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RIAA Equalization Amplifiers – Part One

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[1] Shown above: Michel Orbe SE from J.A. Michel Engineering, London, England

Introduction

In the June 1979 Journal of the Audio Engineering Society, Stanley Lipshitz, published his now famous 'On RIAA Equalization Networks' paper. Lipshitz, a mathematics professor based at Ontario University who originally hailed from South Africa, applied the full rigor of his profession to develop the transfer equations for 4 popular RIAA equalization networks, exploring both inverting, non-inverting and passive configurations. At the time, many RIAA disc pre-amplifiers fell short of meeting the RIAA equalization curve. Lipshitz was quick to discover, and point out in his paper, that this was primarily due to designers making simplifying assumptions about the networks that were incorrect, leading to errors as measured on high end commercial equipment of 3-4dB across the audio band. In a dense and equation heavy 25 page paper, he also addressed the impact of finite amplifier gain bandwidth, sensitivity analysis, all of this culminating in a set of design procedures for each of the various configurations.

I bought his paper from the AES[2], and wrote a spread sheet to facilitate the design of RIAA networks. I have done a cursory investigation into RIAA spreadsheets published on the web; there are some truly monumental efforts out there and I applaud the authors. My effort is humble by comparison, but it is easy to drive, and delivers the correct results, thanks of course entirely to the Lipshitz equations I've used.

Before beginning the discussion, lets first define the amplifier topologies we are going to talk about:-

1. Active RIAA – implements as a minimum T2-T5 (see Fig2) by means of the feedback network connected between the amplifier output and its inverting input. These types of designs often use a *secondary post filter* (R31[3] and C3 in Fig 1) usually set at 1-2 octaves above the audio band, to fine tune the overall transfer function to achieve accurate RIAA conformance in the upper audio octave – see for example Denis Colin's LP797 RIAA equalizer published in Audio Express here LP797.

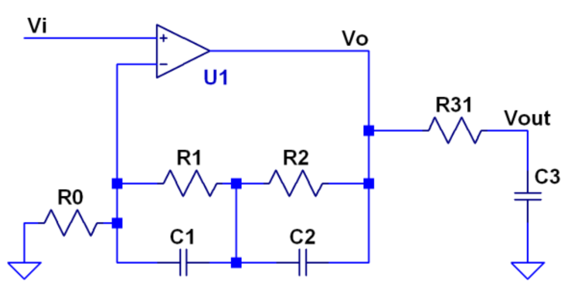

Figure 1 - Active RIAA Amplifer using Lipshitz's 'network 1a'.

2. Active/Passive RIAA – T2, T3 and T4 are implemented in the feedback network, while T5 (the 2122Hz pole) is implemented by means of a *primary post filter*. This configuration has a major impact on HF overload capability – again, details later in the text. T6 is usually not required in active/passive designs, although some do incorporate it for fine tuning. T1 is often implemented between the output and a subsequent buffer stage. Like the active topology, modern active/passive designs may sometimes also

[2] You can obtain a copy from the AES for about $20. The document is copyright, so I am not able to reproduce it or put it up on www.Hifisonix.com unfortunately.

[3] I have purposefully not annotated R31 in Fig 1 as 'R3' so as not to create any confusion with the R3 used in Lipshitz's networks and equations. He placed R3 (which is not used the the designs presented here) between the lower junction of R1 and C1 and R0 and he took feedback off the top end of R0. R0-R2 and C1-C2 are consistent with his annotation.

make use of servo's in order to avoid electrolytic capacitors directly in the signal path and maintain DC accuracy at the output.

2. Passive RIAA. The signal from the pickup is amplified by a fixed, high gain amplifier, which then drives a passive equalization network, after which it is then further amplified.

I am not going to discuss passive equalization on the following pages, and for noise considerations, will confine the thrust of this article to non-inverting topologies and Lipshitz's network 1a as shown in Fig 1 above, ignoring his 1b to 1d networks. The resistor and capacitor assignments R0-R2 and R31 along with C1-C3 will be consistent throughout this document, other than for the active/passive Baxandall design, where I will use his original annotation. We will also assume that the op-amp U1 has sufficient open loop gain such that it will not affect the time constants T1-T6, as would be the case if the loop gain was inadequate. For clarification, R1,R2, C1 and C2 form the main feedback network, with R0 used to determine the stage gain, whilst R31 and C3 form the secondary post filter. T2-T5 are the IEC 98 (1964) RIAA time constants as detailed in the table below:-

Time constant		uSeconds	RIAA Breakpoint Frequency Hz	Comments
T1	Zero	-	Select by Design	Used in the design procedure to set system gain
T2	Pole	7950	20Hz	
T3	Pole	3180	50.05	
T4	Zero	318	500.5	
T5	Pole	75	2122	
T6	Pole	By Design	Usually >100Khz	Per Lipschitz, this is a pole – see text for explanation. Not to be confused with the 'von Neumann' break point mentioned in other texts[4]

The 20Hz breakpoint was added as an amendment and published in 1976 to address disc warp and arm resonances, though its effectiveness in this regard has been questioned by some experts.

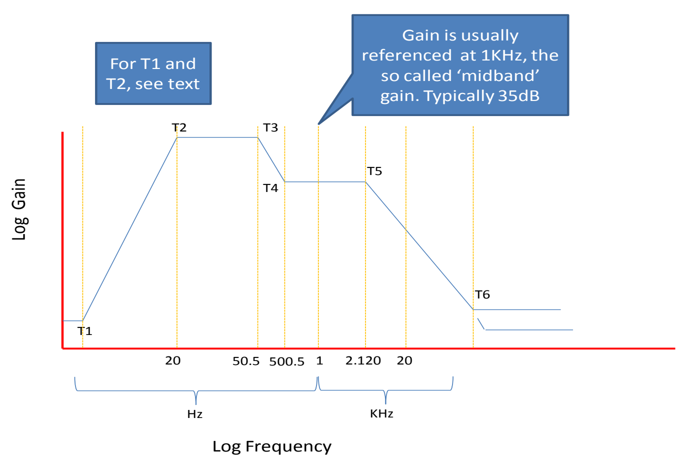

Figure 2 - RIAA Equalization Curve (T1-T6 follow Lipshitz's annotation)

All the time constants interact with each other in RIAA amplifiers where T2-T5 are fully implemented in the feedback network, as is the case in Fig 1. Lipshitz pointed out that you cannot simply calculate the R's and C's from RxCx=Tn because of the way the poles and zero's interact as T1-T6 above interact

[4] According to Self, there is no such thing as a 'von Neumann pole'. See 'Small Signal Audio Design' page 168 for details.

based on the constraint T2.T3.T5=T1.T4.T6. In active/passive designs, the interaction can be minimized or completely obviated, at the expense of overload margin, increased noise or both.

Active/Passive RIAA Equalization

In February of 1981, Baxandall's response was published in the JAES (I paid $20 for copy of that as well), and he took Lipshitz to task about the complexity of his design process, demonstrating a simple, accurate and easily designed network. Lipshitz came at the problem as a mathematician, while Baxandall solved it like an engineer. Lipshitz showed that designing an active RIAA correctly was not a trifling task, while Baxandall demonstrated that through *correct* simplifying assumptions and the use of the active/passive topology, very accurate RIAA was possible without recourse to long equations.

Lets take a look at some of these designs on the following pages starting out with Baxandall's procedure (refer to Fig 3):-

1. Select R6 – Baxandall selected 68k in his original design, which we will see in step 2 below is a very convenient value
2. Calculate C6 from C6R6=3180us – so, for R6 = 68k, this gives a value of 46.76nF – use 47nF standard value
3. Next, take the required midband (i.e. at 1KHz) gain, and calculate the 'Zero Frequency' or 'ZF' gain by multiplying it by 9.898. By way of an example, if the 1KHz gain is 50x, the ZF gain will be 494.9x. The ZF gain is the target gain below the the 50Hz break frequency, but before the 20Hz break frequency.
4. Now calculate R4 from R4 = (10xR6)/(9xZF Gain), which for our example calculates out as 153 Ohms – use 150+3.3 Ohms standard values
5. Calculate R12 from R12 = (R6/9)-R4 which gives 7.4k
6. For the HF pole formed by R3 and C3, select a reasonable value for C3 and calculate R3 from C3R3=75us. If we select 0.1uF for C3, R3 calculates out at 750 Ohms, a standard E24 series resistor.

Due to disc warp and rumble already mentioned, some form of LF attenuation is very desirable. Baxandall indicated a capacitor between R4 and ground, which was used to develop the low frequency T2 pole – usually set at 20Hz. We know from Cyril Bateman's investigation into capacitor distortion published in Wireless world during the 1980's that if an AC voltage is allowed to develop across an electrolytic coupling capacitor, it will cause significant distortion at those frequencies where this occurs. In modern RIAA designs like this, the capacitor is oversized (usually by a factor of 10) to avoid this problem, and a separate capacitively coupled buffer stage used to establish the LF pole using a non-electrolytic capacitor – typical values would be 1uF (use a good quality, tight tolerance poly cap) followed by 15k to ground for example. There will be those that argue that an electrolytic has no place in the signal path of any

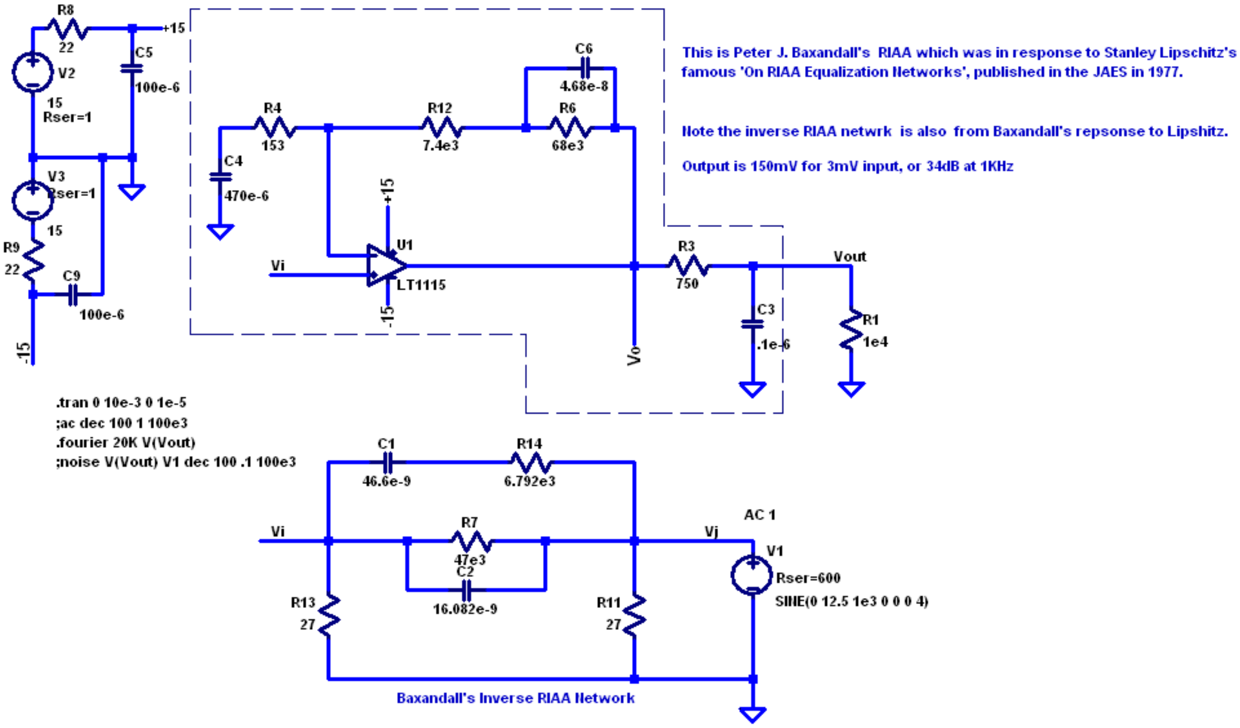

Figure 3 – Baxandall's Active/Passive RIAA Equalizer Amplifier

amplifier dealing with sub mV signals which leaves you with either direct coupling and large offsets to deal with, or, with some sort of servo[5]. Each of these options requires tradeoffs that make a straightforward recommendation impossible. For my part, in the practical implementations of the designs presented here, I will go with a 10x oversized electrolytic, paralleled with some smaller value non-electrolytic caps to deal with electrolytic ESR and ESL at higher frequencies where this may cause issues. Using Baxandall's method, with R3 set to 153.3 Ohms, C3 calculates out at 52.1uF. However, this results in an LF response that is fully 3dB down at 20Hz as shown in Fig 4, a figure some may find unacceptable, to say nothing of the LF distortion this will generate if an electrolytic is going to be used. If C3 is made very large, say 470uF, then the response is flat down to 20Hz as shown in Fig 5. Over the band 2KHz to 20KHz it deviates +0.6dB wrt to the midband value – such is the nature of the interaction between the T1-T6.

Baxandall's simple design utilized the active stage to equalize the 50.05Hz and 500.5Hz breakpoints, leaving the 2122 Hz breakpoint implementation to a primary post filter (R3 and C3 in Fig 3) at the op-amp's output. The penalty for this was a very limited overload margin of around 200mV at 20KHz. If one assumes a nominal 3mV input at 1kHz this will mean about 30mV at 20KHz due to the RIAA inverse

[5] If we assume a nominal 35dB gain at 1KHz, then assuming direct coupling, the gain at DC will be c. 20dB higher than this, so around 55dB. With a worst case 4mV offset on an SA5532 op-amp, this amounts to an output offset of about 2.25V. Too large to ignore.

equalization curve. This translates into a scant 16.5dB overload margin at 20KHz, which is considered low. If you are using a high output MM cartridge – say 5 or 6mV at 1KHz, this translates into about 12dB overload margin. Clicks and pop's from the record surface are likely to cause the op-amp to momentarily clip, causing problems with the sonics due the length of time it takes to recover. 12dB is definitely not enough headroom. The other option here is to then reduce the system gain, gaining perhaps another 6dB of headroom in the RIAA. However, this gain loss has to be made up further down the signal chain, and the cost is a reduction in the noise performance.

Fig 6 shows more clearly the problem with RIAA designs that use 'post' passive equalization like this one – in this case Vo (the output of the op-amp before the filter) – is plotted and it can be seen at 20KHz to be 18dB higher than at 1KHz – this translates directly into loss of headroom at frequencies starting well below 1KHz. It is for this reason that active/passive designs that place the 2122Hz pole at the output of the equalizer are not advised if you are looking for good input overload specifications. I happen to think that high overload margins are important, so for me, this is a key design consideration. Of course, the primary post equalizer filter does have the benefit of reducing the amplifier stage noise through passive attenuation, but this is also accomplished by placing the 2122 pole around the active gain stage where the noise reduction is achieved through reducing gain at HF – you then get low noise and high input overload capability. A further important consideration is the increased distortion of the active/passive configuration arising from increasing high levels of HF content at the output as frequency increases – something active equalization designs neatly avoid.

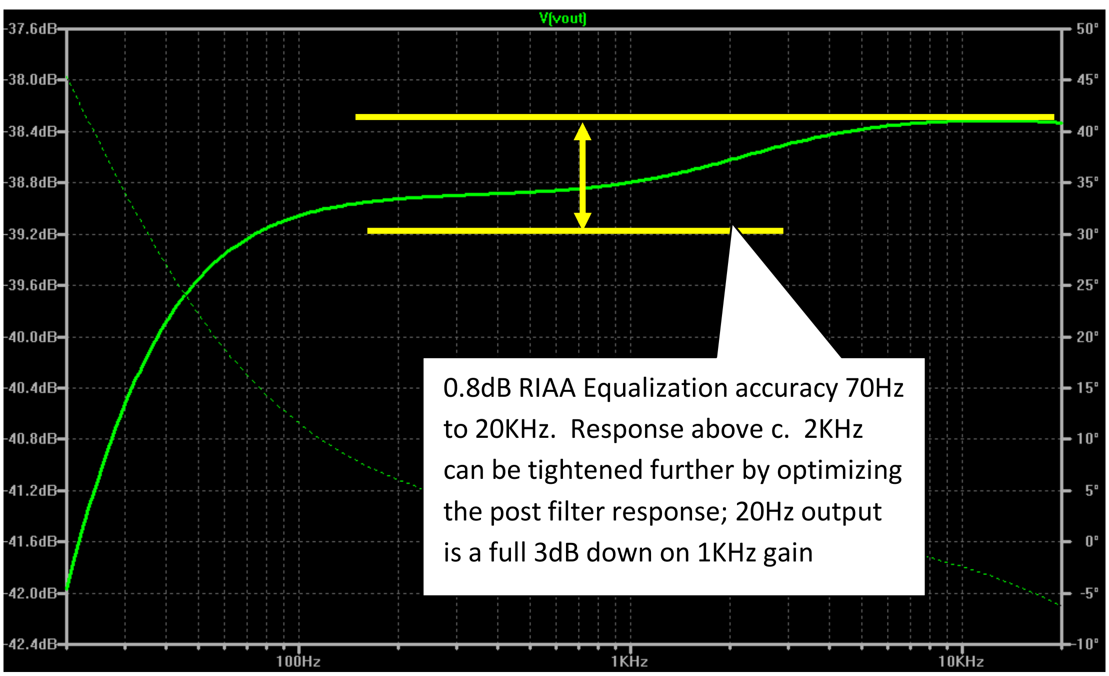

Figure 4 - Baxandall's Active/Passive RIAA Deviation From Ideal Response with T2 implemented per IEC 98 1976 amendment

Although the music signal on the disc may be bandwidth limited during the recording process, surface ticks and pops are not. The headroom loss continues getting worse at frequencies beyond 20KHz, so that once you get to 50KHz or so, another 9db of headroom has gone.

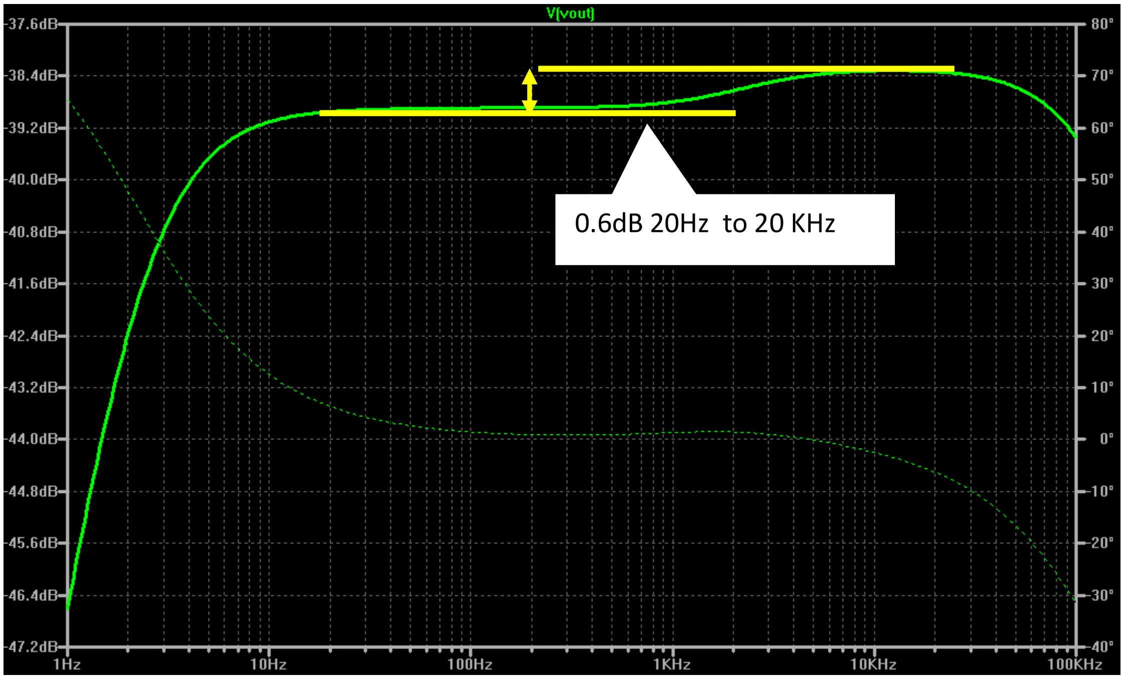

Figure 5 - Baxandall RIAA response with C3 set to 470uF (vs 52.1uF per the original design procedure)

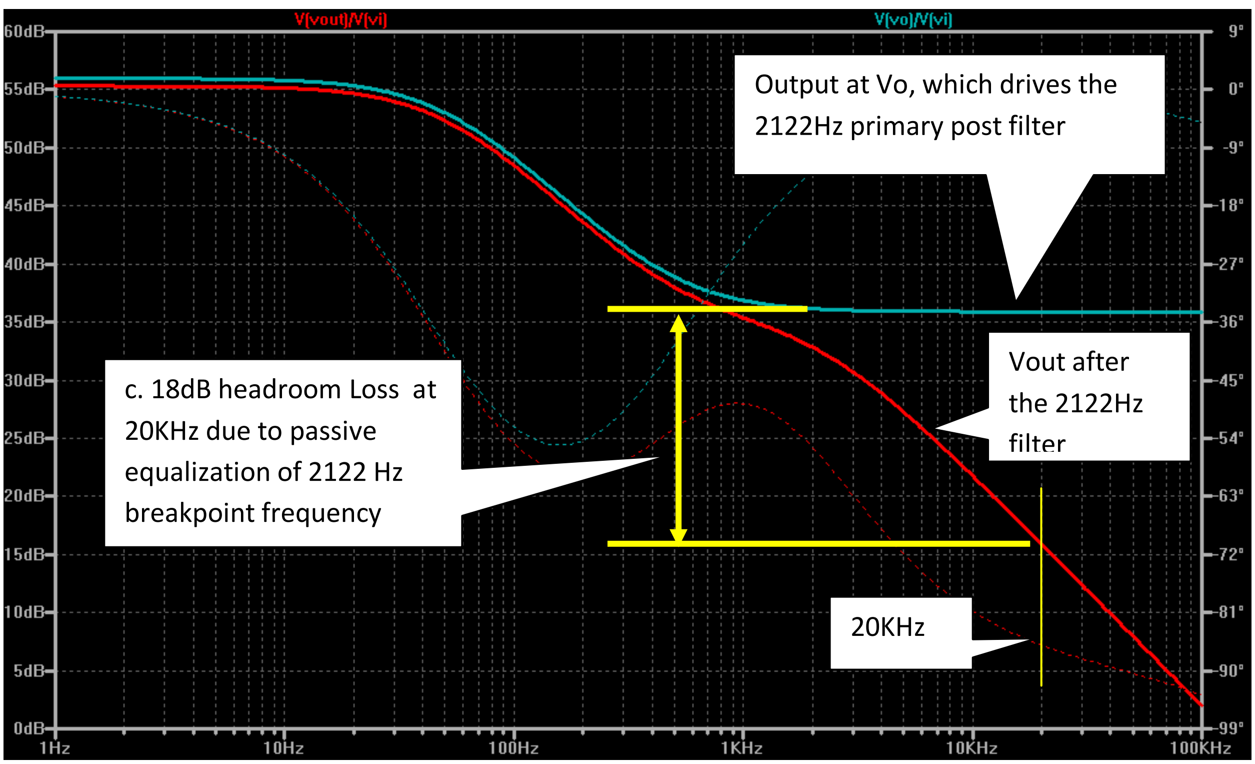

Figure 6 - Overload Margin - Baxandall Active/Passive RIAA Equalizer (DC coupled)

Another option here would be to place the T5 2122Hz pole in the first stage and then do the T2 through T4 breakpoints in the second stage. The penalty for this is noise, since we are placing a high gain stage after an active first stage. Active/passive RIAA pre-amplifiers it must therefore unfortunately be concluded, are suboptimal in terms of overload performance, distortion and noise.

Active RIAA Equalization

Next, we will take a look at non-inverting design that implements the 2122Hz pole in the feedback network, rather than by means of a primary post filter, which effectively deals with the overload margin problem. Clearly, this type of design is a lot more difficult because of the pole zero interactions you have to deal with during the design stage. However, Prof. Lipshitz has done the leg work for us, and his equations have stood the test of time, so we can proceed confidently.

The advantage of the all active approach is that at HF, greater amounts of feedback are applied around the op-amp, reducing the closed loop gain as frequency rises, and usefully, reducing distortion as a side benefit. Assuming 12Vpk output, and a 1KHz gain of 30dB, and an input of 3mV, the overload figure is 190mVpk representing a 36dB overload margin. This overload margin is maintained right across the audio band (reference the expected frequency dependent signal levels at the input) , whereas in the active/passive pre, the overload margin in reduced by 36dB-18dB=18dB at 20KHz.

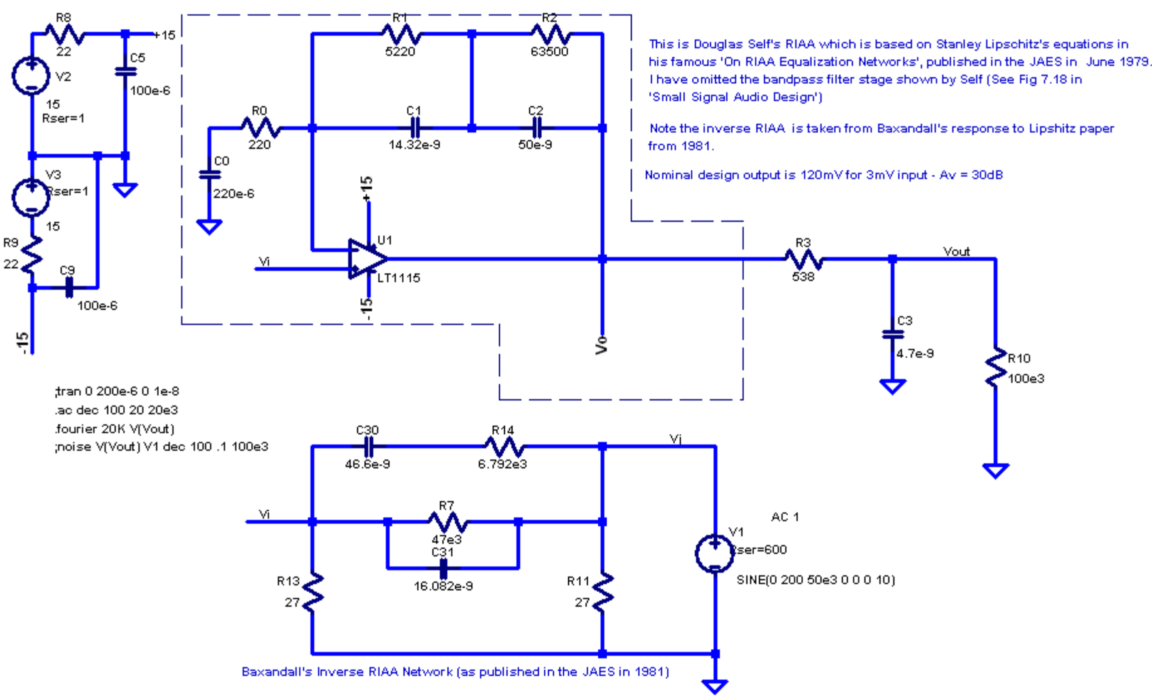

Figure 7 – Active RIAA Amplifier based on Douglas Self's design values

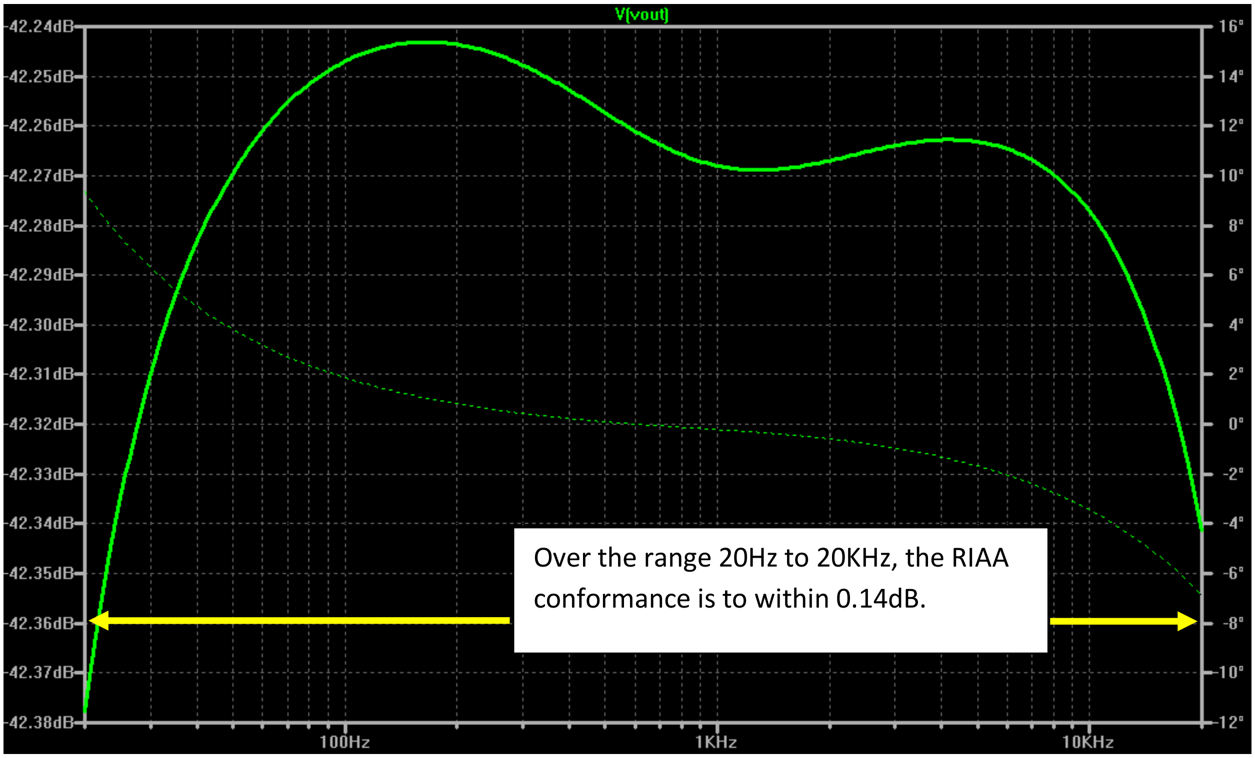

Figure 8 - RIAA Conformance of the Circuit in Fig 7

The T2 through to T5 breakpoints are set by R1, C1, R2, C2 and R0. A secondary post filter is required in order to pull the upper audio band octave into line with the RIAA curve, and is implemented by R31 and C3 in this design.

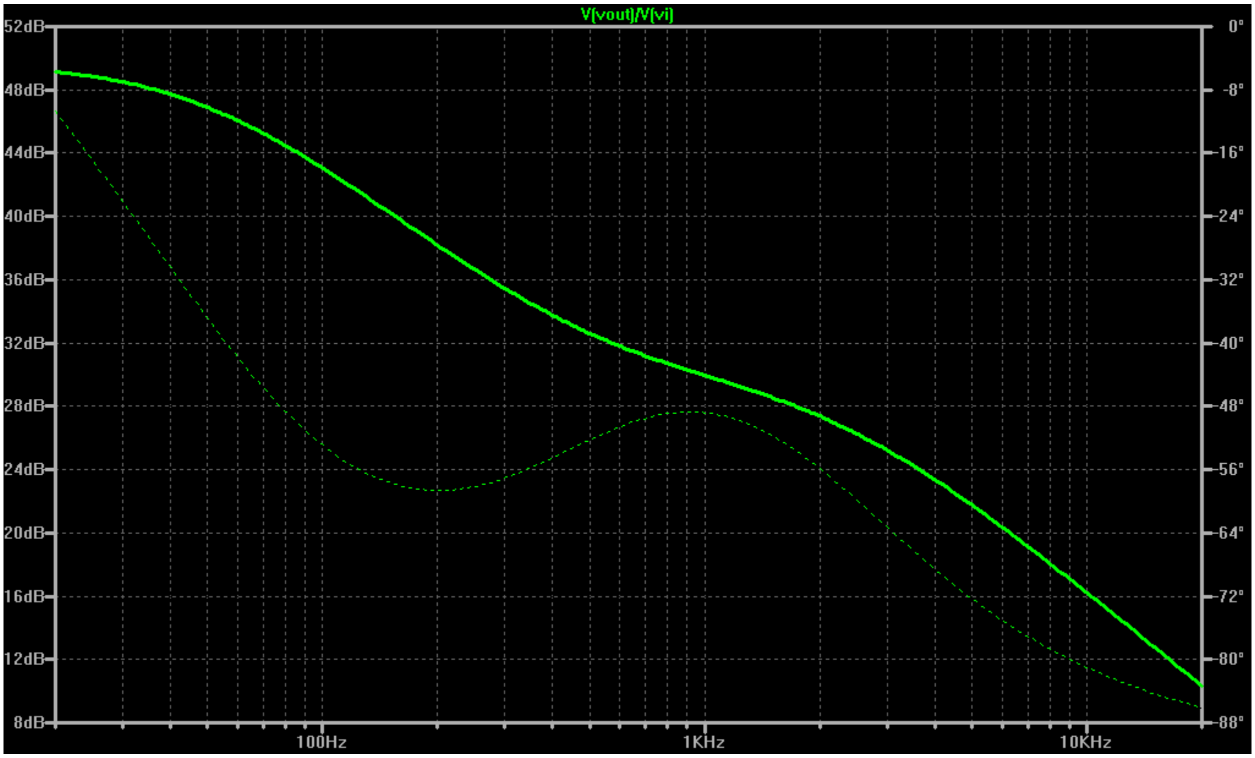

Figure 9 - RIAA Response of The circuit in Figure7 – 20Hz to 20KHz

This secondary post filter is needed because it is not possible to accurately meet the T2-T5 breakpoints using any of the Lipshitz networks using only 2 resistors and 2 capacitors in the main feedback network. The design process therefore targets an accurate fit on T2-T4, with T5 coming in a little higher then actually required at 20KHz – and a little bit higher means 0.5dB or thereabout, so we are talking about a quite small deviation within the audio band. The secondary post filter then applies just a little more HF correction in order to pull the curve back into line such that the overall equalization stage accurately conforms to the standard. Because this frequency pole is usually placed at least 2-3 octaves above the audio band, it has only a very small effect on the HF overload margin – usually no more than around 1 dB at 20KHz, and usually, much less. Nothing to worry about. Fig 9 shows the performance – RIAA accuracy is to within 0.14dB across the audio band. Once the values have been calculated to give an accurate approximation like this, it is possible to tweak the values in a spice model and get even tighter conformance. But, getting RIAA conformance to 0.01dB really isn't a practical idea, since the component tolerances required are unrealistically tight and +-0.2dB is not going to be audible.

RIAA Network Load on Op-Amp Output

A concern of many designers is the low load impedance presented by the equalization network connected between the op-amps output and the inverting input. A quick check using LTSpice reveals that at the onset of clipping at 20KHz, the peak current demanded from the op-amp to drive both the feedback network and the output filter (R3C3) is 20mA, just prior to clipping (c. 12V pk). The NE5534 is quite capable of driving output currents at this level at 20KHz, but at the cost of much increased distortion. If this is acceptable (e.g. in a budget equalizer) you can assume that additional buffers between the op-amp output and the input to the feedback network (Vo) and output filter are not required . You can design for a higher overall network impedance to reduce the load on the op-amp output, but this will come at the cost of noise. Pick your poison.

At higher frequencies, the current demanded by the feedback network increases, so that at 50KHz it is around 25mA peak with the equalization network shown, again just before the onset of clipping. However, you are unlikely to be seeing input signals at this level, and we are now talking about pure specmanship. Of greater concern is the fact that much above 2-3V output at 10 to 20KHz and above, op-amp distortion starts to increase dramatically if called upon to drive heavy loads[6]. There's not much music energy above about 5KHz, but on an LP, there are the inevitable click's and pop's due to surface noise with high energy content that can range into the many 10's of KHz. You don't want these types of signals generating harmonics and causing problems with the sonics lower down in the audio band. For these reasons, high end, cost no object active RIAA designs can really benefit from a decent output buffer stage to drive the feedback network.

Figure 10 is a development of Fig 7 that really does provide the ability to fully drive the feedback network all the way out to 50KHz cleanly whilst significantly reducing the load on the op-amp output stage, and thus improving the HF distortion performance. Although using op-amps, the entire circuit is operated in Class A to avoid any cross over distortion artifacts when driving the feedback network or the output load. In the interests of a small improvement in noise, R0 is lowered from 220 Ohms to 50 Ohms which means the overall feedback network impedance is also lowered, but we can easily drive this because of the output stage buffer.

[6] Current mode amplifiers fair much better in this regard, but there are no ultra low noise current mode amplifiers suitable for RIAA MM amplifiers.

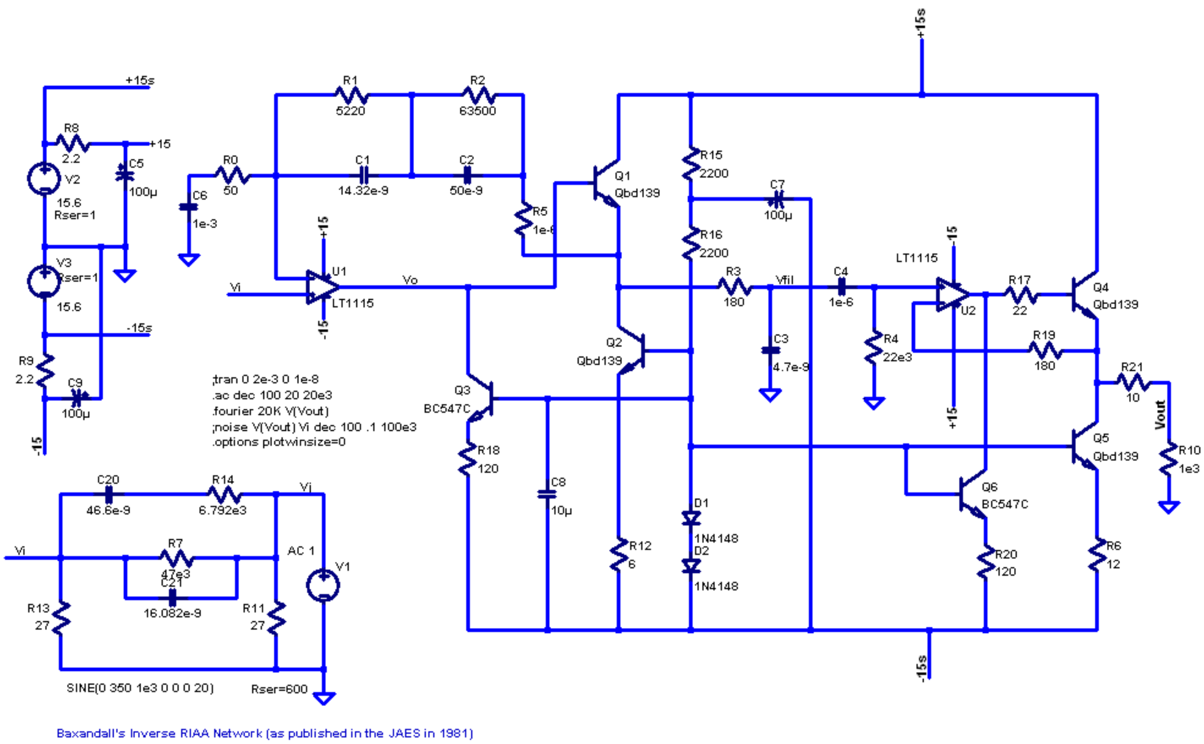

Figure 10 – Complete High End All Class 'A' RIAA Amplifier

Q1 emitter follower buffers the op-amp output, with Q2 providing a 100mA constant current load. A green LED can be substituted for D1 and D2 to provide improved temperature compensation of the constant current sources. Q3 provides a constant current load of around 5mA on the output of U1 and U2 such that they run in class A permanently. The peak current demand required to drive the feedback equalization network is >30mA at 50KHz, which is less than a third of the constant current load provided by Q2, so it is running well in the class A region all the time. The dissipation in Q1 and Q2 amounts to 3.1W on +-15.6V supplies (these are the supply voltages in my preamp), and about 1.5W on Q4 and Q5, so some decent heatsinking is required. The output of the equalizer feeds a class A buffer stage[7]. R0-R3 and C1-C3 provide the RIAA equalization. C4 and R4 provide simple low frequency roll off. You could of course turn U2 into an active rumble/disc warp high pass filter – see Douglas Self's 'Small Signal Audio Design' for an excellent example. However, I have not gone that far in this design.

The gain at 1KHz in this design is 30db. At 12V pk output swing, this translates to 3.8V pk on the input at 20KHz, yielding an input overload of 42dB ref 3mV, and at 1KHz, the corresponding overload is also 42dB; Higher overload margins can only come at the expense of lower gain and increased noise (because we will then have to amplify the signal further along the signal chain), or increased supply voltages. It is very unlikely that signals coming off the cartridge are going to exceed the figures above. Even with high output cartridges, the overload margins are going to be well over 35dB. What about

[7] There is no point in claiming an all class A configuration, only to go and then use a class B op-amp buffer on the output.

signal to noise ratio? With the input short circuited it is -84dB reference 3mV. I think we are good to go on this design.

Distortion on this equalizer with an output voltage of 20V pk-pk at 20KHz is about 17ppm using the LT1115 devices shown[8].

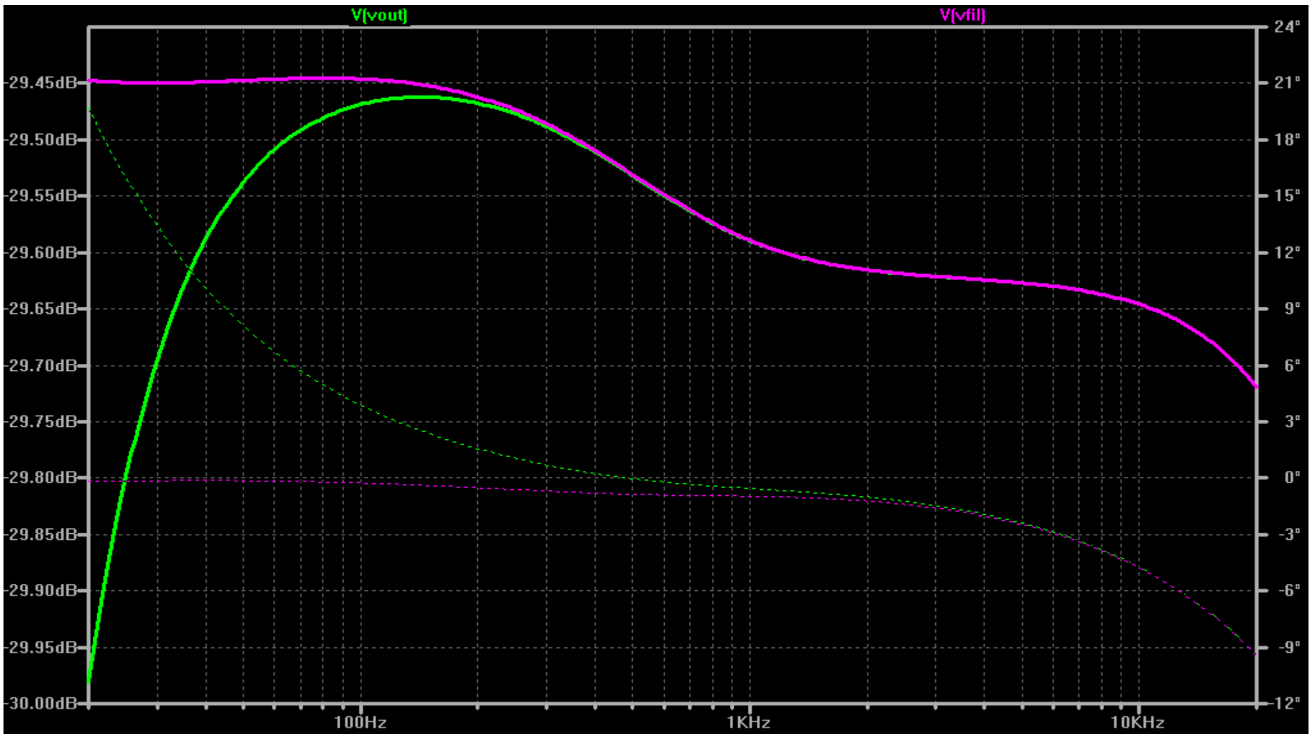

Figure 41- RIAA Deviation for the Circuit in Fig 9 20Hz to 20KHz

[8] I used the LT1115 I this design because the model is readily available in the LTSpice library. However, the NE5534 would also do well here, as would the LM4562.

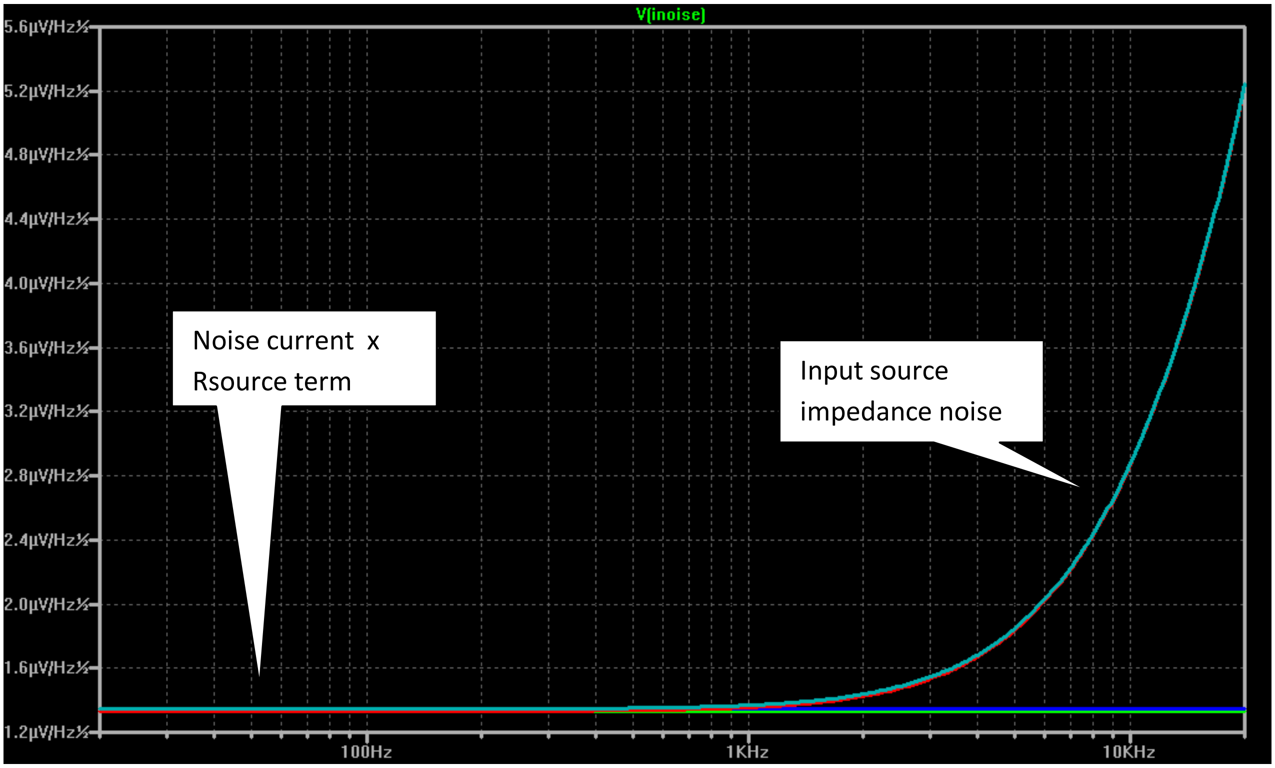

Figure 52 -Noise for the Amplifier in Figure 9

Noise Considerations

No discussion about RIAA equalizer amplifiers can be complete without some thoughts about noise. Books have been written on the subject[9]. LTSpice provides some useful insights if applied with an understanding of the limits of the models and Fig 12 shows a plot of the noise for the design in Fig 11. The dark blue/green bottom trace is the shows the input referred noise (incorporating both voltage and current noise sources) assuming the input to the amplifier is shorted: about 1.3uV and emanating almost entirely from the LT1115. The 50 Ohm R0 resistor and the feedback network contributes only a very small amount to this figure, and given the nominal input of 3mV at 1KHz, can be ignored (not so if we were talking about a MC head amp!). The light blue upper trace shows what happens when you connect a typical cartridge to the input, and here I assumed 600mH inductance and 200 Ohms resistance. The noise increases at 6dB per octave such that at 20KHz, it is approaching 5.6uV and this is due to the amplifier noise current, which when flowing through the source resistance (i.e. DC coil resistance + the inductive impedance at higher frequencies), gives rise to an additional noise term over and above the amplifier intrinsic noise voltage of about 1.3uV. The only way to improve on this noise performance would be to look for a lower noise amplifier solution. For this reason, on MM amplifiers,

[9] See for example Burkhard Vogel's 'The Sound of Silence' for an in depth investigation on the subject.

you often see JFET input stages, which offer much lower current noise. A good example of such a design is Dennis Colins LP797 amplifier published in the Audio Xpress some years back. However, JFET's generally suffer from higher noise voltage, so a careful tradeoff game is required. Currently, insofar as op-amp based designs are concerned, despite being a 30 year old op-amp design, the NE5534A offers the best performance vs cost solution.

Because the of the frequency response shaping RIAA network, and the fact that the input source impedance is largely inductive, simply applying the √bandwidthHz*Vi(noise tot) equation won't give the correct answer. The best approach here is to break the bandwidth up over a number of segments, calculate the noise of the bandwidth, and then root mean square sum them.

Input Loading Networks

Not shown at this stage of the design is any form of input loading network. Cartridge manufacturers often recommend a loading network, and this is done in order to achieve the flattest response from the cartridge, and minimize any ringing. However, in practice, you need to know what your cable capacitance is (typically 50-100pF per meter) so that this can be factored in. In an ideal world, you would feed a square wave into the pickup from a test disc, and then adjust the loading at the output of the pre-amplifier for minimal overshoot and ringing on an oscilloscope – the latter a piece of equipment not everyone has to hand. In most cases, the best way forward here is to use the best instrument you have to hand to adjust the cartridge loading: your ears.

Part Two of this article will detail a practical design based on the principles discussed above, along with a PCB layout, test measurement results and listening tests