

# Some variants of Elvee's single op-amp filter

Version 1, Marcel van de Gevel, 11 February 2023

## 1. Introduction

In the thread <https://www.diyaudio.com/community/threads/any-idea-to-improve-this-electronic-filter.395321/> Elvee looks for ways to filter a -5 kV, low current supply without needing very large series resistors or very large high-voltage capacitors. His filter configuration is shown in Figure 1.

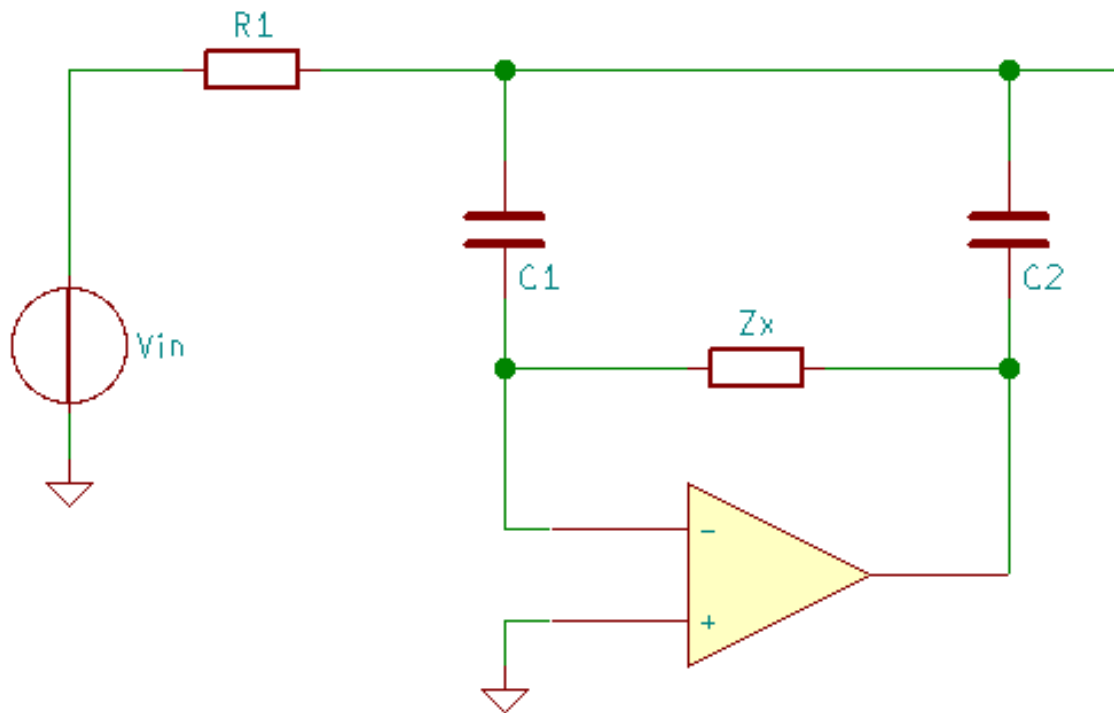


Figure 1: Elvee's single-op-amp filter configuration. In the original,  $Z_x$  is a 22 M $\Omega$  resistor,  $R_1 = 1$  M $\Omega$ ,  $C_1 = C_2 = 1.5$  nF.

A problem with the circuit is its ringing when the feedback resistance is much greater than  $R_1$ .

Later in the thread, he explains that the circuit should ideally mimic a first-order low-pass filter with a capacitance of about 300 nF, which is 200 times as large as each of the two filter capacitors.

However, to get exactly that behaviour (exact when all components including the op-amp behave ideally),  $Z_x$  should be a capacitor of  $1.5 \text{ nF} / 198 = 7.57575757 \dots \text{ pF}$ . The voltage variations across  $C_2$  are then 199 times as large as those across  $C_1$ , so the total capacitance is amplified by  $199 + 1 = 200$ . This leads to the impractical circuit of Figure 2; impractical, because it won't bias properly.

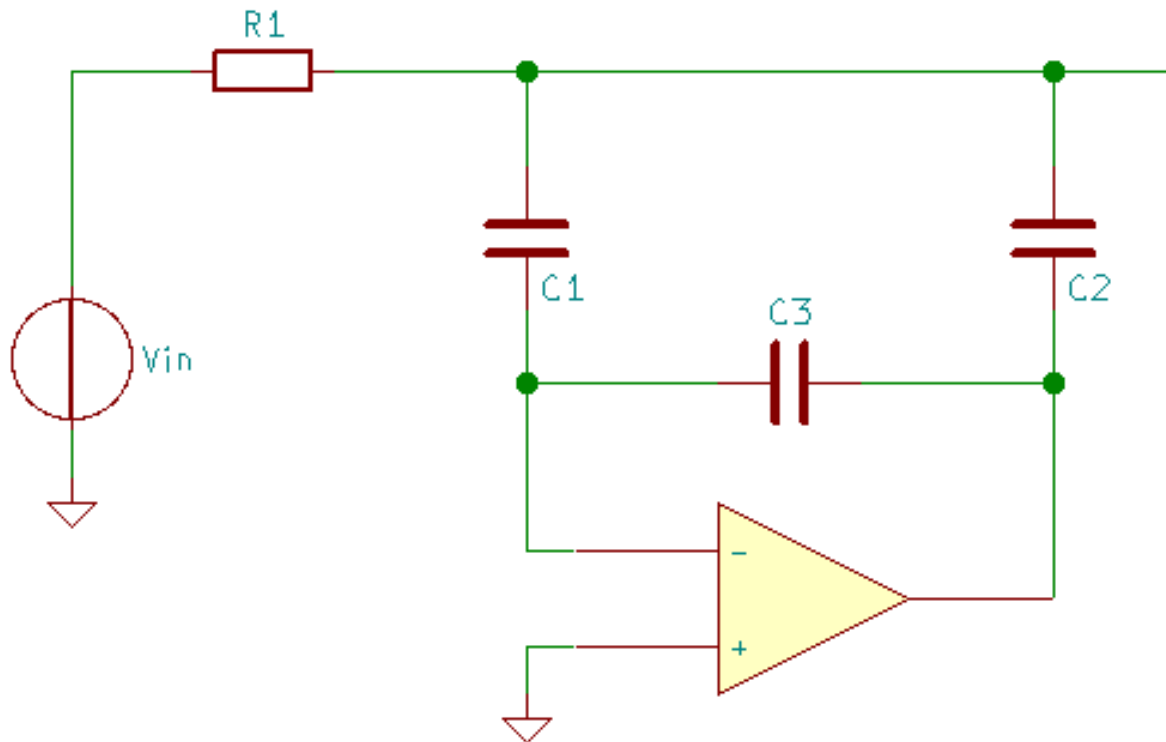


Figure 2: Impractical version that would behave as a first-order low-pass with increased capacitance, if it were biased correctly.

As a first attempt at a solution to the biasing issue, one could connect a resistor in parallel with the feedback capacitor. However, the feedback capacitor is 7.5757575757... pF and the time constant of the intended 300 nF capacitance with the 1 M $\Omega$  filter resistor is 300 ms. If the circuit is to behave like 300 nF over a time scale of the order of 300 ms, the parallel resistor has to be so large that its time constant with 7.5757575757... pF is well above 300 ms, corresponding to a resistance greater than 39.6 G $\Omega$ . Instead of using a resistor of that value, one could also use a T-network having the same voltage-to-current transfer; for example, a voltage divider that attenuates more than 1800 times followed by a 22 M $\Omega$  resistor. Needless to say, offset voltages will then be amplified more than 1800 times and a few hundreds of picoamperes of bias current will clip the op-amp output.

To solve this, one could connect some circuit that behaves inductively across the capacitor - not necessarily a real inductor (which would need to have an impractically large inductance), but something that has a relatively large voltage-to-current transfer at low frequencies that drops with a first-order slope at higher frequencies. For example, an RCR T-network. However, something inductive in parallel with something capacitive can lead to ringing/peaking again. In fact, when you look at the voltage-to-current transfer of the feedback path via  $C_2$ ,  $R_1$  and  $C_1$ , it initially increases with the square of the frequency. It therefore has FDNR-like behaviour, which further aggravates ringing and oscillation issues.

I think you can solve this by combining an FDNR-like, a capacitive, a resistive and an inductive

voltage to current transfer in the feedback, see Figure 3. I've called the proportionality constant of the admittance of the FDNR  $D$ , I haven't a clue what it is usually called.

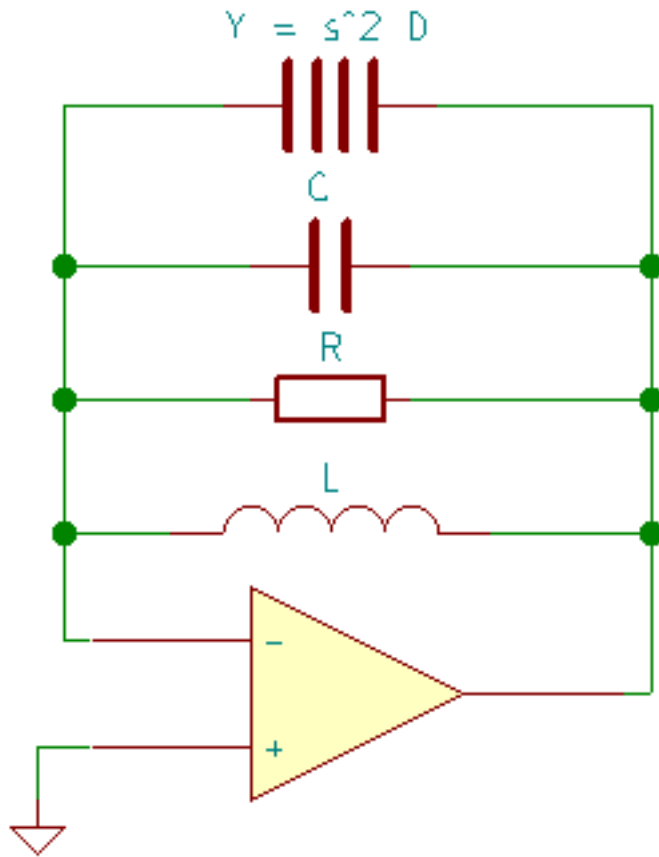


Figure 3: Op-amp with FDNR, capacitive, resistive and inductive feedback

When the op-amp is ideal, the current-to-voltage transfer of the whole circuit is the reciprocal of the voltage-to-current transfer of its feedback network. The zeros of the voltage-to-current transfer of the feedback network are then the poles of the whole circuit.

The voltage-to-current transfer of the feedback network is simply the sum of the admittances of the four branches:

$$Y = \frac{1}{sL} + \frac{1}{R} + sC + s^2 D = \frac{1}{sL} \left( 1 + s \frac{L}{R} + s^2 LC + s^3 LD \right)$$

The term between parenthesis is the characteristic polynomial of the whole circuit. To get Butterworth pole positions,

$$\frac{L}{R} = 2\sqrt[3]{LD}$$

$$LC = 2(\sqrt[3]{LD})^2$$

$D$  is known, as  $D = C_2 R_1 C_1$ , and  $C$  is the desired feedback capacitance (7.575757... pF). For Butterworth pole positions, one can now derive that

$$LC = 2L^{2/3} D^{2/3} \rightarrow L^{1/3} = \frac{2D^{2/3}}{C} \rightarrow L = \frac{8D^2}{C^3}$$

and

$$R = \frac{L}{2\sqrt[3]{LD}}$$

The resistive and inductive terms could be realized with a common T-network, as shown in Figure 4.

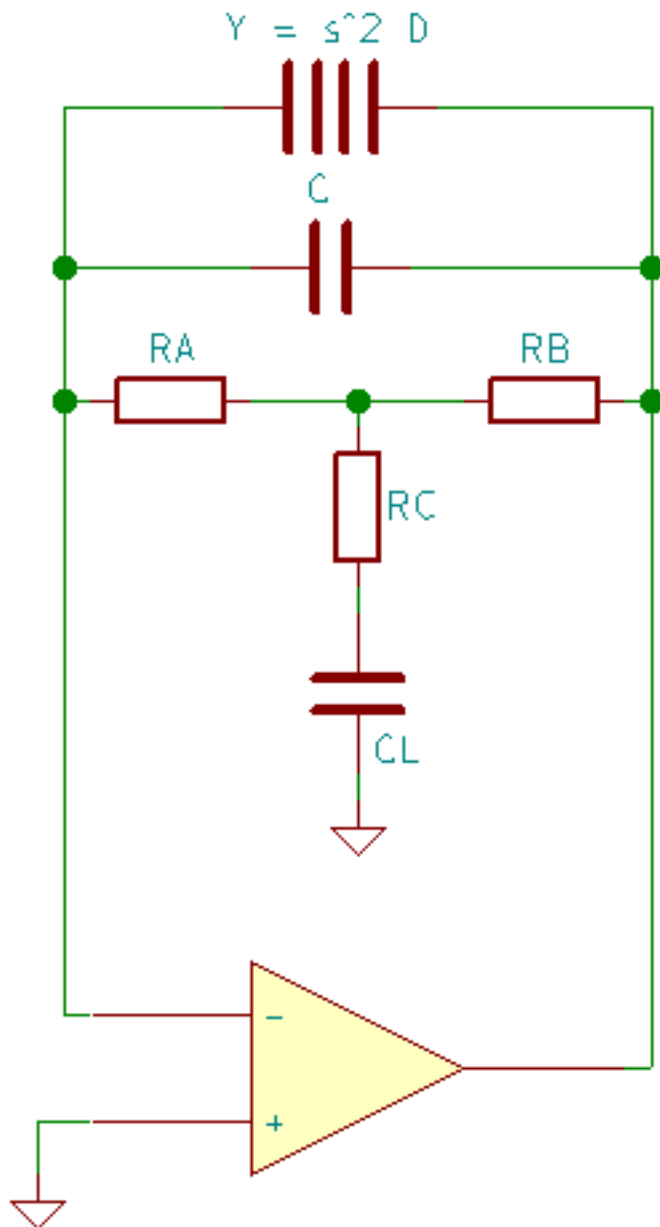


Figure 4: Practical implementation of the resistive and inductive branches. As mentioned in the text, the FNDR models  $C_2$ ,  $R_1$  and  $C_1$ .