

Fig. 5. The layout adopted for the final design.

full output may be confidently expected. The first problem to be dealt with is the very low value of R_{E2} ; this must be as low as possible (say 10ohms) as it is effectively in series with the input source resistance and will degrade the noise performance accordingly. This means that C_F must be very large, of the order of 2200 μ F, to preserve the l.f. response. A 3R3 resistor in the R_{E2} position demands 4700 μ F to give -3dB at 10Hz; this is not elegant. The capacitance C_F cannot be dispensed with, since there is a d.c. level of +0.6V on the emitter of the input device, leading to a wholly impossible offset at the output of the op-amp.

One solution to this is the use of a differential pair, as in Fig.4(e). This cancels out the V_{be} of the input transistor Tr_1 , at the cost of some degradation in the noise performance of the circuit, and hopefully the d.c. offset is so much smaller that, if C_F is omitted and the offset is amplified by the full a.c. gain, it will not seriously reduce the output voltage swing. In effect, the second transistor Tr_2 is an emitter follower transferring the feedback signal to the emitter of Tr_1 , and such a circuit element introduces a small but inescapable amount of extra noise.

In this case, with the component values shown, the degradation is about 2.8dB.

A possibly more serious objection to this circuit is that the offset at the output is non-negligible, about 1 volt, much of which is due to the base bias current flowing through R_{in} . A d.c.-blocking capacitor on the output is essential, and if it is an electrolytic there may be some doubt as to which way round to put it, as the exact level of input pair balance is unpredictable.

After practical trials, it was decided that a 3dB noise penalty was too great, and that a way had to be found to use a single-ended input.

A NEW APPROACH

The new method evolved is shown in the block diagram Fig.5. There is no C_F in the feedback loop, and indeed no overall d.c. feedback at all. The two halves of the circuit, the input transistor and the op-amp, each have their own d.c. feedback systems. The transistor relies on simple shunt negative feedback via d.c. loop 1, while the op-amp has its output held precisely to a d.c. level of 0V by the integrator A_2 . This senses the mean output level, and sets up a voltage on the non-inverting input of A_1 that is very close to the level set on Tr_1 collector, such that the output stays firmly at zero; its time-constant is made large enough to ensure that an ample amount of open-loop gain exists at the lowest audio frequencies. Failure to do this results in a rapid rise of distortion as the frequency is lowered. Any changes in the direct voltage on Tr_1 collector are completely uncoupled from the output. However, a.c. feedback passes through R_{F1} as usual and ensures that the linearity of the compound arrangement is near-perfect, as is often the case with transistor op-amp hybrid circuits. Due to the high open-loop gain of A_1 the a.c. level on Tr_1 collector is very small and so a.c. feedback through d.c. loop 1 does

not significantly affect the input impedance of the amplifier, which is about 8 k Ω .

The device chosen for the input transistor was the 2N4403, a type that has been acknowledged as superior for low-noise applications for some years. The R_b is quoted as about 40 ohms⁵. More modern purpose-designed devices such as the 2SB737 will improve the noise performance by up to 1 dB, but the extra cost is significant.

A single device used in the circuit of Fig.6 gives an e.i.n. of -138dB with a 4mA collector current, which is certainly not bad, but it was consistently found that putting devices in parallel without any current-sharing precautions whatever always resulted in a significant improvement in noise performance. On average, adding a second transistor reduced noise by 1.2dB, and adding a third reduces it by another 0.5dB. Beyond this the law of diminishing returns sets in and, since further multiplication was judged unprofitable, a triple-device input was settled on. The current-sharing under these conditions was checked by measuring the voltage across ohm resistors inserted in the collector paths. Using 3.4mA as the total current for the array it was found after much device-swapping that the worst case of imbalance was 0.97mA in one transistor and 1.26mA in another. No attempt was made to ensure that all the devices came from the same batch. It therefore appears that, for this device at least, matching is good enough to make simple paralleling worthwhile, and it was therefore decided to use three devices in parallel in the final circuit.

There now remains the problem of setting the gain. Usually it would be simple enough to alter R_{F1} or R_{E2} , but here it is not quite so simple. The resistance R_{E2} is not amenable to alteration, as it must be kept to the lowest practicable value of 3.3 ohms, and R_{F1} must be kept up to a reasonable value so that it can be driven to a full voltage swing by an op-amp output. This means a minimum of

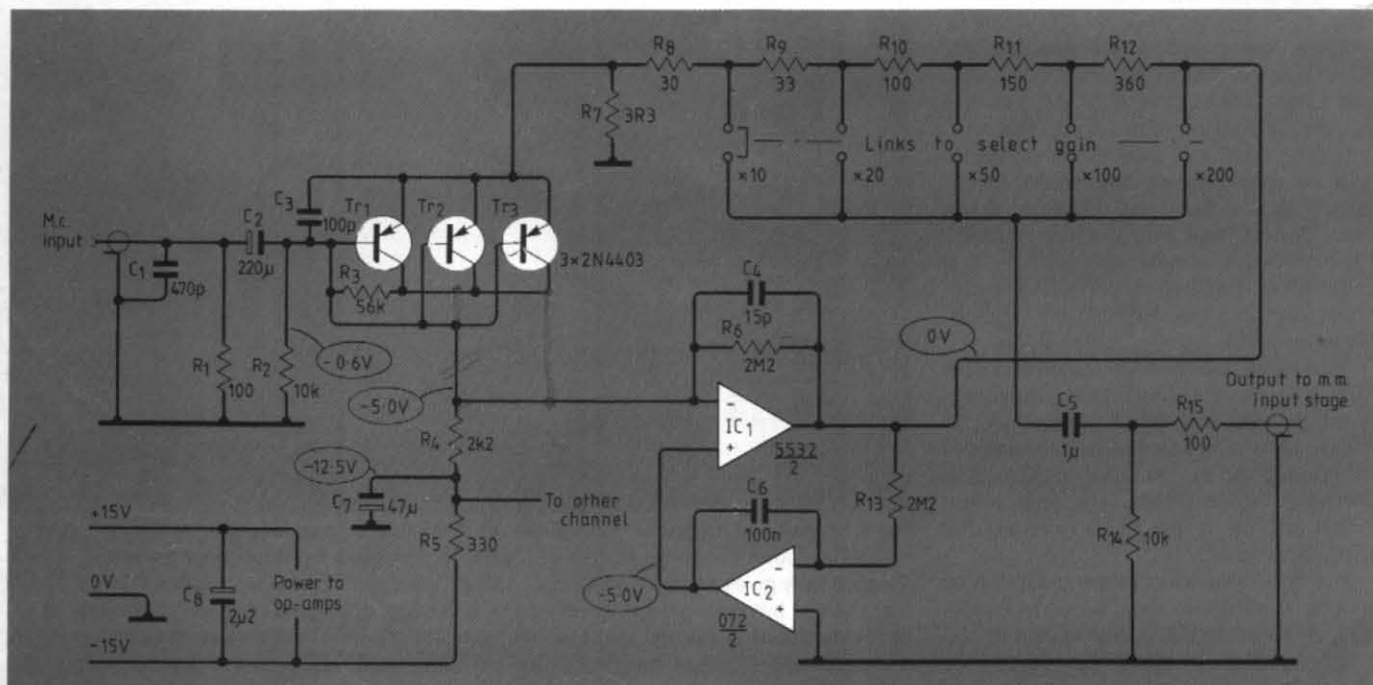


Fig. 6. Complete circuit diagram of the moving-coil head amplifier, intended to drive the moving-magnet input of a preamplifier.

500 ohms if the op-amp is to be of an easily obtainable type such as the 5534. (It is paradoxical that amplifiers whose output is measured in millivolts are required to chuck around so much current).

These two values fix a minimum closed-loop gain of about 44dB, which is far too high for all but the most insensitive cartridges. The only solution is to use a ladder output attenuator to reduce the overall gain; this would be anathema in a conventional signal path, because of the loss of headroom involved, but since an output of 300mVr.m.s. would be enough to overload virtually all m.m. inputs, we can afford to be prodigal with it. If the gain of the head amplifier is set to be a convenient $200\times$ (+46dB) then attenuation to reduce overall gain to a more useful +20dB still allows a maximum output of 480mVr.m.s.; this comfortably exceeds the input capability of the intended host preamplifier, though one previous design would accept it all and come back for more⁷. Smaller degrees of attenuation to provide intermediate gains allow greater outputs, and these are summarized in the specification. The Ortofon MC10 was used with +26dB of gain, to give similar output levels to m.m. cartridges driving the precision preamplifier RIAA stage direct.

The last constraint is the need to provide a low output impedance to the succeeding m.m. input stage, so that it can give a good noise performance; it is likely to have been optimized to give of its best with a source impedance of 500 ohms or less. This implies that the ladder attenuator will need low resistor values, imposing yet more loading on the unfortunate op-amp, so this problem has been side-stepped by making the ladder an integral part of the a.c. feedback loop, as shown in Fig.6. This is only practicable because it is known that the load resistance presented by the next stage will be too high at 47 k Ω to cause any significant gain variations.

THE FINAL CIRCUIT

This is shown in Fig.6, and most closely follows the configuration of Fig.4(d), with the exception that the input devices have suddenly multiplied themselves by three. Capacitor C_1 is soldered on the back of the m.c. input phono sockets and is intended for r.f. filtering rather than modification of the cartridge response. If the need for more capacitive or resistive loading is felt, then extra components may be freely connected in parallel with R_1 . If R_1 is raised in value, then load resistances of up to 5 k Ω are possible, as the impedance looking into C_2 is about 8k Ω . Capacitor C_2 is large to give the input devices the full benefit of the low source impedance, and its value should not be altered. Resistors R_2, R_3 make up d.c. loop 1, setting the d.c. operating conditions of $Tr_{1,2,3}$, while R_4 is the collector load, decoupled from the supply rail by C_9 and R_5 , which are shared between the two channels. Op-amp IC_1 provides the main a.c. open-loop gain, and is stabilized at h.f. by C_4 ; R_6 has no real effect on normal operation, but is included to give IC_1 a modicum of negative feedback and hence tidy behaviour at power-up, when this would otherwise be lacking

due to the charging time of C_2 , the other op-amp, IC_2 , is the integrator that makes up d.c. loop 2, its time-constant carefully chosen to provide plenty of open-loop gain from IC_1 at low frequencies, and to avoid a peaking in the l.f. response that can occur due to the second time-constant of C_2 .

The ladder resistors R_8-R_{12} make up the combined feedback-network and output-divider, overall gain being selected by a push-on link in the prototype. A rotary switch could be used instead, but this will produce loud clicks when moved with the volume up, since the emitter current of Tr_1-Tr_3 flows through R_7 , and a small current therefore flows down the divider chain. The output resistor R_{15} ensures stability when driving long screened cables, and C_5 is included to eliminate any trace of d.c. offset from the output because the stage might find itself driving a horribly vulnerable 'esoteric' input stage with direct coupling and possibly substantial gain at d.c. Anything is possible these days.

COMPARING PERFORMANCE PARAMETERS

These are given in the specification, and I think there will be few opportunities to quibble. On the vital question of noise it would be instructive to compare it with other preamplifiers – not easy because the noise performance of m.c. head amplifiers is specified in so many different ways it is virtually impossible to reduce them all to a similar form, particularly without knowing the spectral distribution of the noise. Noise performances are specified with and without CCIR-ARM weighting, over different bandwidths, and with different source impedances. This article has dealt throughout with unweighted noise referred to the input, over a 400Hz-30kHz bandwidth, and with RIAA equalisation not taken into account. Without getting bogged down in invidious comparisons, I can only say that it is my belief that the design given here is quieter than most current designs, being within 6dB of the theoretical minimum.

When using this design with the precision preamplifier, it was noted with some surprise that it was so quiet that the m.m. RIAA stage actually caused the noise performance to deteriorate by about 3dB. Since the RIAA stage is itself very quiet (s/n ratio -81dB referred to 5mVr.m.s. input it is considered that the design goals were met.

PRACTICE

P.c.b. layouts require some care if the full performance is to be realised. Firstly, the grounding should be carefully planned, as it must be realised that with such low impedances as R_7 (3R3) playing a vital role, the resistance of tracks can be significant. It is suggested that a single star ground point be chosen on the p.c.b., and critical paths (input ground, R_1, R_7) all connected to this, to prevent signal currents causing voltage drops where they are least wanted. It is vital to avoid making loops in the input path that will pick up 50Hz magnetic fields.

It is essential to place the decoupling capacitor C_8 next to IC_1 to prevent insidious

SPECIFICATION

Careful earthing is needed if the noise and crosstalk performance quoted is to be obtained.

Gain	Gain(dB)	Max output (r.m.s.)
10 \times	+20dB	480mV
20 \times	+26dB	960mV
50 \times	+34dB	2.4V
100 \times	+40dB	4.6V
200 \times	+46dB	10V

Input overload level. 48mVr.m.s.

Equivalent input noise. -139.5 dBu, unweighted, no RIAA.

T.h.d. Less than 0.002% at 7Vr.m.s. output, (maximum gain) at 1kHz. Less than 0.004% 30Hz-20kHz.

Frequency response. +0, -2 dB 20Hz-20kHz.

Crosstalk. Less than -90dB 1kHz-20kHz (layout dependent).

Power requirements.

20 mA at $\pm 15V$, for both channels.

h.f. oscillation which makes its presence known only by severely impaired linearity. When interfacing the head amplifier to an existing design, note that about 8mA flows down the ground connection.

References

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BOOKS

Dictionary of Electronics by S.W. Amos, second edition. Butterworth Scientific, 324 pages, hard cover, £25. Covers the subject from A-battery and accumulator to zener and z parameter, with extensive cross-references and numerous line illustrations. About 300 new definitions have been added to this edition and 200 existing entries have been revised. But the price seems rather high for a non-specialist title.

Industrial Control Handbook by E.A. Parr. Volume 2 (of 3): techniques. Blackwell Scientific Publications, 453 pages, hard cover, £45. Practical approach to industrial practice for the student or working engineer. Main subject headings include d.c. amplifiers, rotating machines and power electronics, computers in control, hydraulics, pneumatics and process control valves, recording and display devices, maintenance, fault-finding and safety. Volume 1 dealt with transducers and volume 3 will cover the underlying theory and applications.