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.

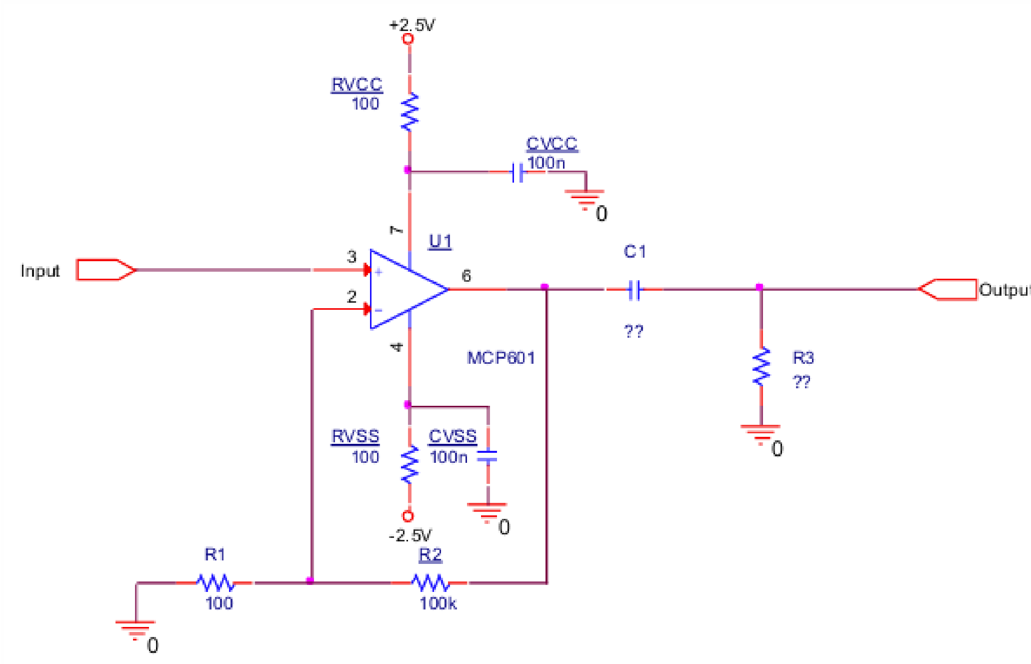


Figure 1: non-inverting amplifier with 1001 times gain

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

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

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

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

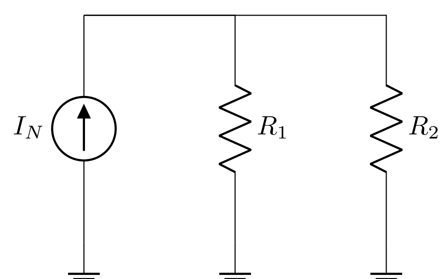
$$E_N = e_{nw}\sqrt{f_c e \ln\left(\frac{f_H}{f_L}\right) + f_H - f_L} = 0.061\mathrm{mV} \tag{7}$$

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

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:

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

1.2 simulation

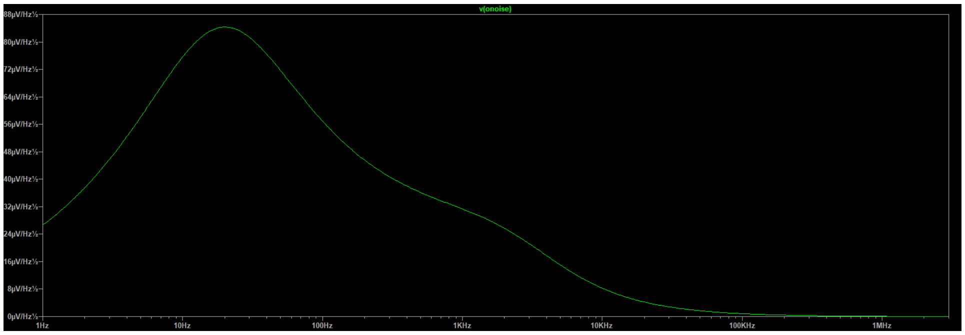


Figure 2: siumlation of the first circuit configuration.

1.3 realisation

Following the schematic and building the circuits lead to a measurement of $AC_{RMS} = 1.730$mV. To increase the accuracy of the measurement, the bench top multimeter was used, as it has better accuracy in AC-RMS scenarios. Whereas the accuracy of the oscilloscope is limited, due to the fact it integrates the signal over a shorter time period, usualy only the entire "screen".

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:

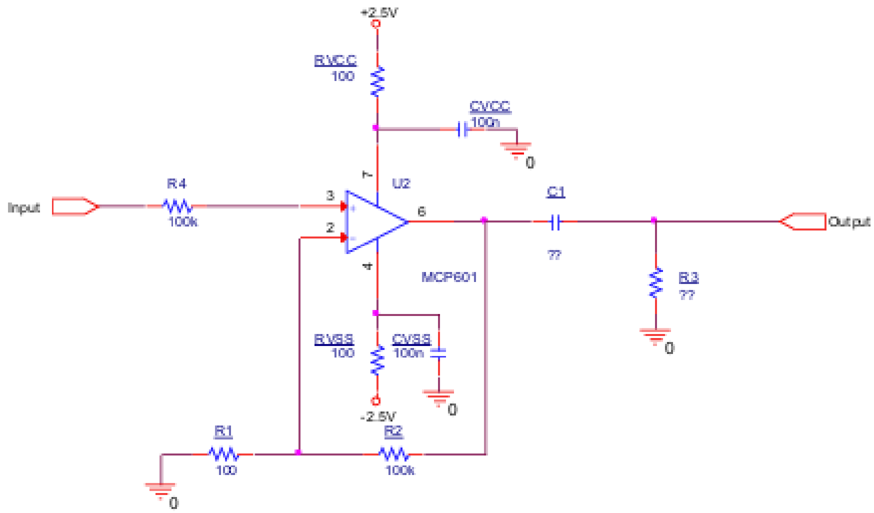


Figure 3: same non-noninverting amplifier from previous section, this time with a large input resistance, which acts a generator resistance

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.

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

And the thermal noise:

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

The total RMS voltage with generator resistance term added

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

2.2 simulation

2.3 realisation

Same as the previous circuit, however this time with the added resistor. We can an $AC_{RMS_100k} = 3.285\text{mV}$ and $AC_{RMS_200k} = 4.437\text{mV}$

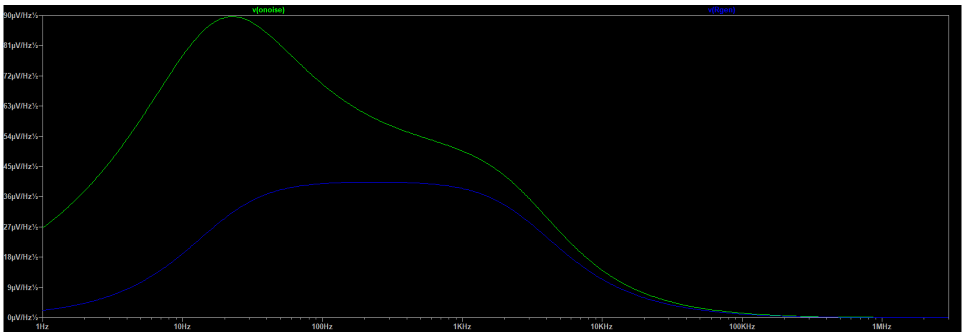


Figure 4: simulation result of the noise spectrum of the non inverting amplifier with a generator resistance of 100k

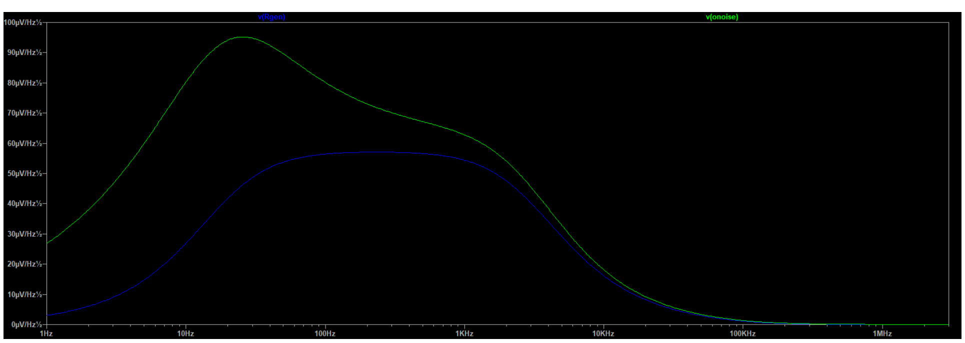


Figure 5: simulation result of the noise spectrum of the non inverting amplifier with a generator resistance of 100k

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

3.1 theory

Optical receiver circuits, use transimpedance amplifiers to make an 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.

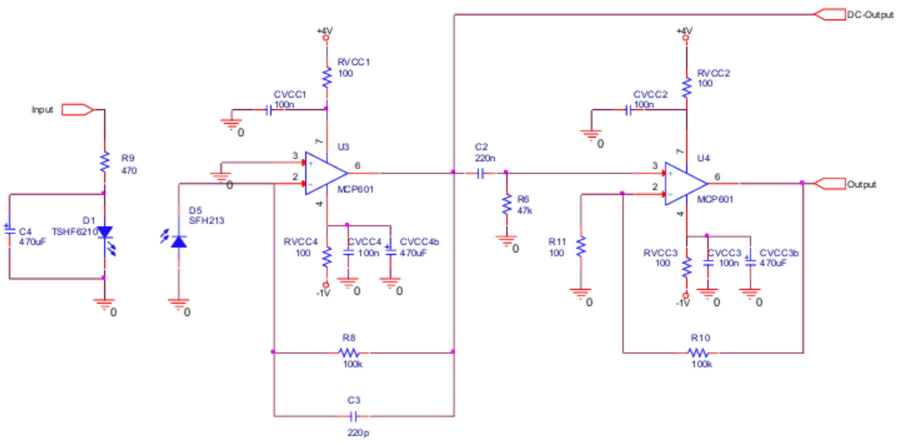


Figure 6: photo detector circuit, a transimpedance amplifier, filtered into a noninverting amplifier with a gain of 1001

Above is the overall circuit, which contains subcircuits, the transimpedance amplifier implemented by op amp U3, the high pass filter $C2R6$ and a non-inverting amplifier U4. This also means there are multiple noise bandwidths. For U3 the bandwidth is determined by the high-pass filter for the lower limit aswell as the gain-bandwidth product for the upper limit. As for U4, the lower limit is determined by the measurement equipment, for an analog discovery the AC RMS is 3Hz, and the upper limit is determined by U4's gain bandwidth product. As will be shown later, the bandwidth of the op amp is above 1KHz, thus the value for $e_{nw} = 21\frac{\text{nV}}{\sqrt{\text{Hz}}}$
Determine the noise gains: Since the transimpedance amplifier, is mainly dominated by current noise, the voltage noise gain is set to 1

$$A_{NU3} = 1 \tag{20}$$

For op amp U4 the noise gain is determined by the closed loop gain:

$$A_{NU4} = 1 + \frac{R10}{R11} = 1001 \tag{21}$$

This gain is also the same as the signal gain.

The upper and lower limit for U3's noise bandwidth:

$$f_{U3l} = \frac{1}{2\pi R6C2} \cdot \frac{\pi}{2} = 24.178\text{Hz} \tag{22}$$

$$f_{U3h} = \frac{f_t}{A_{NU4}} \cdot \frac{\pi}{2} = 4.394\text{kHz} \tag{23}$$

The main noise component of noise from a photodiode is shot-noise. Its noise originating from charge carriers when they cross a potential barrer such as a diode junction. It is defined as $\sqrt{2qI_d}$ where $q = 1.602 \cdot 10^{-19}$C, is the electron charge.
It is possible to create an equation relating voltage noise from current as a function of the diode current I_d

$$V_{Ndiode}(I_d) = R8 \cdot A_{NU3} \cdot A_{U4} \cdot \sqrt{2qI_d \cdot (f_{U3h} - f_{U3h})} \tag{24}$$

Voltage noise from op amp U3:

$$V_{NU3} = e_{nw} \cdot A_{NU3} \cdot A_{U4} \cdot \sqrt{f_{ce} \cdot \ln\left(\frac{f_{U3h}}{f_{U3l}}\right) + f_{U3h} - f_{U3l}} = 2.057\text{mV} \tag{25}$$

Current noise from op amp U3:

$$V_{NIU3} = A_{NU3} \cdot A_{U4} \cdot i_{ni} \cdot R8\sqrt{(f_{U3h} - f_{U3h})} = 3.97\mu\text{V} \tag{26}$$

Thermal noise from R8:

$$V_{NR8} = A_{NU3} \cdot A_{U4}\sqrt{4kt \cdot R8 \cdot (f_{U3h} - f_{U3h})} = 2.693\text{mV} \tag{27}$$

Noise bandwidth from U4:

$$f_{U4l} = 3\text{Hz} \cdot \frac{\pi}{2} = 4.712\text{Hz} \tag{28}$$

$$f_{U4h} = f_{U3h} = 4.934\text{kHz} \tag{29}$$

Next is all the noise that is related to the op amp U4, starting with noise from R6, however this is limited from lowpass relationship

$$V_{NR6} = A_{U4} \cdot \sqrt{4kt \cdot R6 \cdot (f_{U3l} - f_{U4l})} = 123.239\mu\text{V} \tag{30}$$

Voltage noise from U4:

$$V_{NU4} = e_{nw} \cdot A_{NU4} \cdot A_{U4} \cdot \sqrt{f_{ce} \cdot \ln\left(\frac{f_{U4h}}{f_{U4l}}\right) + f_{U4h} - f_{U4l}} = 2.227\text{mV} \tag{31}$$

Current noise from U4 positive and negative inputs:

$$V_{NIU4+} = A_{NU} \cdot i_{ni} \cdot R6 \cdot \sqrt{(f_{U4h} - f_{U4l})} = 1.87\mu\text{V} \tag{32}$$

$$V_{NIU4-} = A_{NU} \cdot i_{ni} \cdot \frac{R10R11}{R10 + R11} \cdot \sqrt{(f_{U4h} - f_{U4l})} = 3.975\text{nV} \tag{33}$$

And lastly thermal noise from U4 feedback network:

$$V_{NR10R11} = A_{NU4} \cdot \sqrt{4kt \cdot \frac{R10R11}{R10 + R11} \cdot (f_{U4h} - f_{U4l})} = 85.318\mu\text{V} \tag{34}$$

The individual terms can then be squared, rooted and summed together to get the entire noise contribution of the entire circuit. This total noise is a function of the diode current

$$V_{Ntotal}(I_d) = \sqrt{V_{Ndiode}(I_d)^2 + V_{NU3}^2 + V_{NIU3}^2 + V_{NR8}^2 + V_{NR6}^2 + V_{NU4}^2 + V_{NIU4+}^2 + V_{NIU4-}^2 + V_{NR10R11}^2} \tag{35}$$

Graphing this functing yields:

3.2 simulation

3.3 realisation

Building the circuit according the schematic and shielding it as good as possible. During the measurements the trasmitter supply voltage, was increased in descrete steps of 200mV, to firstly scan for large DC-output swings, then swept again at a higher resolution. The DC-output of the transimpedance amplifier is used to determine the receiver diode current, and the final output of the system, to determine overall system noise.

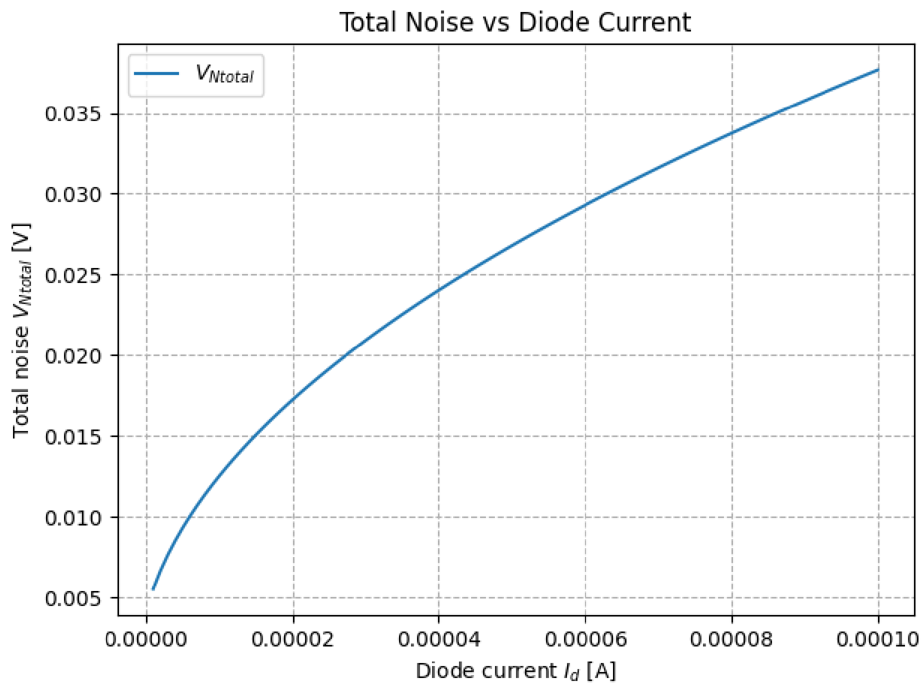


Figure 7: total system voltage noise as a function of diode current. Note that the noise is proportional to the squareroot of the current, this is due to the fact that shotnoise $V_{N,shot} = \sqrt{2qid}$, thus as the diode current increases, the shotnoise does not increase linearly.

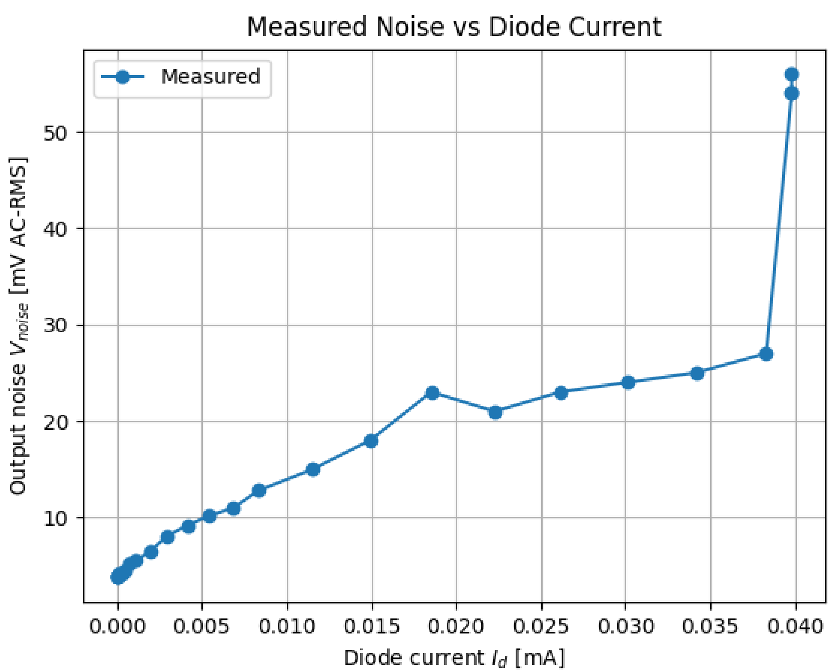


Figure 8: circuit measurement of AC RMS noise from the output vs the diode current. It can be seen that the curve resembles the theoretical one, due to its decreasing nature, due to the squareroot of the shotnoise. the abrubt and sudden increase in noise, is the result of the transimpedance amplifiers output, saturating at roughly 3,980V due to voltage supply limitations.