Precision preamplifier '96

Part I

Douglas Self has thoroughly analysed the requirements for a no-compromise audio preamplifier making the most of today's high-performance op-amps. This first article covers the preamp's overall configuration and focuses on disc replay.

A new preamp design is timely. There is more variation in audio equipment than ever before, so to a greater extent preamps are required to be all things to all persons. High source resistance outputs and low-impedance inputs must be catered for, as well as ill-considered and exotic cabling with excessive shunt capacitance. The last preamp design I placed before the public was in 1983[1], extended in facilities by the moving-coil head amp stage published in 1987[2].

In the last ten years, small-signal analogue electronics has undergone few changes. Most circuitry is still made from *TL072*s, with resort to *5532*s when noise and drive capability are important. In this period many new op-amps have appeared, but few have had any impact on audio design; this is largely a chicken/egg problem, for until they are used in large numbers the price will not come down low enough for them to be used in large numbers. Significant advantage over the old faithfuls is required.

This new design uses the architecture established in reference 1, which has not been improved upon so far. The already low noise levels have been further reduced. The tone controls were fixed-frequency, and proved inflexible compared with the switched-turnover versions in my previous designs[3,4], so these frequencies are now fully variable, and a non-interrupting tone-cancel facility provided.

This preamplifier is designed to my usual philosophy of making it work as well as possible, by the considered choice of circuit configurations etc, rather than the alternative approach of specifying exotic components and hoping for the best.

The evolution of preamplifiers

Minimal requirements are source selection and level control, as in **Fig. 1**a; an RIAA disc

Adding tape facilities and tone control

There are two basic architectures for tape record/replay handling. The simpler, in Fig. **1d**, adds a tape output and a tape monitor switch for off-tape monitoring on triple-head machines.

The more complex version in Fig. **1e** allows any input to be listened to while any input is being recorded, though how many people actually do this is rather doubtful. This method demands very high standards of crosstalk inside the preamp. There is usually no tape return input or tape monitor switch as there is now no guarantee that the main path signal comes from the same original source as the tape output.

The final step is to add tone controls. They need a low-impedance drive for predictable equalisation curves, and a vital point is that most types – including the Baxandall – phase-invert. Since the maintenance of absolute polarity is required, this inversion can conveniently be undone by the active gain control, which also uses shunt feedback and phase-inverts. The tone-control can be placed before or after the volume control, but if afterwards it generates noise that cannot be turned down. Putting it before the volume control reduces headroom if boost is in use, but since maximum boost is only +10dB, the preamp inputs will not overload before 3Vrms is applied; domestic equipment can rarely generate such levels. Figure **1f** shows the final architecture.

preamp stage is one input option. This sort of 'passive preamplifier' (a nice oxymoron) is only practical if the main music source is a low-impedance high-level output like cd.

The only parameter to decide is the resistance of the volume pot; it cannot be too high because the output impedance, which reaches a maximum of one quarter the track resistance at –6dB, will cause high-frequency roll-off with the cable capacitance. On the other hand, if the pot resistance is too low, the source equipment will be unduly loaded. If the source is valve equipment, which does not respond well to even moderate loading, the problem starts to look insoluble.

Adding a unity-gain buffer stage after the selector switch, Fig. **1b**, means the volume control can be reduced to 10kΩ, without loading the sources. This still gives a maximal output impedance of 2.5kΩ, which allows you only 5.4 metres of 300pF/m cable before the response is 1.0dB down at 20kHz. For 0.1dB down at 20kHz, only 1.6 metres is permissible.

The input *RC* filters found on so many power-amps as a gesture against transient intermodulation distortion add extra shunt capacitance ranging from 100pF to 1000pF, and can cause additional unwanted hf rolloff.

Unfortunately only a cd source can fully drive a power amplifier. Output levels for tuners, phono amps and domestic tape machines are of the order of 150mV rms, while power amplifiers rarely have sensitivities lower than 500mV. Both output impedance and level problems are solved by adding a second amplifier stage as Fig. **1c**, this time with gain. The output level can be increased and the output impedance kept down to 100Ω or lower.

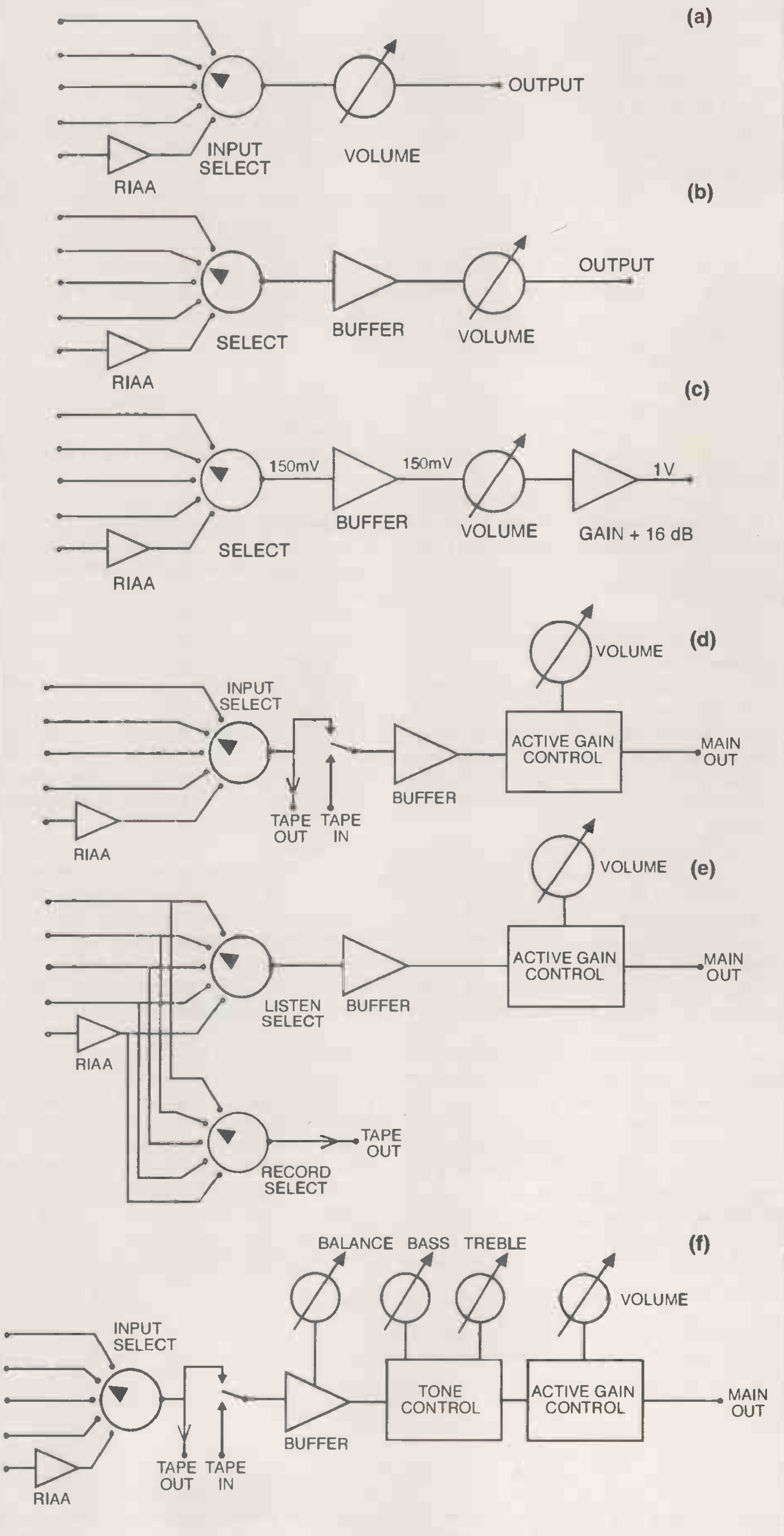
This amplifier stage introduces its own difficulties. Nominal output level must be at least 1V rms (for 150mV in) to drive most power amps, so a gain of 16.5dB is needed. If you increase the full-gain output level to 2Vrms, to be sure of driving exotica to its limits, this becomes 22.5dB, amplifying the input noise of the gain stage at all volume settings. Noise performance thus deteriorates markedly at low volume levels – the ones most of us use most of the time.

One answer is to split the gain before and after the volume control, so that there is less gain amplifying the internal noise. This inevitably reduces headroom before the volume control. Another solution is double gain controls – an input-gain control to set the internal level appropriately, then an output volume control that requires no gain after it.

Input gain controls can be separate for each channel, doubling as a balance facility[3]. However this makes operation rather awkward. No matter how attenuation and fixed amplification are arranged, there are going to be trade-offs on noise and headroom.

All compromise is avoided by an active gain stage, ie an amplifier stage whose gain is variable from near-zero to the required maximum. You get lower noise at gain settings below maximum, and the ability to generate a quasi-logarithmic law from a linear pot. This gives

Fig. 1. The course of preamp evolution, as impedance and level matching problems are dealt with.



Requirements for the RIAA network

- The RIAA network must use series feedback, as shunt feedback is 14 dB noisier.
- Correct gain at 1kHz. Sounds elementary, but you try calculating it.
- Accuracy. The 1983 model was designed for ±0.2dB accuracy 20–20kHz, which was the limit of the test gear I had access to at the time. This is tightened to ±0.05dB without using rare parts.
- It must use obtainable components. Resistors will be *E24* series and capacitors *E12* at best, so intermediate values must be made by series or parallel combinations.
- R_o (Fig. 2), must be as low as possible as its Johnson noise is effectively in series with the input signal. This is most important in moving-coil mode.
- The feedback network impedance to be driven must not be low enough to increase distortion or limit output swing – especially at high frequencies.
- The resistive path through the feedback arm should ideally have the same dc resistance as input bias resistor R_{18} (Fig. 8), to minimise offsets at A1 output. The circuitry here meets all these requirements.

excellent channel balance as it depends only on mechanical alignment.

Design philosophy

There is great freedom of design in small-signal circuitry, compared with the intractable problems of power amplification. Hence there is little excuse for a preamp that is not virtually transparent, with very low noise, crosstalk and thd.

Once all the performance imperatives are addressed, the extra degrees of freedom can be used to, say, make components the same value for ease of procurement. Opamp circuitry is used here, apart from the hybrid moving-coil stage. The great advantage is that all the tricky details of distortion-free amplification are confined within the small black carapace of a *5532*.

One route to low noise is low-impedance design. By minimising circuit resistances the contribution of Johnson noise is reduced, and hopefully conditions set for best semiconductor noise performance. This notion is not exactly new – as some manufacturers would have you believe – but has been used explicitly in audio circuitry for at least fifteen years.

In the equalisation and AGS stages, gains of much less than one are sometimes required. In these cases, avoiding the evils of attenuation-then-amplification (increased noise) and amplification-then-attenuation (reduced headroom) requires the use of a shunt feedback configuration. In the classic unity-gain stage, the shunt amplifier works at a noise gain of ×2, as opposed to unity, so using shunt feedback introduces a noise compromise at a very fundamental level.

Absolute phase is preserved for all input and outputs.

The preamp gain structure

Compared with ref. 1, the moving-magnet disc amplifier gain has been increased from +26 to +29dB (all levels are at 1kHz) to bring the line-out level up to 150mV nominal. This is done to match equipment levels that appear to have reached some sort of consensus on this value. The input buffer has a gain of +1.0dB with balance central.

The maximum gain of the AGS is therefore reduced from +26 to +22dB, to retain the same maximum output of 2V. This affects only the upper part of the gain characteristic.

Disc input

While vinyl as a music-delivery medium is almost as obsolete as wax cylinders, there remain many sizable album collections that it is impractical to either replace with cds or transfer to digital tape. Disc inputs must therefore remain part of the designer's repertoire for the foreseeable future.

The disc stage here accepts a moving-coil cartridge input of 0.1 or 0.5mV, or a moving-magnet input of 5mV. It also includes a third-order subsonic filter and the capability to drive low impedances. The moving-coil stage simply provides flat gain, of either 10 or 50 times, while the moving-magnet stage performs the full RIAA equalisation for both modes.

Moving-coil input criteria

This stage was described in detail in ref. 2. The prime requirement is a good noise figure from a very low source impedance – here 3.3Ω to comply with, for example, the Ortofon *MC10* cartridge. The circuit features

- triple low-r_b input transistors
- two separate dc feedback loops
- combined feedback-network and output-attenuator.

The very low value of R_6 means that a series capacitor to reduce the gain to unity at dc is impracticable; there is no dc feedback through R_7, R_{10} around the global loop. Local dc negative feedback via R_2, R_3 sets input transistor conditions, and dc servo IC_2 applies whatever is needed to IC_1 non-inverting input to bring $IC1$ output to 0V.

The two gains provided are 10× and 50×, so inputs of 0.5mV and 0.1mV will give 5mVrms out. The equivalent input noise of the moving-coil stage alone is –141dBu, with no RIAA. Johnson noise from a 3.3Ω resistor is –147dBu, so the noise figure is a rather good 6dB. Resistor R_6 is also 3.3Ω. This component generates the same amount of noise as the source impedance, which only degrades the noise figure by 1.4dB, rather than 3dB, as transistor noise is significant.

If discrete transistors seem like too much trouble, remember a *5532* stage here would be at least 15dB noisier.

The moving-magnet input stage

The first half of Morgan Jones's excellent preamp article[5] appeared just after this preamp design was finalised. While I thoroughly endorse most of his conclusions on RIAA equalisation, we part company on two points. Firstly, I am sure that 'all-in-one-go' RIAA equalisation as in **Fig. 2a** is definitely the best method for IC op-amp designs at least. In my design the resultant loss of high-frequency headroom is only 0.5dB at 20kHz, which I think I can live with.

Secondly, I do not accept that the difficulties of driving feedback networks with low-impedance at hf are insoluble. I quite agree that 'very few preamps of any age' meet a +28dB ref 5mV overload margin, but some exceptions are ref. 1 with +36dB, ref. 3 with +39dB, and ref. 4 with a tour-de-force +47dB. My design here gives +36dB across most of the audio band, falling to +33dB at 20kHz

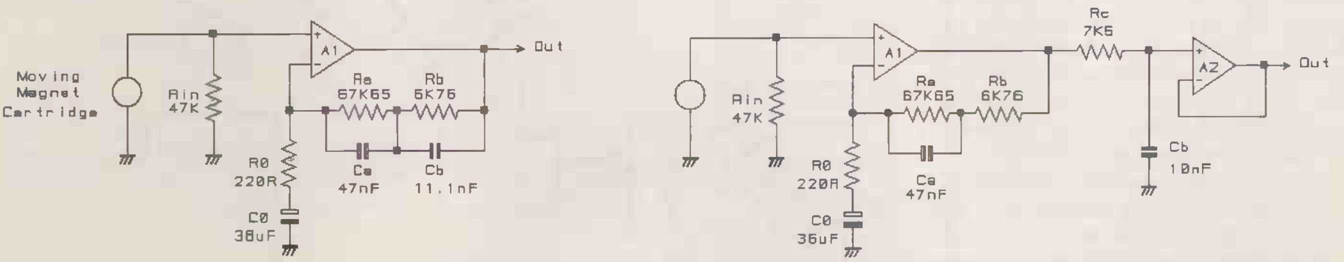


Fig. 2. The basic RIAA configurations. Fig. 2a is the standard 'all-in-one-go' series feedback configuration; the values shown do not give accurate RIAA equalisation. Fig. 2b is the most common type of passive RIAA, with a headroom penalty of 14dB at 10kHz.

(due to hf pole-correction) and +31dB at 10Hz (due to the IEC rolloff being done in the second stage).

Many contemporary disc inputs use an architecture that separates the high and low RIAA sections. Typically there is a low-frequency RIAA stage followed by a passive hf cut beginning at 2kHz, **Fig. 2b**. The values shown give a correct RIAA curve.

Amplification followed by attenuation always implies a headroom bottleneck, and passive hf cut is no exception. Signals direct from disc have their highest amplitudes at high frequencies so this passive configuration gives poor hf headroom. Overload occurs at A1 output before passive hf cut can reduce the level.

Figure 3 shows how the level at A1 output (Trace B) is higher at hf than the output signal (Trace A). Trace C shows the difference, ie the headroom loss; from 1dB at 1kHz this rises to 14dB at 10kHz and continues to increase in the ultrasonic region. The passive circuit was driven from an inverse RIAA network. Using this, a totally accurate disc stage would give a straight line just below the +30dB mark.

A related problem is that A1 in the passive version must handle a signal with much more hf content than A1 in Fig. 2a. This worsens any difficulties with slew-limiting and hf distortion. The passive version uses two amplifier stages rather than one, and more precision components.

Another difficulty is that A1 is more likely to run out of open-loop gain at hf. This is because the response plateaus above 1kHz, rather than being steadily reduced by increasing negative feedback. Passive RIAA is not an attractive option.

Alternatively there may be a flat input stage followed by a passive hf cut and then another stage to give the lf boost, which has even more headroom problems and uses yet more bits. The 'all-in-one-go' series feedback configuration in Fig. 2a avoids unnecessary headroom restrictions and has the minimum number of stages.

In search of accurate RIAA

I have a deep suspicion that such popularity as passive RIAA has is due to the design being much easier. The time-constants are separate and non-interactive; only the simplest of calculations are required.

In contrast the series-feedback system in Fig. 2a has serious interactions between its time-constants and design by calculation is complex. The values shown in Fig. 2a are what you get if you ignore the interactions and simply implement the time-constants as $R_a \times C_a$ equals 3180µs, $R_b \times C_a$ equals 318µs, and $R_b \times C_b$ equals 75µs. The resulting errors are ±0.5dB ref 1kHz.

Empirical approaches (cut-and-try) are effective if great accuracy is not required, but attempting to reach even ±0.2dB by this route becomes very tedious and frustrating. Hence the Lipshitz equations[6] have been converted to a spreadsheet, and used to synthesise the

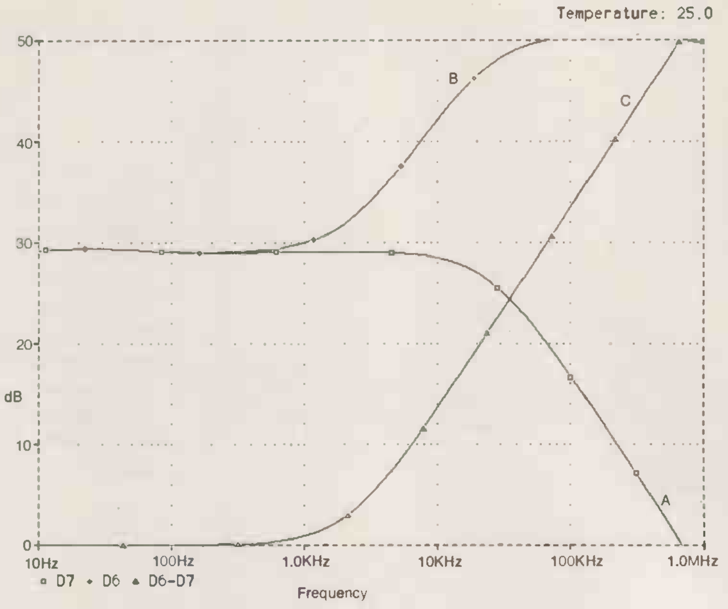


Fig. 3. Headroom loss with passive RIAA equalisation. The signal at A1 (Trace B) is greater than A2 (Trace A) so overload occurs there. The headroom loss is plotted as Trace C.

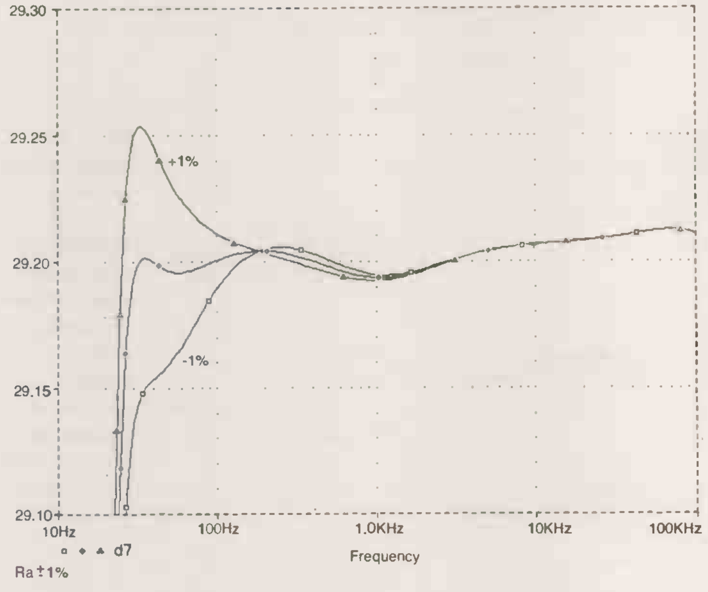


Fig. 4. The effect on RIAA accuracy of a ±1% variation in R_a. Worst-case is 0.05dB, only significant below 100Hz.

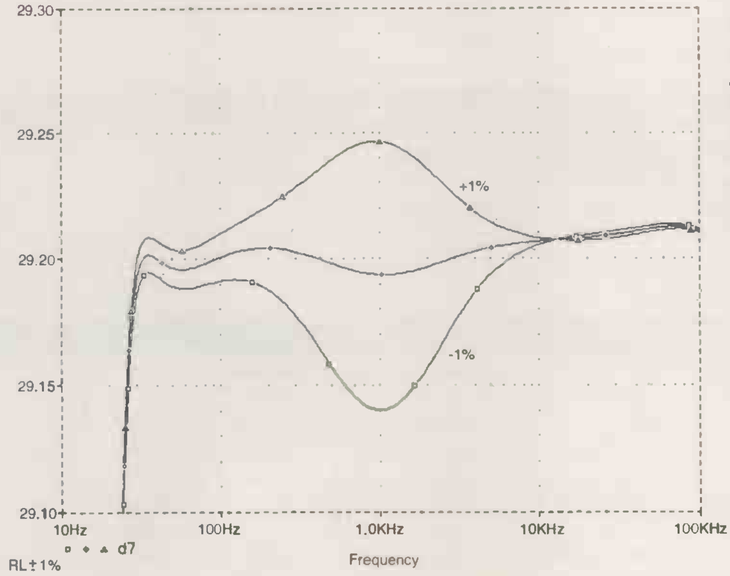


Fig. 5. The effect on RIAA accuracy of a ±1% variation in R_b. Worst-case 0.05dB around 1kHz.

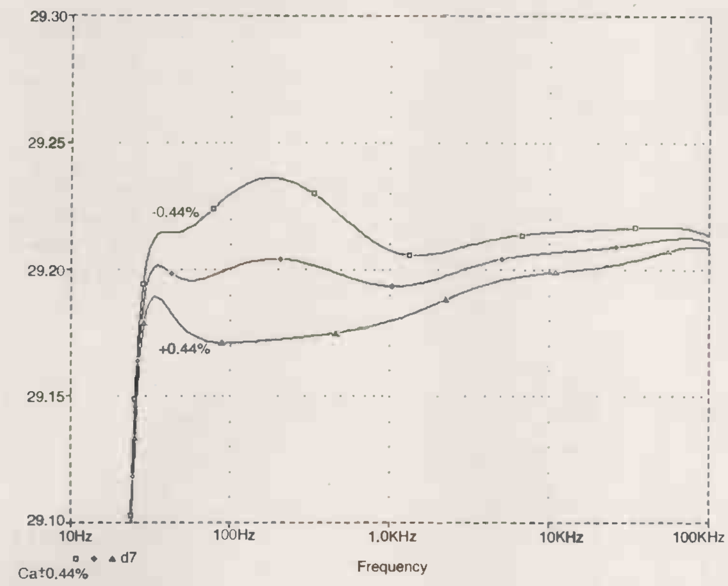


Fig. 6. The effect on RIAA accuracy of a ±0.44% variation in C_a. Effect is less than ±0.05dB at low frequencies, with a small effect on the upper audio band.

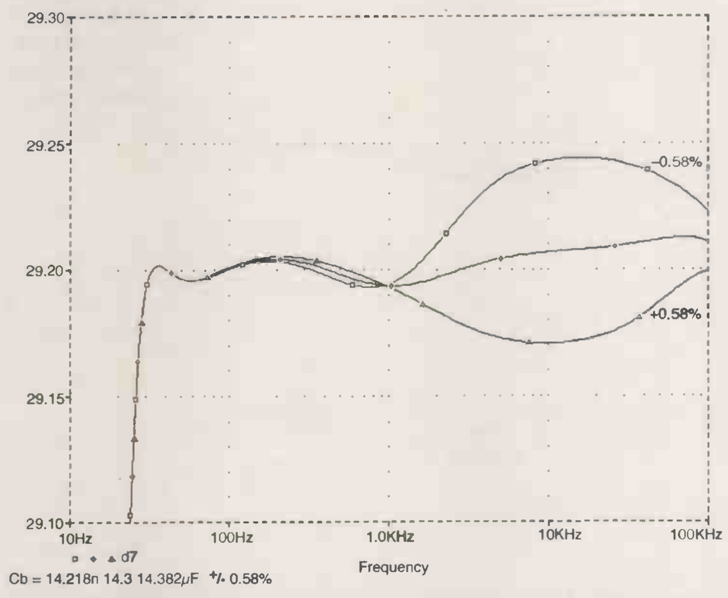


Fig. 7. The effect on RIAA accuracy of a ±0.58% variation in C_b. Effect is less than ±0.05dB on top four octaves. Smaller variation is permissible in the capacitors for the same RIAA error.

Table 1. Measured noise results, showing the *5532's* superiority.

Z_{source}	*TL072*	*5532*	*5532* benefit	*5532* EIN
1k	–88.0	–97.2dBu	+9.8dB	–126.7dBu
Shure M75ED	–87.2	–92.3dBu	+5.1dB	–121.8 dBu

(Preamp gain +29.55dB at 1kHz. Bandwidth 400-22kHz, rms sensing)

Table 2. Calculated minimum noise results.

Case	e_n nV/√Hz	i_n pA/√Hz	R_{in}	R_0	Output dBu	S/N ref 5mV dB	EIN dBu	
1 Noiseless amp	0	0	1000M	0R	–104.0	–89.7dB	–133.5	A
5 Noiseless amp	0	0	47k	0R	–97.1	–82.8dB	–126.5	C
7 Noiseless amp	0	0	47k	220R	–96.7	–82.4dB	–126.2	
11 *2SB737*, I_c=70μA	1.7	0.4	47k	220R	–95.3	–81.0dB	–124.8	
16 *5532*	5	0.7	47k	220R	–92.5	–78.2dB	–122.0	
18 *TL072*	18	0.01	47k	220R	–86.9	–72.6dB	–116.5	

design in **Fig. 8**.

A great deal of rubbish has been talked about RIAA equalisation and transient response, in perverse attempts to render the shunt RIAA configuration acceptable despite its crippling 14dB noise disadvantage. The heart of the matter is that the RIAA replay characteristic apparently requires the hf gain to fall at a steady 6dB/octave forever. A series-feedback disc stage with relatively low gain cannot make its gain fall below one, and so the 6dB/octave fall tends to level out at unity early enough to cause errors in the audio band. Adding a high-frequency correction pole – ie low-pass time constant – just after the input stage makes the simulated and measured frequency response identical to a shunt-feedback version, and retains the noise advantage.

At this level of accuracy, the finite gain open-loop gain of even a *5534* at hf begins to be important, and the frequency of the hf pole is trimmed to allow for this.

What RIAA accuracy is possible without spending a fortune on precision parts? The best tolerance readily available for resistors and capacitors is ±1%, so at first it appears that anything better than ±0.1dB accuracy is impossible. Not so. The component-sensitivity plots in **Figs 4, 5** show the effect of 1% deviations in the value of R_a, R_b; the response errors never exceed 0.05dB, as there are always at least two components contributing to the RIAA response.

Sensitivity of the RIAA capacitors is shown in **Figs 6, 7** and you can see that tighter tolerances are needed for C_a and C_b, than for R_a and R_b to produce the same 0.05dB accuracy. The capacitors have more effect on the response than the resistors.

Finding affordable close-tolerance capacitors is not easy; the best solution seems to be, as in 1983, axial polystyrene, available at 1% tolerance. These only go up to 10nF, so some parallelling is required, and indeed turns out to be highly desirable. The resistors are all 1%, which is no longer expensive or exotic, though anything more accurate certainly would be.

For C_a, the five 10nF capacitors in parallel reduce the tolerance of the combination to 0.44%. This statistical trick works because the variance of equal summed components is the sum of the individual variances. Thus for five 10nF capacitors, the standard deviation (square root of variance) increases only by the square root of five, while total capacitance has increased five times. This produces an otherwise unobtainable 0.44% close-tolerance 50nF capacitor.

Similarly, C_b is mainly composed of three 4n7 components and its tolerance is improved by root-three, to 0.58%.

Noise considerations

The noise performance of any input stage is ultimately limited by Johnson noise from the input source resistance. The best possible equivalent input noise data for resistive sources, for example microphones with a 200Ω source resistance, i.e. –129.6dBu, is well-known, but the same figures for moving-magnet inputs are not.

It is particularly difficult to calculate equivalent input noise for moving magnet stages as a highly inductive source is combined with the complications of RIAA equalisation[7]. The amount by which a real amplifier falls short of the theoretical minimum equivalent input noise is the noise figure, NF. I often wonder why noise figures are used so little in audio; perhaps they are a bit too revealing.

The noise performance of disc input stages depends on the input source impedance, the cartridge inductance having the greatest influence. It is vital to realise that no value of resistive input loading will give realistic noise measurements.

A 1kΩ load models the resistive part of the cartridge impedance. But it ignores the fact that the 'noiseless' inductive reactance makes the impedance seen at the preamp input rise very strongly with frequency, so that at higher frequencies most of the input noise actually comes from the 47kΩ loading resistance. I am grateful to Marcel van de Gevel[8] for drawing my attention to this point.

Hence, for the lowest noise you must design for a higher impedance than you might think, and it is fortunate that the RIAA provides a treble roll-off, or the noise problem would be even worse than it is. This is not why it was introduced. The real reason for pre-emphasis/de-emphasis was to discriminate against record surface noise. **Table 1** shows the two most common audio op-amps, the *5532* being definitely the best and quieter by 5dB.

To calculate appropriate EINs, I built a spreadsheet mathematical model of the cartridge input, called MAGNOISE. The basic method is as in ref. 9. The audio band 50-22kHz is divided into nine octaves, allowing RIAA equalisation to be applied, and the equivalent generators of voltage noise (e_n) and current noise (i_n) to be varied with frequency.

Noise generated by the 47kΩ resistor R_{in} is modelled separately from its loading effects so its effect can be clearly seen. I switched off the bottom three octaves to make the results comparable with real cartridge measurements that require a 400Hz high-pass filter to eliminate hum, and 1/f effects are therefore neglected. No psychoacoustic weighting was used, and cartridge parameters were set to 610Ω+470mH, the measured values for the Shure *M75ED*.

The results match well with my *5532* and *TL072* measurements, and I think the model is a usable tool. **Table 2** shows some interesting cases; output noise is calculated for gain of +29.55dB at 1kHz, and signal-to-noise ratio for a 5mVrms input at 1kHz.

I draw the following conclusions. The minimum equivalent input noise from this particular cartridge, without the extra thermal noise from the 47kΩ input loading, is –133.5dBu, no less than 7dB quieter than the loaded cartridge. (Case 1) It is the quietest possible condition. The noise difference between 10MΩ and 1MΩ loading is still 0.2dB, but as loading resistance is increased further to 1000MΩ the EIN asymptotes to –133.5dBu. A 47kΩ loading is essential for correct cartridge response.

With 47kΩ load, the minimum EIN from this cartridge is –126.5dBu. (Case 5) All other noise sources, including R_0, are ignored. This is the appropriate noise reference for this preamp design.

Resistor R_0, the 220Ω resistor in the bottom arm of negative feedback network, adds little noise. The difference between Case 5 and Case 7 is only 0.3dB.

A disc preamp stage using a good discrete bipolar device such as the remarkable *2SB737* transistor (r_b only 2Ω typ) is potentially 2.8dB quieter than a *5532*, when the noise from R_0 and the input load are included. Compare Cases 11 and 16.

The calculated noise figure for a *5532* is 4.5dB. Measured noise output of the moving magnet stage is –92.3dBu (1kHz gain +29.5dB) and so the equivalent input noise is –121.8dBu, and the real noise figure is 4.7dB,

Filtering subsonics

This stage is a third-order Butterworth high-pass filter, modified for a slow initial rolloff that implements the IEC amendment. This is done by reducing the value of $R_{27}+R_{28}$ below that for maximal flatness. The stage also buffers the high-frequency correction pole, and gives the capability to drive a 600Ω load, if you can find one.

Capacitor distortion[10] in electrolytics is – or should be – by now a well-known phenomenon. It is perhaps less well known that non-electrolytics can also generate distortion in filters like these. This has nothing to do with Subjectivist musicality, but is very real and measurable.

The only answer appears to be using the highest-voltage capacitors possible; 100V polyester generates ten times less distortion than the 63V version.

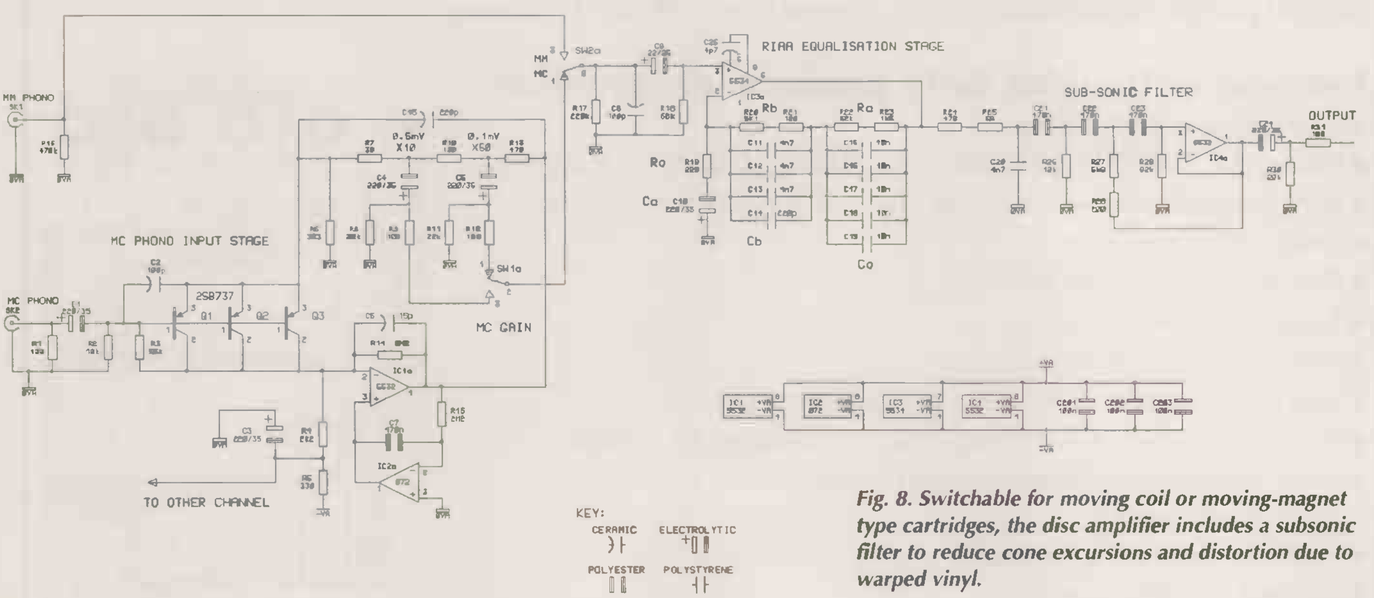


Fig. 8. Switchable for moving coil or moving-magnet type cartridges, the disc amplifier includes a subsonic filter to reduce cone excursions and distortion due to warped vinyl.

which is not too bad. Noise from the subsonic filter is negligible.

Taking e_n and i_n from data books, it looks as though the *5534/5532* is the best op-amp possible for this job. Other types – such as *OP-27* – give slightly lower calculated noise, but measure slightly higher. This is probably due to extra noise generated by bias current-cancellation circuitry[8].

There is an odd number of half-*5532*s, so the single *5534* is placed in the moving-magnet stage, where its slightly lower noise is best used. The RIAA-equalised noise output from the disc stage in moving-coil mode is –93.9dBu for 10× times gain, and –85.8dBu for 50× times. In the 10× case the moving-coil noise is actually 1.7dB lower than moving-magnet mode.

Circuit details

The complete circuit of the disc amplifier and subsonic filter is Fig. 8. Circuit operation is largely described above, but a few practical details are added here. Resistors R_9 and R_{12} ensure stability of the moving-coil stage when faced with moving-magnet input capacitance C_8, while R_8 and R_{11} are dc drains.

The *5534* moving-magnet stage has a minimum gain of about 3×, so compensation should not be required; if it is, a position is provided (C_{26}) for external capacitance to be added; 4.7pF should be ample. The moving-magnet stage feedback arm R_{20-23} has almost exactly the same dc resistance as the input bias resistor R_{18}, minimising the offset at the output of IC_3. The hf correction pole is R_{24}+R_{25} and C_{20}.

Capacitor C_{24} is deliberately oversized so low loads can be driven. Resistor R_{31} ensures stability into high-capacitance cables. ■

References

1. D. Self, 'A precision preamplifier', *Wireless World*, Oct 1983, p 31.
2. D. Self, 'Design of moving-coil head amplifiers', *Electronics & Wireless World*, Dec 1987, p 1206.
3. D. Self, 'An advanced preamplifier design', *Wireless World*, Nov 1976, p 41.
4. D. Self, 'High-performance preamplifier', *Wireless World*, Feb 1979, p 43.
5. M. Jones, 'Designing valve preamps', *Electronics World*, March 1996, p 192.
6. S. Lipshitz, 'On RIAA equalisation networks', *Journ Audio Eng Soc*, June 1979, p 458.
7. H.P. Walker, 'Low noise amplifiers', *Wireless World* May 1972, p 233.
8. M van de Gevel, Private communication, Feb 1996.
9. J. Sherwin, 'Noise specs confusing?', National Semiconductor Application Note AN-104, Linear Applications Handbook 1991.
10. D. Self, 'Capacitor distortion', 'Views', *HiFi News & RR*, Nov 1985, p 23.