RIAA Equalization Amplifiers

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Updated and Expanded

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

!CAUTION!

These designs incorporate mains powered circuitry. If you do not have the skills or knowledge to wire up mains powered circuitry, ask someone who does to undertake that part of the project for you.


OVATION


Accuphase
stereo phono amplifier

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 3-4 dB 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 with 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 eff orts 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.
2. Active/Passive RIAA – T2, T3 and T4 are implemented in the feedback network, while T5 (the 2122 Hz 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 make use of servo's in order to avoid electrolytic capacitors directly in the signal path and maintain DC accuracy at the output.

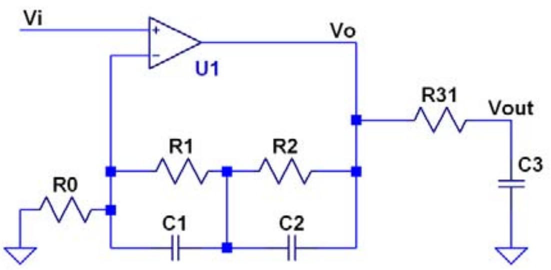


Figure 1 - Active RIAA Amplifier using Lipshitz's 'network 1a'.

[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 http://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 fee dBack off the top end of R0. R0-R2 and C1-C2 are consistent with his annotation.

3. 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 document 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 feed back 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	See the design procedure
T2	Pole	7950	20Hz	
T3	Pole	3180	50.05	
T4	Zero	318	500.5	
T5	Pole	75	2122	
T6	Pole	By Design	Usually >100 kHz	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.

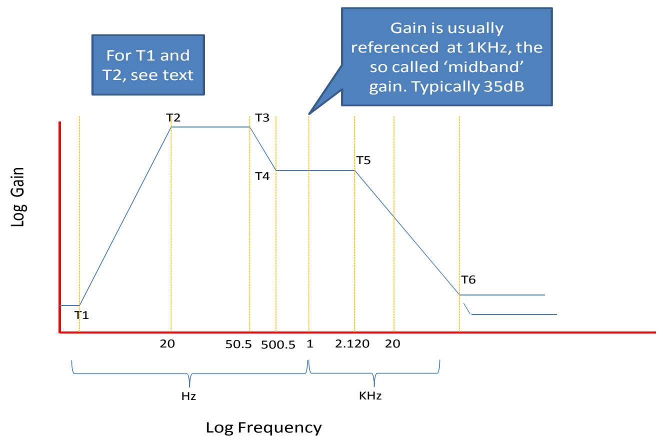


Figure 2 - RIAA Equalization Curve (T1-T6 follow Lipshitz's annotation)

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 based on the constraint T2.T3.T5=T1.T4.T6. Thus, if you start out designing for T4 say, this will affect the locations T2, T3 and T5. Likewise, if you set out to design for T2, it will have an impact on the

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

location of T1, T4 and 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 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 68 k in his original design, which we will see in step 2 below is a very convenient value
2. Calculate C6 from C6R6=3180 us – so, for R6 = 68 k, this gives a value of 46.76 nF – use 47 nF standard value
3. Next, take the required mid band (i.e. at 1 kHz) gain, and calculate the 'Zero Frequency' or 'ZF' gain by multiplying it by 9.898. By way of an example, if the 1 kHz gain is 50x, the ZF gain will be 494.9x. The ZF gain is the target gain below the 50 Hz break frequency.
4. Now calculate R4 from R4 = (10xR6)/(9xZF Gain), which for our example calculates out as 153 – use 150+3.3 Ω or 120 Ω + 33Ω standard values
5. Calculate R12 from R12 = (R6/9)-R4 which gives 7.4 k
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 0.1 uF for C3, R3 calculates out at 750 Ω, a standard E24 series resistor.

Due to disc warp and rumble already mentioned, some form of LF attenuation is very desirable.

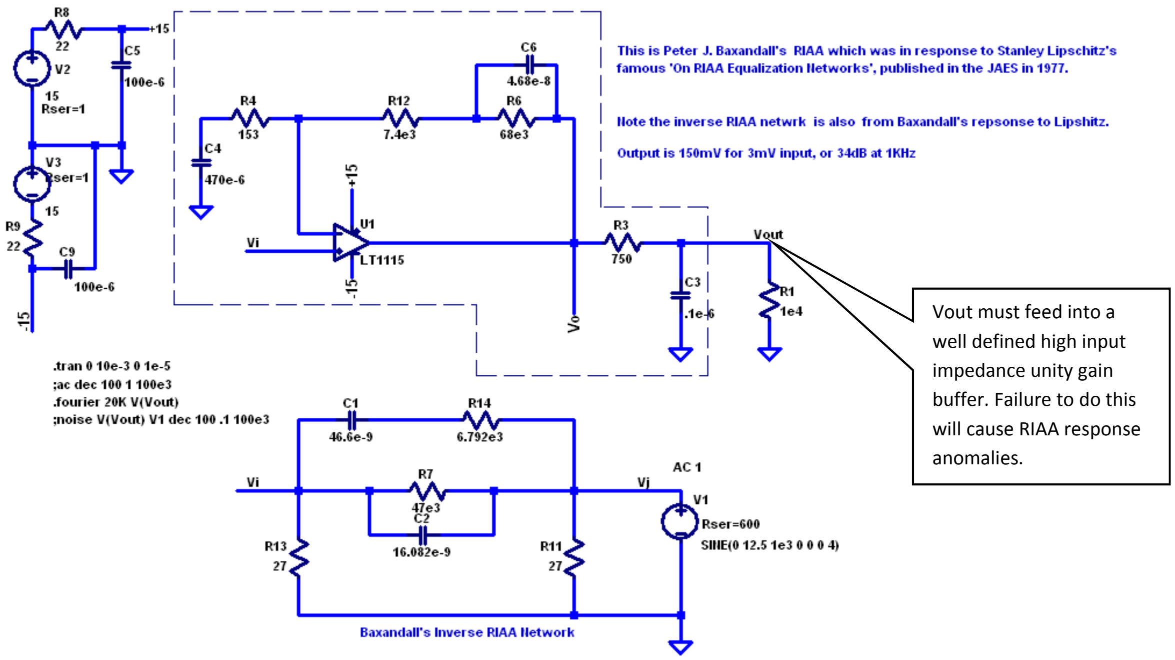


Figure 3 - Baxandall's RIAA Equalizer Solution from his 1981 critique of Lipshitz's design methodology

Baxandall indicated a capacitor between R4 and ground, which was used to develop the low frequency T2 pole – usually set at 20 Hz. 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 1 uF (use a good quality, tight tolerance poly cap) followed by 15 k to ground for example. There will be those that argue that an electrolytic has no place in the signal path of any 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 Ω, C3 calculates out at 52.1 uF. However, this results in an LF response that is fully 3 dB 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 470 uF, then the response

[5] If we assume a nominal 35 dB gain at 20 kHz, then the gain at DC will be at least 20 dB higher than this, do around 55 dB. With a worst case 4mV offset on an SA5532 op-amp, this amounts to an output offset of about 2.25 V. Too large to ignore in my view.

is flat down to 20 Hz as shown in Fig 5. Over the band 2 kHz to 20 kHz it deviates +0.6 dB wrt to the mid band value – such is the nature of the interaction between the T1-T6 breakpoints.

Baxandall's simple design utilized the active stage to equalize the 50.05 Hz and 500.5 Hz 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 20 kHz. If one assumes a nominal 3 mV input at 1 kHz this will mean about 30 mV input at 20 kHz due to the RIAA inverse equalization curve. This translates into a scant 16.5 dB overload margin at 20 kHz, which is considered low. If you are using a high output MM cartridge – say 5 or 6 mV at 1 kHz, this translates into about 12 dB overload margin. 12 dB is definitely not enough headroom. The other option here is to then reduce the system gain, gaining perhaps another 6 dB 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 20 kHz to be 18 dB higher than at 1 kHz – this translates directly into loss of headroom at frequencies starting well below 1 kHz. It is for this reason that active/passive designs that place the 2122 Hz pole at the output of the equalizer are not advised if you are looking for good input overload specifications. I happen to think that overload margins >30 dB 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 Hz 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.

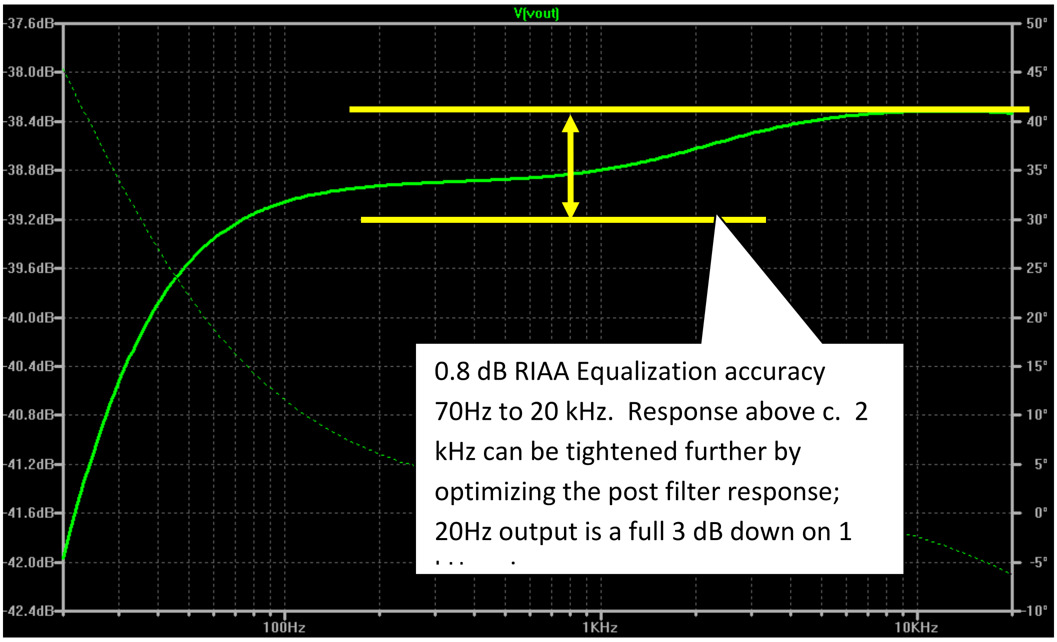


Figure 4 - Baxandall's Active/Passive RIAA Deviation From Ideal Response with T2 implemented per IEC 98 1976 amendment

The headroom loss continues getting worse at frequencies beyond 20 kHz, so that once you get to 50 kHz or so, another 9 dB of headroom has gone. Of course, one would assume that there is very little music energy up at these frequencies. However, as alluded to earlier, surface ticks and pops remain a problem, and they do have a lot of HF energy.

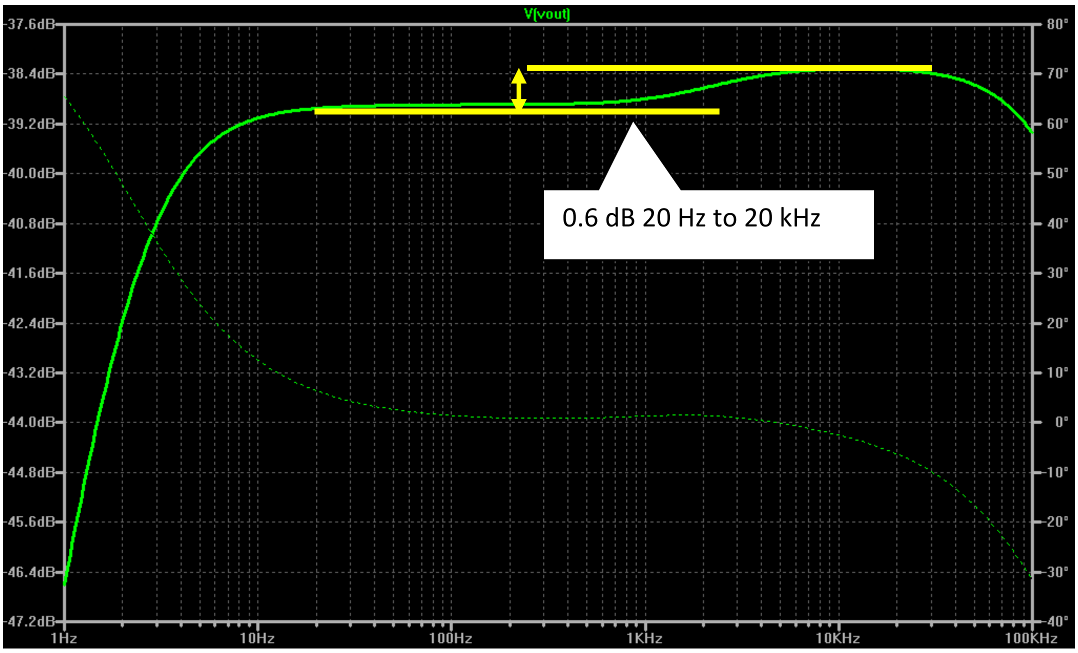


Figure 5 - Baxandall RIAA response with C3 set to 470 uF (vs 52.1 uF per the original design procedure)

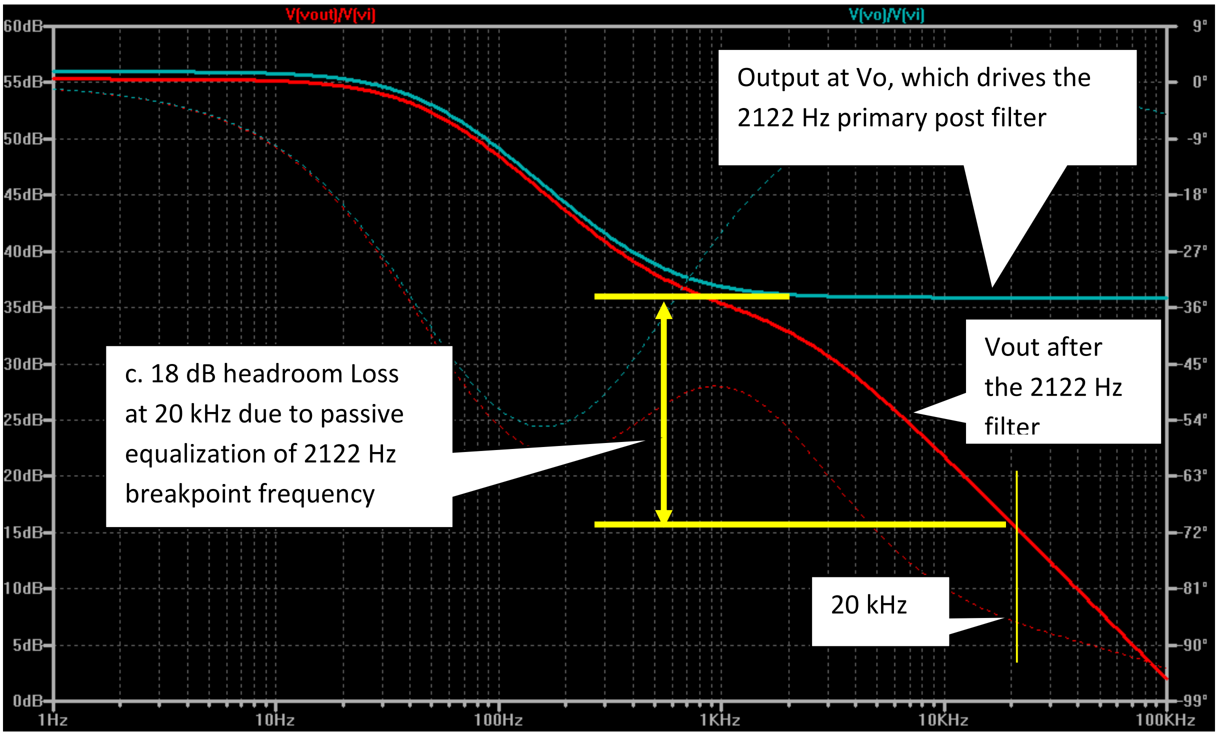


Figure 6 - Overload Margin - Baxandall Active/Passive RIAA Equalizer (DC coupled)

Another option here would be to place the T5 2122 Hz 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. When using opamps, active/passive RIAA pre-amplifiers are sub-optimal in terms of overload performance, distortion and noise.

Active RIAA Equalization

Next, lets take a look at non-inverting design that implements the 2122 Hz pole in the feed back 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 1 kHz gain of 30 dB, and an input of 3 mV, the overload figure is 190 mV peak representing a 36 dB 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 36 dB-18 dB=18 dB at 20 kHz.

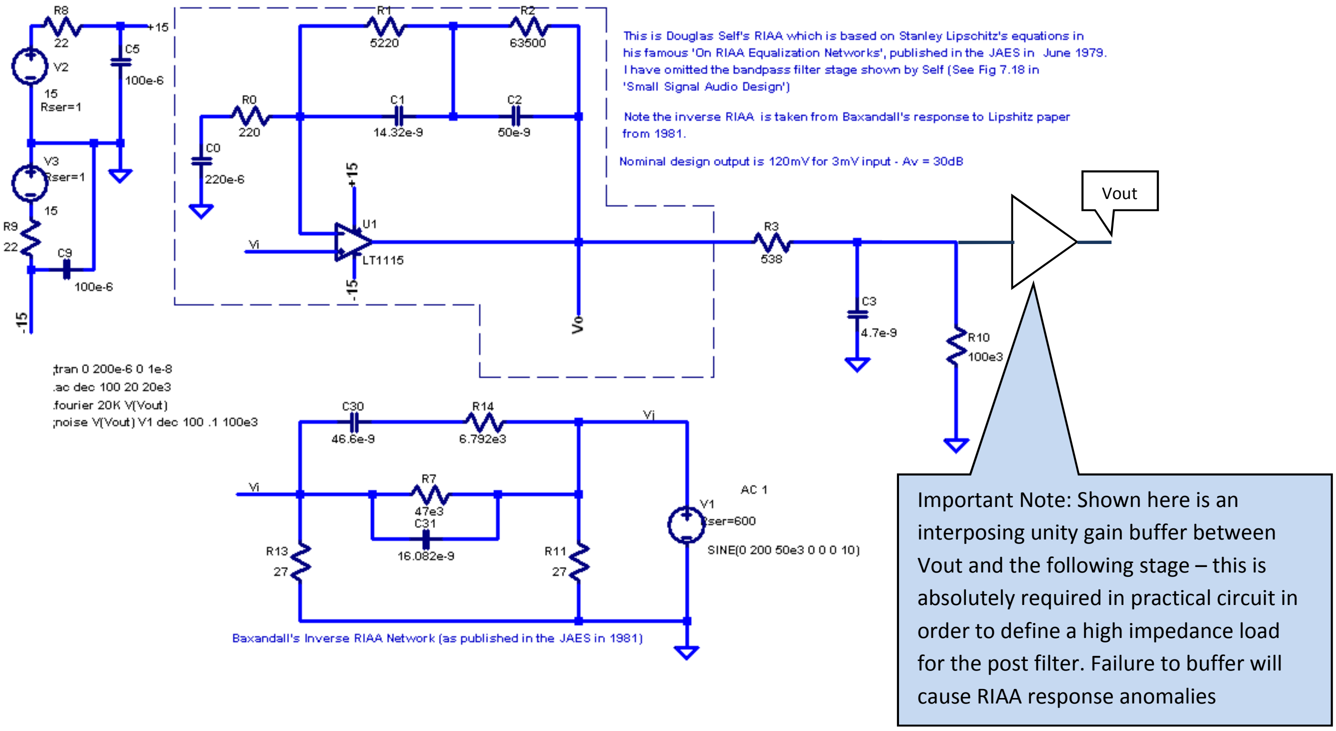


Figure 7 – Active RIAA Amplifier based on Douglas Self's design values

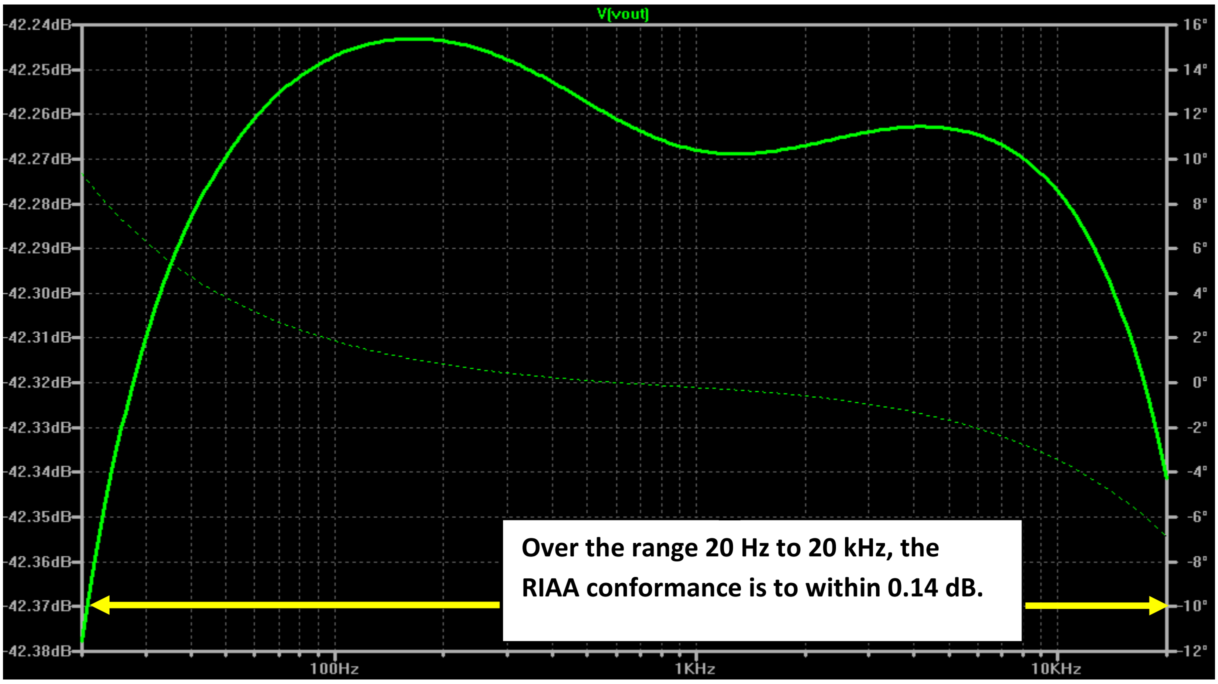


Figure 8 - RIAA Conformance of the Circuit in Fig 7

Table 1 below gives some idea of the overload margins achievable with the design described here assuming 150 mV line stage input sensitivity and +-12 V pk opamp output.

Table 1 - Gain and Overload Margins for 3 and 5 mV Cartridges

Cartridge nominal rated OP at 1 kHz >>>>	5 mV	3 mV
System Gain dB	30 dB	35 dB
Magnitude	31.6x	56x
Output for 3mV input	0.0948 V	0.168 V
Output for 5mV input	.158 V	0.28 V
Overload margin 3mV input	42 dB	37 dB
Overload margin 5mV input	37.6 dB	32.6 dB

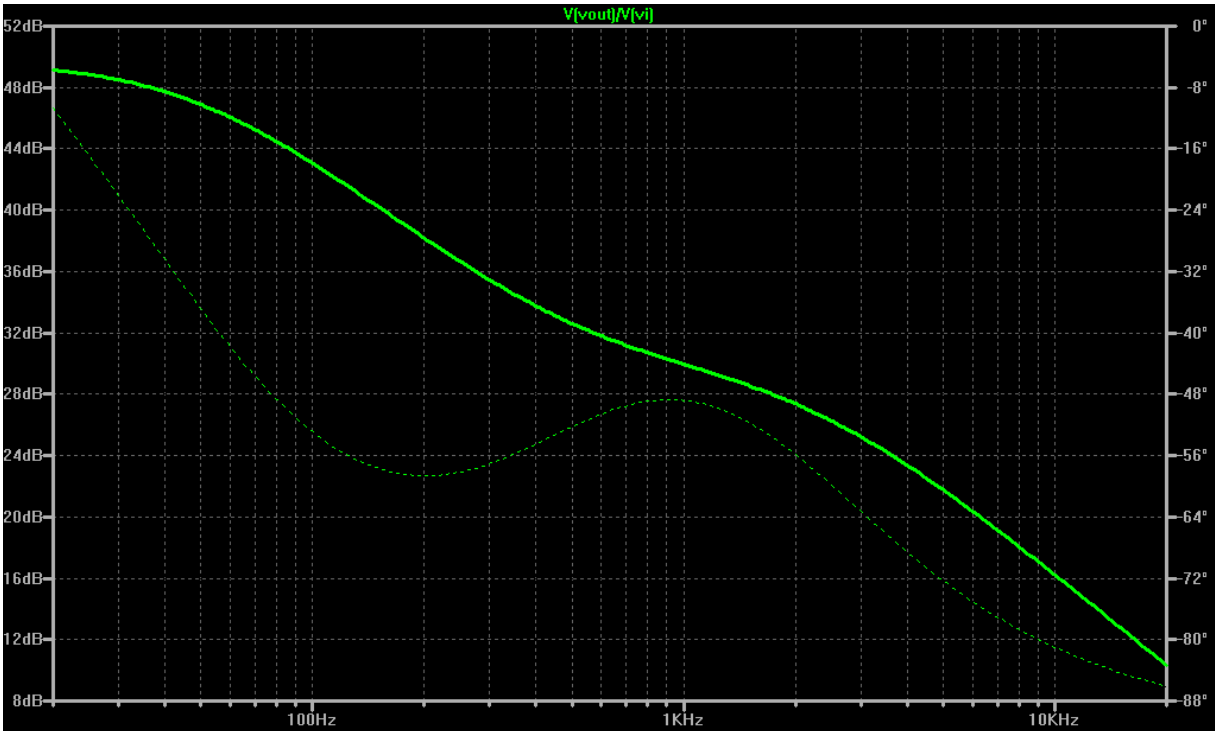


Figure 9 - RIAA Response of the circuit in Figure7 – 20 Hz to 20 kHz

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. 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 fee dBack network. The design process therefore targets an accurate fit on T2-T4, with T5 coming in a little higher then actually required at 20 kHz – and a little bit higher means 0.5 dB 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 20 kHz, and usually, much less. Nothing to worry about. Fig 9 shows the performance – RIAA accuracy is to within 0.14 dB across the audio band.

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 20 kHz, the peak current demanded from the op-amp to drive both the feedback network and the output filter (R3C3) is 20mA and c. 12V pk. The NE5534 is quite capable of driving output currents at this level at 20 kHz, 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) 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 50 kHz it is around 25 mA 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 2-3 V output at 20 kHz 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 5 kHz, 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 50 kHz 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 Ω to 50 Ω which means the overall feedback network impedance is also lowered, but we can easily drive this because of the output stage buffer.

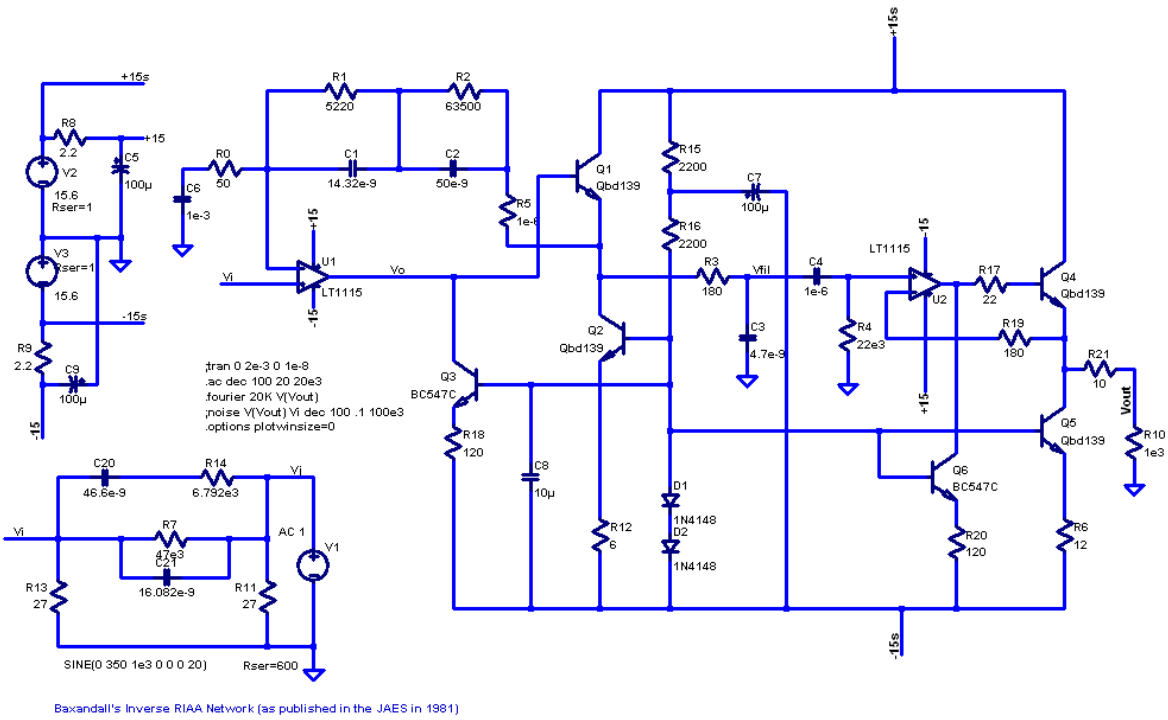


Figure 10 – Complete All Class 'A' RIAA opamp based equalizer amplifier

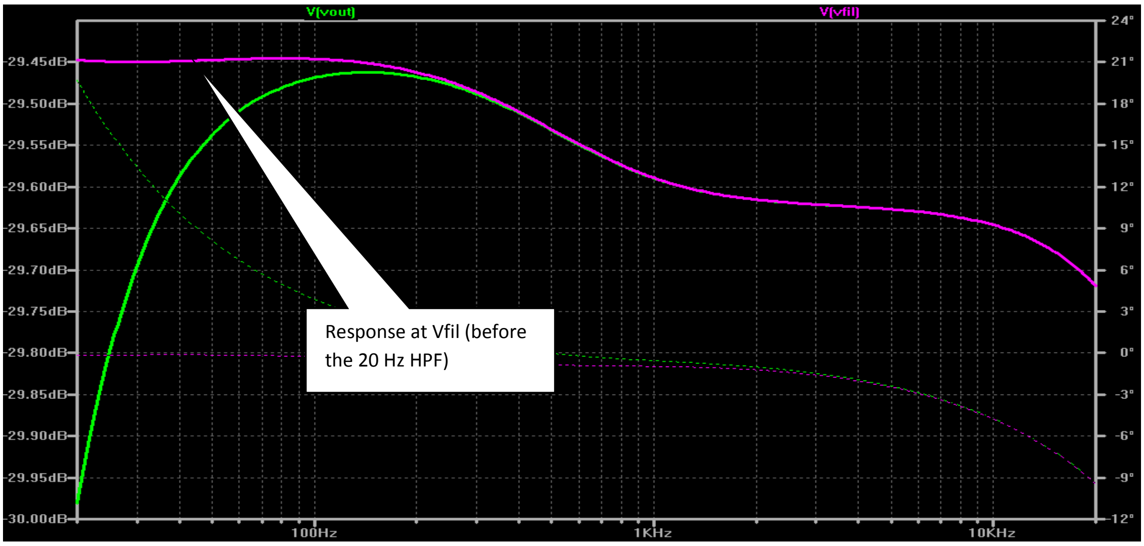
[6] Current mode amplifiers fair much better in this regard, but there are no ultra low noise current mode amplifiers suitable for RIAA MM amplifiers.

Q1 emitter follower buffers the op-amp output, with Q2 providing a 100 mA 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 5 mA 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 >30 mA at 50 kHz, 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 3.1 W on +-15.6 V supplies (these are the supply vol~~tages in my preamp)~~, and about 1.5 W 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.

The gain at 1 kHz in this design is 30 dB, and at 20 kHz, the gain is 10 dB. At 12 V pk output swing, this translates to 3.8V pk on the input at 20 kHz, yielding an input overload of 42 dB ref 3 mV, whilst at 1 kHz, the corresponding overload is also 42 dB; 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 35 dB. What about signal to noise ratio? With the input short circuited it is -84 dB reference 3mV. I think we are good to go on this design.

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

Figure 11- RIAA Deviation for the Circuit in Fig 9 20Hz to 20 kHz



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

[8] I used the LT1115 I this design because the model is readily available in the LTSpice library. However, the NE5534A would also do well here (although distortion is likely to be in the region of 30-40 ppm), as would the LM4562.

Discrete RIAA Equalizers

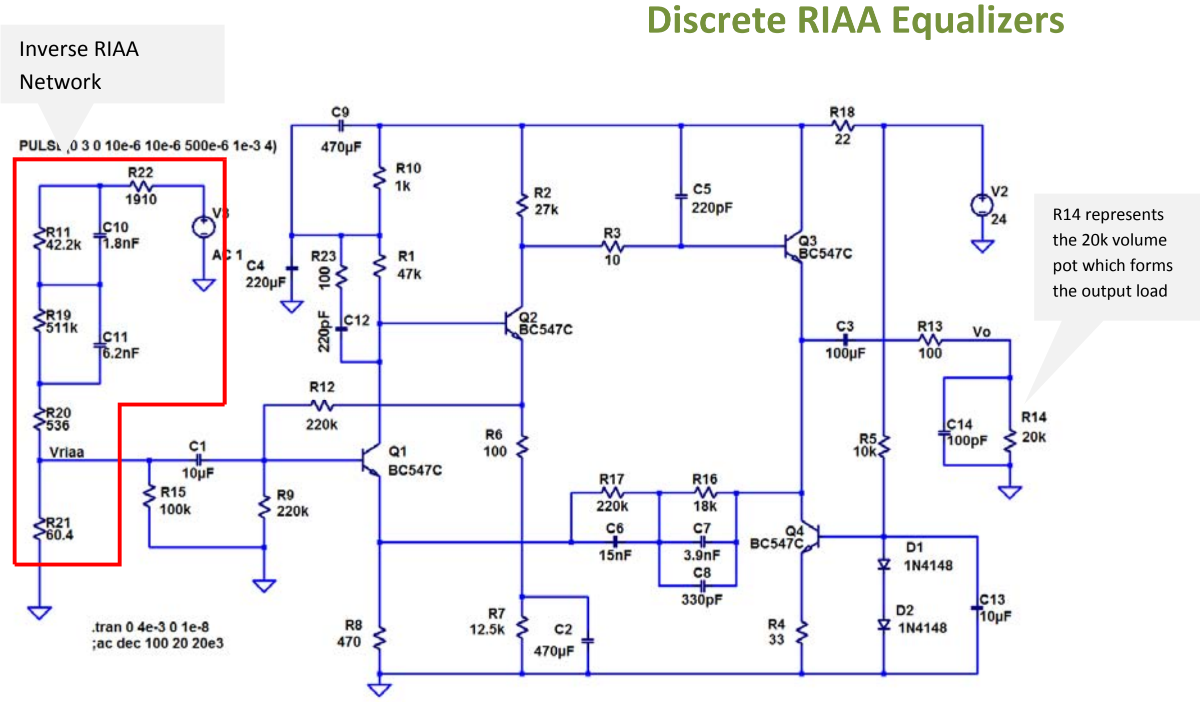


Figure 4 - 4 Transistor discrete 'Bonsai' RIAA equalizer dating from 1979

The RIAA stage pictured in Fig. 4 was designed in 1979 and based on a simple all NPN BC107 4 transistor circuit (I've used BC547's here because models for the original BC107's are not available – however it will work perfectly as shown). Q1 and Q2 are configured as a simple two transistor gain stage, which is then buffered by Q3 emitter follower loaded with a constant current source Q4. There are two feedback paths in this design – a DC control loop running from the emitter of Q2 via R12 which sets the operating point of Q1 such that its collector voltage is about 13 V, and then the RIAA shaping feedback network from the emitter of Q3. In the original design, R7 consisted of 10k in series with a 5k ten turn pot which was used to adjust the voltage at Vout to 12V. R18, C9 and R10 and C4 ensure that any rail noise is shunted to ground, helping to preserve the reasonable noise performance of this simple circuit. This circuit will work with almost any small signal transistors, and modern BC560 low noise types are good candidates. C5 and C12 along with R3 and R23 provide HF compensation to ensure that the amplifier is stable. Both the input and output are capacitively coupled due to the single rail design.

The whole equalizer is powered off a 24V supply (originally an LM7824 regulator, but a well decoupled LM317 regulator would be adequate in a modern spin of this design), drawing about 25mA.

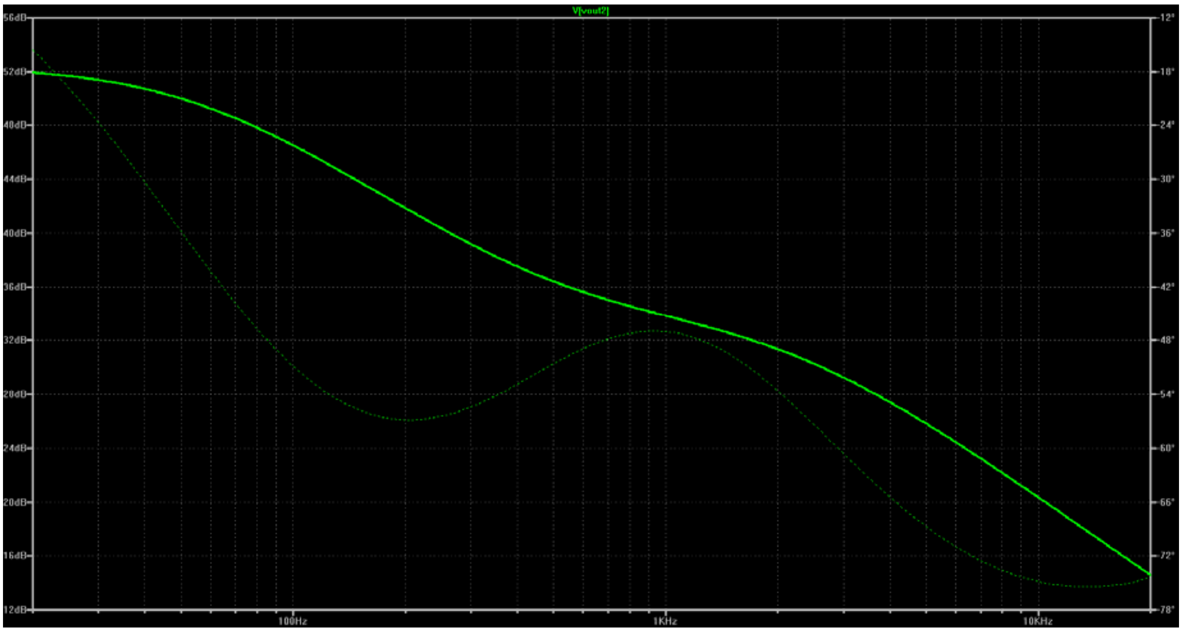


Figure 5 – 4 Transistor RIAA response

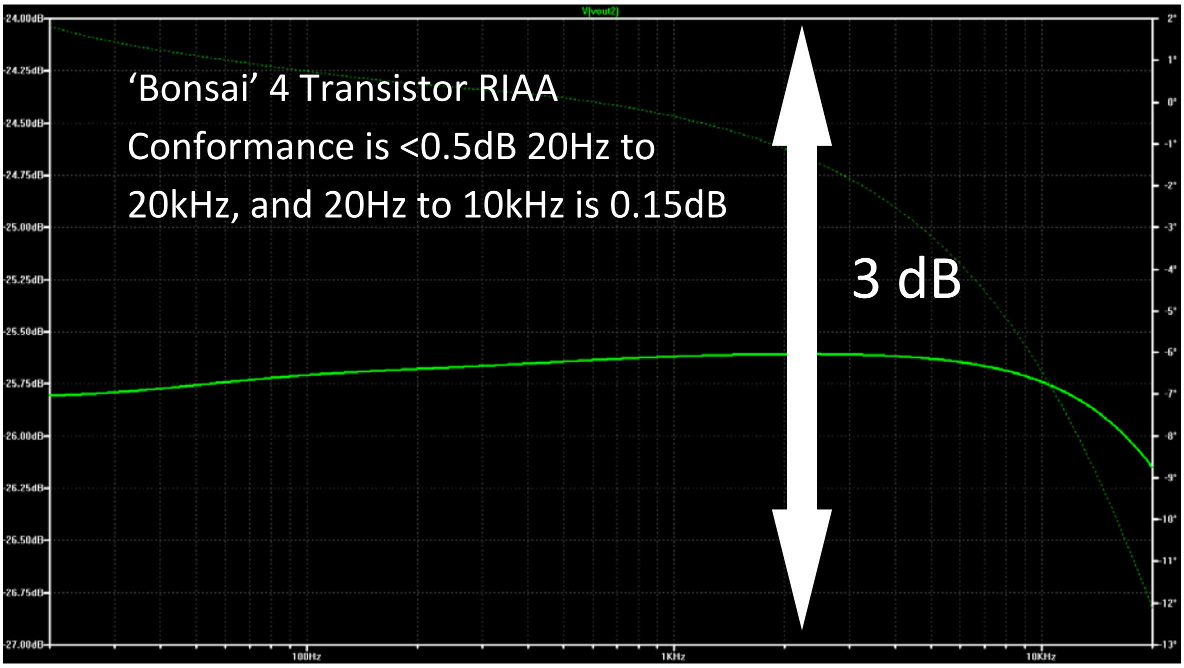


Figure 6 – 4 Transistor RIAA Conformance

Spice was not generally available in 1979 (the PC was still 3 or 4 years away), and the design was simply optimized by injecting a signal on the input and plotting the output response as read on the scope display – as you can imagine, the margin for error was significant – however in the circuit above, the RIAA response was optimized recently using LTspice and is shown in Fig. 5 and the conformance in Fig. 6.

The overload capability on this design is very good at > 30 dB at 1 kHz ref 5mV input, and reflects the advantages of the active feedback approach in low supply voltage designs, of which this one is an example[9]. Very few of these simple discrete types of RIAA were exceptional from the perspective of noise (or distortion) ref a 5mV input, with s/n typically ranging from about 65 dB through to 75 dB or so on the very good designs. This should be contrasted with some of today's better designs, using discrete JFET input stages or opamps, that can achieve 80 dB ref 3mV, with some really exceptional designs even topping out at a few dB higher than this. With hindsight, the loop gain on my design at low frequencies could have been improved somewhat by increasing the collector load in the second stage, and taking a further hit on the noise of course. The input impedance of this design is about 50 k Ω, but this drops at HF because the reflected emitter impedance drops at HF as a result of the RIAA feedback network so that at 20 kHz it is in the region of ~10 k Ω, and one of the reasons a transistor with high hFE is preferred for Q1 - in practice, this reduction in input impedance does not really affect the sonic performance. Incidentally, this is an issue on all discrete designs with similar feedback arrangements to the input device emitter node.

[9] In tube designs, passive equalization makes a lot of sense as the higher supply voltages. However, tube designs are in general much more noisier than good solid state designs.

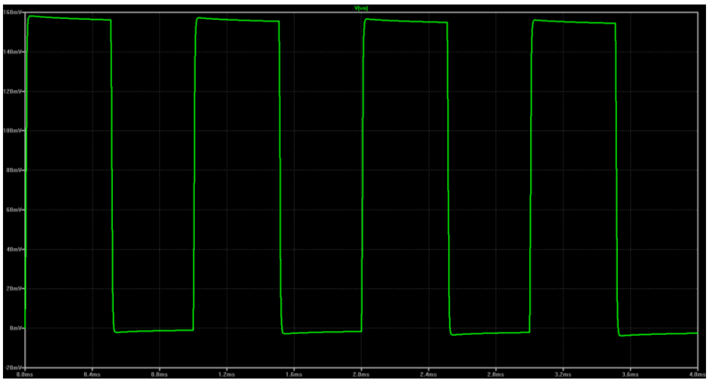


Figure 7 - Square wave Performance if the 4 Transistor RIAA Equalizer

Fig 7. Shows the squarewave performance. This is done by injecting a 1 kHz signal into the inverse RIAA network and plotting the output. The inverse RIAA differentiates the source signal, so the input to the equalizer is a series of 'spikes' – each one corresponding to the edges of the square wave. If the RIAA is accurate, the output should be a square wave, and indeed this is the case, confirming that the conformance is good.

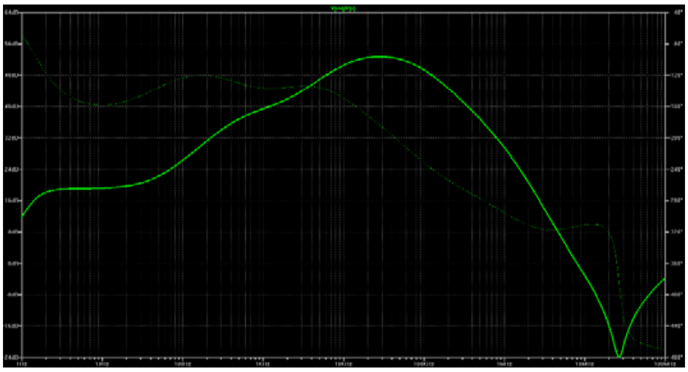


Figure 8 - Loop Gain - 4 Transistor RIAA equalizer

Figure 8 plots the loop gain of the equalizer. At LF, where there is significant boost required, the loop gain is rather low as would be expected – at 100 Hz is it about 25 dB, whilst in the critical mid-band at 1 kHz it's a respectable 40 dB. These figures are much lower than can be expected in an opamp based design, but nevertheless indicate quite good performance for a simple, discrete design.

With rbb' – the main contributor to voltage noise in bipolars – of about 40 Ω for the devices shown, the equivalent input noise is in the region of 3 nV √Hz, or 0.5uV of wideband input noise at 1 kHz. With a cartridge input of 200 Ω + 200mH (typical figures), the en is 3.55 nV √Hz, rising rapidly above 1 kHz to reach 20nV √Hz at 20 kHz. This increase in noise is dominated by the input current noise term – an area where JFET's are better by as much as 9 dB. However, as already mentioned, this is a simple RIAA equalizer and is quite representative of the type of performance one could expect in 1970's and 1980's commercial equipment.

If building this design, effort in sourcing very low rbb' NOS devices like the Rohm 2SD786 devices, which have an rbb' of 4 Ω will pay dividends in terms of noise performance.

An Ultra-Low Noise Discrete MM/MC Equalizer

If you recall, the challenge with MM cartridges lies with their rapid increase in source impedance with frequency due to the pickup coil inductance, which ranges from about 250mH to greater than 600mH, while the specific cartridge DC resistance ranges from 100 to 300 Ohms. As we saw in the previous section, this source resistance plus inductive component completely swamps an otherwise respectable 1nV/√Hz amplifier noise specification by the time we get to a few kHz and may be as much as 20 nV/√Hz at 20 kHz. Its clear then that if the input source impedance is high (i.e. above a few hundred Ω), or is rising with frequency due to the inductive component, we need to keep i_{en} as low as possible. Examining the two device types best suited to low noise amplification, bipolar devices and JFET's, we see that bipolars feature low input noise voltage e_{ein} and high input noise current, while the opposite generally applies to JFET's – the noise voltage is higher, but the input noise current i_{ein} is much lower. For this reason, bipolars tend to be applied in the front end of MC amplifiers, where the source impedance is 10-50 Ohms, while on MM inputs, top of the range designs exploit the low i_{en} of JFET's.

Following the publication of Denis Colin's LP797 design in AudioXpress, there was a resurgence in single ended designs, driven almost entirely by a desire to get the very best noise performance possible. It is well known that if you parallel devices, the noise, which is random in nature, reduces by √2 – you thus gain a 3 dB reduction by paralleling 2 devices, and a 6 dB reduction with 4 devices in parallel and so forth. Douglas Self published a single ended design in 'Small Signal Audio Design' exploiting this technique using bipolar transistors. Ovidiu Popa, a designer from Canada who at one stage was very active on the DIYaudio.com forum, published a number of high performance designs, one of them paralleling BF862 JFET devices from NXP to achieve low noise.

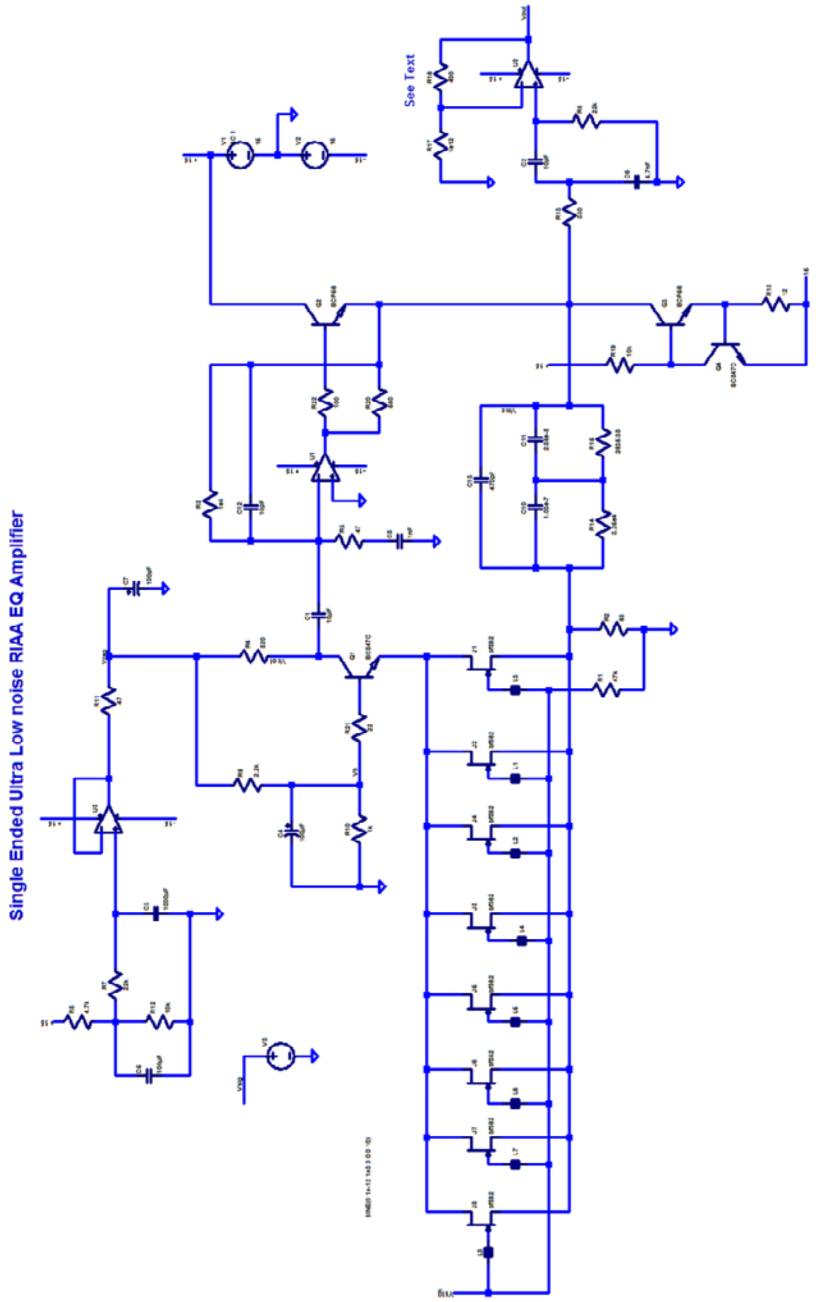


Figure 9 - Discrete Ultra Low Noise MM/MC RIAA

Fig. Fig. 9 shows a design that borrows in part from the work of the aforementioned practitioners. 8 low noise BF862 JFETS are paralleled to realize a wideband input noise voltage of well below 1 nV, which for a 3mV input signal, translates into a 79 dB S/N ratio, and with a 5 mV input, 84 dB – these numbers are state of the art (2014) and can only be bettered slightly by using LSK389/489 devices in place of the BF862 – however, that would be an expensive proposition and poses technical challenges wrt input capacitance and potential oscillation. The Q1 cascode reduces the gate-drain capacitance modulation, and hence distortion of the front end stage and extends the open loop frequency response. Each of the JFET gates is damped with an SMD ferrite bead to prevent HF parasitic oscillation – a problem noted by Popa, and discussed on DIYaudio forum. The front end JFET's are cascoded via Q1 into the drain load R4, which is set at 820 Ohms. C2, a 10 uF film capacitor, couples the front end stage to the opamp U1 (AD797) , where most of the gain is provided. R3 provides bias for the opamp input, while C12, R5 and C8 provide overall loop stability. Without the noise gain compensation network R5 and C8, the loop will oscillate at a few MHz. I put quite some effort into trying to remove these components by using alternative schemes, but it seems that Denis Colin also must have come to the conclusion that this was the best way to solve the problem – so I have also adopted it.

Importantly, unlike competing designs, a single stage active feedback arrangement is used, which for low supply rail designs of +-15 V like this one, offers superior overload capability. I used my Excel RIAA design tool to determine the RIAA frequency shaping components C10, C11, C13 and R14 and R15, with R2 at 68 Ohms setting the overall amplifier gain to 50x at 1 kHz. R13 and C6 form a post filter, and help to keep the overall response conformity for this design to within ~0.1 dB 20 Hz to 20 kHz. Since there are likely to be offsets of up to 1 V at the emitter of Q2 due to the high value of R3, C2 and R6 are required to provide DC blocking and also some attenuation of subsonic frequencies. R6 in the final design was reduced to 2.2 k Ω to reduce the LF energy below 10 Hz, and this also helped to improve the conformity to better than 0.1 dB by removing a rather annoying response hump of c. 0.1 dB below about 100 Hz.

Competing designs DC couple the front end stage to U1, and use a servo to zero the output. I have elected here to go for the simpler approach and capacitive couple the two stages. This has the advantage of also being quieter – and especially so at LF where a servo will inject noise. Since this capacitor is inside the overall loop, the AC voltage appearing across it is *minute*, the distortion is thus negligible – nevertheless, a good quality film capacitor is recommended.

U3 (half LM4562) forms a low noise regulator that leverages the opamp's PSRR - typically 120 dB at LF, and a still very respectable 100dB at 20 kHz. This phono amp is designed to be powered off the Symphony preamp power supply, which has a wide band noise output of under 5 uV – so, the noise injected into the front end JFET stage via R4 is a few tens of pico volts – i.e. negligible and a few orders of magnitude below the intrinsic noise floor of the front end itself – See Fig. 10 for the overall PSRR performance of this RIAA EQ amplifier.

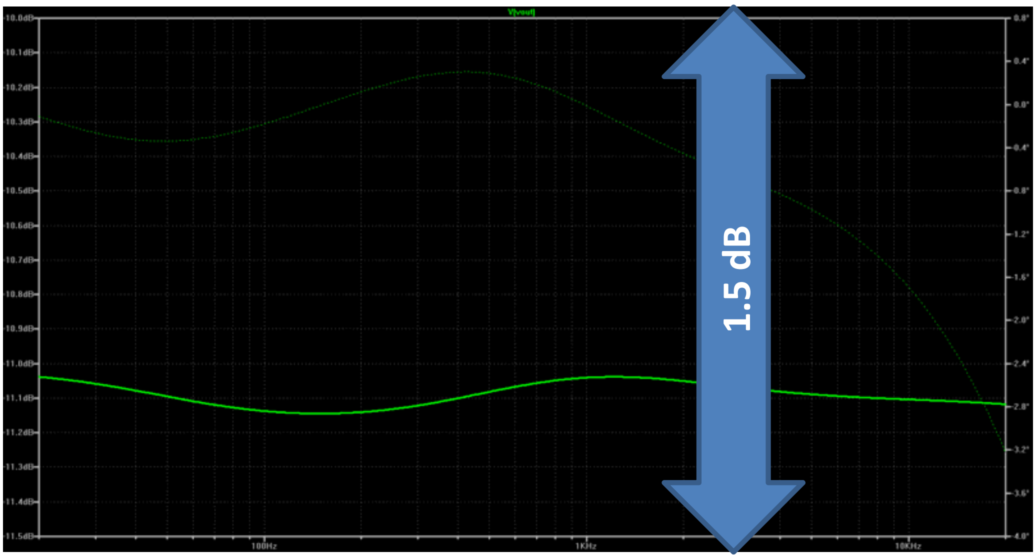


Figure 10 -Discrete JFET RIAA Conformity for the design in Fig.9 is ~0.1 dB

The second half of the LM4562 is used for U2, which buffers the post filter (R13 and C6) and DC blocking via C2 and R6, and to boost the equalizer output by +14 dB for MC cartridge inputs.

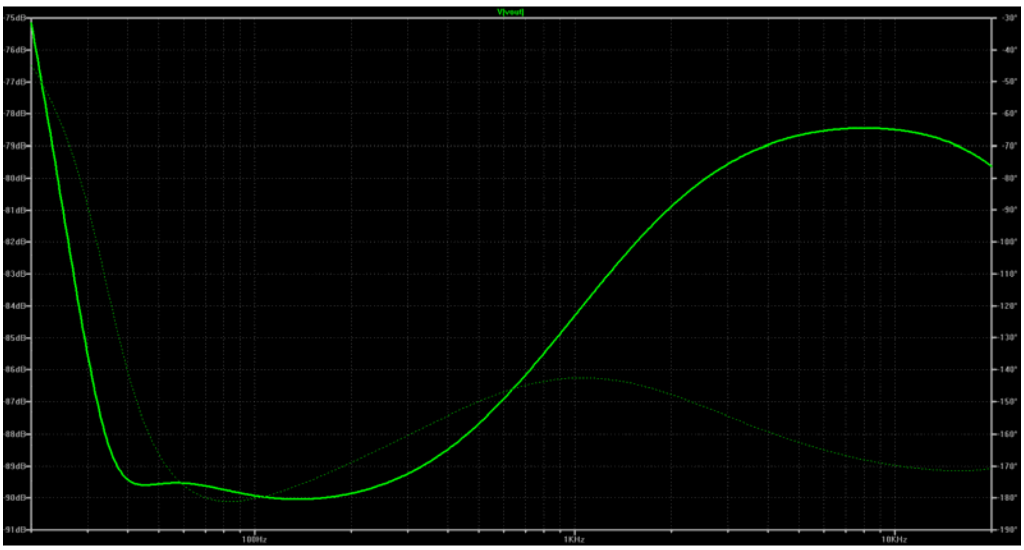


Figure 11 - PSRR of the Complete Equalizer

Fig. 11 shows that the conformity of the RIAA amplifier approaches ~0.1 dB 20 Hz to 20 kHz – this of course with ideal components. This is a very good result, but, as we see in Fig. 12, 1% component spreads easily turn good theoretical performance into something altogether more pedestrian – a factor that may necessitate 0.1% components in some circuit locations, selection, or paralleling of critical passive circuit elements in order to capitalize on the 1/√n tolerance spread reduction this technique brings.

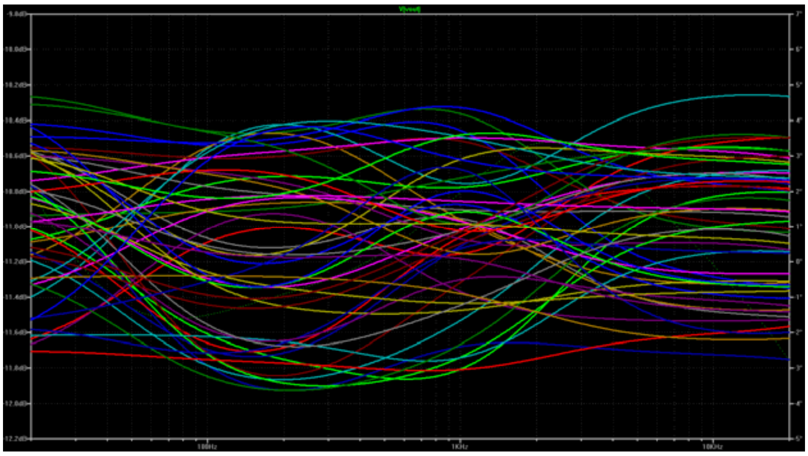
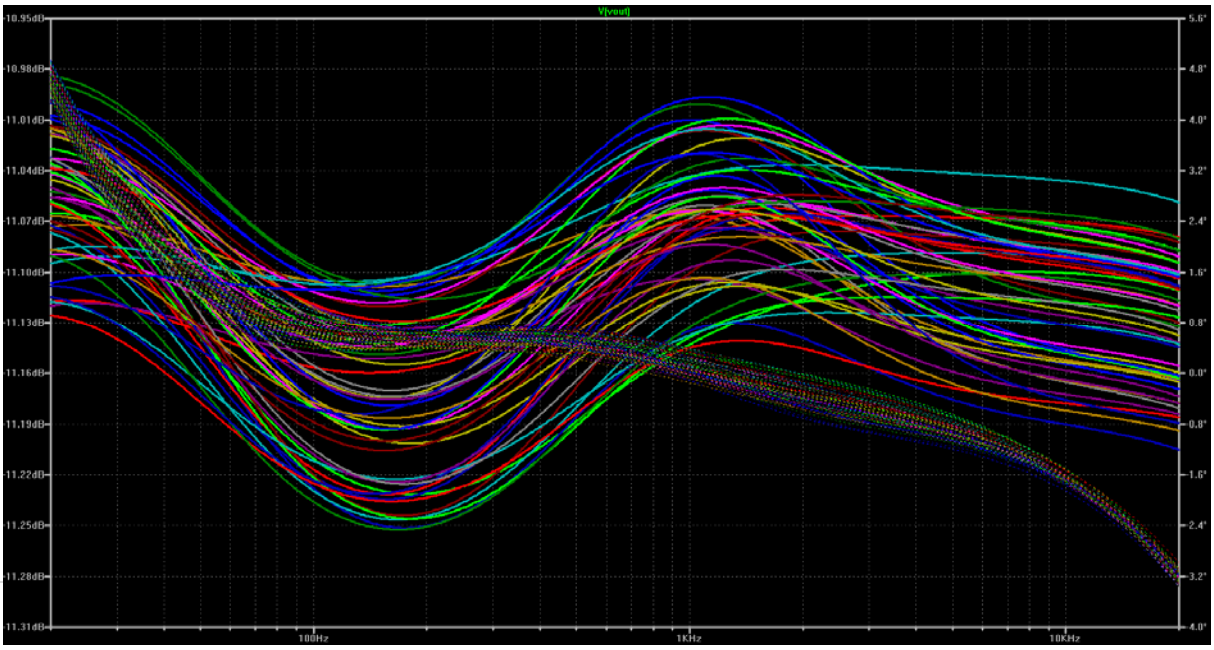


Figure 12 – 50 run Monte Carlo analysis with 1% components

The worst case conformity on a single run is ~1.4 dB, while the total conformity spread approaches 1.8 dB. This result makes it clear that 1% components are <u>not</u> good enough if one expects tight conformity. Even with 0.1% components (an expensive proposition), the spread is around 0.3 dB across the audio band (see Fig. 13).

A smart way to deal with this problem is to parallel 1% components - for each doubling of the number of parallel devices, a reduction in spread of 1/√2 can be expected; thus 4 parallel 1% resistors can yield a 0.36% tolerance composite resistor (the rule applies to series connection of the same value as well). For production, a combination of paralleling and selection is the best way forward. Note that this problem applies to any type of RIAA network design – it's not a problem unique to this design. Fig. 13 displays the conformity using all 0.1% components.

Figure 13 - Conformity Spread with 0.1% component tolerances is about 0.3 dB



The final point to discuss is that of noise performance, and here we see this design is quite capable of achieving 79 dB in the mid-band (Fig. 14) with a 3 mV input reference with a real world cartridge attached to the front end (in this case, a Shure V15 model is used as the source).

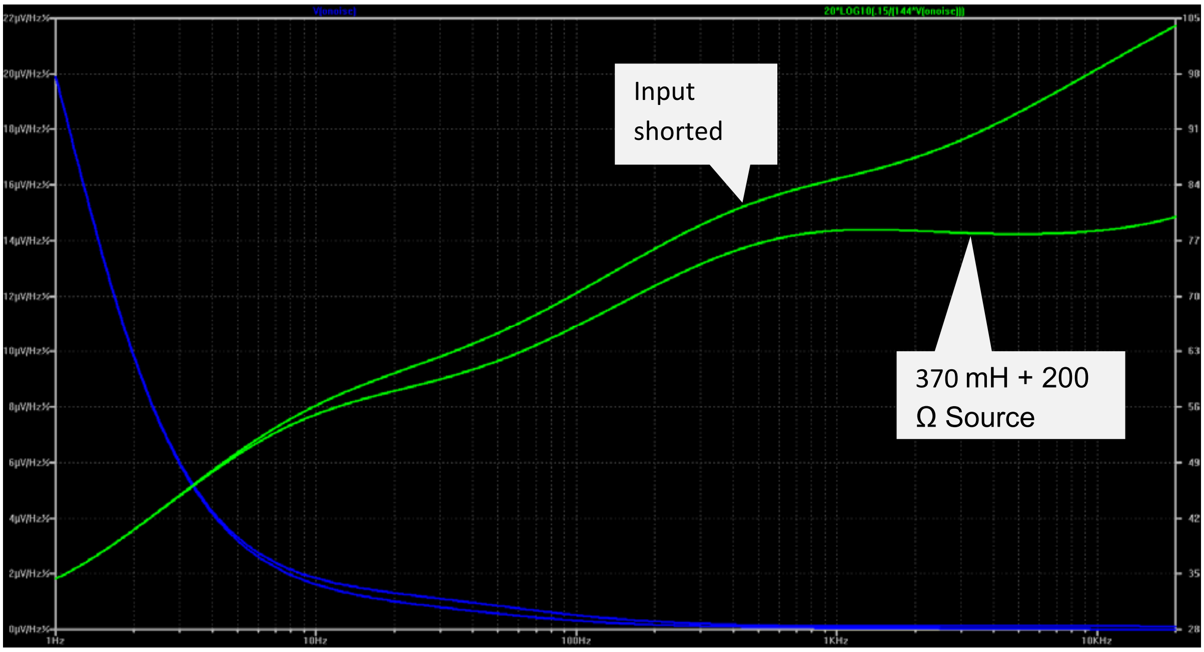


Figure 14 - Noise Performance with a 370mH + 200 Ω Source ref 150mV output (3mV input at 1 kHz). Green trace is S/N ratio (RHS axis) The blue trace is the equivalent output noise expressed as uV/√Hz – LHS axis

Equalization and 1/f noise cause the S/N ratio to degrade quickly below 1 kHz, but luckily the ears sensitivity to LF noise is low so the perception is still that the equalizer is exceptionally quiet. This type of S/N curve is typical of RIAA equalizers – they all degrade severely at LF because below 500 Hz, the signal is boosted by 20 dB/decade as part of the EQ process. This is another reason manufacturers almost always quote S/N ratio at 1 kHz (without stating so), or as a total equivalent input noise voltage, exploiting the lower spectral densities of LF noise in the process. Note that these noise performance figures are with the Shure V15 370 mH +200 Ω equivalent input source resistance – i.e. realistic. If we short the input (something Stereophile's John Atkinson does when he tests phono amplifiers), the S/N ratio at 1 kHz is 85 dB ref 150mV out and 3mV in. Since most RIAA equalizers are specified this way, this places this design in the top performance bracket in this regard. For a 5 mV input signal, the 1 kHz corresponding S/N ratio is 89 dB – an excellent figure by any measure.

Figure 15 - MC Noise Performance

On the output buffer opamp (U2), the gain can be switched between 0 dB and +14 dB. This allows the design to cater for MC pickups with nominal outputs of around 500 uV. Figure 15 shows that the noise performance is still outstanding at 71 dB at 1 kHz. For these tests, the assumed cartridge source resistance is 50 Ω while the load is 100 Ω, and the output 150mV.

High output MC cartridges offer up to 3 mV outputs, but couple this with very low inductive source impedances and these give the very best noise performance. In the case of this design, at 1kHz this would yield a S/N ratio of around 85dB and at 20 kHz, in excess of 100 dB

By way of contrast, Fig. 16 shows the S/N ratio of an opamp only RIAA equalizer – so no discrete JFET front end. With the input shorted, the performance approaches that of the discrete design, and would rank as 'state of the art' in a Stereophile review test – the 1 kHz S/N ratio is 85 dB. However, with our reference Shure V15 connected to the input, the S/N degrades to 69 dB at 1 kHz, and at HF declines to about 65 dB. It's clear here that the opamp current noise is interacting with the source impedance to cause a significant noise problem at HF. This plot was generated using an LM1115, which is promoted by LT as a 'low noise' opamp suited to RIAA EQ applications. Using an AD797, one will get better performance than shown above, but even with the best opamps available today, they will still fall short of the NE5532/34 which would be about 3 dB better than the figures shown here. The reason of course is that the NE5532/34 has about half the noise current of competing designs (at the expense of *very high input bias current* I have to add). So, although the discrete JFET design presented here is complex, the benefit is ~12 dB gain in S/N ratio over straight opamp based designs at 1 kHz, and about 6 dB better than the NE5532/34. Well worth the effort in my book.

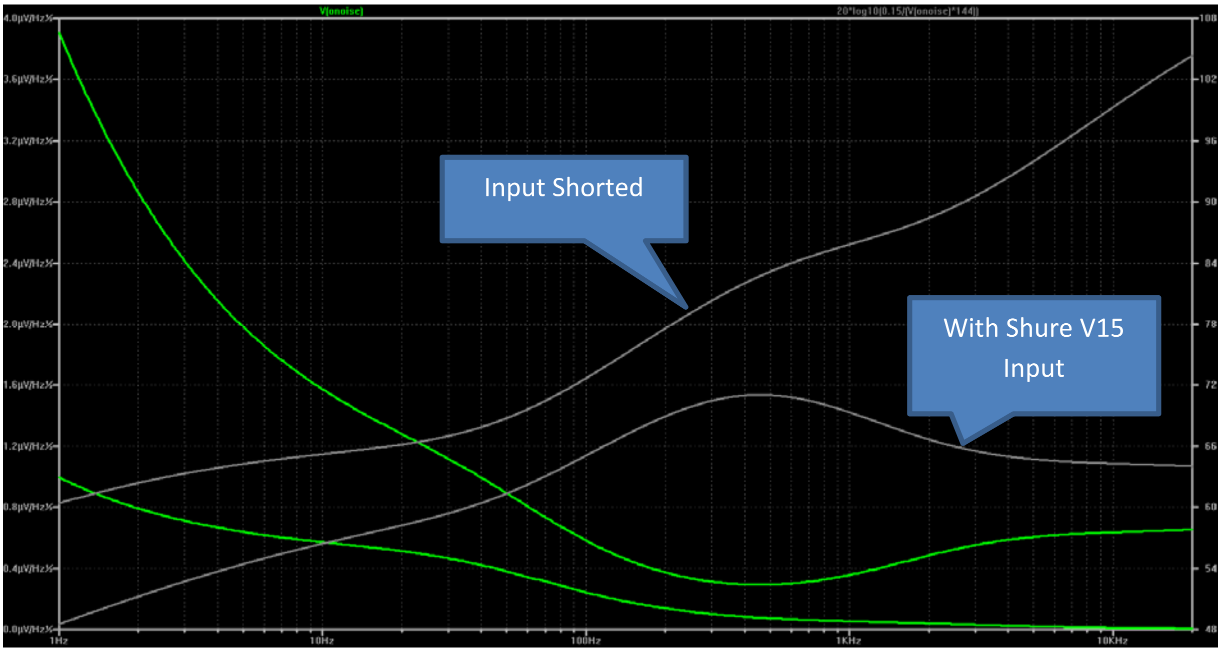


Figure 16 – LT1115 Opamp RIAA EQ Noise Performance (shown for comparison purposes)

Some General Thoughts on Power Supplies

The equalizer amplifiers presented here are designed to work off 3 terminal regulated power supplies – no need for fancy shunt regulators or anything like that. The standard LM317/337 regulators can be easily turned into *very low noise* power supplies by simply adding a capacitance multiplier on the output. Since all of the designs presented here run in class A, the load is fairly constant, and there are no issues with load regulation.

Further, given the very good performance of standard 3 terminal regulators, along with VFA op-amp PSU rejection capability, most of the PSU design effort should be on wiring and layout: that will have a much bigger impact on sonics. I generally heavily decouple my op-amps with a series 22 Ohm/100 uF capacitor on both supply rails and placed as close as possible to the op-amp supply pins, which provides *heavy damping* of the rails. Short, high current demands from the op-amp, in as much as they exist in a line level preamplifier, required to drive the load are kept in a tight loop around the op-amp and trace inductances are then no longer a problem, since it is only the low frequency and DC current flowing into the local decoupling capacitors through the series resistor. This decoupling method also provides substantial filtering of regulator wideband noise. I see a lot of effort put into designing 'stiff' tightly regulated supplies for op-amp circuitry and most of it is misguided. If you have a 'stiff' regulator, this will require a low output impedance, and it also mean that the wideband noise from the regulator will be emanating from a low source impedance. Further, class AB output amplifier stages dump a lot of wideband current mode hash onto the rails and tightly regulated power supplies, coupled with the inevitable trace inductances, can lead to ringing. Kendall Castor-Perry did a comprehensive series of articles on the subject with lots of great advice for how to get it right - clearly, for digital circuitry and high speed data conversion, frequency generation and so forth, you do need stiff, low source impedance supplies. But, for line and source level audio you do not – in fact, they only serve to make matters worse for the most part.

The supply rails on an op-amp for audio work can move around by a few tens or hundreds of mV *as long as it is at a very low frequency (single digit Hz)*, where PSRR is very good. It's the HF PSU noise that causes the problem, but this is easily taken care of by the filter or ripple eater mentioned above. If you bias the output stage into class A as shown in the designs discussed in this document, supply current harmonics will be low order 2nds with a little 3rds which are altogether more easy to deal with – the sonic benefits of class A aside.

Given this philosophy, you can see why I don't believe in 'ultra' regulators. It is important in the PCB layout, to keep the supply rails and decoupling components in a tight, small loop to reduce noise radiation and minimize any common impedance coupling. On a typical pre-amp, trace inductance will start to make itself felt as frequencies approach 100 kHz over trace lengths as short as 10cm's.

Generally, I like to use split secondary transformers, and build completely separate + and - supplies, coupling them after the regulators to form the split supply rails. Using this approach makes it very easy to isolate the filter capacitor charging currents from the rest of the circuitry, the only cost being an additional bridge rectifier – of no consequence in a high end design. Furthermore, it means that an LM317 +ve 3 terminal regular can be used for both rails – the LM317 noise, line and load regulation performance is considerably better than its LM337 -ve 3 terminal regulator complement.