Audio mixer design

Audio mixers are relatively unsophisticated yet present many conflicting design requirements: headroom, noise contribution, linearity and current consumption. Resolving these stretches analogue design talent to the limit. By Richard Brice.

Once a sequence of electronic keyboards has been built up at a multi-track recording session, one may wish to add conventional instruments to the mix, for instance guitar, vocals or a saxophone solo. Since these instruments can be added live to the synthesiser parts it is both a shame to sacrifice quality and a costly luxury to use a multi-track to piece together the final touches to the mix. Far better to use a mixer to add the saxophone or vocals and guitar to the synthesiser parts and record the result as the master-tape.

The requirement for such a mixer might be: six (mic or line level) input to two output; effect-send sub-mix (in my case to an echo chamber); some form of tone control and clear metering of all inputs (including effect-return); effect-send sub-mix and stereo output.

The block diagram for the mixer filling the specification is shown in **Fig. 1.** The tone control circuit is a unity gain circuit (all controls flat) so it may be substituted after each input pre-amp to achieve this alternative configuration.

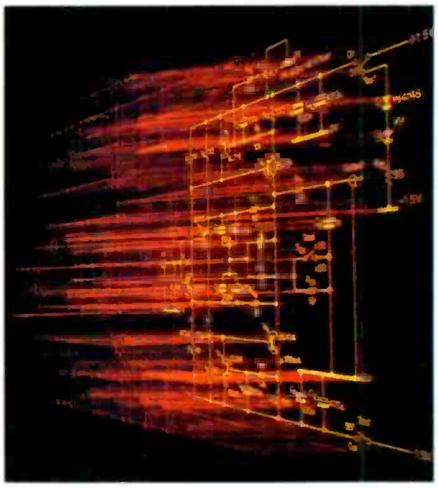
The main justification for using discrete components lies in the decision to rely on a battery supply for the mixer. A further design decision was the choice of the domestic/semi-professional signal amplitude level of 0VU, equivalent to −10dB(V). The VU is a unit intended to express the level of a complex wave in terms of decibels above or below a reference volume, and it implies a complex wave — a programme waveform with high peaks. 0VU reference level therefore refers to a complex-wave power-reading on a standard VU meter.

The usual convention is to assume that the peaks of the complex wave will be 10dB higher than the peak value of a sine wave adjusted to give the same reference reading on the VU meter. In other words, if we adjust a music or speech signal to give a reading of 0VU on a VU meters the system must have at least 10dB headroom over the level of a sine wave adjusted to give the same reading if the system is not to clip the programme audio signal.

In this mixer, 0VU is set to be equivalent to −10dB(V). The peak-to-peak value of a −10dB(V) sine wave is:

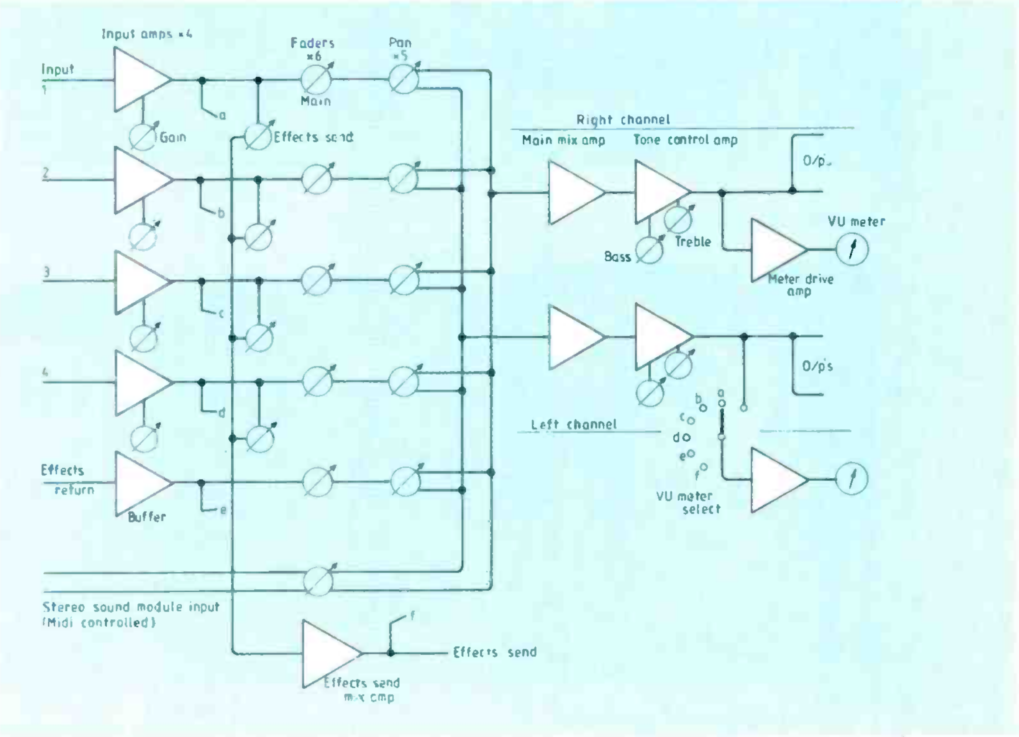
$$2\sqrt{2(1\mathrm{x}10^{-10/20})V}$$

or 894mV. A complex music or speech wave will therefore have a peak-to-peak value of 10dB (or 3.16 times) higher — 2.82V pk-pk. The mixer provides more than the stated 10dB headroom since, in practice, more headroom is occasionally necessary.



Well-designed audio circuits should certainly be able to swing 3V pk-pk when running on a single 9V rail. On the grounds of cost I decided that it would be ideal if the mixer could run from a single PP3-type battery for at least eight hours in continuous use. Low-noise op-amps like the NE5534, which would be suitable for the microphone input stages, have a typical supply current of 4mA on a 9V supply. However, the worst-case figure is

Fig. 1. Mixer block diagram.



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8mA at 25°C. This would give an Iq for the microphone input stages of 32mA which, running on a PP3 with about 100mA/h capacity, would give just three hours of use simply supplying the mic amplifiers.

Alternative op-amp series commonly used for audio, the LF355 or the TL071, have typical current drains of 1.5mA and may be considered suitable alternatives. However, they are less suitable for the mic input amplifiers. This is because (being fet input) they have a relatively large input noise voltage generator and virtually no input noise current generator, and are therefore relatively noisy when matching a low-impedance source like a microphone.

Input amplifiers

Each of the four input-stage amplifiers is formed from a transistor ring of three. The design requirement is for good headroom and a very low noise figure.

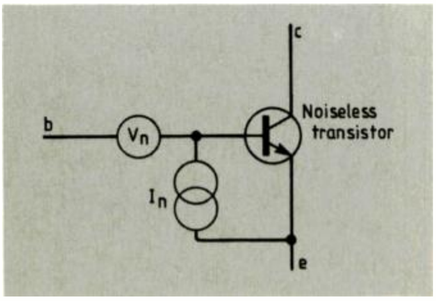


Fig. 2. The noise generator model of a bipolar transistor showing both current and voltage noise sources.

Figure 2 shows noise sources (V_n and I_n) added to a perfect transistor. V_n has the RMS value of:

$$V_n=\sqrt{4kT(r_{b'b}+\frac{1}{2g_m})B} \qquad \text{volts}$$

I_n has the RMS value of:

$$I_n=\sqrt{\frac{4kTB}{(2ß/g_m)}}$$

Because g_m appears in both equations we can sketch what will happen to V_n and I_n as the value of g_m changes (which is the same thing as saying, as collector current changes) — see **Fig. 3**. Now it is apparent that the total noise which appears as a signal at the base of the transistor will depend on the source resistance. If this is high then the noise current generated by the current noise generator must be made small because this signal will flow through the large resistance and generate a large voltage signal at the transistors base. If the source resistance is low, then the magnitude of the current generator is not important, but the magnitude of the voltage generator must be made small.

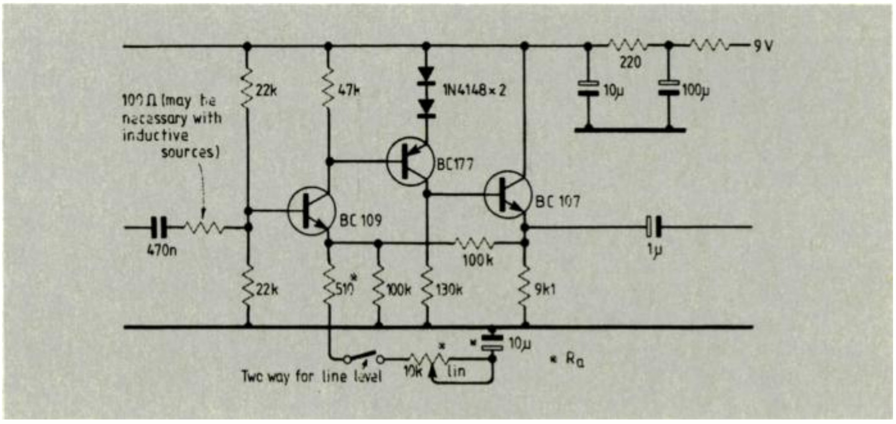


Fig. 4. Ring-of-three microphone channel amplifier.

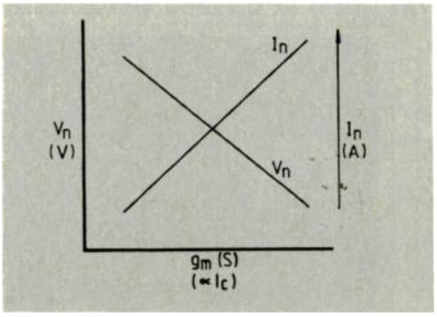


Fig. 3. Sketch of voltage and current noise sources as a function of collector current and transconductance.

Referring to **Fig. 3** we can see that a low source resistance would seem best matched by a high collector current input stage, and a high source resistance would be best matched by a low collector current input stage. The choice of optimum g_m for low noise is given in a simple equation:

$$g_m=\sqrt{ß/R_s} \qquad \text{(Ref. 1)}$$

Matching the input stage to a microphone would seem to require a collector current in mA to match the low impedance of a typical microphone. Paradoxically, in the final design (**Fig. 4**) the first-stage collector current is about 40µA. All the above would be very straightforward were it not for flicker noise. Furthermore, while it is necessary to consider only the resistive part of the source impedance when designing noise-voltage generators, the full source impedance must be taken into account when calculating the effect of the current noise generator.

Flicker noise is a type of noise that increases its magnitude with decreasing frequency, hence the alternative name of 1/f noise. It may be modelled as an increase in the current noise generator below a certain frequency (f_f). This frequency, often called the flicker-noise corner-frequency, is dependent on the transistor type and sample. The magnitude of I_n is better modelled as:

$$I_n=\frac{4kTB}{2ß[g_m(\omega_f/\omega+1)]} \qquad \text{where } \omega_f=2\pi f_f$$

The effect of 1/f noise is further complicated because not only does its magnitude decrease with decreasing collector current (as shown in the above equation) but the flicker-noise corner-frequency itself falls with decreasing collector current. Because an audio signal contains frequencies down to 20Hz, flicker noise must be considered and its reduction ensured by the choice of a low flicker-noise type transistor, like the BC109, operating at a low standing current.

Substituting in the above equations for I_n and V_n with assumptions for the following values: k = Boltzmann's constant = 1.38×10^{-23}; T = average room temperature = 290K; B = bandwidth = 20kHz. Typical figues for a BC109: Beta = 200; $r_{b'b}$ = 200Ω; g_m = 2mA/V for an emitter current of 50µA. So V_n, which is the predominant generator for a low source resistance, is:

$$V_n=\sqrt{3.2\times10^{-16}(450)}$$

or 380nV RMS or −128dB(V). The important term here is the equivalent resistance term ($r_{b'b}$ + $1/2g_m$). Its typical value of 450Ω is not much larger than the magnitude of the real part of the impedance across the output terminals of a moving-coil microphone (230Ω). This means that the thermal noise generated in the resistive part of the microphone impedance produces a significant proportion of the total input noise. If we take a typical output voltage from the microphone to be 1mV RMS then this noise will be −68dB.

The magnitude of the current noise generator is:

$$I_n = \sqrt{\frac{(3.2 \times 10^{-16}) \times (2 \times 10^{-3})}{400}}$$

or 40pA RMS.

The total calculated noise-figure for the amplifier fed from a resistive source of 200Ω (and with the gain set to maximum) is about 12dB, this being a measure of the amount by which the total noise from the amplifier exceeds what it would be if the amplifier were totally noiseless. The noise figure of the amplifier gradually deteriorates at lower gains due to the thermal noise generated in the feedback resistors.

It might be thought that the current noise generator term derived above is so vanishingly small that we do not need to consider the increase in this generator due to flicker noise. This would indeed be the case if, as in the above calculation of noise figure, the noise current simply flowed in the microphone resistance.

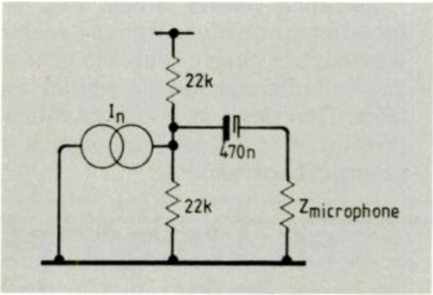


Fig. 5. Reverse view model of the microphone input stage.

However, if we draw the input stage the other way around, as in **Fig. 5**, it becomes very clear that the noise signal, which may be considered in every way just like the wanted audio signal, is AC-coupled into the microphone load. So, at frequencies below about 1.7kHz, the current generator develops its voltage signal not simply across the microphone impedance but across the potential divider which forms the bias supply for the input transistor. This has an equivalent resistance of 10kΩ, and the noise current flowing across it will generate a noise voltage signal of 124nV in a bandwidth of 1.7kHz — this figure ignores the increased effect of flicker noise at extremely low frequency.

The 10kΩ source resistance at low frequency requires an optimum operating g_m of 1.4mA/V and an operating current of 35μA; thus the final design, like so many other engineering solutions, is a compromise. A low collector current is used to ensure low flicker noise by matching the input stage to the source impedance at low frequency with some sacrifice of noise figure in the mid and high frequency bands.

Fig. 6. Mixer summing amplifier.

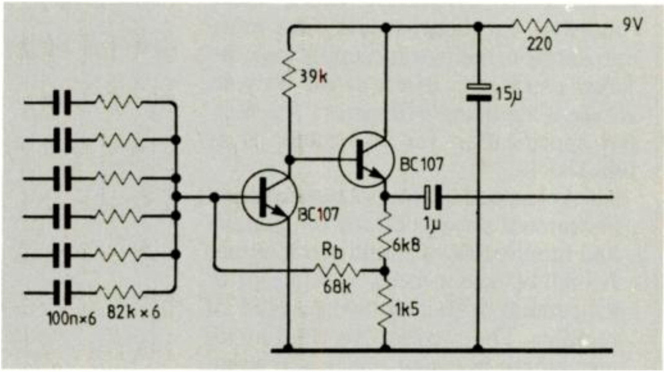


Fig. 7. Fader and panning controls.

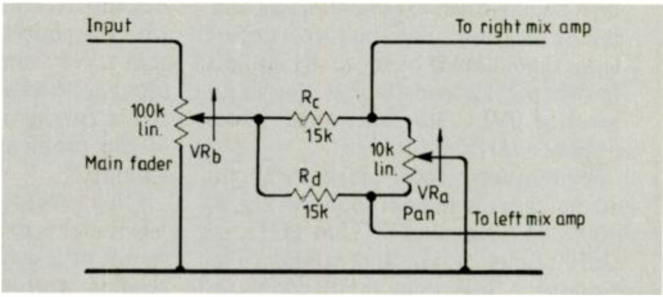
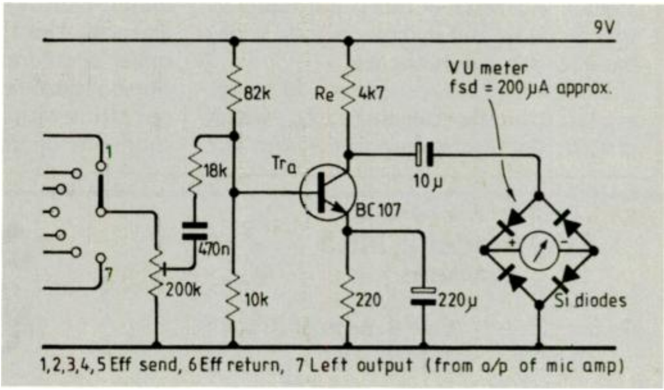


Fig. 8. VU meter circuit fed from current source to reduce effect of diode offset voltage.



The amplifier shown in Fig. 4 has a calculated input noise density of 3nV/√Hz and a calculated input noise current density of 0.3pA/√Hz, ignoring flicker noise.

The frequency and phase response remain much the same regardless of gain setting. This seems to go against the intractable laws of gain-bandwidth product: as we increase the gain we must expect the frequency response to decrease, and vice-versa. In fact, the ring-of-three circuit is an early form of the current-mode-feedback amplifier currently very popular in video applications.

The explanation for this lies in the variable gain-setting resistor R_a. This not only determines the closed-loop gain by controlling the proportion of the output voltage fed back to the inverting port, but also forms the dominating part of the emitter load of the first transistor and consequently the gain of the first stage. As the value of R_a decreases, so the feedback diminishes and the closed-loop gain rises. At the same time the open-loop gain of the circuit rises because Tr^1's emitter load falls in value. The current consumption for all four mic pre-amps is 3mA.

Mix amplifiers

The mix amplifiers shown in **Fig. 6** are based on a conventional transistor pair circuit. The only difficult decision in this area is the choice of the value for R_b. It is this value, combined with the input resistors, that determines the total contribution each input may make to the final output. In the end, I opted for a value that allowed for unity gain so that an input registering 0VU on the input meter will register 0VU on the ouput meter with the channel fader fully open. For there to be unity gain through the system there must be some gain because of the lossy fader and pan arrangement shown in **Fig. 7**.

Other circuitry

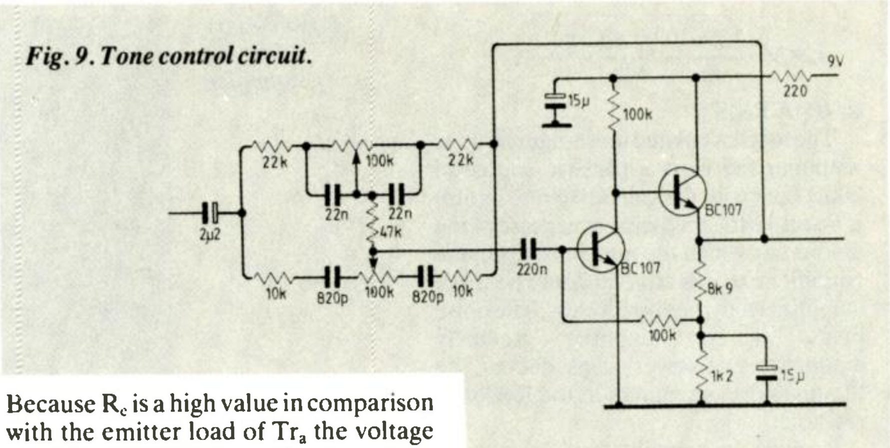
Good, clear metering is a real advantage. But led bar-graph displays and the associated ICs to drive them, while otherwise excellent, would take more current from the battery supply than the mixer electronics itself, so the obvious choice is a moving-coil meter. The written specification for VU meter is as follows:

> The volume indicator is a standardised instrument developed for the control and monitoring of sound programme. It shall be a root-mean-square type of instrument with a full-wave type of rectifier. The rectifier law shall be intermediate between linear and sqare law, having of an exponential of 1.2 ± 0.2. The sensitivity shall not depart from that at 1kHz by more than 0.2dB between 35Hz and 10kHz at an input level of 0VU, nor more than 0.5dB between 25Hz and 16kHz.

The frequency response of the circuit and meter arrangement shown in **Fig. 8**, relative to a signal of 0VU at 1kHz, is: −3dB points, 15Hz and 60kHz. The response is less than 0.5dB down at 50Hz and 20kHz — this is not really to specification, but it shows a good wideband response nonetheless.

Notice that the rectifiers and meter are fed from the collector of Tr_a, which is a current source in parallel with R_c. Because R_c is a high value in comparison with the emitter load of Tr_a the voltage gain is very small during the part of the input cycle when the rectifier diodes are not in conduction. This alleviates most of the problem of the Si diode offset voltage.

Fig. 9. Tone control circuit.



On a practical level remember that a VU meter is an indicator of the average power of a waveform; it is not a peak-reading instrument like a BBC PPM meter. Failure to appreciate this (and on a practical level this means allowing the meter needle to swing into the red section on transients) will mean the mixer is operating with inadequate system headroom. In operation the meter needles should only very occasionally swing above the 0VU reference level on complex programme. Provided input trim levels and main faders are set with this rule in mind, clipping will very rarely be encountered, even under conditions of moderate abuse!

The tone control circuit, **Fig. 9**, is based on the nearly universal Baxandall tone-control circuit, with the gain element a buffered common-emitter amplifier. The circuit values are modified to give the more gentle response at extremes of boost and cut. ■