# Noise and moving-magnet cartridges

**Marcel van de Gevel investigates noise optimisation in RIAA amplifiers for moving magnet cartridges and presents a high-performance design example.**

As you can read in any textbook on low-noise design, the noise of an amplifier can be described with two equivalent input noise sources. These are an equivalent input noise voltage source in series with the input of the amplifier and an equivalent input noise current source shunted across its input[1].

Which of the two has the largest influence depends on the impedance of the signal source driving the input. For example, when the magnitude of the source impedance is 12kΩ, a noise current of 1pA/√Hz has the same influence as a noise voltage of 1pA/√Hz×12kΩ=12nV/√Hz.

Unfortunately, some of the measures that a designer can take to reduce the equivalent input noise-voltage density lead to an increase of the equivalent input noise current density and vice versa. This applies especially to the selection of the input device and its biasing conditions.

It is important for a designer of low-noise electronics to know the source impedance, so that he or she can choose the device and biasing that gives the smallest total noise for that specific source impedance. Because noise optima are usually rather broad, there is no need to know the source impedance with a high accuracy; a reasonable estimate will do.

In the case of RIAA amplifiers for moving magnet (MM) cartridges, the source impedance varies enormously over the band of interest. A typical cartridge with 1kΩ of DC resistance and 0.5H of inductance has an impedance of 1kΩ at low frequencies and almost 63kΩ at 20kHz. The question is: for what impedance should a RIAA amplifier be optimised if we want to get the smallest amount of audible noise? It will be shown in this article that the cartridge impedance at 3852Hz is a good estimate. Also, an easy way to account for 1/*f*-noise will be given.

When the input stage and feedback network are properly designed, the thermal noise of the 47kΩ resistor shunted across the input is usually the largest remaining noise contribution in the RIAA amplifier itself.

Just leaving this resistor out is not an option, as this would reduce the damping of the resonant circuit consisting of the cartridge inductance and the load capacitance, seriously affecting the response between 10kHz and 20kHz. However, a lower noise 47kΩ input resistance can be realised by using combinations of series and parallel feedback.

This technique is also known as the active input impedance or 'electronic cooling' technique. Two examples of such RIAA amplifiers, one discrete circuit and one with op-amps, will be given.

Finally, the improvement which can be obtained by replacing the 47kΩ resistor with an active input resistance will be estimated for a typical cartridge.

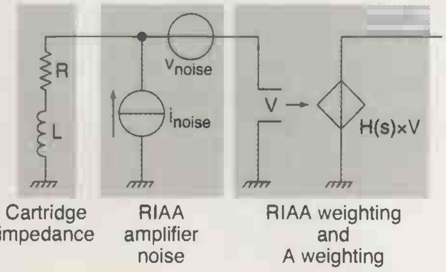



*Fig. 1. Simple model of a cartridge connected to an RIAA amplifier.*

## The 3852Hz rule

**Figure 1** is a simple model of a cartridge connected to an RIAA amplifier, followed by an A-weighting filter. The L-R series network models the impedance of a moving magnet cartridge with inductance L and DC resistance R.

The noise of the RIAA amplifier is represented by the noise voltage source $v_{noise}$ and the noise current source $i_{noise}$. These noise sources are assumed to be white and uncorrelated. The noise of the cartridge itself is not included in this model, because the amplifier designer cannot optimise this anyway. The voltage controlled voltage source with transfer H($s$) models the RIAA- and the A-weighting.

Note that a real cartridge has a frequency-dependent effective series resistance, unlike a simple L-R series network. However, this mainly affects the phase of the cartridge impedance, while only the magnitude will be used in the following calculation.

The contribution of $v_{noise}$ to the total integrated noise at the output is:

$$A = \int_0^\infty |H(j2\pi f)|^2 S(v_{noise}) df = \int_0^\infty |H(j2\pi f)|^2 df \bullet S(v_{noise}) \quad (1)$$

The second equality sign applies when $S(v_{noise})$ is independent of frequency, that is, when the noise is white. $S(v_{noise})$ is the noise voltage spectral density, that is, the number of squared volts per hertz. A is the part of the mean-square noise voltage (square of the RMS noise voltage) at the output which is caused by $v_{noise}$. $|H(j2\pi f)|$ is the magnitude of the RIAA and A filter transfer at frequency $f$.

The contribution of $i_{noise}$ equals:

$$B = \int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) S(i_{noise}) df = \int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) df \bullet S(i_{noise}) \quad (2)$$

assuming the current noise to be white. $S(i_{noise})$ is the noise-current spectral density, that is, the number of squared amperes per hertz. $B$ is the part of the mean-square noise voltage (square of the RMS noise voltage) at the output which is caused by $i_{noise}$.

Hence,

$$\frac{B}{A} = \frac{\int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) df}{\int_0^\infty |H(j2\pi f)|^2 df} \bullet \frac{S(i_{noise})}{S(v_{noise})} = \left( \frac{\int_0^\infty |H(j2\pi f)|^2 f^2 df}{\int_0^\infty |H(j2\pi f)|^2 df} \bullet 4\pi^2 L^2 + R^2 \right) \frac{S(i_{noise})}{S(v_{noise})} \quad (3)$$

At a single frequency $f=f_1$, the ratio of the contribution of $i_{noise}$ to the contribution of $v_{noise}$ equals:

$$\frac{D}{C} = (4\pi^2 f_1^2 L^2 + R^2) \bullet \frac{S(i_{noise})}{S(v_{noise})} \quad (4)$$

Here, $D$ is the contribution of $i_{noise}$ and $C$ is the contribution of $v_{noise}$ to the total noise density at $f_1$.

Equations (3) and (4) are equal when:

$$f_1 = \pm \sqrt{\frac{\int_0^\infty |H(j2\pi f)|^2 f^2 df}{\int_0^\infty |H(j2\pi f)|^2 df}} \quad (5)$$

Equation (5) can be evaluated numerically. This can easily be done with PSpice and Probe, using an analogue behavioural model of an idealised RIAA amplifier with an A-weighting filter behind it.

You can first simulate the AC transfer and use Probe to evaluate the denominator integral. Then put a differentiator in the circuit with unity transfer at 1Hz, and do the same for the numerator. The result is $f_1 \approx 3852$Hz. I used the IEC modified RIAA response, which has an extra roll-off below 20Hz, but this should make very little difference.

This proves that the total integrated weighted noise is a constant factor times the noise in a 1Hz bandwidth around 3852Hz. 'A constant factor' means that this factor is independent of $S(v_{noise})$ and $S(i_{noise})$. Hence, the combination of $S(v_{noise})$ and $S(i_{noise})$ which is optimal at 3852Hz is also optimal for the integrated weighted noise.

In other words, optimising the integrated A- and IEC modified RIAA-weighted noise is the same as optimising the noise at 3852Hz. When $R$=1kΩ and $L$=494mH, which are typical values for a MM cartridge, the impedance is about 12kΩ at this frequency. We will therefore assume a 12kΩ cartridge impedance in the remainder of this article, unless otherwise noted.

## 1/f-noise and noise densities relationships

The calculations above apply only to white noise, while many active devices produce substantial $1/f$-noise in the audio frequency band. To what white noise level does a given amount of $1/f$-noise correspond? Further, if you measure the total A-weighted noise level, how can you compare the results to the theoretical noise densities?

Calculations which are similar to those in the previous section show that $1/f$ voltage noise has the same influence on the total RIAA/IEC- and A-weighted noise level as white voltage noise with the same density at 1169Hz. In other words, if you determine the noise voltage density at 1169Hz, it doesn't matter what part is white and what part is $1/f$ noise.

The situation is a bit more complicated for $1/f$ current noise. If the cartridge impedance were constant, the same rule that applies to $1/f$ voltage noise would also apply to $1/f$ current noise. If the magnitude of the cartridge impedance were exactly proportional to frequency, a similar rule would apply, but at a frequency of 5398Hz instead of 1169Hz.

The frequency dependence of the magnitude of the impedance of a real cartridge is something in between these two cases, closer to being proportional to frequency than to being constant. Hence, the current noise density should be determined at a frequency of about 5kHz.

In order to be able to compare measured A- and RIAA/IEC-weighted RMS noise levels to theoretical noise voltage and current densities, one needs to know the noise bandwidth of the cascade of a RIAA/IEC and an A filter. Simulations show that this bandwidth is 3219Hz referred to the transfer at 1kHz. In other words, if one divides the A-weighted RMS output noise voltage by the voltage gain at 1kHz and by the square root of 3219Hz, one finds the average equivalent input number of volts per root Hz.

In order to find the equivalent input noise voltage and the equivalent input noise current, one can measure the output noise voltage twice, once with short-circuited and once with open input.

The equivalent input noise voltage per root hertz can then be calculated directly from the measurement with shorted input, as explained above. As long as the input capacitance is so small that the input impedance can be assumed to be 47kΩ over the entire audio band, the equivalent input noise current is simply the input noise voltage measured with open input, divided by 47kΩ.

## 'Electronic cooling'

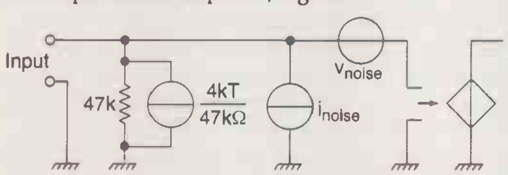
The thermal noise of a resistor can be modelled with a noise voltage source with a spectral density of $4kTR$ (RMS value $\sqrt{(4kTR\Delta f)}$ over a bandwidth Δf) in series with a noiseless resistor (the Thévenin equivalent). In this equation, $T$ is the absolute temperature and k is

Boltzmann's constant ($1.38065\times10^{-23}$J/K).

Alternatively, the noise can be modelled with a noise current source with a spectral density $4kT/R$ (RMS value $\sqrt{(4kT/R)\Delta f}$ over a bandwidth $\Delta f$) in parallel with a noiseless resistor (the Norton equivalent). Both models behave in exactly the same way at their terminals. A designer can choose whichever model is the most convenient for his or her calculations.

Modelling the 47kΩ resistor with its Norton equivalent, it becomes clear that this resistor contributes $4kT/47k\Omega\approx3.445\times10^{-25}A^2/\sqrt{Hz}$, or 0.5869pA/√Hz, to the equivalent input noise current, **Fig. 2**. With a typical effective cartridge impedance of 12kΩ, 0.5869pA/√Hz has the same influence as 7.043nV/√Hz, which is substantial compared to a carefully optimised input stage.

Fortunately, there's a simple solution. Build an accurate and low-noise inverting amplifier with a high input impedance and a voltage gain of –K times. Then connect a resistor with a value of (K+1)×47kΩ between the input and output of this amplifier, **Fig. 3**.




Note: $i_{noise}$ is noise from the rest of the RIAA amplifier.

*Fig. 2. RIAA amplifier with a noisy 47kΩ resistor.*

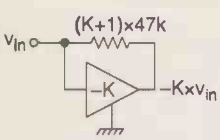



*Fig. 3. Low-noise active input resistance.*

The voltage across the resistor is K+1 times the input voltage and its value is K+1 times 47kΩ. Hence, the current flowing through the resistor equals the current that would flow through a 47kΩ resistor connected between the input and ground.

If the noise of the inverting amplifier is negligible though, the noise current spectral density is K+1 times as small. This technique is known as the active input impedance technique or as 'electronic cooling', as it has the same influence on the thermal noise as cooling the resistor down to a K+1 times lower absolute temperature.

Both terms can cause confusion, the first because it has a completely different meaning in network theory and the second because the resistor is not really cooled down.

## Discrete RIAA amplifier

**Figure 4** depicts a circuit in which the inverting amplification and the RIAA equalised amplification are combined[2]. Block $N_1$ is a two-port with very high voltage, current, transimpedance and transadmittance gain.

Because of the negative feedback around this high-gain block, the voltage $v_{in}$ at the amplifier input also occurs at $R_1$. As a result, a current $i=v_{in}/R_1$ flows through $R_1$, through the RIAA equalising network $R_2$, $C_2$, $R_3$, $C_3$ and through the output port of the amplifying block. Resistors $R_1$ and $R_4$ are normally made much smaller than $R_5$. Hence, the voltage on the upper side of $R_4$ becomes approximately,

$$-R_4 \times i = -\frac{R_4}{R_1} v_{in}$$

realising a negative voltage gain $-R_4/R_1$.

A more accurate analysis shows that if the amplifying block has infinite gain and if the current flowing into one output pin exactly equals the current coming out of the other, the input resistance equals:

$$R_{in} = \frac{R_4 + R_5}{\frac{R_4}{R_1} + 1} \tag{6}$$

Unfortunately, the input impedance changes when a load

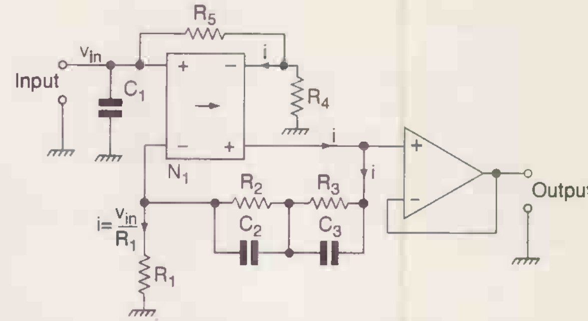



*Fig. 4. RIAA equalised amplifier with active input resistance according to reference 2*

is connected to the output of the two-port. That is why there is a voltage follower connected behind the RIAA equalising stage.

A discrete implementation is shown in **Fig. 5**. I use it as a part of a wholly discrete preamplifier with ±14V regulated power supply voltages, but it should also work at the more usual ±15V. It has an IEC modified RIAA response, that is, it includes a first-order roll-off below 20Hz.

If desired, this corner frequency can be lowered by increasing the values of the bipolar electrolytic capacitors, although lowering it too much can make the start-up time rather long. The input resistance increases from 47kΩ to 10MΩ for subsonic frequencies. This is no problem, as the purpose of the 47kΩ input resistance is to damp a resonance in the 10kHz to 20kHz frequency range.

The voltage follower at the output is a simple emitter follower. The high-gain block is made of two differential pairs. For reasons of noise optimisation, the input pair is asymmetrical, consisting of a JFET and a bipolar transistor.

It can be shown[1] that the optimal bias current for minimum noise is,

$$I_{C,opt} \approx \frac{\frac{kT}{q}\bullet\sqrt{h_{FE}}}{|Z_S + r_b|} \tag{7}$$

when a bipolar transistor is driven from a source impedance $Z_S$, assuming that $1/f$ noise is negligible and that the frequency lies well below $f_T/\sqrt{h_{FE}}$. In this equation, $q$ is the electron charge ($1.6022\times10^{-19}$C), $h_{FE}$ the DC current gain factor and $r_b$ the parasitic base resistance of the transistor. The corresponding transconductance equals,

$$g_{m,opt} \approx \frac{\sqrt{h_{FE}}}{|Z_S + r_b|} \tag{8}$$

and the total contribution to the noise is,

$$S(v_{n,bip,inc.currentnoise}) \approx 4kT\left(\frac{|Z_s + r_b|}{\sqrt{h_{FE}}} + r_b\right) \tag{9}$$

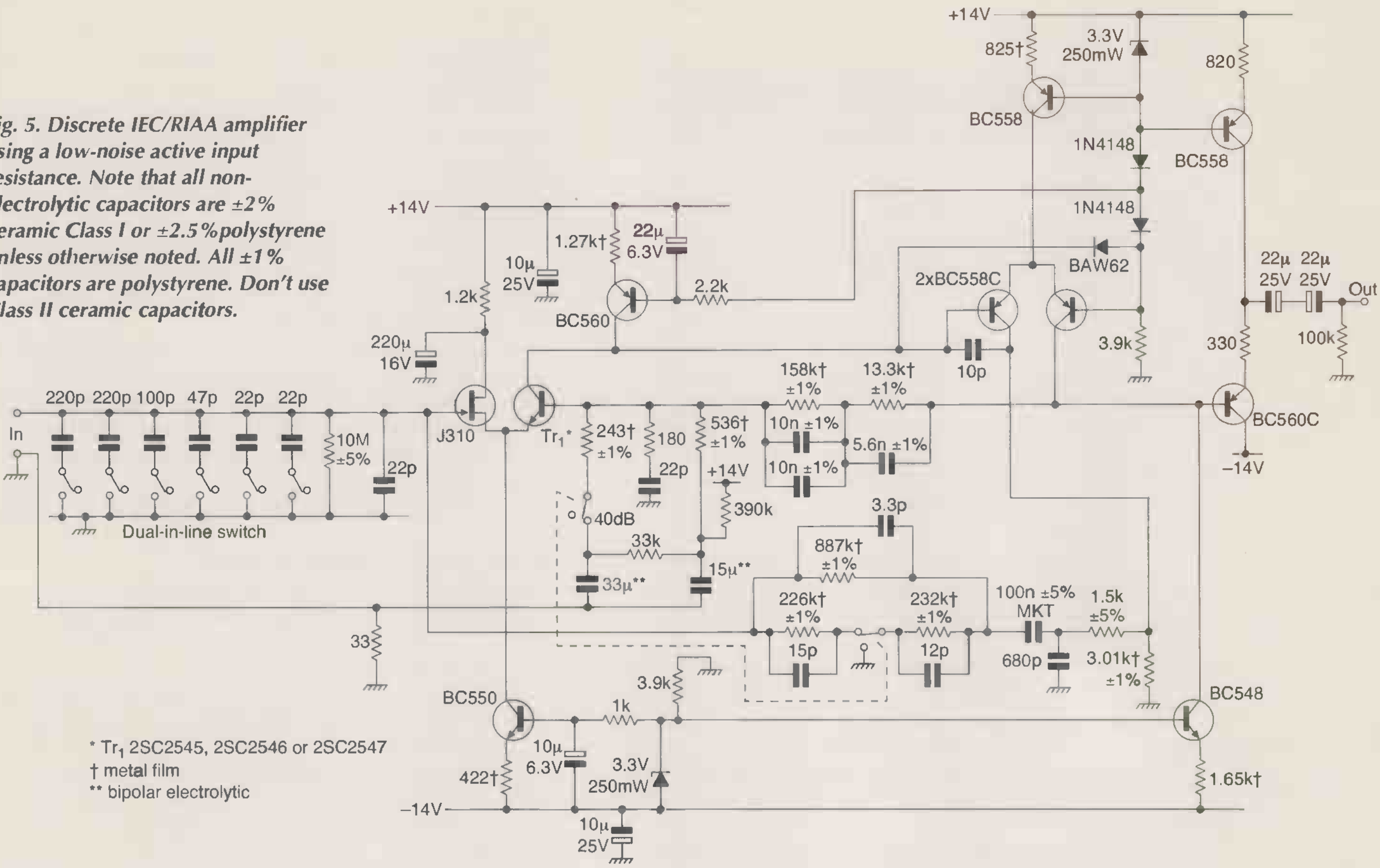
This includes the current noise term, which has been transformed into an equivalent noise voltage across the given source impedance $Z_S$. The 2SC2545, 2SC2546 and 2SC2547, which are good low-noise low-frequency transistors, have $r_b\approx14\Omega$ and $h_{FE}\approx600$.

The input device is driven from a source impedance of about 12kΩ, which is largely reactive. If a 2SC2545 were used here, the resulting values would be:

$I_{C,opt}\approx51.56\mu A$
$g_{m,opt}\approx2.041mS$
$S(v_{n,bip,inc.currentnoise})\approx8.158\times10^{-18}V^2/Hz$ or 2.856nV/√Hz

According to the graphs in reference 3, a high

*Fig. 5. Discrete IEC/RIAA amplifier using a low-noise active input resistance. Note that all non-electrolytic capacitors are ±2% ceramic Class I or ±2.5%polystyrene unless otherwise noted. All ±1% capacitors are polystyrene. Don't use Class II ceramic capacitors.*




transconductance JFET such as the J310 biased at several milliamperes should be able to outperform this bipolar transistor, having a noise of about 2nV/√Hz at 1169Hz (for a large part $1/f$).

At the same time, the JFET provides more transconductance and a better high-frequency behaviour than a 2SC2545 biased at about 52µA. The J310 in Fig. 5 is biased at about 3.3mA. A higher bias current (10mA to 15mA) would be somewhat better.

The situation is different for the transistor driven from the feedback network. In the 40dB gain position, its base is driven from a source impedance of about 165Ω, while its emitter 'sees' the output resistance of the source follower, about 145Ω. This results in a 1.913mA optimal bias current and a 0.6637nV/√Hz total noise contribution. This cannot be outperformed by a single J310 – although it should still be possible with a group of paralleled J310s.

The 887kΩ, 226kΩ and 232kΩ feedback resistors have been shunted with capacitors. This has been done to swamp the few tenths of a picofarad of capacitance associated with a discrete resistor.

The corner frequency lies at about 52kHz. The 680pF capacitor attenuates the voltage on the feedback resistors above 52kHz, eliminating the Miller effect which would otherwise multiply the influence of the capacitances across the feedback resistors.

Gain at 1kHz can be switched between about 30 and 40dB to account for different cartridge sensitivities. The input resistance is designed to be about 3% too low in the 30dB position. It should damp the resonance of the cartridge inductance and load capacitance a bit more than normal, resulting in a small high-frequency loss. This should more or less compensate for the error of +0.37dB at 20kHz caused by the fact that the gain drops to unity rather than zero at high frequencies.

At low frequencies, the maximum output signal level without clipping is about 16.7V peak to peak, or 5.9V RMS in the case of a sine wave. The tail current of the output stage and the feedback network impedance determine the headroom at high signal frequencies.

The output stage can deliver about 1.33mA peak to the feedback network. In the 40dB position, the resistance between the base of the 2SC2545 and ground is about 167Ω, corresponding to 222mV maximum peak input voltage at high audio frequencies. This changes into 711mV peak in the 30dB position.

To put these figures in perspective: my record player produces peak signal levels of 30mV to 35mV when it plays popular music records with a large high-frequency content, but according to reference 4, peak signal levels up to 100mV can occur at the output of sensitive cartridges playing loud records.

## Measured results

Several measurements have been made using an HM1505 oscilloscope with cursors and a ±3% vertical accuracy specification. A 1:1 probe has been used whenever this could be expected to give the most accurate results.

When possible, relative measurements have been made using the same oscilloscope channel to minimise inaccuracies. In most cases, the test signals have been generated with a CD player playing a test CD and with 0.01% accurate resistors to attenuate the signal or to convert it into a current (for the impedance measurements). For the noise measurements, a separate 100 times amplifier and A-weighting filter have been used.

Response when driven from a low source impedance has been measured at 15 different frequencies and compared to the theoretical IEC modified RIAA response. The voltage gain at 1kHz has been measured to be 30.17dB in the 30dB-mode and 40.27dB in the 40dB-mode.

Compared to 1kHz, the response was accurate to within

*Table 1. Equivalent parallel resistances at various frequencies calculated from measurement of input impedance magnitude and phase.*

| Frequency | $R_{in}$, 30dB | $R_{in}$, 40dB |
|---|---|---|
| 1kHz | 46.64kΩ | 48.85kΩ |
| 10kHz | 45.54kΩ | 46.99kΩ |
| 15kHz | 45.5kΩ | 46.76kΩ |
| 20kHz | 45.79kΩ | 46.3kΩ |

±0.1dB for frequencies between 30Hz and 20kHz in the 40dB-mode. With 30dB nominal gain, the response was accurate to within ±0.14dB between 30Hz and 10kHz, and at 20kHz, the measured error was +0.24dB. Theoretically, this should be +0.37dB. At 20Hz, the error was +0.18dB in the 30dB and +0.25dB in the 40dB position, probably due to the relatively large tolerances of the electrolytic capacitors.

The input impedance magnitude and phase have been measured at various frequencies and the equivalent parallel resistance has been calculated from the measurement results, see the **Table 1**.

Input capacitance is in the order of 30pF at the minimum input capacitance setting – i.e. just the RIAA amplifier, not including the probe capacitance.

A-weighted noise has been measured with a shorted input and with an open input with 30dB and 40dB mid-band gain, in all cases with the minimum input capacitance setting.

The measurements can be about 20% off, because I had to estimate the RMS noise from the quasi peak to peak values measured with the oscilloscope. Besides, the noise current (open input) measurement has been affected by interference from the mains.

In the 30dB mode, the A-weighted noise at the RIAA amplifier output was – fortunately – 27.67μV RMS with open input and 9.85μV RMS with shorted input. This corresponds to 5.381nV/√Hz and 0.3216pA/√Hz at the input.

In the 40dB mode, the results were 26.83μV A-weighted RMS at the output with shorted input and 43.67μV RMS at the output with open input, corresponding to 4.729nV/√Hz and 0.1638pA/√Hz at the input.

All measured results are reasonably close to the expected values, except for the equivalent input noise voltage, which is higher than expected. With the source of the J310 AC shorted to the input ground, it drops to somewhere between 2.2 and 2.9nV/√Hz in the 40dB gain mode.

Most probably, the unexpectedly high noise voltage is due to the 1/*f* noise of the J310 being higher than expected.

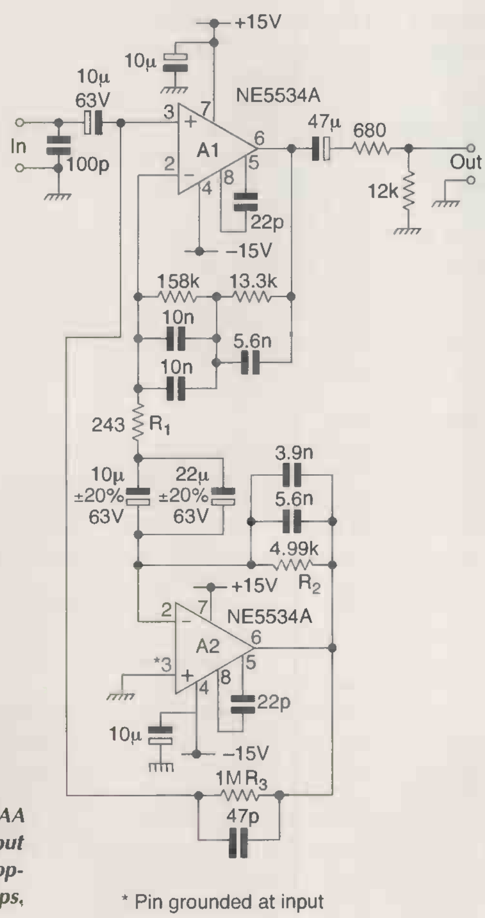



***Fig. 6. IEC modified RIAA amplifier with an active input resistance made from op-amps.***

## RIAA amplifier using op-amps

The network in Fig. 4 is not very suitable for implementation with op-amps. After all, an op-amp does not have a pin carrying a current which is equal but opposite to the output (signal) current.

Although it is possible to use floating power supplies to solve this problem[5], it is much simpler to use a different topology, **Fig. 6**. A $-R_2/R_1$ times amplified version of the input signal occurs at the output of A2, resulting in an input resistance of $R_3/(1+R_2/R_1)$. I have only subjected this circuit to a functional test, not to measurements. It seems to work fine.

Noise can be slightly reduced by replacing A2 with an ultra-low noise op-amp such as the LT1028, and adjusting its frequency compensation. The LT1028 may need a compensation capacitor between pin 5 and its output when the output signal is fully fed back to the inverting input at high frequencies. The data sheet is not very clear about the required value[6].

On the other hand, replacing A1 with an ultra low noise op-amp would make the noise much worse. Ultra-low noise op-amps usually have a bipolar input stage biased at a high bias current. This results in a low input-noise voltage density, but also in a high input noise current density.

To make matters worse, they often feature a base current compensation circuit which injects large fully correlated noise currents in the inverting and non-inverting input terminals.

In the case of the LT1028, the input noise voltage is specified as 0.9nV/√Hz, with a noise current of 1pA/√Hz typical at 1kHz[6]. Carefully reading the small print and the applications information shows that the noise current value only applies when both inputs are driven from exactly the same impedance.

In the case of an RIAA amplifier, this would mean that a dummy cartridge would have to be included in the feedback network. Fortunately, there is also a graph of the total noise versus unmatched source impedance, which shows that the actual input noise current is about 3.25pA/√Hz.

With a source impedance of 12kΩ, 3.25pA/√Hz has the same influence as 39nV/√Hz, showing that even a 741 would be better suited for a moving-magnet RIAA amplifier.

## How much improvement can be obtained?

As a best-case approximation for the improvement due to using an active input resistance, I'll compare the sum of

*Table 2. Impedance for a Shure V15 III cartridge and for an LR network.*

| f | $\|Z_{measured}\|$ | $\|Z_{theory}\|$ | $\varphi_{measured}$ | $\varphi_{theory}$ | R* | G* | F* | $F_{theory}$ |
|---|---|---|---|---|---|---|---|---|
| 2kHz | 6kΩ | 5934Ω | 72° | 76.96° | 1.85kΩ | 51.5μS | 1.5dB | 1.93dB |
| 5kHz | 14kΩ | 14513Ω | 72° | 84.71° | 4.33kΩ | 22.1μS | 2.93dB | 6.38dB |
| 7kHz | 20kΩ | 20276Ω | 70° | 86.21° | 6.84kΩ | 17.1μS | 3.51dB | 8.77dB |
| 10kHz | 28kΩ | 28934Ω | 66° | 87.35° | 11.4kΩ | 14.5μS | 3.92dB | 11.55dB |
| 14kHz | 37kΩ | 40486Ω | 61° | 88.1° | 17.9kΩ | 13.1μS | 4.19dB | 14.32dB |
| 20kHz | 50kΩ | 57821Ω | 53° | 88.67° | 30.1kΩ | 12μS | 4.42dB | 17.33dB |

*calculated from measured results

the cartridge noise and the amplifier noise for two cases.

The first case is with a RIAA amplifier with the 47kΩ resistor as its only noise source, the second with a RIAA amplifier which does not generate any noise at all.

In both cases, assume that the noise coming out of the cartridge is only thermal noise. This implies that the record player is not playing a record, otherwise there would also be record noise and rumble.

If the cartridge had a frequency-independent effective series resistance, calculating the difference in A-weighted integrated noise level would be easy. For the example of a cartridge with a 1kΩ effective series resistance and 12kΩ impedance at 3852Hz:

Cartridge noise: 4.024nV/√Hz
Noise due to 47kΩ: 7.043nV/√Hz
Total of the cartridge and resistor: 8.111nV/√Hz
Difference between total and cartridge noise: 6.089dB.

However, the effective series resistance of a real cartridge rises with frequency. Richard Visée[7] measured the impedance of a Shure V15 III cartridge using an HP4194A impedance gain and phase analyser. The DC resistance and the inductance were 1.3388kΩ and 460mH, respectively. See **Fig. 7** and **Table 2**.

In Table 2, the value $|Z_{measured}|$ is the magnitude of the measured impedance, $\varphi_{measured}$ is its phase. $R$ is the effective series resistance at a frequency $f$, while $G$ is the effective parallel conductance. $F$ is the ratio of the thermal noise of the cartridge and a 47kΩ load to the noise of the cartridge alone at a frequency $f$, expressed in dB.

For comparison, $|Z|$, $\varphi$ and $F$ have also been calculated for a theoretical cartridge with 460mH of inductance and a constant 1.3388kΩ series resistance. Obviously, only the magnitude of the impedance can be reasonably modelled with a simple *LR* series network.

It is clear that the improvement averaged over the audio band lies somewhere around 3dB, rather than the 6.089dB calculated before.

Similar results have been found for a Marantz cartridge by measuring its A and IEC/RIAA weighted noise using the amplifier in Fig, 5. After correcting for the amplifier noise, this relatively low-impedance cartridge with 349mH inductance and 350Ω DC resistance turned out to have a noise equivalent to roughly 2215Ω weighted average effective series resistance. The noise of the cartridge and a passive 47kΩ resistor together would have been 2.27dB higher than the thermal noise of the cartridge itself.

Whether about 3dB best-case improvement is worth the effort depends on your point of view. It will not be immediately noticed by the average user. But then again, it is common practice to try to keep the distortion, frequency response errors and noise levels introduced by audio amplifiers well below the errors introduced in other parts of the audio signal chain or well below the threshold of audibility.

In this way, the final quality is determined by the parts of the audio chain which are the hardest to get right: microphones, loudspeakers, cartridges, the acoustics of the listening room and the bad habit of some recording and broadcasting people to use unnecessarily high amounts of dynamic compression. ■

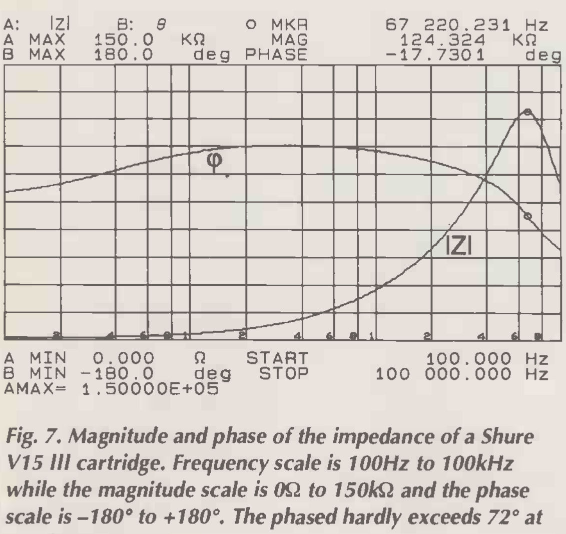



***Fig. 7. Magnitude and phase of the impedance of a Shure V15 III cartridge. Frequency scale is 100Hz to 100kHz while the magnitude scale is 0Ω to 150kΩ and the phase scale is –180° to +180°. The phased hardly exceeds 72° at any frequency.***

**References**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