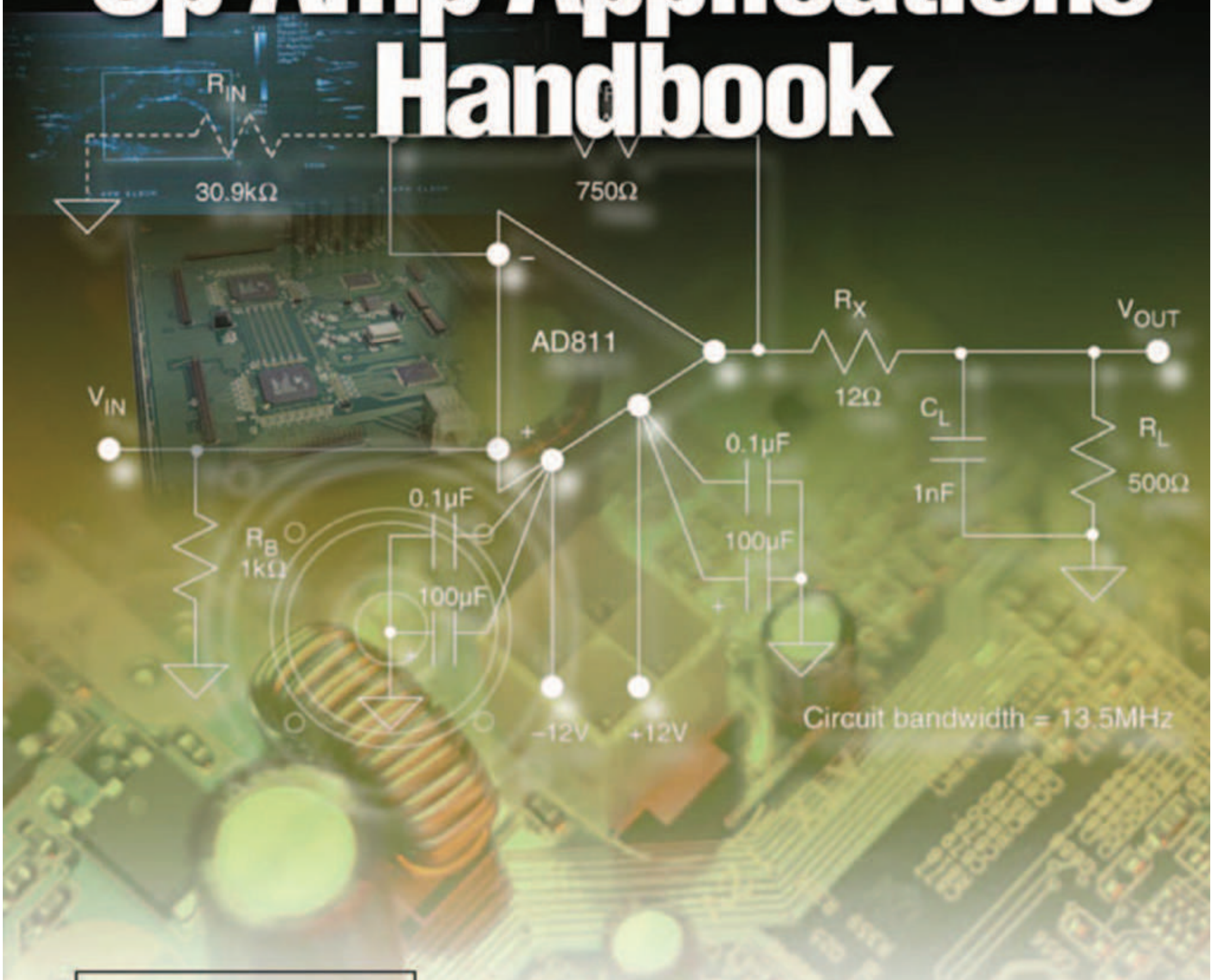
ELSEVIER

Op Amp Applications Handbook



Analog Devices, Inc.

By Walt Jung

ANALOG DEVICES

Newnes

The basic circuit as shown is single-ended with V_{OUT} taken from R8. However, a transformer can be simply added, as an option for driving balanced lines. When this is done, a nickel core type is suggested, for lowest distortion. One type suitable would be a Jensen JT-11-DM (or similar). It is coupled to the U1 output via a 10 Ω resistor.

Just as shown the circuit is suited for local, higher impedance loads of 1 kΩ and more. For very high levels of output drive or to drive long lines, a dedicated high current output driver should be used with U1, as generally described in the "Line Drivers" section. This can be most simply implemented by making U1 a composite amplifier, using a AD797 input section plus a follower-type output stage. A good choice for this would be a BUF04 IC, connected between Pin 6 of the AD797 and the remaining circuitry. The buffer will isolate the U1 stage, allowing it to operate with highest linearity with difficult loads. Note also that ±17 V supplies won't be necessary with the AD797 unless extreme voltage swings are required. More conventional (±15 V) supplies will minimize the U1 heating.

References: Microphone Preamplifiers

1. C. D. Motchenbacher, F. C. Fitchen, **Low-Noise Electronic Design**, Wiley, New York, 1973, ISBN 0-471-61950-7.
2. G. Bore, "Powering Condenser Microphones," **db**, June 1970.
3. "ANSI Standard 268-15 (Revision 1987, amendments 1989, 1990, 1991)." American National Standards Institute, 11 W. 42nd St., New York, NY, 10036.
4. Steve Hogan, "Standard Mic Input Application," Jensen Transformers Application note JT99-0003, November, 1992.
5. W. Jung, M. Stephens, C. Todd, "Slewing Induced Distortion & Its Effect on Audio Amplifier Performance—With Correlated Measurement/Listening Results," **presented at 57th AES convention**, May 1977, AES preprint # 1252.
6. W. Jung, M. Stephens, C. Todd, "An Overview of SID and TIM," Parts 1–3, **Audio**, June, July, August, 1979.
7. Walter G. Jung, **Audio IC Op Amp Applications, 3rd Ed.**, Howard W. Sams & Co., 1987, ISBN 0-672-22452-6.
8. W. Jung, A. Garcia, "A Low Noise Microphone Preamp with a Phantom Power Option," **Analog Devices AN242**, November 1992.
9. Walt Jung, "Audio Preamplifiers, Line Drivers, and Line Receivers," within Chapter 8 of Walt Kester, **System Application Guide**, Analog Devices, Inc., 1993, ISBN 0-916550-13-3, pp. 8-1 to 8-100.
10. Walt Jung, "Microphone Preamplifiers for Audio," **Analog Dialogue**, Vol. 28, No. 2, 1994, pp. 12–18.

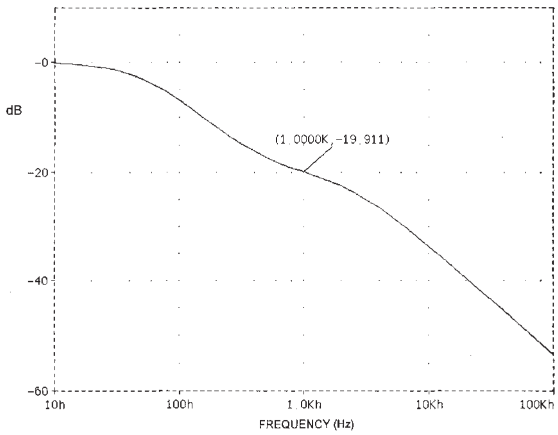
RIAA Phono Preamplifiers

An example of an audio range preamplifier application requiring equalized frequency response is the RIAA phono preamp. While LP record sales have faded with the establishment of new digital media, for completeness equipment is still designed to include phono playback stages. RIAA preamp stages, as amplifiers with predictable, nonflat frequency response, have more general application connotations. The design techniques within this section are specific to RIAA as an example, but they are also applicable to other frequency dependent amplitude designs in general. The techniques are also useful as a study tool, considering the various approaches advanced to optimize the function of high performance gain with predictable equalization (EQ). These last two points make these discussions useful in a much broader sense.

Some RIAA Basics

The RIAA equalization curve (see Reference 1) is shown in Figure 6-7, expressed as it is relative to dc. This curve indicates maximum gain below 50 Hz (f1), with two high frequency inflection points. Above f1, the gain rolls off at 6 dB/octave until a first high frequency breakpoint is reached at 500 Hz (f2). Gain then remains relatively constant until a second high frequency breakpoint is reached at about 2.1 kHz (f3), where it again rolls off at 6 dB/octave through the remainder of the audio region and above.

Figure 6-7: Ideal RIAA de-emphasis (time constants of 3180 μs, 318 μs, 75 μs)



Use of a low frequency roll-off (f0, not shown) is at the option of the designer. Frequency response can be extended towards dc, or, alternately, rolled off at a low frequency below 50 Hz. When applied, this roll-off is popularly called a "rumble" filter, as it reduces turntable/record related low frequency disturbances, lessening low frequency driver overload. This roll-off may or may not coincide with a fourth time constant (below).

However, gain at the frequencies f1, f2, and f3 describes the basic RIAA curve. In the standard, this is described in terms of three corresponding time constants, T1, T2, and T3, defined as 3180 μs, 318 μs, and 75 μs, respectively (Reference 1, again). The T1–T3 are here described as they correspond to *ascending frequency*, the reverse of the terminology in Reference 1 (however, the time constants themselves are identical). In some literature one may occasionally find the frequencies corresponding to T1, T2, and T3 referenced. These exact frequencies can be found simply by the basic relationship of:

$$f = 1/(2 \bullet \pi \bullet T) \qquad \text{Eq. 6-4}$$

So, for the three time constants specified, the frequencies are:

$$f1 = 1/T1 = 1/(2 \bullet \pi \bullet 3180E-6) = 50 \text{ Hz}$$

$$f2 = 1/T2 = 1/(2 \bullet \pi \bullet 318E-6) = 500 \text{ Hz}$$

$$f3 = 1/T3 = 1/(2 \bullet \pi \bullet 75E-6) = 2122 \text{ Hz}$$

An IEC amendment to the basic RIAA response adds a fourth time constant of 7950 μs, corresponding to an f0 of 20 Hz when used (see Reference 2). Use of this roll-off has never been standardized in the US, and isn't treated in detail here.

The characteristic gain in dB for an RIAA preamp is generally specified relative to a 1kHz reference frequency. For convenience in evaluating the RIAA curve numerically, Figure 6-8 is a complete 10 kHz–100 kHz relative decibel table for the three basic RIAA time constants. From these data several key points

FREQ	VDB(6)[1]	VDB(5)[2]
1.000E+01	1.974E+01	–1.684E-01
1.259E+01	1.965E+01	–2.639E-01
1.585E+01	1.950E+01	–4.109E-01
1.995E+01	1.928E+01	–6.341E-01
2.512E+01	1.895E+01	–9.654E-01
3.162E+01	1.847E+01	–1.443E+00
3.981E+01	1.781E+01	–2.103E+00
5.012E+01	1.694E+01	–2.975E+00
6.310E+01	1.584E+01	–4.067E+00
7.943E+01	1.455E+01	–5.362E+00
1.000E+02	1.309E+01	–6.823E+00
1.259E+02	1.151E+01	–8.398E+00
1.585E+02	9.877E+00	–1.003E+01
1.995E+02	8.236E+00	–1.167E+01
2.512E+02	6.645E+00	–1.327E+01
3.162E+02	5.155E+00	–1.476E+01
3.981E+02	3.810E+00	–1.610E+01
5.012E+02	2.636E+00	–1.727E+01
6.310E+02	1.636E+00	–1.828E+01
7.943E+02	7.763E–01	–1.913E+01
1.000E+03	**8.338E–07**	**–1.991E+01**
1.259E+03	–7.682E–01	–2.068E+01
1.585E+03	–1.606E+00	–2.152E+01
1.995E+03	–2.578E+00	–2.249E+01
2.512E+03	–3.726E+00	–2.364E+01
3.162E+03	–5.062E+00	–2.497E+01
3.981E+03	–6.572E+00	–2.648E+01
5.012E+03	–8.227E+00	–2.814E+01
6.310E+03	–9.992E+00	–2.990E+01
7.943E+03	–1.184E+01	–3.175E+01
1.000E+04	–1.373E+01	–3.365E+01
1.259E+04	–1.567E+01	–3.558E+01
1.585E+04	–1.763E+01	–3.754E+01
1.995E+04	–1.960E+01	–3.951E+01
2.512E+04	–2.158E+01	–4.149E+01
3.162E+04	–2.357E+01	–4.348E+01
3.981E+04	–2.557E+01	–4.548E+01
5.012E+04	–2.756E+01	–4.747E+01
6.310E+04	–2.956E+01	–4.947E+01
7.943E+04	–3.156E+01	–5.147E+01
1.000E+05	–3.356E+01	–5.347E+01

Notes: (1) Denotes 1 kHz 0 dB reference
(2) Denotes dc 0 dB reference

Figure 6-8: Idealized RIAA frequency response referred to 1 kHz and to dc

can be observed: If the 1 kHz gain is taken as the zero dB reference, frequencies below or above show higher or lower dB levels, respectively (Note 1, column 2). With a dc 0 dB reference, it can be noted that the 1 kHz gain is 19.91 dB below the dc gain (Note 2, column 3).

Expressed in terms of a gain ratio, this means that in an ideal RIAA preamp the 1 kHz gain is always 0.101 times the dc gain. The constant 0.101 is unique to all RIAA preamp designs following the above curve, therefore it can be designated as "K_{RIAA}", or:

$$K_{RIAA} = 0.101 \qquad \text{Eq. 6-5}$$

This constant logically shows up in the various gain expressions of the RIAA preamp designs following. In all examples discussed here (and virtually all RIAA preamps in general), the shape of the standard RIAA curve is fixed, so specifying gain for a given frequency (1 kHz) also defines the gain for all other frequencies.

It can also be noted from the RIAA curve of Figure 6-7 that the gain characteristic continues to fall at higher frequencies. This implies that an amplifier with unity-gain stability for 100% feedback is ultimately required, which can indeed be true, when a standard feedback configuration is used. Many circuit approaches can be used to accomplish RIAA phono-playback equalization; however, all must satisfy the general frequency response characteristic of Figure 6-7.

Equalization Networks for RIAA Equalizers

Two equalization networks well suited in practice to RIAA phono reproduction are illustrated in Figure 6-9a and 6-9b, networks N1 and N2. Both networks with values as listed can yield with high accuracy the three standard RIAA time constants of 3180 µs, 318 µs, and 75 µs as outlined by network theory (see References 3-6). For convenience, both theoretical values for the ideal individual time constants are shown at the left, as well as closest fit standard "no trim" values to the right. Designers can, of course, parallel and/or series RC values as may be deemed appropriate, adhering to network theory.

There are of course an infinite set of possible RC combinations from which to choose network values, but practicality should rule any final selection. A theoretical starting point for a network value selection can begin with *any* component, but in practice the much smaller range of available capacitors suggests their

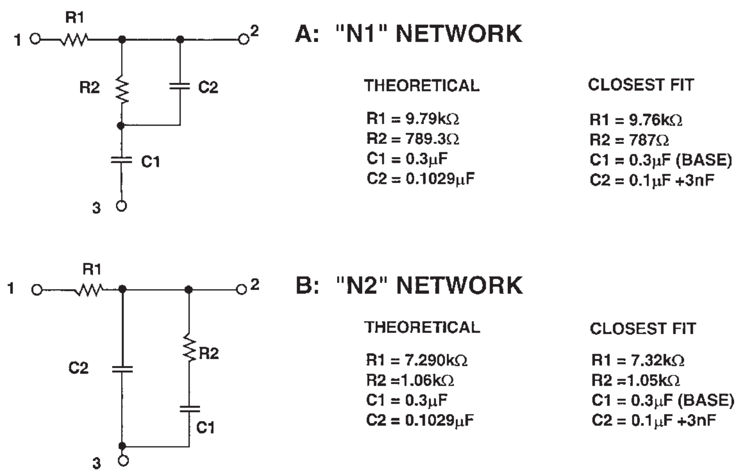


Figure 6-9: Two RIAA EQ networks (T1 = 3180 µs, T2 = 318 µs, T3 = 75 µs)

selection first, then resistors, since they have a much broader span of (stock) values. Note that precision film resistors can in fact be obtained (on special order) in virtually *any* value, up to several megohms. The values listed here are those taken as standard from the E96 series.

Very high standards of EQ accuracy are possible, to tolerances of noticeably better than ±0.1 dB (see for example data from Reference 8, also quoted in 6). In the design process, there are several distinct general aspects of EQ component selection which can impact the ultimate accuracy. These are worth placing in perspective before starting a design.

The *selection tolerance* of the component defines how far an ideal (zero manufacturing tolerance) component deviates from the theoretical value. A good design will seek to minimize this error by using either carefully selected standard values, or series and/or shunt combinations, so as to achieve selection tolerance of less than 1%, preferably zero.

The *manufacturing tolerance* of the component defines how far an otherwise ideal component deviates from its stated catalog value, such as ±1%, ±2%, and so forth. This can obviously be controlled by tighter specifications, but usually at some premium, particularly with capacitors of ±1% or less. Note that a "hidden" premium here can be long delivery times for certain values. Care should be taken to use standard stock values with capacitors—even to the extent that multiple standard values may be preferable (three times 0.01 µF for 0.03 µF, as an example).

Topology-related parasitics must also be given attention, as they can also potentially wreck accuracy. Amplifier gain-bandwidth is one possible source of parasitic EQ error. However, a more likely error source is the parasitic zero associated with active feedback equalizers. If left uncompensated below 100 kHz, this alone can be a serious error.

In any event, for high equalization accuracy to be "real," once a basic solid topology is selected, the designer must provide for the qualification of components used, by precise measurement and screening, or tight purchase tolerances. An alternative is iterative trimming against a reference standard such as that of Reference 9, but this isn't suited for production. An example is the data of Reference 8, derived with the network of Reference 9. If used, the utility of such a trim technique lies in the reduction of the equipment accuracy burden. While the comparator used needs to have high *resolution*, the accuracy is transferred to the network comparison standard used.

It should be understood that an appropriately selected high quality network will allow excellent accuracy, for example either N1 or N2 with the "closest fit" (single component) values of exact value yield a broadband error of about ±0.15 dB. Accuracy about three times better than this is achieved with the use of N1 and the composite C2, as noted. The composite C2 is strongly suggested, as without it there is a selection error of about 3%.

It is also strongly recommended that only the highest quality components be employed for use in these networks, for obvious reasons. Regardless of the quality of the remainder of the circuit, it is surely true that the equalization accuracy and fidelity can be no better than the quality of those components used to define the transfer function. Thus only the best available components are used in the N1 (or N2) RC network, selected as follows:

Capacitors—should have close initial tolerance (1%–2%), a low dissipation factor and low dielectric absorption, be noninductive in construction, and have stably terminated low-loss leads. These criteria in general are best met by capacitors of the Teflon, polypropylene and polystyrene film families, with 1%–2% polypropylene types being preferred as the most practical. Types to definitely avoid are the "high K" ceramics. In contrast, "low K" ceramic types, such as "NP0" or "COG" dielectrics, have excellent dissipation factors. See the passive component discussions of Chapter 7 on capacitors, as well as the component-specific references at the end of this section.

Resistors—should also be close tolerance (≤1%), have low nonlinearity (low voltage coefficient), be temperature stable, with solid stable terminations and low-loss noninductive leads. Types that best meet these criteria are the bulk metal foil types and selected thick films, or selected military grade RN55 or RN60 style metal film resistor types. See the passive component discussions of Chapter 7 on resistors, as well as the component-specific references at the end of this section.

It should be noted also that the specific component values suggested might not be totally optimum from a low impedance, low noise standpoint. But, practicalities will likely deter using appreciably lower ones. For example, one could reduce the input resistance of either network down to say 1 kΩ, and thus lower the input referred noise contribution of the network. But, this in turn would necessitate greater drive capability from the amplifier stage, and raise the C values up to 1 μF–3 μF, where they are large, expensive, and most difficult to obtain. This may be justified for some uses, where performance is the guiding criterion rather than cost effectiveness, or the amplifiers used are sufficiently low in noise to justify such a step. Regardless of the absolute level of impedance used, in any case the components should be adequately shielded against noise pickup, with the outside foils of C1 or C2 connected either to common or a low impedance point.

These very same N1/N2 networks can suffice for both active and passive type equalization. Active (feedback) equalizers use the network simply by returning the input resistor R1 to common, that is jumpering points 1–3, and employing the network as a two-terminal impedance between points 1+3, and 2. Passive equalizers use the same network in a three-terminal mode, placed between two wideband gain blocks.

RIAA Equalizer Topologies

Many different circuit topologies can be used to realize an RIAA equalizer. Dependent upon the output level of the phono cartridge to be used, the 1 kHz gain of the preamp can range from 30 dB to more than 50 dB.

Magnetic phono cartridges in popular use consist of two basic types: moving magnet (MM) and moving coil (MC). The moving magnet types, which are the most familiar, are suitable for the first two circuits described. The moving coil cartridge types are higher performance devices; they are less commonplace but still highly popular.

Functionally, both types of magnetic cartridges perform similarly, and both must be equalized for flat response in accordance with the RIAA characteristic. A big difference in application, however, is the fact that moving magnet types have typical sensitivities of about 1 mV of output for each cm/s of recorded velocity. In moving coil types, sensitivity on the order of 0.1 mV is more common (for a similar velocity). In application then, a moving coil RIAA preamp must have more gain than one for moving magnets. Typically, 1 kHz gains are 40 dB–50 dB for moving coils, but only 30 dB–40 dB for moving magnets. Noise performance of a moving coil preamp can become a critical performance factor however, because of low-output voltage and low impedance involved—typically this is in the range of just 3 Ω–40 Ω. The following circuit examples illustrate techniques that are useful to these requirements.

Actively Equalized RIAA Preamp Topologies

The most familiar RIAA topology is shown in general form in Figure 6-10, and is called an active feedback equalizer, as the network N used to accomplish the EQ is part of an active feedback path (see References 10, 11). In these and all of the following discussions it is assumed that the input from the pickup is appropriately terminated by R_t–C_t, which are selected for flat *cartridge* frequency response driving U1. The following discussions deal with the *amplification* frequency response, given this ideal input signal.

Assuming an adequately high gain amplifier for U1, the gain/frequency characteristics of this circuit are determined largely by the network. The gain of the stage is set by the values of the network N and R_3, and

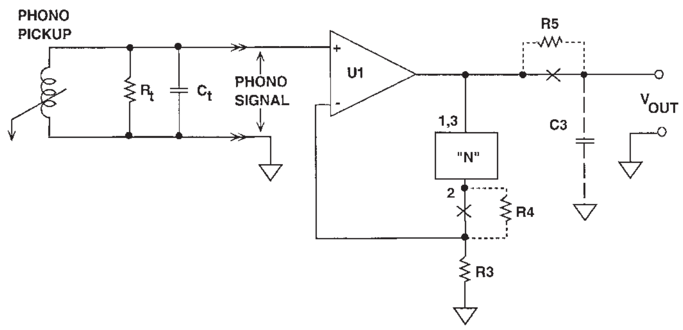


Figure 6-10: Active feedback RIAA equalizer

the U1 output is a low impedance, V_{OUT}. The 1 kHz gain of this stage is defined by the RIAA curve and resistors R_1 and R_3, and is:

$$G = 0.101 \bullet \left[1 + \left(R_1/R_3\right)\right] \qquad \text{Eq. 6-6}$$

where 0.101 is the constant K_{RIAA}. R_1 is within N; R_4 and R_5 are discussed momentarily.

As noted previously, an ideal RIAA response continues to fall with increasing frequency, and can in fact be less than unity at some high frequency (Figure 6-7, again). But, the basic U1 topology of Figure 6-10 can't achieve this, as the minimum gain seen at the output of U1 approaches unity at some (high) parasitic zero frequency, where the network equivalent series capacitive impedance of C_1 and C_2 is equal to R_3. At this zero frequency, the response from U1 simply levels off and ceases to track the RIAA curve.

However, in terms of practical consequence the error created by this zero may or may not be of significance, dependent upon where the zero falls (as determined by gain). If well above audibility (i.e., ≥ 100 kHz), it will introduce a small equalization error at the upper end of the audio range. For example, if it falls at 100 kHz, the 20 kHz error is only about 0.3 dB. Fortunately, this error is easily compensated by a simple low-pass filter after the amplifier, R_5–C_3. The filter time constant is set to match the zero T_4, which is:

$$T4 = R_3 \bullet C_{EQUIV} \qquad \text{Eq. 6-7}$$

where R_3 is the value required for a specific gain in the design.

C_{EQUIV} is the series equivalent capacitance of network capacitors C_1 and C_2, or:

$$C_{EQUIV} = \left(C_1 \bullet C_2\right)/\left(C_1 + C_2\right) \qquad \text{Eq. 6-8}$$

Here the C_{EQUIV} is 7.6 nF and R_3 200 Ω, so T4 = 1.5μs The product of R5 and C4 are set equal to T4, so picking a R5 value solves for C4 as:

$$C4 = T4/R5 \qquad \text{Eq. 6-9}$$

The 1.5 μs R_5-C_4 time constant is realized with R_5 = 499 Ω and C4 = 3 nF. This design step increases the output impedance, making it more load susceptible. This should be weighed against the added parts and loading. In general, R_5 should be low, i.e., ≤1 kΩ.

In some designs, a resistor R_4 (dotted in Figure 6-10) may be used with N (for example, for purposes of amplifier stability at a gain higher than unity). With R_4, T4 is calculated as:

$$T4 = \left(R_3 + R_4\right) \bullet C_{EQUIV} \qquad \text{Eq. 6-10}$$

The R_5–C_3 product is again chosen to be equal to this T4 (more on this below).

The next two schematics illustrate variations of the most popular approach to achieving a simple RIAA phono preamp, using active feedback, as just described. Figure 6-11 is a high performance, dc-coupled version using precision 1% metal film resistors and 1% or 2% capacitors of polystyrene or polypropylene type. Amplifier U1 provides the gain, and equalization components R1-R2-C1-C2 form the RIAA network, providing accurate realization with standard component values. N1 is the network, with 1 and 3 common.

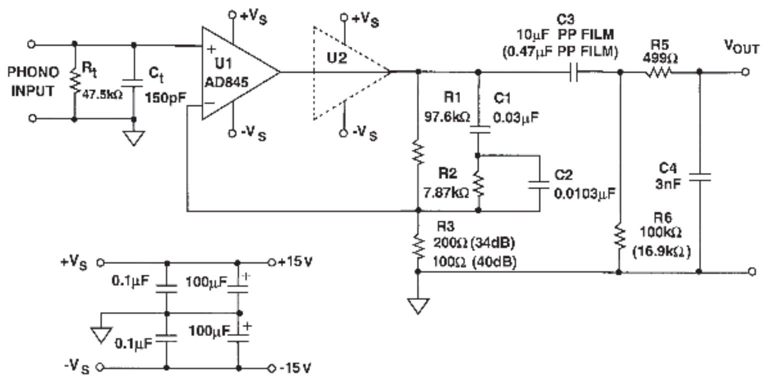


Figure 6-11: A dc-coupled active feedback RIAA moving magnet preamp

As mentioned, input RC components R_t-C_t terminate the moving magnet cartridge with recommended values (shown as typical). In terms of desired amplifier parameters for optimum performance, they are considerably demanding. For lowest noise from a cartridge's inductive source, the amplifier should have an input voltage noise density of $5\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$ or less (favoring a bipolar), and an input current noise density of $1\ \mathrm{pA}/\sqrt{\mathrm{Hz}}$ or less (favoring a FET). In either case, the 1/F noise corner should be as low as possible.

For bipolar-input amplifiers, dc input-bias current can be a potential problem when direct coupling to the cartridge, so in this circuit only a very low input bias current type is suggested. If a bipolar input amplifier is used for U1, it should have an input current of <<100 nA for minimum dc offset problems (assuming a typical phono cartridge of ≈1 kΩ resistance). Examples are the OP27, OP270 families. FET-input amplifiers generally have negligible bias currents but also tend typically to have higher voltage noise. FET-input types useful for U1 are the AD845 and OP42, even though their voltage noise is not as low as the best of the bipolar devices mentioned. On the plus side, they both have a high output current and slew rate, for low distortion driving the feedback network load (approximately the R3 value at high frequencies). Of the two, the OP42 has lower noise, the AD845 higher output current and slew rate.

For high gain accuracy at high stage gains, the amplifier should have a high gain-bandwidth product; preferably >5 MHz at audio frequencies. Because of the 100% feedback through the network at high frequencies, the U1 amplifier must be unity-gain-stable. To minimize noise from sources other than the amplifier, gain resistor R_3 is set to a relatively low value, which generates a low voltage noise in relation to the amplifier.

RIAA accuracy is quite good using the stock equalizer values. A PSpice simulation run is shown in Figure 6-12 for the suggested gain of 34 dB. In this expanded scale plot over the 20 kHz–20 kHz range, the error relative to the 1 kHz gain is less than ±0.1 dB.

As can be noted from Figure 6-12, the relative amplitude is expanded, to easily show response errors. A perfect response would be a straight line at 0 dB, meaning that the circuit under test had exactly the same gain as an ideal RIAA amplifier of the same 1 kHz gain. This high sensitivity in the simulation is done via the use of a feature in PSpice allowing the direct entry of Laplace statements (see Reference 10). With this evaluation tool, the ideal transfer function of an RIAA equalizer can be readily generated. The key parameters are the three time constants described above, and the ideal dc gain.

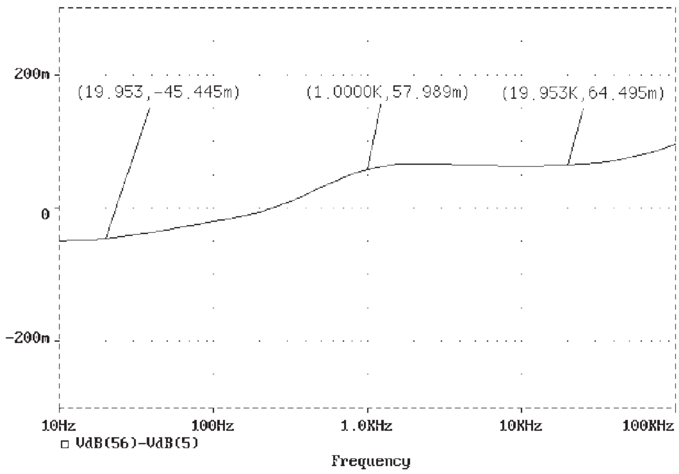


Figure 6-12: Relative error (B) versus frequency for dc-coupled active feedback RIAA moving magnet preamp, gain of 34 dB

The syntax to enable this mode of comparison is contained in the listing of Figure 6-13, which is the PSpice CIR file for the circuit of Figure 6-11. The Laplace details are all contained within the dotted box, and need only the editing of one value, "ENORM," for gain normalization from one circuit to another (see boldface). In this case ENORM is set to 490.7, to match the ideal R1 and R3 values of Figure 6-13. When the analysis is run, a difference display of the circuit-under-test and the ideal outputs (i.e., VdB(56)–VdB(5)) shows the relative response (Figure 6-12). Vertical axis scaling is easily adjusted for sensitivity, and is ±300 mB as displayed in Figure 6-12.

```
RIAA34LP: 34 dB gain RIAA preamp with AD845
*
.OPT ACCT LIST NODE OPTS NOPAGE LIBRARY
.AC DEC 10 10 100KHZ
.LIB D:\PS\ADLIB\AD_RELL.LIB
.PRINT AC VDB(5) VDB(56)
.PROBE
VIN 1 0 AC 1E-3
VCC   52  0  +15V
VEE   53  0  -15V
* ---------- V(5) = idealized RIAA frequency response -----------------
*
* Uses Laplace feature of PSpice Analog Behavioral option
* for frequency response reference.
* ENORM = ideal U1 DC gain = 1+(R1/R3) Use ideal values for R1, R3
* T1 - T3 are time constants desired (in µs).
* Input = node 1, Laplace Output = node 5
.PARAM ENORM  = {490.7}
.PARAM T1 = {3180}      ; Reference RIAA constants, do not alter!
.PARAM T2 = {318}       ; Reference RIAA constants, do not alter!
.PARAM T3 = {75}        ; Reference RIAA constants, do not alter!
*
ERIAA 5 0 LAPLACE {ENORM*V(1)}={(1+(T2*1E-6)*S)/((1+(T1*1E-6)*S)*(1+(T3*1E-6)*S))}
RDUMMY5 5 0 1E9
*
* --------------------------------------------------------------------
*
*    (+) (-) V+  V-  OUT
XU3   1  21  52  53  55 AD845
*  Active values         Theoretical values
R1 55  21  97.6K          ; 97.9k
R2 21  8   7.87K          ; 7.8931563k
C1 55  8   30NF           ; 30nF
C2 21  8   10.3NF         ; 10.2881nF R3 21  0   200            ; 199.9148
C3 55  100 10E-6
R6 100 0   100K R5 100 56  499
C4 56  0   3.0000E-9
.END
```

Figure 6-13: An example PSpice circuit file that uses the Laplace feature for ideal RIAA response comparison

The 1 kHz gain of this circuit can be calculated from Eq. 6-6. For the values shown, the gain is just under 50 times (≈34 dB). Higher gains are possible by decreasing R_3, but gains >40 dB may show increasing equalization errors, dependent upon amplifier bandwidth. For example, R_3 can be 100 Ω for a gain of about 100 times (≈40 dB). Note that if R_3 is changed to 100 Ω, C_4 should also be changed to 1.5 nF, to satisfy Eq. 6-9.

Dependent upon the amplifier in use, this circuit is capable of very low distortion over its entire range, generally below 0.01% at levels up to 7 V rms, assuming ±15 V supplies. Higher output with ±17 V supplies is possible, but will require a heat sink for the AD845. U2 is an optional unity-gain buffer useful with some op amps, particularly at higher gains or with a low-Z network. But this isn't likely to be necessary with U1 an AD845.

For extended low-frequency response, C_3 and R_6 are the large values, with C_3 preferably a polypropylene film type. If applied, the alternate values form a simple 6 dB per octave rumble filter with a 20 Hz corner. As can be noted from the figure's simplicity, C_3 is the only dc blocking capacitor in the circuit. Since the dc circuit gain is on the order of 54 dB, the amplifier used must be a low offset-voltage device, with an offset voltage that is insensitive to the source. Since these preamps are high gain, low level circuits (≥50 dB of gain at 50 Hz/60 Hz), supply voltages should be well regulated and noise-free, and reasonable care should be taken with the shielding and conductor routing in their layout.

Alternately, an inexpensive ac-coupled form of this circuit can be built with higher bias current, low noise bipolar op amps, for example the OP275, I_B = 350 nA(max), which would tend to make direct coupling to a cartridge difficult. This form of the circuit is shown in Figure 6-14, and can be used with many unity gain stable bipolar op amps.

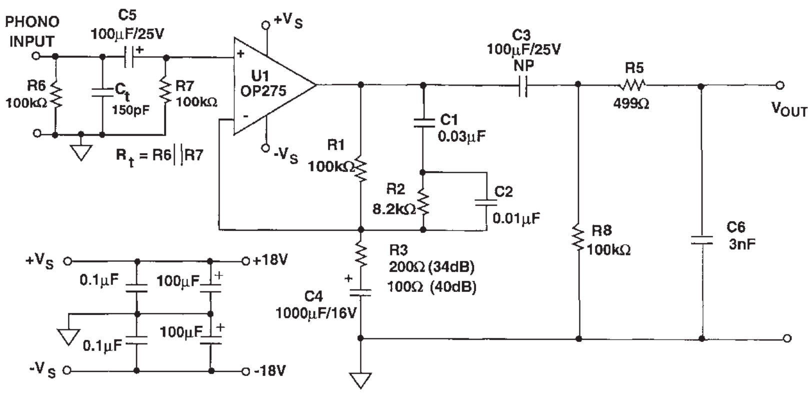


Figure 6-14: An ac-coupled active feedback moving magnet RIAA preamp

Here input ac coupling to U1 is added with C_5, and the cartridge termination resistance R_t is made up of the R_6-R_7 parallel equivalent. R_3 of the feedback network is ac-grounded via C_4, a large value electrolytic. These measures reduce the dc offset at the output of U1 to a few mV. Nearest 5% values are also used for the network components, making it easily reproducible and inexpensive. C_3 is a nonpolar electrolytic type, and the R_3-C_4 time constant as shown provides a corner frequency of ≤1 Hz at the 34 dB gain.

Frequency response of this version (not shown) isn't quite as good as that of Figure 6-11, but is still within ±0.2 dB over 20 Hz–20 kHz (neglecting the effects of the low frequency roll-off). If a tighter frequency response is desired, the N1 network values can be adjusted. With a higher rated maximum supply voltage for the OP275, the power supplies of this version can be ±21 V if desired, for outputs up to 10 V rms.

There is another, very useful variation on the actively equalized RIAA topology. This is one that operates at appreciably higher gain and with lower noise, making it suitable for operation with higher output moving coil (MC) cartridges. In this design example, shown in Figure 6-15, the basic circuit is used is quite similar to that of Figure 6-11. The lower R_t and C_t values shown are typical for moving coil cartridges. They are of course chosen per the manufacturer's recommendations (in particular the resistance).

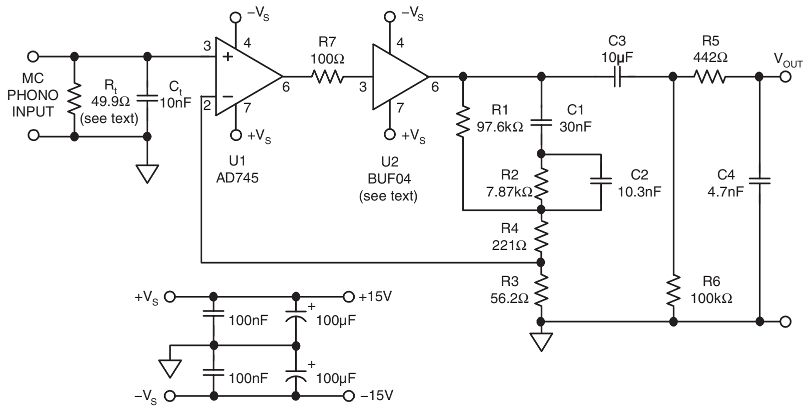


Figure 6-15: A low noise dc-coupled active feedback RIAA moving coil preamp with 45 dB of gain

To make it suitable for a high-output MC cartridge, a very low-noise FET op amp is used for U1, the AD745. The AD745 is stable at a minimum gain of five times, as opposed to the unity-gain stable op amps of the prior examples. This factor requires a modification to gain resistors R1-R3. This is the inclusion of an extra resistor, R4. With the ratio shown, R3 and R4 form a 5/1 voltage divider for the voltage seen at the bottom of network N (the R1-R2-C2 node). This satisfies U1's gain-of-five stability requirement.

In this gain setup, R3 is still used for the gain adjustment, and R1-R2-C1-C2 still form the basic N1 RIAA network. With R4 used, Eq. 6-9 is used to calculate the T4 time constant. With C4 chosen as a standard value, R5 is then calculated. With these N1 network values and a 45 dB 1 kHz gain, R3 is 56.2 Ω, which is still suitable as a low noise value operating with either an AD745 or an OP37 used for U1.

Some subtle points of circuit operation are worth noting. The dc gain of this circuit is close to 1800, which can result in saturation of U1 if offset isn't sufficiently low. Fortunately, the AD745 has a maximum offset of 1.5 mV over temperature, making the output referred offset always less than 3 V. While this may limit the maximum output swing some due to asymmetrical clipping, 5 V rms or more of swing should be available operating from ±15 V supplies. Coupling capacitor C3 decouples the dc output offset at U2, so any negative consequences of dc-coupling the U1 gain path are minimal.

For minimal loading of U1 and maximum linearity at high gains, the unity-gain buffer amplifier U2 is used, a BUF04. The BUF04 is internally configured for unity-gain operation, and needs no additional components. Note that this buffer is optional, and is not absolutely required. Other buffer amplifiers are discussed later in this chapter.

This Figure 6-15 circuit was analyzed with PSpice using the Laplace comparison technique earlier described, and the results are displayed in Figure 6-16. As was true previously, the vertical scaling of this

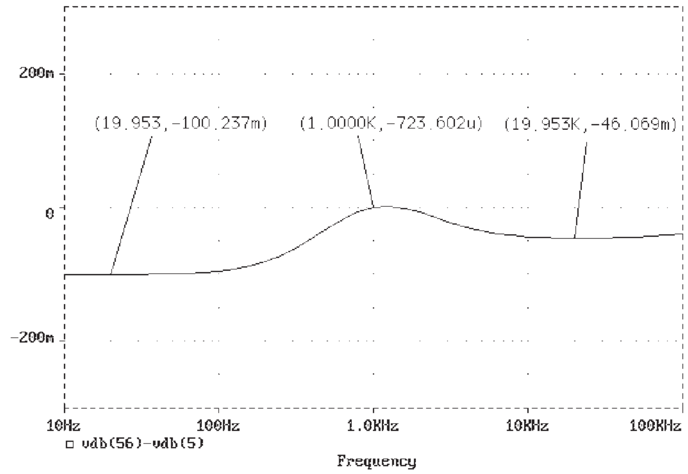


Figure 6-16: Relative error (B) versus frequency for dc-coupled active feedback RIAA moving coil preamp, gain of 45 dB (simulation)

display is very sensitive; ±300 mB (or ±0.3 dB). Thus, placed in context, gain errors relative to 1 kHz over 20 kHz–20 kHz are extremely small, ≈0.1 dB. Lab measurements of the circuit were also consistent with the simulation. Of course in terms of audible effects, errors of ±0.1 dB or less aren't likely to be apparent.

Distortion/noise measurements of the circuit are essentially dominated by noise (as opposed to actual distortion) measuring ~0.01% THD + N or less, over output levels ranging from 0.5 to 5 V rms, from 20 Hz–20 kHz. Of course, as with any high gain circuit, layout and lead dress into the circuit are extremely critical to noise, and must be arranged for minimum susceptibility. Supply voltages must be low in noise, and well regulated.

This exercise has illustrated both the basic design process of the active RIAA equalizer, as well as a convenient SPICE analysis method to optimize the design for best frequency response. It is not suggested that the exact network values shown of the examples are the only ones suitable. To the contrary, great many sets of values can be used with success comparable to that shown above.

This final active equalizer circuit example is the best of the bunch, and has a virtue of being easily adapted for other operating conditions; i.e., higher gain, other networks, and so forth. For example, note that even lower noise MC operation is possible, by using the $\leq 1\ \text{nV}/\sqrt{\text{Hz}}$ AD797 for U1, and scaling the N1 RC components further downward. This will have the desirable effect of making R3 lower than 50 Ω, which minimizes the R1–R4 network's noise. Note that gains of 50 dB or more are also possible, suitable for very low output moving coil cartridges (given suitable attention to worst-case U1 offsets).

Passively Equalized RIAA Preamp Topologies

Another RIAA design approach is the so-called *passively equalized* preamp (see Reference 11). This topology consists of two high quality, wideband gain blocks, separated by a three terminal passive network, N (N can be either network N1 or N2). The gain blocks are assumed very wide in bandwidth, so in essence the preamp's entire frequency response is defined by the passive network, thus the name passively equalized.

A circuit topology useful for such RIAA phono applications is shown in Figure 6-17. This circuit consists of two high-quality wide bandwidth gain blocks, U1 and U2, as discussed above. Selection of these amplifiers and their operating conditions optimizes the preamp for gain, noise, and overload characteristics. The circuit can be set up for either MM or MC operation by simple value changes and op amp selection.

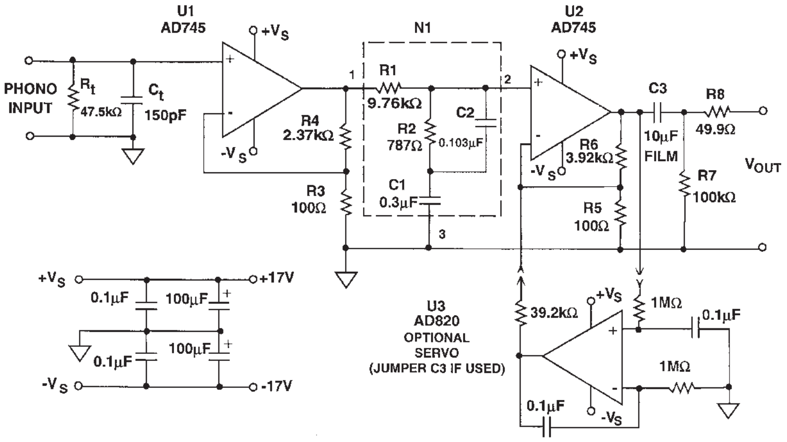


Figure 6-17: A passively equalized RIAA preamp with 40 dB gain

The gain stages are set up for the required total gain, via R_4-R_3 and R_6-R_5. In general, the total 1 kHz gain of this circuit G is:

$$G = 0.101 \bullet \left[1 + (R_4/R_3)\right] \bullet \left[1 + (R_6/R_5)\right] \qquad \text{Eq. 6-11}$$

The op amp gain blocks could be made identical for purposes of simplicity but are not necessarily so for the following reasons. A preamplifier topology such as this must be carefully optimized for signal-handling capability, both from an overload standpoint and from a low-noise viewpoint. Stage U1 is chosen for a gain sufficiently high that the input-referred noise will be predominantly due to this stage and the cartridge, but not so high that it will readily clip at high level high frequency inputs. Amplifiers with a ≈10 V rms output capability allow U1 to accept ≈400 mV rms at high frequencies using ±18 V supplies, while still operating with useful gain (about 25 times).

The gain of the two blocks are set by R_4-R_3 and R_6-R_5, as defined by Eq. 6-11. The gain values shown yield a 1 kHz gain that is the product of the U1-U2 stage gains (24.7 times 40.2), times that of the interstage network N (0.101). This yields an overall 40 dB 1 kHz gain. Other gains are realized most simply by changes to R_5 or R_3.

As previously noted, a passively equalized preamplifier such as this must be carefully optimized both from an overload standpoint and from a low noise viewpoint. Stage U1 is chosen for a gain sufficiently high that the input-referred noise will be predominantly due to this stage (and the cartridge, when connected), but not so high that it will readily clip at high level high frequency inputs. To aid this objective, maximum supply voltage and a high output capability amplifier should be used for U1.

Note that U1 operates at relatively high gain, but it needn't be unity gain stable. Decompensated low noise op amps such as the OP37 and the FET input AD745 will provide best signal/noise ratio here. For other FET-input types, the AD845, as well as the OP17 family types, will also yield good performance, but with higher noise levels.

In general, the preceding factors dictate that gain distribution between U1 and U2 be LOW/HIGH from an overload standpoint, but HIGH/LOW from a noise standpoint. Practically, these conflicting requirements can be mitigated by choosing the highest allowable supply voltage for U1, as well as a low noise device. Because of nearly 40 dB loss in the network N at 20 kHz, the output overload of the circuit will be noted at high frequencies first. With the gain distribution shown, the circuit allows a 3 V rms undistorted output to 20 kHz with ±15 V supplies, or more with higher supply voltages.

The equalization network N following U1 should use the lowest impedance values practical from the standpoint of low noise, as the noise output at Pin 2 of the network is equivalent to the input referred noise of A2. The network of Figure 6-17 uses the "N1" RC values of R_1-R_2-C_1-C_2 of Figure 6-9a. As noted, scaling can be applied to either network of Figure 6-9 for component selection, as long as the same ratios are maintained.

Noise in amplifier U2 is less critical than U1 at low frequencies, but is still not negligible. A low voltage noise device is very valuable to the U1 and U2 positions, as is a relatively low input current noise. If extremely low noise performance is sought, such as for a moving coil preamp, the N1 values can be reduced further, and R3 be lowered for lower noise and additional gain. For example, a 45 dB gain preamp could be realized by just dropping R3 to 56.2 Ω, and using an OP37 for U1.

As mentioned before, a low bias current device is appropriate to U1 using bipolar amplifiers. With a 100nA or less bias current device, direct coupling to a moving magnet phono cartridge is practical. For example, the 80 nA (maximum) bias current of the OP37 will induce only an additional 80 µV–160 µV input voltage offset at U1 for a typical 1 kΩ–2 kΩ cartridge resistance. For lower dc resistance MC cartridges, this will be much less of course. Similarly, the bias current induced offset voltage of U2, from the 10 kΩ dc resistance of R_1 will also be low relative to the amplified offset of U1. As a result, the worst-case overall output dc offset using two AD745s can be held to under 2 V for a 40 dB gain, allowing a single C_3 coupling capacitor for dc blocking purposes.

Frequency response of this passively equalized preamp tends to be better than that of the active versions, because of less interaction with the amplifier(s) as compared to the active preamps. It can approach the inherent accuracy of the network components in the audio range, with potentially greater errors at higher frequencies.

Figure 6-18 illustrates this point, in a simulation of the Figure 6-17 circuit using the OP37 models. The midband error is on the order of ±0.02 dB with the N1 network composite values. For practical purposes

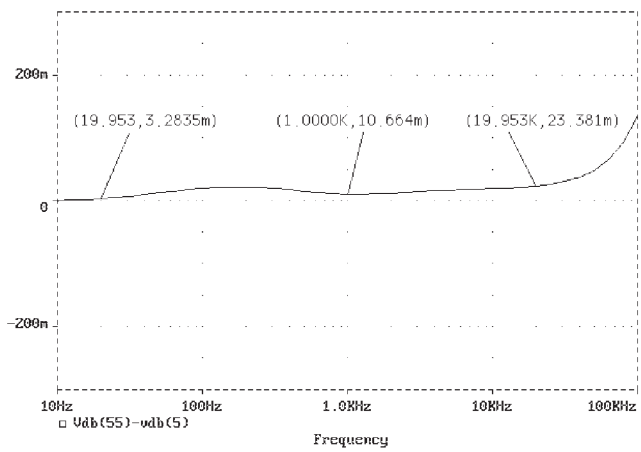


Figure 6-18: Relative error (B) versus frequency for passively equalized RIAA preamp, gain of 40 dB (simulation)

then, the frequency response errors of this circuit will be governed by the tolerances of the network components used within it.

This circuit also can be optionally adapted to servo control of the output offset. This is accomplished by deleting coupling capacitor C_3, substituting a jumper in its place, and using the noninverting servo integrator U3 around stage U2. This is shown as an option within Figure 6-17. A general-purpose noninverting servo can be used for U3, along with a low-offset op amp, such as the AD820, or the OP97.

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