
2.1 General Design of Amp3

The two central Amps 3 & 4 are the heart of the whole amplifier arrangement. Therefore, I will start my explanations with these, followed by the output stage Amp5 and the two input stages Amps 1 & 2.

Before I entered into the here presented design of Fig. 2.1, I had some tests on the most useful balanced triode driven solution, however, always in conjunction with the placement of the RIAA network. In addition, the solution should be as low-noise as possible, thus, increasing the noise level of the preceding gain stage by not more than point B.2. of Chap. 1 would allow. On the other hand, it should have outstanding CMRR and easy balance trimming.

Based on the findings in the Differential Gain Stage (DIF) chapter of the 2nd edition of my *How to Gain Gain* book (HTGG-2)¹ I opted for a DIF input stage followed by a CF (cathode follower) output stage. The easy handling of a CCSCF gain stage (Common Cathode gain Stage CCS followed by a CF) led to the shown Fig. 2.1 configuration without RIAA networks. I used such a CCSCF as output stage in the triode driven Module 4 phono-amp of Engine I in TSOS-2.² To get a rather high CMRR the DIF stage's DC current comes from a solid-state current generator (a sink here), formed by two BJTs. It creates a very high dynamic resistance between t1 & t2 cathode and ground.

The input section is the DIF formed by a g_m -selected low-noise double-triode E88CC/6922 (E188CC/7308 work well too). In each triode system the anode current is equal and trimmed to 2 mA by P2 of the current sink T1 & T2 (480 mV between test points TP3 & TP4). Trimming of P5 optimizes CMRR further (calculated appr. 100 dB). It ensures equal signal levels at the cathodes of t4 & t3. Each of the following CF stages is powered by appr. 90 V/2 mA too. Here, instead of the shown

¹“How to Gain Gain”, 2nd ed., B. Vogel, (HTGG-2), Chap. 30.

²TSOS-2, Chap. 17.

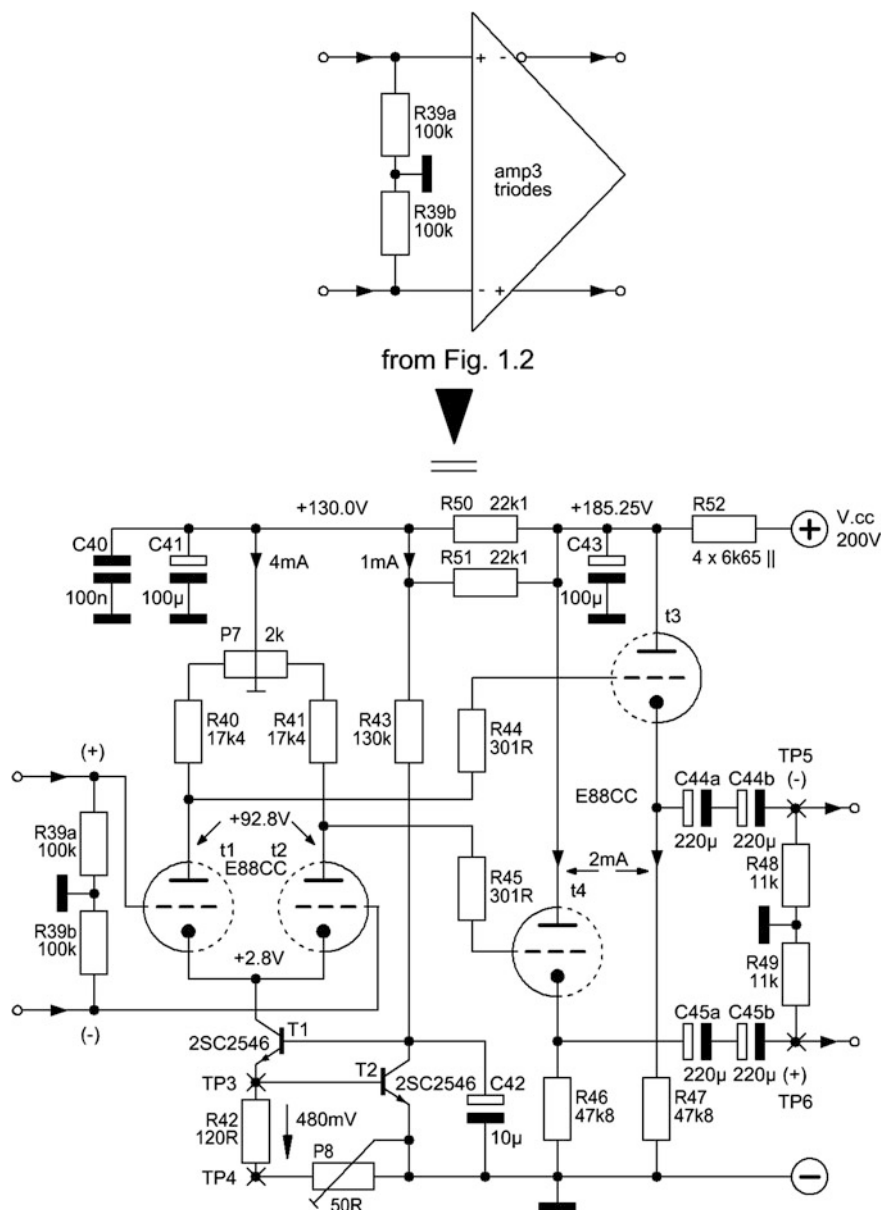


Fig. 2.1 Amp3 without RIAA networks

E88CC I also tried non-selected NOS 7308 s. Their noise level is rather low and their triode systems do not differ very much. Finally, I took the 7308 s.³

The gain G_{amp3} becomes measured appr. 16 and the whole design looks rather simple. In addition, with a perfect trim we can get a 1 kHz THD $\leq 0.010\%$ (stronger d2 than d3) and IMD always $< 0.010\%$ (I've measured 4 different 80 %/20 % frequency pairs). My Clio sinus generator offers a min. 1 kHz THD level of 0.002 % rounded⁴ through my un-balanced to balanced converter (see Footnote 4), strictly THD only and not THD + N! I could calculate the real 1 kHz THD with distortion spike level figures taken from the FFT diagram: 0.00159 % (more about distortions etc. see Chap. 12—Engine II Performance).

The gain stage fulfils the overload goal. I measured 46 V_{pp} before soft clipping. With an input signal level of 100 mV_{rms} + 20 dB overload margin = 1 V_{rms} and a gain of 16 we need a max. voltage swing of $16 \text{ V} * 2 * \sqrt{2} = 45.255 \text{ V}_{\text{pp}}$.

Based on the following considerations we can roughly check the extra-generated noise level of the sequence of Amp3 & Amp5: with input loaded by 20 Ω , Amp2 (its SN looks worse than the one of Amp1) alone generates a measured (_m) output referred non-equalized (_{ne}) $\text{SN}_{\text{ne.o.m}} = -73.1 \text{ dB ref. } 100 \text{ mV}/B_{20k}$, almost white noise. It includes a tiny amount of 1/f-noise.⁵ Multiplication by 10 (theoretically through a no-noise amp-stage) leads to a total output referred $\text{SN}_{\text{ne.o.tot}} = -73.1 \text{ dBV}$ at the engine's output. Now, after application of the B_{20k} RIAA function and A-weighting SN-improvement figure $\text{SN}_{\text{ar}} \approx -8 \text{ dB}$ ⁶ for purely white noise generating devices we obtain the guessed output referred A-weighted and equalized (_{ariaa}) $\text{SN}_{\text{ariaa.o}} = -81.1 \text{ dBV(A)}$. With the sequence of 20 Ω + Amp2 + Amp3 + Amp5 + Trafo at the output of the Engine I've measured $\text{SN}_{\text{ariaa.o.m}} = -79.9 \text{ dBV(A)}$ for the left channel and -80.2 dBV(A) for the right one. With that, the goal of an input referred $\text{SN}_{\text{ariaa.i}} = -79.0 \text{ dBV(A)}$ won't get into trouble.

Nevertheless, via shorted external input, the output referred SN of the amp sequence Amp3 + Amp5 + Trafo becomes measured (calculated) -99.0 dBV(A) (-100.2 dBV(A)). Figure 2.2⁷ shows the curve of the noise voltage density at the output of the before given sequence, based on data-sheet data. It also shows a kind of 1/f-noise characteristic. It is generated by two sources:

- (a) by an assumed 1/f-noise corner frequency of $f_{\text{c.e1.2}} = 1 \text{ kHz}$ of the DIF input triodes (high influence on the overall noise voltage) and an $f_{\text{c.e3.4}} = 10 \text{ kHz}$ of the two output CFs (rather low influence on the overall noise voltage), and

³I deeply have to thank my friend Klaus Burosch (www.burosch.de) for his courteous support concerning his huge collection of NOS and brand new valves. All used (and many more) valves had to pass the test arrangement I've presented in Jan Didden's Linear Audio Vol. 4 "The Glowing NoiseMaker—on the demystification of triode noise" or in HTGG-2, Sect. 2.3.

⁴Details see Chap. 15.

⁵Additionally see my remarks on Amp2's SN in Chapter 10.

⁶TSOS-2, Chapter 15, TSOS-1 Chapter 6.

⁷Details see next Chapter and MCD-WS 3.1.

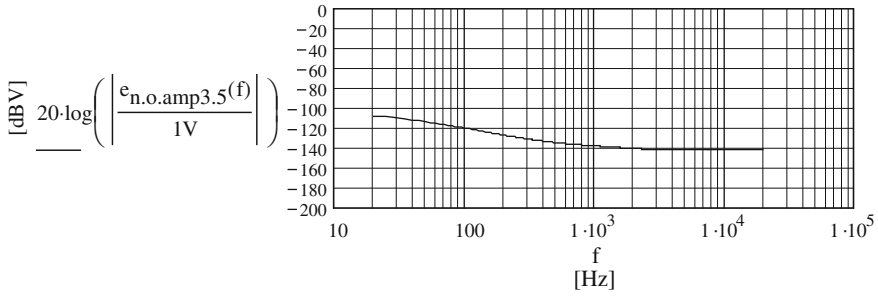


Fig. 2.2 Output noise voltage density of the amp sequence Amp3 + Amp5 + Trafo with input shorted

- (b) by the RIAA network effect of the 318 μ s/3180 μ s network at the output of Amp3. The 75 μ s input network has practically no effect on the Amp3 noise generation. It only filters the incoming noise voltage from preceding gain stages.

To calculate the component values for the T1/T3 RIAA network we need the differential o/p resistance $R_{o.cf,dif}(f)$ of t3 & t4 (see (1.6)–(1.8) in the previous chapter). Because of C44a–C45b it is frequency dependent and in consideration of R48 & R49 it is the sum of the equal o/p resistances $R_{o.cf3}(f) + R_{o.cf4}(f)$. The relevant equations look as follows:

$$\begin{aligned}
 R_{o.cf3}(f) &= R_{o.cf4}(f) \\
 &= \left[\frac{1}{R48} + \frac{1}{R_{o.cf3} + \left(2j\pi f \frac{C44a}{2}\right)^{-1}} \right]^{-1}
 \end{aligned} \tag{2.1}$$

$$R_{o.cf,dif}(f) = R_{o.cf3}(f) + R_{o.cf4}(f) \tag{2.2}$$

$$\begin{aligned}
 R_{o.cf3} &= \frac{r_{a3}R47}{r_{a3} + (1 + \mu_3)R47} \\
 &= R_{o.cf4}
 \end{aligned} \tag{2.3}$$

According to the goals C44a–C45b must be chosen of a size that should not hurt a flat frequency and phase response in B_{20k}. Then, with $r_{a3} = r_{a4} = 8.836$ k Ω , $g_{m3} = g_{m4} = 3.5$ mS, and $\mu_3 = \mu_4 = 29$ we'll get $R_{o.cf,dif} = 549.2$ Ω . I've chosen Panasonic FC 63 V types. With them, the deviation from the flatness becomes a calculated -0.025 dB/ $+0.2^\circ$ at 20 Hz only. Figure 2.3 shows the calculated deviation from the exact RIAA transfer if we would consider the RIAA networks. The measured frequency and phase response will be given in Chap. 12.

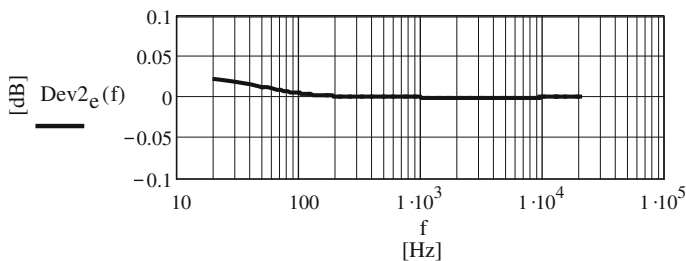


Fig. 2.3 Calculated deviation from the exact RIAA transfer

Figure 2.1 also shows the constant current sink around BJTs T1 & T2. The actual noise voltage of this CCsi is of minor importance. What hits the DIF most is the noise current mainly produced by T1's collector current. Multiplied by the cathode input resistance of the DIF we have an enormous noise voltage that is amplified by the here effective grounded grid gain stages(CGS) formed by t1 & t2, in this case leading to 100 % correlated noise voltages of equal amplitude at the anodes of t1 & t2. Hence, at the differential output of Amp3 we find the CCsi generated noise voltage with a doubled level! To suppress it we need a following Amp5 with rather high CMRR. Chapter 6 gives the details of Amp5.

2.2 Gain and Noise Calculations

The Mathcad worksheet (MCD WS-3.1) of the next chapter gives all the details of a rather extensive calculation course. All results are based on data-sheet data. I've also gone through the calculation with actual data. Selected low-noise triodes should have very low 1/f-noise corner frequencies and far better (higher) g_m -values than the ones of the data-sheets. Fortunately, they do not differ very much from the ones gained by application of data-sheet data together with the assumed data for the 1/f-noise corner frequency. I guess it is clear that higher 1/f-noise corner frequencies will automatically lead to worsened SNs.

The complete calculation of the gain and noise production of a DIF can easily be studied in HTGG-2, Chap. 30. However, for a better understanding I will repeat the equivalent circuit and the main equations here.

2.2.1 Gain of a DIF Followed by Two CFs

The DIF's idle gain $G_{0,dif}$:

$$G_{0,dif} = G_{0,t1.2} = -\mu_1 \frac{R40 + 0.5P7}{r_{a1} + R40 + 0.5P7} \quad (2.4)$$

The CF's frequency dependent gain with output load $R_L(f)$ (because of its tiny influence here the impedance of Fig. 1.2's C12 is set to 0 Ω):

$$G_{cf4}(f) = \mu_4 \frac{R47}{r_{a4} + (1 + \mu_4)R47 + \frac{r_{a4}R47}{R_{L,t4}(f)}} \quad (2.5)$$

$$R_{L,t4}(f) = \frac{1}{2j\pi f 0.5C44a} + \left[\frac{1}{R48} + \frac{1}{R34 + ([0.5R35]^{-1} + R36^{-1})^{-1}} \right]^{-1} \quad (2.6)$$

$$= R_{L,t3}(f)$$

$G_{0,dif}$ is the DIF's idle gain because its anode has an infinite load by the following t3/t4 grids. $R_{L,t4}(f)$ is the frequency dependent load at the cathode of t4. The same applies to $R_{L,t3}(f)$.

$$G_{cf3}(f) = G_{cf4}(f) \quad (2.7)$$

$$G_{cf}(f) = G_{cf3}(f)G_{cf4}(f) \quad (2.8)$$

\Rightarrow The frequency dependent DIFCF gain $G_{difcf}(f)$ thus becomes:

$$G_{difcf}(f) = G_{0,dif}G_{cf}(f) \quad (2.9)$$

\Rightarrow The balanced gain $G_{op1,2}$ of the two op-amps OPs 1 & 2 is 1. Hence, the Amp3 gain $G_{amp3}(f)$ without RIAA transfer becomes:

$$G_{amp3}(f) = G_{difcf}(f)G_{op1,2} \quad (2.10)$$

2.2.2 RIAA Transfer Function

From Figs. 1.4 and 1.5 and (2.1)–(2.3) we can derive the frequency dependent and RIAA transfer loaded gains $G_{T2}(f)$ and $G_{T1,3}(f)$ of the Amp3 input and output networks as follows:

$$G_{T2}(f) = \frac{M}{M + R25 + R26 + P1 + 2\left(R_{o,op1}^{-1} + R23^{-1}\right)^{-1}} \quad (2.11)$$

$$M = \left[2j\pi f(C9 + C_{i,dif}) + (R39a + RR39b)^{-1}\right]^{-1}$$

$$G_{T1.3}(f) = \frac{\left[\left([2j\pi f C12]^{-1} + R35 \right)^{-1} + R_{L.dif}^{-1} \right]^{-1}}{\left[\left([2j\pi f C12]^{-1} + R35 \right)^{-1} + R_{L.dif}^{-1} \right]^{-1} + R_{T1.eff}(f)} \quad (2.12)$$

$$R_{L.dif} = \left([R1_{amp5} + R2_{amp5}]^{-1} + [R36 + R37]^{-1} \right)^{-1} \quad (2.13)$$

$$R_{T1.eff}(f) = R_{o.cf.dif}(f) + P3 + R33 + R34 \quad (2.14)$$

⇒ With C9 and C12 carefully selected according to Figs. 1.4 and 1.5 the transfer function $T_{amp3}(f)$ of the whole Amp3, including RIAA transfer function, thus becomes:

$$T_{amp3}(f) = G_{amp3}(f)G_{T2}(f)G_{T1.3}(f) \quad (2.15)$$

2.2.3 Noise and SN Calculations According to Fig. 1.2

The calculation of the noise voltage of the DIFCF alone makes no sense, as long as there are influential factors at its input (OPs 1 & 2 + $T2(f)$) and at its output ($T1(f) + T3(f) + \text{Amp5}$). All together, they generate a noise voltage that can be measured at the output of Amp5, and thus be compared with the calculated results. The calculation course follows the mathematical course given in MCD-WS 3.1, “6. Noise and SN calculations”. However, here comes the short version. It tackles the major factors.

To calculate the output noise voltage density $e_{n.o.amp3.5}(f)$ at the o/p of Amp5 and with Amp3 input shorted the rather complex looking equation looks as follows:

$$e_{n.o.amp3.5}(f) = G_{amp5} \sqrt{\left[\left[\begin{array}{l} \left(\begin{array}{l} e_{n.o.op1.2}(f)^2 G_{T2}(f)^2 \\ + e_{n.Z.T2}(f)^2 \\ + 2e_{n.rN1}(f)^2 \end{array} \right) |G_{0.dif}|^2 \\ + 2e_{n.Rgg3}^2 + 2e_{n.Ra.eff}(f)^2 \\ + 2e_{n.rN3}(f)^2 \\ + 2e_{n.Rc.eff}(f)^2 \end{array} \right] G_{cf}(f)^2 \right] G_{T1.3}(f)^2 + \left(\frac{2e_{n.ccsi} G_{cgs.1} G_{cf}(f) G_{T1.3}(f)}{CMRR_{amp5}} \right)^2 + e_{n.Z.T1.3}(f)^2 + i_{n.i.amp5}^2 Z_{T1.3}(f)^2 + e_{n.i.amp5}^2 \right]} \quad (2.16)$$

$$G_{\text{cgs.1}} = (1 + \mu_1) \frac{R40 + 0.5P7}{r_{a1} + R40 + 0.5P7} \quad (2.17)$$

$$= G_{\text{cgs.2}}$$

According to (2.16), Figs. 1.2 and 2.1 it includes the following 100 % un-correlated noise sources:

• OP1 & OP2 (frequency dependent = fd):	$e_{n.o.op1.2}(f)$
• T2(f) network (fd):	$e_{n.Z.T2}(f)$
• t1 & t2 noise (fd) ^a :	$e_{n.rN1}(f) = e_{n.rN2}(f)$
• $R_{gg3} = R44 + R45$:	$e_{n.Rgg3}$
• $R_a = R40 + R41$ (incl. fd excess noise):	$e_{n.Ra.eff}(f)$
• t3 & t4 noise (fd) ^b :	$e_{n.rN3}(f) = e_{n.rN4}(f)$
• $R_c = R46 + R47$ (incl. fd excess noise):	$e_{n.Rc.eff}(f)$
• T1(f) & T3(f) network (fd):	$e_{n.Z.T1.3}(f) \text{ \& } Z_{T1.3}(f)$
• Amp5 i/p noise current	$i_{n.i.amp5}$ (no 1/f-noise!)
• Amp5 input referred noise voltage:	$e_{n.i.amp5}$ (=average value in B_{20k})

^aI've chosen a low-noise double-triode here that has, in both systems, a 1/f-noise corner frequency of 1 kHz; a change to 10 kHz would worsen the calculated output referred SN of -100.2 dBV(A) by appr. 3.5 dB, a change to 100 Hz would improve SNs by appr. 0.6 dB

^bFor calculation purposes, I've chosen a noisy double-triode here that has in both systems a 1/f-noise corner frequency of 10 kHz; a change to 1 kHz would improve the result of (2.16) by 0.014 dB only

Note: By integration over B_{20k} and division by $\sqrt{B_{20k}}$ the frequency dependency of the input referred noise voltage density of Amp5 (see respective MCD-WS in Chap. 7) can be turned into one single average density value. Hence, and in other words, we gain the rms value of the noise voltage in B_{20k} and after division by $\sqrt{B_{20k}}$ we'll get the average density value, however, guilty in B_{20k} only!

- Generated by the noise current of the constant current sink CCs_i and its BJTs T1 & T2 and multiplied by the equal gains $G_{\text{cgs.1}}$ of the Common Grid Stages (CGS) t1 & t2 the 100 % correlated noise voltage $e_{n.ccs_i}$ is damped by the CMRR of Amp5 (see Chaps. 6 and 7). Because there is 100 % un-correlation between this term and all the other ones, it is integrated into (2.16) too.

I must point out that some terms in (2.16) do not add significant values to the total sum underneath the root. Nevertheless, I keep them for universal usage with other than the chosen components.

The resulting noise voltage density multiplied by the A-weighting function, referenced to 1 V_{rms} nominal signal output voltage, and integrated over the bandwidth of B_{20k} , will lead to the A-weighted output referred $SN_{a.o.amp3.5}$ in B_{20k} , expressed in dBV(A).

With the exception of the DIF, the detailed calculation approaches of the different terms in (2.16) can completely be studied in TSOS-2. TSOS-1 is not a help at all because it doesn't cover the triode math approaches.

2.2.4 A Look into the Content of MCD-WS 3.1

MCD-WS 3.1 shows some additional interesting results:

- Very important for external amplifiers Point 6.5.2 shows the calculation of the Amp3 CMRR.
- Point 7. covers the math of an extremely low-noise input load of the Fig. 2.1 arrangement with Amp3. Here, I've chosen a pre-amp with a gain of 200, an i/p referred noise voltage density of 0.2 nV/rtHz and noise current density of only 2.4 pA/rtHz. The i/p load is 20 Ω . Now we can compare the A-weighted and RIAA equalized SN result ($=-82.523$ dBV(A)) with the one of Point 8.
- Point 8. covers the math of the Point 7. low-noise Amp1, followed by a no-noise arrangement à la Fig. 2.1. The SN result becomes -82.582 dBV(A).
- Hence, the difference is appr. 0.06 dB only. It is nothing else but the Noise Figure. In other words: a further chase for extremely low-noise solutions makes no sense for input loads ≥ 20 Ω . We will see later on in Chap. 10 what it will mean for input loads < 20 Ω .
- Point 9. and 10. show calculations of the Noise Figure NF of the amp chain Amp3 + Amp5, fed by a lowest-noise input amp: 9. for MC and 10. for MM cartridge purposes. These NFs are all < 0.1 dB. Hence, together with its input and output loads the noise impact of the here presented Amp3 is completely ignorable.
- Point 11. gives up the shorted input and replaces the shortage by an output resistance of a preceding gain stage, here 1 k Ω . Because of the 75 μ s lp at the input, the noise impact becomes marginal too.

Balanced Phono-Amps

An Extension to the 'The Sound of Silence' Editions

Vogel, B.

2016, XXXVII, 424 p. 281 illus., 12 illus. in color.,

Hardcover

ISBN: 978-3-319-18523-1