Split active RIAA correction with limited loop gain

Marcel van de Gevel, version 3, 31 January 2024

1. Introduction

In the thread https://www.diyaudio.com/community/threads/good-resource-for-learning-how-to-figure-riaa-eq-in-a-feedback-loop-active-riaa.408061/ rongon came up with the idea to make a moving-magnet phono preamplifier with split RIAA correction, where the 3180 μs and 318 μs part is done actively by shunt feedback. chip_mk wrote that if the first stage would be a plain old common-cathode stage with the 75 μs implemented with a capacitor across its output, the capacitor would very much limit the Miller effect. I did some calculations on this configuration.

The derivation of approximate expressions for the tranfer are in section 2. In section 3, a proposal for a procedure to dimension the circuit is made. In section 4, an example is given, which is checked with a pole-zero extraction program. The conclusion is in section 5.

2. Transfer function

Suppose you wanted to implement the split RIAA amplifier with active RIAA correction via shunt feedback in the second stage, while using only a single triode in each stage. You would then end up with something like the upper schematic of these three:

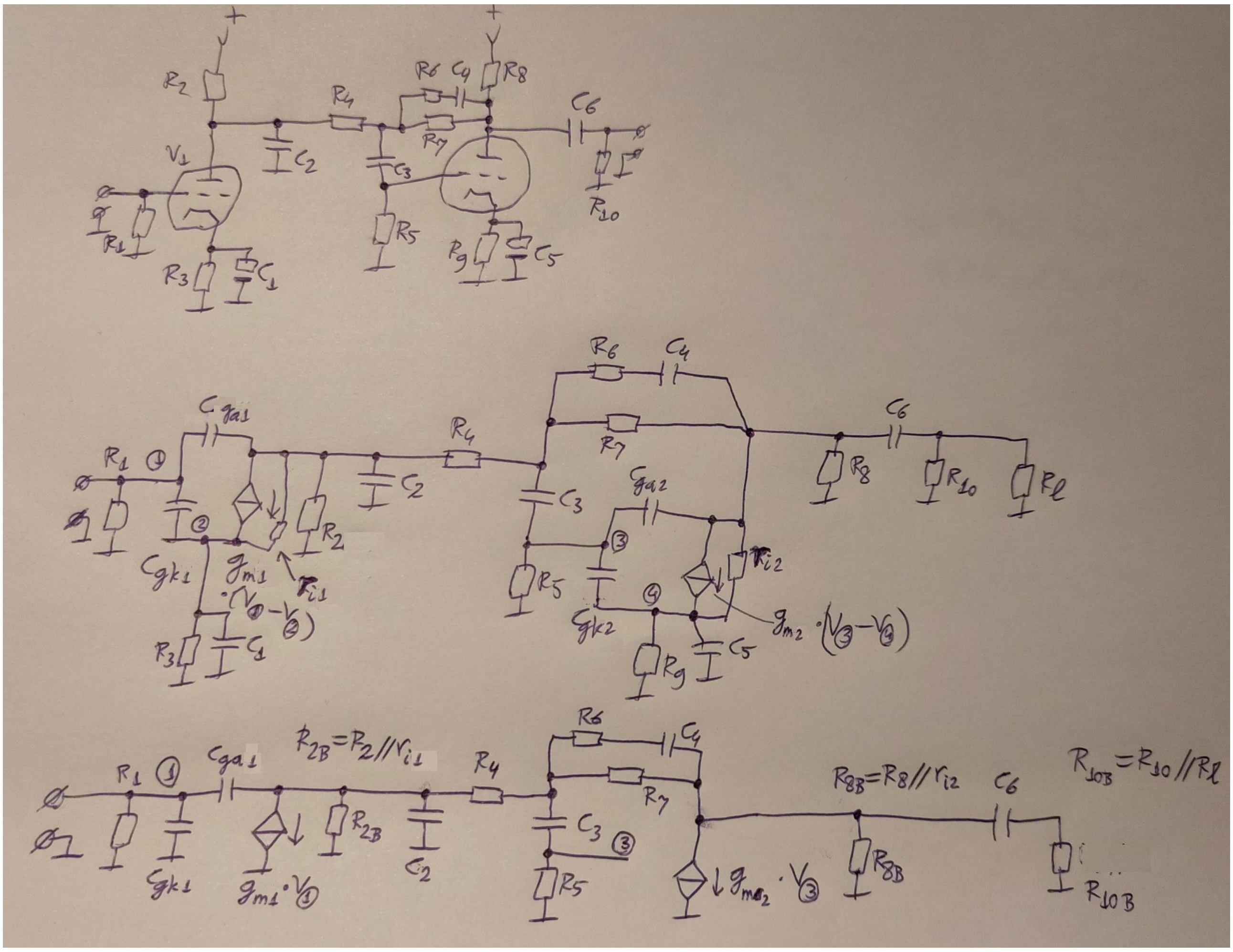


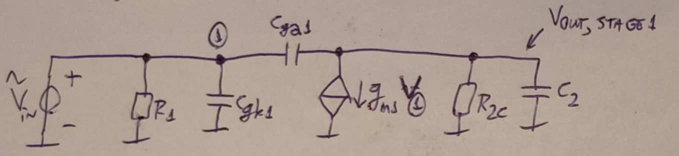
Figure 1: Amplifier schematic, network model and first simplification of the network model. The symbol // stands for "in parallel with".

When you simply replace each resistor and each capacitor with an ideal one and each triode with the usual small-signal models, you end up with the middle schematic. It's ninth order (there are ten capacitors plus zero inductors, but four of the capacitors are in a loop, reducing the order to nine). The order reduces to eight when you connect a voltage source to the input, but that is still quite high, and everything influences everything, making it impractical to use this schematic for hand calculations.

Assuming that the cathode decoupling capacitors can be made so large that their effect can be neglected and that the internal capacitances of the second valve are small enough not to do much harm, you end up with the third schematic, which is still fifth or fourth order with everything influencing everything.

Further simplifications are possible by assuming that the AC coupling capacitors C_3 and C_6 will also be made big enough, which is an assumption that can always be met but is not always met, and that C_4 will be a reasonably good short at 2122 Hz, while C_2 and the capacitances of the first valve will be a reasonably good open branches at 50 Hz. When calculating the required value of C_2 and the resistors around it, the imperfect virtual ground of the second stage can then be modelled as having a constant input impedance Z_{st2} for frequencies in the vicinity of 2122 Hz. Conversely, the capacitances of the first valve and C_2 can then be neglected when calculating the values of C_4, R_6 and R_7.

STAGE 1.



$R_{2C} = R_2 // r_{i1} // (R_4 + Z_{ST2})$, WHERE // MEANS IN PARALLEL WITH.

r_{i1}: INTERNAL RESISTANCE FIRST TRIODE

Z_{ST2}: INPUT IMPEDANCE "VIRTUAL GROUND" SECOND STAGE, ASSUMED ROUGHLY CONSTANT AROUND 2122 Hz

REPLACE THE CONTROLLED SOURCE WITH AN INDEPENDENT ONE WITH STRENGTH $g_{m1}\tilde V_{IN}$ AND APPLY SUPERPOSITION: FIRST CALCULATE THE EFFECT OF THE CURRENT SOURCE WITH THE VOLTAGE SOURCE SET TO 0, THEN THE EFFECT OF THE VOLTAGE SOURCE WITH THE CURRENT SOURCE OFF, THEN ADD THE RESULTS.

$$\tilde V_{OUT,STAGE1} = \underbrace{-g_{m1}\tilde V_{IN}\frac{R_{2C}}{sR_{2C}(C_2+C_{ga1})+1}}_{\text{CURRENT THROUGH THE CURRENT SOURCE TIMES THE IMPEDANCE IT FLOWS INTO WHEN THE INPUT VOLTAGE SOURCE IS 0}} + \underbrace{\tilde V_{IN}\frac{\frac{R_{2C}}{sR_{2C}C_2+1}}{\frac{R_{2C}}{sR_{2C}C_2+1}+\frac{1}{sC_{ga1}}}}_{\text{VALUE OF THE VOLTAGE SOURCE TIMES THE TRANSFER OF THE VOLTAGE DIVIDER } C_{ga1},\, R_{2C},\, C_2 \text{ WHEN THE CURRENT SOURCE IS OFF}}$$

THE EQUATION FOR $V_{OUT,STAGE1}$ CAN BE SIMPLIFIED BY MULTIPLYING THE NUMERATOR AND DENOMINATOR OF THE SECOND TERM BY $sC_{ga1}(sR_{2C}C_2+1)$. THIS RESULTS IN

$$V_{OUT,STAGE1} = -\tilde V_{IN}\frac{g_{m1}R_{2C} - sR_{2C}C_{ga1}}{sR_{2C}(C_2+C_{ga1})+1}$$

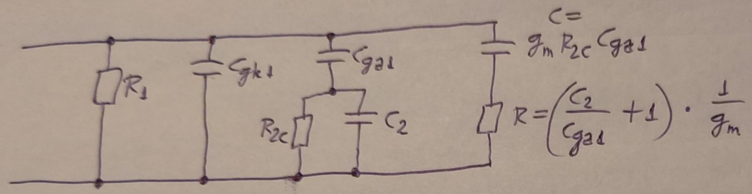
HENCE, THERE IS A POLE AT $s = -\dfrac{1}{R_{2C}(C_2+C_{ga1})}$.

FOR PROPER RIAA CORRECTION, IT HAS TO LIE AT $s = -\dfrac{1}{75\,\mu s}$,

SO $R_{2C}(C_2+C_{ga1}) = 75\,\mu s$.

THERE IS ALSO A ZERO AT $\dfrac{g_{m1}R_{2C}}{R_{2C}C_{ga1}} = \dfrac{g_{m1}}{C_{ga1}}$. IT IS UNDESIRED, BUT IT IS USUALLY SO FAR AWAY THAT ITS EFFECT IS NEGLIGIBLE.

USING THE SAME METHOD TO ANALYSE THE INPUT IMPEDANCE OF STAGE 1 RESULTS IN:



WHERE $g_m R_{2C} C_{ga1}$ IS ESSENTIALLY THE MILLER CAPACITANCE. IT GETS A SERIES RESISTANCE SUCH THAT THE RC-PRODUCT $g_m R_{2C} C_{ga1} \left(\frac{C_2 + C_{ga1}}{C_{ga1}}\right) \cdot \frac{1}{g_m} = R_{2C}(C_2 + C_{ga1}) = 75\ \mu s.$

IN THE FREQUENCY RANGE WHERE IT MATTERS MOST, 10 kHz ... 20 kHz, THE RESISTANCE DOMINATES. YOU CAN CORRECT FOR IT BY SLIGHTLY INCREASING R_1.

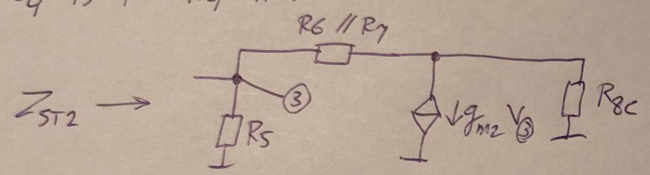
When input capacitance is not that critical while noise is, you could add an RC parallel network (with a DC blocker capacitor in series) with $R(C + c_{ga1}) = 75$ µs between anode and grid of the first valve to make an electrically cold input resistance. R_{2C} then has to be slighly increased, such that $R_{2C} C_2 = 75$ µs (instead of $R_{2C} (C_2 + c_{ga1}) = 75$ µs). I haven't looked into that in more detail, though.

Regarding the virtual ground impedance,

Z_{ST2}

THE EQUATIONS FOR STAGE 1 DEPEND ON Z_{ST2}, THE IMPEDANCE OF THE VIRTUAL GROUND OF THE SECOND STAGE AROUND 2122 Hz.

C_4 IS ROUGHLY A SHORT AT 2122 Hz, SO



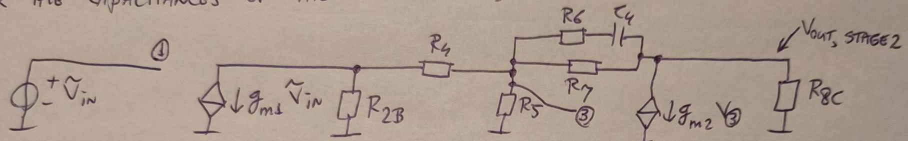
USING THE SAME TRICK AS FOR STAGE 1,

$$Z_{ST2} = \frac{1}{\frac{1}{R_5} + \frac{1}{\frac{R_6 R_7}{R_6 + R_7} + R_{8C}} + g_{m2} \cdot \frac{R_{8C}}{\frac{R_6 R_7}{R_6 + R_7} + R_{8C}}} = \frac{1}{\frac{1}{R_5} + \frac{g_{m2} R_{8C} + 1}{\frac{R_6 R_7}{R_6 + R_7} + R_{8C}}}$$

Here, $R_{8C} = R_8 // r_{i2} // R_{10} // R_L$

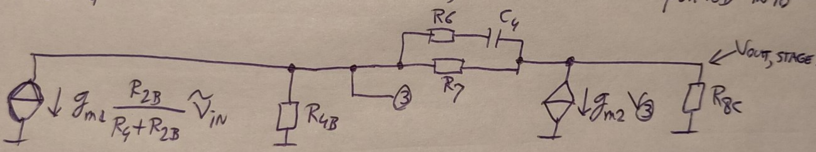
SECOND STAGE

ASSUMING C_2, WHICH SETS THE $75\,\mu s$ TIME CONSTANT, DOESN'T HAVE MUCH IMPACT ON THE $3.18\,ms$ TIME CONSTANT, ~~[illegible]~~ AND THAT THE SAME HOLDS FOR THE CAPACITANCES OF THE FIRST AND SECOND VALVES WE CAN USE THIS MODEL:



$R_{8C} = R_8 // r_{i2} // R_{10} // R_l$

AS R_{2B} AND R_4 FORM A CURRENT DIVIDER, IT CAN BE SIMPLIFIED INTO



$$R_{4B} = (R_4 + R_{2B}) // R_5$$

AS THERE IS ONLY ONE REACTIVE COMPONENT, THIS NETWORK CAN ONLY HAVE ONE POLE. IT SUFFICES TO CALCULATE ANY TRANSFER TO FIND THAT POLE, AS ALL TRANSFERS HAVE THE SAME POLES.

THERE IS A ZERO IN THE TRANSFER TO THE OUTPUT WHERE THE PATH THROUGH g_{m2} CANCELS THE PATH THROUGH THE PARALLEL CONNECTION OF R_7 AND R_6-C_4. THAT IS, THE ZERO IS THE VALUE OF s THAT MAKES $\frac{1}{R_7} + \frac{sC_4}{sR_6C_4+1} = g_{m2}$.

HENCE, $$g_{m2} - \frac{1}{R_7} = \frac{sC_4}{sR_6C_4+1}$$

$$s\left(g_{m2} - \frac{1}{R_7}\right)R_6C_4 + g_{m2} - \frac{1}{R_7} = sC_4$$

$$sC_4\left(R_6\left(g_{m2} - \frac{1}{R_7}\right) - 1\right) = -g_{m2} + \frac{1}{R_7}$$

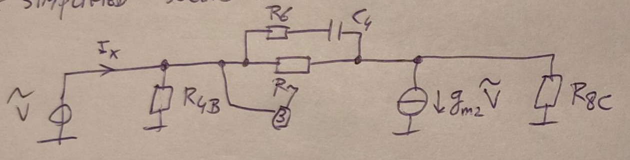
$$s = -\frac{g_{m2} - \frac{1}{R_7}}{\left(R_6\left(g_{m2} - \frac{1}{R_7}\right) - 1\right)C_4} = -\frac{1}{\left(R_6 - \frac{1}{g_{m2} - \frac{1}{R_7}}\right)C_4}.$$

TO GET THE ZERO AT THE PROPER LOCATION $s = -\frac{1}{318\,\mu s}$,

$$\left(R_6 - \frac{1}{g_{m2} - \frac{1}{R_7}}\right)C_4 = 318\,\mu s$$

FOR LARGE TRANSCONDUCTANCE, $g_m \to \infty$, THIS TENDS TO $R_6C_4 = 318\,\mu s$.

TO FIND THE POLE OF THE SECOND STAGE, WE CAN CALCULATE THE TRANSFER FROM INPUT CURRENT TO ANYTHING, FOR EXAMPLE, TO THE INPUT VOLTAGE. THAT IS, THE POLE OF THE TRANSFER IS THE POLE OF THE INPUT IMPEDANCE. AS THE INPUT IMPEDANCE IS THE RECIPROCAL OF THE INPUT ADMITTANCE, IT IS ALSO THE ZERO OF THE INPUT ADMITTANCE OF THE SECOND STAGE OR OF THE SIMPLIFIED SECOND STAGE.



$$Y = \frac{I_x}{\tilde{V}} = \frac{1}{R_{4B}} + \frac{1}{R_{8C} + \frac{1}{\frac{1}{R_7} + \frac{sC_4}{sR_6C_4+1}}} + g_{m2} \cdot \frac{R_{8C}}{R_{8C} + \frac{1}{\frac{1}{R_7} + \frac{sC_4}{sR_6C_4+1}}}$$

$$= \frac{1}{R_{4B}} + \frac{g_{m2}R_{8C}+1}{R_{8C} + \frac{R_7(sR_6C_4+1)}{sR_6C_4+1+sC_4R_7}} = \frac{1}{R_{4B}} + \frac{(g_{m2}R_{8C}+1)(s(R_6+R_7)C_4+1)}{R_{8C}(s(R_6+R_7)C_4+1) + R_7(sR_6C_4+1)}$$

$$= \frac{\frac{R_{8C}}{R_{4B}}(s(R_6+R_7)C_4+1) + \frac{R_7}{R_{4B}}(sR_6C_4+1) + (g_{m2}R_{8C}+1)(s(R_6+R_7)C_4+1)}{R_{8C}(s(R_6+R_7)C_4+1) + R_7(sR_6C_4+1)}$$

THE NUMERATOR OF Y IS 0 AT THE SAME VALUE OF s WHERE THE SECOND STAGE HAS ITS POLE.

THAT IS,

$$\left(g_{m2} R_{8C} + \frac{R_{8C}}{R_{4B}} + 1\right)\left(s(R_6+R_7)C_4+1\right) + \frac{R_7}{R_{4B}}\left(s R_6 C_4 + 1\right) = 0$$

$$s\left(\left(g_{m2} R_{8C} + \frac{R_{8C}}{R_{4B}} + 1\right)(R_6+R_7)C_4 + \frac{R_7}{R_{4B}} R_6 C_4\right) = -\left(g_{m2} R_{8C} + \frac{R_{8C}}{R_{4B}} + 1 + \frac{R_7}{R_{4B}}\right)$$

$$s = -\frac{g_{m2} R_{8C} + \dfrac{R_7 + R_{8C}}{R_{4B}} + 1}{\left(\left(g_{m2} R_{8C} + \dfrac{R_{8C}}{R_{4B}} + 1\right)(R_6+R_7) + \dfrac{R_7 R_6}{R_{4B}}\right) C_4}$$

THE CORRECT LOCATION FOR RIAA-CORRECTION IS $s = -\dfrac{1}{3.18\text{ ms}}$,

SO

$$\frac{\left(g_{m2} R_{8C} + \dfrac{R_{8C}}{R_{4B}} + 1\right)(R_6+R_7) + \dfrac{R_7 R_6}{R_{4B}}}{g_{m2} R_{8C} + \dfrac{R_7 + R_{8C}}{R_{4B}} + 1} C_4 = 3.18\text{ ms}$$

FOR LARGE TRANSCONDUCTANCE, $g_m \to \infty$, THIS WOULD TEND TO $(R_6+R_7)C_4 = 3.18\text{ ms}$.

3. Dimensioning the circuit

In principle, one could choose values for all components but three, three that occur in the equations for the three RIAA time constants, for example R_4 (or R_{4B}), R_6 and R_7. These three are then regarded as three unknowns in a system of three equations. Hopefully, manipulating the equations will then lead to a solution for all three of them.

For simplicity, I will propose a more straightforward but less accurate method. One approach to dimensioning the circuit would be:

First stage:

1. Choose a triode with low noise and a high μ, for example an ECC83 (12AX7).

2. Choose the anode current for minimum noise. A too low anode current worsens white noise, a too high current worsens $1/f$ noise, somewhere in between there is an optimum. This optimum depends on the way the output noise is weighted.

The noise optimum can be measured directly or can be calculated when sufficient data about white and $1/f$ noise are available - data which are never found in valve datasheets. See Merlin Blencowe, "Noise in triodes with particular reference to phono preamplifiers", *Journal of the Audio Engineering Society*, vol. 61, no. 11, November 2013, pages 911...916, for a most interesting article about triode noise modelling with measured data for ECC81 (12AT7), ECC82 (12AU7), ECC83 (12AX7), ECC88 (6DJ8) and 6J52P valves. The data for the ECC83 are for two triodes in parallel, see http://www.diyaudio.com/forums/tubes-valves/251154-flicker-noise-dominates-triode-noise-audio-aes.html How to calculate noise optima using Merlin Blencowe's data is explained in my article "Grammophone preamplifier noise calculations - the 3852 Hz rule revisited" in *Linear Audio* volume 8.

The results for the ECC81 and ECC88 and for my own measurements for triode-connected EF86 valves are in 1.

Table 1 Optimal bias currents and weighted average noise voltage densities (that is, white noise level that would give the same weighted total noise as the actual valve) for a few common audio valves. The values for the ECC83 and ECC88 were calculated using data from Merlin Blencowe, the value for an EF86 and RIAA plus A-weighting was directly measured by me using a very limited number of EF86 valves, the value for an EF86 and RIAA plus brick wall weighting is an educated guess.

Valve type	RIAA and 20 Hz to 20 kHz brick wall	RIAA and A-weighting, 1 Hz to 1 MHz	RIAA and ITU-R 468
Triode-connected EF86	1 mA	1.75 mA 10.2 nV/√Hz	
ECC88 (single triode)	1.25 mA 14.3 nV/√Hz	2.5 mA 5.75 nV/√Hz	4.5 mA 4.4 nV/√Hz
ECC83 (single triode)	0.25 mA 12.1 nV/√Hz	0.75 mA 6.05 nV/√Hz	0.9 mA 5.26 nV/√Hz

Once the anode current of the first stage is chosen, the corresponding transconductance g_{m1} and internal resistance $r_{i1} = \mu_1/g_{m1}$ should follow from the valve datasheet.

3. Look up in the datasheet what negative grid-cathode voltage is required to get a negligibly small grid current. Choose R_3 accordingly, that is,

$$R_3 = -\frac{V_{GK1}}{I_{A1}}$$

4. Look up in the datasheet what anode voltage belongs with this grid voltage and anode current.

5. Choose R_2 such that you end up at the desired anode voltage when the current is what it should be and there is no DC current flowing through R_4. That is,

$$R_2 = \frac{V_B - V_{AK1} + V_{GK1}}{I_{A1}}$$

when the supply voltage is V_B. The choice to have no DC current flowing through R_4 is a rather arbitrary choice to simplify things, you can make a different choice if that works out better.

6. Choose an initial target value $R_{4,\,\mathrm{target}}$ for R_4, for example a few times smaller than the parallel connection of R_2 and the internal resistance of the first valve r_{i1}. A too small value of R_4 will make the nominally 75 µs time constant very sensitive to the transconductance of the second stage, while a too high value of R_4 will make it very sensitive to the internal resistance (and, hence, to the transconductance, as $r_i = \mu/g_m$) of the first stage.

7. Calculate C_2 assuming an ideal virtual ground in the second stage and round it of to a practical value, for example an E3, E6 or E12 value. That is,

$$C_2 = \left(\frac{75 \mu s}{R_2 // r_{i1} // R_4} - c_{ga1} \right)$$ rounded to the nearest standard value

8. Now calculate the R_4 that would belong to the rounded C_2 if Z_{st2} (the "virtual ground" impedance of the second stage) were 0. We will call this $R_{4\infty}$ as it is the value R_4 would have if the transconductance of the second stage were infinite.

$$R_{4\infty} = R_4 + Z_{st2} = \frac{1}{\frac{C_2 + c_{ga1}}{75 \mu s} - \frac{1}{r_{i1}} - \frac{1}{R_2}}$$

As soon as Z_{st2} is known, the real R_4 can be calculated from this.

9. Calculate

$$R_1 = \frac{1}{\frac{1}{47\ k\Omega} - g_{m1} \frac{c_{ga1}}{c_{ga1} + C_2}}$$

Second stage:

10. Choose a valve with sufficient voltage gain and transconductance.

11. Choose a high anode current for maximum transconductance.

12. Choose the same anode voltage as for stage 1. This is needed due to the rather arbitrary choice of trying to let 0 DC current flow through R_4.

Steps 11 and 12 together actually conflict with steps 2 and 5 if the same type of valve is used for both stages! This could be a reason not to make the arbitrary choice of not having DC current flow through R_4. Alternatively, one could connect an extra AC coupling capacitor in series with R_4, so anode DC voltage differences don't matter anymore.

13. Look up the corresponding grid-cathode voltage and dimension R_9 correspondingly:

$$R_9 = -\frac{V_{GK2}}{I_{A2}}$$

14. Make

$$R_8 = \frac{V_B - V_{AK2} + V_{GK2}}{I_{A2}}$$

15. Use the recommended grid leakage resistance value for R_5, usually 1 MΩ or so.

16. Choose a high value for R_{10}, 1 MΩ or so.

17. Calculate

$$R_{8C} = R_8 // r_{i2} // R_{10} // R_L$$

where R_L is the input impedance of whatever is driven by the output of the phono preamplifier.

Adding a cathode follower with its grid DC coupled to the anode of the second stage would eliminate R_L as well as R_{10} from the equation and make the RIAA correction more predictable - at the expense of an extra triode.

18. Choose the gain at 1 kHz that you would like, and choose an appropriate target value $R_{6,\text{ target}}$ for R_6. For example, if the target value for R_4 in step 6 was chosen as half the parallel value of R_2 and the internal resistance of the first valve r_{i1}, about 2/3 of the current of the first stage ends up in the second stage. The gain at 1 kHz is then roughly $\frac{2}{3} g_{m1} 0.9 R_6$, where the factor 0.9 is supposed to account for the effect of R_7 and finite loop gain. This is only a very rough approximation, but presumably the precise value of the gain doesn't matter.

19. Approximately calculate C_4 and round it to a convenient value, an E3, E6 or E12 value for example.

$$\left(R_6 - \frac{1}{g_{m2} - \frac{1}{R_7}}\right) C_4 = 318\ \mu\text{s} \quad \text{, hence} \quad C_4 = \frac{318\ \mu\text{s}}{R_6 - \frac{1}{g_{m2} - \frac{1}{R_7}}} \approx \frac{318\ \mu\text{s}}{R_6 - \frac{1}{g_{m2}}} \quad \text{, so take}$$

$$C_4 = \left(\frac{318\ \mu\text{s}}{R_{6,\text{target}} - \frac{1}{g_{m2}}}\right) \text{rounded to the nearest standard value}$$

20. Now that the actual C_4 is known, we can approximately calculate the real value of R_6:

$$\left(R_6 - \frac{1}{g_{m2} - \frac{1}{R_7}}\right) C_4 = 318\ \mu\text{s} \quad \text{, hence}$$

$$R_6 = \frac{318\ \mu\text{s}}{C_4} + \frac{1}{g_{m2} - \frac{1}{R_7}} \approx \frac{318\ \mu\text{s}}{C_4} + \frac{1}{g_{m2}}$$

21. Approximate R_{4B}:

$$R_{4B} = (R_4 + R_{2B}) \mathbin{//} R_5 \approx (R_{4\infty} + R_{2B}) \mathbin{//} R_5$$

22. The equation for the 3.18 ms time constant can be rearranged to become an expression for R_7. If I did the calculation properly:

$$R_7=\left(\frac{3.18\text{ ms}}{C_4}-R_6\right)\frac{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1}{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1+\frac{R_6}{R_{4B}}-\frac{3.18\text{ ms}}{R_{4B}C_4}}$$

23. Calculate

$$R_4=R_{4\infty}-Z_{st2}=R_{4\infty}-\frac{1}{\frac{1}{R_5}+\frac{g_{m2}R_{8C}+1}{\frac{R_6R_7}{R_6+R_7}+R_{8C}}}$$

24. You could do a new iteration by recalculating step 20 without neglecting the effect of R_7, recalculating step 21 without approximating R_4 with $R_{4\infty}$ and repeating steps 22 and 23.

4. Example

Suppose we want to make a RIAA amplifier with (per channel) two ECC83 triodes that doesn't need to drive any R_{10} or any load, for example because there is a third triode DC coupled to the anode of the second triode, and this third triode is used as a cathode follower.

The first ECC83 triode will be biased at 0.9 mA of anode current for minimum ITU-R 468- and A-weighted noise. If we target a zero DC current through R_4 and no AC coupling capacitor in series with it, and equal cathode voltages, we automatically end up with also biasing the second ECC83 triode at 0.9 mA.

The minimum required negative grid voltage for an ECC83 was originally specified as -1.3 V for 300 nA of grid current, but later revised to -0.9 V, see http://www.r-type.org/pdfs/ecc83.pdf pages 7 and 8. As 300 nA is on the high side, we will target -1.3 V. Using 1.5 kΩ for R_3 and R_9 will lead to -1.35 V. According to the graph on page 10, the corresponding anode voltage is about 165 V - with respect to the cathode, so 166.35 V with respect to ground when the grid is at 0 V (not that such small differences matter).

Assuming a 250 V supply, R_2 = (250 V - 166.35 V)/0.9 mA = 92.9444... kΩ. Nearest E96 value:

R_2 = 93.1 kΩ

Since the first and second stages are biased equally:

R_8 = 93.1 kΩ

Don't forget to check whether the resistors can handle 250 V!

At 0.9 mA, an ECC83 should have a transconductance of about 1.4537 mS and an internal resistance of about 68.79 kΩ, extrapolated from the data at 1.2 mA using Child's law. The number of digits has nothing to do with the actual accuracy.

The parallel value of R_2 and r_{i1} is therefore about 39.55988 kΩ. Aiming for an R_4 of the order of 20 kΩ, C_2 must be about 5.64426 nF (assuming an ideal virtual ground of the second stage and after correction for the 1.6 pF anode-to-grid capacitance). The nearest E12 value is 5.6 nF, so

C_2 = 5.6 nF

Hence,

$$R_{4\infty} = R_4 + Z_{st2} = \frac{1}{\frac{C_2 + c_{ga1}}{75\mu\text{s}} - \frac{1}{r_{i1}} - \frac{1}{R_2}} \approx 20238.87344 \ \ \Omega$$

and

$$R_1 = \frac{1}{\frac{1}{47\ \text{k}\Omega} - g_{m1}\frac{c_{ga1}}{c_{ga1} + C_2}} \approx 47935.48683 \ \ \Omega$$

Rounding this to the nearest E96 value results in

R_1 = 47.5 kΩ,

the exact same E96 value we would have found without correcting for the remaining Miller effect. The only difference is that we are rounding the resistance down rather than up.

Let's choose R_5 = 1 MΩ, even though the datasheet even allows twice that value.

As written at the beginning of this section, R_{10} and R_L are assumed to be non-existent (infinite values). Hence,

R_{8C} = 93.1 kΩ // 68.79 kΩ ≈ 39559.88017 Ω

Suppose we want a gain of about 100 at 1 kHz. $R_{6,\text{ target}}$ then has to be about 120 kΩ. Hence, C_4 has to be around 2.6653 nF, the nearest E12 value is 2.7 nF.

C_4 = 2.7 nF

$$R_6 = \frac{318\ \ \mu\text{s}}{C_4} + \frac{1}{g_{m2} - \frac{1}{R_7}} \approx \frac{318\ \ \mu\text{s}}{C_4} + \frac{1}{g_{m2}} \approx 118465.6776 \ \ \Omega$$

The nearest E96 value is

R_6 = 118 kΩ

Using

$$R_{4B} = (R_4 + R_{2B}) \,//\, R_5 \approx (R_{4\infty} + R_{2B}) \,//\, R_5$$

where R_{2B} is R_2 in parallel with the internal resistance of the first triode, one finds

R_{4B} ≈ 56424.631 Ω

and

$$R_7=\left(\frac{3.18\text{ ms}}{C_4}-R_6\right)\frac{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1}{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1+\frac{R_6}{R_{4B}}-\frac{3.18\text{ ms}}{R_{4B}C_4}}\approx 1.55214376\text{ M}\Omega$$

nearest E96 value R_7 = 1.54 MΩ,

and

$$R_4=R_{4\infty}-Z_{st2}=R_{4\infty}-\frac{1}{\frac{1}{R_5}+\frac{g_{m2}R_{8C}+1}{\frac{R_6R_7}{R_6+R_7}+R_{8C}}}\approx 17695.93916\ \Omega$$

nearest E96 value R_4 = 17.8 kΩ

Taking step 24 from section 3, that is, doing another iteration, the values change into:

$$R_6=\frac{318\ \mu\text{s}}{C_4}+\frac{1}{g_{m2}-\frac{1}{R_7}}\approx\frac{318\ \mu\text{s}}{2.7\text{ nF}}+\frac{1}{1.4537\text{ mS}-\frac{1}{1.54\text{ M}\Omega}}\approx 118465.985\ \Omega$$

which still becomes 118 kΩ when rounded to an E96 value.

$$R_{4B}=(R_4+R_{2B})\,//\,R_5\approx 54248.20938\ \Omega$$

where R_{2B} is still R_2 in parallel with the internal resistance of the first triode.

$$R_7=\left(\frac{3.18\text{ ms}}{C_4}-R_6\right)\frac{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1}{g_{m2}R_{8C}+\frac{R_{8C}}{R_{4B}}+1+\frac{R_6}{R_{4B}}-\frac{3.18\text{ ms}}{R_{4B}C_4}}\approx 1581254.49\ \Omega$$

After rounding to the nearest E96 value, the new R_7 becomes 1.58 MΩ.

$$R_4=R_{4\infty}-Z_{st2}=R_{4\infty}-\frac{1}{\frac{1}{R_5}+\frac{g_{m2}R_{8C}+1}{\frac{R_6R_7}{R_6+R_7}+R_{8C}}}\approx 17692.57504\ \Omega$$

Nearest E96 value: still R_4 = 17.8 kΩ.

That is, the only change is a one E96 step increase of R_7, to 1.58 MΩ.

All in all, we find these values:

ECC83, 250 V supply, no R_{10} or R_L

R_1 = 47.5 kΩ

R_2 = 93.1 kΩ

R_3 = 1.5 kΩ

R_4 = 17.8 kΩ

R_5 = 1 MΩ

R_6 = 118 kΩ

R_7 = 1.58 MΩ

R_8 = 93.1 kΩ

R_9 = 1.5 kΩ

C_2 = 5.6 nF

C_4 = 2.7 nF

The cathode decoupling capacitors C_1 and C_5 and coupling capacitor C_3 (and C_6, if there were one) have to be large enough to be neglected, whatever that means.

Checking the result with a pole-zero extraction program, assuming infinite decoupling and coupling capacitors:

```
DCgain =    1.031 E  3

Number of poles:  4                    Number of zeros:  3

  n     pole                             n     zero
        [rad/s]                                [rad/s]

  1  -975.408 E  6                       1   908.562 E  6
  2    -6.185 E  6                       2   902.870 E  6
  3  -310.233                            3    -3.157 E  3
  4   -13.467 E  3
```

The gain is about right and the RIAA correction poles and zero are within 1.4 % of their correct locations. The correct locations would have been -314.4654... rad/s and -13333.333... rad/s for the poles and -3144.654... rad/s for the zero, the actual locations are -310.233 rad/s, -13467 rad/s and -3157 rad/s. There are some extra poles and zeros that affect the response from about 1 MHz onward.

Assuming 47 μF cathode decoupling, so C_1 = 47 μF and C_5 = 47 μF, but C_3 still infinite:

```
DCgain =  320.947

Number of poles:  6                    Number of zeros:  5

  n     pole                              n     zero
       [rad/s]                                 [rad/s]

  1  -975.408 E  6                        1   908.562 E  6
  2    -6.185 E  6                        2   902.870 E  6
  3   -13.474 E  3                        3    -3.157 E  3
  4  -310.404                             4   -14.184
  5   -37.451                             5   -14.172
  6   -17.214
```

The reduced DC gain is due to the series feedback by the cathode resistors. There are some deep subsonic poles and zeros due to the cathode decoupling, but the RIAA correction poles and zero are hardly affected.

Also adding C_3, the lowest RIAA pole shifts to -307.845 rad/s (2.11 % too low) when C_3 = 100 nF and to -309.248 rad/s (1.66 % too low) when C_3 = 220 nF. The complete result with 100 nF is:

```
DCgain =   -1.125

Number of poles:  7                    Number of zeros:  6

  n     pole                              n     zero
       [rad/s]                                 [rad/s]

  1  -975.408 E  6                        1   908.562 E  6
  2    -6.186 E  6                        2   902.870 E  6
  3   -13.474 E  3                        3    -3.157 E  3
  4  -307.845                             4    13.953 E -3
  5   -37.931                             5   -14.184
  6   -15.745                             6   -14.181
  7    -4.336
```

5. Conclusion

The simpler the RIAA correction amplifier, the more complicated the calculations seem to get.