EIC

Integrerede kredsløb

Exercise 3 - NOISE, calculattion, simulation and lab measurement

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1 2.1 Amplifier with great amplification

1.1 theory

An amplifier is given with, in figure: 1, it is a non-inverting amplifier with a gain of a 1001. The upper cutoff frequency is to be determined by the op amp itself but the lower cutoff is to be determined by the RC-circuit on the output of the amplifier.

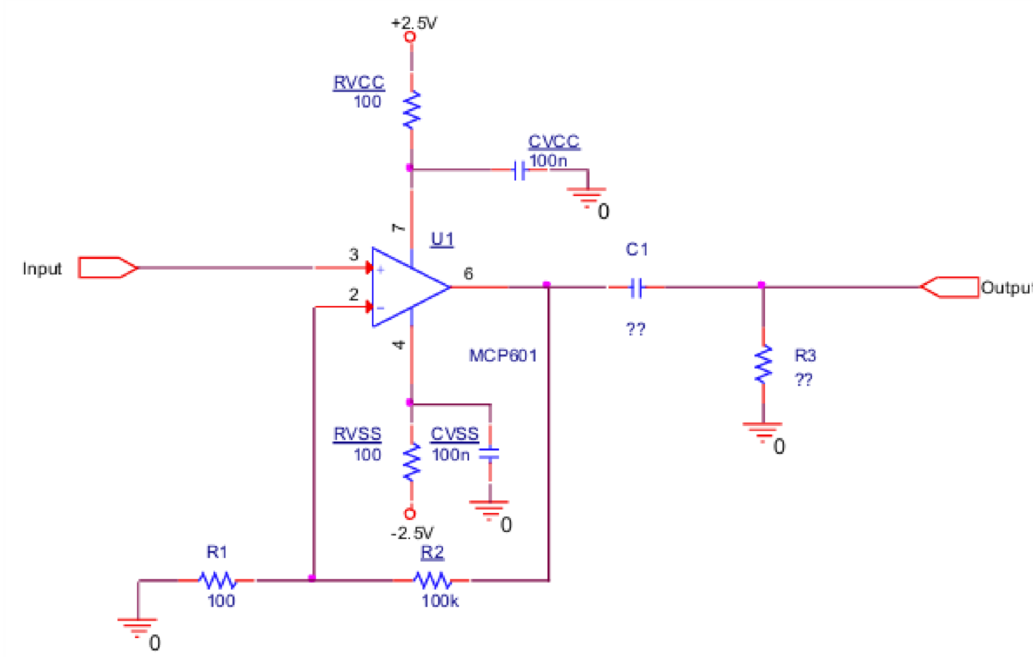


Figure 1: non-inverting amplifier with 1001 times gain

Since the filter is supposed to be designed for the human ear, the cutoff frequency should be somewhere arund the 20Hz range. Choosing a suitable standard value capacitor, say 47nF, we can solve for the Required resistance

$$R_3 = \frac{1}{2\pi C_1 R_3} = 20 \text{ Hz} \Rightarrow \frac{1}{1880 \cdot nF \cdot Hz} = 169 \text{ k}\Omega \tag{1}$$

we choose 180kΩ, this yields a cut-off:

$$f_c = \frac{1}{2\pi C_1 R_3} = 18.8 \text{ Hz} \tag{2}$$

Next is to determine the cut-off frequency of the operation amplifier. This is determined by the system gain and the *open-loop gain, phase vs frequency* which is figure: 2-21 in the MCP601 datasheet, this is also known as the *gain-bandwidth product (GBP)*. Which in this case is $GBP = 2.8$ MHz

As stated in the exercise, the gain of the system is 1001, this is determined by the feedback network:

$$A_N = 1 + \frac{R2}{R1} \Rightarrow 1 + \frac{100\text{k}\Omega}{100\Omega} = 1001 \tag{3}$$

thus the bandwidth of the amplifier is:

$$f_B = \frac{GBP}{A_N} \Rightarrow \frac{2.8 \text{ MHz}}{1001} = 2.8 \text{ KHz} \tag{4}$$

Thus this circuit is not suited as an audio amplifier, due to the fact that the cutoff is an order of magnitude below the upper human hearing range of 20KHz

To calculate the noise of the amplifier, we have to identify individual contributors to this nois These are *resistor noise*,*current noise* and the *voltage noise*. To calculate this, some of the factors are stated in the datasheet

From the datasheet, the voltage noise is $e_{nw} = 21\,\frac{\mathrm{nV}}{\sqrt{\mathrm{Hz}}}$ and the current noise is $i_{nw} = 0.6\,\frac{\mathrm{fA}}{\sqrt{\mathrm{Hz}}}$, these values are both at $f_{CE} = f_{CI} = 1\,\mathrm{KHz}$.

From the bandwidth defined by the output filter and the op amp itself, it is possible to create a brick-wall equivalent. which acts as an analogy for the white noise, when we pass $\frac{1}{f}$ noise through a first-order filter.

We create this equivalent by having two frequencies which the noise is "contained" within

$$f_L = f_C \cdot \frac{\pi}{2} = 29.5\ \mathrm{Hz} \tag{5}$$

$$f_H = f_B \cdot \frac{\pi}{2} = 4.4\ \mathrm{KHz} \tag{6}$$

The bandwidth provided by these two frequencies are also known as *equivalent noise bandwidth*
The noise voltage and current can then be calculated as

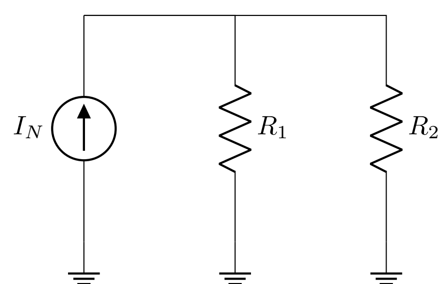
$$E_N = e_{nw}\sqrt{f_c e \ln\left(\frac{f_H}{f_L}\right) + f_H - f_L} = 0.061\mathrm{mV} \tag{7}$$

$$I_N = u_{nw}\sqrt{f_c e \ln\left(\frac{f_H}{f_L}\right) + f_H - f_L} = 1.26\mathrm{pA} \tag{8}$$

the final op amp voltage noise then gets amplified by the noise gain

$$V_{outN_V} = A_N \cdot E_N = 61\mathrm{mV} \tag{9}$$

The current noise, is the current that is entering the inverting input, from a loop, using superposition, the voltage source in the output of the op-amp gets grounded, meaning that the current source essentially sees two resistances in parallel



The equivalent resistance $R_{EQ} = R1||R2 \Rightarrow (\frac{1}{100\cdot 10^3\Omega} + \frac{1}{100\Omega})^{-1} = 99.9\Omega$.
voltage from the current noise then becomes:

$$V_{outN_I} = I_N \cdot R_{EQ} \cdot A_N = 0.126\mu\mathrm{V} \tag{10}$$

The final contribution is the thermal- or *johnsonnoise* from the feedback network and from the resistor in the filter:

$$V_{out_{Req}} = \sqrt{4kt \cdot R_{EQ} \cdot (F_H - F_L)} \cdot A_N = 2.701\mathrm{mV} \tag{11}$$

$$V_{out_{Rfilter}} = \sqrt{4kt \cdot 180\mathrm{k}\Omega \cdot (F_L - 0\mathrm{Hz})} = 0.306\mu\mathrm{V} \tag{12}$$

The total RMS voltage then becomes the sum of all the sources:

$$V_{TotRMS} = \sqrt{V^2_{outN_V} + V^2_{outN_I} + V^2_{out_{Req}} + V^2_{out_{Rfilter}}} = 61.02\mathrm{mV} \tag{13}$$

1.2 simulation

1.3 realisation

2 2.2 Thermal noise from resistor

2.1 theory

Reusing the amplifier from section 2.1, it is desired to see the thermal noise effect from a large resistor on the input of the amplifier.

The circuit is:

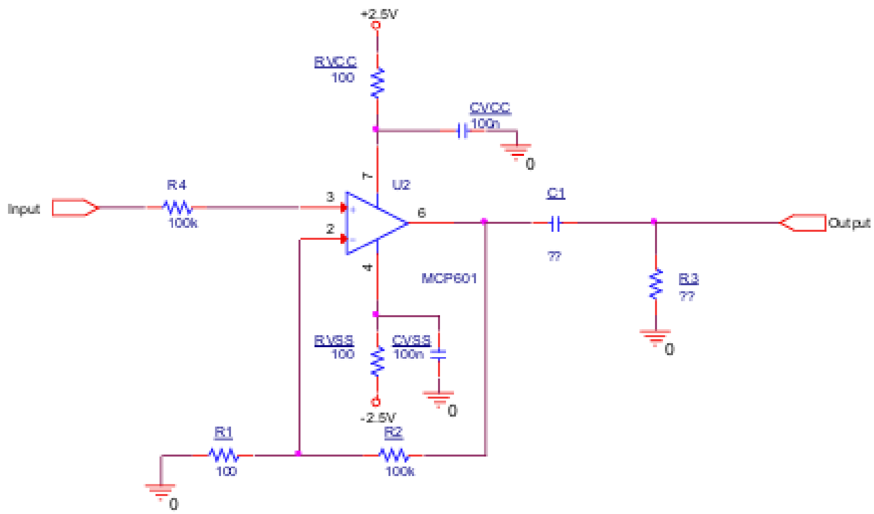


Figure 2: same non-noninverting amplifier from previous section, this time with a large input resistance, which acts a generator resistance

Due to the fact that a generator resistance now is present at the non-inverting input, this means that the thermal- and currentnoise terms generated here, must be added to the sum of noise sources.

As for the current noise, the bandwdith is still determined by the output of the filter and the limitation of the op amp itself.

$$V_{out_{NnonInv100k}} = I_N \cdot 100\text{k}\Omega \cdot A_N = 0.126\text{mV} \quad (14)$$
$$V_{out_{NnonInv200k}} = I_N \cdot 200\text{k}\Omega \cdot A_N = 0.252\text{mV} \quad (15)$$

And the thermal noise:

$$V_{out_{NonInvThermal100k}} = \sqrt{4kt \cdot 100\text{k}\Omega \cdot (F_H - F_L)} \cdot A_N = 0.085\text{mV} \quad (16)$$
$$V_{out_{NonInvThermal200k}} = \sqrt{4kt \cdot 200\text{k}\Omega \cdot (F_H - F_L)} \cdot A_N = 0.121\text{mV} \quad (17)$$

The total RMS voltage with generator resistance term added

$$V_{TotRMS} = \sqrt{V^2_{outN_V} + V^2_{outN_I} + V^2_{out_{Req}} + V^2_{out_{Rfilter}} + V^2_{out_{NnonInv100k}} + V^2_{out_{NonInvThermal100k}}} = 105\text{mV} \quad (18)$$
$$V_{TotRMS} = \sqrt{V^2_{outN_V} + V^2_{outN_I} + V^2_{out_{Req}} + V^2_{out_{Rfilter}} + V^2_{out_{NnonInv200k}} + V^2_{out_{NonInvThermal200k}}} = 135.375\text{mV} \quad (19)$$

2.2 simulation

2.3 realisation

3 2.3 circuit illustrating the effect of shotnoise(diode noise)

3.1 theory

Optical receiver circuits, use transimpedance amplifiers to make a I/V conversion, however this is prone to noise from the current source and parasitic elements of the diode, such as parasitic capaticance and resistance.

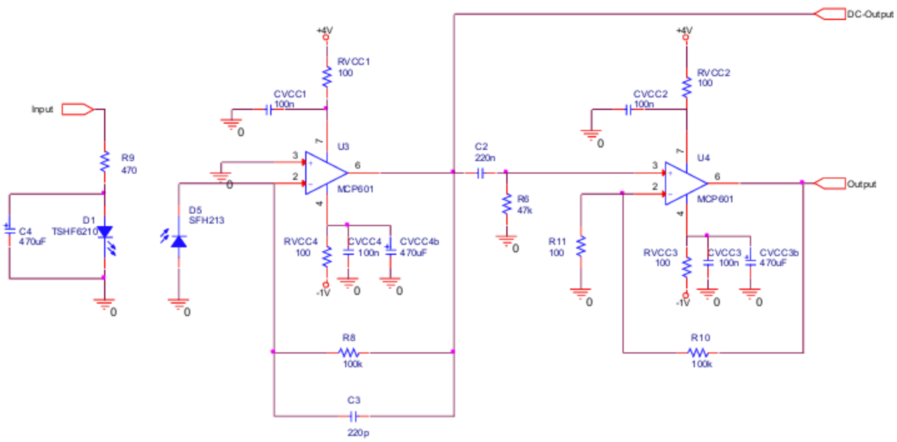


Figure 3: photo detector circuit, a transimpedance amplifier, filtered into a noninverting amplifier with a gain of 1001

The diode in question is a SFH213 and accoring to its datasheet is has a capaticance of $C_D = 11\text{pF}$ The resistance of the diode is not given in the datasheet, however it is often multiple orders of magnitude larger than the feedback resistance, so $R_D = 1\text{G}\Omega$ is chosen.
From the circuit we also have a feedback network $R8 = 100\text{k}\Omega$ and $C3 = 220\text{pF}$ We can then create expressions for the two impedances $Z1$ and $Z2$ which will determine the bandwidth of the amplifier

$$Z1(j\omega) = \frac{R1}{1 + j\omega R1C1} \tag{20}$$

$$Z2(j\omega) = \frac{R2}{1 + j\omega R2C2} \tag{21}$$

To determine the DC-gain of the amplifier, which is defined as $A_N = \frac{DC_{output}}{I_{detector}}$ relating voltage to current, the gain becomes $100\text{k}\Omega$. To its size, we ignore the diode resistance on the input.

The total

The transfer characteristics of the high-pass filter before the non-inverting amplifier can be calculated as:

$$V_o = V_{in} \cdot G(s) \tag{22}$$

$$G(s) = \frac{R6}{\frac{1}{sC2} + R6} \Rightarrow \frac{s}{s + \frac{1}{C2R6}} \tag{23}$$

with a cutoff frequency at:

$$f_{C2} = \frac{1}{2\pi C2R6} = 15.392\text{Hz} \tag{24}$$

the transfer characteristics of the op amp is set by its GBWP, which is unity as 2.8MHz

3.2 simulation

3.3 realisation