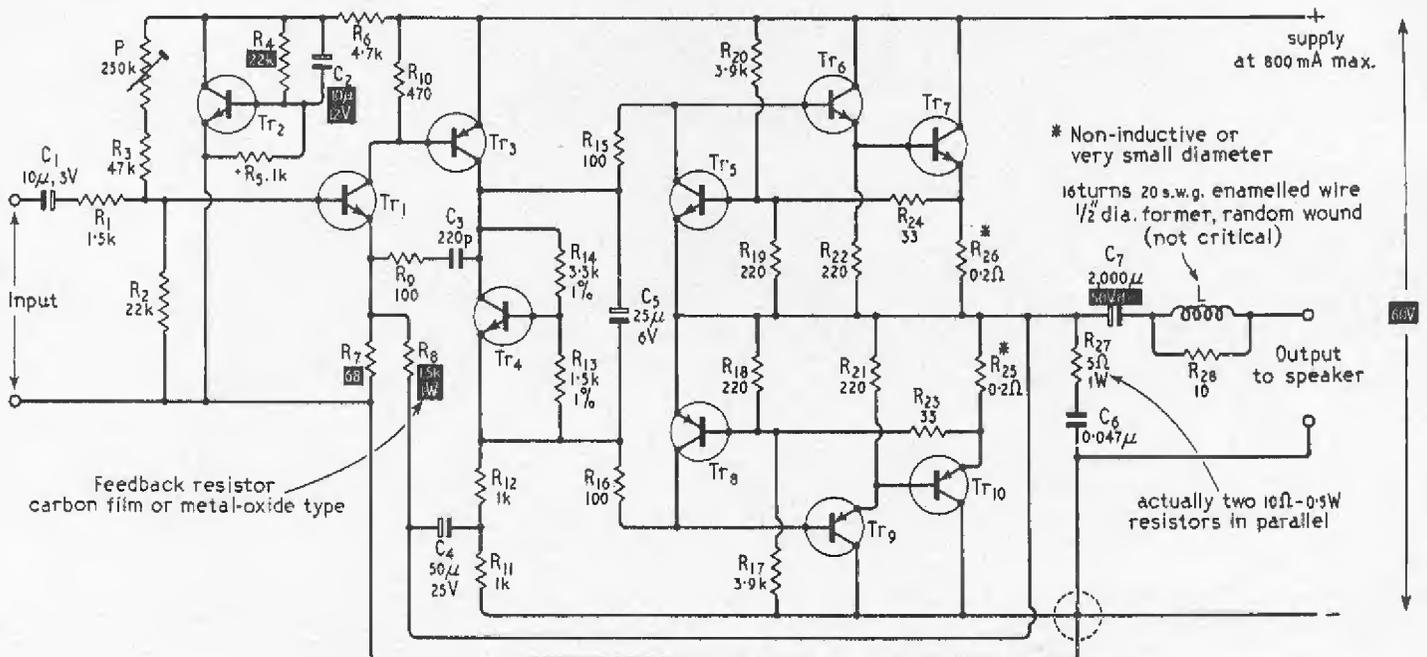


Fig. 2. A suitable power supply for the modified circuit.

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. 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.



* Non-inductive or very small diameter
16 turns 20 s.w.g. enamelled wire
1/2 dia. former, random wound
(not critical)

actually two 10 Ω -0.5W resistors in parallel

Feedback resistor
carbon film or metal-oxide type

Baxandall tone control revisited

Improvements for greater flexibility in tailoring audio signals

by M. V. Thomas, B.A.

The main disadvantage of the Baxandall tone control circuit is that if boost and cut are required at a particular frequency a much greater effect occurs at the extremes of the audio range. Modifications are described to limit the degree of this effect.

The Baxandall configuration has for some time been the almost universal choice of audio amplifier manufacturers for their tone control circuits. This is due in no small measure to its simplicity of construction and ease of use, and it is difficult to envisage any improvement of the circuit whilst retaining only two controls. However, once a decision to increase the number of controls is made, the field becomes wide open, the most obvious development being to have each control affecting the level of a limited band of frequencies. Such circuits are obviously much more elaborate than the basic Baxandall type, and require careful design to prevent excessive interaction between the controls. This article describes a modification to the basic Baxandall circuit which greatly increases its versatility whilst maintaining its simplicity.

The main disadvantage of the basic circuit is that it has its greatest effect at the extremes of the audio range, as shown in Fig. 1. For example, if a 6dB boost is required at 4kHz, one must simultaneously tolerate a much greater boost of perhaps 18dB at 16kHz. Furthermore, the turnover frequency of the bass control depends on its setting, but this does not apply to the treble control! This effect is also shown in Fig. 1, and the reason can be seen in Fig. 2, which shows the basic circuit. At very low frequencies the impedances of the capacitors are high and the circuit is essentially resistive. The bass control then acts as a simple gain control and the treble control has no effect. As the frequency is increased, the bass control is progressively decoupled by C_1 and C_2 , the relevant time constants being C_1R_3 and C_2R_2 . But the relative values of R_2 and R_3 obviously depend on the setting of the control, this causing the variation in turnover frequency as shown in Fig. 1. For example, to increase the bass boost, R_2 must be increased, thereby increasing the C_2R_2 time constant and increasing the turnover frequency. At higher frequencies (above 1kHz) the bass control is completely decoupled by C_1 and C_2 , and the impedance of C_3 has fallen to a value where the treble control begins to

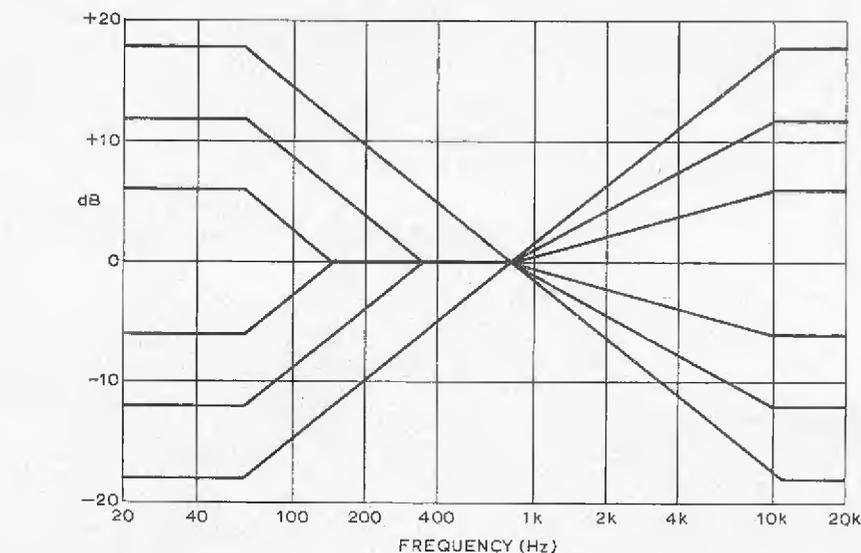


Fig. 1. Idealized responses of the Baxandall-type tone control circuit.

have a significant effect (capacitor C_3 is often replaced by two capacitors, one at each end of the track of the treble control, but the effect is basically the same). Resistor R_5 prevents the bass control from loading the treble control, and the time constant which primarily decides the treble turnover frequency is C_3R_5 , which is independent from the control settings. At very high frequencies (above 10kHz) the treble control acts as a gain control. Resistors R_1 , R_4 , R_6 and R_9 serve to limit the effect of the controls at their extreme settings, and are comparatively small in value, so they do not affect the basic operation of the circuit.

Whether this difference in mode of action of the bass and treble controls, is advantageous is a difficult question, and one could no doubt argue either way. However, it is about all one can do with a simple RC network, unless completely separate boost and cut controls are used¹. Using this approach it is possible to synthesise "step" responses as shown in

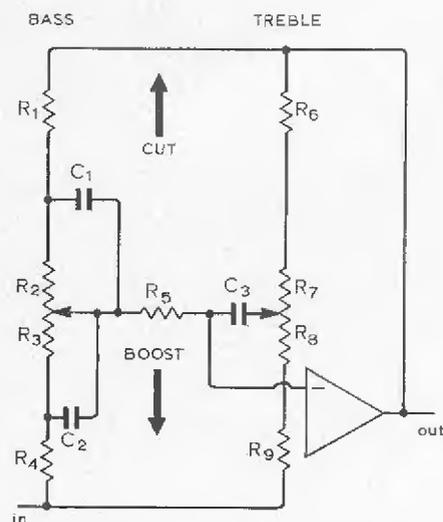


Fig. 2. Baxandall-type tone control network.

Fig. 3. Synthesis of "shelf" responses using completely separate boost and cut networks. In the example shown, the maximum overall boost is +12dB (curve C), but at 20kHz this overall boost is obtained by applying 40dB boost and 28dB cut (curves A and B), thereby causing overload or noise problems, depending on which operation is performed first.

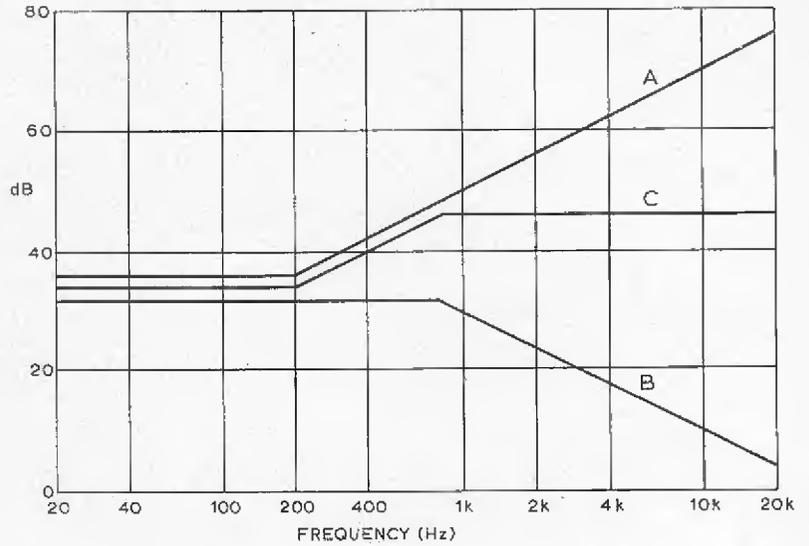


Fig. 3, but in order to obtain a flat response it is necessary to simultaneously boost and cut the signal, which can cause overload and noise problems as the two operations are performed in different parts of the circuit, in contrast to the Baxandall arrangement.

Possible modifications of the circuit to limit its effect at the extremes of the audio range were therefore considered. It was decided that the most useful addition would be of separate "effect" controls for bass and treble, which would limit (in a symmetrical fashion) the maximum degree of boost and cut obtainable from the bass and treble controls, and further reference to Fig. 1 shows how this may be done. The maximum boost and cut of the bass control are decided by R_1 and R_4 respectively, so the desired result could be obtained by replacing these with variable resistors, but this arrangement has two disadvantages. Firstly, two controls are required, and secondly, changing the values of these resistors will cause some change in the turnover frequency of the bass control. However, the same result can be achieved with neither of these disadvantages, by connecting a single variable resistor directly across the bass control, as shown in Fig. 4. This control

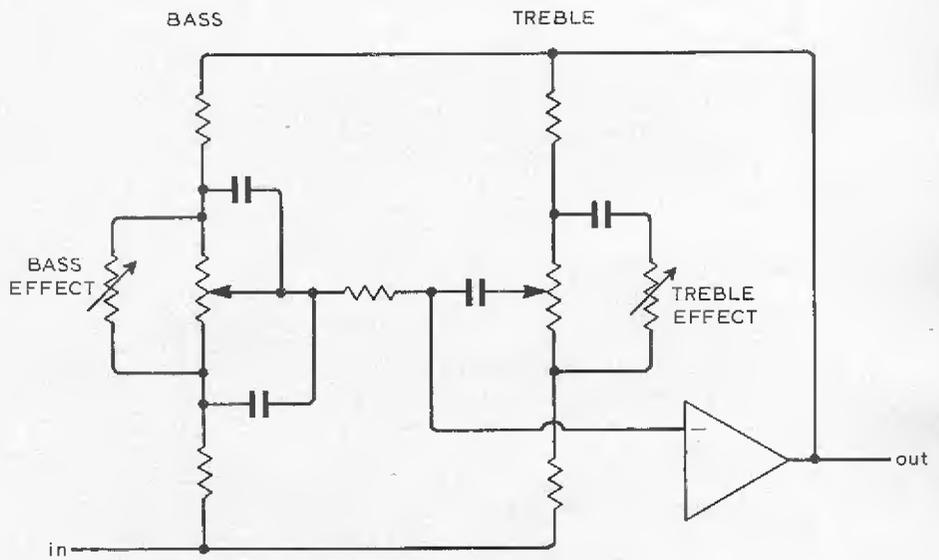


Fig. 4. As Fig. 2, but with the addition of the two effect controls.

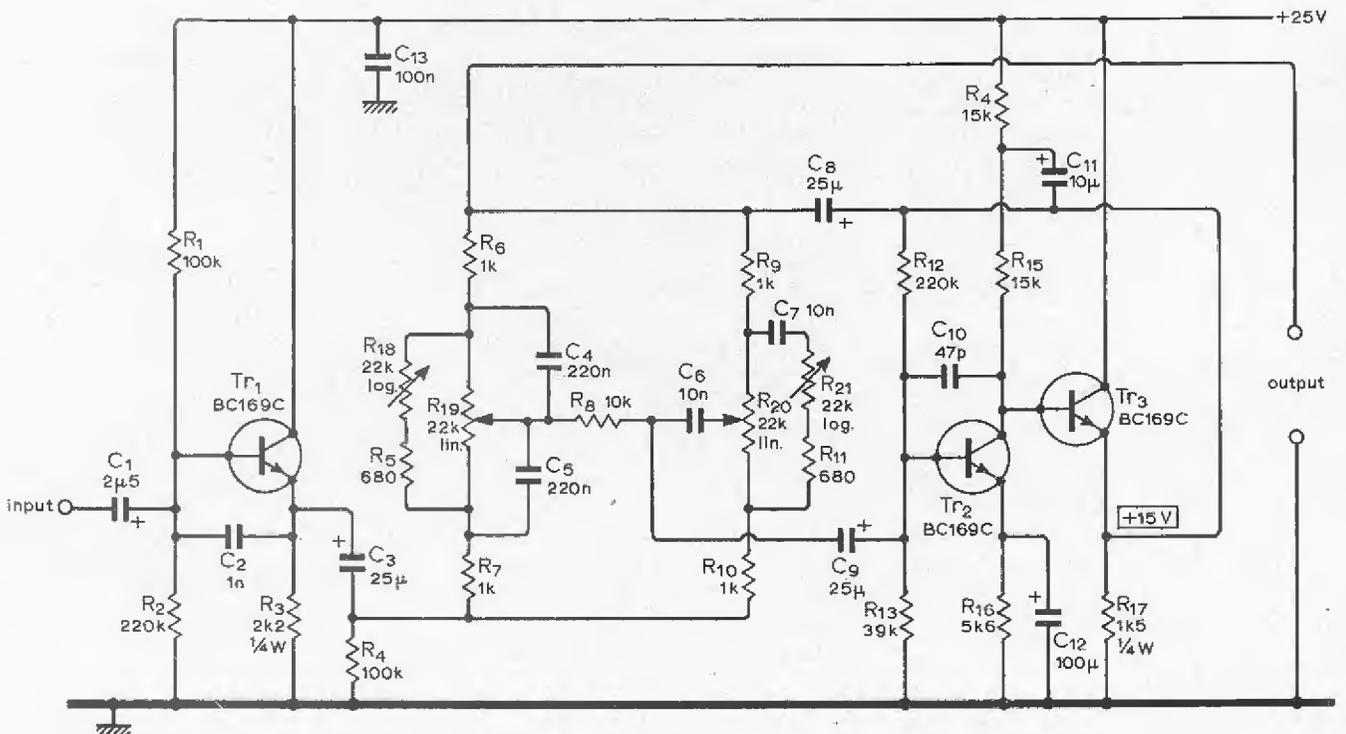


Fig. 5. Complete circuit. Resistors can be $\frac{1}{8}$ watt unless shown otherwise.

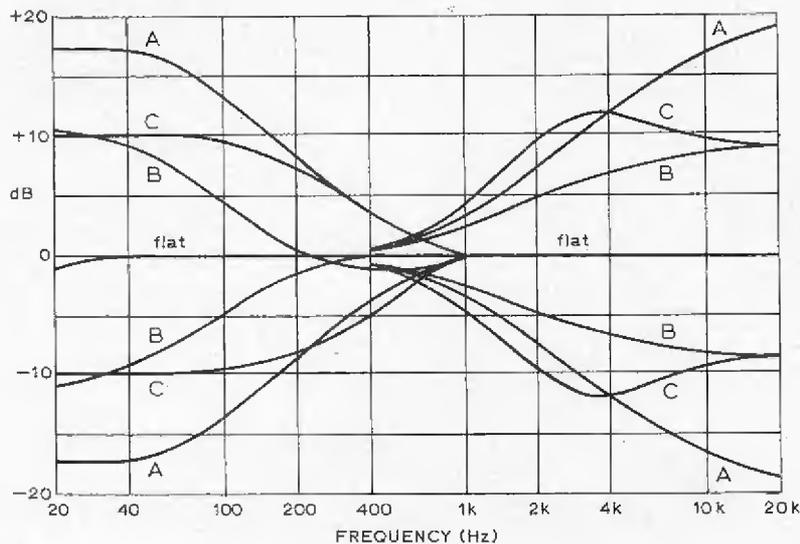


Fig. 6. Selection of the frequency responses obtainable with the circuit. See text for further details.

simply acts as a potential divider in conjunction with R_1 and R_4 . As the resistance of the control is reduced, the fraction of the input and feedback signals appearing across the bass control is also reduced, thereby limiting its maximum effect. A similar modification to the treble control will not have the desired result, because of the fixed turnover frequency of this control; it would indeed reduce the maximum boost and cut of the control, but one could obtain exactly the same frequency response by removing the "effect" control and by having the treble control at a less extreme setting! This state of affairs can be prevented by including a series capacitor, as shown in Fig. 4, so that the "effect" control is operative only above a turnover frequency decided by the values of this capacitor, R_6 and R_9 .

The complete circuit incorporating these modifications is shown in Fig. 5, and apart from the additions it is quite conventional. The only extra precaution necessary is to ensure that it has a reasonably high drive current capability, as the impedance of the control network can be comparatively low at some control settings. However, the worst-case maximum output of the circuit is approximately four volts r.m.s. before clipping, which should be perfectly adequate. The transistors used in the prototypes were BC169Cs; these have a V_{ces} of 30V, which allows the use of a fairly high supply voltage for the circuit. Transistor Tr_1 provides a low impedance drive to the network while Tr_2 and Tr_3 form a bootstrapped amplifier. The low distortion and low output impedance of this configuration make it an ideal choice for this application². The circuit has a gain of unity with the controls set flat, and C_2 and C_{10} reduce the r.f. gain to prevent instability. Resistors R_{12} and R_{13} provide d.c. feedback to hold the emitter of Tr_3 at 15V, this being a useful point to check when testing the circuit. Logarithmic pots are recommended for the two effect controls. The "top" end of each track should be left unconnected so that the

controls will then have their smallest modifying effect when fully clockwise. Resistors R_5 and R_{11} in Fig. 5 prevent the effect controls from completely swamping the bass control and treble controls when the former are fully anticlockwise; the values shown set the limits at the audio extremes to ± 4 dB.

Fig. 6 shows a selection of the frequency response curves which can be obtained from the circuit. The set of curves marked "A" obtained with the effect controls fully clockwise shows the responses with the bass and treble controls at their extreme settings, and set "B" is similar except that this shows the responses with the bass and treble controls at approximately half maximum settings. These curves are almost identical to those obtainable from a conventional circuit—compare for example with those of the tone control circuit in ref. 3. The only difference is that the bass and treble responses have been deliberately arranged to overlap rather more than usual, so as to take fuller advantage of the effect controls. Set "C" is obtained with the bass and treble controls at their extreme settings but with the effect controls set to limit the responses at the audio extremes to the same as those of set "B", and this clearly shows the advantages of the extra controls. Basically, by the use of these controls it is possible to alter the level of a band of frequencies far more uniformly than with a conventional circuit, and this advantage is very noticeable in use, being considered well worth the extra complexity.

References

1. Hutchinson, P. B., "Tone control circuit", *Wireless World*, Nov. 1970, pp.538-540.
2. Quilter, P. M., "Low distortion tone control circuit", *Wireless World*, April 1971, pp.199-200.
3. Walker, H. P., "Stereo mixer", *Wireless World*, June 1971, Part 2, pp.295-300.

Active crossover networks

Using a single bass speaker in a stereo system

by D. C. Read, B.Sc.

The author finds that active crossover networks, giving better speaker damping and improved transient response, sound better than passive ones as well as being easier to adjust. But the system first designed (December 1973) needed six amplifiers. This article therefore describes three ways to reduce size and cost, by using a common bass unit with 1-cu.ft sealed enclosures, by using the bass unit with 0.33-cu.ft enclosures, and by omitting the bass unit. Alterations to active filter circuit values for these options are given together with a modification to the "full" six-unit system.

The first and most obvious economy in the previous active crossover design was to follow the well-tried example of Baxandall* using the fact that low-frequency sounds are non-directional and have a mixed mono and stereo system. Fortunately, for modern living rooms of average size, this non-directional effect extends well into the audio band. By combining a single bass speaker with a pair of relatively small mid-range/tweeter boxes working in stereo there is an immediate saving of one power amplifier, its driving filter circuit and one large enclosure.

Although one filter and amplifier assembly would be unchanged, the other would be smaller, with about one-third of the components removed from the filter (front board in photograph) and two power amplifier boards instead of three. An added practical advantage is that the bass speaker can be placed in any convenient position to suit the general room layout.

Choice of bass crossover

Having decided on a combined bass arrangement, the first requirement was to determine where in the audio band to make the change between mono and stereo operation. In doing this work, the ease of adjustment afforded by active filters was most helpful.

The single bass unit, KEF type B139, was installed in a box of about 5cu.ft and then subjected to listening tests in a normal-size living room. For experimental purposes, the middle and upper frequency sounds were produced by units in the transmission-line enclosures described in the previous article. Simple

A-B switching allowed repeated comparison between this combined bass set-up and the complete 2 × 3-way system, i.e. between mixed mono and stereo, and full audio-band stereo.

Initially, the bass speaker was fed with combined signals from the two channels up to about 490Hz: this was the bass crossover frequency already in use for the transmission-line speakers.

But stereo separation was not then maintained over the whole band: the single source of low frequencies could be identified and located when moved. The crossover point was then lowered by successive amounts until, with it set at 150Hz, the mono bass effect disappeared. Thus, when compared against the full 2 × 3-way stereo it was found that the l.f. speaker could be placed virtually anywhere in the room, facing in any direction, without noticeable change in the stereo image.

The 150-Hz crossover frequency determined as above provided for satisfactory

(i.e. undetectable) mono bass for a stereo installation working in a room measuring about 15 feet square. In a listening room with dimensions considerably greater than this, however, the presence and location of the single source of l.f. may become apparent, and it will then be necessary to lower the stereo-to-mono changeover point.

To move the changeover point either way, the response curves of both the mid-range band-pass filter (high-pass) and the bass-unit low-pass filter must be modified by altering the roll-off frequencies of the 6dB and 12dB/octave circuits (the circuits surrounding IC_3/Tr_5 and IC_4/Tr_6 in Fig. 2). As the change of frequency in each instance is likely to be relatively small, the appropriate new component values can be deduced by simple linear scaling. For readers who wish to re-calculate circuit constants, perhaps to satisfy the requirements of different units or considerably changed listening conditions, the pro-

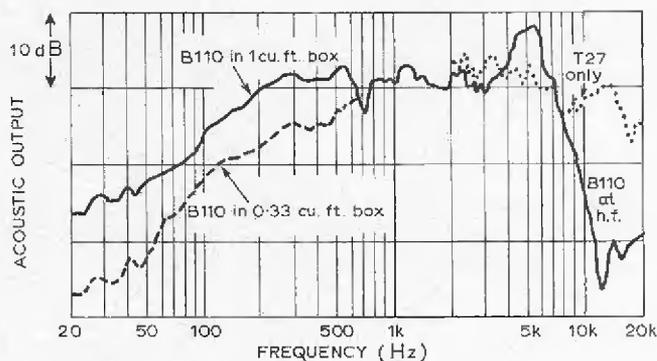


Fig. 1. For the common-bass system the mid and upper-ranges use B110 and T27 drive units, as in the original design. Graph shows axial response of B110 in a 1-cu.ft sealed box (upper solid line). T27 curve (broken line) shows an upper crossover of about 4kHz is suitable. Lower B110 curve shows axial response for a small, 0.33-cu.ft sealed box, for which slope correction is needed (see Fig. 5). Acoustic response curves taken with a writing speed of 80mm/s and a paper speed of 3mm/s.

*Low-cost high-quality loudspeaker, (part 2) by P. J. Baxandall, *Wireless World* vol. 74 1968, pp.316-9.

cedure is given in a section at the end of this article.

The component-scaling process is straightforward. In all instances, suppose the existing crossover frequency is f_1 , and the wanted frequency is f_2 . Suppose also, for example, that the required frequency shift is that necessary to lower the crossover point (to raise it each multiplying factor would, of course, be inverted). In the feedback circuit to input 3 of IC_3 (Fig. 2), either the two C values (33nF) or the R values ($2 \times 56k\Omega$ and $36k\Omega$) must be increased in the ratio f_1/f_2 . For the Tr_5 input circuit increase either the series C (to $f_1/f_2 \times 0.033\mu F$) or the two shunt R s (both to $f_1/f_2 \times 62k\Omega$).

Similarly, in the IC_4 circuit, new values are found for the input 3 components such that either the $100k\Omega$ and $15k\Omega$ resistors or the 33 and $68nF$ capacitors are increased by the factor of f_1/f_2 . The Tr_6 input circuit is modified in the same way as for Tr_5 .

As a further practical point relating to the use of a single, combined bass system as suggested here, note that the power amplifier described in the first article (Fig. 4 of the December article) is

easily able to provide the necessary drive for a single unit taking l.f. signals from both channels, or even from four channels.

Smaller mid and upper range stereo units

With the optimum changeover point between stereo and mono working now established, it was necessary to construct a pair of small enclosures with speakers to produce the mid and upper audio frequencies. Identical units to those already used (KEF types B110 and T27) were used.

An obvious minimum requirement here was that the bass response of the mid-range unit must extend at least to 150Hz. Past experience suggested that for this an enclosure volume of about 1cu.ft would be suitable. Accordingly, a sealed box of this size was made, with dimensions $10 \times 20 \times 9$ in. Damping was provided by a thick coat of car underseal on all internal surfaces and the box was filled with about $\frac{1}{2}$ lb of long-fibre wool: an inside layer of 2-in. foam plastics material would be a suitable alternative.

With the units installed, the assembly was tested in non-reverberant surroundings. As the B110 axial response given in

Fig. 1 shows, the arbitrary estimation of enclosure volume was about right and obviates the need for slope equalization in the mid-range band-pass filters. The T27 response curve, drawn dotted in Fig. 1, shows that an upper crossover frequency of 3.5 to 4kHz was suitable here; the overall 6dB/octave slope is easily offset by making the high-pass filter l.f. roll-off more gradual.

Fig. 3 illustrates the three active filter responses as finally set for the mixed system. These curves are the voltage outputs from the full circuit given in Fig. 2, with $R_2 = 33k\Omega$, R_7 shorted and R_3, R_6, C_1 omitted. Note that two each of the band-pass and high-pass filters shown in Fig. 2 are needed but only one low-pass circuit

Overall response with combined bass

To check the overall acoustic output of the total system, the bass speaker together with a 1-cu.ft mid-range/tweeter assembly was tested in non-reverberant surroundings. Fig. 4 shows the complete response.

Note that only one of the stereo pair was used to obtain the curve of Fig. 4 because large differences in phase response

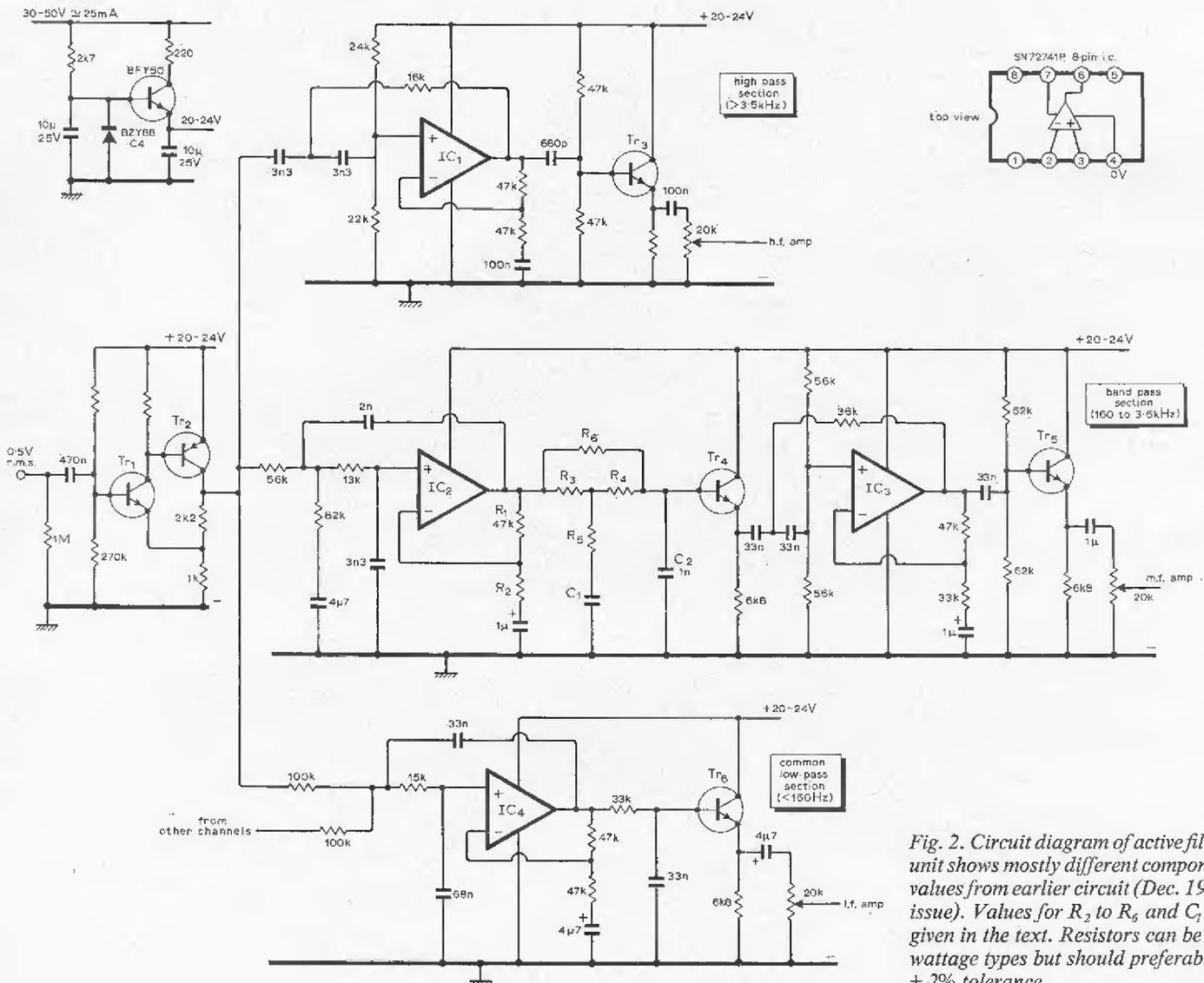


Fig. 2. Circuit diagram of active filter unit shows mostly different component values from earlier circuit (Dec. 1973 issue). Values for R_2 to R_6 and C_1 are given in the text. Resistors can be low-wattage types but should preferably be $\pm 2\%$ tolerance.

Fig. 3. Active filter responses for common-bass system. Curves are voltage outputs from Fig. 2 circuit, with $R_2 = 33k\Omega$, R_3 shorted and R_5, R_6, C_1 omitted.

Fig. 4. Axial acoustic output for B139 unit in separate l.f. enclosure and 1-cu.ft box using B110 and T27, in non-reverberant surroundings.

Fig. 5. Correction circuit used in bandpass filter to compensate for response of 0.33-cu.ft box in Fig. 1. Other components in (a) give part of the 18dB/octave roll-off at the top of the passband (C_2, R_4) and R_6 determines in-band slope (see text for value of R_2).

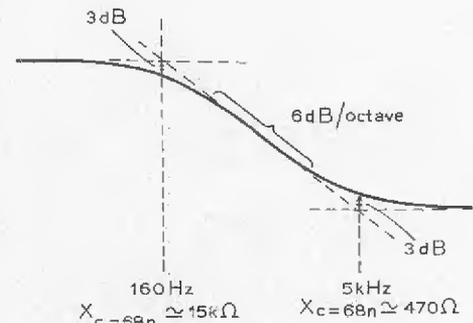
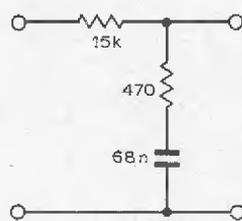
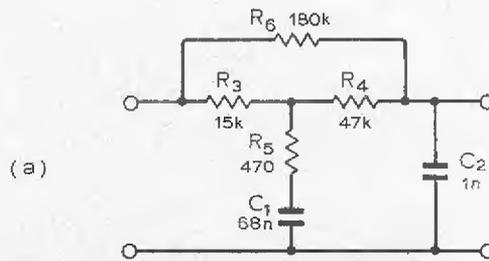
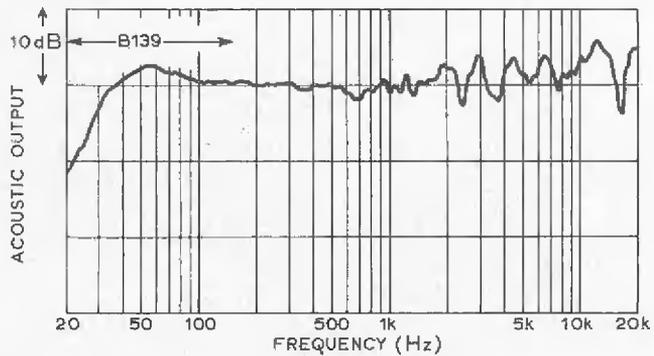
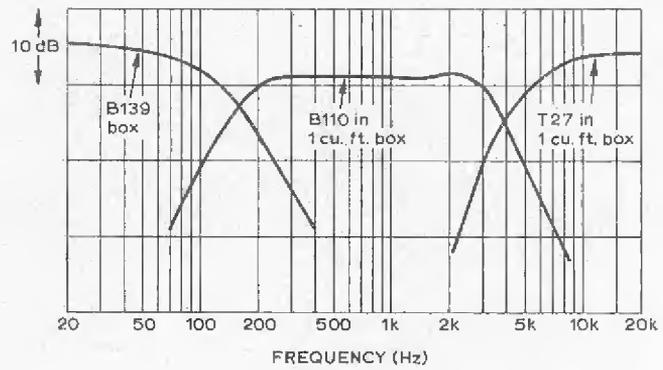
between individual units of the same type (e.g. as exhibited by different examples of the B110 unit) resulted in a spikey response with almost complete cancellation at some points. Such disparity in phase performance is, of course, a well-known failing of cone radiators; plane-driven electrostatic speakers are much more consistent in this respect. However, in normal domestic surroundings, multiple reflection tends to smooth out such variations in combined response and therefore largely overcomes the problem.

Active-filter phase response

While on the subject of phase performance, it should be mentioned that this was an aspect of active-filter operation which was carefully considered at an early stage. For the full stereo installation outlined in the December article, the various filter phase responses were calculated and their combined effect on acoustic performance tested. Specifically, this was compared to performance with the conventional passive-network equivalent. Detailed results of these tests have not been given because they showed only that even the largest differences in phase characteristic, particularly those occurring in the crossover regions, had little effect on sound output. The one observable change was a frequency redistribution, without increase in amplitude, of troughs and peaks in the overall response.

Even smaller stereo speakers

Although the combined-bass speaker enclosure used in the mixed system can be of any suitable shape and put in any convenient out-of-the-way corner, the mid- and upper-range assemblies must of necessity occupy particular positions, divided and directed so as to obtain the intended stereo effect. But, in rooms where space is at an absolute premium, even the 1-cu.ft enclosures used as above may be too large. To meet this objection and to exploit other properties of active filter, notably their possible use in correcting speaker response deficiencies, a pair of 0.33-cu.ft boxes was built (dimensions $8 \times 6 \times 12$ in) to form part of a modified stereo/mono system using the same units



as before, i.e. one B139 and two each B110, T27.

Mid-range unit slope correction

The l.f. axial response of a B110 in a 0.33-cu.ft enclosure is given by the lower, dashed, curve up to 1kHz in Fig. 1. This shows that slope correction is needed in the band-pass filters which provide input signals for the mid-range units. An R-C network feeding T_r can be used for such equalization. In Fig. 5a, the correction network is re-drawn with appropriate values. Here, R_3, R_5 and C_1 give a 6dB/

octave bass lift over the required frequency range as shown in Fig. 5b.

Components R_4 and C_2 provide a similar slope as part of the overall 18dB/octave roll-off at the top end of the pass band. The value of R_6 determines the amount of in-band slope obtained for the whole circuit.

Using the values shown in Fig. 5a, the band-pass filter output response is as shown full line in Fig. 6. The other two filter responses remain as before but are included here to show how they relate to the modified band-pass curve.

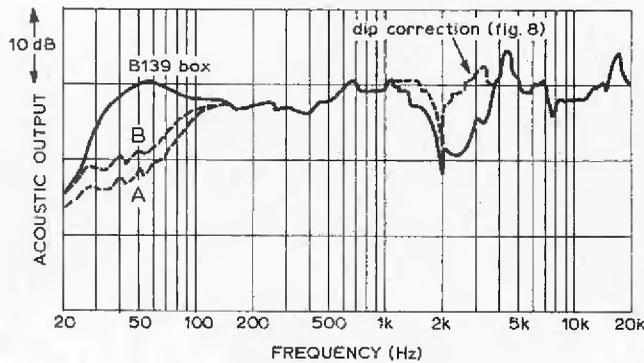
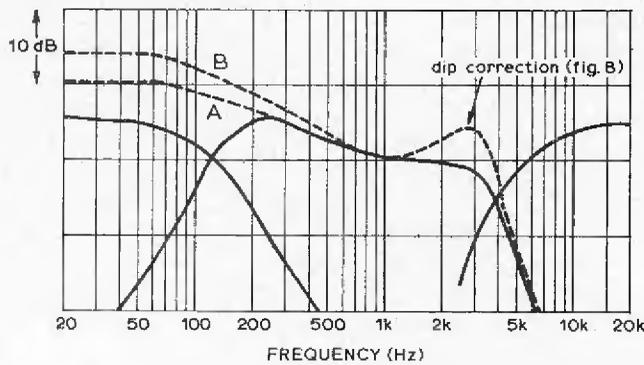


Fig. 6. Filter response for small enclosure, using values of Fig. 5 (a). Curves A and B are obtained with high-pass part of band-pass filter removed (by taking output from Tr_4 emitter) and apply for an inexpensive system using only the 0.33-cu.ft enclosures (see Fig. 7).

Fig. 7. Axial acoustic response for B139 unit in separate enclosure and 0.33-cu.ft box, in non-reverberant surroundings, showing dip at 2kHz before and after correction. Curve A applies when common-bass unit is not used, and high-pass filter of bandpass section is removed (to the right of Tr_4). Improvement at B is obtained by making C_1 100nF and R_6 270k Ω .

Mid-range dip correction

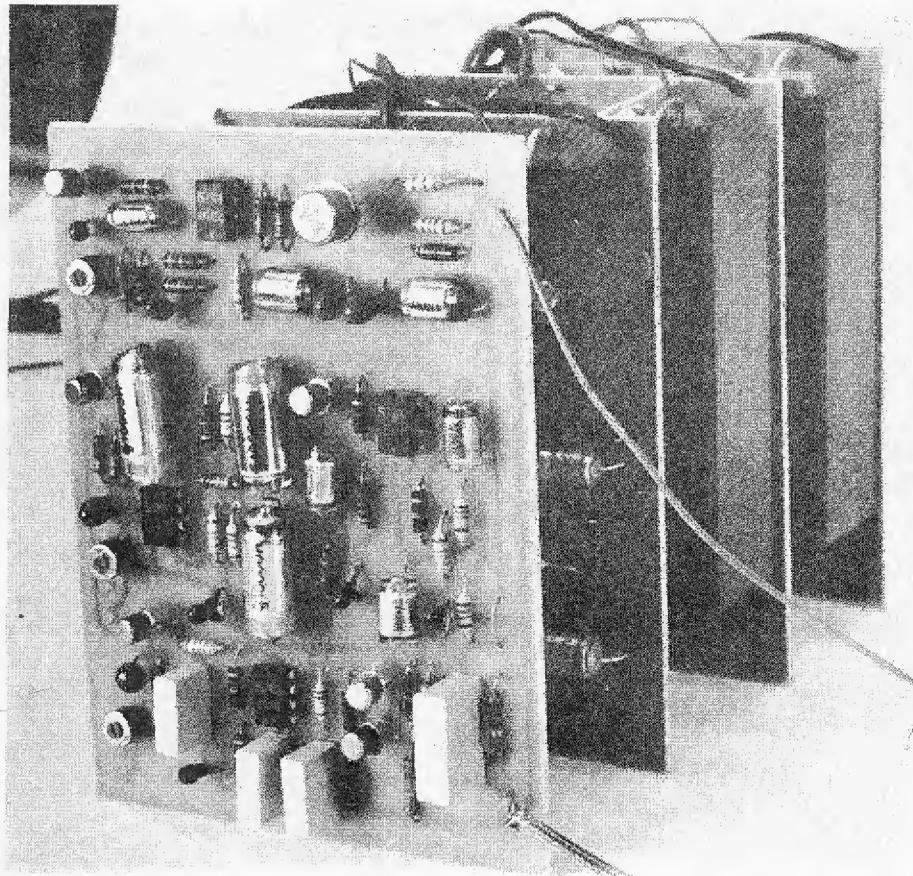
Fig. 7 gives the axial response in non-reverberant surroundings of the smaller mixed system, i.e. of the single B139 bass unit together with the B110/T27 combination in a 0.33-cu.ft enclosure. It shows that the bass response has been considerably improved, but there remains a large dip around 2kHz. Fortunately, since the response region requiring correction occupies a narrow range of frequencies mainly below the h.f. crossover point, the dip can largely be filled by creating a suitable peak in advance of the upper band-pass filter roll-off. This is done by adjustment of feedback over the first of the 741P amplifiers (IC_2 in the low-pass section of the band-pass filter). Fig. 8, which shows the roll-off characteristic for a 741P gain setting of 2.74, achieved with $R_1 = 47k\Omega$ and $R_2 = 27k\Omega$ (gain = $1 + R_1/R_2$). (Peak height can be changed by altering gain slightly.)

With this modification to the band-pass filter circuit the overall response appears double-humped as illustrated in Fig. 6 by taking the upper dotted curve. The resulting acoustic response of the system is then as shown in Fig. 7 with the broken dip-corrected curve replacing the full-line one between 1.2kHz and 3.8kHz.

Miniature full stereo arrangement

A simpler stereo system was set up by using only the B110 and T27 units in the 0.33-cu.ft boxes, i.e. without the large combined bass enclosure. This economy also dispenses with about half of the filter circuitry; the parts now required are just the low-pass sections of the original band-pass filters and the two unmodified high-pass sections for the T27 tweeters. Further equalization must be added to give the B110 units extra l.f. response and thereby compensate for the absence of the B139 speaker.

In Fig. 6, the dotted line from 230Hz to 20Hz (curve A) shows the output from Tr_4 in Fig. 2. Because the IC_3/Tr_5 circuit is not included in this reduced-scale system, it is therefore the low-pass response applicable to signals feeding the B110 units. The high-pass filters and output characteristics remain as before.



Complete amplifier for one channel. Front board is modified active filter, including l.f. section and voltage regulator, not needed for the other channel. Note reduced-size heat dissipator for h.f. power amplifier.

The axial response of one B110/T27 assembly (0.33-cu.ft box) as used for this full stereo arrangement is given in Fig. 7 by taking the full-line curve plus dip correction above 200Hz and adding to it broken curve A below this frequency. Evidently, bass is still lacking below 100Hz. Curve B gives a possible 4dB maximum improvement here, and corresponds with the curve B filter response illustrated in Fig. 6. For such additional l.f. output, C_1 in the correction network (Fig. 2 or Fig. 5a) becomes 0.1 μ F and R_6 is increased to 390k Ω .

More bass boost than is shown by curve B in Fig. 7 could be obtained by other changes to the R-C network values, but, whenever extra l.f. response is sought from the B110 in this way, there is a danger that excursion of the cone driving-coil might become too great; limiting and consequent distortion is then inevitable. For this reason such increased bass output must be accompanied by a restriction in overall listening level, especially for music with a heavy bass component.

Note that both the A and B filter curves give steady l.f. increase down to about 80Hz after which the response gradually levels off and the B110 characteristic takes over. This is a deliberate compromise. It not only helps to overcome the problem of excessive cone movement but also gives some protection against spurious low-frequency signals such as those caused by rumble.

Use of combined bass for quadraphony

A mono-bass arrangement using active-filter circuits to the design described in this and the earlier article has been applied to a quadraphonic installation. When considering the extension of a split-band, multiple-speaker stereo system to quadraphony, the prospect of doubling-up yet again on most of the equipment is a daunting one, and any possible saving in outlay—especially if this can be done without reduction in performance—is an

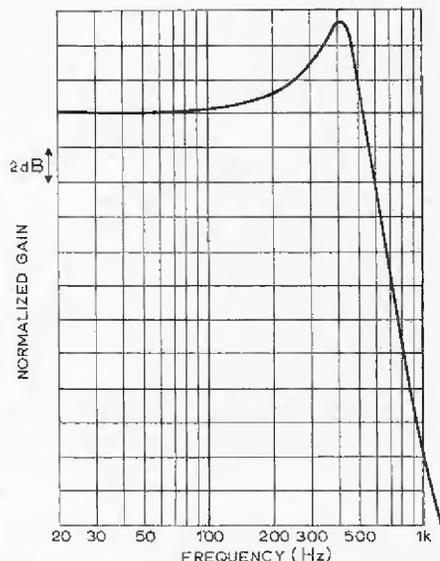


Fig. 8. Peak of appropriate height to compensate dip in Fig. 7 obtains for $R_2=27k\Omega$, with $R_1=47k\Omega$ in IC_2 circuit.

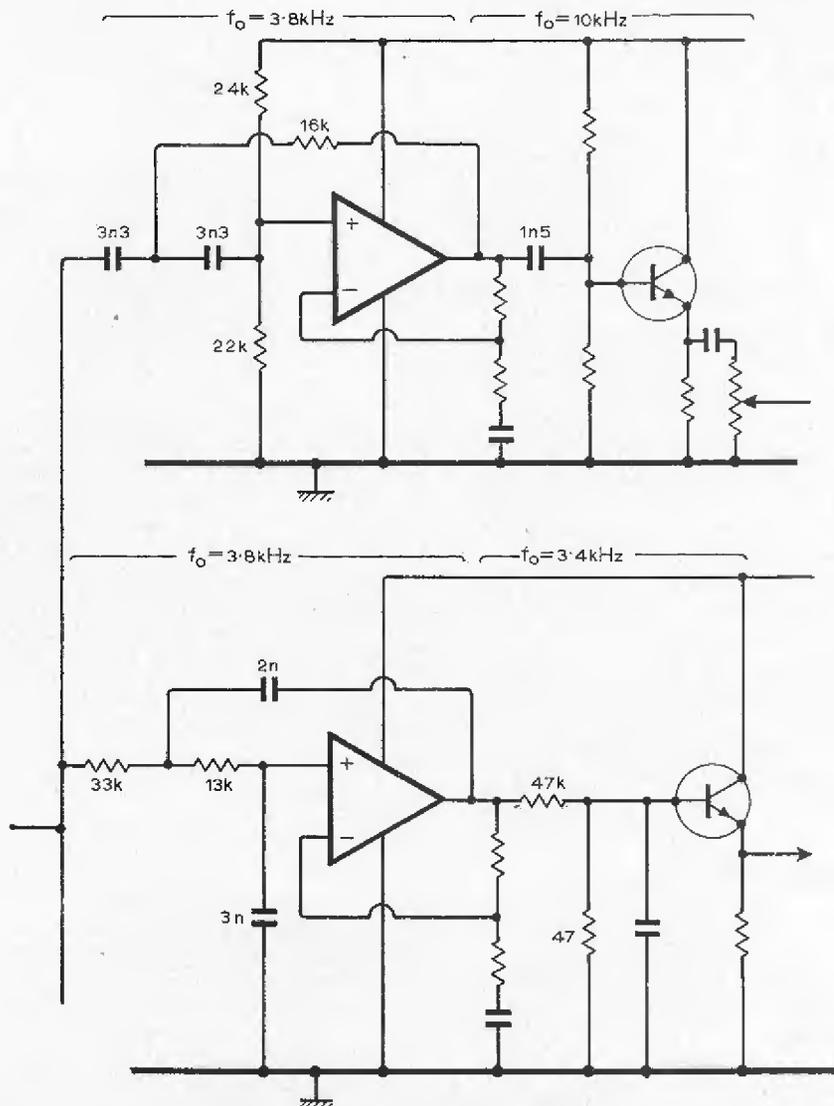


Fig. 9. Original active crossover design (December) can be improved by lowering upper crossover frequency to 4kHz. New component values shown above.

attractive proposition. The use of a single bass speaker offers just such an economy.

Obviously, twice the (stereo) complement of filters and amplifiers serving the mid and upper frequency ranges must be provided. For this quadraphonic application, however, the only change necessary in the basic circuit of Fig. 2 is that four mixing resistors, each of 200k Ω , are now needed to combine the l.f. signals at the input to IC_1 . Although this input point is not a virtual earth, there is no danger of crosstalk between the four contributing channels because their outputs are fed from constant-voltage sources as represented in Fig. 2 by the circuit containing Tr_1 and Tr_2 . Low output-impedance amplifiers in this position were, of course, already a necessity to prevent interaction between the stereo channels and, equally important, between the filters.

Although the foregoing outline suggests that bass combination is as easy to achieve for quadraphony as for stereo, there is a difficulty which may arise concerning the matrix system used and its effect on channel combination at low frequencies. In particular, the SQ coding system, and more especially the better

QS system, which makes use of phase as well as amplitude relationships to improve separation, could create interaction problems at frequencies where the channels are combined. If this happened, it would be difficult to predict the subjective result; it would certainly depend on the programme content.*

Modification to original system

Readers who have either built or considered building the full 2 x 3-way system may be interested in a modification to the filter circuit given in Fig. 2 of the December 1973 article.

The change concerns the upper crossover frequency which has been lowered to 4kHz. In the original design, a 6kHz crossover was chosen so that a direct comparison of performance could be made with the Radford FN10 passive network. But as mentioned in passing (p.576 of that issue), the subjective effect was improved by shifting the crossover down by 2kHz; the reason for this improvement

*In mono reproduction of SQ and QS records, sounds intended for the centre back are heavily attenuated, and with QS records, left back and right back sounds would be 7.7dB down on the front sounds—Ed.

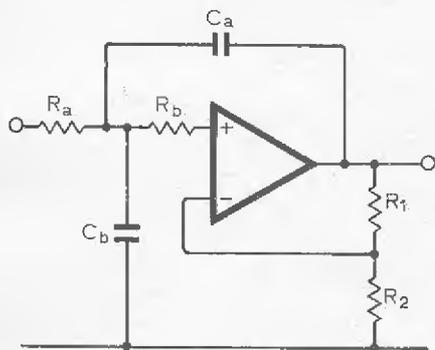
was not obvious at the time. However, subsequent discussion with the maker of the KEF B110 unit revealed that coupling between coil and cone becomes increasingly less stable at frequencies above about 3kHz (hence the response peak at 5kHz as shown in Fig. 1 of this article). By reducing the mid-range response in this region and making more use of the available lower-end output from the T27 tweeter, adverse effects from the B110 are avoided.

Fig. 10 shows the relevant parts of the circuit, taken from Fig. 3 of the December 1973 article, showing amended values.

Component layout. Photocopies of printed board positive transparencies, together with component layouts and a response curve for the modified crossover, are available from the *Wireless World* editorial office. Please send a stamped and addressed envelope for these.

Appendix

Circuit below applies to a low-pass filter where C_a , C_b and R_a , R_b are the component values to be determined.

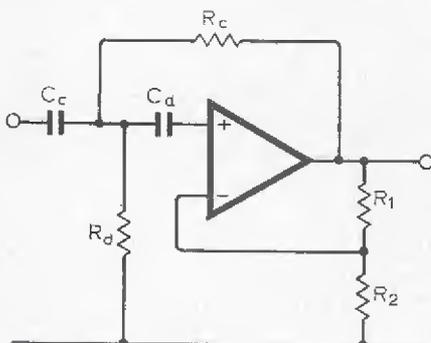


First, choose a suitable value for C_a (in μF) and thus find a value for a number, k , where $k=2\pi C_a f_o$. In this expression, f_o is the 3dB-down frequency which can be related to a wanted crossover frequency (normally at the 6dB-down point) by inspection of the appropriate curve.

Next, calculate

$$m = \frac{\alpha^2}{4} + (G-1), \text{ where } \alpha = \sqrt{2}$$

maximally flat response, overall gain $G=1+R_1/R_2=2$ (nominal value). Hence,



$$\left. \begin{aligned} C_b &= m \cdot C_a \\ R_a &= \frac{2}{\alpha k} \\ R_b &= \frac{\alpha}{2mk} \end{aligned} \right\} \text{giving the remaining values.}$$

In the high-pass case C_c (μF) is chosen to find k from f_o as before. Assuming $\alpha = \sqrt{2}$ and $G=2$
 $C_d = C_c$ (as chosen)

$$R_c = \frac{\alpha + \sqrt{\alpha^2 + 8(G-1)}}{4k}$$

$$R_d = \frac{4}{\alpha + \sqrt{\alpha^2 + 8(G-1)}} \cdot \frac{1}{k}$$

Having thus established the basic curve shapes, further adjustments will have to be made to achieve the optimum performance. For example, Fig. 1 shows that the T27 tweeter output tends to fall at the higher frequencies and this tendency is counteracted by moving f_o for the 6dB/octave section of the high-pass filter to about 10kHz. This 'softens' the roll-off shape and also gives the wanted rising characteristic.

A similar technique was used to compensate for unit deficiencies at the l.f. end of the band. Other forms of response adjustment, such as roll-off "corner sharpening" and in-band dip correction are achieved as already described by altering the appropriate op-amp gains.

The above points are mentioned as a guide to the various ways in which active filters can be tailored to fit particular requirements. But, obviously, there are as many combinations and permutations of adjustment method as there are conditions to satisfy; it is largely a matter of trial and error which form is chosen to obtain the best result.

Letters to the Editor

ACTIVE CROSSOVER NETWORKS

As a designer of p.a. systems using active crossover networks, I was very interested to read the article by D. C. Read in your November, 1974, issue. However, I feel that readers may be confused by the graphs for the B110 unit in Fig. 1, and in particular by the conclusions drawn from them.

The graphs are obviously drawn for the free field responses, and predominantly show the effect not of internal cabinet volume, but of external cabinet dimensions. In fact the sound power output response for this unit is *improved* by the smaller cabinet, the 0.33cu. ft enclosure being near the optimum for the B110, the enclosure Q being just under 1 at around 75Hz, giving a response flat down to around 80Hz and dropping at 12dB/octave below this. Response curves for small speakers such as this should always

be taken under the same conditions as their normal use, in this case almost certainly up against a wall, corresponding roughly to half-space conditions. Under these conditions the B110 will show a flat response up to around 600Hz, rising after this up to about 2kHz, where it levels out again. This unit should not be used on bass signals in as large a cabinet as 1cu. ft, as there will be almost no stiffness control over the low-frequency cone excursions.

A worthwhile improvement with these small enclosures is to provide, electronically, deliberate cross-talk between stereo channels below about 150Hz, thus cutting down considerably on cone excursions with bass signals predominantly in one channel or the other, unfortunately quite common with modern recording techniques. To the best of my knowledge, the first company to do this commercially was Servo-Sound.
 K. C. Gale,
 Poole,
 Dorset.

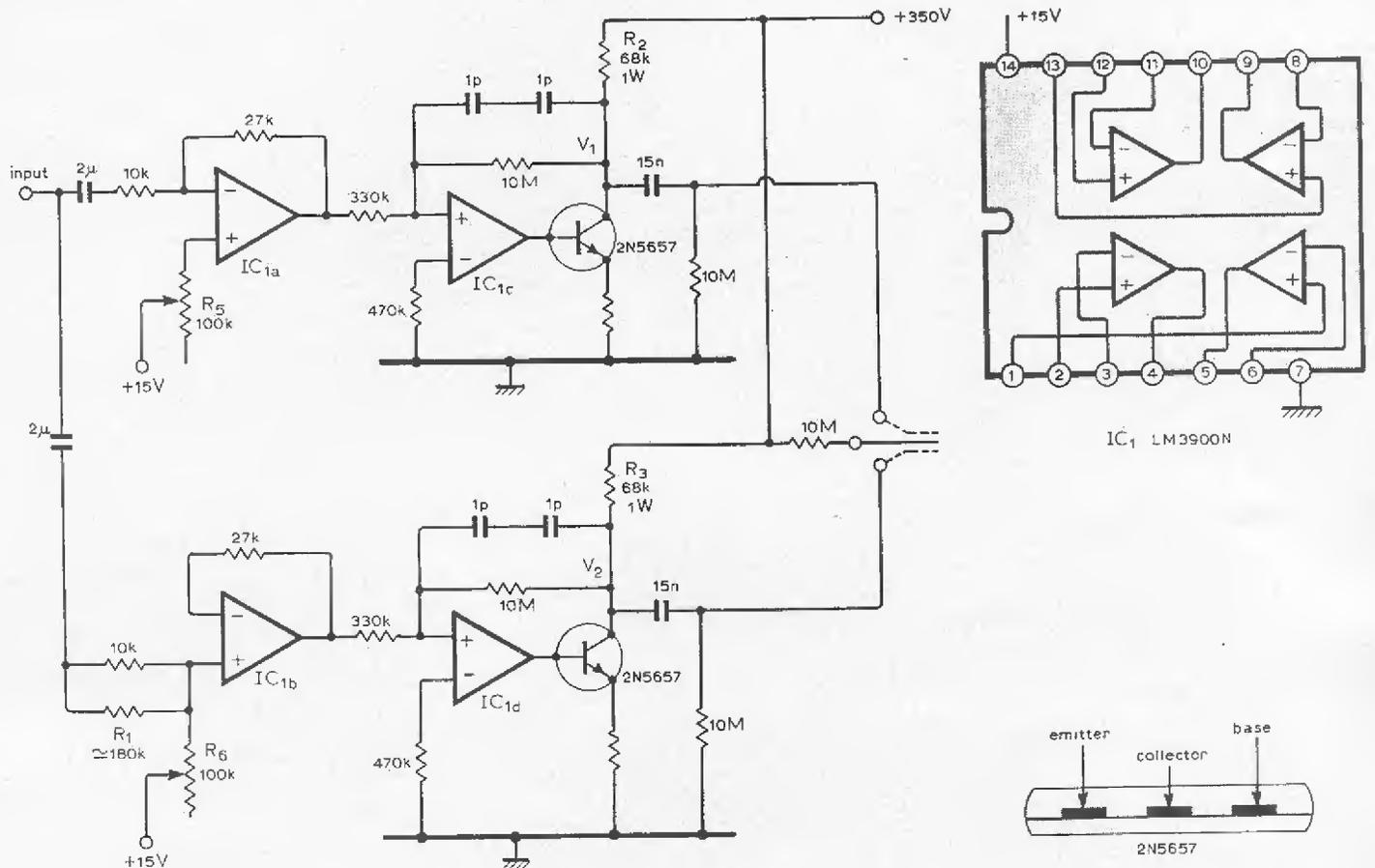
Electrostatic headphone amplifier

This circuit has been used successfully with a pair of headphones based on the W.W. design Dec. 1968. The amplifier can be driven from the headphone output of most power amplifiers. Po-

tentiometers R_5 and R_6 are used to set V_1 and V_2 at half the supply voltage. Resistor R_1 is required to compensate for the small signal resistance of a diode in the non-inverting input of IC_{1b} . If headphones of greater capacitance than 150pF are used it is necessary to reduce R_2 and R_3 to maintain the power bandwidth. It may then be necessary to heat sink the power transistors. The +15V bias supply for IC_{1a} and IC_{1b} must

be well filtered. The amplifier has a small signal frequency response of (-3dB) 10Hz to 40kHz, a power bandwidth of 10Hz to 15kHz, and a total harmonic distortion at 1kHz (almost entirely second harmonic) of 0.1% at 50V pk-pk and 1.0% at 300V pk-pk output.

N. Pollock,
Sandringham,
Australia.

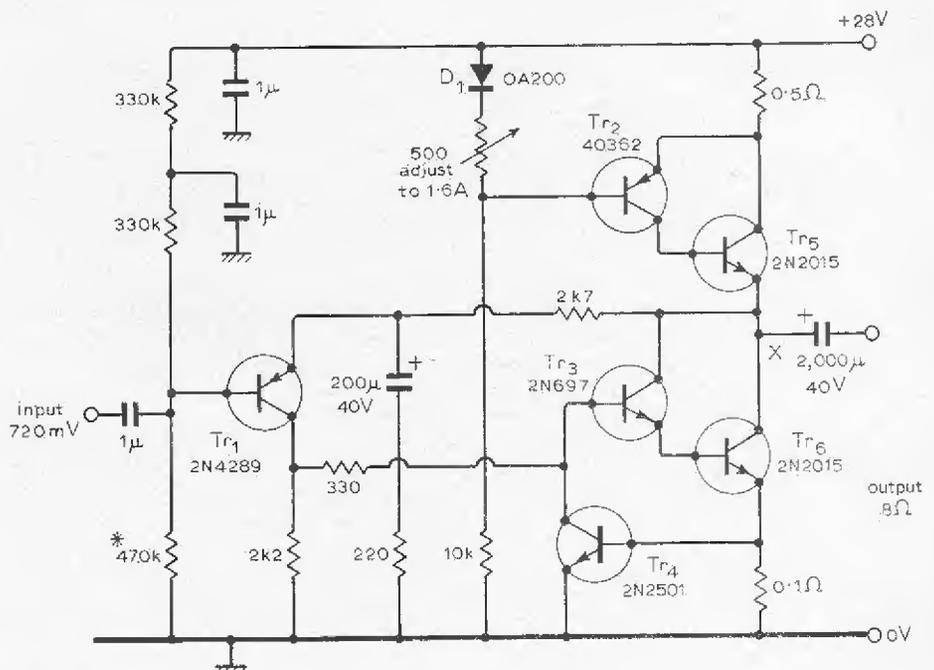


Class A power amplifier

This circuit was designed using easily available components. Its operating conditions are unconditionally stable, unaffected by individual transistor parameters and the transistor types are not critical. Short-circuit protection is provided by the constant-current source D_1 , Tr_2 , Tr_5 and Tr_4 . Transistor Tr_4 may be omitted if short-circuit protection is not required. The amplifier may be altered for a different load-impedance (R) or power output (W) by putting the output-stage current $I = 2W/R$, and supply voltage equal to $2 + 0.61 + 2\sqrt{2WR}$.

Dissipation in the output transistors is roughly $W \times 4$ watts, so the heat sinks need to be adequate. The amplifier shown will deliver 10W into an eight-ohm load. Diode D_1 and Tr_2 should be in thermal contact.

A. H. Calvert, London NW1.



* adjust to make point X half the supply voltage (if necessary)

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 $C = 33\mu\text{F}$, $r = 75\Omega$, $R = 30\text{k}\Omega$. 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

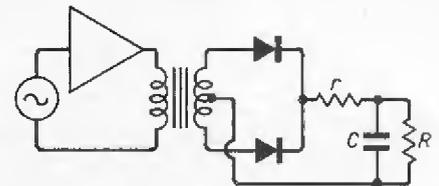


Fig. 1. Conventional PPM using centre-tapped transformer and full-wave rectifier.

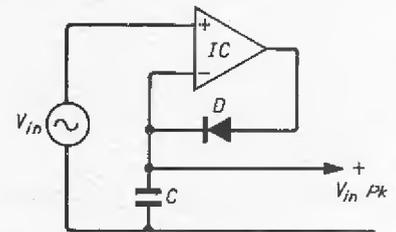


Fig. 2. Peak detecting circuit.

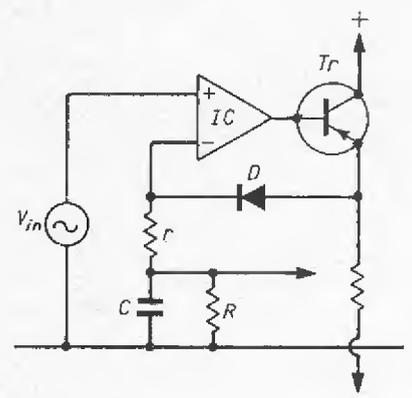


Fig. 3. Peak detecting circuit with time constants added.

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.

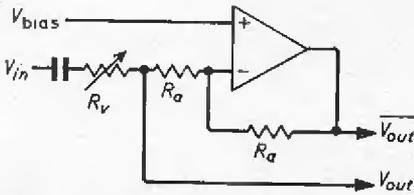


Fig. 4. Unity gain inverting circuit.

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.c.s, 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 \bar{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

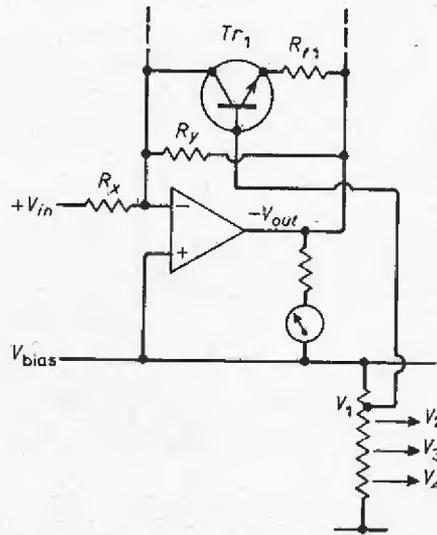


Fig. 5. Use of a transistor in the feedback path to provide law correction of transfer characteristic.

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.

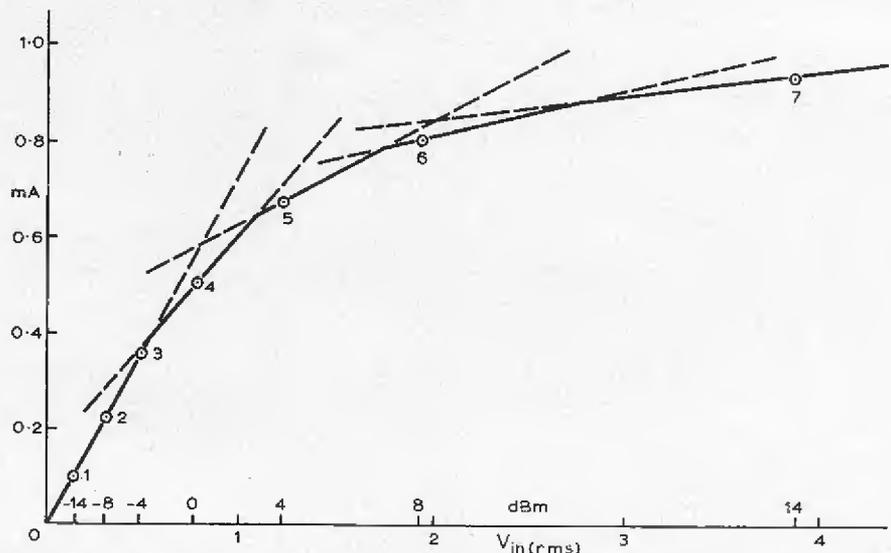


Fig. 6. Low corrected transfer curve approximated by a five section linear characteristic.

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 $600\Omega \pm 5\%$ so that with its series resistor of 4.7k Ω (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

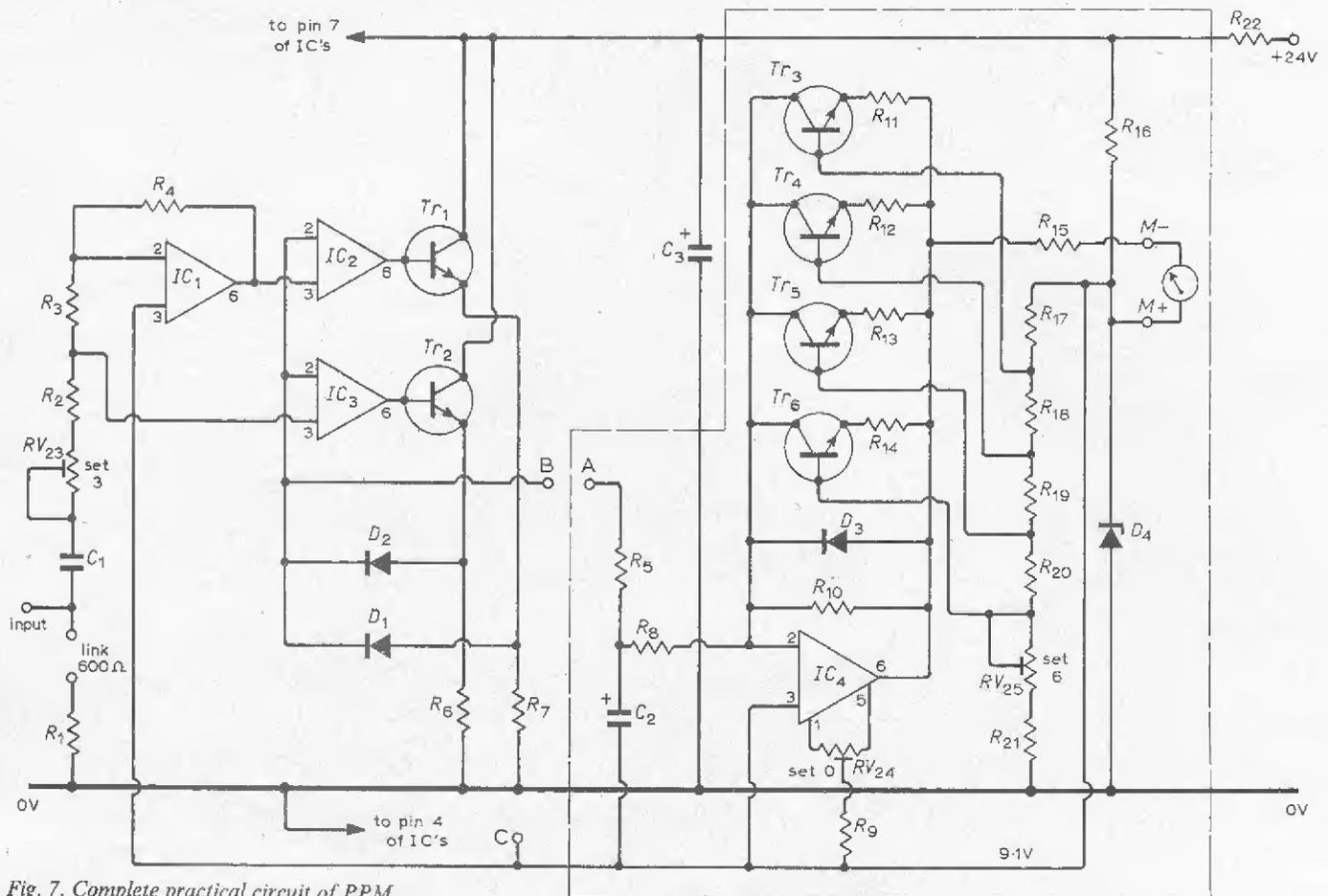


Fig. 7. Complete practical circuit of PPM.

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 $16k\Omega$.

Setting-up and performance

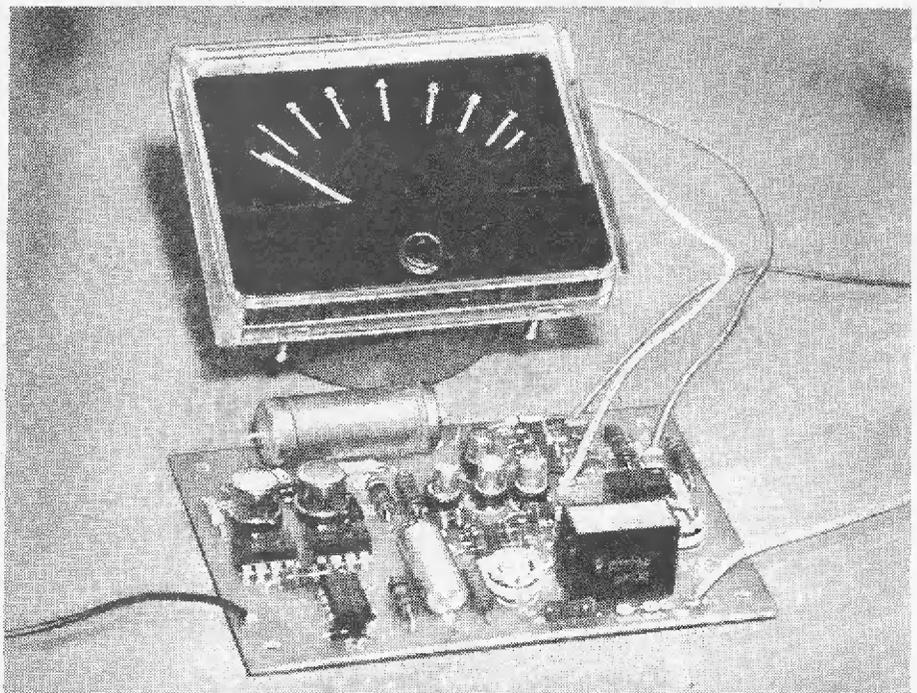
The procedure for setting-up the PPM is a simple one. First, with zero input voltage, the zero is set ('Set 0' control) RV_{24} . Next a level corresponding to $-4dBm$ (reading 3 on the PPM scale) or $490mV$ r.m.s. sine wave, $690mV$ peak, is applied and the 'Set 3' control (RV_{23}) is adjusted to bring 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 $30^\circ C$ rise in temperature (from $17^\circ C$) gave only about $10\mu A$ 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



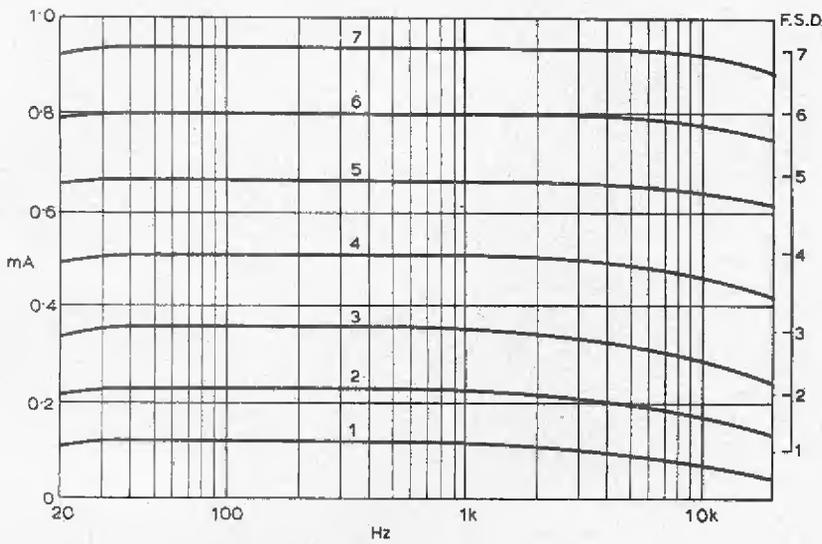


Fig. 8. Frequency response of prototype PPM.

Fig. 9. Configuration of circuit using alternative 748C op-amps.

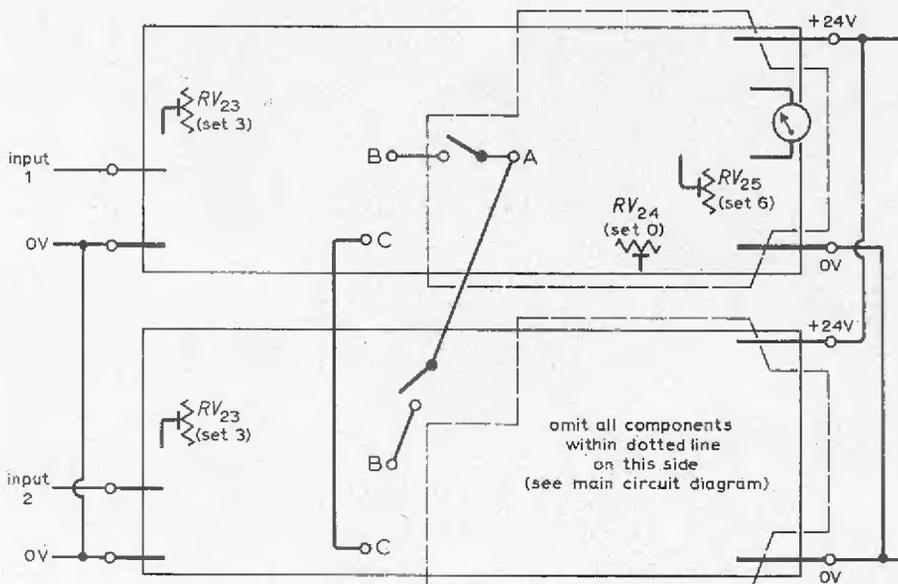
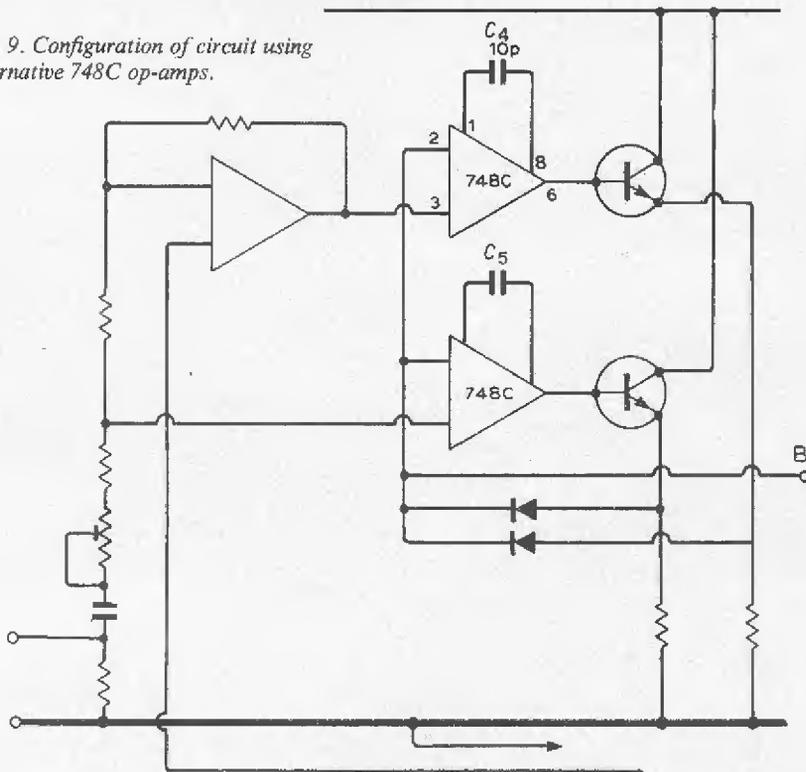


Fig. 10. Outline of printed circuit panel for mono or stereo.

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 $+8\text{dBm}$.

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 Depart-

ment 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 MRS or Electrofil 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 $\frac{1}{4}W$ 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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LM340-5 1.35	
LM340-12 1.35	
LM340-15 1.35	
LM340-18 1.35	

TRANSISTORS

AC126	0.15	BC181	0.38	BD131	0.36	BFW60	0.36	CG72	0.22	2N3705	0.10*
AC127	0.15	BC168	0.09*	BD132	0.40	BFX29	0.22	CG84	0.40	2N3706	0.10*
AC128	0.15	BC169	0.12*	BD135	0.36*	BFX30	0.30	CG138	1.30	2N3707	0.10*
AC128K	0.25	BC169C	0.14*	BD136	0.38*	BFX84	0.23	CG140	1.30	2N3708	0.09*
AC141	0.22	BC170B	0.12*	BD137	0.40*	BFX85	0.25	CG170	0.23	2N3709	0.08*
AC141K	0.35	BC170B	0.12*	BD138	0.48*	BFX86	0.25	TIP29A	0.64	2N3710	0.10*
AC142	0.18	BC172B	0.12*	BD139	0.58*	BFX87	0.20	TIP30A	0.52*	2N3711	0.10*
AC142K	0.32	BC182	0.11*	BD144	2.20	BFX88	0.20	TIP31A	0.54	2N3715	1.70
AC176	0.16	BC182L	0.12*	BD157	0.60	BFX89	0.90	TIP32A	0.64	2N3716	1.80
AC176K	0.32	BC183	0.10*	BD181	0.86	BFY11	1.10	TIP41A	0.58	2N3771	1.80
AC187	0.18	BC183L	0.10*	BD182	0.92	BFY18	0.72	TIP42A	0.72	2N3772	1.90
AC187K	0.35	BC184	0.11*	BD183	0.97	BFY20	0.60	2N404	0.40	2N3773	2.10
AC188	0.18	BC184L	0.12*	BD184	1.20	BFY41	0.60	2N896	0.20	2N3819	0.28*
AC188K	0.32	BC186	0.20*	BD232	0.80	BFY50	0.20	2N897	0.20	2N3904	0.16*
AD149	0.45	BC187	0.24	BD233	0.48	BFY51	0.18	2N706	0.15	2N3906	0.11*
AD181	0.45	BC207B	0.10*	BD237	0.60	BFY52	0.18	2N718	0.22	2N4123	0.14*
AD182	0.35	BC212	0.11*	BD238	0.60	BFY53	0.25	2N929	0.16	2N4124	0.14*
AF114	0.20	BC212L	0.12*	BD410	0.60	BFY64	0.35	2N1131	0.15	2N4290	0.14*
AF115	0.20	BC213	0.12*	BDX32	2.30	BFY90	0.90	2N1132	0.16	2N4291	0.14*
AF116	0.20	BC213L	0.14*	BDY10	1.50	BLY15A	0.20	2N1302	0.40	2N4292	0.14*
AF117	0.20	BC214	0.14*	BDY11	2.00	BSX19	0.18	2N1303	0.40	2N4347	1.10
AF118	0.50	BC214L	0.15*	BDY20	0.80	BSX20	0.18	2N1304	0.45	2N4348	1.20
AF124	0.25	BC237	0.16*	BDY38	0.60	BSX21	0.20	2N1305	0.45	2N4870	0.35*
AF125	0.25	BC238	0.16*	BDY60	1.70	BSX76	0.30	2N1306	0.50	2N4871	0.35*
AF126	0.25	BC300	0.34	BDY61	1.65	BSX77	0.30	2N1307	0.50	2N4919	0.60*
AF139	0.35	BC301	0.32	BDY62	1.15	BSX78	0.35	2N1308	0.60	2N4919	0.70*
AF239	0.37	BC302	0.40	BDY96	2.14	BSY52	0.28	2N1309	0.60	2N4920	0.50*
AL102	1.45	BC303	0.46	BDY98	4.96	BSY53	0.28	2N1310	0.60	2N4922	0.58*
AL103	1.30	BC307	0.15*	BDY97	2.45	BSY54	0.33	2N1202	0.44	2N4923	0.46*
AL107	3.30*	BC308	0.16*	BF127	0.90	BSY55	0.74	2N2217	0.30		
AL110	1.75*	BC309	0.18*	BF157	0.80	BSY65	0.30	2N2369	0.14		
AL113	1.80*	BC310	0.20*	BF177	0.25	BSY76	0.20	2N2369A	0.14		
BC107	0.12	BC317	0.12*	BF178	0.28	BSY78	0.75	2N2483	0.20		
BC107B	0.12	BC319	0.13*	BF178	0.30	BSY84	0.16	2N2484	0.16		
BC108	0.12	BC320	0.18*	BF180	0.30	BU105	1.80*	2N2646	0.50		
BC108B	0.12	BC321	0.18*	BF181	0.30	BU105/02	1.90*	2N2711	0.20		
BC109	0.12	BC323	1.60*	BF182	0.30	BU108	3.00*	2N2712	0.15		
BC109B	0.12	BC327	0.18*	BF185	0.30	BU109	2.50*	2N2804A	0.20		
BC109C	0.15	BC328	0.16*	BF184	0.20	BU125	1.80*	2N2905	0.18		
BC117	0.15*	BC337	0.17*	BF185	0.20	BU133	1.60*	2N2905A	0.22		
BC119	0.25*	BC336	0.17*	BF194	0.10*	BU204	1.60*	2N2906	0.18		
BC125	0.18*	BC407	0.22*	BF196	0.12*	BU205	1.90*	2N2925	0.14*		
BC126	0.20*	BC408	0.22*	BF197	0.12*	BU208	2.40*	2N2926	0.09*		
BC140	0.25	BCY30	0.55	BF200	0.40	BU208	2.80*	2N2928R	0.10*		
BC141	0.28	BCY31	0.55	BF206	0.30	MJ480	0.80	2N2929	0.09*		
BC142	0.23	BCY32	0.60	BF218	0.30	MJ481	1.05	2N2926G	0.10*		
BC143	0.23	BCY33	0.55	BF220	0.28	MJ490	0.90	2N3053	0.50		
BC144	0.23	BCY34	0.55	BF224J	0.18*	MJ491	1.15	2N3055	0.20		
BC147	0.09*	BCY38	0.50	BF244	0.17*	MJ5240	0.40*	2N3133	0.30		
BC148	0.09*	BCY39	0.50	BF257	0.30	MJ5250	0.45	2N3134	0.20		
BC149	0.09*	BCY40	0.75	BF258	0.35	MJ5251	0.55	2N3137	0.10		
BC152	0.25*	BCY42	0.30	BF259	0.48	OC43	0.95	2N3440	1.55		
BC153	0.18*	BCY54	1.60	BF336	0.35*	OC44	0.32	2N3442	1.20		
BC157	0.09*	BCY70	0.12	BF337	0.32*	OC45	0.32	2N3870	3.60		
BC158	0.09*	BCY71	0.18	BF338	0.45*	OC46	0.20	2N3702	0.10*		
BC159	0.12	BCY72	0.12	BFW30	1.25	OC70	0.30	2N3703	0.10*		
BC160	0.32	BD115	0.15	BFW69	0.30	OC71	0.35	2N3704	0.10*		

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	(105)	(200)	(400)	(600)	(1000)	(1500)	(2000)
200	0.35	0.50	0.45	0.40	0.58	0.60	0.66
400	0.40	0.60	0.50	0.45	0.68	0.88	0.98
600	0.65	0.85	0.70	—	1.09	1.19	1.26
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TRIACS - Plastic TO-220 Package Isolated Tab

RV	4A	6.5A	8.5A	10A	15A
	(a)	(b)	(c)	(d)	(e)
100V	0.60	0.60			

Horn loudspeaker design

An article summarizing the development of design theories and concluded with two systems for construction

by J. Dinsdale, M.A., M.Sc.
Cranfield Unit for Precision Engineering

After a period in the infancy of the gramophone when it was universally employed, the horn loudspeaker has fallen from popularity, due probably to its relatively large size, complexity of manufacture and hence high cost. Although full-range horn systems are used today only by a small number of enthusiasts, most experts are unanimous in acclaiming their virtues as loudspeaker enclosures, especially their high degree of realism and "presence". These articles examine briefly the history of the exponential horn loudspeaker and discuss the theory of horn-loading and the technical requirements of a good design. Comprehensive data are included for a wide range of horns, together with outline designs for a large and a small horn, suitable for domestic use.

The ideal exponential horn consists of a straight circular tube whose cross-sectional area increases logarithmically along its length from a small throat (at which is mounted the loudspeaker) to a large mouth. Extreme bass notes demand a mouth of very large area (20 to 30 sq. ft) and a horn at least 20ft in length, whereas extreme treble notes require a horn with dimensions of only a few inches. For this reason most wide-range horn systems will incorporate a number of separate loudspeakers, each with its individual horn of appropriate length and mouth area. To accommodate these horn combinations within a cabinet of reasonable size, the bass and middle horns are generally of square cross-section and are "folded" into a complicated pattern. Unfortunately, the inevitable restrictions and compromises introduced by these departures from a straight axis and circular section can cause serious variations in the frequency response, and much of the art of horn design is concerned with achieving a product of reasonable overall size and cost, without sacrificing any of the astonishing realism which is obtainable from the ideal horn.

The efficiency of a horn system will be typically between 30 and 50%, a figure to be compared with 2 or 3% for a bass-reflex enclosure and less than 1% for a totally-enclosed box.

The principal reasons for the evident lack of popularity of the horn probably lie in its dimensions and cost. The overall size of a bass horn, even when folded into a cabinet of reasonable shape, will be larger than a bass-reflex or infinite baffle enclosure of

comparable specification. But although one reads occasionally of straight horns up to 20ft long, excellent results may be obtained from horns of more moderate dimensions; for example a complete horn system may be folded into an attractive cabinet of volume only 6 cu. ft, a not unreasonable size for domestic listening. The cost of horn enclosures is often considered to be prohibitive, and it is true that there is considerably more work in constructing a folded horn than in other enclosures; furthermore, this is work best performed by craftsmen and not easily adapted to "production-line" methods. Nevertheless, the building of a folded horn is by no means outside the capability of a competent do-it-yourself enthusiast, and it is to these individuals that the practical designs will be directed.

Although the early acoustical gramophones or phonographs employed horns of one type or another to couple the diaphragm to the listening room, and the early electrical reproducers of the 1920s and '30s also used horns, thereafter the horn suffered a setback from which it has never recovered. Certainly, a few companies market horn loudspeaker enclosures, and the occasional articles in the technical press^{1,2} stir up a passing interest, but unless one resorts to the masterly academic treatises by Olson³ or Beranek,⁴ or reverts to pre-1940 publications, there is very little information available for the enthusiast who wishes to both design and construct a horn. Recent experience gained by Telfer and others^{5,6} has reinforced the author's opinion that there are many audio enthusiasts who would be interested in constructing a horn enclosure.

After a brief historical survey, these articles examine the theory behind the horn-loaded loudspeaker enclosure and explain the basic points to consider when designing horns. The various compromises adopted by different workers are discussed, especially in the area of folding techniques, and the effects of these compromises on audio quality are studied. Finally, outline designs for two domestic horns are given: a "no-compromise" horn to suit the most fastidious (and enthusiastic) listener, and a "mini-horn" which provides a more limited performance for those with smaller living rooms (and bank balances), and which, while no more obtrusive than most commercial loudspeaker cabinets, will provide extremely clear and natural reproduction.

Background

It has been known for many thousands of years that when sound is passed through a tube with a small throat and a large mouth, it experiences an apparent amplification, and from Biblical times man has used rams' and similar naturally occurring horns both as musical instruments and as megaphones. Thomas Edison attached a tin horn to his primitive phonograph in 1877 to couple the minute vibrations of the diaphragm to the air load in the listening area, and to the majority, the term "gramophone horn" conjures up an image of the early gramophones or phonographs designed between about 1890 and 1912, all of which utilised an external horn.

A variety of expansion contours were employed for these early horns, mainly straight conical horns in the earliest machines, but the later gramophones of this period employed large flaring horns with either straight or curved axes depending on the overall length of the horn and the general design of the complete equipment. An analysis of these early horns, carried out in the light of modern acoustic knowledge, reveals a lack of understanding at that time of the operation of the horn as an acoustic transformer. This is surprising since Lord Rayleigh had analysed the "transmission of acoustic waves in pipes of varying cross-section" in Articles 265 and 280 of his classic treatise "Theory of Sound", published in 1878.⁷

Lord Rayleigh gave the analysis in Art.281 for the passage of sound through a conical pipe, and he also made the interesting statement that "when the section of a pipe is variable, the problem of the vibrations of air within it cannot be generally solved". For some years after publication, Lord Rayleigh's results were purely of academic interest, but more general interest was aroused about the turn of the century by the early gramophones, most of which used external conical horns, as in the early HMV "dog" models.

After 1912, a number of manufacturers introduced internal horns with a degree of folding to enable cabinets of reasonable size to be used, and these models held the consumer market during the following 12 years, on account of their compactness and suitability as pieces of furniture. (Even in those early days, the enthusiast must have had

problems in persuading his wife to provide house-room for a large unfolded external horn.)

In the early 1920s a number of designers carried out theoretical analyses based initially on the work of Lord Rayleigh, but extending the work to be more applicable to the full audio range at domestic listening levels. Among these early analyses must be mentioned the work in America by A. G. Webster⁸ in 1920, by C. R. Hanna and J. Slepian⁹ in 1924 and by P. B. Flanders¹⁰ in 1927. In Britain independent analyses were carried out by P. Wilson in 1926 writing in *The Gramophone* magazine and later with A. G. Webb in "Modern Gramophones and Electrical Reproducers", and also by P. G. A. H. Voigt¹² in 1927.

All of these analyses, except the last, were based on an exponential contour, and were derived from a statement in Art.265 of Rayleigh's treatise. Webster had worked out an approximate theory for other types of horn and had deduced that the exponential was the optimum contour. All these analyses made the assumptions that (a) the cross-section is circular, (b) the axis is straight, and (c) all wavefronts are plane.

However, while it may be reasonable to assume plane wavefronts at the throat of the horn, it is clear that the wavefront at the mouth will be curved (as if a balloon were emerging from the horn, being inflated at the same time). Wilson, who had independently derived the analysis of the exponential horn in 1926 working from Rayleigh's treatise, later published a modified form on the assumption that the wavefront would assume a spherical shape always cutting the

contour of the horn and its axis at right angles.

This assumption, that the curvature of the wavefront would gradually increase from zero (the initial flat wavefront at the throat), satisfies also the condition specified by Hanna and Slepian and later by I. B. Crandall¹³ that the wavefront as it emerges from the open end will be equivalent to that provided by a spherical surface, as opposed to that produced by a flat piston. Voigt, however, had commenced his analysis on the assumption that wavefronts within the horn will be spherical and of the same radius throughout their progression through the horn. This assumption leads to a tractrix curve for the horn contour, and both theoretical considerations and very careful listening tests by the author and others tend to support the claims of the tractrix as the optimum horn contour. The mathematical basis of the exponential and tractrix curves is discussed in a later section of this article.

During the 1920s, 30s and 40s a large number of experimenters investigated methods of folding horns into small enclosures for domestic gramophone reproducers, and the records of the Patents Office bear witness to the ingenuity of man at overcoming conflicting conditions in the search for perfect sound reproduction. These designs for folded horns enjoyed a greater or lesser degree of success according to a number of factors including the performance of the loudspeaker motor. Nevertheless, it must be repeated that they were almost invariably of square or rectangular cross-section, and the axis was no longer straight and thus any resemblance between their actual perform-

ance and theoretical considerations was to some extent coincidental.

The advent of the moving coil loudspeaker in 1927 and electrical amplification stimulated further advances in the design of horns, which, because they now no longer had to be connected to the acoustical tone-arm, were freed of many of the earlier constraints. Many loudspeaker motor units were designed specifically for horn loading, and it was not until World War II that interest in the horn lapsed in favour of the bass reflex, infinite baffle and other types of loading systems which, although they had the peripheral advantages of smaller physical size, greater ease of design and manufacture and hence lower cost, were decidedly inferior in terms of musical realism.

During this time the designs of Voigt in Britain and of Klipsch¹⁴⁻¹⁸ in America continued to attract considerable support, especially the ingenious method evolved by the latter in adapting a doubly-bifurcated bass horn design to utilize the acoustic advantages inherent in corner positioning, a design which has now become a classic. Others at this time were experimenting with horn-loaded loudspeakers, notably J. Enoch and N. Mordaunt (whose design was subsequently incorporated in the Tannoy "Autograph" and "GRF" enclosures). Lowther (using a modern version of Voigt's high-flux motor unit) and J. Rogers (whose horn-loaded mid-frequency ribbon is still regarded by many as the ultimate in sound reproduction in this range) and one must not overlook the contributions of H. J. Crabbe¹⁹ and R. Baldock²⁰ in more recent times.

However, it must be emphasised that the multiple reflections, absorptions, resonances and changes of direction inherent in folded horns, together with the uncertainty of function of non-circular sections must inevitably alter the performance of such horns from that of the straight, circular-section horn on which the design may have been based.

Recent years have seen a minor resurgence in the popularity of the horn, caused perhaps by the search for "perfect sound reproduction", and there are many who hope that this trend will continue.

A very readable account of the early history of the horn loudspeaker has been given recently by P. and G. L. Wilson.²¹

General theoretical principles

The following section deals principally with the exponential contour, which is the basic expansion curve used in most high quality horn loudspeakers, and the tractrix, which has a more complicated formula, but with a dominant exponential component—indeed the two curves are virtually identical from the throat to about midway down the horn.

Determination of flare contour

The theory of the conical horn was originally worked out by Lord Rayleigh, but the first serious attempts to establish a practical working formula for the exponential horn were not made until 1919 and the years following. The basic formulae for the transmission of sound waves through horns have been given in modern terms by V. Salmon²²

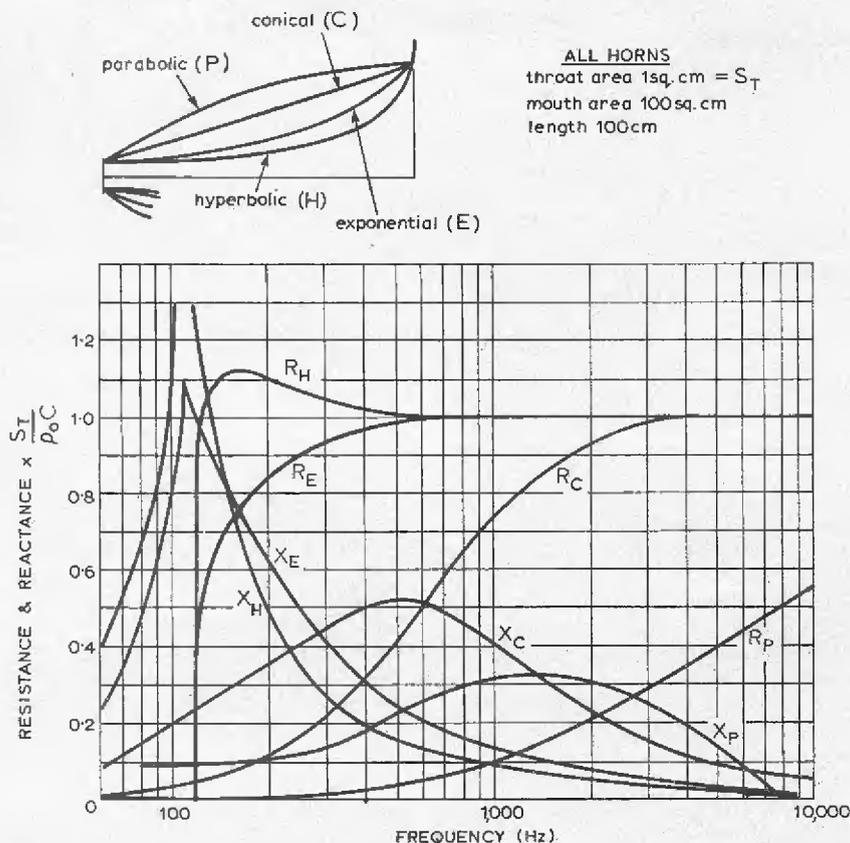


Fig. 1. Acoustical resistance and reactance against frequency at the throats of a series of infinite horns of different contour.

and others. Beranek⁴ has plotted the acoustical resistance and reactance against frequency at the throats of a series of infinite horns of different contour with identical cross-sectional areas at the throat and at a given point along the axis of the horn, and the resulting curves are shown in Fig. 1. For optimum loading of the loudspeaker motor, it may be shown that the impedance presented by the throat of the horn should be entirely resistive and of constant value throughout the working frequency range, i.e. the sound transmission should be of unity "power factor". Examination of the curves in Fig. 1 shows that the exponential and hyperbolic contours satisfy this condition most closely.

However, a further condition to be satisfied is that of minimum distortion at the throat of the horn, caused by "air overload". When a sound wave is propagated in air, a series of harmonics will be produced, thereby distorting the waveform. This occurs because if equal positive and negative changes in pressure are impressed upon a mass of air, the resulting changes in volume will not be equal; the volume change due to an increase in pressure is less than that due to an equal decrease in pressure. The rapid expansion and compression of air caused by the propagation of sound waves takes place adiabatically, i.e. there is no net transfer of heat, and the pressure and volume are related by the formula $pV^\gamma = \text{constant}$, where

- p = pressure
- V = volume
- γ = adiabatic gas constant (approx. 1.4 for air under normal room conditions)

This curve has been plotted in Fig. 2, together with a superimposed large sinusoidal change in pressure to illustrate the corresponding distorted change in volume.

If the horn were a long cylindrical pipe, distortion would increase the further the wave progressed towards the mouth. However, in the case of a flaring horn, the amplitude of the pressure wave decreases as the wave travels away from the throat, so for minimum distortion the horn should flare out rapidly to reduce the pressure amplitude as early as possible after the sound wave has left the throat. From this viewpoint it is apparent that the parabolic and conical contours will generate the least distortion due to air overload, and that distortion will be highest for the hyperbolic horn, because the sound wave must travel a further distance before the pressure reduces significantly.

Further inspection of Fig. 1 shows that the acoustical resistance of the hyperbolic horn lies within 10% of its limiting value over a larger part of its working frequency range than that of the exponential horn, and for that reason the hyperbolic horn provides rather better loading conditions to the loudspeaker motor. However, in view of the considerably higher air-overload distortion of the hyperbolic horn, the exponential or one of its derivatives is generally chosen as a satisfactory compromise between the hyperbolic and conical contours.

In cases where the advantages of a long

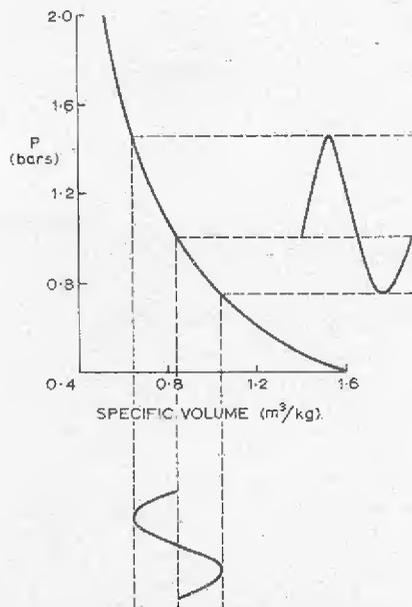


Fig. 2. Adiabatic pressure/volume relationship for air.

slow flare rate are required without the attendant high air-overload distortion, Olson³ has shown that a horn can be made up of a series of manifold exponential sections, commencing with a very short stub of high flare rate at the throat (to minimize distortion) which leads into a longer section of lower flare rate and thence to the main horn of very low flare rate. Klipsch has referred to this technique as the "rubber throat" in his paper on corner horn design.¹⁴ The mouth acoustical impedance of each exponential section is designed to match the throat impedance of the preceding section, right along the chain. Practically any acoustical impedance relationship with frequency may be obtained by this technique, but the procedure is complicated, and the additional effort cannot generally be justified for domestic horns.

Determination of mouth area

The acoustical resistance and reactance of the exponential horn have been plotted on a normalized scale in Fig. 3, which shows that

the acoustic impedance is entirely reactive below a frequency given by

$$f_c = \frac{mc}{4\pi}$$

where c = speed of sound; m = flare constant which appears in the basic exponential horn formula

$$S_x = S_T e^{mx}$$

where S_x is the area at distance x from throat; S_T is the area at the throat.

The frequency f_c , known as the cut-off frequency, is the lowest frequency at which the horn will transmit acoustical power, and thus the flare constant defines the lower frequency of transmission by a given horn. The flare constant may be calculated for any given cut-off frequency, and the horn profile may then be constructed. The above statement refers strictly only to horns of infinite length. In horns, as in cylindrical tubes, wavefronts of sounds whose wavelength is large compared with the mouth diameter tend to be reflected back into the horn where they interfere with successive wavefronts. Just as the loading of the loudspeaker motor by the throat of the horn must be largely resistive over the working frequency range for the smooth efficient transfer of acoustical energy, so must be the loading presented to the mouth of the horn by the surrounding air. Beranek has shown⁴ that for the radiation impedance of the mouth to be mainly resistive, the relationship $C/\lambda > 1$ must hold, where C is the circumference of the mouth of the horn and λ is the wavelength of the lowest note to be transmitted. If the mouth of the horn is not circular, it will behave in a similar way for equal mouth areas, i.e. if $C = 2\pi r_m > \lambda_c$ is the limiting condition

$$\text{and } S_m = \pi r_m^2 > \frac{\lambda_c^2 c}{4\pi} r_m > \frac{\lambda_c}{2\pi}$$

where λ_c = cut-off wavelength; r_m = mouth radius; S_m = mouth area.

Thus a horn of square section may be employed provided the mouth area exceeds $\frac{\lambda_c^2}{4\pi}$. Hanna and Slepian had examined from

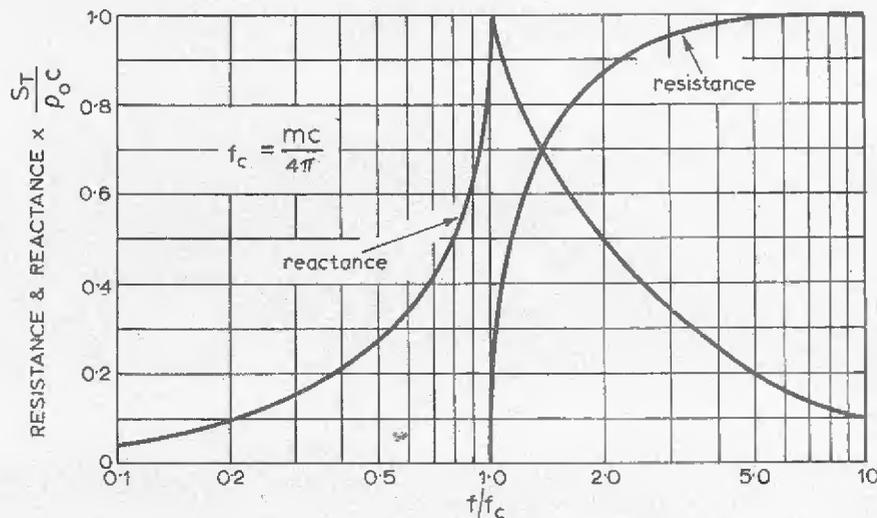


Fig. 3. Acoustical resistance and reactance of an exponential horn.

a different standpoint the behaviour of wavefronts at the mouth of the horn, and deduced that reflection was a minimum when the slope of the profile was 45° (i.e. included angle of 90°). This will be so where the mouth circumference equals the cut-off wavelength of the horn. It also illustrates the importance of distinguishing between the values of flare constant used for calculating exponential increase in area, and in plotting the profile of the actual horn. Fig. 4 (after Olson) illustrates the effect of foreshortening the horn to a length less than

the ideal. When the mouth circumference becomes less than the cut-off wavelength, reflections at the mouth cause objectionable peaks and troughs in the frequency response at frequencies near to cut-off, and if, in a given design, the mouth dimensions are restricted, it is generally preferable to increase the cut-off frequency to a value which allows the correct mouth area to be adopted, rather than to accept the uneven bass response illustrated in Fig. 4.

Plane and curved wavefronts

Hitherto, the assumption has been made that successive wavefronts remain plane throughout their propagation through the horn. However, along a straight circular section horn the wavefront must be normal to the axis, and also normal to the walls. (If the wavefront were either approaching or receding from the walls, energy would be either absorbed or supplied; alternatively, the composite wavefront resulting from the original wavefront and its reflection will itself be normal to the walls.) Thus wavefronts transmitted along a cylindrical tube will be plane, while wavefronts transmitted down a conical horn will be spherical. It is therefore clear that the wavefront emerging from an exponential horn will possess a degree of curvature, and that the conventional calculations made on the assumption of the exponential increase of plane wavefronts will be in error (in practice, the actual cut-off frequency will be somewhat altered from that derived theoretically, and the profile errors of the horn are not excessive).

The correct approach to the design of a horn in which the areas of successive wavefronts expand according to a true exponential law is not certain, since any horn profile chosen will *per se* determine the contour of the wavefronts within it, and in general this contour will be different to that originally assumed. Wilson¹¹ decided to assume spherical wavefronts of increasing curvature from zero (plane wavefronts) at the throat of the horn, and on this basis he cal-

culated a modified contour which lies just inside and very close to the true exponential. Fortunately, if a papier mâché horn is made on a solid former designed to a true exponential contour, the shrinkage of the papier mâché when drying converts the horn very closely to Wilson's modified form. Nevertheless, the prime assumption has been made that wavefronts are spherical and of changing curvature, and it is by no means certain that this is the case.

The tractrix contour

Voigt, in his 1927 patent, had proceeded on the more elementary assumption that the wavefronts within the horn must be spherical and of the same radius throughout their propagation through the horn. He based this assumption on the reasoning that if the curvature increases from plane waves (zero curvature) at the throat to a certain curvature at the mouth, then a point on the axis must travel at a faster rate than a point at the wall. Since the entire wavefront must travel at the speed of sound (assumed to be constant throughout the horn) the wavefront has no alternative but to be spherical and of constant radius. This requires that the horn contour should be the tractrix.

The tractrix is the involute of the catenary (the curve adopted by a uniform heavy chain suspended between two points at the same level) and is the curve traced out by a load being dragged along by a man moving in a straight line not passing through the load. It is not the "pure pursuit" curve traced by a missile which always travels towards an escaping target, as is often mistakenly supposed. The length *x* of a tractrix horn of mouth circumference λ_c , may be expressed as

$$x = \frac{\lambda}{2\pi} \log_e \frac{\frac{\lambda}{2\pi} + \sqrt{\left(\frac{\lambda}{2\pi}\right)^2 - y^2}}{y} - \sqrt{\left(\frac{\lambda}{2\pi}\right)^2 - y^2}$$

where *y* is the radius and λ the cut-off wavelength

cf. the equivalent exponential,

$$x = \frac{\lambda}{2\pi} \log_e \left(\frac{\lambda}{2\pi y} \right)$$

Both these curves are shown in Fig. 5.

It will be seen that the tractrix has a dominant exponential term which becomes less significant towards the mouth; in fact for the first 50% of their length the exponential and tractrix contours for a given cut-off frequency and throat area are virtually identical, but thereafter the tractrix flares at an increasingly greater rate until it attains its fully developed mouth at 180° included angle. In view of the complex nature of the formula, the best way to construct a tractrix is by graphical means, as shown in Fig. 6. The curve thus derived may be used to provide ordinates for the tractrix horn, after some smoothing of the slight discontinuities inherent in the graphical construction.

Whereas the tractrix terminates when the angle between the horn and the axis is 90° (180° included angle), the true exponential goes on to infinity in both directions. The

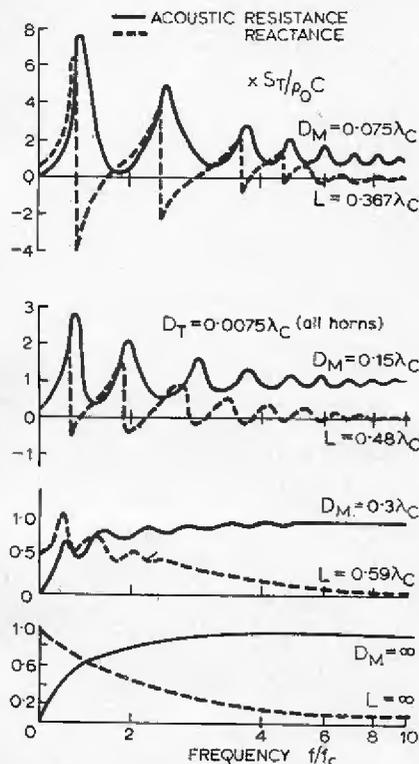


Fig. 4. Performance of foreshortened horns. Reflections at the mouth cause peaks and troughs in the frequency response near to cut-off.

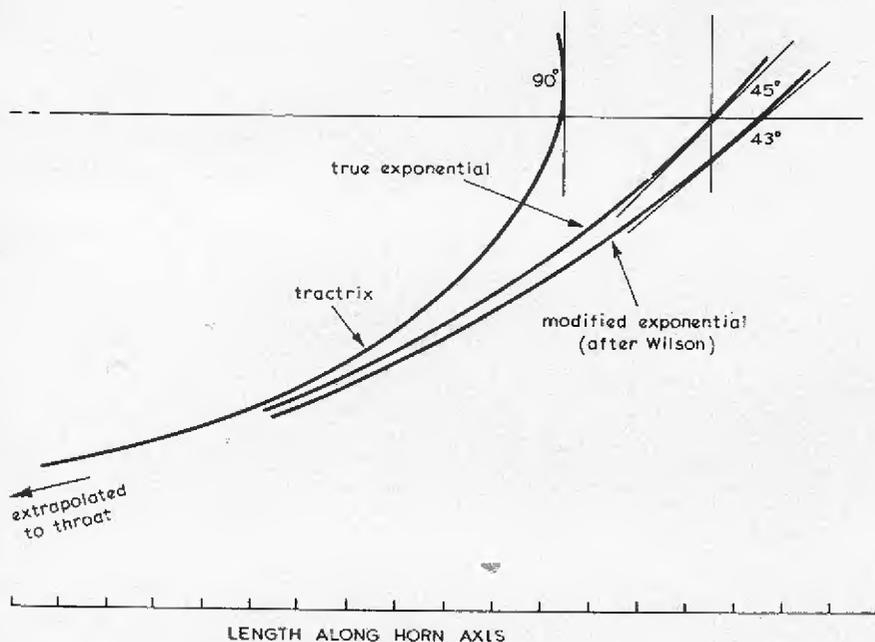


Fig. 5. Comparison of the exponential and tractrix contours.

tractrix horn for given throat and mouth dimensions is thus shorter than the equivalent exponential. It has been suggested that with the full tractrix terminating in a mouth of 180° included angle, the sound appears to originate from a point just inside the mouth, where the included angle is only 90° . There is thus some evidence that the tractrix may be terminated prematurely at this point, and if this is done, the mouth perimeter will be 90% of the wavelength at cut-off, as shown in Fig. 5, which compares the true and modified exponentials and the tractrix contours.

Efficiency

The efficiency of an exponential horn loudspeaker is determined by a large number of parameters, and a comprehensive treatment has been provided by Olson.³ Typical efficiencies of bass horns can be as high as 50%, while mid-frequency and treble horns can have efficiencies of over 10%, and these figures compare very favourably with bass-reflex enclosures (efficiency 2 to 5%) and infinite baffles (efficiency generally less than 1%). The extremely high efficiency of the horn is not necessarily of value in enabling amplifiers of lower output power to be used. Indeed, some class B output stages may produce a higher distortion level in horns because they need only be operated within the first 10% of their capability, at which low levels the effects of crossover distortion are more pronounced.

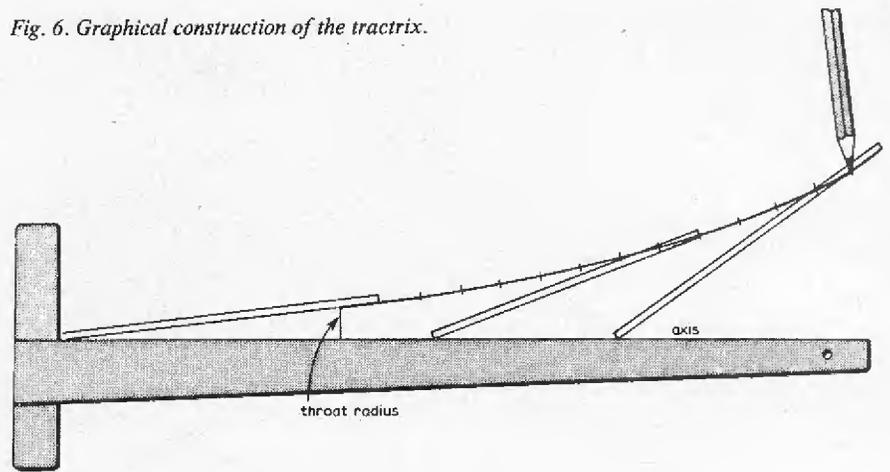
The principal advantage conferred by the horn's high efficiency is that for a given loudness the amplitude of movement of the loudspeaker motor is appreciably less than with other enclosures. The effects of nonlinearities in the magnetic field and suspension are therefore greatly reduced, and there is less tendency for "break-up" of the cone to occur. Thus the relatively high distortion products normally produced by the loudspeaker motor will be minimized, and, provided the horn itself does not introduce distortion, extremely high quality sound can be radiated.

A further advantage resulting from this reduction in amplitude of movement of the cone is that a form of inter-modulation distortion, caused by variation of the volume of the cavity between the loudspeaker cone and the throat of the horn, may be reduced to negligible proportions.

Tuning the throat cavity

The cavity, which must inevitably exist between the loudspeaker diaphragm and the throat of the horn, plays an important function in the design of horn systems, since it can be used to limit the maximum frequency to be transmitted. Although the lower frequency limit may be set with some precision by the flare rate of the horn, in conjunction with the mouth area, the upper frequency limit is ill-defined, being determined by a combination of (a) unequal path lengths between different parts of the diaphragm and the throat of the horn, (b) internal cross reflections and diffraction effects within the horn, especially when the horn is folded, (c) the high frequency characteristics of the motor unit itself, and (d) the effective low-pass filter characteristic presented by the cavity between diaphragm and throat.

Fig. 6. Graphical construction of the tractrix.



Using a straight edge of length equal to the final mouth radius, the tractrix curve is constructed of a series of tangents, length not greater than $\frac{1}{10}$ the mouth radius, starting at the throat.

It may be shown that a cavity of fixed volume behaves as an acoustic reactance of value

$$\frac{S_D^2 \rho c^2}{2\pi f V}$$

where S_D = area of diaphragm, V = volume of cavity, ρ = density of air, c = speed of sound, f = frequency.

When the cavity is placed between the diaphragm and throat, it behaves as a "shunt capacitance" across the throat itself, and thus by choosing the correct parameters, the cavity/throat combination acts as a low-pass filter at a frequency which may be set by making the cavity impedance equal to the throat impedance at the desired frequency,

$$\text{i.e. } \frac{S_D^2 \rho c^2}{2\pi f V} = \frac{\rho c S_T^2}{S_T}$$

where S_T = throat area, f = desired upper frequency limit, whence

$$V = \frac{c S_T}{2\pi f}$$

The volume of the cavity may therefore be calculated to provide high-frequency roll-off at a point before the poorly-defined effects (a) to (c) stated above become significant (Fig. 7).

A further benefit resulting from the use of a cavity tuned to prevent mid and high frequencies from entering a bass horn at the rear of a loudspeaker is that the efficiency of transmission of these frequencies by the opposite side of the loudspeaker is greatly increased, thus improving the performance of a mid/high frequency horn mounted at the front of the loudspeaker.

The considerations affecting the practical determination of the upper and lower frequency limits of a particular horn will be considered in more detail.

Loading the rear of the loudspeaker motor

Mention has already been made of distortion resulting from the non-linear expansion/compression characteristics of air. This effect is accentuated when a loudspeaker is horn-loaded on one side only, because the

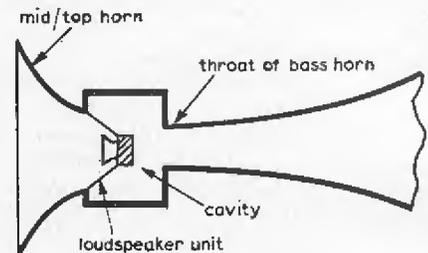


Fig. 7. Effect of the throat cavity in limiting high frequency performance.

constant resistance characteristic of the throat acts only against excursions of the cone in the forward direction; when the cone moves back it is against a far lower load and hence the excursion will be larger. The ideal way of eliminating this distortion is to load both sides of the loudspeaker by equal horns, or to employ a bass horn for loading the rear of the cone and a middle/top frequency horn to load the front. The design of the mini-horn, to be described, utilizes this feature.

An alternative solution favoured by many designers is to load the rear of the loudspeaker by a sealed compression chamber, the effect of which is to provide a loading similar to the horn. The compression chamber thus reduces the effects of non-linearity due to uneven loading on each side of the loudspeaker diaphragm, and also presents a better resistive load to the diaphragm because a closed chamber on the opposite side of the diaphragm to the horn itself acts as a "capacitive" reactance which tends to balance the "inductive" reactance presented by the mass reactance of the throat impedance at low frequencies.

Klipsch states¹⁴ that the volume of this cavity is given by the throat area multiplied

by the speed of sound divided by 2π times the cut-off frequency. This is readily shown as follows:

The air chamber reactance is given by

$$\frac{S_D^2 \rho c^2}{2\pi f_c V}$$

where S_D = diaphragm area, V = volume of air chamber.

The throat reactance at cut-off is

$$\frac{\rho c S_T^2}{S_T}$$

where S_T = throat area.

Equating these,

$$V = \frac{c S_T}{2\pi f_c}$$

However, some observers claim that the use of a compression chamber detracts from the realism of the reproduced sound, and advocate either double horn-loading or a combination of horn-loading with direct-radiation from the other side of the diaphragm; in other words, the most realistic reproduction occurs when both sides of the diaphragm are allowed to radiate.

Summary

In summarizing this section, it is clear that there is no universal formula applicable to any aspect of horn design. The reason for mentioning the alternative approaches and for providing a comprehensive list of references is to stimulate others to experiment in those areas where to a large extent results must be evaluated subjectively by very careful comparative listening tests *a posteriori*.

To quote Wilson:²¹ "It cannot legitimately be assumed that a horn incorporated in a cabinet has the precise characteristics of any particular type of straight horn, whether exponential, hyperbolic, catenary or tractrix, even though their dimensions have been used as guides in its construction. The multiple changes of direction, coupled with reflections and absorptions and internal resonances, are always such as to destroy any legitimate comparison. Every internal (horn) enclosure construction must be judged on its merits as revealed by measurement and by listening tests."

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Basic design procedure

The previous sections have outlined the physical principles underlying the operation of horns, and have shown how, provided certain basic rules are followed, sound reproduction of startling clarity and realism is possible from horns. However, it will also be clear by now that, unless one is prepared to accept extremely large and costly structures, it is all too easy to lose many of the potential qualities of horns through attempts to reduce the size to more acceptable proportions. This section now considers the procedures to be adopted in designing a domestic horn enclosure.

It has already been stated that the horn behaves as a transformer, converting acoustic energy at high pressure and low velocity at the throat to energy at low pressure and high velocity at the mouth. As with the analogous electrical transformer in which electrical voltage and current correspond to acoustical pressure and velocity, the basic requirements of the acoustical horn are that: (a) the primary (throat) should be correctly matched to the signal source (loudspeaker motor); (b) the secondary (mouth) should be correctly matched to the load (listening room); (c) the horn should be designed to handle the specified power level and frequency range. There are four principal parameters of the horn, namely mouth area, throat area, flare contour, and axial length. Any three of these will determine the fourth, and hence the characteristics of the horn itself. Once non-circular cross-section and non-linear axes are adopted, the problem becomes far more complex, and mathematical and physical concepts are no longer sufficient to design a horn. Nevertheless, the basic characteristics even of folded horns are determined to a large extent by known acoustic principles, and the most effective method of design is to work from these principles, ensuring that any deviation from

theory is made on scientific grounds where possible.

Flare profile

Previous sections discussed the most commonly considered flare profiles, and it was concluded that a contour which allowed an exponential increase of the area of the wavefront as it travelled from throat to mouth provided the best compromise between the extremely gradual expansion of the hyperbola (giving optimum loading of the motor, but excessive throat distortion) and the rapid expansion of parabolic and conical horns (giving minimum throat distortion but poor loading of the motor). However, the exact shape of the wavefront within a horn of curved profile is uncertain, and therefore assumptions have to be made, ranging from Wilson's modified exponential (lying a little inside the true exponential) to Voigt's tractrix, (which commences in a virtually identical manner to the true exponential, but departs substantially outside it in the region of the mouth). Which contour one adopts must be largely a matter of personal preference based preferably on careful listening experience.

Mouth geometry

The mouth of the horn couples the horn itself to the listening room. One of the commonly raised disadvantages of horns is that they require a very large mouth area if bass notes are to be properly reproduced. To some extent this is true; one cannot get a double bass out of a piccolo. However, there are a number of ways in which the mouth area may be reduced to manageable proportions without significantly sacrificing bass response.

As a sound wavefront travels up the steadily increasing bore of the horn, it should not meet any major discontinuity. However,

it is clear that, unless the length and mouth diameter of the horn are infinite, there must be some discontinuity as the wavefront emerges and is no longer constrained by the walls of the horn. Although the cut-off frequency of the exponential horn is determined by the flare constant, the linearity with frequency of the acoustical resistance and reactance are determined by the mouth area, which, for a given throat area and flare constant will also determine the overall length of the horn. Strictly speaking, for no discontinuity, the mouth should have infinite area. However, Olson³ has shown that provided the perimeter of the mouth of an exponential horn is greater than four times the cut-off wavelength,

$$\text{i.e. } p_m > 4\lambda_c$$

there will be no significant deviation of mouth resistance from that of the infinite horn.

A more important result is that for only 6dB variation in acoustic resistance, the mouth perimeter may be made equal to the cut-off wavelength, i.e. mouth area = $\lambda_c^2/4\pi$ where λ_c is the cut-off wavelength. As the mouth area is reduced below this value, the non-linearity of the acoustical resistance and reactance will increase.

Now these figures refer to the situation where the horn is suspended in free space, i.e. it radiates into an angle of 4π solid radians. In practice, this situation never occurs: even if the horn were placed on the ground at the centre of an infinite field, the mouth would only radiate into half a solid angle, or 2π solid radians; against the centre of a wall the mouth would be loaded by π solid radians, and in a corner formed by two walls and the floor the mouth will be loaded by only $\pi/2$ solid radians. The significance of this is that, whereas the minimum mouth area for a circular horn has been shown to be $\lambda_c^2/4\pi$ when loaded by 4π solid radians, this value may be divided by a factor of two each time the solid angle is halved. Thus the mouth area may be reduced to a size more in keeping with domestic conditions, e.g. a horn with a cut-off frequency of 56Hz (wavelength 20ft) would require a mouth area of 32 sq ft in space, but only 8 sq ft against a wall and 4 sq ft in a corner position, to give variations in loading of less than 6dB.

This situation which is illustrated in Fig. 8 may be compared with the mouth of a single horn placed at the intersection of eight rooms: four on the ground floor and four on the first floor. The bass response of the original horn will not be impaired, even though a listener in each room will see only one eighth of the total mouth area. One seldom gets anything for nothing in this world, and those who adopt corner speaker positioning in order to obtain a purer extended bass response from as small an enclosure as possible, may have to live with the eigen-tones such a position produces.

A plan view of a corner horn shows that the room itself provides a natural extension of the horn mouth. Many listeners have observed that corner horns can provide bass notes from fore-shortened horns, well below

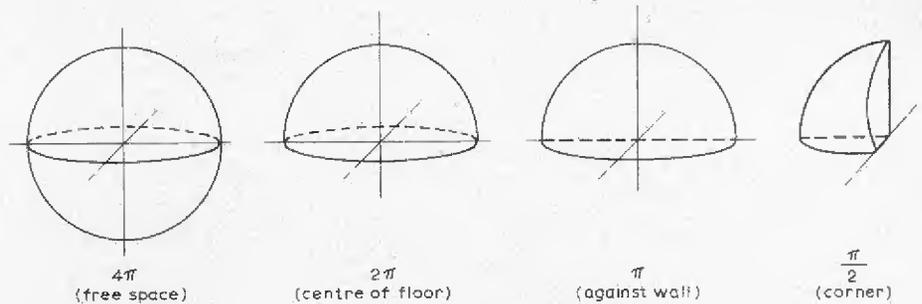


Fig. 8. Solid angles presented to a horn in different positions.

the limit dictated by the mouth area²⁵. It is tempting to reduce the mouth area still further below the 3dB limit established earlier and rely instead on the corner placement itself to supply the additional mouth area and horn length. In the author's experience, this technique cannot be justified because although the bass response is undoubtedly there, careful listening reveals an uneven response over the first two octaves above the cut-off frequency which will often detract from the realism offered by the horn. It is therefore recommended that in cases where overall enclosure size is a limitation, a correctly-designed horn with a cut-off frequency of (say) 80Hz will give a more satisfying and linear response than a fore-shortened horn whose expansion constant has been set to 40Hz but whose length has been limited to give a mouth area corresponding to 80Hz.

Most domestic horns will be of rectangular cross-section for ease and cheapness of manufacture. The foregoing comments regarding horns of circular section apply also to rectangular sections, although it is clear that the wavefront must behave in a most complex way at the corners, thereby reducing slightly the effective cross-sectional area. Provided that the ratio between the major and minor axes at the mouth does not exceed 4:3, rectangular horns may be employed to good effect.

Tabular design data is given for horns of both round and square section, with mouth areas computed for both corner positioning ($\pi/2$ solid radians) and wall positioning (π solid radians).

Throat geometry

The throat of the horn couples the wavefronts from the loudspeaker, which should ideally be plane at this point, to the horn itself. It has previously been shown that the horn is an acoustic transformer, converting acoustic radiation of high pressure/low velocity at the throat to low pressure/high velocity at the mouth. It is clearly of advantage to have a high pressure (and hence a low velocity) at the throat, because the low velocity will result in smaller movement of the loudspeaker cone, thus reducing the distortion produced by non-linearities in the magnetic field and the suspension. One way of increasing the pressure, and also of ensuring a higher degree of "plane-ness" of the wavefronts is to employ a throat area substantially smaller than that of the loudspeaker itself. Tests carried out on a number

of loudspeakers have shown that the "equivalent piston area" is approximately 70% of the speech cone area, i.e. the loudspeaker diaphragm in the shape of a truncated cone gives the same acoustic output as a plane piston with 70% of its area.

There are a number of practical reasons why modern loudspeakers are not manufactured as plane pistons; one of the unfortunate results of employing conical diaphragms is that the resulting wavefronts are in general not planar. However it has been found empirically that a throat area of between one quarter and one half the "equivalent piston area" of the loudspeaker provides satisfactory coupling between the loudspeaker and the horn, and also gives an approximation to plane wavefronts at wavelengths well in excess of the throat dimensions. It must be emphasized that for higher frequencies, where the wavelengths are of the same order as the physical dimensions of the loudspeaker diaphragm, the throat area should be made the same as that of the loudspeaker, and the horn should be of circular section, at least at the throat, so as to minimize standing waves across the horn itself.

The phenomenon of air overload distortion is caused by the non-linear relationship between pressure and volume of the air in the throat of the horn as it undergoes adiabatic compression and expansion. Beranek⁴ has derived the relationship for 2nd harmonic distortion at the throat of an infinite exponential horn as:

$$\begin{aligned} \% \text{ 2nd harmonic distortion} \\ = 1.73(f/f_c)M_p^{1/2} \times 10^{-2} \end{aligned}$$

where f = driving frequency f_c = cut-off frequency I_t = intensity (watts/sq m) at the throat.

This expression is also closely correct for finite horns because most of the distortion occurs near the throat. This expression has been plotted in Fig. 9 from which the throat area for given power and distortion may be obtained.

It is important to appreciate that the acoustic power radiated by musical instruments is extremely small²⁶, and that the higher the frequency the lower is the acoustic power to give the same subjective effect at the human ear. With the exception of full orchestra and pipe organ, which in the author's view it is futile to attempt to reproduce in domestic surroundings at anything approaching normal volume level, the acoustic power levels are extremely small,

and an aim-point of (say) 3 watts and 1% distortion at the cut-off frequency, reducing to 0.05 watts and 0.5% distortion at four times the cut-off frequency, is likely to prove entirely satisfactory for domestic listening²⁷

The above proposals for power and distortion give a throat area of around 10 sq cm, from Fig. 9, which compares not unfavourably with the effective piston area of 43 sq cm for a 3½ in loudspeaker, one quarter of which is a little over 10 sq cm. Of course, if the throat area is increased, as would be the case with larger loudspeakers, the available power for a given level of distortion will also increase.

Having established the throat and mouth areas and the flare profile, the length of the horn and hence its area at any point may be obtained mathematically or graphically.

The horn as a filter

The foregoing sections have indicated how the horn can act as a bandpass filter—the lower pass frequency of which is determined by the expansion coefficient and the upper by the volume of the cavity between the loudspeaker and the throat of the horn. It is important that the response should be as linear and free from distortion as possible over this passband, and as far as the lower frequencies are concerned, careful choice of mouth area, in conjunction with a knowledge of the solid angle into which the horn will radiate and the flare constant, can ensure that non-linearities in the frequency response are kept to a satisfactorily low level.

However, with regard to higher frequencies, non-linearities of increasing amplitude become apparent at frequencies exceeding about four times the cut-off frequency, due to internal cross-reflections and standing waves set up within the horn itself. These non-linearities will be more serious if the material of which the horn is constructed can resonate, and they are also accentuated

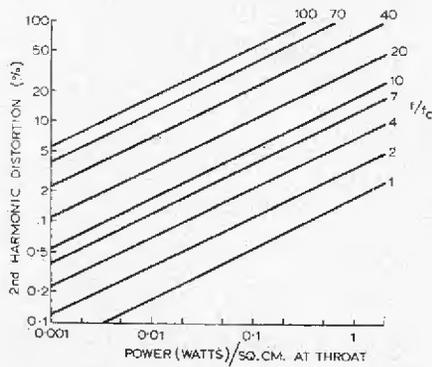


Fig. 9. Distortion caused by air overload at the throat.

if the horn is folded, when wavefronts at the higher frequencies will be distorted at bends. In fact, there is also a practical limit beyond which folding becomes undesirable: folding should not occur beyond the point at which the lowest wavelength (highest frequency) to be transmitted exceeds 0.6 of the diameter of the horn. More will be said of this limitation during the discussion on folding, but it clearly points to a practical limit on the highest frequency a horn may accurately transmit.

Yet a further limitation becomes apparent from the graph of throat distortion versus frequency (Fig. 9). As the frequency increases, the percentage distortion for a given power density at the throat will also increase, and although it is generally true that in the majority of complex musical sounds the energy level reduces with increasing frequency there will still be a frequency above which throat distortion becomes unacceptable.

A commonly used and quite adequate rule of thumb is that a horn should not handle frequencies higher than four octaves above its cut-off frequency, although purists

may prefer to limit at only three octaves in order to ensure lower distortion levels.

The complete multi-horn system

The maximum frequency range to be handled by a wide-range high-quality loudspeaker is about 9 octaves, i.e. 40Hz to 20kHz. This is clearly too wide a range to be handled by a single horn, for the reasons already noted, but it can conveniently be divided into three ranges, i.e. 40Hz to 320Hz, 320Hz to 2.5kHz and 2.5kHz to 20kHz. In practice, a 10% overlap should be allowed to ensure that there are no troughs in the response at the crossover points, and a case could be made for a four-horn system to cover a wider range.

It is also worth considering a more modest instrument. If the cut-off frequency is limited from 80Hz to 18kHz and a two-horn system is considered with each horn handling a little under four octaves, the frequency ranges become 80Hz to 1.2kHz and 1.2kHz to 18kHz. Again, about 10% frequency overlap should be allowed.

The great attraction of a two-horn system is that only a single loudspeaker is required: the bass horn will be loaded from the rear of the loudspeaker; while the middle and treble horns will be loaded from the front of the loudspeaker, to eliminate interference and diffraction effects caused by the frame and magnet assembly at lower wavelengths. It has already been emphasized that the throat of the horn should match exactly the loudspeaker dimensions at these higher frequencies, and this arrangement is particularly attractive if a twin-cone speaker is employed. Treble wavefronts may be prevented from going down the bass horn by the cavity. To show the ease and utility of this approach, this article includes the design of a "mini-horn" utilising both sides of a single loudspeaker in a cabinet of

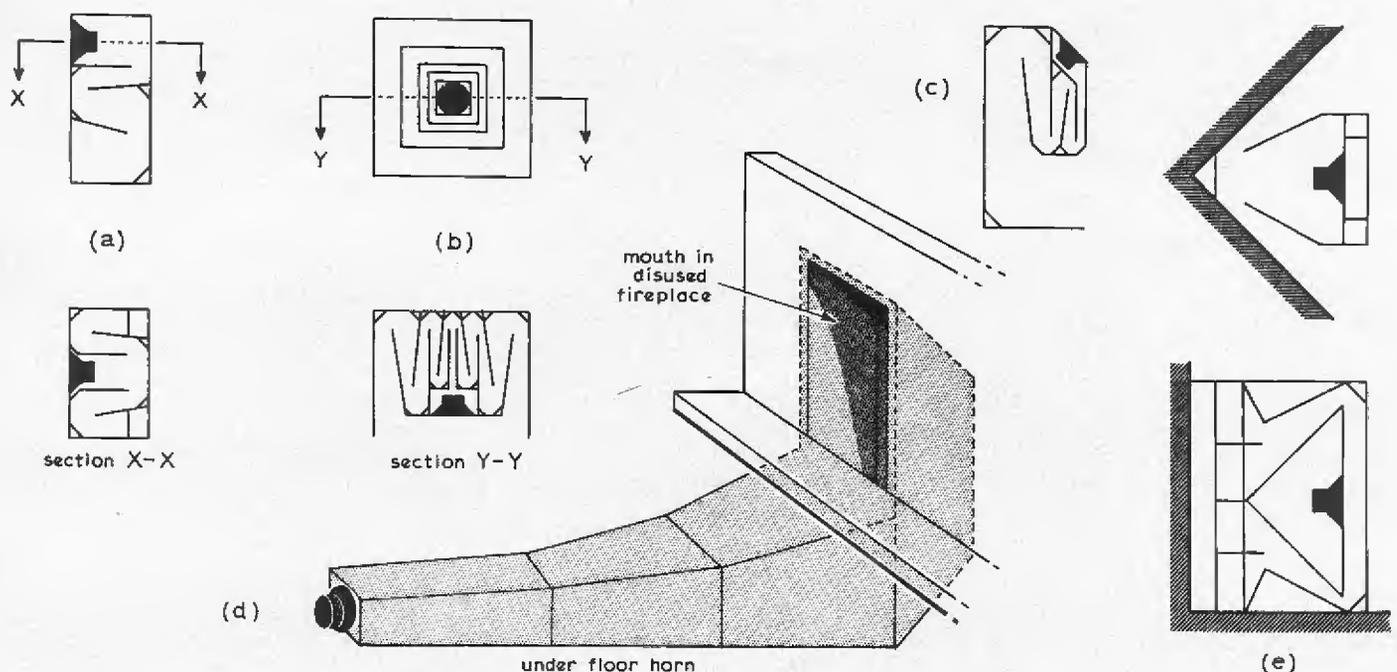


Fig. 10. Methods of folding horns (a) Olson, (b) Olson and Massa, (c) Lowther, (d) Newcombe, (e) Klipsch.

reasonable size and cost for small domestic living rooms.

Purists may claim that the curtailed frequency range of 80Hz to 18kHz is inadequate. It is however the author's experience that the flat relatively distortionless response between these limits, together with the sense of presence afforded by the horn's transformer action, make the mini-horn sound more attractive than many commercial loudspeaker systems of similar size but two or three times its price.

Once one adopts a multi-horn approach, there will be a number of frequencies which fall within the compass of two horns, i.e. 320Hz and 2.5kHz in the case of the three-horn system, and 1.2kHz for the two-horn system. It is essential that the radiation from the relevant pair of horns should be reasonably in phase at the crossover frequency, to avoid the presence of troughs in the frequency response, because the bass horn will be folded to bring its mouth adjacent to the other horns (it is not normally necessary or desirable to fold the middle and treble horns). This requirement places a restriction on the length of the horn, which has until now apparent that the difference in lengths of the lower horn of each pair should be either an odd or even number of half wavelengths of the crossover frequency, depending on whether the radiation wavefronts at the throats of the two horns are respectively in or out of phase.

Thus, if separate loudspeakers are used and the voice coils are connected in phase, the difference in lengths of the horns from the loudspeakers to the plane of the mouths should be an even number of half wavelengths. Conversely, if a single loudspeaker is used to feed two horns, the radiation from the front and rear of the cone will be out of phase and the difference in lengths of the two horns should be an odd number of half wavelengths. In practice, the lower horn will be considerably longer than the upper, and will effectively determine the design.

Folding, cabinets and room placing

Hitherto, discussion has been confined to ideal horns, of circular cross-section and straight axis, constructed of very stiff material. Although typical dimensions for practical horns have not been calculated formally, it will be clear from many of the tables and diagrams that the dimensions of bass horns are almost certainly too large for comfortable accommodation in an average living room. Two further points must therefore be added to the design procedure, adoption of rectangular sections and folding the horn into a compact size.

Rayleigh showed that bends in tubes of constant cross-section will have no effect on transmitted sounds if the wavelength is large compared with the diameter, but that any cross vibrations set up will have a fundamental wavelength of 1.7 times the tube diameter. Wilson¹¹ has summarized the three principal rules of folding horns as follows: the wavefronts must not be twisted

across the horn; the horn diameter (or width if rectangular) must be less than 0.6 times the lowest wavelength to be transmitted by that horn; the wavefront should be accelerated round bends to preserve its form.

As soon as one departs from the straight horn of circular cross-section, the scientific design principles described cease to be so relevant and become of more approximate value, although the three basic rules quoted above, together with the choice of a suitably stiff material for construction, provide very acceptable results.

A folding technique which twists the wavefront across the horn is difficult to achieve in practice, and may be eliminated by folding always in one plane. The requirement to "accelerate the wavefront around bends to preserve its form" is difficult to achieve when more than one fold is involved, since it requires a rectangular cross-section before the bend to become trapezoidal around the bend itself¹¹, and then revert to a different rectangular section after the bend. If one considers a multi-fold horn, concertina-fashion within an overall rectangular enclosure, this is not really a practical proposition, and is unnecessary because subsequent bends correct the waveform. But for single bends it can be adopted, and the mini-horn design described later could utilize this feature.

Examination of the Patent Office records for folded horn designs registered during the 1920s and 30s provides a fascinating monument to the ingenuity of acoustical designers, and Fig. 10 illustrates a number of the more well-known methods of folding. The restriction of horn width at a bend to 0.6 times the highest wavelength to be transmitted suggests initially that folding can only be attempted over the first few feet of the length of a horn; after that point the width will have reached the limiting value. However, this limitation may be overcome by bifurcating the horn (splitting into two equal channels) at each point when the width limits. Thus the mouth of a horn may comprise four equal mouths (brought together for convenience and to ensure audio realism) and the four "quarter-horns" may be folded far closer to the mouth than would otherwise be possible. Rayleigh has shown⁷ in Art. 264 that bifurcating a conduit will have no effect on the transmission of sound provided the lengths of the two portions are equal and the sum of their areas

at corresponding points is equal to that of the original conduit.

In many cases, the front side of a loudspeaker, whose reverse side is horn loaded, will be physically close to the mouth of the horn itself, and it is commonly feared that there will be cancellation at certain frequencies caused by interference between the two radiations in anti-phase. However, the direct radiation from the unloaded front of the cone is only a few percent of that through the horn, and so the amount of cancellation is negligible.

Frequency handling

Although it has been shown that each horn acts as an acoustic bandpass filter, the lower cut-off frequency being determined by the expansion coefficient and the upper cut-off frequency by the throat cavity, there are important reasons why the full audio signal should not be applied directly to all horns regardless of their frequency handling capability. At the low frequency end of the spectrum, examination of Fig. 3 (Part 1) shows that the horn provides the loudspeaker with no resistive acoustic loading below its cut-off frequency. Thus any applied signals below this frequency will cause excessive movement of the loudspeaker diaphragm, which will be constrained only by the mechanical and electro-magnetic factors. This excessive movement can cause unpleasantly high intermodulation distortion, and can also lead to further non-linear distortion when the loudspeaker moves outside its linear range. At the upper frequency end, signals of excessive power can also give rise to distortion products due to deficiencies in the cavity/throat relationship. It is therefore beneficial to restrict the bandwidth of the electrical signal reaching each loudspeaker to match the acoustic bandwidth of its corresponding horn.

Although most commercial multi-unit loudspeaker systems use passive LC crossover networks between power amplifier and loudspeaker to route signals of the appropriate bandwidth to each loudspeaker, careful comparative listening tests show that these networks undoubtedly introduce a "dullness" or loss of "brilliance" into the audio output. Many explanations have been offered for this situation; in the author's opinion, the most likely reason being the loss of "direct drive" from the output of the amplifier, allied with a significant reduction in the degree of electro-magnetic damping afforded

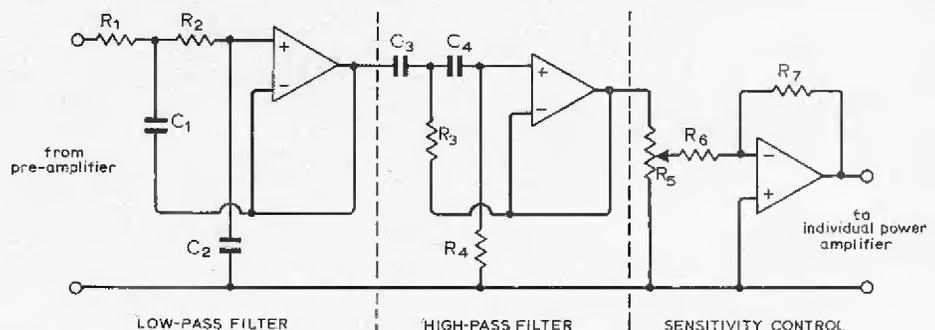


Fig. 11. Circuit for an active filter network. See appendix for component details.

by the low output impedance of the amplifier.

Recent correspondence in *Wireless World*²⁸ and elsewhere has extolled the virtues of splitting the frequency range at low signal level, and employing a separate power amplifier directly coupled to each loudspeaker. The author has devised such a circuit, which consists of three (or four) parallel frequency-selective channels comprising Sallen & Key active filters giving preset low and high-pass characteristics in series in each channel, together with some gain adjustment to allow for the inevitable differences in sensitivity of each loudspeaker/horn combination. The active filters provide 2nd order Butterworth characteristics, a response which appears to give the least displeasing effects at the cross-over frequencies. (There will inevitably be phase-shifts associated with any filter circuit, and the effects of these on transients can produce a marked difference in their character.) This circuit is in Fig. 11 and the Appendix.

Thus, some form of electrical cross-over is generally necessary in addition to the acoustic cross-over provided by the horn itself. An exception is of course the case where a single loudspeaker drives two horns: one loading the front and one loading the rear of the diaphragm. In this situation, some compromise will be necessary in the acceptable distortion level and bandwidth of the loudspeaker system.

Directional horns

This article has extolled the ability of the horn to propagate wavefronts that are nearly plane at its mouth. However, there are situations where it is desirable to propagate wavefronts with different characteristics in the vertical and horizontal planes, particularly when middle and treble horns are used in stereophonic systems; it is often desirable to spread the wavefronts in the vertical plane while preserving more of a "point-source" in the horizontal plane. There are a number of different techniques for achieving this, based on diffraction and refraction effects at the horn mouth with the comparatively short wavelengths (a few inches or less) with which these high frequency horns are concerned.

The design and manufacture of multicellular horns, distributed-source horns, diffraction horns and reciprocal-flare horns is beyond the scope of this article, and with the exception of the first two mentioned is probably outside the capability of most amateur constructors. Those interested should refer to the papers by Smith²⁹, Winslow³⁰ and to the relevant chapters by Olson³ and Cohen⁵.

Klipsch^{16,17} has described the design of his high frequency horn, in which the length/breadth ratio of the (rectangular) mouth assumes a value in excess of 4:1 c.f. the ratio of near unity advocated for bass horns). The optimum dimensions, length/breadth ratio, and apportionment of flare to the long and short axes depend on a number of complex factors, however, an aspect ratio between 2:1 and 4:1 with the flare apportioned in similar ratio has been

found to give good practical results, and these parameters have been adopted for the "no-compromise horn" to be described. Although the high frequency horn of the "mini-horn" system is specified as circular (in view of its handling the relatively large wavelengths at 1kHz) an alternative rectangular mouth with aspect ratio of 2.5:1 has also been described.

Detailed design procedure

The previous sections have dealt in some detail with the basic theory of the horn, and the essential design procedures have been outlined for a series of horns which can cover the complete audio range. The final sections will consider the detail design of two horns: a "mini-horn" and a "no-compromise horn".

Because all horns are designed to slightly different requirements, and inevitably many readers will wish to "bend" the specification to a greater or lesser extent in order to satisfy their own needs, the designs are presented here by means of tables so that they represent a comprehensive design code for a wide range of horns.

Bass horn design

The bass horn should be examined initially, commencing with the mouth. Tables 1, 2 and 3 indicate the relationship between minimum frequency and mouth dimensions for horns positioned in free air (4π solid radians) at a wall (π solid radians), and in a corner ($\pi/2$ solid radians). In table 1, the speed of sound has been taken as 1125 ft/sec, and the mouth perimeter as the wavelength. The mouth areas in tables 2 and 3 are equal to $\frac{1}{4}$ and $\frac{1}{8}$ respectively of the mouth area in free air, and the dimensions for the circular, square and rectangular mouths are derived from these areas. It is tempting to reduce the areas of the square and rectangular horns so as to give a perimeter equivalent to the wavelength (suitably scaled for wall or corner positioning) but this is not recommended. However, the shorter side of the rectangular horn has been derived in this way (i.e. a square horn with this side would have the appropriate perimeter).

After settling the mouth dimensions, the throat may be determined from the chosen loudspeaker unit. Table 7 gives suggested throat areas for five typical mean loudspeaker sizes. In some designs, the choice of loudspeaker will be influenced by considerations of overall size (the length of the horn is greatest for the smallest loudspeaker) and whether the loudspeaker is to perform as both bass and mid/top driver, using two separate horns on either side. Many loudspeakers will possess different dimensions, and in these cases table 7 will be of little value. The effective area (piston area) has been taken as 0.7 of the area derived from the mean (quoted) diameter, and the throat area as 0.3 of the effective area. Although there is obviously scope for experiment here, the quoted dimensions should give very acceptable results.

Having decided the throat and mouth areas, tables 8 and 9 give the overall lengths

of horns with true exponential and tractrix contours for both wall and corner placing for horns with the five derived throat areas at each of the cut-off frequencies specified in table 1. The factor of 1.2 applied to the cut-off frequency in table 5 when calculating the flare coefficient is to ensure a fairly linear frequency relationship throughout the working range of the horn. The flare coefficient m is thus given by

$$m = (4\pi/c)(f/1.2)$$

where c is the speed of sound (1125ft/sec) and f is the lowest frequency to be reproduced.

The area increase is given by $(e^m - 1)\%$ and the doubling distance by $(\log_e 2)/m$ for each frequency. The length of the exponential horn is given by $(1/m) \log_e S_m/S_T$ for each specified set of areas, and the length of the tractrix horn will be $r_m(1 - \log_e 2)$ shorter than the true exponential, where S_m = mouth area, S_T = throat area, r_m = mouth radius.

N.B. The tractrix lengths given in tables 8 and 9 are approximations, being based on the fully developed tractrix referred to the flare cut-off frequency, whereas the mouth radius is referred to the lowest bass frequency to be reproduced.

Middle top horn design

Attention should now be directed to the middle and high frequency horns. The mouth perimeter should not be less than the wavelength of the lowest working frequency, and in practice a perimeter of 1.5 times the lowest working frequency has been found to give good results. Table 4 is based on this factor of 1.5, and gives the recommended minimum mouth dimensions for free air loading. It is safest to assume free air loading to apply at these higher frequencies, because diffraction and reflection effects at short wavelengths prevent true wall or corner loading from being achieved, and it is for this same reason that the perimeter has been specified at 1.5 times the wavelength of the lowest working frequency. The dimensions of square and rectangular horns have been derived in the same way as those in tables 2 and 3. The throat dimensions of middle and high frequency horns should match the drive unit directly, and may be taken as the mean diameter and area of the chosen loudspeaker, shown in table 7. Tables 6 and 10 give the flare constants and lengths of exponential horns assuming the throat and mouth dimensions of tables 7 and 4 respectively.

Integration of multiple horns

It has been emphasized that the radiation from the mouths of each pair of horns at their common crossover frequency should be in-phase. Assuming that the mouths of all the horns will lie in the same plane, the difference in length of each pair of horns should be compared with the multiples of half wavelengths of the crossover frequency set out in table 11. If the drive signals at both throats are in-phase (separate loudspeakers), the difference in length should be an even number of half-wavelengths; if the drive sig-

Table 1

Freq. (Hz)	Wave-length (ft)	Diameter (ft)	Area (sq. ft)
30	37.5	11.94	111.98
40	28.13	8.95	62.92
50	22.5	7.16	40.27
60	18.75	5.97	28.0
70	16.07	5.12	20.59
80	14.06	4.48	15.77
90	12.5	3.98	12.44
100	11.25	3.58	10.07
110	10.23	3.25	8.30
120	9.38	2.98	6.98

Table 1. Minimum mouth dimensions for bass horn (free loading).

Table 5

Freq. (Hz)	Cut-off freq. (Hz)	Flare coeff. (ft ⁻¹)	Area increase (% ft ⁻¹)	Doubling dist. (ft)
30	25	.278	32	2.49
40	33	.366	44	1.89
50	42	.466	59	1.49
60	50	.555	74	1.25
70	58	.644	90	1.08
80	66	.733	108	.945
90	75	.833	130	.832
100	84	.932	154	.744
110	92	1.02	178	.679
120	100	1.11	205	.624

Table 5. Exponential constants for bass horn.

Table 7

Nom. dia. (in)	Area (sq. in)	Effective area (sq. in)	Throat area (sq. in)	Throat area (sq. ft)
3½	9.62	6.74	2.02	.014
5	19.64	13.75	4.12	.029
6½	33.19	23.23	6.97	.048
8	50.27	35.19	10.56	.073
10	78.55	54.99	16.50	.114

Table 7. Throat dimensions.

Table 2

Freq. (Hz)	Area (sq. ft)	Dia. (ft)	Sq. side (ft)	Rect. sides (ft)
30	28	5.97	5.29	4.69 5.97
40	15.73	4.47	3.96	3.52 4.47
50	10.07	3.58	3.17	2.81 3.58
60	7.0	2.98	2.64	2.34 2.98
70	5.15	2.56	2.27	2.01 2.56
80	3.94	2.24	1.98	1.76 2.24
90	3.11	1.99	1.76	1.56 1.99
100	2.52	1.79	1.58	1.41 1.79
110	2.07	1.62	1.44	1.27 1.62
120	1.74	1.49	1.32	1.17 1.49

Table 2. Minimum mouth dimensions for bass horn (wall position).

Table 6

Freq. (Hz)	Cut-off freq. (Hz)	Flare coeff. (in ⁻¹)	Area increase (% in ⁻¹)	Doubling dist. (in)
200	166	.155	17	4.48
250	208	.193	21	3.59
300	250	.233	26	2.97
350	292	.271	31	2.56
400	330	.307	36	2.26
450	375	.349	42	1.98
500	420	.391	48	1.77
550	458	.426	53	1.63
600	500	.465	59	1.49
700	580	.539	71	1.29
800	660	.614	85	1.13
900	750	.698	101	.993
1000	840	.781	118	.887
1100	920	.855	135	.810
1200	1000	.930	153	.745
1300	1083	1.01	175	.686
1400	1166	1.08	196	.642
1500	1250	1.163	218	.596
2000	1660	1.54	368	.450
2500	2080	1.93	590	.359

Table 6. Exponential constants for mid/top horn.

Table 10

Freq. (Hz)	3½	5	6½	8
200	30.9	26.3	22.9	20.3
250	22.5	18.8	16.1	14.0
300	17.1	14.0	11.8	10.0
350	13.6	10.9	8.98	7.46
400	11.1	8.78	7.07	5.73
450	9.09	7.05	5.55	
500	7.58	5.77	4.42	
550	6.51	4.84		
600	5.56			
700	4.24			
800	3.31			
900	2.58			

Table 10. Length of mid/top horn (in), free loading. Since the mouth perimeter equals 1.5 times the highest working wavelength, the tractrix cannot be used. Tractrix contours can however be incorporated if the mouth perimeter is reduced to one wavelength.

Table 3

Freq. (Hz)	Area (sq. ft)	Dia. (ft)	Sq. side (ft)	Rect. sides (ft)
30	14.0	4.22	3.75	3.32 4.22
40	7.87	3.16	2.81	2.49 3.16
50	5.03	2.53	2.24	1.99 2.53
60	3.5	2.11	1.87	1.66 2.11
70	2.57	1.80	1.60	1.42 1.80
80	1.97	1.58	1.40	1.24 1.58
90	1.55	1.41	1.25	1.10 1.41
100	1.26	1.27	1.12	0.995 1.27
110	1.04	1.15	1.02	0.904 1.15
120	0.87	1.05	0.93	0.829 1.05

Table 3. Minimum mouth dimensions for bass horn (corner position).

Table 8

Freq. (Hz)	3½		5		6½		8		10	
	Ex	Tr								
30	27.3	25.1	24.7	22.5	22.9	20.7	21.4	19.2	19.8	17.6
40	19.2	17.6	17.2	15.6	15.8	14.2	14.7	13.1	13.5	11.9
50	14.1	12.8	12.6	11.3	11.5	10.2	10.6	9.3	9.62	8.30
60	11.2	10.1	9.88	8.78	8.98	7.88	8.22	7.12	7.42	6.32
70	9.17	8.23	8.05	7.11	7.25	6.31	6.60	5.66	5.92	4.98
80	7.69	6.83	6.70	5.84	6.01	5.15	5.44	4.58	4.83	3.97
90	6.48	5.75	5.61	4.88	5.00	4.27	4.50	3.77	3.97	3.24
100	5.57	4.91	4.79	4.13	4.25	3.59	3.80	3.14	3.32	2.66
110	4.90	4.30	4.18	3.58	3.69	3.09	3.28	2.68	2.84	2.24
120	4.34	3.79	3.68	3.13	3.23	2.68	2.85	2.30	2.46	1.91

Table 8. Length of bass horn (ft) for different flare constants, wall position. Ex-exponential, Tr-tractrix. N.B. The tractrix lengths are approximate.

Table 4

Freq. (Hz)	Wave-length (in)	Dia. (in)	Area (sq. in)	Sq. side (in)	Rect. sides (in)
200	67.5	32.2	815.4	28.6	25.3 32.2
250	54.0	25.8	522.9	22.3	20.3 25.8
300	45.0	21.5	365.1	19.1	16.9 21.5
350	38.57	18.4	265.9	16.3	14.5 18.4
400	33.75	16.1	203.6	14.3	12.6 16.1
450	30	14.3	160.6	12.7	11.3 14.3
500	27.0	12.9	130.7	11.4	10.1 12.9
550	24.55	11.7	107.5	10.4	9.2 11.7
600	22.5	10.7	89.9	9.5	8.4 10.7
700	19.28	9.2	66.5	8.2	7.2 9.2
800	16.88	8.1	51.5	7.2	6.3 8.1
900	15	7.2	40.7	6.4	5.6 7.2
1000	13.5	6.4	32.2	5.7	5.1 6.4
1100	12.27	5.9	27.3	5.2	4.6 5.9
1200	11.25	5.4	22.9	4.8	4.2 5.4
1300	10.38	4.9	18.8	4.3	3.9 4.9
1400	9.64	4.6	16.6	4.1	3.6 4.6
1500	9	4.3	14.5	3.8	3.4 4.3
2000	6.75	3.2	8.0	2.8	2.5 3.2
2500	5.40	2.6	5.3	2.3	2.0 2.6

Table 4. Minimum mouth dimensions for mid/top horn (free loading).

Table 9

Freq. (Hz)	3½		5		6½		8		10	
	Ex	Tr								
30	24.8	22.6	22.2	20.0	20.4	18.2	18.9	16.7	17.3	15.1
40	17.3	15.7	15.3	13.7	13.9	12.3	12.8	11.2	11.6	10.0
50	12.6	11.3	11.1	9.8	9.98	8.66	9.08	7.76	8.12	6.80
60	9.95	8.85	8.64	7.54	7.73	6.63	6.97	5.87	6.17	5.07
70	8.10	7.16	6.96	6.02	6.18	5.24	5.53	4.59	4.83	3.89
80	6.75	5.89	5.75	4.89	5.07	4.21	4.50	3.64	3.89	3.03
90	5.65	4.92	4.78	4.05	4.17	3.44	3.67	2.94		
100	4.83	4.17	4.05	3.39	3.51	2.85				
110	4.22	3.62	3.51	2.91						
120	3.72	3.17								

Table 9. Length of bass horn (ft) for different flare constants, corner position. Ex-exponential, Tr-tractrix. N.B. The tractrix lengths are approximate.

nals are out-of-phase (single speaker horn-loaded at both front and rear) the difference in length should be an odd number of half wavelengths. If necessary, small changes may be made to the crossover frequency (with subsequent re-design of the higher frequency horn) to ensure optimum conditions at crossover.

The complete design

The bass horn will generally be folded. Originally it was intended to provide a table giving the maximum permitted length of horn before folding should cease because the horn diameter has become equal to 0.6 times the lowest wavelength to be transmitted. However, examination has shown that at frequencies up to five times the bass cut-off frequency (i.e. 4 octaves bandwidth) this restriction does not apply to the corner-positioned horn (due to the small mouth dimensions) and with the wall-positioned horn the limitation lies between 92% and 95% of the full exponential length. It may therefore be assumed that provided the wall-positioned horn is not folded within the final 10% of its length, the problem of cross reflections will not arise.

Finally, the cavities at the throats of the lower frequency horns should be designed in accordance with the formula already given, remembering to allow for the loss of cavity volume due to the frame, magnet and cone assembly of the loudspeaker itself.

The design procedure laid down in this part has been applied to two different designs of horn to follow, and further examples of overall horn design are given in refs 34 to 37, and also in ref. 5.

Appendix

A variable bandpass active filter for feeding a 3 horn loudspeaker system (see Fig. 11):

Low-pass filter

Frequency (Hz)	R_1 (k Ω)	R_2 (k Ω)	C_1 (pF)	C_2 (pF)
200	59	59	20,000	10,000
1k	12	12	20,000	10,000
2k	59	59	2,000	1,000
10k	12	12	2,000	1,000
6k	59	59	680	330
30k	12	12	680	330

N.B. R_1 & R_2 to be realized as 12k in series with 47k log pots.

High-pass filter

Frequency (Hz)	R_3 (k Ω)	R_4 (k Ω)	C_3 (pF)	C_4 (pF)
25	28	57	160,000	160,000
100	7	14	160,000	160,000
250	28	57	16,000	16,000
1k	7	14	16,000	16,000
4k	28	57	1,000	1,000
16k	7	14	1,000	1,000

N.B. R_3 to be realized as 6.8k in series with 22k log pot. R_4 to be realized as 12k in series with 47k log pot.

All i.c.s to be Signetics N5741V, etc. R_5 10k log, R_6 22k, R_7 100k.

Table 11. Multiples of half wavelengths (ft).

Freq. (Hz)	$y/2$	y	$3y/2$	$2y$	$5y/2$	$3y$	$7y/2$	$4y$	$9y/2$	$5y$	$11y/2$	$6y$
200	2.81	5.62	8.44	11.24	14.1	16.87	19.7	22.4	25.3	28.1	30.9	33.6
250	2.25	4.5	6.75	9.0	11.3	13.5	15.8	18.0	20.2	22.5	24.7	27.0
300	1.87	3.75	5.63	7.5	9.4	11.25	13.1	15.0	16.9	18.7	20.6	22.5
350	1.61	3.21	4.82	6.42	8.05	9.64	11.2	12.84	14.4	16.1	17.7	19.26
400	1.41	2.81	4.22	5.62	7.05	8.44	9.84	11.25	12.6	14.1	15.5	15.9
450	1.25	2.50	3.75	5.0	6.25	7.5	8.75	10.0	11.3	12.5	13.7	15.0
500	1.125	2.25	3.38	4.5	5.63	6.75	7.87	9.0	10.14	11.25	12.4	13.5
550	1.02	2.04	3.07	4.08	5.1	6.14	7.16	8.16	9.21	10.2	11.24	12.24
600	.938	1.87	2.81	3.75	4.69	5.63	6.56	7.5	8.43	9.38	10.3	11.26
700	.803	1.61	2.41	3.21	4.01	4.82	5.62	6.42	7.23	8.03	8.9	9.64
800	.703	1.41	2.11	2.81	3.51	4.22	4.92	5.62	6.33	7.03	7.7	8.44
900	.625	1.25	1.88	2.50	3.12	3.75	4.38	5.0	5.64	6.25	6.87	7.50
1000	.562	1.125	1.68	2.25	2.81	3.38	3.93	4.5	5.04	5.62	6.18	6.76
1100	.511	1.02	1.53	2.04	2.56	3.07	3.58	4.08	4.59	5.11	5.62	6.14
1200	.468	.938	1.41	1.87	2.34	2.81	3.28	3.75	4.23	4.68	5.15	5.62
1300	.433	.865	1.29	1.73	2.16	2.59	3.03	3.46	3.87	4.33	4.75	5.18
1400	.402	.803	1.21	1.61	2.01	2.41	2.81	3.21	3.63	4.02	4.41	4.82
1500	.375	.75	1.12	1.50	1.87	2.25	2.62	3.0	3.36	3.75	4.12	4.50
2000	.281	.562	.84	1.125	1.41	1.68	1.96	2.25	2.52	2.81	3.09	3.36
2500	.225	.45	.68	.90	1.12	1.35	1.57	1.80	2.04	2.25	2.47	2.70

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Constructional principles

Much has been written about the best methods of constructing loudspeaker enclosures, especially regarding rigidity and the prevention of resonances and leaks, and as far as the horn is concerned, these points are equally important. The horn enclosure has to stand up to considerable acoustic stress, and any shortcomings in its manufacture are liable to cause more serious aural distress than would be the case with some other enclosures.

Ideally, the horn should be cast in circular section, but this form of construction is only practical with small middle and high frequency horns. The technique best adopted by the home constructor is to cast in plaster-of-Paris using a plywood mould and reinforcements as necessary. The calculated profile should be set out using plywood templates held in place by stringers, which will be buried within the casting itself as "reinforcement". It is also a good plan to provide a $\frac{1}{2}$ in panel at the throat end for mounting the loudspeaker, and a further panel surrounding the mouth which will help in securing the complete horn assembly and fixing any decorative cloth finish.

Bass horns are almost invariably constructed of flat panels cut so as to approxi-

mate to the correct flare profile. Plywood, chipboard or blockboard are satisfactory, either $\frac{3}{4}$ or $\frac{1}{2}$ in thick. Composite sand-filled panels consisting of two thin walls of $\frac{1}{8}$ in plywood spaced $\frac{3}{4}$ or 1in apart and filled with dry sand provide extremely rigid resonance-free enclosures. Great care should be taken to prevent any particles of sand, sawdust, etc. from entering the speech-coil gap of the loudspeaker; a recent demonstration mounted by the author at St. Albans was spoiled by distortion caused by a minute particle of wood in one of the treble loudspeakers. Care must also be taken to ensure that the sand is dry, otherwise the wood panels will rot. It is advisable to bake the sand on shallow trays in an oven to ensure that all moisture is removed. After filling, a few minutes of organ music will help the sand to settle down ready for "topping-up".

The wooden panels should be fixed together using wood-screws and with a liberal application of a liquid glue along all mating faces. This not only adds strength, but also makes all the joints air-tight. Further strength should be provided by triangular corner fillets and "glue-blocks" placed at intervals along the longer joints. In addi-

tion, smaller reflecting plates should be placed at the outside of all sharp corners to "ease" the wavefronts around bends, and to help preserve the steady exponential increase in horn area, as indicated in Fig. 10. Manufacturing tolerances should not exceed $\pm \frac{1}{16}$ in at the throat but errors of $\frac{1}{2}$ in at the mouth of the bass horn are unlikely to have any noticeable effect on the performance. It is worth bearing in mind that the velocity of sound, on which all design calculations are ultimately based, itself varies by as much as 5% at climatic extremes.

A vital detail is to ensure that the loudspeakers can be fitted and removed easily, maintaining an overall air-tight construction by means of thin rubber gaskets if necessary. It should of course be remembered that the highest pressures occur at the throat, and the greatest effort to ensure rigidity and absence of leaks should be made in this area. As the cross-sectional area of the horn increases, it is a good plan to fit longitudinal stiffening panels, made of $\frac{3}{4}$ in plywood, across the centre of the horn, thereby converting the horn into two symmetrical adjacent ducts. This reduces air turbulence effects at bends and makes the bends themselves less critical in addition to providing extra cross-bracing between panels that might otherwise resonate. It is worth fitting longitudinal stiffeners for the final 25% of each bass horn.

Unlike the majority of loudspeaker enclosures, there is no need to provide any sound absorbent material within the enclosure itself (except within the compression chamber, if fitted, which may be lined with acoustic wadding, long-hair wool, etc., to absorb high frequency sound). The interior of the horn should have all sharp edges removed with sandpaper and all internal corners filled with putty or a similar setting

plastic compound and smoothed down by means of a finger. This practice, which is not mandatory, also has the effect of sealing any remaining air leaks. The whole interior surface should be treated with a thick coat of gloss paint.

Design of a "mini-horn"

The intention of the mini-horn is to provide as many as possible of the benefits of horn-loading within an enclosure which is sufficiently small for use in a small living room, where the overall size is of course especially important when a quadrasonic or even a stereophonic installation is under consideration. The room for which this particular mini-horn was originally designed imposed limitations of 20 in as the maximum intrusion into the room, and an overall height of 4 ft; fortunately, corner positioning was acceptable.

It was clear that only one loudspeaker could be used, and after some thought, the Eagle FR65 was chosen. This is a co-axial twin-cone loudspeaker in which the inner (tweeter) cone is itself shaped as a small horn. This subsidiary cone will handle the extreme top of the frequency range and beam it out axially through the treble horn. The loudspeaker has a nominal diameter of 6.5 in, a frequency range of 35 to 18,000 Hz and power handling capacity of 10 watts. It is clear that, since top frequencies will be dealt with by the tweeter cone, the bass horn need only cover 3 octaves, i.e. from 70 Hz to 560 Hz, and the middle-frequency horn can take over at (say) 500 Hz. This middle-frequency horn will be most efficient for 4 octaves, i.e. up to about 8 kHz, at which point the tweeter cone will already be taking over. The complete frequency spectrum will therefore be:

Bass horn	70 Hz to 650 Hz
Middle horn	500 Hz to 8000 Hz
Top horn	
(tweeter cone)	8 kHz upwards

The other design consideration at this stage is the power handling capacity. A bass power of 0.3 watts at a distortion level of up to 1% was decided.

Bass horn

In order to derive the greatest benefit from corner-positioning, the mouth of the bass horn should be at floor level and should stretch horizontally from wall to wall. A mouth consisting of a quadrant of a circle of 19 in radius was considered, giving a horizontal arc of 2.48 ft. Examination of Table 3 (Part 2) shows a minimum mouth area of 2.56 sq.ft for a horn capable of reproducing down to 70 Hz, and dimensions of 2.48 ft \times 1.03 ft (29.7 in \times 12.5 in) were therefore chosen for the bass horn. Table 7 suggests a throat area of 0.048 sq.ft (i.e. 45 sq.cm) and Fig. 9 indicates that for 1% distortion at 7 times the cut-off frequency (i.e. 490 Hz) the power at the throat will be 0.007 watts/sq. cm, giving 0.3 watts total, which is the specified value.

Table 9 shows that the bass horn will have a length of 6.18 ft (exponential contour) or 5.24 ft (tractrix contour). It was decided to adopt the tractrix so as to give a shorter overall length, and the complete contour has been constructed in Fig. 12.

Treble horn

The treble horn will load the front of the loudspeaker, commencing at the nominal diaphragm area of 23 sq.in, which is thus the throat area for this horn. The lowest frequency to be handled is 500 Hz, and from Table 4 the mouth area is 130.7 sq. in, which may conveniently be realized as 10.1 in \times 12.9 in. Table 10 now gives the horn length as 4.42 in.

It is also possible to adopt a circular format for this horn, in which case the mouth diameter will be 12.9 in or altern-

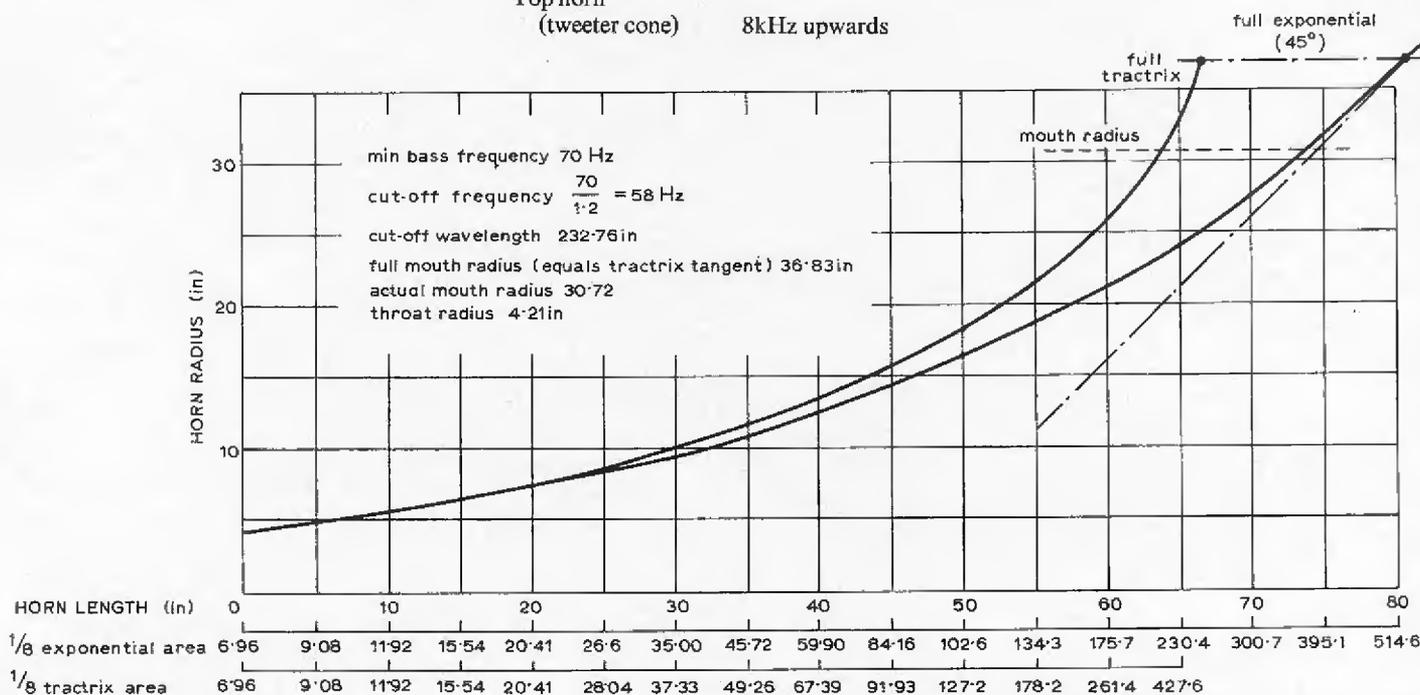


Fig. 12. Flare contour for the bass section of the mini-horn.

actively some degree of horizontal directivity may be introduced by adopting an aspect ratio of 2.5:1 (larger dimension horizontal). In this event the mouth dimensions become 18.08in × 7.23in, and the flare contours should be arranged to give the appropriate expansion (see Fig. 13).

Integration and complete design

Radiation at the throats of the mid-frequency and bass horns will be in anti-phase, since these horns load the front and back respectively of the single loudspeaker. Since the mouths of both horns will be in the same vertical plane, the difference in length should be an *odd* number of half-wavelengths at the crossover frequency to ensure that radiation from both mouths is in phase at this frequency. The lengths are 5.24ft and 0.37ft, giving a difference of 4.87ft. At 580Hz this length corresponds closely to five half-wavelengths as shown by Table 11. This design thus includes a satisfactory combination of horns.

The cavity which couples the rear of the loudspeaker diaphragm with the throat of the bass horn should now be designed to cut off radiation from the bass horn at 650Hz.

$$V = cS_t / 2\pi f$$

where V = cavity volume, S_t = throat area, f = cut-off frequency, whence $V = 23$ cu.in. The volume taken up by the magnet, etc, of the loudspeaker is approximately 21 cu.in (obtained by direct measurement) and thus the overall cavity volume will be some 44 cu.in. In

many cases the minimum space necessary to enclose the rear of the bass loudspeaker apparently exceeds the optimum cavity volume for giving the correct upper cut-off frequency, often by a factor of four or five times. Since the cut-off frequency is inversely proportional to the cavity volume, this will have the effect of giving a serious "trough" in the overall frequency response before the mid-frequency horn takes over. The answer is to reduce the cavity to the correct volume by means of a circular plaster or wood moulding leading from the rear of the loudspeaker diaphragm to the throat of the horn. This technique has been well described by Crabbe (ref. 19), and is illustrated in Figs 15 and 16. The parameters of the mini-horn have been summarized in Table 12.

Finally, the bass horn should be folded into a suitable shape, and the two horns integrated together. In view of the limited space available and the desire for simplicity, a design with only one major fold may be adopted, shown in Fig. 14(b) (*overpage*). The mouth of the bass horn is at the bottom of the enclosure between the two walls, making contact with the floor. The mouth of the mid-horn is placed immediately above this, in the same plane, leading back to the loudspeaker itself. The loudspeaker is mounted on a small baffle board which supports the middle-frequency horn at the front and the cavity coupling to the throat of the bass horn at the rear. The bass horn bends vertically upwards almost immediately after the throat to a point some 4ft high at which it doubles back on itself down the corner of the room to form the mouth. The cross sectional area may conveniently be made trapezoidal, but the design shown will not preserve "plane-ness" of wavefronts around the bends. Fig. 14(b) (*overpage*) illustrates the general arrangement only, as readers may well wish to make modifications for personal reasons. The material used should be $\frac{3}{4}$ in chipboard, etc, except the side and front panels of $\frac{1}{2}$ in plywood.

Design of a no-compromise horn

Following the many qualifications already stated in this article, it must be clear that "no-compromise" is in itself a misnomer—horn design consists largely of making the most effective compromises between conflicting requirements. However, this design is aimed at the situation where the best possible practice can be followed without being unduly hampered by limitations of either size or position. Nevertheless, it would be somewhat pointless to design an enclosure which cannot be built without professional tools, facilities and materials, so the design has been conducted with a large living room (or small hall) in mind, and directed towards the competent do-it-yourself enthusiast.

In order to cover a wide frequency range, a three-horn design has been adopted, using three separate loudspeaker units.

Bass Horn	40Hz to 400Hz
Mid-frequency Horn	350Hz to 3.8Hz
Treble Horn	3.5Hz to 20Hz

A total acoustic power handling capacity of 1W at a distortion level up to 2% has been chosen as being more than adequate for the situation envisaged. A wall-mounting design is adopted, using a tractrix bass horn, to simplify the eventual positioning of the system. (Of course, there is no reason why a design intended for wall-mounting should not be placed in a corner but it is not recommended that corner designs be placed against a wall.)

The choice of loudspeaker drive units is not straightforward; there are many commercial units worth using in horns, although individual designers inevitably have their own favourites. For this design it was decided to use:

Horn	Unit	Power(W)	Frequency	Diaphragm
			Range (Hz)	Area (Nominal sq.in)
Bass	KEFB139	30	20-1k	42
Middle	KEFB110	15	30-5k	13
Treble	KEFT27	6	3k-30k	1.5

Bass horn

The mouth area for a minimum frequency of 40Hz is given in Table 2 as 15.73 sq.ft, realized as 47in by 48in. The diaphragm area of the B139 loudspeaker is approximately 42 sq.in (the diaphragm is oval in shape) which corresponds closely with the 8in speaker of Table 7. This table gives a recommended throat area for the bass horn of 0.073 sq.ft (68 sq.cm). The highest frequency to be handled by this horn will be 400Hz, i.e. 10 times the cut-off frequency. Fig. 9 shows that for 2% distortion at 10 times the cut-off frequency, the throat will handle 0.01 watts/sq.cm, which for a mouth area of 68 sq.cm gives 0.68 watts total power. The length of the bass horn, from Table 8, is 13.1ft using the tractrix contour, and this curve may be constructed to a suitable scale in the same way as that produced for the mini-horn. The form of the tractrix should commence with a throat radius equivalent to four times the chosen throat area (eight times in the case of the corner horn) and terminate at a mouth radius giving a perimeter equal to the cut-off wavelength. The area at a series of points along the horn (e.g. every 6in) may be obtained by reading the radius from the graph and taking one quarter of the corresponding area.

Middle horn

Attention should now be directed to the mid-frequency horn. The cut-off frequency of 350Hz, together with the area of the chosen loudspeaker (13 sq.in) result in a mouth area of 266 sq.in, a throat area of 13.75 sq.in and a length of 10.9in (exponential contour). Again the contour should be constructed to a suitable scale (which may be 1:1 for the mid-frequency and treble horns). Since the throat areas of the

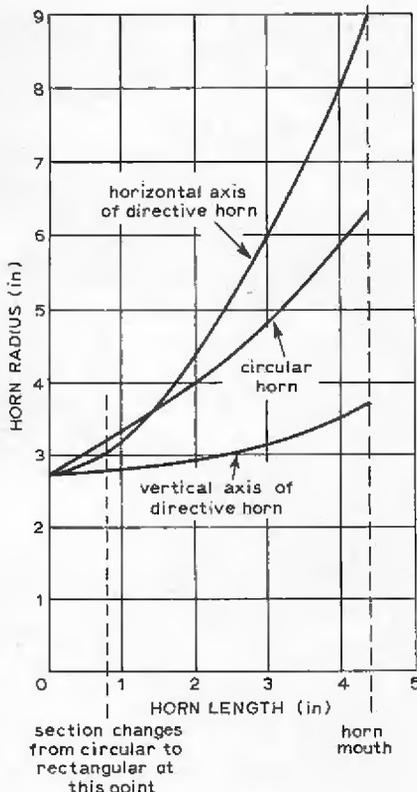


Fig. 13. Flare contour for the treble section of the mini-horn.

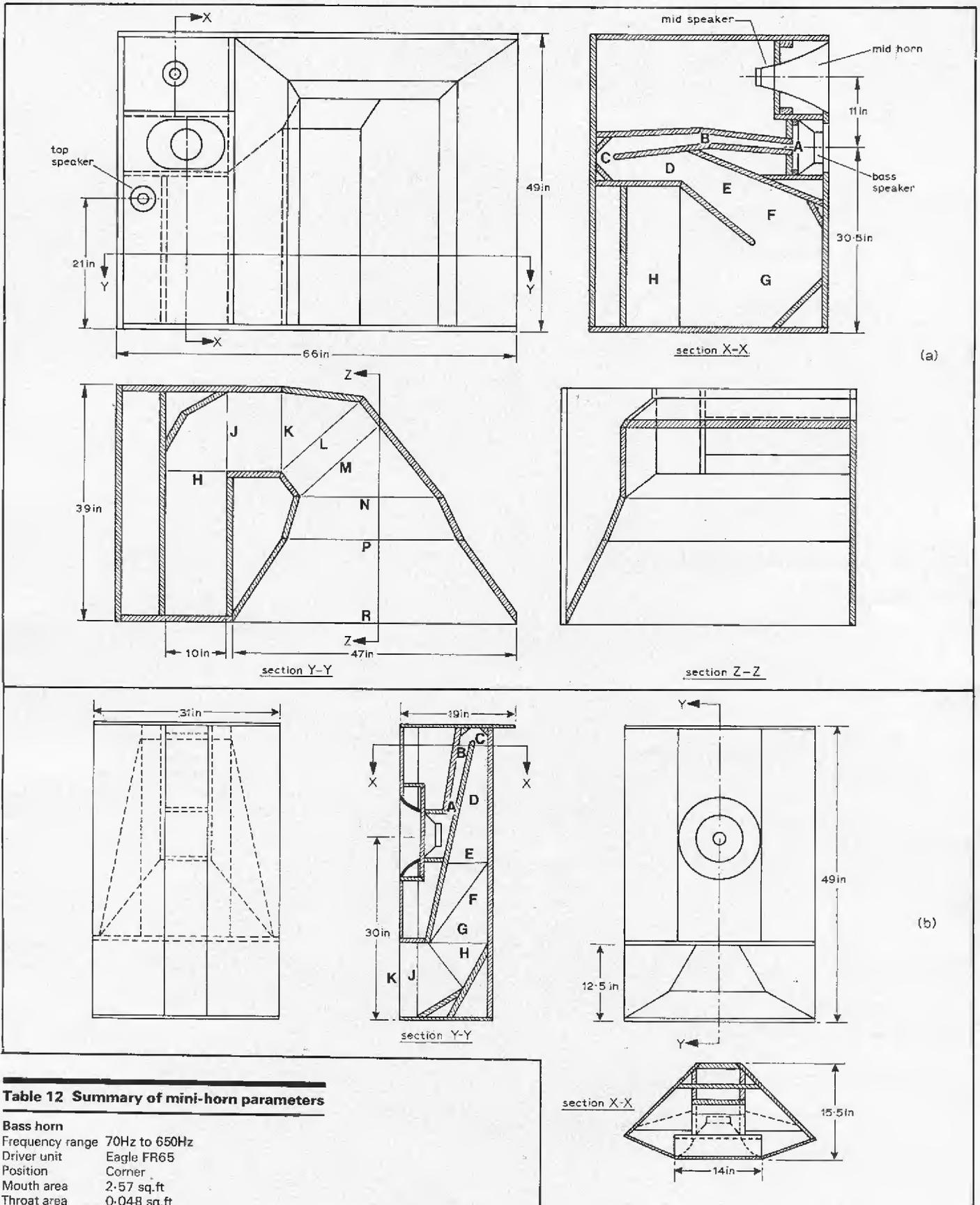


Table 12 Summary of mini-horn parameters

Bass horn	
Frequency range	70Hz to 650Hz
Driver unit	Eagle FR65
Position	Corner
Mouth area	2.57 sq.ft
Throat area	0.048 sq.ft
Contour	Tractrix
Length	5.24ft
Cavity volume	44 cu.in (including 21 cu.in loudspeaker volume)

Treble horn	
Frequency range	500Hz upwards
Driver unit	Eagle FR65
Mouth area	130.7 sq.in
Throat area	23 sq.in
Contour	Exponential
Length	4.42in

Fig. 14 (a) Suggested realization of the no-compromise horn (b) suggested realization of the mini-horn. Cross sectional details for the mini bass horn are given in Table 14 and for the no-compromise bass horn in Table 15.

middle and treble horns are made equal to their respective loudspeaker diaphragm areas, there is no problem regarding air overload distortion—one could do little if there were.

Treble horn

In view of the throat area of 1.5 sq.in for this horn, it is suggested that a mouth area of (say) 30 sq.in is adopted, giving an exponential length of 1.1in.

However, at these frequencies it is quite acceptable to mount the loudspeaker directly onto a flat baffle board without any horn.

Integration and complete design

The three loudspeaker drive units should drive their respective horns in phase. Initially it will be assumed that, whereas the middle and treble horns must load the front of the loudspeakers (to avoid diffraction effects caused by the frame and magnet assembly at the rear), the bass horn will in fact load the rear of the loudspeaker. This implies that the bass loudspeaker must be connected in anti-phase to the middle and treble loudspeakers. If examination of the behaviour of the bass and middle

Table 13

Summary of no-compromise horn parameters

Bass horn	
Frequency range	40Hz to 400Hz
Driver unit	KEF B139
Position	Wall
Mouth area	15.73 sq.ft
Throat area	0.073 sq.ft
Contour	Tractrix
Length	13.1ft
Cavity volume	56 cu.in (directly at front of loudspeaker)
Middle horn	
Frequency range	350Hz to 3.8kHz
Driver unit	KEF B110
Mouth area	266 sq.in
Throat area	13.75 sq.in
Contour	Exponential
Length	10.9in
Treble horn	
Frequency range	3.5kHz upwards
Driver unit	KEF T27
Mouth area	30 sq.in
Throat area	1.5 sq. in
Contour	Exponential
Length	1.1in

Table 14

Cross section details for the mini bass horn

Sec.	Length (in)	Area (sq.in)	Realized (in)
A	0	7.0	1
B	10	11.9	1.7
C	15	15.5	2.2
D	25	28	4
E	35	49	7
F	40	67	} complex section
G	45	92	
H	50	127	
J	60	261	
K	63	370	
			26 x 10 high
			29.7 x 12.5 high

horns at their mutual crossover point reveals that the radiation is in anti-phase, the bass horn can be arranged to load the front of its loudspeaker, which will then have to be connected in phase with the other two. In fact the difference in length of the bass and middle horns is 12.2ft and Table 11 shows that this is nearly equivalent to an even number of half-wavelengths at 370Hz. The bass loudspeaker may therefore be reversed so that the horn loads the front of the diaphragm; this will also make the design of the acoustic cavity much simpler.

The cavity between the bass loudspeaker and its horn should be designed to give a cut-off frequency of 400Hz. Applying the aforementioned formula gives a volume of 56 cu.in. There is no real need to employ a similar cavity at the crossover between the middle and treble horns; the fact that these horns are not folded, together with their large throats, reduces distortion at high frequencies to negligible proportions.

Finally, the three horns must be combined into a composite enclosure. As with the mini-horn there are many ways of achieving this, and it would be invidious to specify a particular design to the exclusion of all others. However, certain basic rules apply, and the following suggestions may be of value.

The rectangular mouth of the bass horn should be placed at floor level, with the mouths of the middle and treble horns placed above it, in the same plane. If the back-to-front depth of the complete structure is at a premium, the middle and treble horns may be mounted on top of the complete folded bass horn, giving a very high cabinet. If, however, height is at a premium, then the bass horn may be folded behind the middle and top horns, thus minimizing the overall height but increasing the width. This latter approach is shown in Fig. 14(a), and the complete design of the no-compromise horn is summarized in Table 13. Material used for construction is 1in block-board, plywood, etc, and all joints should be screwed and glued to make them airtight.

When converting from a basic parameter design, as described in this section, to complete working drawings, the temptation is often to press on rapidly and adopt certain compromises. Unfortunately, the final construction is a "once only" event, and horn structures cannot easily be modified if major audio deficiencies (e.g. resonances or "holes" in the frequency spectrum) become apparent during listening tests. It

is therefore strongly recommended that the final design takes place over an extended period, with several alternative approaches being worked on simultaneously until one of them emerges as the right solution for the parameters and overall concept in mind.

The three loudspeaker units must be connected via suitable filters so that each handles frequencies only within its appropriate pass-band. The simplest way of achieving this is by means of passive

Table 15

Cross section details for the no-compromise bass horn

Sec.	Length (in)	Area (sq.in)	Realized (in)
A	0	10.6	1.06
B	15	17.0	1.7
C	30	27.8	2.78
D	40	38.0	3.8
E	50	54.0	5.4
F	60	75.0	7.5
G	75	122	12.2
H	94	230	23
J	104	316	13.5 x 23 high
K	113	426	13.5 x 31.5 high
L	120	540	17 x 31.5 high
M	125	630	17 x 37 high
N	132	790	21 x 37 high
P	139	1125	28 x 40 high
R	153	2265	47 x 48 high

crossover networks at the output of the power amplifier. However, this method reduces the beneficial effects of electromagnetic damping of the loudspeaker movement afforded by the low output impedance of the amplifier, and a better method is to use three separate power amplifiers whose inputs are fed via active high and low pass filters, as outlined in Part 2.

It is well-worth experimenting over an extended listening period until the optimum bandwidth and sensitivity of each horn has been realized, paying particular attention to the crossover points.

Conclusions

This article has taken the form of a critical review of work which took place largely between 20 and 50 years ago. The author of such reviews benefits from hindsight, but inevitably loses much of the excitement and impact of the original work. I have been in contact with many individuals who were personally involved with the development of horns, in both amateur and professional capacities; I thank them all for their advice and comments, and hope that I have done justice to their suggestions.

In spite of the obvious disadvantages of large size and high cost, and the difficulties of realizing an adequate design, the exponential horn loudspeaker still has many enthusiastic users, the present author among them. The clear advantages conferred by the horn in terms of presence, clear bass, low distortion and sheer realism, combine to make horn enthusiasts redouble their efforts to design a better horn rather than to adopt an alternative type of enclosure.

It will be clear to readers of this article that, with the possible exception of straight horns of circular section constructed in a very stiff material, the simple horns described here can only approximate to the ideal performance offered by this genre of reproducer. Although the pioneer development work was conducted between 50 and 70 years ago, engineers are continuously designing new horns and investigating different aspects of their performance, often with the aid of computers to construct a

mathematical model for the analysis of conditions in a practical horn (38, 39). It must be emphasized again that first-class results may be obtained by following the basic design data and constructional advice given in this article. Loudspeakers in general, and horns in particular, are controversial subjects, and I have no doubt that many will wish to challenge some of the statements I have made. I hope that this article, together with any discussion, will stimulate many audio enthusiasts to design and make their own horns, and to write about the results so that all may benefit from their findings.

Finally, I acknowledge with thanks the helpful advice given by Mr Gilbert Telfer, whose experience of the design and manufacture of horns has been a constant encouragement.

Letters to the Editor

Horn loudspeaker design

I read with great interest Mr Dinsdale's first article on horn loudspeakers in the March issue and much appreciated the historical survey given by the author. The later part of the article dealing with horn theory was, however, slightly confusing, not only the obvious lack of correspondence between Fig. 2 and the text accompanying it but more so the statements made under "Loading the rear of the loudspeaker motor". It is hard to believe that the difference in loads presented to the diaphragm between positive and negative excursions should be of major importance. The air load on a diaphragm is the pressure difference across it and, assuming sinusoidal or near sinusoidal pressure variations on both sides, the net difference is the same irrespective of direction of movement, only a change of sign taking place. The pressure on either side of the diaphragm may have any instantaneous value as long as the variations are periodic.

The effect of loading the rear of the loudspeaker with a closed chamber is best seen from the expression for the mechanical impedance at the throat:

$$Z_{MT} = R_{MT} + j \frac{1}{\omega \cdot C_{MT}}$$

The reactance should clearly be termed a negative compliance; thus the positive compliance presented by the diaphragm suspension in parallel with the compliance of the closed chamber can be made to cancel the throat reactance at all frequencies above the cut-off frequency, not only at cut-off. From a loading point of view a closed chamber is superior to a mid-range horn.

A compound horn design must rely on the stiffness of the diaphragm suspension to load it at low frequencies.

Tore Hevrenge,
Lidingö,
Sweden.

The author replies:

I am sorry that my article has confused Mr Hevrenge. Fig. 2 was intended to illustrate the non-linearity of the adiabatic p-v relationship for air, as I tried to explain in the accompanying text. It is clear from Fig. 2 that the change in volume resulting from a compressive stress is less than that resulting from an equal expansive stress, and this does indeed give rise to serious distortion at the throat of a horn, where pressures can be very high.

For example, Fig. 9 (in part 2 of the article, May issue) indicates that when operated at a level of 1 watt/sq.cm at the throat, a horn will give distortion levels of nearly 2% at the cut-off frequency, and 20% at ten times this frequency, due to the non-linear p-v relationship. At lower power levels of course the distortion becomes insignificant.

I am grateful to Mr Hevrenge and a number of readers for pointing out an oversight on my part, that the reactance of the closed compression chamber used by some designers to load the rear of the loudspeaker is capacitive while the reactive component of the throat impedance is inductive. The compliances of the diaphragm suspension and the closed chamber can, of course, be arranged to cancel the throat reactance at all frequencies above cut-off, but examination of Fig. 1 shows that for the exponential and hyperbolic horns this reactance becomes significant only at frequencies near to cut-off.

Finally, the loudspeaker should never be allowed to rely merely on the stiffness of the diaphragm suspension to load it at low frequencies; as I indicate in part 2 of the article the signal bandwidth should be limited electrically (preferably by active filters at low level, rather than by passive crossover networks between power amplifier and loudspeakers), so that the horn is not driven at frequencies below cut-off. If this is done, it is my experience that horns in which the unloaded side of the diaphragm is allowed to radiate give a better performance than those with a compression chamber.

J. Dinsdale.

Horn loudspeaker design

I have been following Mr Dinsdale's series on horn loudspeakers (March, May, June) with great interest as I have recently designed and built a three unit horn system, and I thought my experiences might be of interest to those intending to build the Dinsdale loudspeakers.

Coincidentally, I also chose to base the design on the three KEF units used by

Dinsdale. The low frequency horn is folded in a similar manner to the Klipschorn¹ and uses a compression chamber behind the B139 driver as well as a small air chamber in front of it connected to a rapid initial flare section, as suggested by Klipsch. This exponential horn was designed for corner use and has a flare cut-off frequency of 50Hz and a mouth area of 550sq. in. In use this horn gives an apparently smooth response from 400Hz down to about 35Hz, with very high efficiency and an overall clear, undistorted sound. Any resonances present are less noticeable than the natural resonances of the small rooms in which it has been used.

The problems with the system come at the top end and are caused by two factors: colouration and poor dispersion. The colouration, which takes the form of audible resonances in the mid and high frequency ranges, seems to be due to transverse reflections between the walls of the horn, especially in the throat region where the cross-section is almost square. This was confirmed by using a microphone probe which picked out standing waves across the horn whereas longitudinal resonances were not noticed (the mouth area being sufficiently large to obviate reflections). The T27 tweeter simply refused to sound right with any form of horn loading. In fact if the T27 is mounted flush in a baffle as suggested by KEF and is fed with white noise, audible colouration occurs as soon as any hard object is placed within about six inches of the diaphragm!

The top end horns were intended to give good horizontal dispersion over an angle of 90°. This is necessary to preserve the off-axis response which otherwise falls at high frequencies. To this end the mid-range horn was made with a mouth 10in high and 20in wide and incorporated four splitters in the throat section to give better angular dispersion of the high frequencies. This technique had only moderate success.

It should be noted that the "plane wavefronts" advocated by various authorities must by their very nature give rise to highly directional propagation, especially at high frequencies. This gives poor mono reproduction and a small stereo listening area. For this reason cinema horn loudspeakers invariably employ some form of diffraction on the high frequency unit, either by a multicellular design^{2,3} or by means of an acoustic diverging lens⁴.

The above faults made the system sound characteristically coloured when compared with professional monitor loudspeakers (the Spendor BC1 and Rogers BBC monitor) although it sounded fairly reasonable on its own. For any further development of a horn top end I would personally opt for a drive unit specifically intended for "pressure loading" (which means in effect a smaller, lightweight diaphragm loosely suspended) and work along the lines indicated by Klipsch⁵. Conventional speakers do not seem to take kindly to horn loading.

One point which Mr Dinsdale does not

seem to have covered in his historical survey is the effect of a time delay due to the length of the low frequency horn; if this is several feet long the low frequencies will be delayed by several milliseconds. In the 1930s it was noticed that this can cause audible echos on some transients and thenceforth the high frequency horns of cinema loudspeakers were moved back so that the drivers were in line rather than the horn mouths. While this is not entirely practical for folded domestic systems, the high frequency horns should be set back as far as possible. Phase matching at the crossover frequency is still desirable, of course, taking into account the phase shifts in the crossover network itself.

I would like to end by suggesting that in order to minimize the size of bass horns more research should be done into the design of corner standing units. The corner horn can be thought of as an acoustic coupling between the drive unit and the conical horn formed by the corner of the room. Freehafer⁶ has analysed a horn of this form, the true hyperbolic horn (not to be confused with the more common family of horns characterized by hyperbolic trigonometric functions and often called "hyperbolic" or "hypex" horns). He was able to do this without making the usual plane wave assumptions and found that the low frequency response was much better than that of the conical horn to which it is asymptotic. He states that "... hyperbolic horns favour the low frequencies to a much greater extent than do the corresponding conical ones. Since the hyperbolic horn differs in shape from the conical only in the curvature near the throat, its better performance must be attributed to that curvature. It appears that the ideal horn shape approaches that of a uniform

tube near the throat." This is potentially very interesting for the designer of corner horns as the throat is the only part of the horn over which he has control. Unfortunately due to the complexity of the mathematics involved it would seem that computer simulation of the system is the most promising approach to an optimal corner horn.

D. C. Hamill,
Wimbledon,
London SW19

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The author replies:

Mr Hamill provides some valuable comments on his experiences with horn-loading the KEF units. In my opinion the B139 is the best available driver for bass horns, and even more impressive results can be obtained from using two or even four such units (connected in parallel) at the throat of a suitably-designed horn. For those with limited space, a single bass horn driven by two (or four) B139s, with one (or two) drivers handling the bass range up to (say) 400Hz for each channel, employing acoustic mixing within the horn itself, can provide a useful compromise. There is little stereo information below 1kHz, so this compromise is quite legitimate.

The formation of standing waves across rectangular mid-range horns is all too common an experience, and I feel that the only real solution to this problem is to employ horns of circular section, in spite of the greatly increased difficulty of manufacture. Nevertheless, I have not personally experienced undue distress due to this cause from horns of similar dimensions to those described in my article. Recent experience has now confirmed to me that the Lowther PM6 and PM7 provide the most natural sound in this middle-frequency range, especially when driving circular horns.

I entirely agree with Mr Hamill's comments about the T27, and confirm that it sounds best when mounted flush in a baffle. I would also recommend the Eagle HT21 (which comes complete with its own diecast rectangular horn) as providing a useful addition to the top range.

The point about time delay is an interesting one: clearly the length of the low-frequency horn will cause phase distortion on transients, and I like the idea of setting the high-frequency horns back so that the drivers are in line. Regrettably, as Mr Hamill points out, this is not entirely practical in the domestic situation. I have of course stressed the importance of phase matching at the crossover frequency, and fully agree that phase shifts in the crossover network itself must also be taken into account.

Finally I endorse wholeheartedly Mr Hamill's call for a concerted attack on the optimum design of the corner horn using computer techniques, and I would be pleased to act as a "clearing house" for any ideas and results which readers of *Wireless World* may have on this subject. J. Dinsdale.

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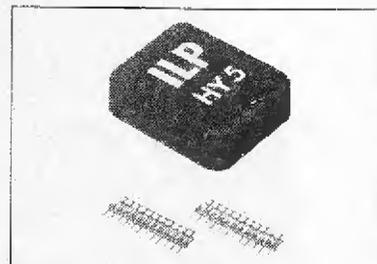
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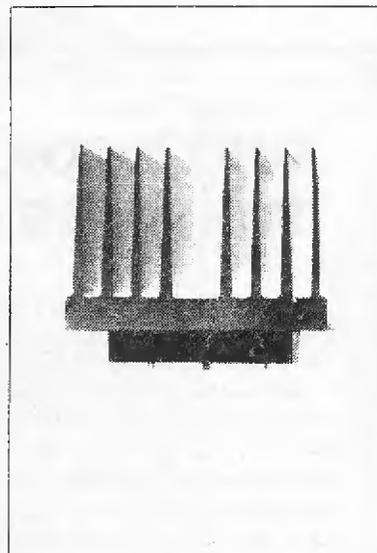
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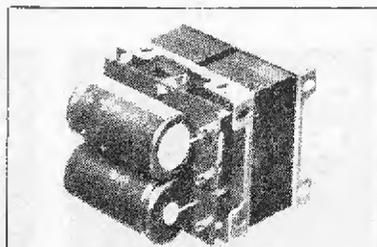
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FEATURES: Thermal shutdown — Very low distortion — Load line protection — No external components.

APPLICATIONS: Public address — Disco — Power slave — Industrial.

SPECIFICATIONS:
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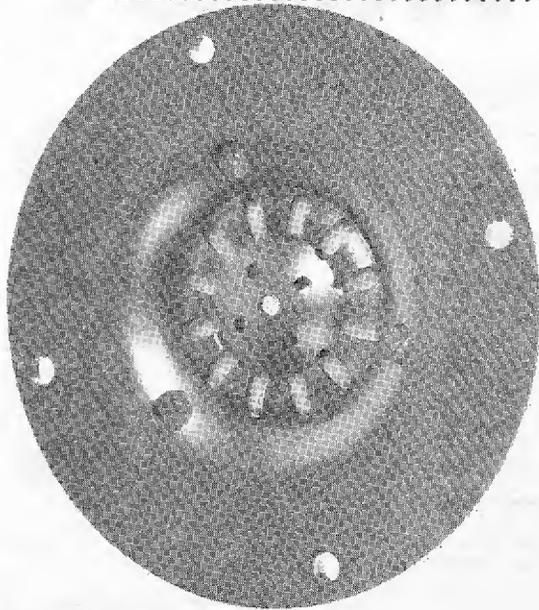
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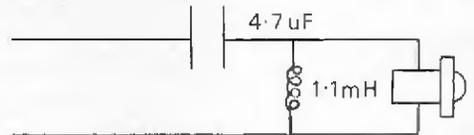
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Dimensions Front Mounting:

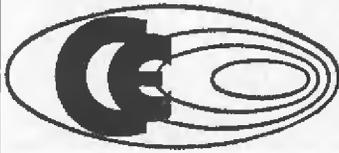
Body 2 1/8" (70.2 m/m) dia. x 1 1/4" (31 m/m) deep
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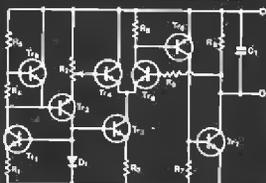
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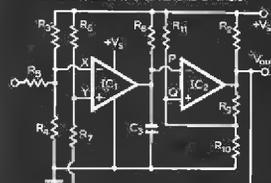
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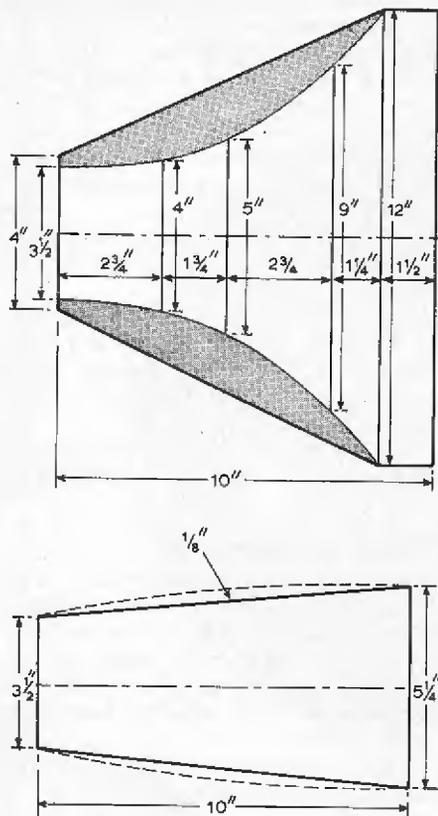


Fig. 2. Details of wood sections used in constructing the treble horn. The side sections should be cut along the curve shown dotted (see text).

mounted from the front against a strip of rubber which should be fixed on the front of the board round the edge of the 11 1/2 in diameter hole to provide an air-tight seal. The front panel does not need a seal but the inset 1 x 1 in mounting frame should be firmly glued in place, and will need chamfering in the top and bottom positions.

Before the triangular corner section is attached, airtight speaker terminals should be fitted. Connecting wire is soldered on and fed through the back-of-cone chamber out through the hole in the speaker board. The wire should be soldered to the tags on the drive unit and the drive unit screwed down tightly.

Four 4 in bracing posts, cut from 1 x 1 in wood should be stuck on to the speaker board, round the drive unit, using p.v.a. adhesive. When the joints are dry paint the ends (with poster paint for example) and press the front panel up against them. Remove the panel, mark its top, and drill through at the centre of each area of transferred paint. The edges of the front panel should also be drilled, as shown in Fig. 1, and the panel screwed into place.

This completes the bass section.

Construction of treble horn

As in the concrete version the horn has flat top and bottom and curved sides. Fig. 2 gives details for the cutting of the sections: Taking the top and bottom pieces first each piece can be cut from a rectangle 10 x 12 in, and 12mm ply should be

used. Each six-sided piece of wood should be marked up as shown and the curves constructed. The areas shown shaded will lie outside the horn when it is finished.

The curved sides of the horn should be cut from a flexible plywood such as bending ply. The cutting should follow the dotted curve so that the side can be glued vertically along one of the curved lines drawn on the top or bottom panel. The joint should be made using an 'impact' adhesive as the wood will be under stress. The remaining flat panel can be matched up and similarly fixed in place.

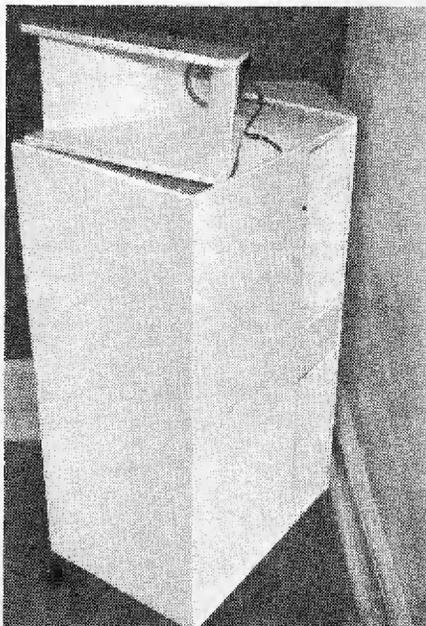
Throat section. Cut a 3 1/2 in diameter hole in a 4 in square piece of 5mm ply (not shown in Fig. 2) and glue it in place over the throat end of the horn. Fill the inside of the throat region with Polyfilla (interior grade of course) the four 'fingers' of plaster stopping about 3 1/2 in from the throat. This procedure provides the correct exponential transition from circular to rectangular cross-section.

When the Polyfilla is dry give the inside of the horn a layer of undercoat and a couple of layers of gloss paint. The Eagle FR4 drive unit can be fitted to complete the treble horn.

Crossover circuit

The modifications to the bass horn have reduced its efficiency. Advantage is taken of this in that the attenuating resistors in series with the FR4 are bypassed with a small capacitor to boost the top and compensate for the somewhat reduced efficiency of the horn at high frequencies. The series network is shown in Fig. 3.

Winding the chokes. A 2 in piece of 3/8 in diameter ferrite rod (with cardboard discs glued on at the ends) can be wound with 37ft 6in of 24 s.w.g. enamelled copper wire. The turns should be close and the



A finished speaker in its corner.

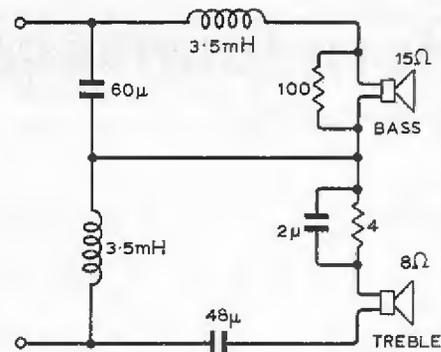


Fig. 3. Crossover circuit using a 1/2-section series network. Resistors can be 1/2W.

layers neat. The ferrite rod is available in 4 in and 6 in lengths from G. W. Smith (Radio) Ltd. To break the rod, first file a shallow notch. Place a pin on a hard surface, such as a metal ruler, and with the notch facing upwards press the ends of the rod downwards with the pin lying exactly below the notch. This should result in a clean break.

Capacitors. The reversible 50V electrolytic capacitors used are available from K.E.F. Electronics Ltd, of Tovil, Maidstone, Kent. 60 microfarad capacitors are available but the 48 microfarad required on the treble side is made up from 3 x 16 microfarad units. The prices are 48p per pair for the 16 microfarad, and 68p per pair for the 60 microfarad units.

Horn performance

The system described (which was demonstrated at the Audio Fair) will work perfectly if the bass horn is placed in the corner up against the two walls. However, a gap of up to four inches will result in no significant change in performance. The treble horn can be turned round to alter the apparent direction of the sound.

Because very few loudspeakers are capable of launching plane waves at mid-range frequencies few listeners ever hear really good stereo. It is a characteristic of a well-designed radial horn (and a well-designed flat electrostatic radiating element) that the pressure contours have flat fronts. The intersection of two such wavefronts, provided by loudspeakers angled in toward the listener, produces a stable stereophonic image. Constructors of a pair of horns for stereo reproduction will find that they can move freely about in their listening room in the same way that they can move about at a live recital.

Of course, the type of radiation pattern described is a *sine qua non* for worthwhile quadruphony.

Overall efficiency of the system is such that it is unlikely to be overdriven in domestic use, even at high sound levels, when used with a 10W-per-channel amplifier.

A provisional patent application has been filed on aspects of the system.

The transmission-line loudspeaker enclosure

A re-examination of the general principle and a suggested new method of construction

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

Since the wool-filled transmission-line loudspeaker enclosure was first described† there has been a steadily increasing interest in its use.

The basic transmission-line enclosure is shown in Fig. 1. Radiation from the back of the driver cone flows down a pipe filled with a low-density sound-absorbing material. Fibrous absorbents such as loose wool, cotton wool and kapok can be used; sound absorption decreasing as the frequency goes down. In general it is very difficult to obtain good absorption if the path length is less than one-quarter wavelength of the sound in free-space; at 30Hz this corresponds to a path length of about 9ft.

If the pipe length is less than 9ft, sound at and below 30Hz will emerge from the open end of the pipe. Due to time delay in the pipe, the sound will not start to cancel the radiation from the front of the cone until the effective pipe-length is less than one-sixth of a wavelength. It is therefore possible to use the radiation from the open end of the pipe to reinforce that from

the front of the loudspeaker cone at low frequencies.

The effect of the wool filling in the pipe is to slow down the wave relative to its velocity in free air. This reduction factor is between 0.7 and 0.8 for the recommended packing density, so the system will operate down to a somewhat lower frequency than would otherwise be expected.

The folding in the original cabinet design caused sound coloration due to reflections at the bends — particularly the first one at the back of the cabinet. The degree of coloration introduced by this first reflection (which, incidentally, is present in all plain box-shaped cabinets) was quite serious with the high crossover frequency of 1500Hz. Certainly the reproduction without it sounded as if an echo had been removed. The reasons for this were investigated.

In a simple closed box, as shown in Fig. 2, a sound impulse generated by the cone will have two components — the direct radiated pulse from the front of the cone and that propagated back into the cabinet. If this latter is assumed to be a plane wave, i.e. sound travelling parallel to the cabinet sides, it will strike the back wall and bounce back to the cone still as an impulse. Some of this energy will radiate through the cone to the outside and the remainder will be reflected back into the cabinet for re-reflection. The net result is a succession of steadily weakening pulses being radiated from the cabinet. The acoustic output will therefore be as shown in Fig. 3.

If the reflection from the back wall of the cabinet is changed so that it is gradual rather than abrupt, then the reflected wave will not be a unit impulse from an initiating unit impulse, but a pulse whose length and shape will depend on the nature of the reflection. This may be more readily appreciated by referring to Fig. 4. Here, the back wall of the cabinet is triangular. Sound radiated from the back of the cone is successively subjected to reflection, the first reflection occurring due to sound from the edge of the cone, and the last being due to that from the centre. As the path lengths for these reflections are very different, the sound in the cabinet

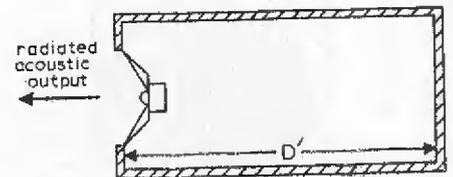


Fig. 2. Unlagged totally enclosed cabinet.

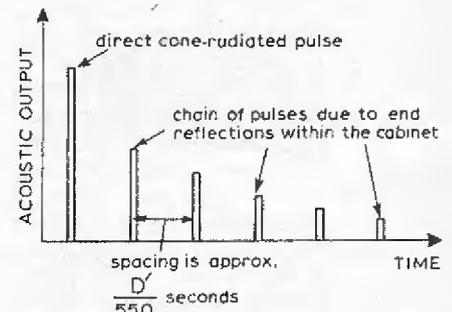


Fig. 3. Pulse-response (idealized) of cabinet shown in Fig. 2.

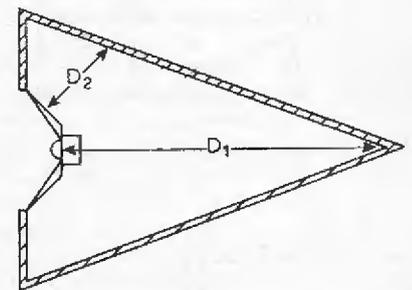


Fig. 4. Triangular cabinet showing different reflecting path lengths.

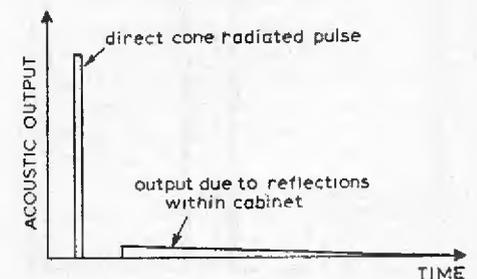


Fig. 5. Approximated output for triangular cabinet excited with impulse from loudspeaker.

* University of Bradford

†Bailey, A. R., 'Non-resonant Loudspeaker Enclosure', *Wireless World*, Oct. 1965.

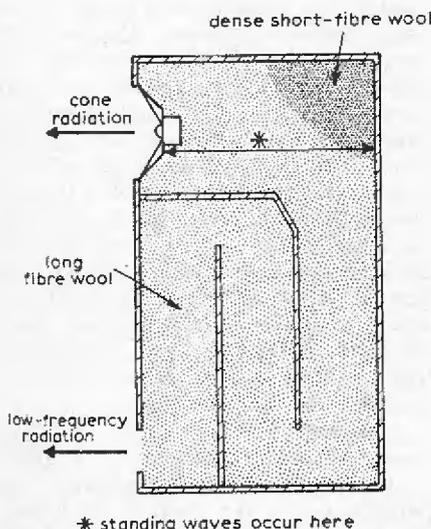


Fig. 1. Original transmission-line cabinet design

will not emerge as a unit impulse but as a much lower amplitude long pulse — more like a continuous low level sound. This is subjectively far less noticeable to the ear than a series of decaying impulses. The two effects are compared in Fig. 5.

Obviously, it would be far better to absorb the sound completely on the back wall of the cabinet. Here we face the impossible, but it is imperative to make the best use of absorbent within the cabinet.

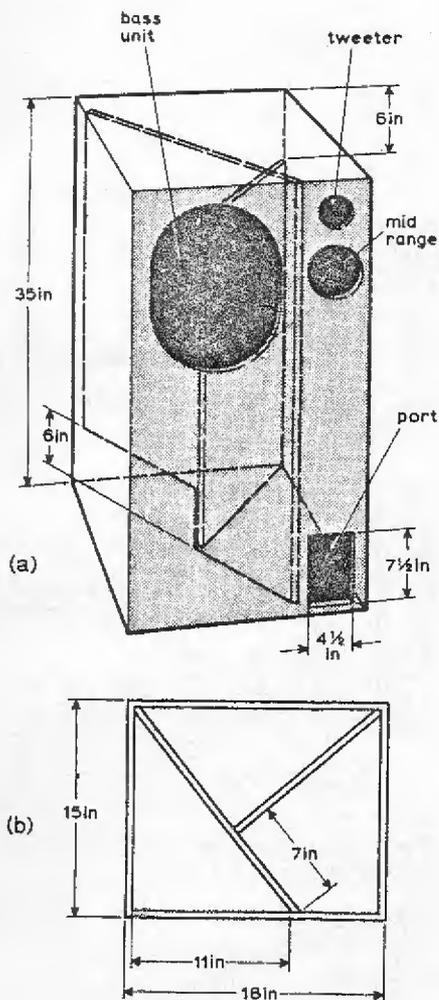


Fig. 6. Final design.

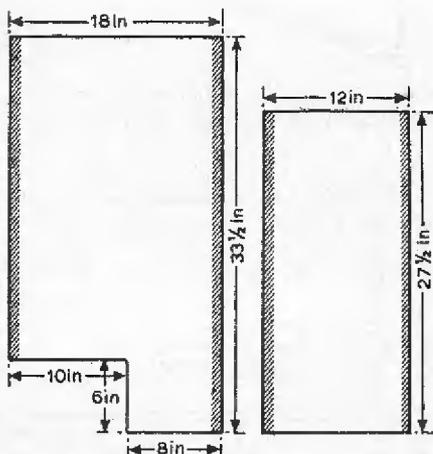


Fig. 7. Dimensions of internal partitions.

Towards a solution

The basic requirement in using a forward facing loudspeaker unit is to guide the sound into the vertical direction without producing bad reflections. In addition the system must be fairly simple, to keep woodworking costs low, and also mechanically strong, to avoid significant panel resonances.

After experimenting with different cabinet shapes, the arrangement shown in Fig. 6 was arrived at. This has several advantages over the original design. First, the pipe is triangular in cross-section, thus giving less audible coloration due to reflections. Secondly, the woodwork is very simple; only two internal partitions are necessary. Thirdly, the front of the cabinet and the large partition are automatically braced so reducing panel resonance. Finally, sufficient area is available on the front of the cabinet to mount a mid-range unit in part of the line remote from the back of the bass driver when internal pressures are not too high. In practice, it has not been found necessary to use a separate enclosure for the mid-range unit.*

Details of the partition sizes are shown in Fig. 7. A three-speaker system is recommended, as there is at present a difficult 'gap' between known low-coloration bass units and tweeters.

Incidentally, it should be noted that many 'high-fidelity' drive units are that in name only. Frequency-response is only one aspect of performance, and the transient response is far more important overall. Pulse or step testing loudspeakers in a long matched acoustic transmission line is most illuminating. Some units will still be radiating appreciable acoustic power 50ms after the exciting pulse has disappeared!

The units specified are known to have good transient response and are available with a suitable crossover network. Such networks are very difficult to design and it is not sufficient to use a general-purpose crossover unit. Unfortunately, loudspeakers do not behave as pure resistance at all frequencies — often quite the contrary. Design of crossovers from this assumption is completely incorrect and it is not uncommon for correct inductor sizes in a network to be double that expected from simple theory. In addition the different phase-angles of speakers at the crossover frequency complicates matters even more, and bad design can lead to abnormally low impedance levels over some parts of the frequency range. In short, crossover networks *must* be designed to operate with the speaker units that they are to be used with or very peculiar results can be obtained.

As previously, long-fibre wool is recommended as the continuous acoustic absorbent that fills the whole of the transmission line pipe. The wool must be well teased out or it loses its effective-

*However, there may be audible improvement if an absorbent filled enclosure is used behind the mid-range unit. ED.

Recommended drive units and crossover networks

Bass driver	BD 25/8	£24.50 + VAT
Mid-range unit	MD 6	£16.50 + VAT
Tweeter	TD 3	£ 7.50 + VAT
Crossover network	FN 10	£11.00 + VAT
Radford Audio Ltd., Bristol BSA3 2HZ.		

Alternative units

Bass driver	B 139	£19.50 + VAT
Mid-range	B 110	£ 8.25 + VAT
Tweeter	T 27	£ 6.00 + VAT
Crossover unit	DN 12	£ 7.00 + VAT
K.E.F. Electronics Ltd., Tovil, Maidstone, Kent. (The performance with the K.E.F. units is improved if the Radford crossover is used in place of the K.E.F. crossover.) Long-fibre wool may be obtained from John W. Pennington (Dowley Gap) Ltd., Midland Wool Warehouses, Briggate, Windhill, Shipley, Yorks. The cost is £1.35/lb including VAT and postage.		

ness. Anchoring the wool is something of a problem as it can compact with transport or use over a period. Nails or dowels projecting from the partitions will serve, but make stuffing difficult. The best suggestion yet made is to use a 'Netlon' core for the wool, fibres being teased through it and left sticking out all round. Using front-mounting loudspeakers as specified, the two front pipes can be loaded through the speaker and port apertures, and the rear pipe is easily filled by removing the back of the cabinet. Alternatively, if the cabinet top is made removable all three pipes can be loaded from the top. A packing density of about 1/2lb per cubic foot is about right. Excess wool will cause back-pressures on the cone, and too little will cause pipe resonances in the low bass region.

What column resonance remains in the system can be reduced by putting 45-degree corner reflectors at the back of the speaker and also at each side of the first bend. These are not critical but should be so arranged that sound from the back of the cone will 'bounce' down the first pipe and then up the second, i.e. consider the sound to be light and the reflectors to be mirrors. The improvement is only about 1dB in frequency response when using reflectors, and as this improvement is only just detectable it may well be decided to omit them. If included, they should be made from 3/4in chipboard or some similar material, and firmly fixed to the cabinet.

The port area is not critical as there is none of the tuning effect that occurs in the base reflex enclosure. Changes in port area of two-to-one ratio produce no noticeable effect, but nevertheless it would be unwise to make the port much smaller than that given in the drawings as this is already considerably smaller in area than the pipe feeding it.

As mentioned previously, response curves should be treated with extreme caution as they represent only part of the performance of the speaker. Nevertheless the overall response curve should be as flat as is reasonably possible. The curve for the complete system when mea-

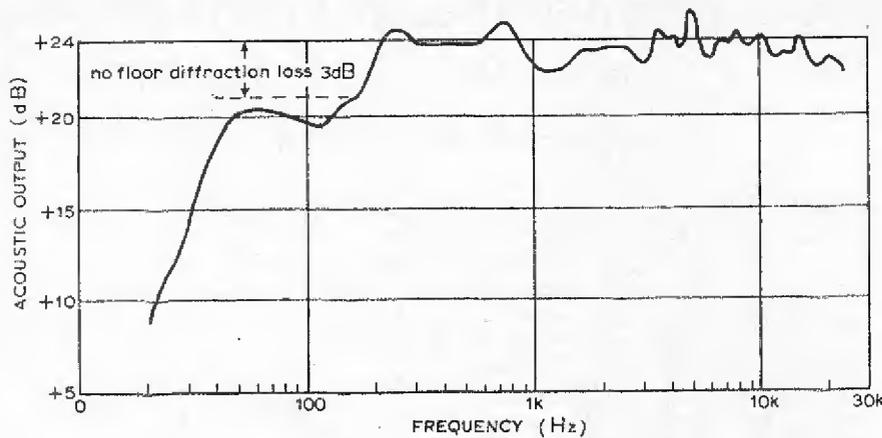


Fig. 8. Anechoic response curve of complete loudspeaker system.

sured in an anechoic chamber is shown in Fig. 8.

In looking at this curve several points must be noted. First, the falling bass response shown will not occur in the same manner in a room. The presence of a floor will give a 3dB lift due to the absence of diffraction in the downward direction. Similarly, walls and ceilings add to the on-axis bass output. In fact a 'flat' response in an anechoic chamber will sound very bass heavy when the system is used in a normal room. Under normal room conditions the bass output of the system described is quite adequate, windows being easily rattled at 30Hz.

The smoothness of the overall curve is the most critical point, the odd decibel of gain or loss in overall response being far less important than a response with a smooth envelope. A jagged response curve means high Q components in the output and these are very noticeable on test. In fact a high Q resonance that lifts the overall response by only 0.2dB may ruin the reproduction of an otherwise excellent speaker. Transient response testing is the ideal answer, but interpretation of the results is very difficult at present except on a rather empirical basis.

The ultimate test is the ear, but one must always remember that personal prejudice can enter into things to a very large extent. For this reason the best test material is not music but such things as pure sine waves (for distortion) and applause, or better still white and pink noise (for transient response).

Regarding the use of pink noise, which incidentally is only white noise attenuated by 3dB per octave with increasing frequency, the following incident happened to the author. A pink noise generator had been built and was being tested with a speaker system. A most noticeable hum was produced and all attempts to find the source in the generator failed. It was finally discovered that the hum was not present in the noise, but was the fundamental resonance present in the speaker system. This speaker (which was of the unlagged reflex type) had previously been used in music tests, and several people had commented on the

good performance, particularly in the bass region.

If the cabinet size is felt to be rather too large, then it is possible to scale all the dimensions shown according to the diameter of the bass driver being used. This will result in a poorer bass performance and is not advised, but a very creditable performance is possible using an 8in driver unit and scaling all the dimensions by four fifths. This factor results from the recommended bass driver having the same effective cone area as a normal 10in unit of circular construction.

In conclusion it must be emphasized that only the system as described is in any way guaranteed. Readers can experiment, of course, but must be prepared to solve their own problems.

Letters to the Editor

Transmission-line speakers

Dr Bailey's transmission-line speaker design in the May issue of *Wireless World* was very interesting. Just before reading it I had been studying his previous design in the October 1965 issue and wondering whether it would be possible to reduce the chance of standing waves inside the enclosure by using angled panels to give triangular cross sections, when, lo and behold, Dr Bailey had done it for me. I was all set to build the new enclosure when I happened to notice the comparison between the published frequency response curves for the new design, the 1965 design, and the KEF Concord totally-enclosed design. For frequencies below 1000Hz the original design response is flat within +0.5-2.0dB down to 25Hz (excellent), the new improved design is flat within +2.0-11.0dB over the same range (not so good). More interesting is the comparison between the new design and KEF's totally-enclosed cabinet as published in their enclosure design leaflet. The two curves are virtually identical being within 2.0dB of each other over

nearly all the range from 1kHz to 20 Hz. One would expect the new transmission-line response to be similar to the original, and yet it is much more like the KEF cabinet which is one-third the volume and based on a different principle.

Sound reproduction is still more art than science, and I feel Dr Bailey's comments on this apparent contradiction would be illuminating and also helpful to people like me.

P. A. Sheppard,
Chandlers Ford,
Hants.

The author replies:

The original transmission-line cabinet was tested under free-field conditions but with the presence of a floor and a reflecting wall. This will give at least 6dB of boost compared with the truly anechoic response curves which were published for the second design. This boost at the low frequencies is due to the prevention of diffraction which otherwise occurs due to the absence of floor, walls or ceiling. In other words, the anechoic response is only a guide to performance, and a loudspeaker that gives a truly flat response under anechoic conditions will sound very bass heavy when used in a normal room.

Frequency response curves are only a guide to performance, and certainly one cannot tell from an examination of response curves alone the respected merits of different loudspeaker systems. In a system using several drive units it is possible to obtain widely differing response curves merely by changing the microphone position being used to measure the loudspeaker characteristics. It is by no means unusual to find a response curve that is sensibly flat on the axis of the tweeter turn into a curve with dips of 10dB or more when measuring the axis of the mid-range or bass unit. In addition, no anechoic chamber is perfect and one can obtain considerable variations in response curves when using the same test object in different anechoic chambers. There is at least one known instance of a speaker being deliberately angled to a very peculiar position in an "anechoic" chamber, because that was the position where a flat response curve was obtained. We are still in a position of not being able to evaluate loudspeakers adequately using axial response curves, and until superior measurement techniques have been evolved, by far the best test is that of careful listening and switching between the various units being evaluated.

I have not tested the KEF Concord that Mr Sheppard mentions, but I feel that the foregoing may partly answer his queries. In addition there are such aspects as power handling capacity and frequency doubling to be taken into account, so I would hesitate to make a definite pronouncement except that in general, large cabinets perform better than small ones.

A. R. Bailey

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Commercial quadraphonic systems

An up-date of two 1972 WW articles

by Geoffrey Shorter Technical Editor Wireless World

Quadraphonic systems, here interpreted as attempting to convey sound direction with four loudspeakers using either two, three or four transmission channels, use matrix circuits to suitably organise the information into the required number of channels.

The approach in the January and February 1972 *WW* articles applies to the practice in recording studios of preparing four-track mix-downs for quadraphonic use. It does not represent the only way of achieving surround sound within a finite number of channels; but the alternate microphone mixing techniques are a comparative rarity. The justification for using such tapes is a commercial one: record companies, having built up stocks of multi-track tapes over the years, will obviously want to remix them into a four-track format and use them with quadraphonic playback in mind.

What this multi-track approach will never do of course, is to capture the "ambience" or reverberant qualities of the live acoustic scene. This ideally requires a microphone technique that will take sounds from all directions in such a way they can be played back with the apparent reproduced directions corresponding with the original directions. This is the aim behind the "ambisonic" proposal of Peter Felgett for instance, whose concepts of course, do have relevance to the multi-microphone, pair-wise mixed master tapes. As Felgett has argued¹, any worthwhile technique that can adequately cope with sounds from any direction should be able to handle these signal sources as well.

Matrixing, in connection with quadraphonic

systems, is one way of combining separate ("discrete") audio channels into a lesser number so that channels are kept sufficiently separate to be retrievable by a suitable "dematrixing" process, albeit with large amounts of crosstalk from other channels.

Matrix circuits are frequently used in situations where information taken from separate signals is required to be re-arranged for some reason (typically to make signal processing easier) without loss of information.

In stereo, for example, left and right-channel signals can be used to form sum and difference signals that are equivalent in that no loss of information has occurred. Circuits that achieve such re-organisation are called 2-2 matrix circuits. Another example of a similar matrix circuit occurs in colour television, where colour subcarriers provide two difference signals, $R-Y$ and $B-Y$, which are fed to a resistive matrix to provide $G-Y$. This is possible because $Y = 0.3R + 0.59G + 0.11B$. (The three colour difference signals are then fed directly to the c.r.t. grids, with a $-Y$ signal at the cathode. Alternatively, $+Y$ is added to the three colour difference signals to provide R, G, B .)

In the mathematical sense a matrix is merely an array of rows and columns of numbers, but it is more in an electrical sense that it is applied in this context. One can think of an electrical matrix as an array of intersections between input and output lines, between some of which may be connected linear or non-linear elements. The result of this arrangement is a coding of the input onto the outputs. A binary-to-decimal coder or decoder is a good example of such an electrical matrix.

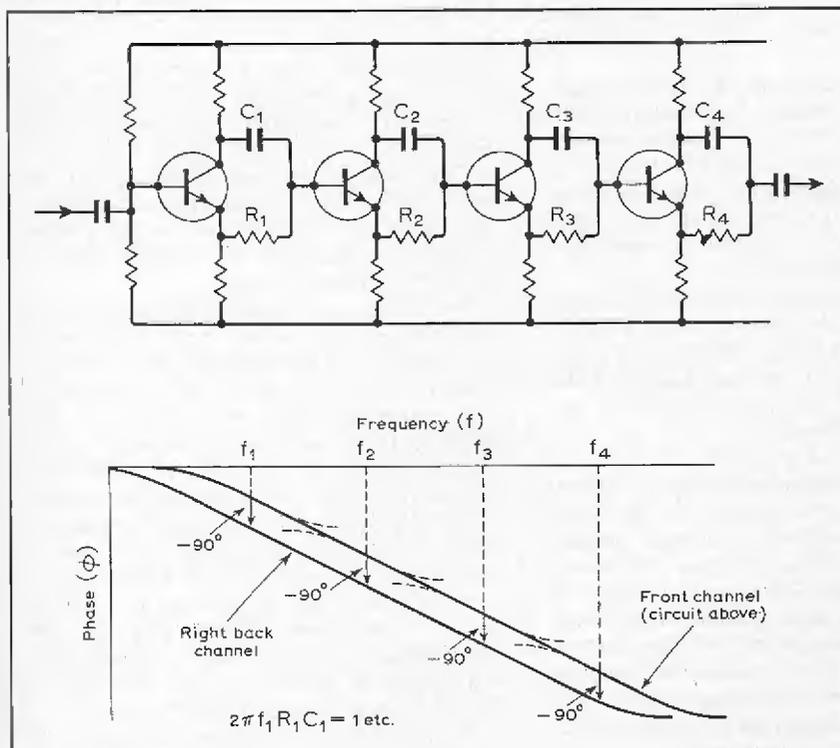
In a simple amplitude matrix the gains between inputs and outputs can be set; in a phase matrix it would be a phase difference between a pair of lines that can be set. Considering first an amplitude matrix, one can visualise a general linear resistive network where four input signals, say A, B, C, D , are fed into two outputs, say L and R , with A contributing an amount a to L and an amount a' to R , and similarly for B, C and D (like that of Fig. 2 in ref. 2). In retrieving the original inputs from the two channels L and R , outputs are derived by mixing different amounts of L and R , a special case being when one takes an amount a of L and a' of R to form one of the four outputs. Thus

$$\begin{aligned} aA + bB + cC + dD &= L \\ a'A + b'B + c'C + d'D &= R \end{aligned}$$

and the four outputs would be

$$\begin{aligned} A' &= aL + a'R \\ B' &= bL + b'R \\ C' &= cL + c'R \\ D' &= dL + d'R \end{aligned}$$

Fig. 1. A 90° phase difference over a wide band between two channels can be realised using two sets of cascaded phase shift circuits of the kind shown. The difference in phase between the two cascaded circuits is made 90° . In practice the phase-difference/frequency response is rippled but the amount of phase ripple can be kept within prescribed limits.



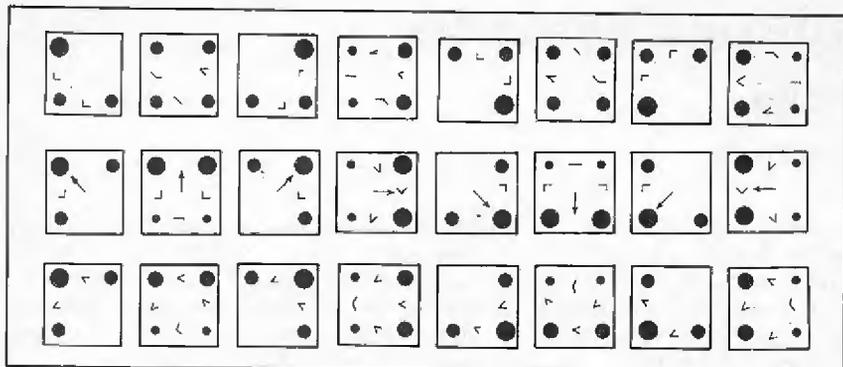


Fig. 2. Speaker outputs for the commercial two-channel matrix systems for eight intended source directions (see arrow in middle). Blob size indicates speaker output and angles represent phase between speakers. Top is the basic SQ matrix, middle is the QS matrix and bottom is the BMX matrix. There are also differences in mono and stereo compatibility of the three systems, see Table

Substituting the equations for L and R in the above gives (for the first output only)

$$\begin{aligned} A' &= a^2A + abB + acC + adD \\ &+ a'^2A + a'b'B + a'c'C + a'd'D \\ &= A(a^2 + a'^2) + B(ab + a'b') + \\ &C(ac + a'c') + D(ad + a'd'). \end{aligned}$$

The first thing one notices is that for an input to this network, or matrix, of A only, the output A' has components from the B , C , and D inputs, i.e. crosstalk. Matrixing works because one can make $(a^2 + a'^2) > ab + a'b'$, for example. It can be shown that if the D contribution is made zero by suitable choice of coefficients, the other contributions cannot be smaller than -3dB .

The trouble with this simple sort of amplitude matrix is the anti-phase component that is always present in four-speaker playback which upsets the apple cart as far as accurate portrayal of direction is concerned (ref. 2). The way round the problem is to use 90° phase difference circuits (two circuits of the kind shown in Fig. 1 can be used) and the three current quadrasonic matrix systems use such circuits to distribute the 180° phase difference in one of three ways (see next section).

Commercial systems

For adequate representation of sound field direction it can be shown that for frequencies where head dimensions are small in relation to sound wavelength, three transmission channels will do; but it is possible to do it with two provided a directional ambiguity is resolved by suitably organising a 180° phase difference. There have been in-between proposals, like that of R. Berkovitz* in which it is proposed to transmit ambiguity-resolving information in the form of a "key" signal, sent with one of the two channels at a high or inaudible frequency.

But current commercial two-channel systems (SQ, QS, BMX) all effectively divide the 180° of phase difference in different ways, the newcomer to the market place being BMX which distributes the phase in such a way as to be least bothersome.

Specifications and main characteristics of the three commercially-produced two-channel matrix systems are compared in the Table and in Fig. 2 which indicates speaker amplitudes and phase difference for sounds intended for eight different directions. It shows clearly the large amounts of crosstalk in all the systems—a fact of life in two-channel systems—the differences between them being one of organising the crosstalk-component amplitudes and phases to best

effect. Taking the basic systems without any embellishment, it would seem that the SQ matrix does the poorest job, while the BMX matrix provides the most clues as to source direction.

The SQ encoding system concentrates on minimising front left-right crosstalk at the expense of mono compatibility and directional anomalies in quadrasonic playback. Image positions around two opposite corners are compressed and centre front-to-back crosstalk is severe (though much improved with the "10%-40%" blend decoder, see Table on p. 86).

The QS/RM system gives speaker amplitudes as shown for BMX in Fig. 2, but with phases that do not seem to be as helpful as in BMX. Mono compatibility of QS records is poor (see Table) and in stereo there is a crosstalk of 7.7dB , narrowing the stereo spread, and which is accompanied by back sounds appearing to come from beyond the two speakers. (Mono compatibility is improved for broadcasting by phase shifting one channel with respect to the other by 90° .)

BMX achieves its performance—about the best that can be expected from a 4-2-4 system in the quadrasonic and mono modes—at the expense of a 90° phase difference between channels in stereo.

Variomatrix. Methods of circumventing the various defects rely either on using variable-gain or blend circuits to suppress crosstalk components in the four-speaker mode or in adding further audio channels. Circuits and operation of variable gain or blend for the SQ system have been published in *Wireless World* before, but at the time of going to press details of the latest developments in SQ decoders were still not available.

The Sansui Variomatrix technique is based on a phenomenon of "directional masking" in which it is argued that the direction of a sound is less easy to perceive in the presence of a more dominant sound, following within 10ms. Sansui say that effects equivalent to those of a "discrete" system are obtainable from a matrix system that reproduces the dominant sound with utmost directionality and the lesser sounds with less precise directionality (within a time span of 10ms), carefully and openly adding the rider, "provided such obscuring of directionality is within limits dictated by the psychoacoustics of four-channel listening which is yet to be studied seriously".

So Sansui use circuits additional to the basic QS matrix to detect the direction of the dominant sound (in level) and derive suitable control signals to alter the matrix parameters and reduce crosstalk components for that direction (within maximum periods of 10ms). The reduction of crosstalk is done in a symmetrical way, both crosstalk components being altered equally, to avoid changes in image position. At the same time crosstalk components of weaker sounds are increased.

While there are similarities in the way control signals are derived with the gain control techniques applied to SQ decoders, the method is different in that weak sounds are not so severely mislocated.

Phase detector circuits determine the intended location of a sound image. Front-

* U.S. patent 3,718, 773, filed May 1970.

Two-channel quadrasonic matrix systems for four-track master tapes

Code	Decode	Four-speaker crosstalk (dB)	Stereo*	Mono
SCHEIBER original system				
$\begin{matrix} L_F \\ R_F \\ L_B \\ R_B \end{matrix} \begin{bmatrix} 0.924 & 0.383 & 0.924 & -0.383 \\ 0.383 & 0.924 & -0.383 & 0.924 \\ 0.924 & -0.383 & 0.924 & -0.383 \\ -0.383 & 0.924 & -0.383 & 0.924 \end{bmatrix}$	$\begin{matrix} L & R \\ L_F & R_F \\ L_B & R_B \end{matrix} \begin{bmatrix} 0.924 & 0.383 \\ 0.383 & 0.924 \\ 0.924 & -0.383 \\ -0.383 & 0.924 \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.71 & 0.71 & 0.00 \\ 0.71 & 1.00 & 0.00 & 0.71 \\ 0.71 & 0.00 & 1.00 & -0.71 \\ 0.00 & 0.71 & -0.71 & 1.00 \end{bmatrix}$		<p>Front crosstalk 7.7dB</p> <p>Back -7.7dB CdB with phase shift mod.</p> <p>$1.31L_F + 1.31R_F$ $+ 0.54L_B + 0.54R_B$ Centre back suppressed</p>
SANSUI QS				
$\begin{bmatrix} 0.924 & 0.383 & j0.924 & j0.383 \\ 0.383 & 0.924 & -j0.383 & -j0.924 \end{bmatrix}$	$\begin{bmatrix} 0.924 & 0.383 \\ 0.383 & 0.924 \\ -j0.92 & j0.38 \\ -j0.38 & j0.92 \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.71 & j0.71 & 0.00 \\ 0.71 & 1.00 & 0.00 & -j0.71 \\ -j0.71 & 0.00 & 1.00 & 0.71 \\ 0 & j0.71 & 0.71 & 1.00 \end{bmatrix}$		<p>Front crosstalk 7.7dB</p> <p>Back -7.7dB Centre back suppressed</p> <p>$1.31L_F + 1.31R_F$ $+ j0.54L_B - j0.54R_B$ Centre back suppressed</p>
CBS SQ				
$\begin{bmatrix} 1.00 & 0.00 & -j0.71 & 0.71 \\ 0.00 & 1.00 & 0.71 & j0.71 \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.00 \\ 0.00 & 1.00 \\ j0.71 & -0.71 \\ 0.71 & -j0.71 \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.00 & -j0.71 & 0.71 \\ 0.00 & 1.00 & 0.71 & j0.71 \\ j0.71 & -0.71 & 1.00 & 0.00 \\ 0.71 & -j0.71 & 0.00 & 1.00 \end{bmatrix}$		<p>Front crosstalk none</p> <p>Centre back signal -suppressed Centre side diff. 7.7dB</p> <p>$L_F + R_F$ $+ L_B(-135^\circ) + R_B(+45^\circ)$ Centre back signal -suppressed Centre side diff. 7.7dB</p>
10%-40% blend				
	$\begin{bmatrix} 1.00 & 0.10 \\ 0.10 & 1.00 \\ 0.71(0.4+j) & -0.71(1+0.4j) \\ 0.71(1+0.4j) & -0.71(0.4+j) \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.10 \\ 0.10 & 1.00 \\ -0.71(1+0.4j) & -0.71(0.4+j) \\ -0.71(0.4+j) & -0.71(1+0.4j) \end{bmatrix}$		<p>Centre front to back crosstalk 7dB</p> <p>As above</p> <p>As above</p>
EVX-44				
	$\begin{bmatrix} 1.00 & 0.20 \\ 0.20 & 1.00 \\ 0.63(0.4+j) & -0.63(1+0.4j) \\ 0.63(1+0.4j) & -0.63(0.4+j) \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.20 \\ 0.20 & 1.00 \\ -0.63(1+0.4j) & -0.63(0.4+j) \\ -0.63(0.4+j) & -0.63(1+0.4j) \end{bmatrix}$		<p>Similar to CBS SQ 10-40 blend decoder, as above</p>
BMX				
$\begin{bmatrix} 0.924 & 0.383 & 0.924 & 0.383 \\ \angle 22.5 & \angle 67.5 & \angle -22.5 & \angle -67.5 \\ 0.383 & 0.924 & 0.383 & 0.924 \\ \angle -67.5 & \angle -22.5 & \angle 67.5 & \angle 22.5 \end{bmatrix}$	$\begin{bmatrix} 0.924 & 0.383 \\ \angle 22.5 & \angle 67.5 \\ 0.383 & 0.924 \\ \angle -67.5 & \angle 22.5 \end{bmatrix}$	$\begin{bmatrix} 1.00 & 0.71 & 0.71 & 0.00 \\ \angle 00 & \angle 45 & \angle -45 & \angle 00 \\ 0.71 & 1.00 & 0.00 & 0.71 \\ \angle -45 & \angle 00 & \angle 45 & \angle 00 \\ 0.71 & 0.00 & 1.00 & 0.71 \\ \angle 45 & \angle 00 & \angle -45 & \angle 00 \\ 0.383 & 0.924 \\ 0.00 & 0.71 & 0.71 & 1.00 \\ \angle -45 & \angle -45 & \angle 45 & \angle 00 \end{bmatrix}$		<p>Front crosstalk 7.7dB $\angle 90^\circ$ $\equiv 15.3dB \angle 0^\circ$</p> <p>Back crosstalk 7.7dB $\angle -90^\circ$ $\equiv 15.3dB \angle 0^\circ$</p> <p>$1.00L_F + 1.00R_F$ $+ 1.00L_B + 1.00R_B$ No points suppressed</p>

* Equations as under 'code'

Table gives data for basic two-channel quadrasonic systems. Data for gain-control systems for SQ, variable blend system for SQ, and the 4-4-4 QMX brother of BMX are not included.

to-back sound source is detected by comparing the phase of the two transmission channels, *L* and *R*, while left-to-right sound positions are determined by comparing the left-channel signal and the sum of the right signal, phase shifted by 45°, with that of the difference between these two signals.

The Variomatrix circuits can be applied to SQ encodings too, with the result that centre front-to-back separation is improved from 0dB to at least 6dB. It can also produce four-speaker playback of stereo records.

Phase-encoded matrix

The other basic way of representing direction of a sound using two channels is by carrying an omnidirectional signal in one channel and another omnidirectional signal in the other channel but phase shifted by an amount that depends on the direction of the sound. The omnidirectional signal, which is the sum of all sound sources, while being by definition an ideal mono signal is not in itself in a convenient form for stereo transmission—neither is the phase-encoded signal—so they can be sum-and-difference matrixed, the new sum

forming a left transmission signal and the difference forming the right signal. A mono pickup would then respond by producing the original omnidirectional signal.

Using the expression $\exp j\theta$ to represent phase angle, to be made the same as the bearing angle, the left and right signals become $1 + \exp(-j\theta)$ and $1 - \exp(-j\theta)$ in the case of sounds encoded to the optimal BMX matrix of Cooper and Shiga. (In these expressions, unity represents the omni or directionless signal.) In decoding, the two are sum-and-differenced matrixed, providing a mono or omni signal and a phase-encoded signal, the last-mentioned of which is phase shifted by an amount ϕ° say, a speaker bearing angle. This is then added to the omni signal to provide a speaker signal.

For a speaker that is at the same angular location as the source the speaker output is a maximum, for an opposing speaker the output is zero; and for speakers at $\pm 90^\circ$ from the source the outputs will be $-3dB$ and phase shifted by $\pm 45^\circ$. The maximum phase difference of 180° that occurs between speakers in the Scheiber system and which is reduced

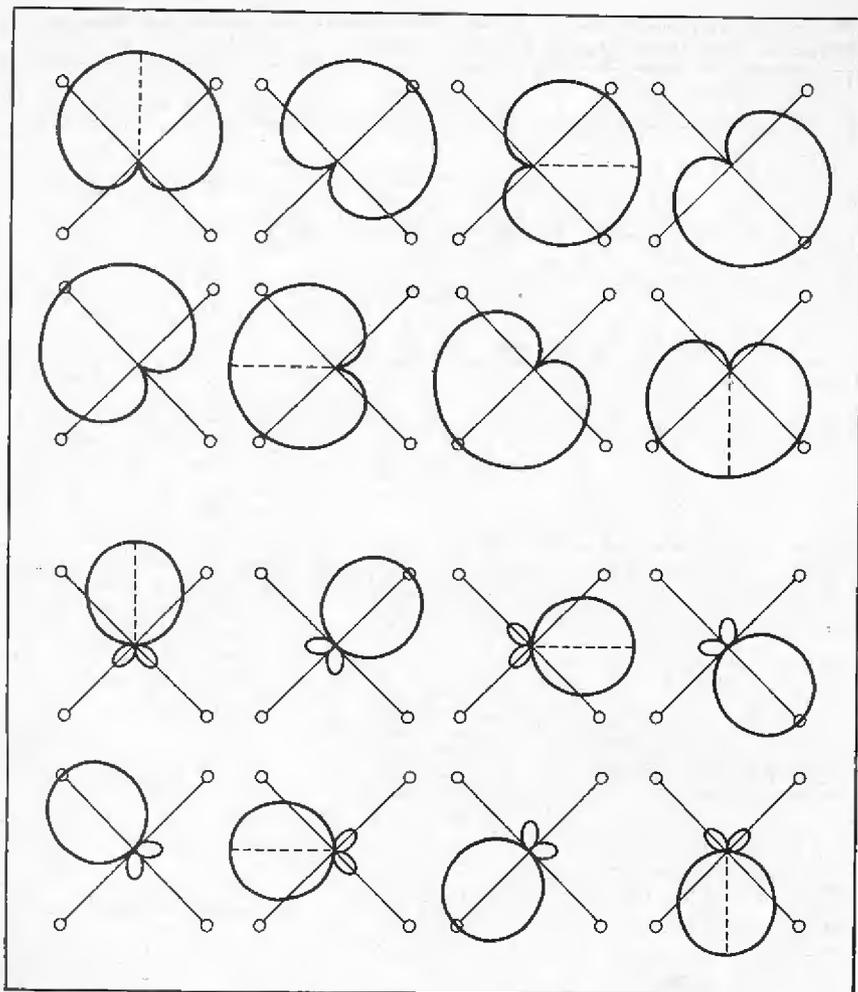


Fig. 3. Directivity patterns obtainable with BMX (top) and QMX (bottom) two- and four-channel systems. Speaker amplitudes are obtained from the intersection of the polar curve with the lines joining the four corners. In the new QMX carrier-channel disc, known as UD-4, the improved directivity of the four channel matrix is obtained after demodulating carrier channels.

to 90° in the Sansui system, is now down to 45° (see Fig. 2).

Signals for the four inputs from a four-channel master tape will appear as shown in the Table. The coefficients, which show similarities with the Scheiber and Sansui QS coefficients, can be seen to result from combining by addition and subtraction the omni signal, $L_F + R_F + L_B + R_B$ with the phase-encoded signal $L_F \exp -j135^\circ + R_F \exp -j45^\circ + \dots$

The inventors of this particular form of phase-encoded matrix approached the design from a viewpoint of bearing-angle harmonic synthesis³, in which directional functions are accurately represented using a limited number of channels and optimised by a statistical technique. The workings of Cooper and Shiga show that as well as the BMX two-channel system there exists an hierarchy of systems, of three, four, five, etc. channels, each being optimum in conveying directionality for that given number of channels and with the virtue that channels can be added or taken away while leaving the system still optimum.

The third and fourth channels turn out to be phase-encoded omni signals, but with the opposite of, and double, the BMX phase angles respectively (effectively the first two, three or four terms of a bearing-angle Fourier series). This four-channel matrix system, called QMX, is the basis of the new discrete system, UD-4 (see next section).

There are many other matrix systems that have been proposed or discussed, like that developed at the BBC Research Department,

the three and four-channel systems listed by Eargle⁴, the tetrahedral system first pointed out by Gerzon, and the periphonic (including height) ones listed by Gerzon⁵ who conceived of an hierarchy of systems using "spin" and spherical harmonics.

None of these have been taken up commercially, except for the Cooper/Shiga hierarchy of horizontal-only systems now marketed by Nippon Columbia in the QMX format as UD-4, which could be considered to be a special case of the Gerzon periphonic family. (Though horizontal in concept the system does lend itself to the addition of height information.)

Discrete systems

Assuming, as almost everyone does, that surround-sound in its quadraphonic form is here to stay, the question of how to get four separate channels of audio from a compatible disc record was bound to raise its head, whichever way those four channels are used. JVC with their CD-4 record system have gone a long way to achieving this end, but there still remain h.f. distortion problems and sensitivity to various inequalities.

Apart from possible mistracking of the complementary noise suppression circuits, there can be differences in phase and amplitude of the baseband and carrier signals, leading to crosstalk and consequent direct image displacements. For instance, with a carrier modulation level greater than baseband level, side images are shifted toward the speakers away from centre and centre-side images become even more unstable than they already are. With a lesser carrier-channel level the side images shift the other way and centre front and back images become less stable.

There can also be h.f. baseband losses at the inner point of the groove spiral, h.f. losses due to fall off in pickup response and crosstalk distortion products due to tracking and tracing difficulties, causing phase modulation of the carrier. While "downtalk" may be negligible, "uptalk" is not because of errors inherent in tracing and tracking.

Such sources of crosstalk are minimised in the QMX carrier-channel system of Cooper, Shiga and Takagi. In the UD-4 disc record system that is based on QMX, the baseband channels are encoded to the BMX matrix, and speaker outputs for this are shown in Fig. 3 (top). When the BMX channels are combined linearly with the carrier-channel matrix, to give the overall QMX matrix, the speaker outputs will be as shown in Fig. 3 (bottom).

The result of inequalities in level and phase between the baseband and carrier modulations is a broadening of the QMX patterns Fig. 3 (bottom) into the BMX patterns (top). This broadening is gradual, continuous and symmetrical, there being no change in rotational or axial symmetry. The result is that there are no direct image shifts, merely an increase in the risk of mislocalisation because of the broadening.

Another important feature of the UD-4 QMX approach is that the carrier-channel bandwidth can be restricted, without affecting overall response, not possible with the CD-4 matrixing, with consequent benefits that accrue from limiting h.f. information in

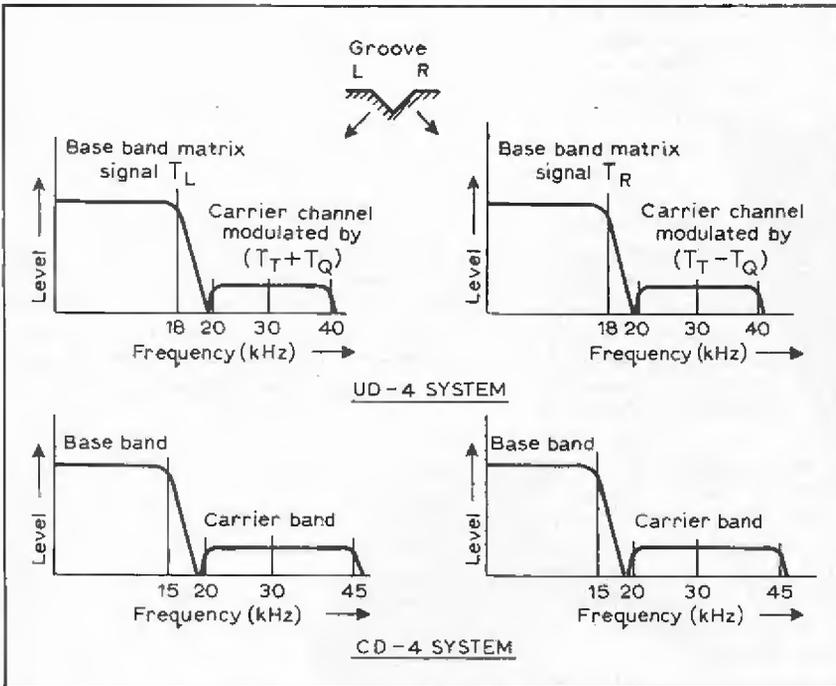


Fig. 4. In the UD-4 system the carrier channels can be narrow-bandwidth without affecting overall response, with a consequent gain in signal-to-noise ratio.

the carrier channels. Localisation tests⁶ on a selected group of young people in an anechoic chamber (with speakers hidden from view) and using spectrally-weighted intermittent noise with bandwidth limited to 130Hz to 3kHz did not show significant degradation in accuracy.

The restriction in bandwidth after demodulation of the carrier channels means that the noise and distortion contribution is reduced. A gain in signal-to-noise ratio of around 10 to 12dB is obtained over a full-band system. In UD-4, the gain is used to dispense with the noise reduction system, reducing complexity, cost and eliminating any errors from mis-tracking.

Actually, recorded bandwidth of the carrier channels is made 6kHz so that when the state-of-the-art allows, 4dB of signal-to-noise ratio could be sacrificed for the extra bandwidth by making the carrier channel bandwidth in playback equipment 6 instead of 3kHz.

Maximum deviation of the 30kHz carrier is limited ± 10 kHz, giving a lower top frequency than on CD-4 records (Fig. 4) with less dependence on the pickup cartridge.

Spherical representation

The simple view adopted earlier in looking at matrix systems is not the only one, though

it is helpful to some people to look at them this way. A more useful approach is the geometric model thought up by Michael Gerzon, who calls it the energy sphere, and Peter Scheiber, who calls it the representative sphere. By conveying a sound source, *S*, say on two channels, *L* say, equal to $S \cos \theta/2$ and the other, *R* say, equal to $S \sin \theta/2$, source direction can be portrayed, defined by $\theta/2 = \tan^{-1} L/R$.

Representing this angle on a horizontal circle is the starting point for introducing phase angle. This corresponds to tilting the circle about a diameter through an angle, say ϕ , the circle ultimately describing a sphere. Thus encoding direction as phase difference between two channels is equivalent to this tilt and if amplitude ratio between channels is unity the phase angle is measured around a vertical circle.

The various ways of coding a sound and its direction onto two channels, i.e. by phase or amplitude coding, or any combination of both, can be defined by the locus of a point on the sphere that makes angles ϕ and θ at the centre with the vertical and horizontal planes.

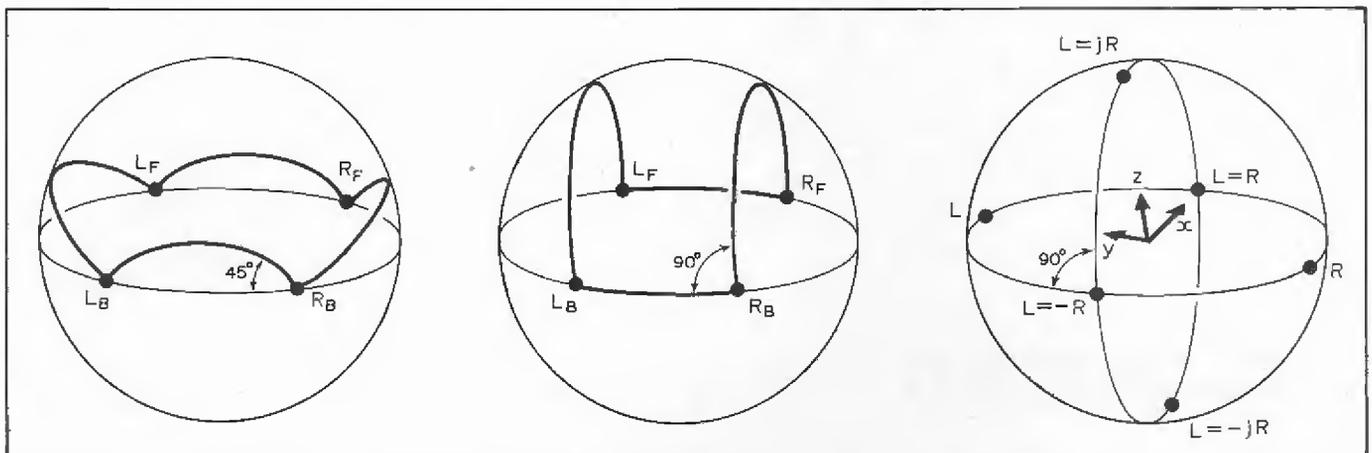
The "regular matrix" is one in which direction is conveyed by amplitude and is defined by a locus that is an horizontal great circle. The locus for the phase-encoded system described earlier is a vertical great circle.

To decode a sound direction, the two channels are added each with a phase and amplitude appropriate to each speaker position. It can be shown that, given a single input, this process leads to outputs the amplitudes of which follow a law governed by a cardioid shape. This cardioid-type directional resolution means that an output from a two-channel system fed to a speaker for which the decoding point on the sphere corresponds to the encoding point will feed a speaker with maximum signal. A decoding point that is diametrically opposite will give zero output (infinite crosstalk), and for points at $\pm 90^\circ$ output will be 3dB down on maximum.

Representing matrix systems on the sphere

The sphere model has geometric properties that relate to matrixing. The transformation on the sphere surface due to matrixing is a conformal one in which angles between curves on the sphere are preserved and in which circles change into other circles on the sphere. The simplest kind occurs with 2-2 matrixing which results in a rotation of the sphere.

Fig. 5. M. Gerzon's energy sphere representation of coding sounds onto two channels. The "pan locus" shown left applies to a pan-potted amplitude-coded system with 45° phase differences between adjacent points—the locus for the phase-encoded BMX system is similar but turned vertically through 90°. Middle diagram shows QS pan-locus, and diagram on right shows points on the sphere corresponding to different phase and amplitude relations between two channels.



The constraint to be met for this to happen is that the sum of the squares of amplitudes coefficients of the L and R channels must be a fixed multiple of the sum of the square for the new matrixed channels. This is clearly met in the case where the L and R signals are matrixed into $L \cos \theta/2 + R \sin \theta/2$ in the left channel and $R \cos \theta/2 - L \sin \theta/2$ in the right, the sphere being turned through θ about the vertical axis. (Incidentally this also corresponds to rotation of the direction of stylus motion in a stereo groove by $\theta/2$.)

The preservation of angles between circles is helpful in visualising pan-potting. In a 4-2-4 matrix system the four inputs are represented by four points on the encoding sphere. In pan-potting between a pair of these inputs, or points, the "pan locus" follows circular arcs on the sphere. If the two inputs, or points, have a phase difference the arc angle at one point is equal to the phase difference and joins the other point at the same angle. If the phase difference is zero, the arc is the shortest great circle; if the difference is 180° the arc is the longest great circle.

The difficulty with early systems like Dynaco/Hafler, ElectroVoice, Scheiber, i.e. of the non-phasor, amplitude-encoded kind—that of a 180° phase difference between rear

channels—shows up as a doubling back of the pan locus, a kind of double horseshoe shape.

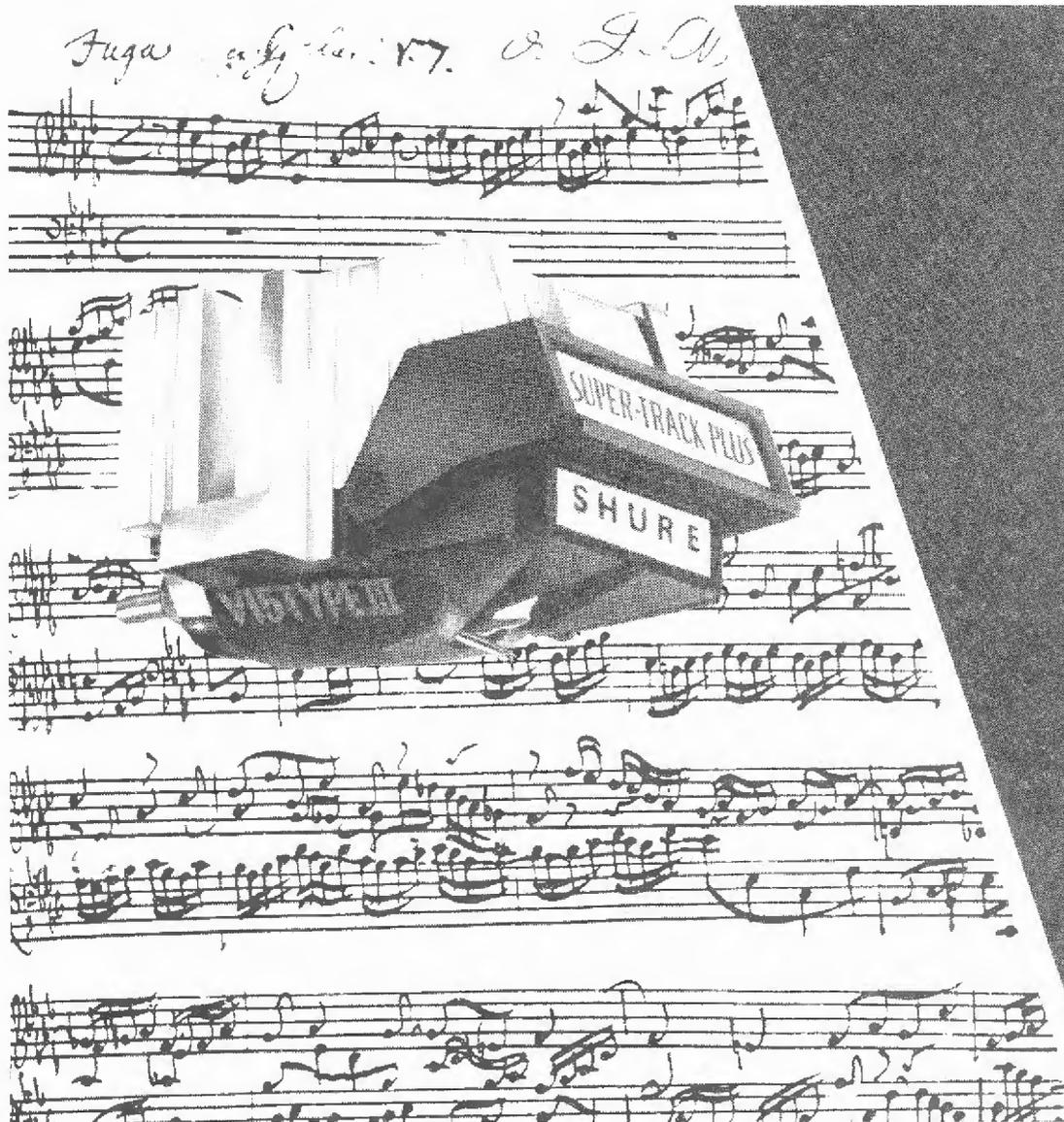
In many ways a great circle locus is best, like the amplitude-coded regular matrix "optimal" specification (an horizontal great circle), the phase-encoded "optimal" system of Cooper and Shiga (a vertical circle) and Gerzons SMQ idea (a great circle tilted at 45°). Any one of such great circle systems can be converted to another by a relative phase shift between channels.

But they require special microphone techniques and panpots that encode directly onto the two channels. For encoding four-track mix-downs compromise raises its head and for two-channel systems, the 180° phase anomaly has to be dealt with.

One way is to distribute this in four lots of 45° suggested by various people around 1971. Gerzon and Eargle thought of doing this to an amplitude matrix (Fig. 5 (left)) while Cooper and Shiga proposed it with a phase-encoded matrix (Fig. 5 (left) with the horizontal axis turned to the vertical). To end this brief look at the energy sphere, Fig. 5 also shows the way Sansui chose to distribute phase in their QS/RM system (middle), and some important points on the sphere (right).

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