Stereo Mixer

A comprehensive range of high-quality input stages with mixing, filtering, and tone-control facilities

by H. P. Walker, B.A.

The stereo mixer to be described, in this and next month's issues, is shown in the block diagram of Fig. 1, and its specifications are given in Table 1. Although the author's equipment was built with five stereo channels, there is in principle no reason why it should not be expanded to many more with a more complex system of mixing and group faders, or simplified to a high-quality pre-amplifier with no mixing facilities. Similarly, the nominal output levels can be altered as required.

Table 1. Input facilities

Source	Max. Sensitivity	Noise	Overload margin	Comments
Magnetic pickup	1.5 mV @ 1kHz	−67.5 dB ref. 1.5mV @ 1kHz	>30dB	Normal R.I.A.A. equalization
Ceramic pickup	15 mV @ 1kHz	−70 dB ref. 15mV @ 1kHz	28dB	Utilizes mechanical equalization, input impedance ~200kΩ
Crystal pickup	70 mV @1kHz	−85 dB ref. 100mV @ 1kHz	>26dB	Economical circuit input impedance = 2MΩ
Microphones	Various depending on type		30dB	Several circuits are described for different requirements
Auxiliary	30mV 100K Ω input impedance	<−70 dB	Infinite	Preset sensitivity control preceding amplifier

Pre-Mixing circuits

For an overload margin of 30dB for the input pre-amplifiers, and a residual noise level of better than 80dB below full output, the nominal signal level at mixing should be 120mV, requiring an output of about 250mV from the pre-mixing circuits. This defines the gain of the input circuits for the basic sensitivities.

The stereo balance and mono/stereo switch, shown in Fig. 2, is a common feature of all the pre-mixing amplifiers. The balance control should be a wire-wound potentiometer to avoid the crosstalk which would result from a high contact resistance at the slider (e.g. with a carbon-track potentiometer). The inclusion of the 4.7kΩ resistors, before the mono/stereo switch, provides a properly mixed version of the two stereo channels for mono operation. If one wishes to mix a monophonic signal stereophonically (e.g. for a point source of sound, movable in a stereo field) the signals must be paralleled at the input or at the presensitivity control.

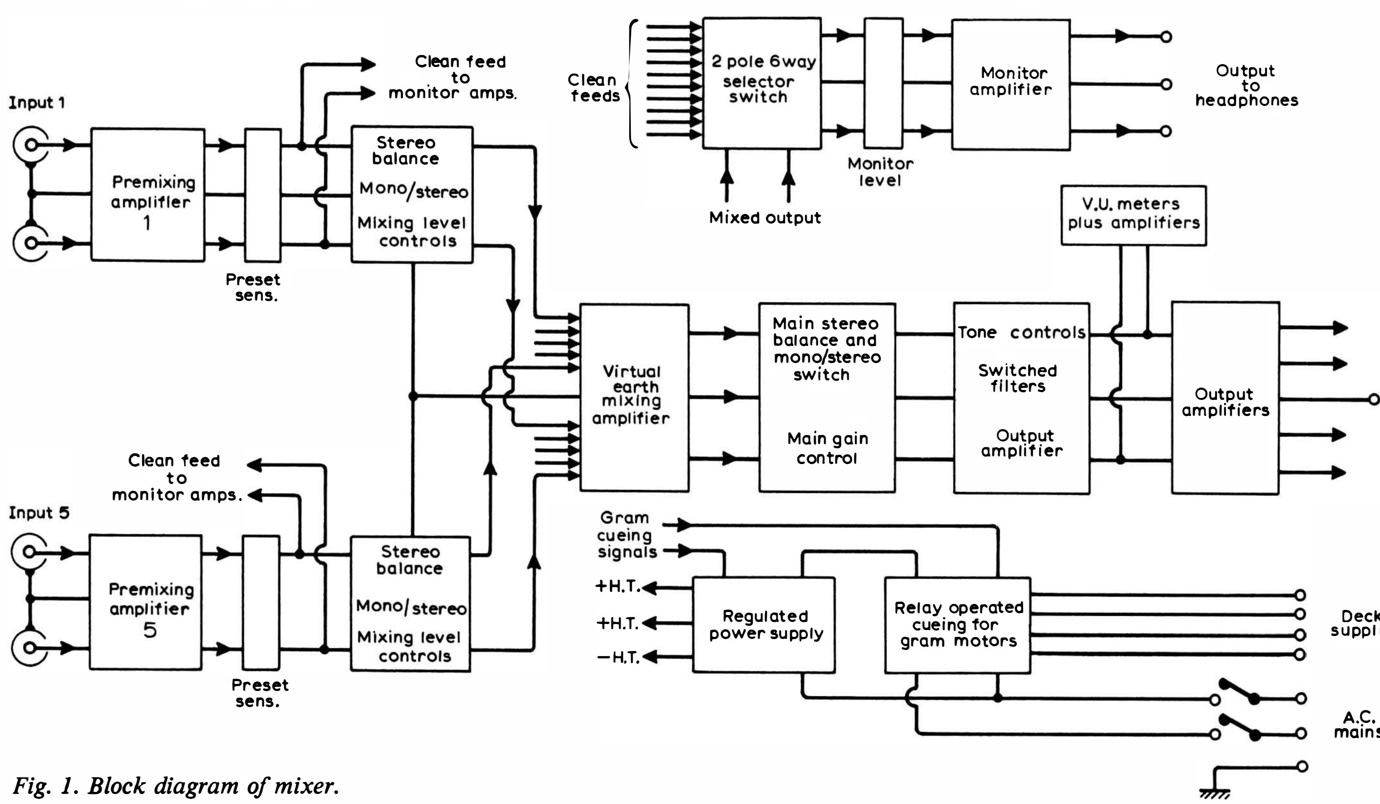


Fig. 1. Block diagram of mixer.

High-quality magnetic and ceramic pickups: While it is possible to design a virtual-earth feedback amplifier suitable for a magnetic pickup, signal-to-noise ratio is inferior to that of the series input of the feedback pair when the source impedance is less than the input resistance. It was required that the input of the mixer be switchable between a microphone and a ceramic pickup so there was a further advantage in using the feedback pair, since input impedance can be changed without altering gain. The final circuit for this amplifier is shown in Fig. 3. and the equalization curve in Fig. 4.

Although a field-effect transistor will give a good noise figure when operated with a ceramic pickup it will give a poor flicker noise performance when the source resistance is low. This is exactly the case with a magnetic pickup and the effect is worsened by the bass-boost of the R.I.A.A. characteristic. It was considered most important to cater for a high-quality magnetic pickup, and for this reason a bipolar transistor is used as the input device. When operating a low-noise silicon transistor at about 80-100μA collector current, the noise figure is 2dB or less over the effective range of source resistances from 1kΩ-50kΩ. However, this optimizes the noise figure for a source resistance of about 5kΩ and as the stage must also be operated with a 200kΩ source resistance for ceramic pickups it was decided, in the interests of low flicker noise, to reduce the standing current to 35-40μA. This increases the noise figure to 3dB at low frequencies with the magnetic pickup.

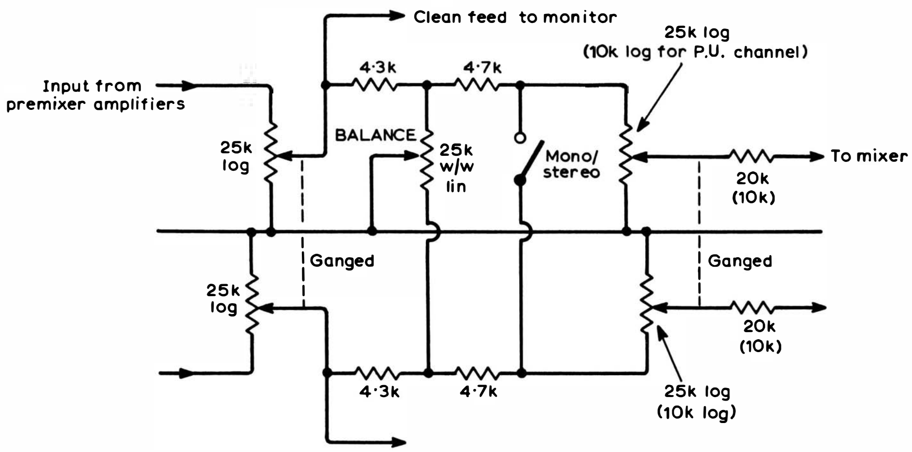


Fig. 2. Stereo balance and mono/stereo switch.

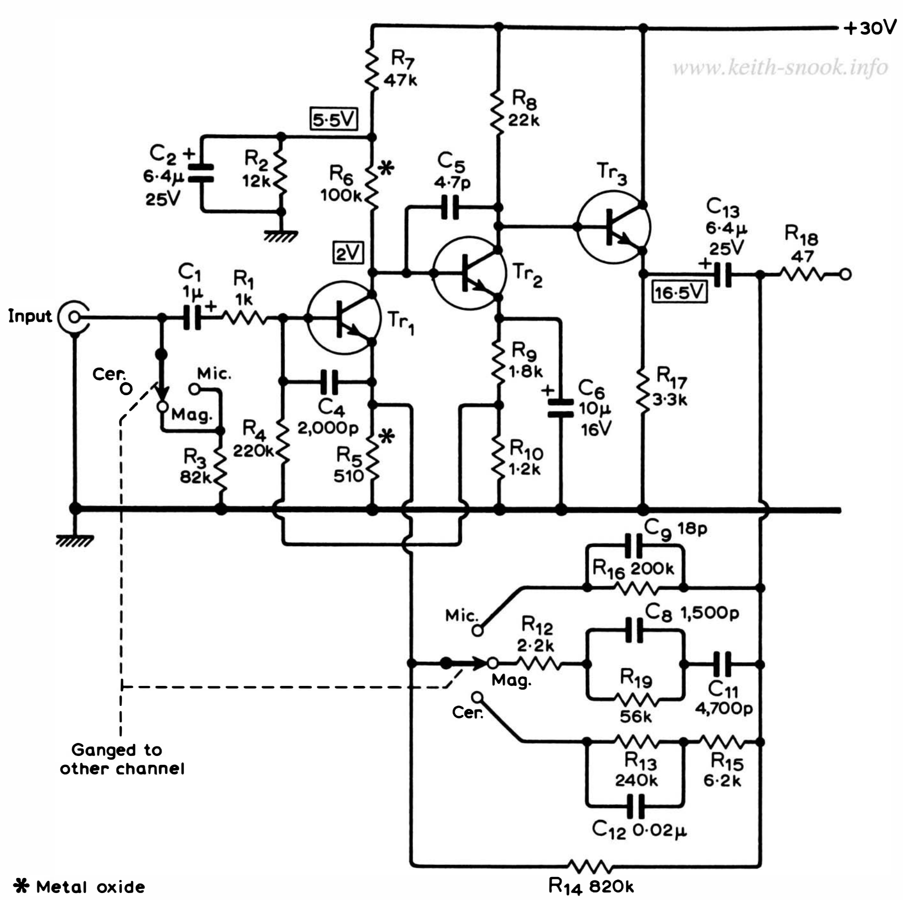


Fig. 3. Gramophone pickup amplifier with switched equalization for magnetic and ceramic cartridges, and for use with a microphone. Tr_1 BC184LC or BC109C; Tr_2 2N3707 or BC167; Tr_3 BC167 or BC107.

Ceramic pickups operating into about 200kΩ require bass boost in the pre-amplifier to balance the falling bass response caused by the input time constant[1]. The component values for C_{12}, R_{15}, and R_4 are suitable for pickups having a self-capacitance of about 600pF—which includes the majority of better cartridges. This results in turnover at about 1.5kHz and is approximately the same as that of the treble tone-control (discussed later) which can be used to compensate for different cartridge capacitances and degrees of mechanical equalization. The input resistance or feedback time constant can also be adjusted to suit other types, provided that the feedback resistor R_{15} is not made less than 5kΩ otherwise serious loading of the emitter-follower will result.

The second transistor is operated under conditions of low distortion; C_5, R_{12} and R_{18} are included to improve the high-frequency stability and only affect the performance outside the audio range. The filter comprising R_1 and C_4 is essential to prevent r.f. appearing at, and being detected by, the base-emitter junction of the first transistor. Inevitably the presence of R_1 in series with the base of Tr_1 causes a poorer noise performance (particularly at low frequencies with a magnetic pickup) but it should not be omitted if a wide variety of gramophone equipment is likely to be used.

The microphone input will match a 50kΩ high-impedance dynamic microphone. In conjunction with a transformer it is suitable for a low-impedance type. This is, however, only a useful secondary function of this circuit and the noise performance will be slightly inferior to circuits specifically designed for this kind of input.

About 20dB of n.f.b. is applied at mid-frequencies, reducing distortion to less than 0.1%. The purpose of C_9 is again to improve the high-frequency stability of the circuit.

The R.I.A.A. feedback-loop time constants are:

$$R_{19}C_8 = 82\mu s \quad R_{19}C_{11} = 240\mu s$$
$$R_{14}C_{11} = 3000\mu s$$

The values shown in Fig. 3 approximate these time constants, with the exception of the l.f. turnover*[2] (sub-audio frequencies being attenuated by a rumble filter later in the mixer), and the gain at 1kHz is given by R_{19}/R_5†. The basic sensitivity (at 1kHz) is then 2mV for 240mV output, allowing more than 30dB of feedback at mid-frequencies, falling to 10dB at very low frequencies.

High-output crystal and ceramic pickups: Pickup cartridges having average outputs in excess of 100mV r.m.s. will overload the ceramic pickup input unless

* It was pointed out by Mr. Linsley Hood[2] that record manufacturers do not boost the frequencies below 50Hz in accordance with the R.I.A.A. characteristic and therefore a larger l.f. time constant will allow the fuller reproduction of these rather 'dubious' signals. Making R_{14}=620kΩ would restore the correct 3180μs time constant.

† Altering R_5 is the simplest way of adjusting the basic sensitivity.

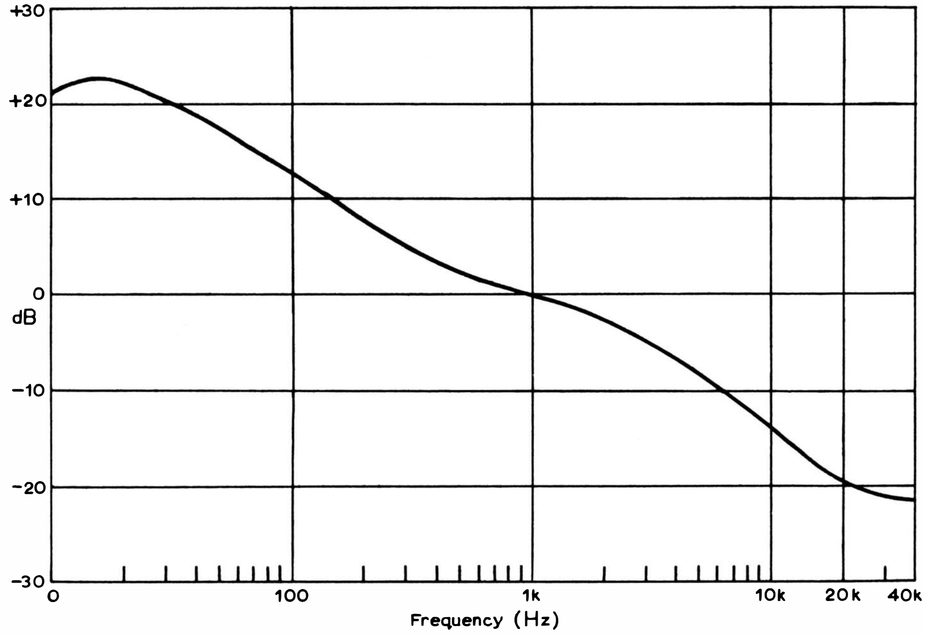


Fig. 4. Measured R.I.A.A. characteristic for magnetic pickup input of Fig. 3.

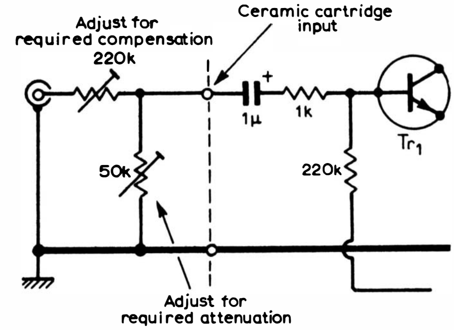


Fig. 5. Attenuator for use with high output ceramic cartridge when employing Fig. 3.

Table 2. Performance characteristics of Fig. 3

Specification:	Magnetic p.u.	Ceramic p.u.	Microphone
Sensitivity	1.4mV r.m.s. at 1kHz	15mV at 1kHz	0.7mV at 1kHz
Input impedance	60kΩ	220kΩ	60kΩ
Distortion at nominal levels	<0.02%	<0.02%	< 0.02%
Overload margin (0.1% dist.)	>30dB at 1kHz & 10kHz	28dB	30dB
Signal-to-noise ratio referred to max. sens.	67.5dB input shunted with 420mH	70dB input shunted with 680pF	42dB input open circuit

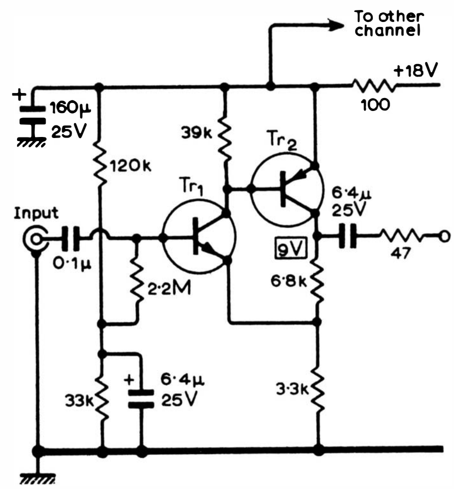


Fig. 6. Amplifier for high-output ceramic cartridge. Tr_1 BC169, BC184LC, etc.; Tr_2 2N4058, 2N4062, etc.

resistive attenuation is placed between the cartridge and the input. A suitable circuit is shown in Fig 5.

The circuit of Fig. 6 has the merit of costing very little to build. The first transistor is run at a collector current of 10-20µA and current-drives the second transistor, whose collector voltage is defined by the d.c. and a.c. feedback to the emitter of the first transistor. Using the normal Darlington connection would achieve the high input impedance but would result in gross distortion because of the effective voltage drive at the base of the second transistor.

Performance

Input impedance (measured) 2MΩ in parallel with 2pF
Max. sensitivity 70mVr.m.s.
Max. input 1.8Vr.m.s. (overload margin > 26dB)
Distortion for an input of 600mVr.m.s. at 1kHz < 0.02% not exceeding 0.1% for input of 1.5Vr.m.s. in audio band
Signal-to-noise ratio ref. to input of 70mV (1000pF on input) > 80dB.

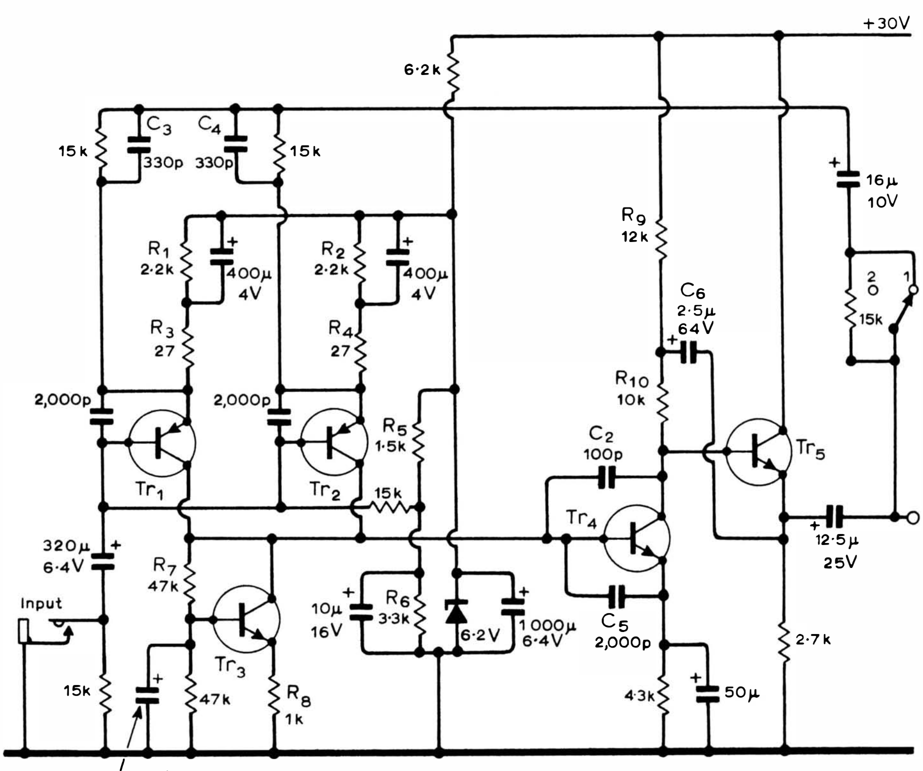
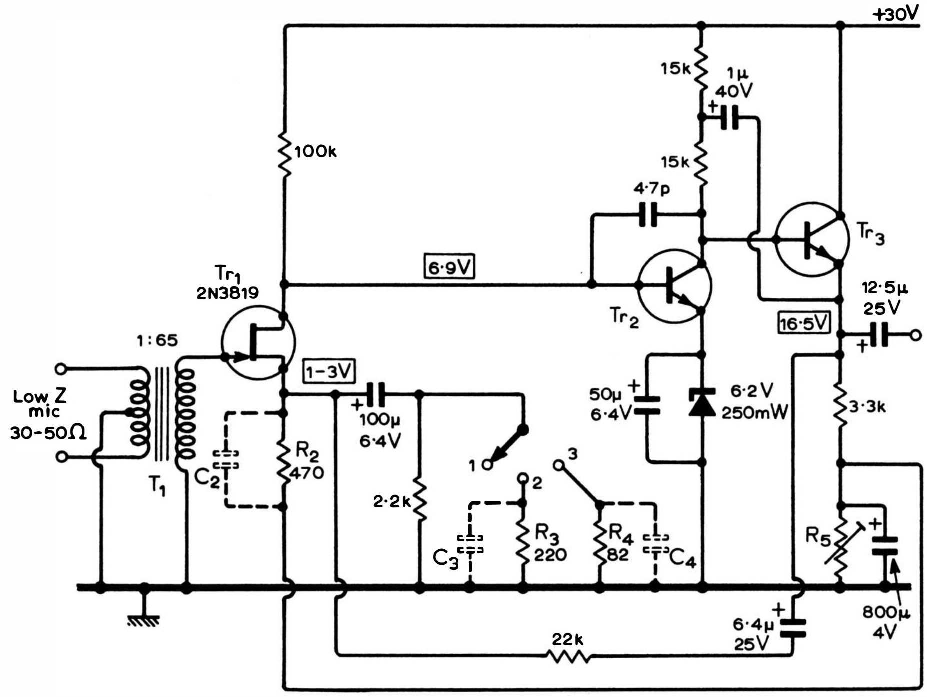


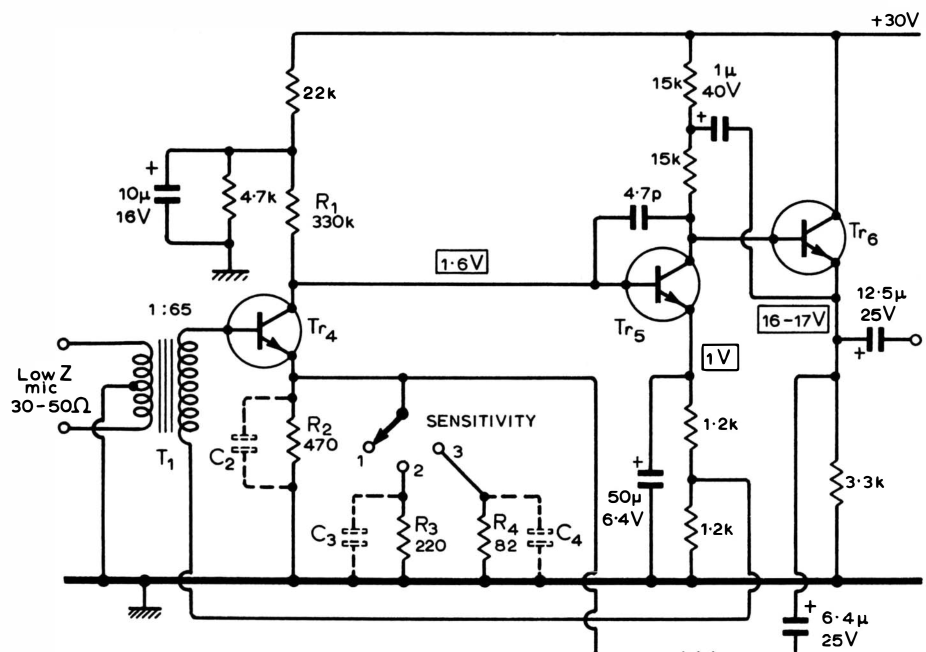
Fig. 7. Amplifier for medium impedance microphone. The approximate maximum sensitivities are 400 to 500µV (600Ω mic.) and 150-200µV (200Ω mic.). The overload margin is 30dB (0.1% distortion). Frequency response is 20Hz-25kHz (—3dB) and s/n ratio not less than 60dB (66dB with 200Ω mic. ref. 45µV @ 1kHz). Tr_1 Tr_2 2N4058, 2N4126 or 2N4289; Tr_3 BC109, etc; Tr_4 2N3707, Tr_5 BC167, etc.

Microphone amplifiers

First, a brief summary of the types of microphones which are likely to be encountered. Crystal microphones will not be considered because of their low sound quality.



(a)



(b)

Fig. 8. Low-impedance microphone amplifiers. The overload margin is 30dB for 0.1% distortion. Sensitivity for a 30Ω input impedance is determined by the three-position switch—(1) 80-100μV, (2) 25-30μV, (3) 12-20μV. Distortion ≈0.02%. S/N ratio 63.5dB. w.r.t. 100μV @ 1 kHz 30Ω source (20kHz bandwidth). T_1 in the prototype was Radiospares 'Hygrade' type. Tr_1 2N3819; Tr_2 and Tr_5 2N3707; Tr_3 and Tr_6 BC167, BC107 or 2N3707; Tr_4 BC169, BC1841C or BC109. C_2, C_3 and C_4 will have to be added to compensate for transformer losses.

Capacitor microphones: These have their own head-amplifier within the microphone case, the output signal of about 1mV being sent down a low-impedance line (typically 200 or 600Ω).

Ribbon microphones: Depending on the step-up transformer supplied with the microphone, the impedance and signal levels could be low Z 30-50Ω (typical open-circuit voltage 100-200μV) or 200-600Ω (200-500μV), med. Z 1k-1.5kΩ (600-800μV), and high Z 50kΩ (2-4mV.)

Moving-coil microphones: Again, the impedance of these depends on the transformer supplied, but generally are either med. Z 200-600Ω (typical open-circuit voltage 200-800μV), and high Z 50kΩ (2-4mV.)

We will now consider various low-noise amplifiers suitable for the last three categories. High impedance types suffer losses at high-frequencies because of the length of cable which can be used between microphone and amplifier. If used, they should be connected directly to the circuits of Fig. 8(a) or 8(b), omitting the transformer but including a resistor (e.g. 220kΩ) in place of the transformer secondary—an isolating capacitor, say 1μF, will be required with circuit of Fig. 8(b).

The medium impedance versions ($200\Omega < Z_{mic} < 1.5k\Omega$) are suitable for direct connection to the circuit of Fig. 7. This rather unusual circuit makes use of the principle suggested by E. A. Faulkner[3] of paralleling several transistors to overcome the limitation on achievable noise-figure for low source resistances. Also, p-n-p transistors have lower "effective" base resistances[4] (which determines the minimum n.f.); the greater circuit complexity results from the d.c. voltage requirements of these devices. The point of interest in this circuit is the collector load of the paralleled Tr_1 and Tr_2, which by means of Tr_3 presents a low d.c. resistance of about $2R_8$ and a high a.c. resistance of R_7. The effectively constant-current collector load for a.c. signals current-drives the base of Tr_4—a requirement for low distortion. The current of about 0.5mA in Tr_1 and Tr_2 is set by R_1, R_2 and the base voltages derived from the 6.2V sub-rail by the potential divider R_5 and R_6.

The switched sensitivity control in the feedback loop makes the circuit suitable for both the medium-impedance versions of the ribbon and moving-coil microphones and the more sensitive capacitor microphones with approximately 1mV output. By paralleling more p-n-p transistors, this method could be extended to the design of a low-impedance microphone amplifier. However, the greater circuit complexity is not justified when one considers the added advantage of a transformer input, namely the cancellation of extraneous signals picked up on a balanced line.

With regard to step-up transformers, mounted close to the microphone amplifier, one is most likely to encounter the high-impedance transformers (1:50) prevalent in the valve days and still made by most firms. The objection to the large turns ratio on these transformers is the degrading effect on the high-frequency response caused by the high secondary impedance, leakage inductance and winding capacitance. However, the circuits in Figs. 8(a) and 8(b), which are suitable for this type of transformer, can offset this limitation by including high-frequency compensation.

The f.e.t. used as the input device in Fig. 8(a), gives a good noise figure with the high secondary impedance of 100-200kΩ, and, because of its high input impedance, results in negligible attenuation of the microphone signals. The d.c. conditions for this circuit are set by the 6.2V zener in the emitter of Tr_2 and by the negative feedback from the tapping in the emitter-follower load (Tr_3) to the source of Tr_1. Unfortunately the value of R_5 must be adjusted for each f.e.t. because of the spreads in V_{GS}/I_{DS} characteristics.

The circuit of Fig. 8(b) is a modified form of the gramophone pickup amplifier shown in Fig. 3. The bipolar transistor, Tr_4, is operated at a very low collector current (≈10μA) to provide a good noise figure. The transformer secondary acts as the d.c. feedback loop, replacing the usual resistor —e.g. R_4 in Fig. 3—and thereby avoiding unnecessary attenuation of input signals. Both circuits employ the 'boot-strapping' technique to improve the linearity; a fact which is reflected in the excellent overload margin. A switched feedback sensitivity control makes the circuits suitable for most low-impedance microphones and the addition of a suitable capacitor in parallel with the feedback resistor, R_2, R_3, R_4, compensates for high-frequency losses in the transformer. The performance of the two circuits is almost identical except that when used in the most sensitive condition, the f.e.t. input is superior because with bipolar transistors the non-linearity of the V_{BE}/I_B characteristic is very marked at low collector currents; ultimately we depend on this part of the circuit to perform the subtraction of feedback from signal voltage.

The most suitable microphone transformer for the bipolar circuit of Fig. 8(b) would have a turns ratio of between 1:15 and 1:30 giving a secondary impedance of 10-30kΩ. For these transformers the collector current in Tr_4 should be set at about 30-40μA by reducing R_1 to 100kΩ. The higher sensitivity required under these conditions can be achieved by halving the values of feedback resistors R_2, R_3, R_4.

Microphone transformers with secondary impedances of 1-1.5kΩ could be used with the circuit of Fig. 7 though now only one p-n-p transistor operating at 0.5mA would be necessary instead of the two in parallel. As a suggestion, readers might like to try paralleling two n-p-n transistors and modifying the circuit of Fig. 8(b).

Finally, two constructional points. When designing a component layout for any of the above circuits, one must take care to keep input leads as short as possible and adequately screened, particularly for the high-impedance transformer-secondary connection to the inputs of Figs. 8(a) and 8(b). Switched jack sockets should be used, and wired so that when not in use the input is shorted.

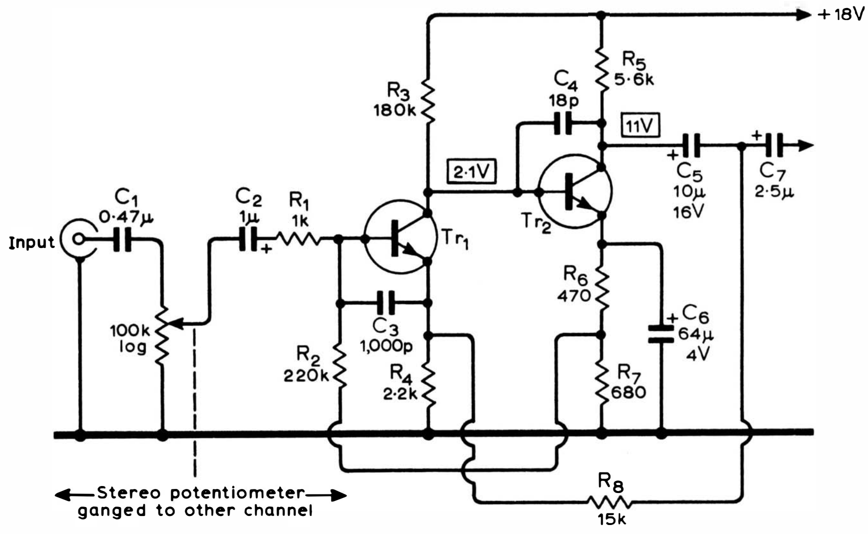


Fig. 9. Auxiliary amplifier. Tr_1 Tr_2 BC109 etc.

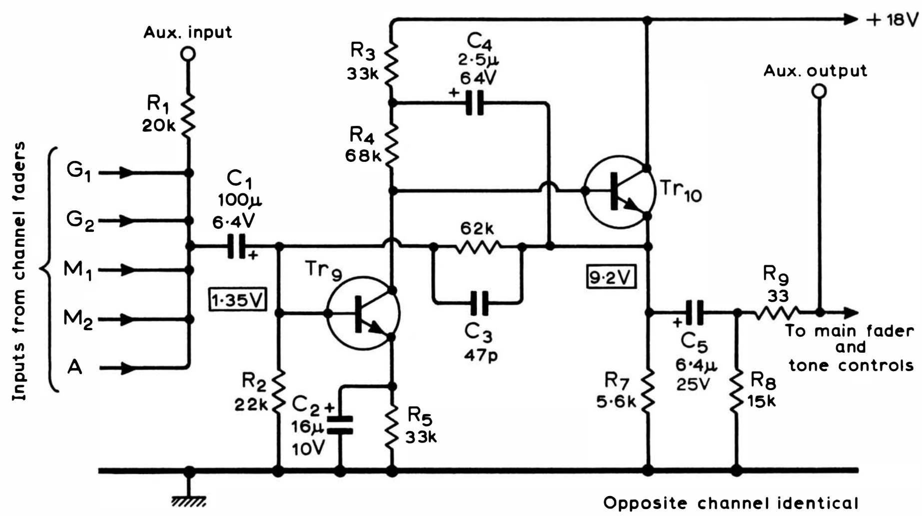


Fig. 10. Virtual earth mixer. Tr_9 Tr_{10} BC109 etc.

Auxiliary amplifier

So that the auxiliary input can handle a very wide range of input signal levels, the preset sensitivity control is placed at the front of the pre-amplifier, as shown in Fig. 9. The amplifier gain is given by $(R_8+R_4)/R_4 \approx 8$ making the basic sensitivity about 30mV. Although the maximum signal-to-noise ratio is no longer obtainable, because of the resistive attenuation, the worst possible noise level is still better than 70dB below a 30mV input.

When large signals are applied to the base of Tr_1 (as is the case when a large amount of feedback reduces the gain to less than ten), insufficient collector-base voltage may cause distortion due to saturation in the first transistor. To avoid this, the d.c. feedback resistor, R_2, is connected to a tapping in the emitter resistor, (R_6+R_7), of Tr_2 to increase the collector-base voltage of Tr_1.

The r.f. filter, R_1 and C_3, is present, as before, to prevent radio breakthrough, and C_4 serves to improve the h.f. stability. A feedback factor of greater than 30dB ensures that the distortion is much less than 0.02% at working output levels over the whole audio frequency range. The minimum input impedance is about 70kΩ.

Virtual-earth mixer

Fig. 10 shows the complete circuit of the mixer. The bootstrap capacitor, C_4, increases the amplifier gain to over 4000 and reduces harmonic distortion to less than 1% for a 3Vr.m.s. output. About 60dB of n.f.b. is applied; this reduces distortion to quite negligible proportions ($<0.01\%$) and ensures proper mixing of the signals from the channel faders with no interaction. The capacitor, C_3, in parallel with the 62kΩ feedback resistor, curtails the very extended high-frequency response which might cause instability with some layouts.

The provision for five stereo channels in the present design should satisfy most requirements; the more versatile system of several virtual-earth mixers and group faders is preferable when mixing a greater number of channels. The nominal signal level at the output is about 350mV and residual noise is 84dB down (measured on a bandwidth of ~ 20kHz).

REFERENCES

1. Burrows, B.J.C., 'Ceramic Pickups and Transistor Pre-amplifiers', *Wireless World,* February 1970.

2. Linsley Hood, J.L., 'Modular Pre-amplifier Design', *Wireless World,* July 1969.

3. Faulkner, E.A., 'Optimum Design of low-noise Amplifiers', *Electronics Letters (I.E.E.),* Vol.2, No.11. pp.426-427 November 1966.

4. Faulkner, E.A. and Harding, D.W., 'Some Measurements on Low-noise transistors for Audio-frequency Amplification'. *The Radio and Electronic Engineer,* 36, No.1. pp.31-33. July 1968.

Stereo Mixer

2—Further circuits and construction notes

by H. P. Walker, B.A.

Post-mixing circuits

In the prototype, the signal from the virtual-earth mixer is taken to the tone-control circuit via a switchable rumble filter and the main gain fader and stereo balance controls. The signal is fed at low impedance from the tone-controls directly to external equipment or passed on to the line and metering amplifiers.

Tone-control and filter circuits. The circuit shown in Fig. 11 can conveniently be divided into two by the main gain and stereo balance controls. These controls are similar to those described for the pre-mixing amplifiers except that the balance control does not fade either channel to zero. The author chose a 10kΩ potentiometer for the main gain fader after finding that a 50kΩ potentiometer caused an audible deterioration in the residual noise level when the control was set at about half full output.

Filter circuits. Prior to the stereo balance control, a switchable rumble filter attenuates frequencies below 25Hz at about 24dB/octave in conjunction with the built-in low-frequency turnover, as shown in the tone-control characteristics, Fig. 12. The high-pass filter is synthesized using emitter-follower sections which are cheap to construct and give a gain of unity in the pass-band. Low-pass filters may also be synthesized in a similar manner to give various turnover characteristics (e.g. maximally flat amplitude or linear phase) and various rates of attenuation.

Although no high-frequency filter is included in the present design, its use is definitely advantageous when gramophone records are being reproduced. The omission is partly due to the prejudice of the author who feels that only a comprehensive h.f. filter system, in which the slope and turnover-frequency are variable, is really useful: the best known is that on the Quad 22 and 33 control units*. The subjective affect of Butterworth (or maximally flat magnitude) low-pass filters is often to increase the noise and distortion they are intended to reduce, while the signal assumes an unpleasant 'nasal' quality due to the transient distortion. This is caused by the abrupt turnover in the amplitude response followed by a steep roll-off at 18dB/octave or more, and is easily demonstrated by examining the theoretical response of such a circuit when excited by a step function[7] (e.g. a square wave).

As a result, the author has never found this type of filter very useful and frequently prefers the original noise and distortion to the coloration caused by the transient

*A useful passive circuit having a fixed turnover frequency and variable slope was described by R. Williamson in *Hi-Fi News*[5] and reproduced in a later article by B. Grossmith[6].

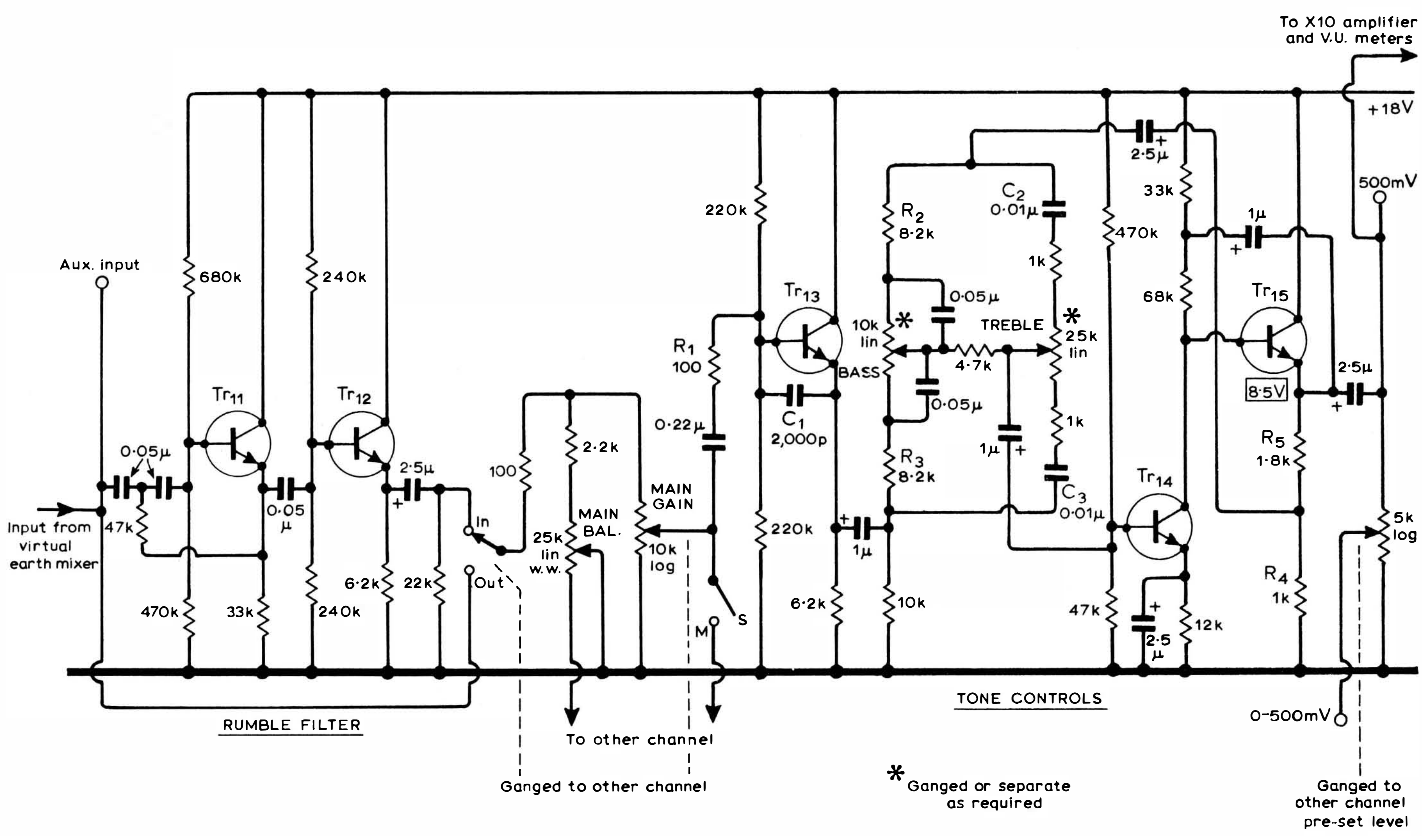


Fig. 11. Tone-control and filter circuit. Residual noise −98dB. Tr_{11}-Tr_{15} BC109 etc.

ringing. However, if the rate of attenuation is reduced to 12dB/octave this transient distortion is aurally less noticeable while tests using a Bessel (linear phase or maximally flat time delay) filter[7] gave, subjectively, a quite pleasing effect even with an 18dB/octave roll-off. This is to be expected from the improved transient response resulting from the more gradual turnover characteristics. A practical circuit, realizing this filter, is shown in Fig. 13 and the frequency response curve for the values shown, in Fig. 14. Changing the turnover frequency involves scaling the values of C_1, C_2 and C_3, but simply halving C_2 to 220pF gives curve 2, which, although not maintaining the true Bessel characteristics, does not seem to cause audible transient distortion.

Tone-control circuit. The shape of the tone-control characteristics, like those of the h.f. filter, are a somewhat subjective problem, though there are some important specifications which must be met by any proposed system:

(i) low interaction between bass and treble controls;

(ii) low distortion even at maximum boost; and

(iii)a truly flat amplitude characteristic and a good transient (i.e. square wave) response when the controls are in the 'flat' position.

The tone-control circuit of Fig. 11 is basically a virtual-earth feedback configuration in which the bass control is of the variable turnover frequency type and the treble control has a fixed turnover frequency, approximately determined by the time constant R_2C_2, and effectively lifts and cuts the whole of the frequency range above 2kHz. The gain/frequency characteristics are shown in Fig. 12. The component values used in the tone control network are the result of experimental measurements and listening tests. In particular the treble control components were carefully chosen so that the control can alter the musical 'brilliance', and it gives a variation which is related to rotation, i.e. there is no 'dead band'.

The nominal signal level of 360mV, leaving the virtual-earth mixer, is attenuated by 6dB in the stereo balance control so that the maximum signal level entering the tone-control stage is about 180mV. As the tone-control network is symmetrical, to obtain a gain of greater than one the feedback connection must be made to a tapping in the output load resistor. The gain is then given by $(R_5+R_4)/R_4$. For a nominal output of 500MVr.m.s. a gain of 2.8 is required; for higher or lower outputs R_4 should be altered (e.g. to 390Ω for 1Vr.m.s. output), remembering of course that the circuit is not designed for outputs much greater than 2Vr.m.s.

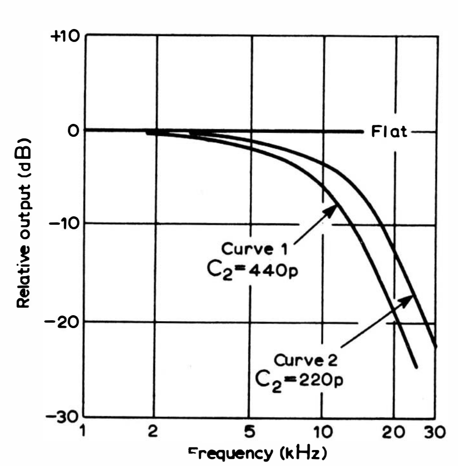


Fig. 14. Characteristics of Fig. 13.

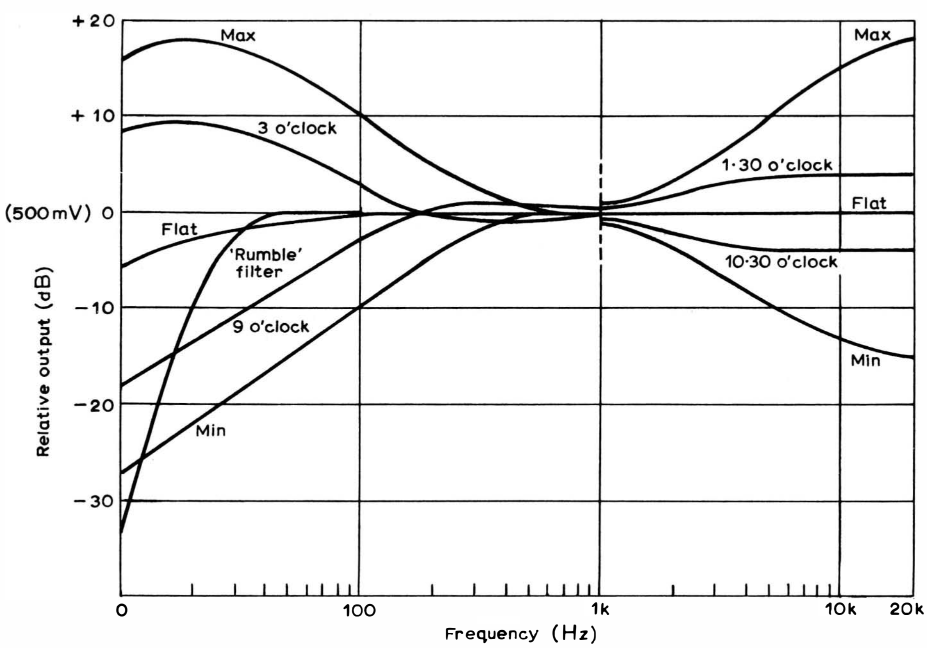


Fig. 12. Characteristics of filter and tone controls.

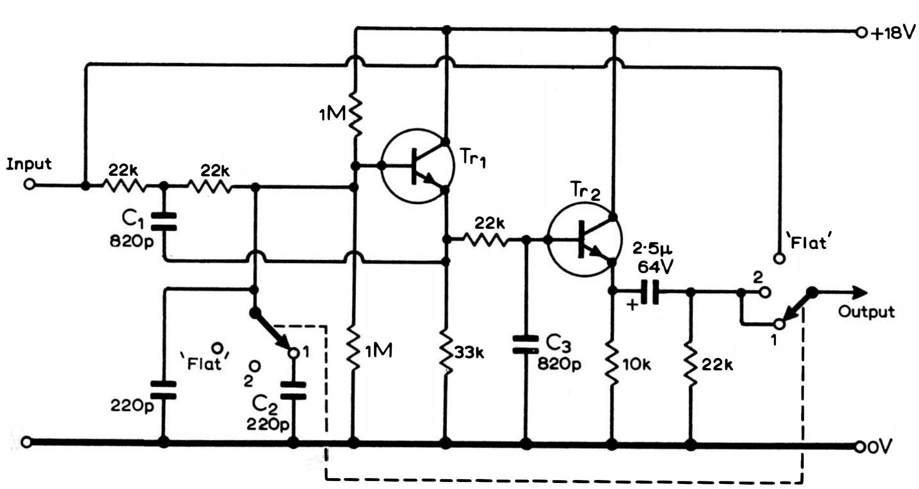


Fig. 13. High-frequency filter. Tr_1 Tr_2 BC109 etc.

Turning to the amplifier itself, the bootstrapping circuit is used which gives a gain without feedback of 2000 and a distortion level of less than 1%. This enables very low distortion to be obtained even at maximum boost. Although the effective source resistance presented by the tone-control network varies with frequency, a suitable average value for design purposes is 5kΩ, which sets the collector current at 100μA in Tr_{14} for a low noise figure. Tr_{13} provides a low impedance drive to the tone-control network, R_1 and C_1 being included to prevent h.f. instability.

The complete circuit has low interaction between the controls and in the 'nominally flat' position the overall response is flat to within 1dB over the audio range and the square-wave performance shows rise and fall times of about 0.5μs with no overshoot. Harmonic distortion for an output of 500mV r.m.s. is less than 0.02% rising to 0.05% at 2V r.m.s. The level of residual noise at the output with the main gain control turned down is approximately −93dB w.r.t. 500mV when measured on a bandwidth of −20kHz.

Line amplifiers: The mixer may often be required to feed into a 600Ω termination at a power level of several milliwatts. The line amplifiers have been designed for this purpose and single-ended and push-pull versions are shown in Fig. 15 and Fig. 16 respectively.

Both circuits have a gain of about ten, determined by the feedback components, but the single-ended stage, operating in class A, has a limited output voltage swing (at low distortion) of 2V r.m.s. into a 600Ω load. When feeding high impedance loads (> 10kΩ), such as insensitive power amplifiers, the maximum output voltage is in excess of 5V r.m.s. at less than 0.05%

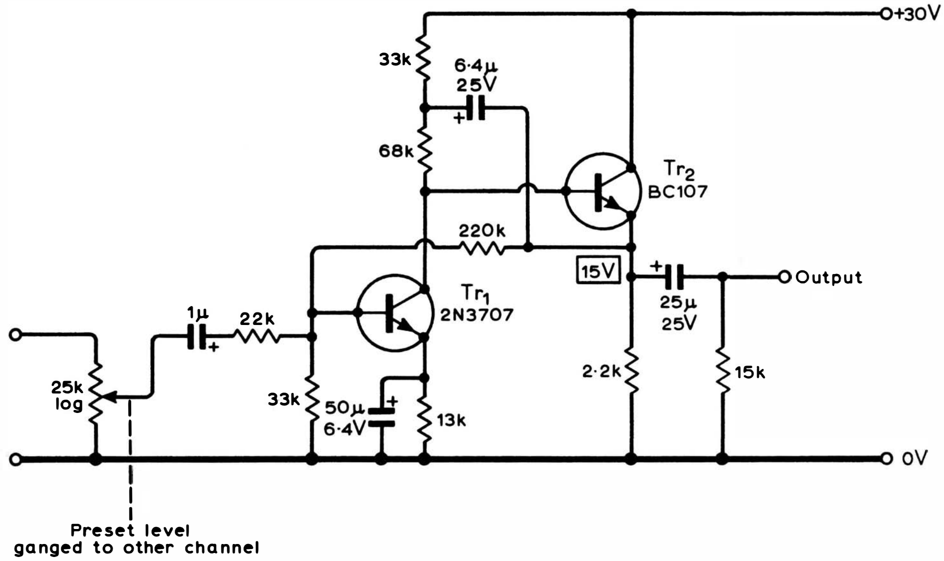


Fig. 15. Single-ended line amplifier. Distortion 0.1% at 2V r.m.s. into 600Ω; 0.05% at 5V into high impedance. Residual noise −82dB w.r.t. 500mV output (20kHz bandwidth).

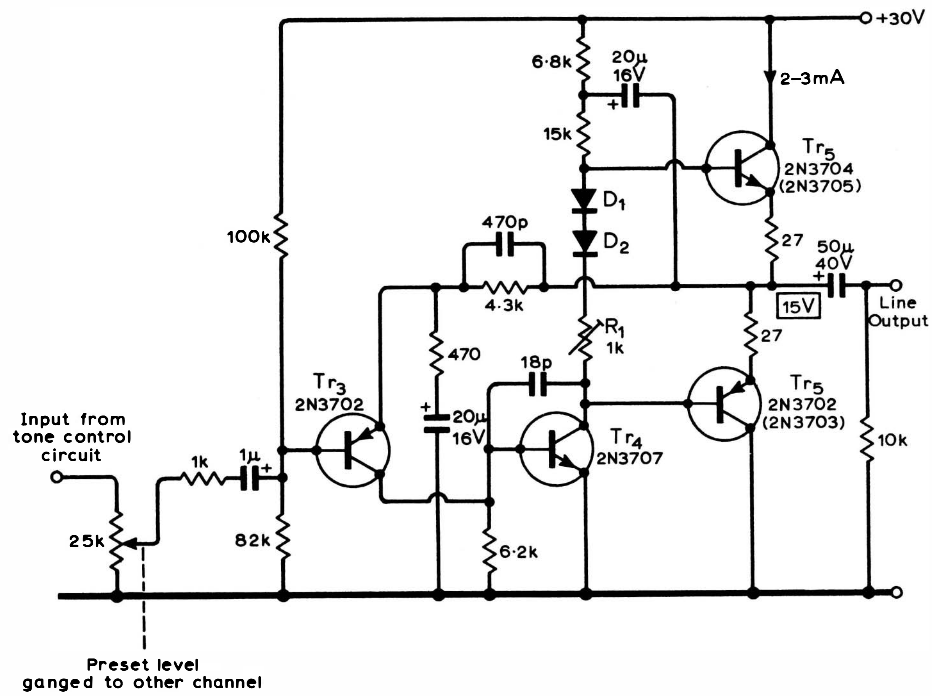


Fig. 16. Push-pull line amplifier. Distortion 0.03% at +20dBm (8V into 600Ω) at 1kHz and 10kHz, and at lower powers (20kHz bandwidth). D_1 D_2 1N914.

distortion. Residual noise is −82dB w.r.t. 500mV output (20kHz bandwidth). The single-ended circuit has the merit of simplicity compared with the more sophisticated design of Fig. 16.

The complementary emitter-follower, Fig. 16, operating in class B, can supply more than +20dBm to a 600Ω load (8-9V r.m.s.) at very low distortion (<0.05%). A quiescent current of 2-3mA in the output circuit, set by R_1, reduces the crossover distortion to a low level (none is detectable on the oscilloscope trace of the residual). The circuit configuration is that commonly used in power amplifiers, the a.c. and d.c. feedback loop being combined as they also are in the single-ended design. Residual noise is less than −95dB measured as above.

Monitor amplifier: The complete circuit is shown in Fig. 17. The clean feeds and mixed output are selected by the two-pole, six-way switch and a 50kΩ log. potentiometer is used as a monitor level control to prevent loading when the individual channels are selected. The monitor amplifier is basically the same configuration as that used in the line amplifier, Fig. 16, except that a higher standing current of 10mA is used in the second transistor and a complementary pair of power transistors, Tr_3 and Tr_4, make the circuit suitable for driving a loudspeaker up to about 2W. The amplifier is, however, primarily meant for headphone monitoring and under these conditions its performance is very good, but the possibility of l.s. operation could be useful for a talk-back facility in live recording. Needless to say the output transistors should have some form of heat sink for continuous loudspeaker operation.

The components R_1, C_1 and C_2 improve the high-frequency stability. The preset, R_4, is adjusted for symmetrical clipping at the output (a centre-line voltage of 10V d.c.) and R_5 is used to set the quiescent current of 10mA in Tr_3 and Tr_4. The diode, D_1, across which the bias voltage is developed, should be mounted close to the power transistors to aid thermal stability. A gain of approximately 20, determined by $(R_2+R_3)/R_3$, is sufficient to cause overload when the amplifier is driven from the clean feeds at a nominal signal level of 240mV. Normally some resistance (e.g. 15Ω) must be connected in series with low-impedance headphones to protect the phones—and also the operator's ears!

Signal-level meters: Some form of metering is essential in a mixer of this complexity if maximum signal-to-noise ratio and low distortion are to be obtained. Apart from the obvious value in setting-up procedures and in monitoring output levels, a properly calibrated built-in meter is very useful for checking pickup cartridges, tone control characteristics, channel balance and so on.

The relative merits of peak programme and VU meters are a controversial subject![8] The use of VU meters in this design does not presume any general preference for this type of metering. In the author's opinion peak programme meters are better as recording level indicators, whereas VU meters seem to give a more realistic indication of sound or signal level. In either case, though, the interpretation placed on meter readings depends on the experience of the user.

In the prototype the metering circuit was connected directly to the output of the tone-control stage, Fig. 11, but naturally it could be switched to read other data in the equipment. The simple one-transistor amplifier of Fig. 18 provides the required gain and any frequency compensation and isolates the meter (and its non-linear loading) from the signal circuit. The meter is the heart of the system and while it is possible to obtain satisfactory results without spending vast sums of money, the constructor should avoid 'cheap' meters which are often poorly damped and inaccurately calibrated and give quite meaningless readings on transient signals. The meters used in the prototype were SEW MR45P, from G. W. Smith & Co. (Radio) Ltd., which give adequate performance for the author's requirements; more expensive models would give more consistent readings. The values of R_C and C_C given in Fig. 18 relate to this particular meter and give a calibration accurate to within 0.5dB at high frequencies.

Power supply

The advantages of a regulated power supply are—

(i) rail voltages independent of mains supply fluctuations,
(ii) mains-borne interference partially suppressed, and
(iii) low-impedance power supply rails,

avoiding possible low-frequency instability due to interaction between circuits.

This last point is particularly relevant in circuits where emitter-follower outputs draw currents of several milliamps. Although the author included an a.c. decoupling network (100—200Ω and 100μF) in series with the power supply rail to most of the individual circuits, this was avoided with the tone-control circuit because the bass-boost control would make the circuit rather sensitive to l.f. crosstalk and instability if maximum boost were used. Electronic overload protection can be built into such a supply and can prevent excessive damage under fault conditions.

The complete power supply circuit shown in Fig. 19(a) is the one used by the author for the prototype but there are several variations possible to suit different requirements. These are mainly determined by the desired output facilities. If it is intended to use the low-power loudspeaker monitoring facility (described in connection with Fig. 17) then the separate 20V regulated supply (Fig. 19(a)) should be used since this provides isolation from the supply rails to the rest of the mixer and incorporates a peak overcurrent protection circuit, R_2 and Tr_6. The moderate current swings required for headphone monitoring allow the monitor amplifier to use the same supply rails as other circuits in the mixer. For this application the series transistor regulator, shown in Fig. 19(b), should be substituted for the shunt zener stabilized circuit.

The series regulators are of conventional design. Transistors Tr_1 and Tr_4 are mounted on heatsinks (the metal chassis is satisfactory) and Tr_8 (Fig. 19(b)) should be fitted with a clip-on, finned, heatsink though currents of more than 50mA are better handled by a chassis-mounted 2N3054 (with R_4 reduced to 2.2kΩ). When the current drawn from the supply is sufficient to develop about 0.6V across R_1 (or R_2), Tr_2 (or Tr_6) begins to conduct and removes a proportion of the current drive from the series regulator to maintain a constant current limit which is set by Vbc_2/R_1.

It may have occurred to readers that, because the 0V rails of the regulator outputs will be a common signal earth, the same control voltage will appear across both R_1 and R_2 and consequently the current limits cannot be set independently. In passing, it should be noted that common power supply rails must be paralleled at the actual supply and *not* by devious routes in the mixer.

Under working conditions R_1 is between 2.7Ω and 1.2Ω and R_2, which will dominate the current limit, should not be less than 0.3Ω thus setting the peak-current limit at 2A. When testing the mixer circuits during construction, the author found it expedient to disconnect the 20V regulated supply and increase R_1 to a much larger value (say 10Ω) as a safeguard against wiring mistakes. The maximum current which can be drawn from the zener-stabilized 20V supply is determined by the value of R_3, such that $I_{max} = 10/R_3$.

If, however, there is likely to be a large variation in the current required from this supply, allowance must be made for the dissipation in the zener diode and the alternative circuit, Fig. 19(b), is preferable.

Little need be said of the rest of the circuit; any type of rectifying diodes and transformer can be used so long as they meet the specifications.

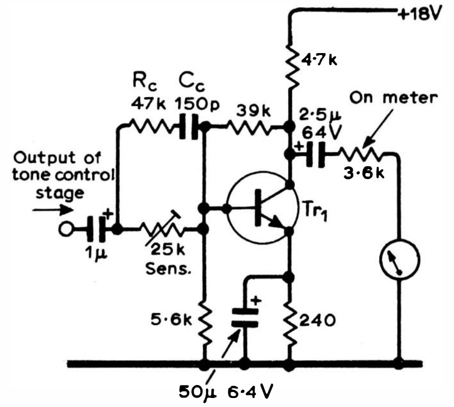


Fig. 18. VU-metering amplifier. Tr_1 BC109 etc.

Earthing

Care must be taken that earth loops do not arise within the mixer or when it is connected to other equipment. When a low-resistance closed loop is formed (e.g. by a multiple earth return of busbar and chassis) a relatively large current can flow for only a small induced voltage. If this current happens to pass through the earth return to the input of a sensitive amplifier, the potential difference developed across this earth return is in series with the signal source and a background hum is produced.

The presence of both power supply and signal earth returns is a potential source of trouble and to prevent it, low-value resistors (1.2—4.7Ω) should be included at suitable points. Fig. 20 shows the method used by the author for earthing the sensitive input circuits and no earth-loop problems were experienced with the prototype. Note that the only low-resistance path is between the input sockets; also that 'dry-joints' in the earthing system can create low-value resistances in the signal-earth return path.

Much the same kind of arguments apply to the interconnection of equipment, namely, the mixer should not be connected to amplifiers, tape machines, etc., via both the mains and signal earths.

To facilitate 'setting-up' procedures the author fitted a switchable mixer earth return (Fig. 20) and also optional turntable earthing switches in case some strange earthing arrangement is encountered with ancillary equipment.

It is very difficult to generalize on earthing problems and although one can become quite speedy at this kind of trouble shooting, it can also be very perplexing at times. However, an intelligent and logical approach from the beginning helps enormously if trouble does occur and one should not tempt providence by placing mains transformers close to input circuits and microphone transformers. Earth loops can usually be identified by the 'edgy' character of the hum they produce (more of a 'buzz' in fact).

Constructional details

The photographs, Fig. 21, show the front and back of unit 1, which contains

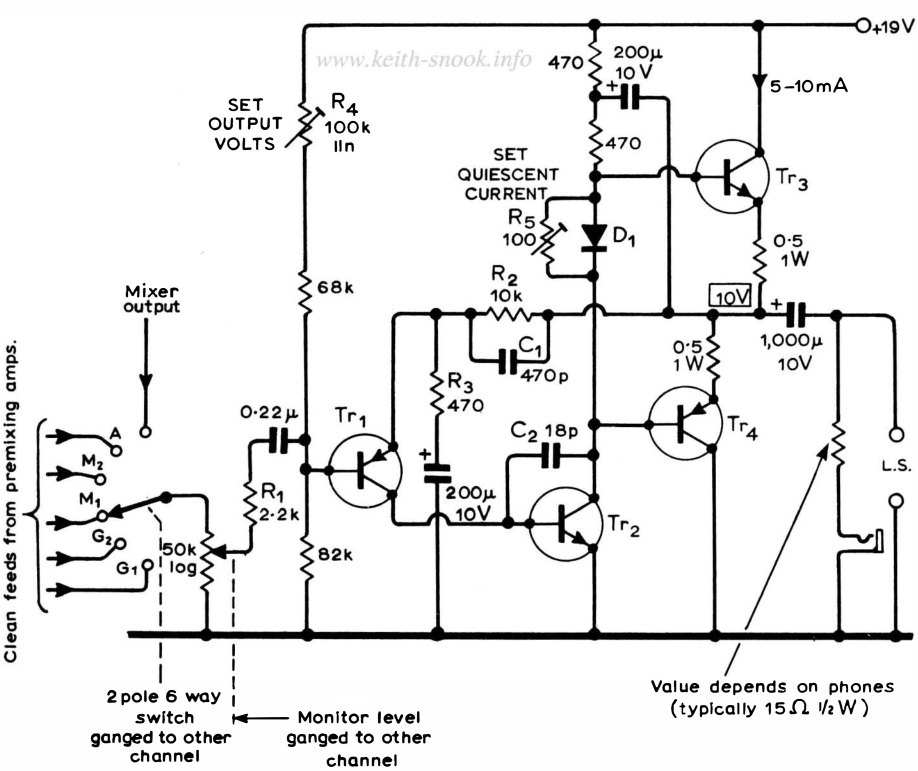


Fig. 17. Monitor amplifier. Tr_1 2N4058, 2N3702 etc; Tr_2 BC109 etc, Tr_3 Tr_4 AD161/162 (matched), D_1 OA95.

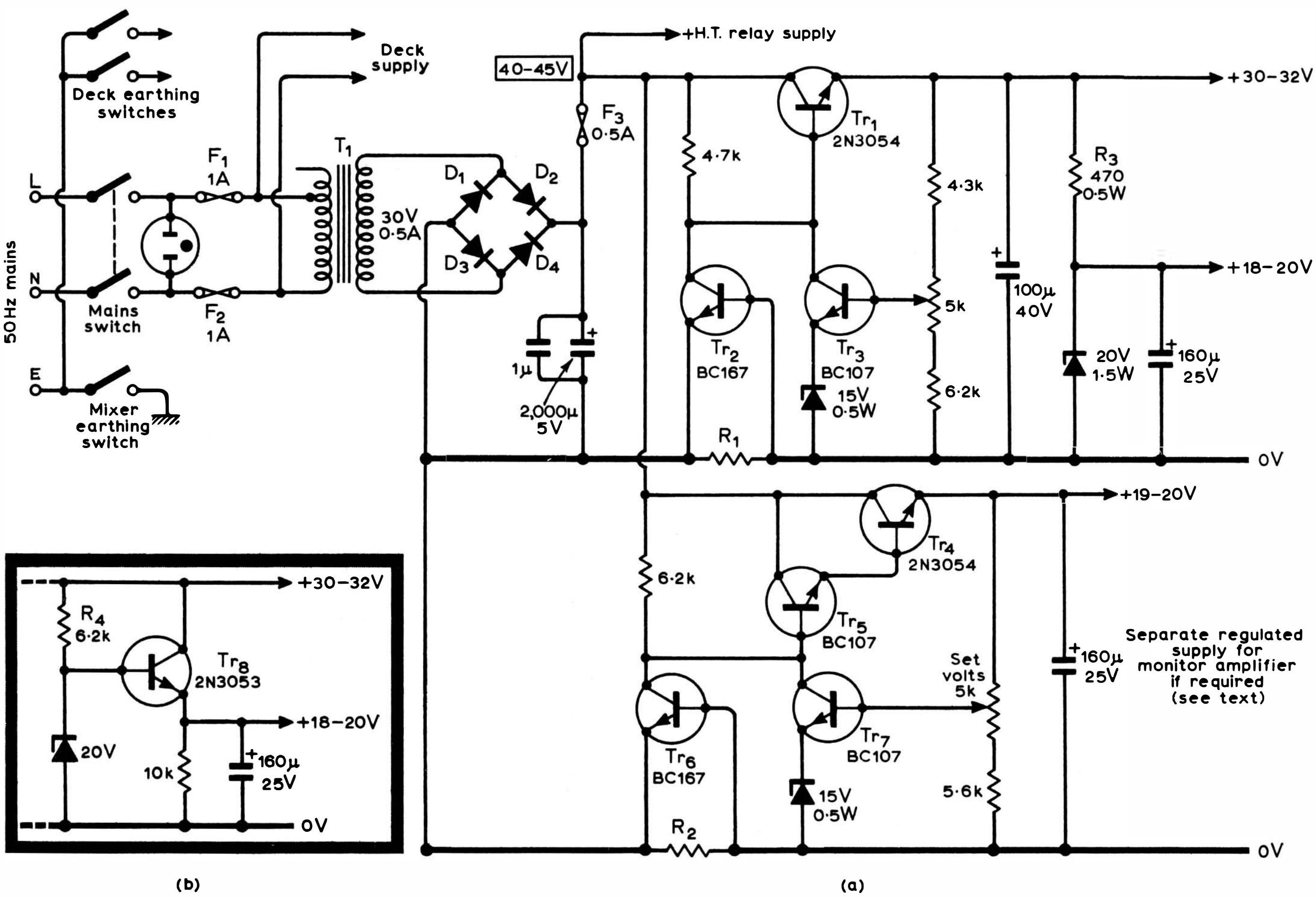


Fig. 19. (a) Comprehensive power supply. $D_1 - D_4$ 100V 0.5A. T_1 30V 0.5A. (b) Alternative to zener diode stabilized 18V rail.

the premixing amplifiers, channel faders and virtual-earth mixer. Interconnection between units is by one 8-way cable carrying signal and power supplies and one 12-way cable for clean-feed monitoring signals. Unit 2 contains tone-controls, output and monitor amplifiers, metering facilities and regulated power supply. The advantages of using two units instead of one are, first, that the sensitive input circuits can be isolated physically and electrically from the a.c. mains wiring; secondly, the greater flexibility compared with the rather cumbersome single unit; and thirdly, that if one wished to use completely different mixing facilities, unit 1 could be changed while still maintaining the facilities of unit 2 in its present form.

All the individual circuits, with the exception of the power supply, were constructed on plug-in printed circuit boards made by the author from $\frac{1}{16}$in copper-clad laminated board. Input facilities of the mixer can then be changed simply by removing one board and plugging-in another. As all connections to a board must be made in close proximity to each other, care must be taken to keep input and output as far apart as possible particularly on high gain, 'wideband' circuits (e.g. microphone amplifiers).

It is logical to layout a stereo pair of circuits as mirror images of one another and then the circuits will still operate if the boards are accidently inserted back to front. The outermost tabs on the edge connector for any board are the power-supply earths, and they connect via the low-value resistors to the signal earth busbar round the perimeter of the board. The positive rail enters on the innermost tabs; the different supply voltages are standardized to enter on different sets of tabs (18V on the middle two and 30V on adjacent tabs) so that even if an 18V board were accidently plugged

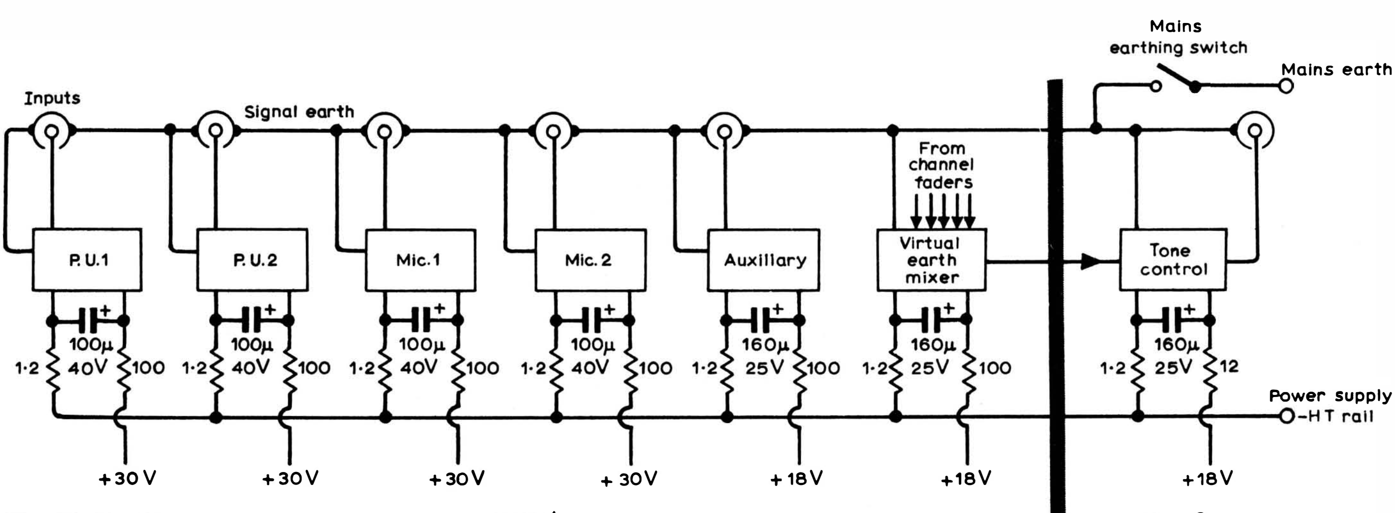


Fig. 20. Earthing arrangements.

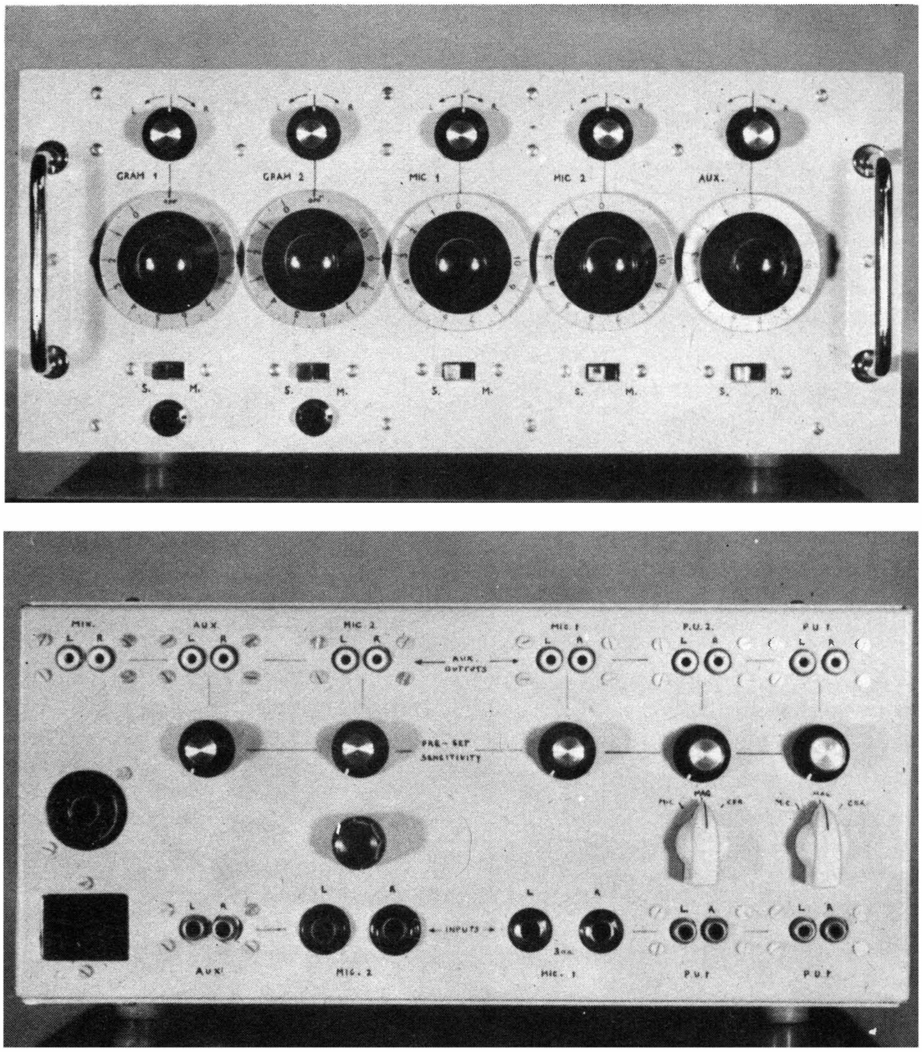


Fig. 21. Front and rear views of unit 1.

into a 30V socket, no damage could result. Signal outputs were normally placed near the edge of the board and inputs near the middle.

For obvious reasons the connections between input sockets and edge connectors should be as short as possible and properly screened. The circuits themselves are sufficiently stable for reasonable lengths of screened wire to be used in connecting them to controls and no great care was exercised in this respect when building the prototype.

So far in this article we have not mentioned the subject of crosstalk because it is a function only of layout, provided that the circuits have been designed with low output impedances to prevent electrostatic induction. For the circuitry from the virtual-earth mixer input to the output, the author obtained readings of -75dB and -66dB (ref. 500mV r.m.s. output) at 1kHz and 10kHz respectively. Crosstalk in the input circuits is very dependent on how the input of the other channel is loaded but typically the electronics will have a crosstalk 30dB better than that of pickup cartridges and tape machines. The worst crosstalk is likely to occur with sensitive microphone inputs though here the measured values for the prototype were considerably better than -40dB even at 10kHz.

Testing

If the constructor does not own the necessary equipment to test the specification of the completed circuits, simply checking the d.c. voltage at the amplifier output (shown on the circuit diagrams) is correct can be taken as an indication that the circuit is functioning properly. This is because the direct coupling between stages requires all d.c. conditions to be right.

Using the mixer

To obtain the highest signal-to-noise ratio at low distortion, some attention must be given to the adjustment of preset sensitivity controls and it is here that the metering facility comes into its own. The steps are as follows:—

(i) Turn-up the main gain fader to 0.75 full output and leave the main balance control at the central position.

(ii) Turn-up the channel fader to the desired working point (e.g. 0.75 full output) and then adjust the preset sensitivity and channel stereo balance controls for maximum outputs of 0 VU when a monophonic tape or record is being played stereophonically. Repeat this procedure for all channels to be used.

The important point is to use full signal level after the channel faders without going to the other extreme of overloading the virtual-earth mixer and obtaining the maximum output of 0 VU with the main gain control turned down to low levels. Conversely if low outputs are required from the mixer (e.g. for tape recorders or sensitive amplifiers) these should be achieved by turning down the main gain fader or preferably the preset output level control and *not* by using a low mixing level with the main gain control turned fully up. The main stereo balance control is intended to compensate for imbalance in output equipment such as loud speakers.

Some readers may think this is stating the obvious, but the author has sometimes found a logical approach the only way of ensuring full use of the mixer when connected to a variety of input and output equipment which may also have gain controls.

References

5. Williamson, R., 'The Twin-twenty Mark 2', *Hi-Fi News,* March 1969.
6. Grossmith, B., 'A Low-distortion Integrated Amplifier', Part 2, *Hi-Fi News,* November 1969.
7. Kuo, F.F., 'Network Analysis and Synthesis', second edition Chapter 13, J. Wiley (1966).
8. Robinson, D., 'P.P.M. or VU', *Tape Recorder,* September 1967.