On Reference RIAA Networks

by Jim Hagerman

You'd think there would be nothing left to say. Everything you need to know about RIAA networks has already been published. However, a few years back I came across an interesting chapter in a vacuum tube book[1] which spoke of a mythical 3.18µs corner in the RIAA response. Huh? That's 50kHz. The seminal works[2][3] never mentioned this, equalization was to appear as Figure 1.
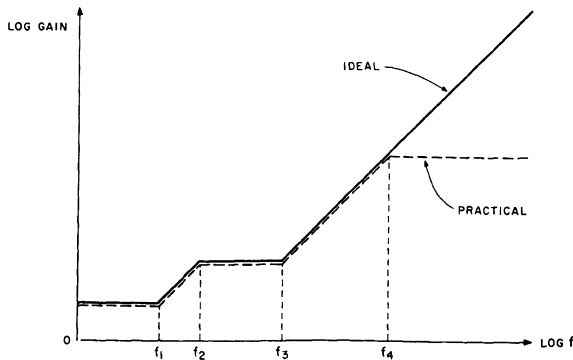

Figure 1. Desired inverse RIAA response from Lipshitz.

Hmmm. The mythical corner frequency is shown as f_4 but ideally should be missing. Or should it? To quote from Allen Wright's book[1]:

> "... look back at the graph of the recording EQ, they cut the LF and boost the HF. But do you really think they continue boosting to way out past whatever? Of course not, they'd burn out their cutter heads or something even more expensive ... This new 3dB point, according to a *Neumann* cutting amp manual, is set at 3.18µs – which equates to 50,048Hz ..."[1]

Not only does the cutting head response have a pole at f_4, but it must also have one at f_5! No amplifier has gain out to infinity. Hopefully, all cutting head manufacturers chose the same 3.18µs corner for limiting gain. Where is all this leading? It means that the RIAA equalization networks in our phono preamplifiers should have a zero at 3.18µs, putting back some gain before finally rolling off at higher frequencies. The legacy reference network[2] shown in Figure 2 has an f_4 pole at 337kHz – too high of frequency.
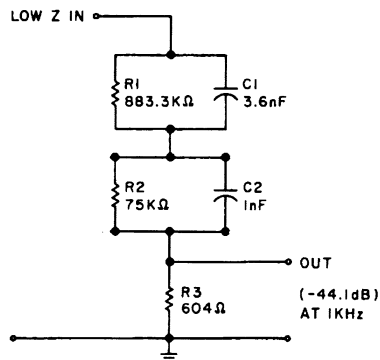

Figure 2. Legacy reference inverse RIAA network from Lipshitz.

Reference Inverse RIAA

What we need is a new modified RIAA reference curve to help us properly design phono equipment. I like using SPICE to simulate filter circuits and decided this would be a good way to generate a new standard. The generic filter section shown in Figure 3 is a simple lag-lead type with a zero and pole in the right places.
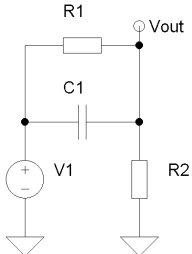

Figure 3. Lag-lead filter section.

Its voltage transfer function is given by

$$H(s) = \frac{s + \dfrac{1}{R_1 C_1}}{s + \dfrac{R_1 + R_2}{R_1 R_2 C_1}}$$

with zero and pole time constants determined by

$$t_z = R_1 C_1$$

$$t_p = \left(\frac{R_1 R_2}{R_1 + R_2}\right) C_1$$

The lower RIAA zero-pole pair is at 3180μs and 318μs. Using an arbitrary capacitance value of 1μF, the resistances are calculated as

$$R_1 = \frac{t_z}{C_1} = \frac{3180\mathit{ms}}{1\mathit{mF}} = 3.18k$$

$$R_2 = \frac{R_1 t_p}{R_1 C_1 - t_p} = \frac{(3.18k)(318\mathit{ms})}{(3.18k)(1\mathit{mF}) - 318\mathit{ms}} = 353.3$$

The values for a second section (zero at 75μs and pole at 3.18μs) are 7.5k and 353.3 ohms respectively using a 10nF capacitor. The final circuit is shown in Figure 4. Note, I used a voltage controlled voltage source (E1) to decouple the responses of the two sections, otherwise the input impedance of the second section would load the first section and alter response.
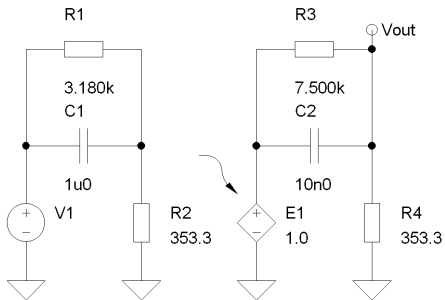
Figure 4. SPICE schematic for generating reference inverse RIAA curve.

Listing 1 is the input text file to my SPICE simulator. Figure 5 shows the resulting frequency response of the circuit, which is also given in tabular form in Listing 2. I offset the data so that gain at 1kHz would be 0dB. I find it helpful to sweep a wide frequency range of 1Hz to 1MHz as it gives a better view of the total response and what occurs outside of the 20 to 20,000Hz "audio band".

```
                PSPICE Input
Inverse RIAA Curve

Vin 1 0 ac 1

R1 1 2 3.18k
R2 2 0 353.33
C1 1 2 1u

E1 3 0 2 0 1.0
R3 3 4 7.5k
R4 4 0 353.33
C2 3 4 10n

.ac dec 20 1 1000k
.print vm(4)
.probe
.end
```

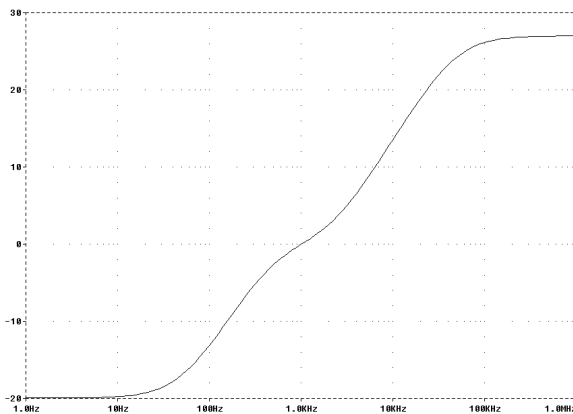**Listing 1. SPICE input file for generating modified reference curve.**

Figure 5. SPICE output of reference inverse RIAA curve with 1kHz set to 0dB.

Modified Inverse RIAA					
frequency	dB	frequency	dB	frequency	dB
10.00	-19.74	223.9	-7.43	5012	8.18
11.22	-19.70	251.2	-6.64	5623	9.04
12.59	-19.64	281.8	-5.88	6310	9.92
14.13	-19.58	316.2	-5.15	7079	10.81
15.85	-19.50	354.8	-4.46	7943	11.72
17.78	-19.40	398.1	-3.81	8913	12.63
19.95	-19.27	446.7	-3.20	10000	13.55
22.39	-19.12	501.2	-2.63	11220	14.46
25.12	-18.94	562.3	-2.11	12590	15.37
28.18	-18.72	631.0	-1.63	14130	16.27
31.62	-18.46	707.9	-1.19	15850	17.17
35.48	-18.16	794.3	-0.77	17780	18.04
39.81	-17.80	891.3	-0.38	19950	18.89
44.67	-17.40	1000	0.00	22390	19.71
50.12	-16.93	1122	0.38	25120	20.50
56.23	-16.41	1259	0.77	28180	21.25
63.10	-15.84	1413	1.17	31620	21.96
70.79	-15.22	1585	1.60	35480	22.62
79.43	-14.55	1778	2.07	39810	23.23
89.13	-13.83	1995	2.57	44670	23.79
100.0	-13.09	2239	3.12	50120	24.28
112.2	-12.31	2512	3.72	56230	24.72
125.9	-11.51	2818	4.36	63100	25.11
141.3	-10.70	3162	5.05	70790	25.44
158.5	-9.88	3548	5.78	79430	25.72
177.8	-9.05	3981	6.55	89130	25.96
199.5	-8.23	4467	7.35	100000	26.16

Listing 2. Results from SPICE simulation.

New Network Design

Figure 2 can be modified to achieve the desired results. In order to shift the high frequency pole the value of R3 must change. Unfortunately, this also changes the gain of the network. However, by moving part of R3 to the input side, we can control both gain and pole frequency independently.

A reference network should interface nicely between test equipment and phono preamplifiers. Therefore, I selected the following design parameters:

- 50 ohm source impedance (many generators do not use the 600 ohm audio standard)
- 600 ohm output impedance
- Dual output gains of –40dB and –60dB @1kHz
- Standard capacitance values.
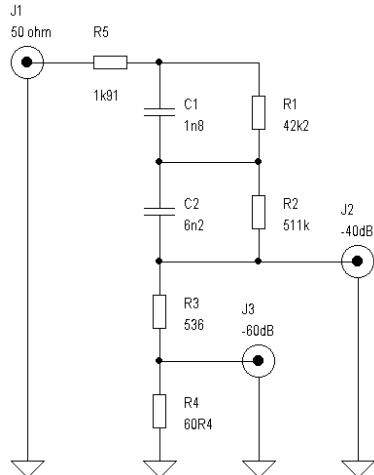
Figure 6 shows my new modified RIAA network.

Figure 6. Modified inverse RIAA circuit.

Optimizing component values is easily done by iterative SPICE simulations, but a good starting point is needed. Approximate values can be calculated by utilizing known boundary conditions. Since output impedance should be 600 ohms, and the two outputs are 20dB apart, we can write

$$R_3 + R_4 = 600$$

$$\frac{R_4}{R_3 + R_4} = -20dB = 0.1$$

Solving we get

$$R_4 = (0.1)(600) = 60$$
$$R_3 = 600 - 60 = 540$$

At high frequency the capacitors appear as short circuits and the network simplifies to a resistor divider comprised of R5, R3, and R4. From Figure 5 we see that high frequency gain is about 27dB higher than at 1kHz. Since desired 1kHz gain is –40dB, our high frequency gain will be about –13dB. The gain equation

$$dB = 20\log A_{HF}$$

is rewritten and solved as

$$A_{HF} = 10^{\left(\frac{dB}{20}\right)} = 10^{\left(\frac{-13}{20}\right)} = 0.22$$

High frequency divider gain is then given by

$$A_{HF} = \frac{R_3 + R_4}{R_3 + R_4 + R_5} = 0.22$$

and we can solve for R5

$$R_5 = \frac{R_3 + R_4}{A_{HF}} - R_3 - R_4 = \frac{600}{0.22} - 600 = 2.1k$$

At low frequency we have the opposite effect and the capacitors appear as open circuits. The divider again is resistive and has a gain of –60dB. This is written as

$$A_{LF} = \frac{R_3 + R_4}{R_1 + R_2 + R_3 + R_4 + R_5} = 0.001$$

and the equivalent series resistance of R1 and R2 is solved as

$$R_1 + R_2 = 999(R_3 + R_4) - R_5 = 597k = R_{eq}$$

Our high frequency pole at 3.18µs is equal to the series resistance of R3, R4, and R5 times the equivalent series capacitance of C1 and C2. This capacitance is given by

$$C_{eq} = \left(\frac{C_1 C_2}{C_1 + C_2}\right) = \frac{3.18\mathbf{\mu s}}{R_3 + R_4 + R_5} = \frac{(3.18\mathbf{\mu s})}{600 + 2.1k} = 1.2nF$$

Finally, we have four equations and four unknowns

$$R_1 C_1 = 75\mathbf{\mu s}$$
$$R_2 C_2 = 3180\mathbf{\mu s}$$
$$C_{eq} = 1.2nF$$
$$R_{eq} = 597k$$

I'll spare you the math, R1 is solved as

$$R_1 = \frac{\frac{(R_1 C_1)(R_2 C_2)}{C_{eq}} - (R_1 C_1)R_{eq}}{(R_2 C_2) - (R_1 C_1)} = 50k$$

and R2 as

$$R_2 = R_{eq} - R_1 = 597k - 50k = 547k$$

and the capacitors as

$$C_1 = \frac{75\mathbf{\mu s}}{R_1} = 1.5nF$$
$$C_2 = \frac{3180\mathbf{\mu s}}{R_2} = 5.8nF$$

These are only starting values. For a best fit real world design I optimized for the nearest 1% resistor and standard capacitor values (as shown in Figure 6).

Comparison: Old vs. New

As a reference network the performance must be pretty good. Figure 7 shows the frequency response errors relative to Listing 2. My new network is accurate to within +/-0.1dB over the full frequency range. For comparison I ran an error simulation of the Lipshitz network in Figure 2. It is extremely good up to about 10kHz where it drops off as expected due to the missing 3.18µs corner. The Lipshitz circuit can, however, be modified by adding a 3.83k resistor as R5 at the input to reposition the high frequency pole.
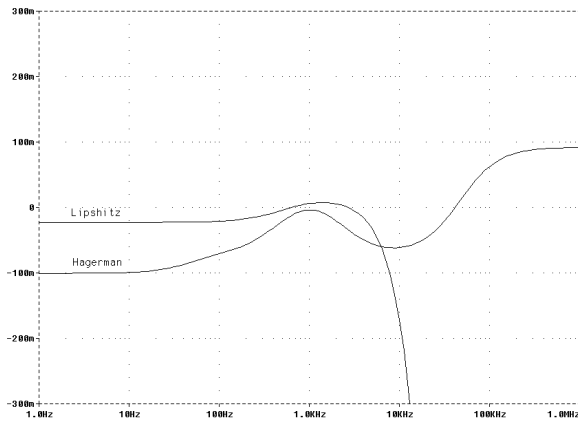
Figure 7. SPICE generated error responses for Lipshitz and Hagerman networks.

To check the sensitivity of component values I ran a few more simulations varying capacitances by 5%. Errors remained within +/-0.3dB.

I hope this information is helpful to designers of high end audio equipment.

References

[1] A. Wright, "The Tube Preamp Cookbook", *Vacuum State Electronics*, 1995.
[2] S. Lipshitz, and W. Jung, "A High Accuracy Inverse RIAA Network", *Audio Amateur*, 1980.
[3] S. Lipshitz, "On RIAA Equalization Networks", *JAES* 1979.
[4] "AN-124: Three High Accuracy RIAA/IEC MC and MM Phono Preamplifiers", *Analog Devices*, 1992
[5] M. Giles, "Audio/Radio Handbook", *National Semiconductor*, 1980
[6] A. Wright, "Secrets of the Phono Stage", *Sound Practices* #15, 1998

Sources

The network of Figure 6 is available as a kit (order #KF-1) from Old Colony Sound Lab for $25. The two channel kit comes complete with PCB, 1% metal film resistors, 2% polypropylene capacitors, connectors, and instructions.

Old Colony Sound Lab
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