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# A Phono PrePreamp for the (SA) CD Age, Part 1

By Norm Thagard

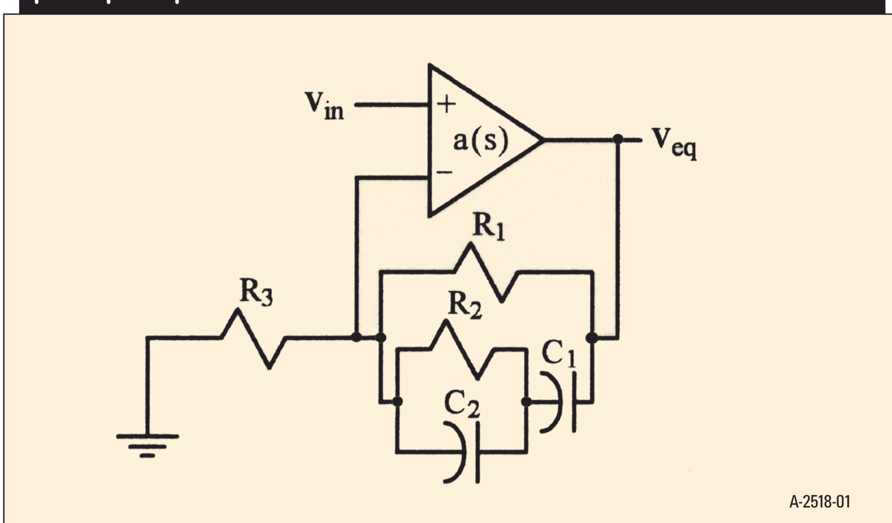
This three-part article takes a look back at a phono pre-preamp design featured in the premiere issue of *audioXpress*.

When re-published in *Electronics World* at the request of Ed Dell, my 2001 *audioXpress* phono pre-preamplifier design article[1] received a scathing condemnation by Douglas Self. The active-passive filtering was his principal stated objection. Brushing aside the example his review provided of the general decline in civility and addressing the objection itself, I decided to approach the design issue in the way that he apparently would. This approach is to generate the necessary poles and zero by active equalization in the feedback network around a gain block.

This article's title notwithstanding, the design described here should not be confused with a so-called "head amp." That term usually connotes a preamplifier with frequency-independent gain of ten or so that is designed to "bump up" the output of a low-output cartridge for input into the MM phono input of a system preamplifier. Like my previously published "pre-preamp," this new offering provides RIAA compensation and, in this case, sufficient gain to drive a power amplifier directly, although most builders will probably opt to apply its output to one of the high-level (e.g., CD or tuner) inputs of a system preamplifier.

Although I made no specific attempt to optimize for input capacitance, another motivation for attempting a new design was knowledge gained from Raymond Futrell's useful article[2] indicating that cartridge interfacing might potentially be improved if input capacitance were reduced. My referenced design employed a complementary differential amplifier at the input, which is not optimal from the standpoint of capacitance. Therefore, I wanted to employ a single-ended topology in the first stage as a design exercise.

FIGURE 1: RIAA compensation by the feedback network of a noninverting phono preamplifier.




As is my usual practice, I shall attempt to explain the design process and principles thoroughly and provide derivation of those design equations that were not addressed in earlier articles. I attempt to limit equations to the algebraic variety unless there is absolutely no choice. I hope that this series of articles will have use as tutorials even if you do not construct the circuits described herein.

Increasingly, I am emboldened to attempt the derivations myself rather than by reference to a text or paper. Despite holding a terminal degree in medicine, I did, after all, begin life as an electrical engineer. But for the closure of the engineering science school at Florida State University in 1972 five quarters into my Ph.D. program, I would today be a Ph.D. electrical engineer rather than an M.D. Even so, with no one to proof these exercises, I expose myself to serious mistake.

I welcome constructive criticism, and it is in all our interests to correct such mistakes quickly for the sake of the reader and experimenter. Whether my explanations of circuit operation are entirely correct, every design to date has been objectively tested on the bench and subjectively evaluated in a quite good stereo system for a reasonable period before publication.

## THE TOPOLOGY, DERIVING THE EQUATIONS

In an article in *TAA*,[3] Reg Williamson looked at methods of RIAA equalization through frequency-dependent feedback networks of gain blocks, including the noninverting configuration depicted in **Fig. 1**. He also provided the necessary formulae for calculating the component values.

I never believe I really understand something unless I can work through the derivations myself. For this reason and because I took a slightly different approach to component selection, I offer the following development of the design steps beginning with the impedance of the $R_1R_2C_1C_2$ network in **Equation 1**.

The feedback fraction—i.e., that portion of the output voltage applied to the inverting input—is shown in **Equation 2**.

Both numerator and denominator can be factored as $(s/s_1 + 1)(s/s_2 + 1)$. These factors, when equated to zero, yield zeros from the numerator and poles from the denominator for the feedback fraction. From analysis of the closed-loop circuit, if open-loop gain $a(s) >> 1$, then closed-loop gain $A(s) \cong 1/\beta(s)$, and the zeros become poles and the poles become zeros in the closed-loop response.

The Laplace complex frequency value corresponding to DC is $s = 0$. Substituting this value in the impedance expression makes $Z$ and the bracketed term in the feedback fraction equation identically one. Thus, DC gain is $A_0 \cong A(0) \cong 1/\beta(0) = (R_1 + R_3)/R_3 = 1 + R_1/R_3$. You should recognize this as the gain of an ideal op amp in a noninverting configuration, and it is simply the inverse of the $R_1$, $R_3$ voltage divider ratio as expected from feedback theory. That this is true is obvious from the schematic because $C_1$ and $C_2$ are open-circuits at DC, thereby effectively removing themselves and $R_2$ from the circuit.

Making the substitution $j\omega$ for $s$ in the numerator yields $(j\omega/\omega_1 + 1)(j\omega/\omega_3 + 1)$ or, alternatively, $(j\tau_1\omega + 1)(j\tau_3\omega + 1)$. In this case, the numerical values of these time constants are known because they are dictated by the RIAA equalization criteria. The numerator should yield the RIAA poles at 50.05Hz ($\tau_1$ = 3,180µs) and 2,122.06Hz ($\tau_3$ = 75µs). You can similarly write the denominator as $(j\omega/\omega_2 + 1)(j\omega/\omega_4 + 1)$ or $(j\tau_2\omega + 1)(j\tau_4\omega + 1)$. This must yield the RIAA zero at 500.5Hz ($\tau_2$ = 318µs).

The second zero, whose time constant is represented by $\tau_4$, is not part of the RIAA equalization but rather is the inevitable result of the impedance network shown. The only way to keep $\tau_4$ from significantly affecting the in-band RIAA response curve is to make it as small as possible—i.e., to place $f_4 = \omega_4/(2\pi)$ as far above 20kHz as practicable.

The task is to derive relationships for the determination of the five component values based on the three known time constants and the desired gain. Although it is the 1kHz gain that is usually specified, it is easier in this approach to specify the DC gain. There is no standard for absolute gain anyway, and the 1kHz gain should be 19.3dB less than the 20Hz gain, or about ten times less than $A_0$.

Following gain specification, $R3$ is often picked as a first component value selection on the basis of a compromise between a small value for minimum noise and a large value to preclude excessive loading of the amplifier's output by low feedback network impedance[4]. Ideally in a voltage amplifier, the feedback network would have infinite input impedance to preclude such loading effects and zero output impedance so that its (the feedback network) output is not loaded down by the impedance of the node to which feedback is applied. Once $R_3$ is selected, then $R_1 = (A_0 - 1)R_3$.

Note that the numerator is as shown in **Equation 3**. From this, it is apparent that:

(A) $\tau_1\tau_3 = R_1R_2R_3C_1C_2$

and (B) $\tau_1 + \tau_3 = R_1C_1 + R_2(C_1 + C_2)$

Since $\tau_1$ and $\tau_3$ are the time constants of the RIAA response poles and $R_1$ would have been previously determined, there are three unknowns and only two equations. However, a similar operation on the denominator gives:

(C)

$$\tau_2\tau_4 = \frac{R_1R_2R_3C_1C_2}{R_1 + R_3}$$

and (D)

$$\tau_2 + \tau_4 = \frac{R_1R_3C_1}{R_1 + R_3} + R_2(C_1 + C_2).$$

Time constant $\tau_2$ corresponds to the RIAA response zero, so there are four independent equations in four unknowns. You can calculate the remaining undetermined component values and time constant $\tau_4$, albeit not without some algebraic manipulation of these equations. For example, you can rearrange equation (A) as $R_2C_1C_2 = \tau_1\tau_3/R_1$, with this substitution then made in equation (C) to yield **Equation 4**.

This is useful because it shows that the second zero, which is not part of the RIAA equalization scheme, can only be pushed to a higher frequency by increasing $A_0$. The values for $R_2$, $C_1$, and $C_2$ remain to be determined.

From equation (B), $R_2(C_1 + C_2) = \tau 1 + \tau 3 - R_1C_1$. You can substitute this result into equation (D) to eliminate unknowns $R_2$ and $C_2$ (**Equation 5**).

Again, from equation (B), $R_2 = (\tau_1 + \tau_3 - R_1C_1)/(C_1 + C_2)$. Substituting this into equation (A) yields **Equation 6**.

Now that only $R_2$ remains to be determined, any of the four equations can be used. Equation (A) is the simplest:

$$R_2 = \frac{\tau_1\tau_3}{R_1C_1C_2}.$$

It is an easy matter to march through the succession of equations just derived until all component values have been determined. The disadvantages of the equalization method used here are the interdependence of component values with gain and the production of a possibly unwanted zero at $f = 1/(2\pi\tau_4)$.

There is an argument that setting $\tau 4 = 3.18\mu s$ compensates nicely for the pole introduced by the cutting lathe. This would dictate the value of $A_0$ with the compensation network used here, but a modification to this network would allow independent setting of $A_0$ and $\tau_4$. I have not elected to do this, so the zero represented by $\tau_4$ must be well above 20kHz in order to minimize its effect upon equalization in the audio band.

For the usual moving-magnet cartridge, $A_0$ = 60dB seems to be a good choice. If it is less than this, then in my limited experience there is a significant disparity in volume control settings between phono and high-level sources. For my original phono preamp, I increased midband gain to 50dB to account for the use of first a Grado Sonata low-output moving-magnet cartridge and then an Audio-Technica OC9/MLII moving-coil cartridge. Even at this level, the volume disparity was significant because both cartridges output only 0.4mV at 5cm/s at 1kHz.

Ultimately, $|Z| \to 0$ and the feedback network will load the preamp's output with $R_3$. This is readily seen by reference once again to the schematic, noting that $C_1$ and $C_2$ become short-circuits as $f \to \infty$, effectively producing a voltage follower whose output is shunted with $R_3$. This is the effect that produces $\tau_4$.

One of the objections to this approach to equalization is exactly that the noninverting configuration becomes a voltage follower. RIAA equalization would allow gain to decrease indefinitely rather than have a minimum value of one. Maybe the various objections to this, that, or the other topology and/or equalization scheme are valid, maybe not.

For example, gain of the inverting configuration can continue to fall indefinitely with increasing frequency, a clear advantage over the noninverting configuration, at least in this regard. The inverting configuration can be realized by removing the ground at the left side of $R_3$ and applying the cartridge output there instead. The noninverting input is then grounded, producing a virtual ground at the inverting input due to feedback. $R_3$ now sets the input resistance of the phono preamplifier. For a moving-magnet design, cartridge interface requirements may dictate a value as high as $R_3$ = 47kΩ, with the noise generated by this high resistance input directly to the preamp.

The input resistor in the noninverting configuration is shunted by the low or relatively low resistance of the cartridge and is therefore not usually a major contributor to noise. This is a clear *dis*advantage of the inverting relative to the noninverting configuration

**Equation 1.**

$$Z = R_1 \left\| \left[ \frac{1}{sC_1} + R_2 \left\| \frac{1}{sC_2} \right. \right] \right. = R_1 \frac{R_2(C_1 + C_2)s + 1}{R_1R_2C_1C_2s^2 + [R_1C_1 + R_2(C_1 + C_2)]s + 1}.$$

**Equation 2.**

$$\beta(s) = \frac{R_3}{R_3 + Z} = \frac{R_3}{R_1 + R_3}\left\{\frac{R_1R_2C_1C_2s^2 + [R_1C_1 + R_2(C_1 + C_2)]s + 1}{\frac{R_1R_2R_3C_1C_2}{R_1 + R_3}s^2 + \left[\frac{R_1R_3C_1}{R_1 + R_3} + R_2(C_1 + C_2)\right]s + 1}\right\}$$

**Equation 3.**

$$\frac{1}{\omega_1}\frac{1}{\omega_3}(j\omega)^2 + \left(\frac{1}{\omega_1} + \frac{1}{\omega_3}\right)(j\omega) + 1 = \tau_1\tau_3(j\omega)^2 + (\tau_1 + \tau_3)(j\omega) + 1.$$

**Equation 4.**

$$\tau_2\tau_4 = \frac{R_1R_3(\tau_1\tau_3/R_1)}{R_1 + R_3} \Rightarrow \tau_4 = \frac{R_3}{R_1 + R_3}\frac{\tau_1\tau_3}{\tau_2} = \frac{1}{A_0}\frac{\tau_1\tau_3}{\tau_2}.$$

when used for moving-magnet cartridges. Even if $R_3$ is small—as it might be to interface with a moving-coil cartridge—there is the problem that $R_3$ is in series with the cartridge impedance $Z_g$, which perhaps could significantly affect gain and equalization.

Practically speaking, the approaches that have been used historically have all worked more or less satisfactorily, and the shortcomings of a given approach, if properly managed, tend to have little significance to the listener, at least on a verifiable basis. Therefore, no one should assume for him or herself the mantle of arbiter of phono preamp design.

## HEADROOM

Another design factor is accommodation of the highest signal levels that will be encountered. Simply put, the phono preamplifier's output must have reasonably low distortion under any and all conditions of normal use.

Stylus velocities produce proportional voltages that are then amplified by the phono preamplifier. To determine whether the preamplifier will potentially have a problem, you must consider the maximum velocities that can occur at all frequencies in the audio band. This requirement results from the RIAA response of the preamplifier.

To understand this better, consider that one author reported that in research in the Shure Labs, velocities of 50cm/s at high frequency were not rare and 25cm/s at 1kHz was frequently seen[5]. The modern Shure V15VxMR MM cartridge has 1kHz output of 3mV RMS at a stylus velocity of 5cm/s, so at 25cm/s it would produce 15mV RMS. The MM preamp variant presented in this article, with approximately 40dB gain at 1kHz, would produce about 1.5V RMS output from this input. The author then states that headroom of 20dB is essential, meaning that the preamp should be capable of producing 15V RMS at 1kHz with reasonably low distortion.

Such a preamplifier must have power supply voltages in excess of $2 \times 15 \times \sqrt{2}$ = 42.5V or ±21.25V. It should be emphasized that this is the 1kHz headroom requirement. Potentially, it could be even higher at other frequencies because preamplifier gain rises below 1kHz while above this frequency velocities would tend to increase, all other things being equal. This can be thought of as a slew rate problem. A sine wave of frequency $f$ would have maximum velocity of $2\pi A_{max} f$, where $A_{max}$ is the highest amplitude (displacement) pro-

**Equation 5.**

$$\tau_2 + \tau_4 = \frac{R_1 R_3 C_1 + (R_1 + R_3)(\tau_1 + \tau_3 - R_1 C_1)}{R_1 + R_3} \Rightarrow C_1 = \frac{(\tau_1 + \tau_3 - \tau_2 - \tau_4)(R_1 + R_3)}{R_1^2}.$$

**Equation 6.**

$$\tau_1 \tau_3 = R_1 \left( \frac{\tau_1 + \tau_3 - R_1 C_1}{C_1 + C_2} \right) C_1 C_2 \Rightarrow C_2 = \frac{\tau_1 \tau_3 C_1}{R_1 C_1 (\tau_1 + \tau_3 - R_1 C_1) - \tau_1 \tau_3}.$$

duced at frequency *f*. Fortunately, it probably is reasonable to assume that satisfaction of the 1kHz criterion is sufficient due to lower recording levels at low frequency and lower preamp gain at high frequency[6].

Given the unappealing power supply voltage requirement, I decided to take another look at the argument. On the Shure website, trackability for the V15V × MR in units of cm/s peak velocity was given as 30cm/s at 400Hz, 46cm/s at 1kHz, 80cm/s at 5kHz, and 60cm/s at 10kHz, all measured at 1 gram tracking force. Shure rightly claims that their cartridge is among the best, if not *the* best, in terms of trackability. In my mind, if these results are accepted, it should be sufficient to base preamp headroom requirements upon them with perhaps some margin. It does not seem reasonable to levy a requirement upon the preamp that significantly exceeds the capability of the cartridge.

The preamp design offered here can produce low distortion outputs in excess of 12V RMS at all audio frequencies. Even assuming output limited to 10V RMS, the MM-optimized variant would have headroom of 11.1dB at 400Hz. This is calculated as follows:

$\text{Gain}_{dB}$ at 400Hz = $\text{Gain}_{dB}$ at 1kHz + 3.8dB (RIAA-specified) ≅ 43.8dB, which is 155 numerically.

Output voltage = numerical gain × cartridge output = 155 × [(3mV RMS/5cm/s) × 30cm/s] = 2.79V RMS.

$\text{Headroom}_{dB}$ = (maximum preamp output/actual output)$_{dB}$ = 20log(10V RMS/2.79V RMS) = 11.09dB.

This result assumes that 400Hz cartridge output is the same as 1kHz output for the same velocity, a fair assumption if the cartridge is, indeed, a velocity-responding transducer. Similar calculations indicate that headroom is 11.2dB at 1kHz, and 14.6dB at 5kHz.

Assuming order-of-magnitude lower moving-coil cartridge voltages and order of magnitude greater MC preamplifier gain, headroom should be approximately the same in both cases. Due to loading of the output at high frequency by the feedback network, the initial moving-coil design could not slew fast enough at 10kHz to provide low distortion output above 8V RMS, and at 20kHz, the useful output was reduced to just 4.5V RMS. It is clear from this pattern that this should be no problem. In addition, studies have shown that there is little acoustic energy above 10kHz.

However, knowing that there would be criticism for this shortcoming, I increased second stage current levels enough to eliminate slew-rate limiting at audio frequencies. Fortunately, the resulting quiescent current levels were not so high as to require heatsinks for the transistors, although they do become quite warm. I chose to use small heatsinks anyway because I'm a conservative at heart. These heatsinks are grounded, thereby requiring insulation of the transistors from the metal sinks.

Obviously, this preamp does not meet the 20dB headroom requirement, but bear in mind that the calculated headroom is in excess of the *cartridge's* trackability limit. Well below the point where velocities are forcing this preamp into clipping, the stylus would have ceased to accurately track the groove and significant distortion would already be apparent. Perhaps tracking distortion is less objectionable than preamp distortion but I doubt it, especially given the "graceful" clipping that is seen on the oscilloscope.

As a matter of practicality, I ran the output of a Shure V15 Type V MR cartridge into the MC version of the preamp described in this article. Despite the very high gain and high cartridge output, on the oscilloscope I saw few peak voltages approach the clipping level (approximately 17V) and heard nothing abnormal in the sound while listening to several records with moderate to high recording levels. The preamp's input resistance was set to 47kΩ per Shure's recommendation prior to the test. This certainly seems to indicate that practical headroom of the MC version will be much more than sufficient in normal use with a low-output cartridge.

Admittedly, some magnetic cartridges have outputs in excess of 3mV at 5cm/s, but probably do not produce usable output significantly exceeding the levels just considered. In any event, it is possible to meet any requirement by either increasing power supply voltages or by reducing gain. The design for the former is more difficult than for the latter.

## MOVING-MAGNET DESIGN

As an example, consider a design suitable for moving-magnet cartridges. $R_3$ = 499Ω is a reasonable value. It follows that for approximately 40dB gain at 1kHz, $R_1 = R_3(A_0 - 1) = 499\Omega(1000 - 1) \cong 499k\Omega$.

From the previously derived formulae, see **Equation 7**.

## MOVING-COIL DESIGN

To accommodate moving-coil cartridges, consider a second design example. I've mentioned without elaboration that noise considerations lead you to make $R_3$ as small as possible. Noise abatement requires more attention not only due to higher gain, but also because cartridge resistance is generally lower. This latter fact means that resistances in the input stage must be kept correspondingly small, otherwise their noise will significantly lower signal-to-noise (SNR) ratio.

The spec sheets for the Audio-Technica OC-9/MLII and the Shelter 501 moving-coil cartridges that I own both indicate that 100Ω input resistance is optimal when used with a "head amp" or 20Ω when used with a step-up transformer. The resistance (d.c.r.) of the OC-9/MLII is about 10Ω. I chose $R_3$ = 5Ω, knowing in advance that this would tax my design skills, but nonetheless I wanted to use a value no more than that of almost any cartridge. $A_0$ = 80dB provides an order of magnitude greater gain than for the moving-magnet design.

The OC-9 and Shelter 501 have 1kHz outputs of 0.4mV at 5cm/sec, about 1/10 that of a Shure V15, so the extra gain should compensate nicely for the lower output. The choice of 5Ω for $R_3$ then dictates $R_1 = R_3(A_0 - 1) = 5\Omega(10{,}000 - 1) \cong 49.995k\Omega$. The closest 1%-tolerance value is 49.9kΩ.

At this point, I should mention that the amount of feedback is given by the difference between open- and closed-loop gain and is therefore not effective unless $|a(s)| \gg |A(s)|$. This requirement is the basis of a major problem with deriving all RIAA poles and zeros by feedback around a single gain block, especially for a moving-coil preamplifier. If $A_0$ = 80dB (10,000), then satisfaction of the above inequality probably means that $|a(j0)|$ should be at least 100dB (100,000). Such gains are easily achieved with some op amps but can pose a problem for simple discrete designs.

Audiophiles usually prefer simple, yet high-performance designs. I accepted as a design challenge the task to realize a well-performing circuit in a two-stage gain block, although admittedly the two stages are both cascodes and therefore *compound* stages. A measure of open-loop gain adequacy at all audio frequencies is accuracy in tracking the RIAA curve.

The design tracks within 0.1dB in the 20Hz – 20kHz band, which is a good sign. However, feedback is squeezed at 20Hz, so it would not be unexpected that distortion might be proportionally higher at low frequency. The remaining component values are calculated as shown in **Equation 8**.

The *RC* time constants must remain the same, so if resistance values fall, capacitance values must rise by the same factor. Of particular interest is the very small value of $\tau_4$ due to the very high gain. This time constant corresponds to a frequency of 2.12MHz. This is so far above the audio band that no significant effect on equalization should be manifest.

As always, there is no opportunity to use custom parts, so the values used in the prototype were $C_1$ = 56nF, $C_2$ = 20nF,

**Equation 7.**

$$\tau_4 = \frac{R_3}{R_1 + R_3}\frac{\tau_1\tau_3}{\tau_2} = \frac{499\Omega}{499k\Omega + 499\Omega}\frac{(3{,}180\mu s)(75\mu s)}{318\mu s} = 750ns,$$

$$C_1 = \frac{(3{,}180\mu s + 75\mu s - 318\mu s - 0.74925\mu s)(499k\Omega + 499\Omega)}{(499k\Omega)^2} = 5.89nF,$$

$$C_2 = \frac{(3{,}180\mu s)(75\mu s)(5.89nF)}{(499k\Omega)(5.89nF)[3{,}180\mu s + 75us\text{-}(499k\Omega)(5.89nF)] - (3{,}180\mu s)(75\mu s)} = 2.04nF,$$

$$R_2 = \frac{(3{,}180\mu s)(75\mu s)}{(499k\Omega)(5.89nF)(2.04nF)} = 39.8k\Omega.$$

**Equation 8.**

$$\tau_4 = \frac{5\Omega}{49.9k\Omega + 5\Omega}\frac{(3{,}180\mu s)(75\mu s)}{318\mu s} = 75\ ns,$$

$$C_1 = \frac{(3{,}180\mu s + 75\mu s - 318\mu s - 74.925ns)(49.9k\Omega + 5\Omega)}{(49.9k\Omega)^2} = 58.9\ nF,$$

$$C_2 = \frac{(3{,}180\mu s)(75\mu s)(58.9nF)}{(49.9k\Omega)(58.9nF)[3{,}180\mu s + 75\mu s\text{-}(49.9k\Omega)(58.9nF)] - (3{,}180 v s)(75\mu s)} = 20.4\ nF,$$

$$R_2 = \frac{(3{,}180\mu s)(75\mu s)}{(49.9k\Omega)(58.9nF)(20.4nF)} = 3.98\ k\Omega.$$

and $R_2$ = 4.02kΩ. There is a greater choice of resistor values than capacitor values, so if "tweaking" is required to achieve closer RIAA tracking, choose the capacitor values and then substitute different resistor values close to the calculated ones. It is always possible to "trim" values by series-parallel combinations of stock components to achieve any desired degree of accuracy.

## NOISE CONSIDERATIONS

Open-circuit resistor "Johnson noise" is given by $V_{nR} = (4kTR)^{1/2}$ V/Hz$^{1/2}$. Since power is proportional to the square of voltage, the equation implies equal power per hertz of frequency. This component of resistor noise is said to be "white noise." The term "this component" was used intentionally, because while any resistance will have associated with it this noise term, resistors may have other noise components as well, producing "excess noise," i.e., noise above and beyond the resistance-dependent white noise just described. If low-noise were a design requirement[7], you would probably choose wirewound and metal film resistors, which exhibit less excess noise than carbon resistors.

While it seems intuitive that resistor noise would be proportional to resistance, at first glance it seems paradoxical that $R_3$ would be one of the bigger noise sources, especially given that it is small in comparison to its feedback network partners, $R_1$ and $R_2$. However, $R_3$ noise voltage, given by $e_{n\text{-}R3} = (4kTR_3)^{1/2} = [4(1.38 \times 10^{-23}$ joules/°K)(300.16 °K)(5Ω)]$^{1/2}$ = 2.88 × $10^{-10}$ V/Hz$^{1/2}$, is applied directly to the inverting input to be amplified by the closed-loop gain from that input. At 20Hz, the $C_1$, $C_2$, $R_2$ branch has little effect, gain is nearly 10,000, and RMS noise level slightly less than 2.88 × $10^{-6}$ V/Hz$^{1/2}$ appears at the output due to $R_3$.

This squares (no pun intended) reasonably well with the PSpice-predicted noise value of 7.909 × $10^{-12}$ V$^2$/Hz. In fact, the PSpice-calculated gain at 20Hz is 9,769 due to the fact that open-loop gain is not infinite and some shunting due to the $C_1$, $C_2$, $R_2$ branch occurs. Using this gain value yields output noise of $(9{,}769 \times e_{n\text{-}R3})^2$ = 7.906 × $10^{-12}$ V$^2$/Hz, giving some comfort that the analysis so far is correct.

Anyhow, to continue, $R_1$'s noise influence is primarily due to noise current $i_{n\text{-}R1} = e_{n\text{-}R1}/R_1$ = 2.88 × $10^{-8}$ /49,900 = 5.76 × $10^{-13}$ A/Hz$^{1/2}$ flowing through $R_3$, where it generates noise voltage, $(e_{n\text{-}R1}/R_1)R_3$ = 2.88 × $10^{-12}$ V/Hz$^{1/2}$. This noise voltage, too, is amplified 9,769 times to yield [9,769 × (2.88 × $10^{-12}$ V/Hz$^{1/2}$)]$^2$ = 7.922 × $10^{-16}$ V$^2$/Hz at the output, which squares with the 7.953 × $10^{-16}$ V$^2$/Hz predicted by PSpice. PSpice models only Johnson noise for resistors, so on that basis it will tend to give lower noise projections than a real circuit can realize.

Amazingly enough, $R_3 \cong R_1/10{,}000$ produces about 10,000 times the noise at the output than that produced by $R_1$. This ratio is true until shunting of $R_1$ noise current by the $C_1$, $C_2$, $C_3$ branch becomes significant at very high frequency. However, the absolute amount of output noise generation will fall with frequency due to falling gain dictated by the RIAA network.

Total output noise is the sum of all

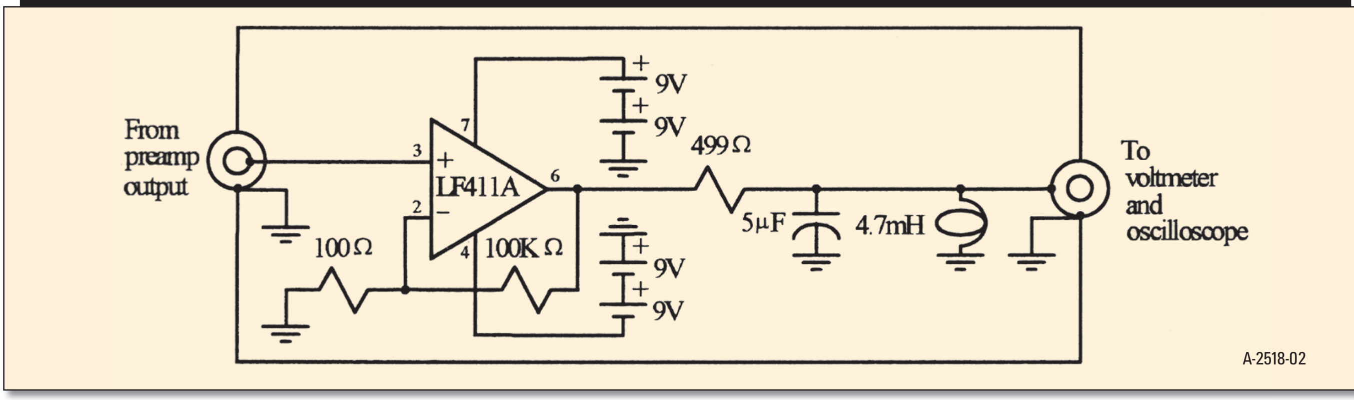
contributions from each component multiplied by the appropriate transfer function (gain) from component to output. You must use the squared amplitudes when summing uncorrelated values. For this example, the values with units $V^2/Hz$ could be summed, but not those with units $V/Hz^{1/2}$.

To determine the RMS output voltage, take the square root of the summed squared values of all the individual contributions. In practice, noise is treated as though the amplifier were noiseless with an exogenous noise source applied to its input. To find this noise source value, divide the total output noise by the closed-loop gain of the amplifier at the frequency of interest.

By the way, RMS noise voltage can be physically measured, but not, strictly speaking, on a $V/Hz^{1/2}$ basis. Rather, it can be measured with a bandwidth-restricted voltmeter as $V_{meter} = v_n B^{1/2}$, where $B$ is the bandwidth. In the limit as $B \rightarrow 0$, the spot-frequency noise could in theory be found. Of course, the voltmeter ultimately responds to power and as $B \rightarrow 0$, the voltmeter reading becomes immeasurably small.

Due to the low levels of noise in high-fidelity audio equipment, amplification of the output of the component under test is typically required if measurements are to be made. In amplifying the noise to be measured, the addition of noise by this extra amplifier must be accounted for. Despite the apparent complexities, meaningful noise measurements can be performed[8]. While I knew this to be true because I devised a system and made a limited number of noise measurements for my master's

**FIGURE 2: Spot noise measurement system.**



thesis[9], I nonetheless was reluctant to make noise measurements.

While I did not need to worry about it in my thesis measurements, electromagnetic interference (EMI) at 60Hz and 120Hz can be a major aggravation in audio system noise measurements. One laboratory in our engineering school had an EMI-proof room (Faraday cage) effected through complete shielding and elimination of AC power lines into the room. A room is not required but it may be necessary to place the component under test inside a metal box such as a large chassis. The box should preferably be steel or some ferromagnetic material because aluminum enclosures for electronic equipment, for example, while effective for electrostatic shielding, are less effective for shielding from magnetic fields[10].

Some of my senior design students, as part of a noise measurement project on a tube-based microphone amplifier under my guidance, made a very nice steel enclosure with only one input and one output connector allowing access to the inside of the enclosure. After their project was completed, I bought the enclosure from them. **Figure 2** shows the noise measurement system, consisting of a battery-powered op amp followed by a 1kHz center frequency, 100Hz (noise) bandwidth RLC bandpass filter placed inside the steel enclosure[11]. Gain-Bandwidth Product of the LF411A is guaranteed to be at least 3MHz, so the measurement should be valid even in the gain-of-1000 configuration used.

With the 10mV scale selected, a HP 427A voltmeter read zero with the op amp input connector shorted. With a shorted RCA plug inserted into the phono preamplifier's input RCA jack and corresponding preamp output connected to the filter/amplifier, the voltmeter reading was about 5.5mV. An oscilloscope tracing on its 2mV scale was compatible with the meter reading.

The HP 427A is a wide-bandwidth meter but it is not a true RMS voltmeter, so I used a 1dB correction factor[12] to conclude that noise was 6.2mV RMS. Adding some feeling of validity to the measurements was the fact that both preamp channels measured essentially the same while the phono preamp of my 2001 article measured significantly worse. Also, measured noise increased somewhat, after removing the short from the preamp input.

Due to component tolerances, filter mid-band frequency was just less than 1kHz. Preamplifier gain measured 1,006 at that frequency. Referred to the input of the phono preamplifier, this becomes 6.2mV/[(op amp gain)(preamp gain)] = 6.2 nV in a 100Hz bandwidth.

Spot noise is found by dividing the input noise by the square root of the bandwidth, which yields 620 $pV/Hz^{1/2}$. This is not too much above the input noise predicted by PSpice at 1kHz. Assuming that this is white noise, which it almost certainly is, this corresponds to a resistor of the value shown in **Equation 9**.

This is not the best that can be done, but it is very respectable. By comparison, the LF411A op amp has input spot noise of about 25 $nV/Hz^{1/2}$ at 1kHz.

With input shorted, unfiltered output noise measured about 2mV on the HP

427A, whose bandwidth is advertised as 1MHz but which measured over 2MHz in testing on my bench. It hardly matters, though, because the preamp gain continues to fall at 20dB/decade from 100 at 20kHz until the zero corresponding to $\tau_4$ is reached. Thus, output noise will be weighted toward lower frequencies due to the preamp's falling gain at higher frequencies, which is exactly the effect intended by RIAA equalization. Compared to a signal output of 12V RMS, signal-to-noise ratio is 20log (12/0.002) = 76dB.

In my view, noise is best evaluated subjectively. In this regard, the preamp is a superb performer. Nothing is audible with the ear placed immediately at the loudspeakers' high-frequency drivers as though the system were not powered. There is an impression with the ear placed immediately in front of the woofers of a low-frequency phenomenon similar to that heard when the ear is placed inside an open tube. This becomes inaudible if the ear is more than a few inches in front of the low-frequency drivers.



**Equation 9.**

$$V_{noise}(rms) = \sqrt{4kTRB} \Rightarrow R = \left[\frac{V_{spot\ noise}(rms)}{\sqrt{4kT}}\right]^2 = \left[\frac{6.2\times10^{-10}}{1.27\times10^{-10}}\right]^2 \cong 24\Omega.$$

This is very likely flicker noise and according to PSpice emulation, the biggest noise contribution at low frequencies is flicker noise from the input JFETs $J_{1A}$– $J_{1D}$. Total noise contribution decreases by roughly 40dB/decade, about ten times faster than the decrease in gain due to RIAA compensation. If you take the effect of falling gain into account, this certainly is the characteristic of "flicker noise," also for obvious reasons termed 1/*f* noise. This is interesting because some texts[13] indicate this should be true while others[14] indicate that Johnson noise should be the primary source of noise for an amplifier with JFET input stage whose driving source, in this instance a moving-coil cartridge, has low resistance.

In any event, paralleling JFETs can reduce noise with the penalties being added circuit complexity, the need to match devices, and increased input capacitance. This latter penalty, in general, can reduce bandwidth and interact with moving-magnet cartridge inductance to alter frequency response. As one text argues[15], $g_m/C_{rss}$, where $g_m = 2(I_D I_{DSS})^{1/2}/V_P$ is the transconductance and $C_{rss}$ is the common-source reverse transfer capacitance, might be considered a high-frequency figure of merit.

Because $C_{rss} = C_{gd}$ is the capacitance that will be multiplied by Miller effect, placing the JFET(s) in a cascode can help by reducing Miller effect. Since $g_m$ at a given drain current value and $C_{rss}$ are both intrinsic properties of a JFET, the device that maximizes $g_m/C_{rss}$ should be chosen if high-frequency performance is to be optimal. Obviously, the application at hand is for audio frequencies, not *r.f.* Still, it is important not to allow Miller effect to adversely impact performance through interaction with the phono cartridge.

Noise reduction is the reason that I used four JFETs at the input in this design. One reason for choosing the 2SK170 was the manufacturer's spec sheet assertion that the device was intended for, among other things, use in moving coil "head amps." Capacitance is not as important a factor because the low source resistance and inductance of moving-coil cartridges (you hope) makes them rather insensitive to capacitive loading. Even so, the topology is that of a BiFET cascode with $Q_1$ as the common-base portion.

Erno Borbely rightly argued that operating the JFETs at high drain currents can improve noise performance and also noted that overzealous application of this fact can actually increase noise[16]. The improvement is proportional to $(I_D)^{-1/4}$, which is not so dramatic to prompt the designer to increase drain current at the expense of degradation of other important properties[17].

It should be admitted that BJTs are generally superior to JFETs in noise performance when driving resistance is small, as it certainly is with most moving-coil cartridges. JFETs, however, are

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generally easier to interface. This is because it is undesirable for bias currents to flow through the cartridge, and BJT bias conditions may require resistors that are large enough to produce significant noise in a circuit where overall impedance levels are too low to allow for effective bypass except with very large electrolytic capacitors. Even so, there are solutions that do allow BJT use while avoiding the above pitfalls[18].

I, myself, designed a common-base input phono preamp with low-frequency noise performance significantly better than the JFET-input design offered here. If the actual circuit performs as well as its PSpice emulation suggests, then perhaps that design could be the subject of a subsequent article.

To quickly march through remaining noise issues, consider that only the input stage is important here. From a noise standpoint, the designer is free to do almost anything with the second stage, including use of MOSFETs if desired.

I chose the first stage BC549 and BC559 BJTs on the basis of noise properties, although there may be other devices even better in this regard. The widely available 2N4401 and 2N4403 types could be directly substituted with only a slight increase in noise. However, the former transistors were readily and cheaply obtained from Mouser, so it is with no trepidation that they are specified.

Design specifics will be covered next month in **Part 2**. *aX*

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