FH JOANNEUM
Elektronik und
Computer Engineering

Analog Circuit Design Laboratory

Lab Report

written by:

Lorenz Buchinger

Agustin Sevil de Llobet

Power Electronics Engineering

FH JOANNEUM – University of Applied Sciences, Austria

Graz, January 1, 1991

Contents

1 Computer Lab

1.1 Noise Analysis with PSpice

1.1.1 Task Description

1.1.2 Schematic

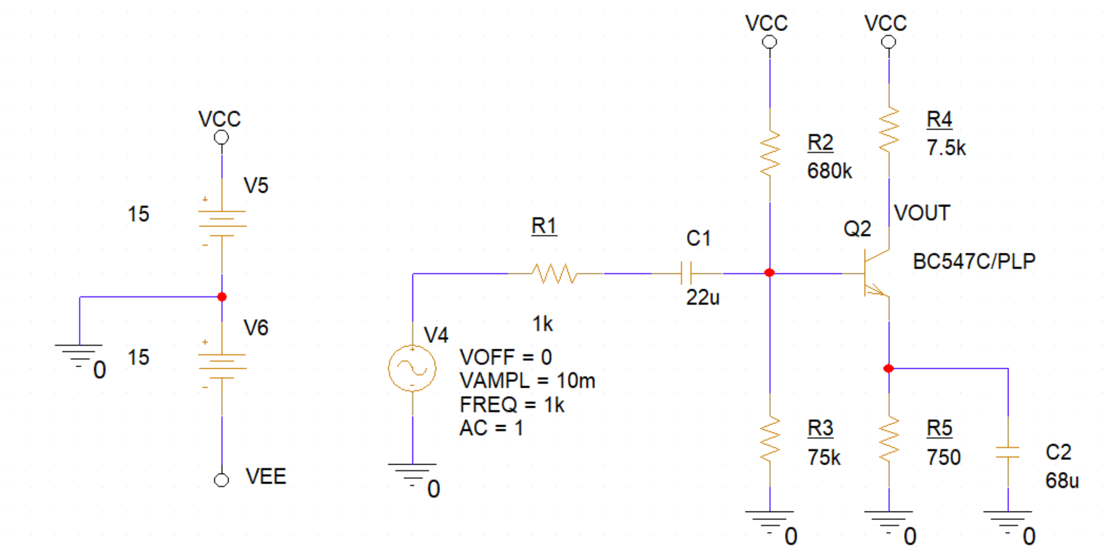


Figure 1.1: Schematic of the single stage emitter circuit used for simulation.

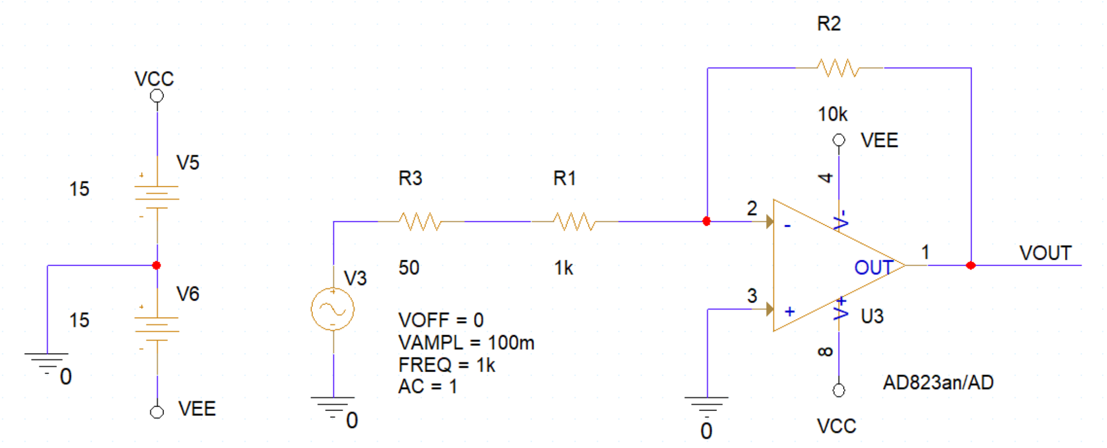


Figure 1.2: Schematic of the inverting amplifier circuit used for simulation.

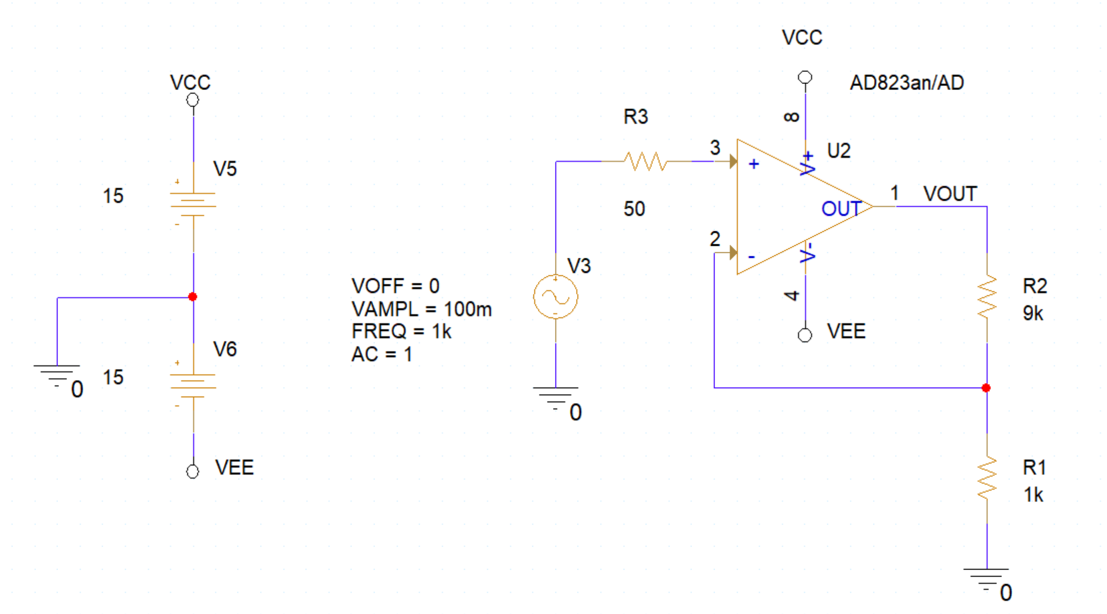


Figure 1.3: Schematic of the non-inverting amplifier circuit used for simulation.

1.1.3 Curves & Diagrams

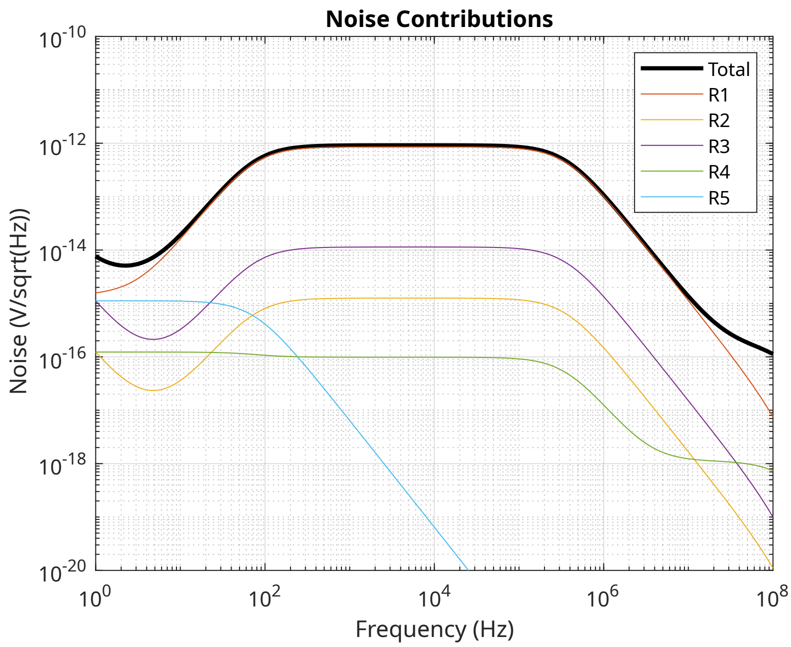


Figure 1.4: Noise spectral voltage density of the single stage emitter circuits output. The simulation results of the noise generated by each individual resistor is also depicted.

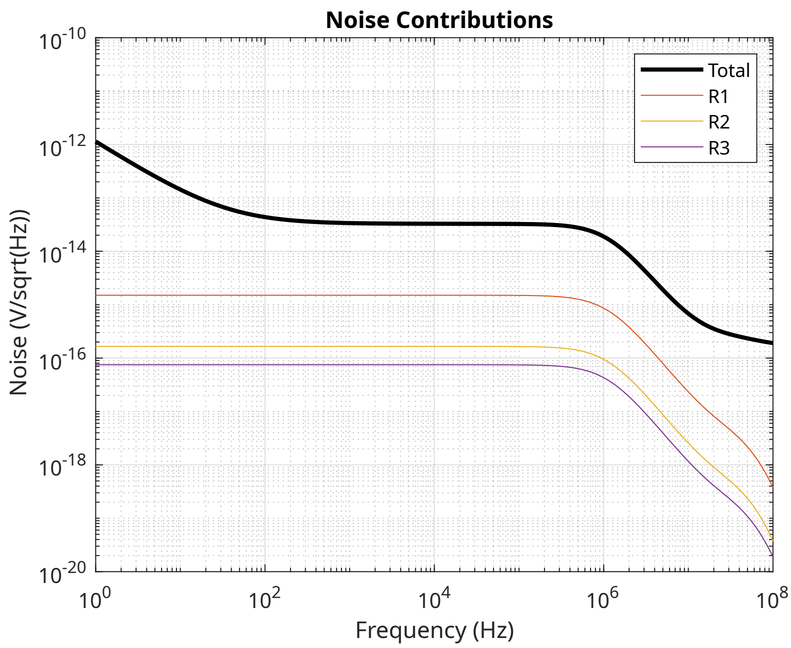


Figure 1.5: Noise spectral voltage density of the inverting amplifiers output. The simulation results of the noise generated by each individual resistor is also depicted.

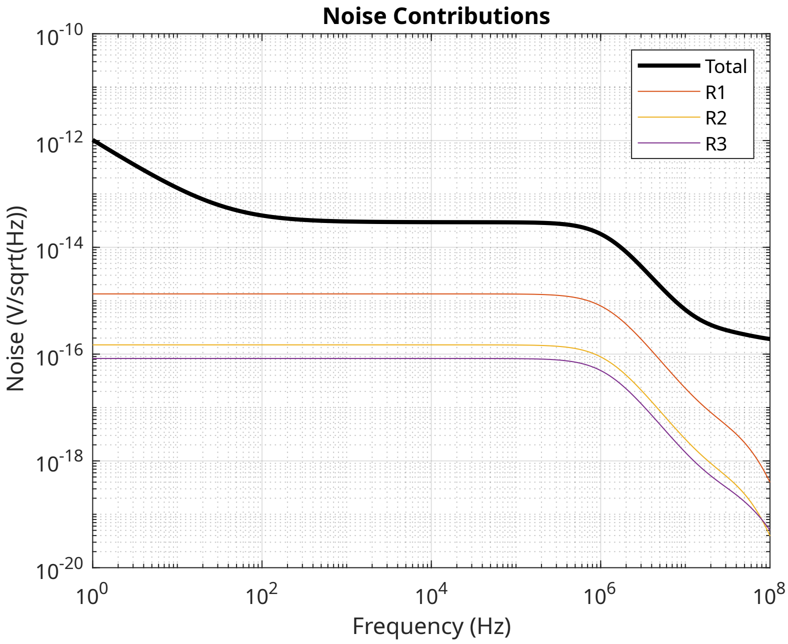


Figure 1.6: Noise spectral voltage density of the non-inverting amplifiers output. The simulation results of the noise generated by each individual resistor is also depicted.

1.1.4 Discussion of Measurement Results

2

Electronic Lab

2.1 Active Band-pass Filter

2.1.1 Task Description

A second-order active bandpass filter was designed using the LM4562 dual operational amplifier by cascading two staggered tuned first-order bandpass stages in multi-feedback topology. The transfer function and component values were calculated manually according to the given specifications (Butterworth and 1 dB Chebyshev, 4–5 kHz, 20 dB gain), using E24 components with 1

The circuit was simulated in PSpice (transient, AC, noise and Monte Carlo analysis), implemented on a prototype board with ±5 V supply, and experimentally characterized. Frequency response, passband gain, 3 dB corner frequencies and step response were measured and compared with simulation results in a combined Bode plot.

Finally, the output noise behavior was analyzed and discussed in the report.

2.1.2 Schematic

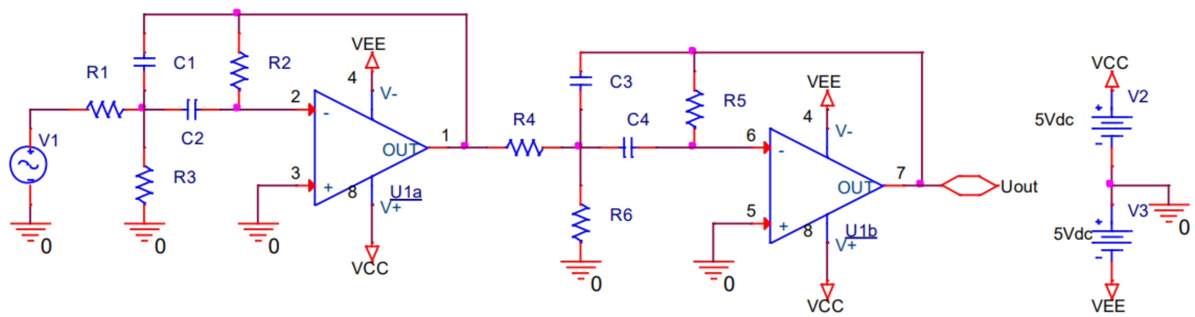


Figure 2.1: Schematic of the second order band pass filter.

2.1.3 Formulas and Calculations

The calculations were done by hand. A scan of the hand written notes can be found in section A.1.1 of the appendix

2.1.4 Table(s) with Measurement Results

Table 2.1: Component values for Butterworth band-pass filter

	Calculated	**Chosen**	**Measured**
R_1 in kΩ	5.448	5.6	5.5831
R_2 in kΩ	48.880	47	46.943
R_3 in Ω	303.59	300	298.59
R_4 in kΩ	4.650	4.7	4.6045
R_5 in kΩ	41.718	43	42.798
R_6 in Ω	259.11	270	269.01
C_1 in nF	-	10	10.03
C_2 in nF	-	10	10.09
C_3 in nF	-	10	10.03
C_4 in nF	-	10	10.10

Table 2.2: Component values for Chebyshev 1 dB band-pass filter

	Calculated	**Chosen**	**Measured**
R_1 in kΩ	7.532	7.5	-
R_2 in kΩ	124.641	120	-
R_3 in Ω	117.89	120	-
R_4 in kΩ	6.492	6.2	-
R_5 in kΩ	107.432	110	-
R_6 in Ω	101.613	100	-
C_1 in nF	-	10	10.03
C_2 in nF	-	10	10.09
C_3 in nF	-	10	10.03
C_4 in nF	-	10	10.10

The measured resistor values show deviations below 1% with respect to the nominal chosen E24 values, which confirms the specified tolerance class. The capacitor measurements also confirm the 1% NP0 specification and therefore only introduce minor shifts in the center frequency.

Table 2.3: Measured frequency response of band-pass filter

Frequency / Hz	**Gain / dB**	**Phase / Degree**	**Remark**
1	-47	-	
10	-45	-	
100	-45	-	
1000	-32	-	
1100	-30	170	
1200	-29.4	165	
1300	-27.9	165	
1400	-25.8	162	
1500	-24.3	160	
2000	-17.6	164	
2600	-10	164	

Table 2.3: Measured frequency response of band-pass filter (continuation)

Frequency / Hz	**Gain / dB**	**Phase / Degree**	**Remark**
3000	-4.3	158	
3300	0.2	152	
3400	1.9	148	
3500	3.6	145	
3600	5.5	142	
3700	7.4	136	
3800	9.5	130	
3900	11.7	123	
4000	14.2	112	
4100	16.5	100	
4200	18.7	82	**-3 dB region**
4230	19.3	76	**-3 dB region**
4300	20.5	62	**-3 dB region**
4400	21.6	39	
4472.13	22.1	22	**Theoretical f_c**
4500	22.2	16	
4570	22.3	0	**Measured f_c**
4600	22.3	-6	
4700	22	-28	
4800	21.2	-50	
4900	19.9	-70	**-3 dB region**
4940	19.3	-76	**-3 dB region**
5000	18.2	-86	**-3 dB region**
5100	16.4	-100	
5200	14.5	-110	
5300	12.7	-119	
5400	11	-126	
5500	9.4	-131	

Table 2.3: Measured frequency response of band-pass filter (continuation)

Frequency / Hz	Gain / dB	Phase / Degree	Remark
5600	8	-136	
5700	6.7	-139	
5800	5.5	-142	
5900	4.3	-145	
6000	3.2	-147	
6100	2.2	-150	
6200	1.2	-152	
6300	0.3	-154	
6400	-0.5	-156	
6700	-2.8	-160	
7500	-7.5	-171	
8000	-9.7	-	
9000	-13.3	-	
10000	-16.4	-	
50000	-32.5	-	

2.1.5 Curves & Diagrams

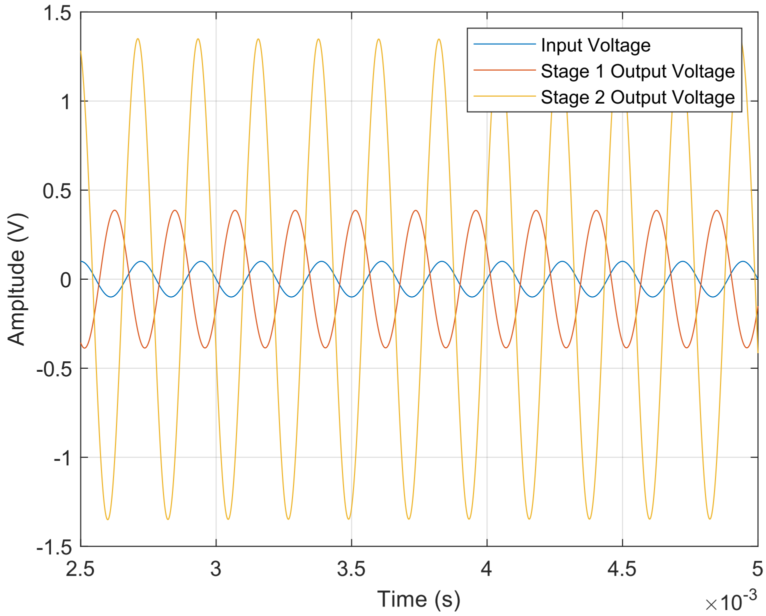


Figure 2.2: Transient simulation showing the sinusoidal input voltage and the output voltage of each of the two amplifier stages.

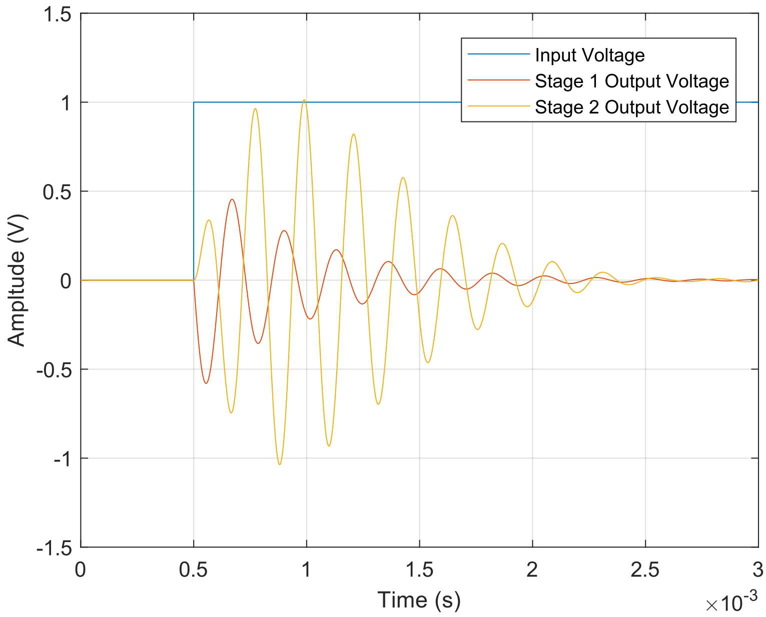


Figure 2.3: Transient simulation showing the step response of both amplifier stages after applying a signal transitioning from 0 V to 1 V to the input.

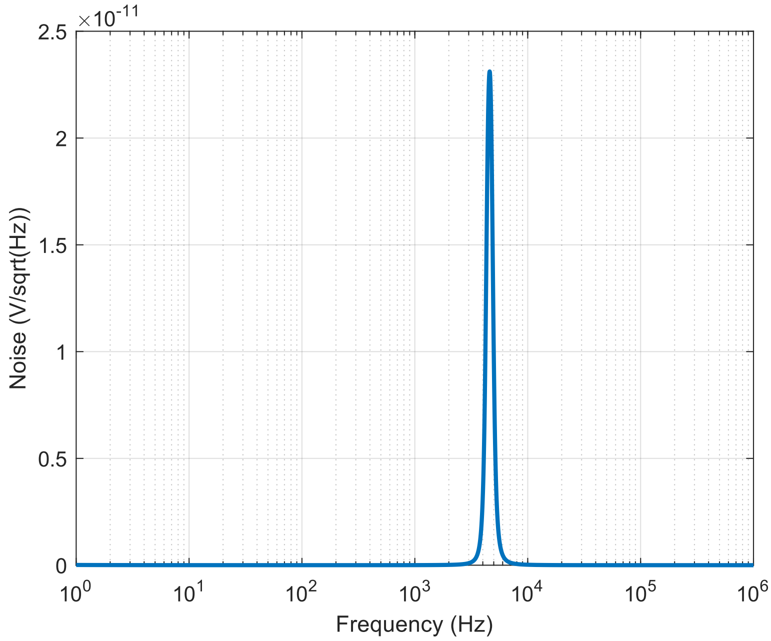


Figure 2.4: Simulation results of the noise spectral voltage density.

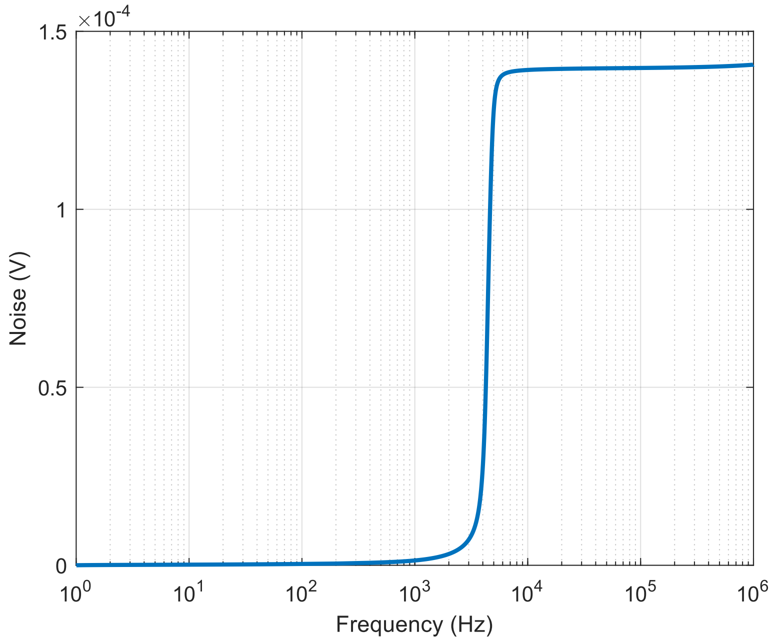


Figure 2.5: Simulation results of the total output noise voltage.

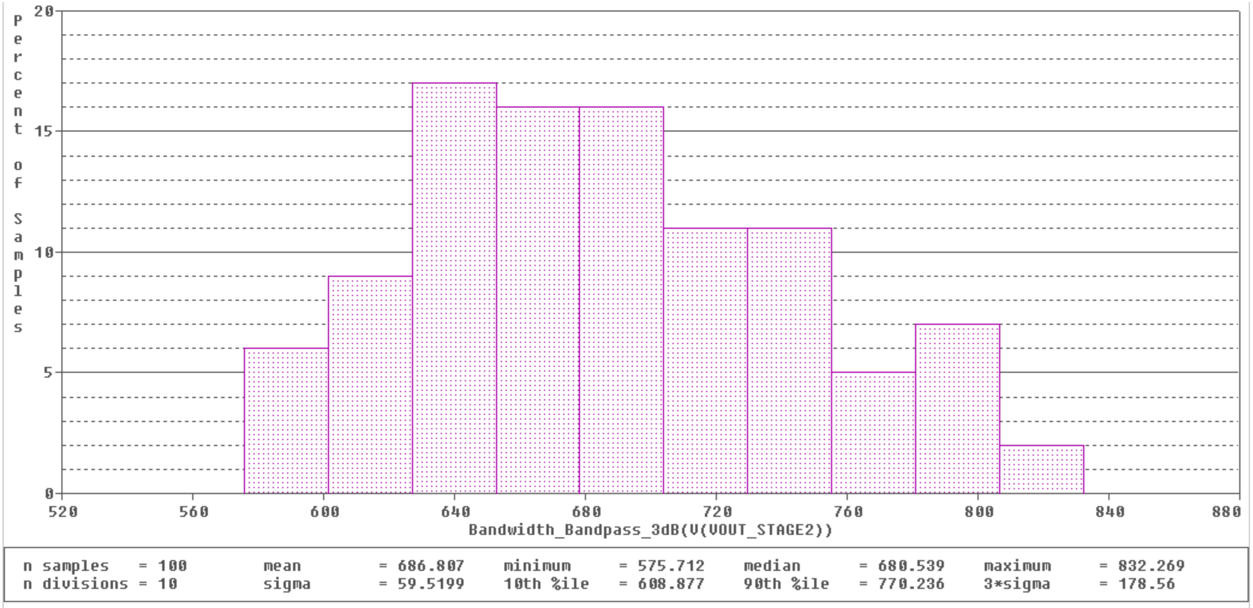


Figure 2.6: Monte Carlo simulation of the 3 db bandwidth. 1 % tolerances were used for both resistors and capacitors.

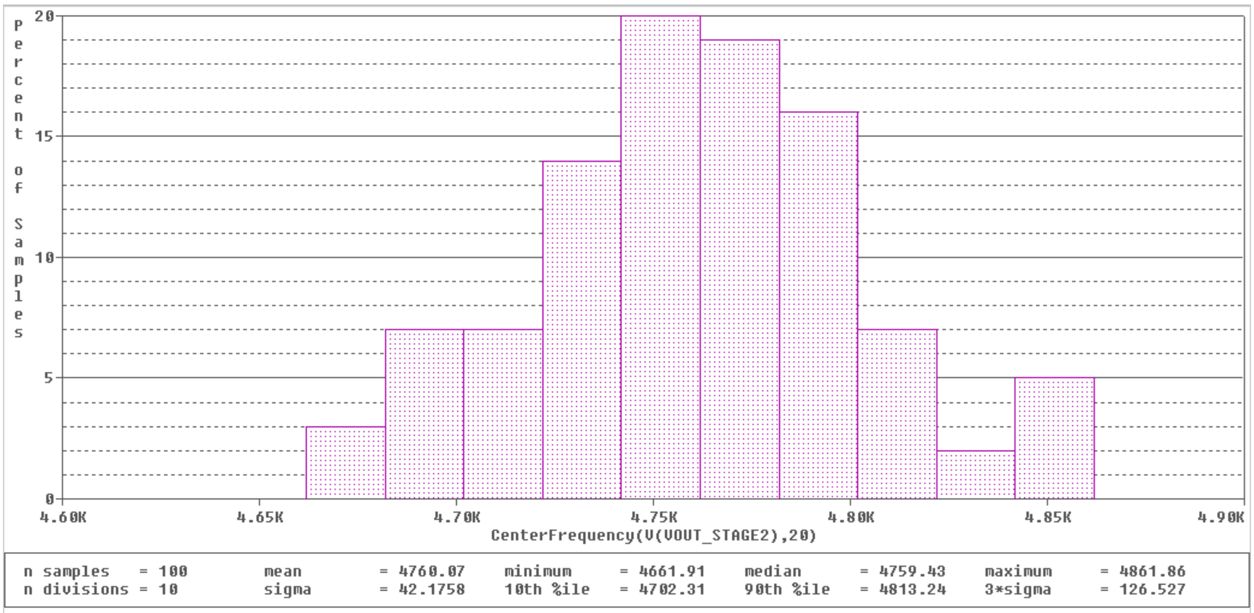


Figure 2.7: Monte Carlo simulation of the center frequency. 1 % tolerances were used for both resistors and capacitors.

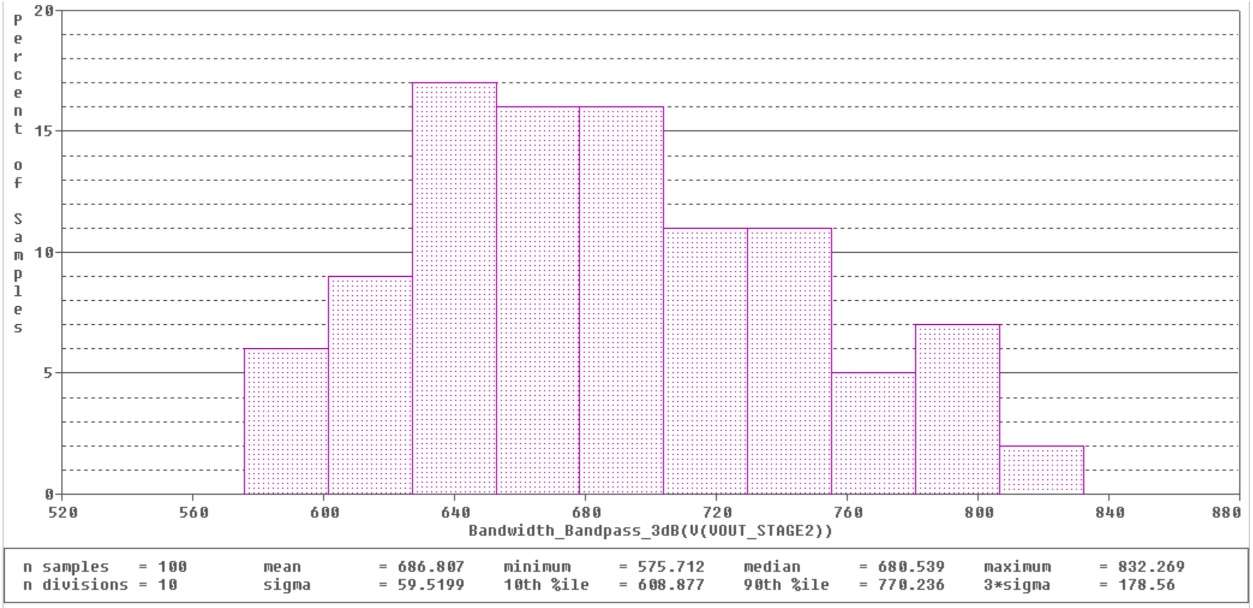


Figure 2.8: Monte Carlo simulation of the gain variation. 1 % tolerances were used for both resistors and capacitors.

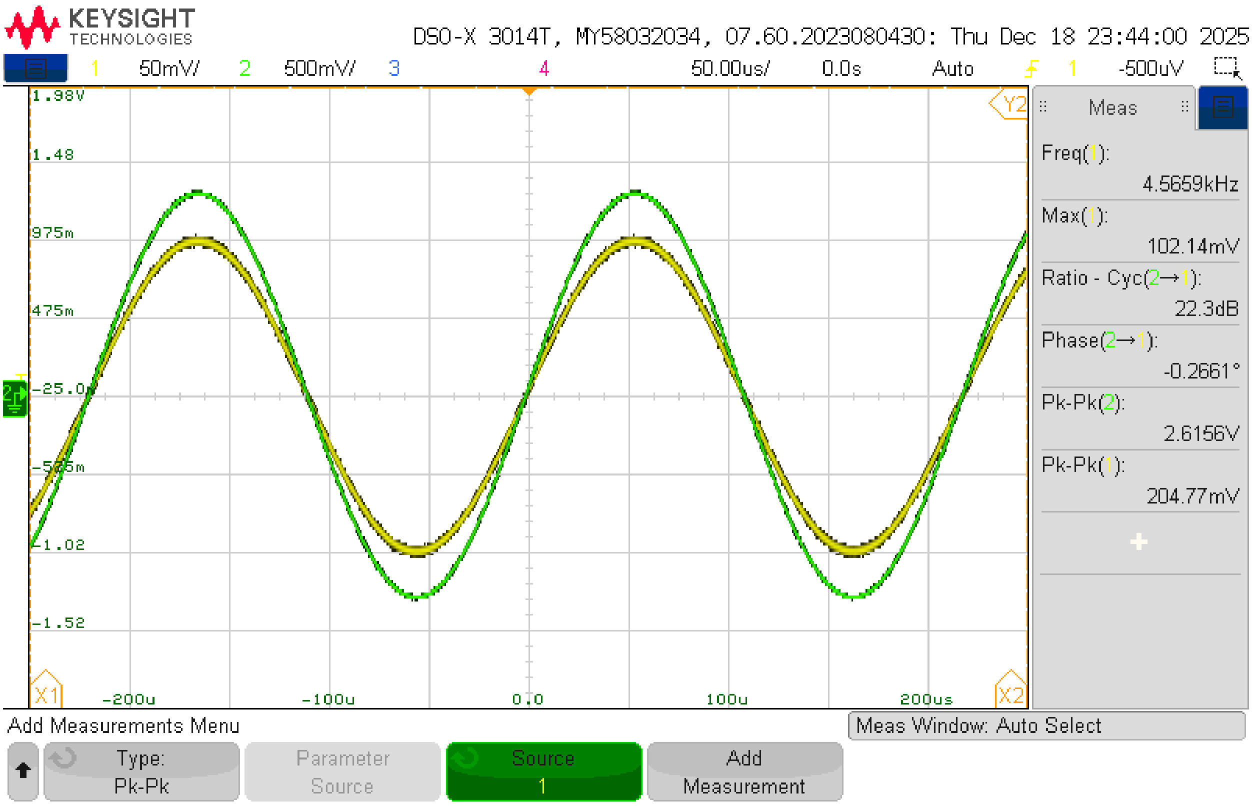


Figure 2.9: Measurement of the passband gain.

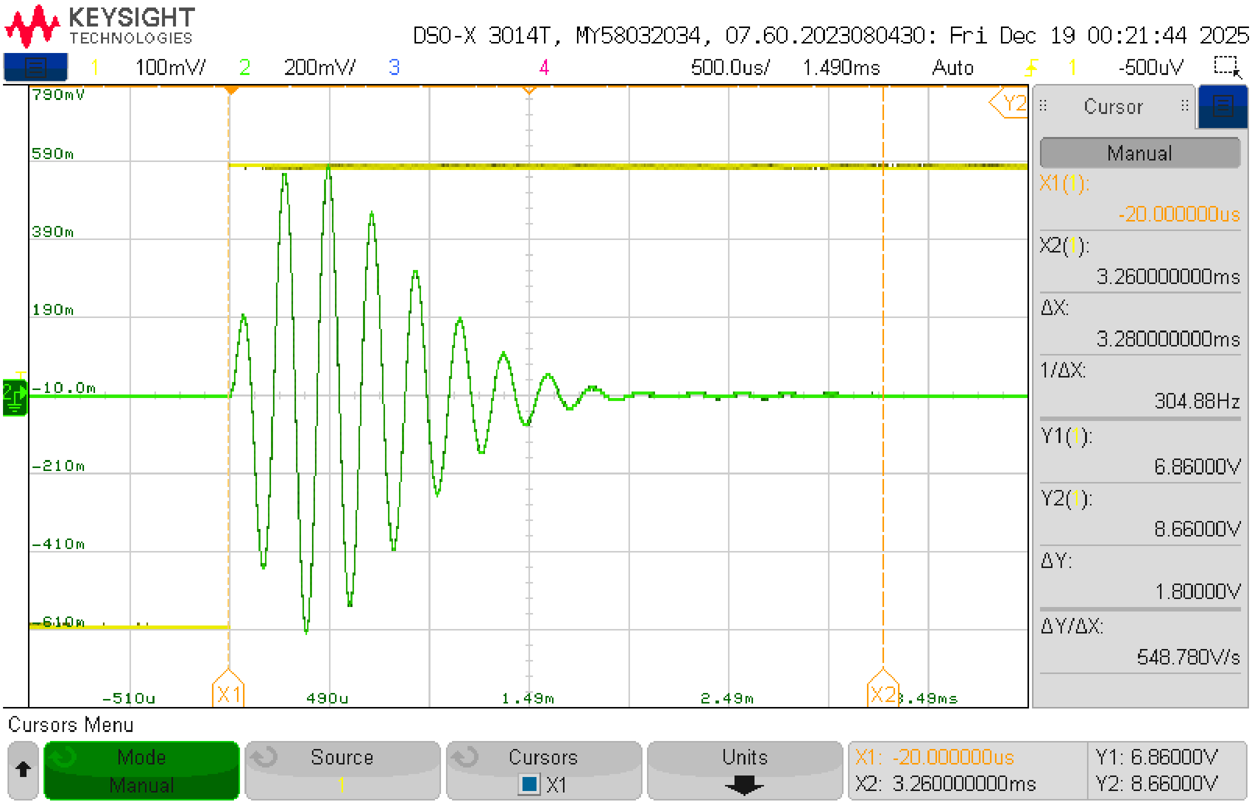


Figure 2.10: Measurement of the filters step response. A 0,5 V square wave was used as the step function.

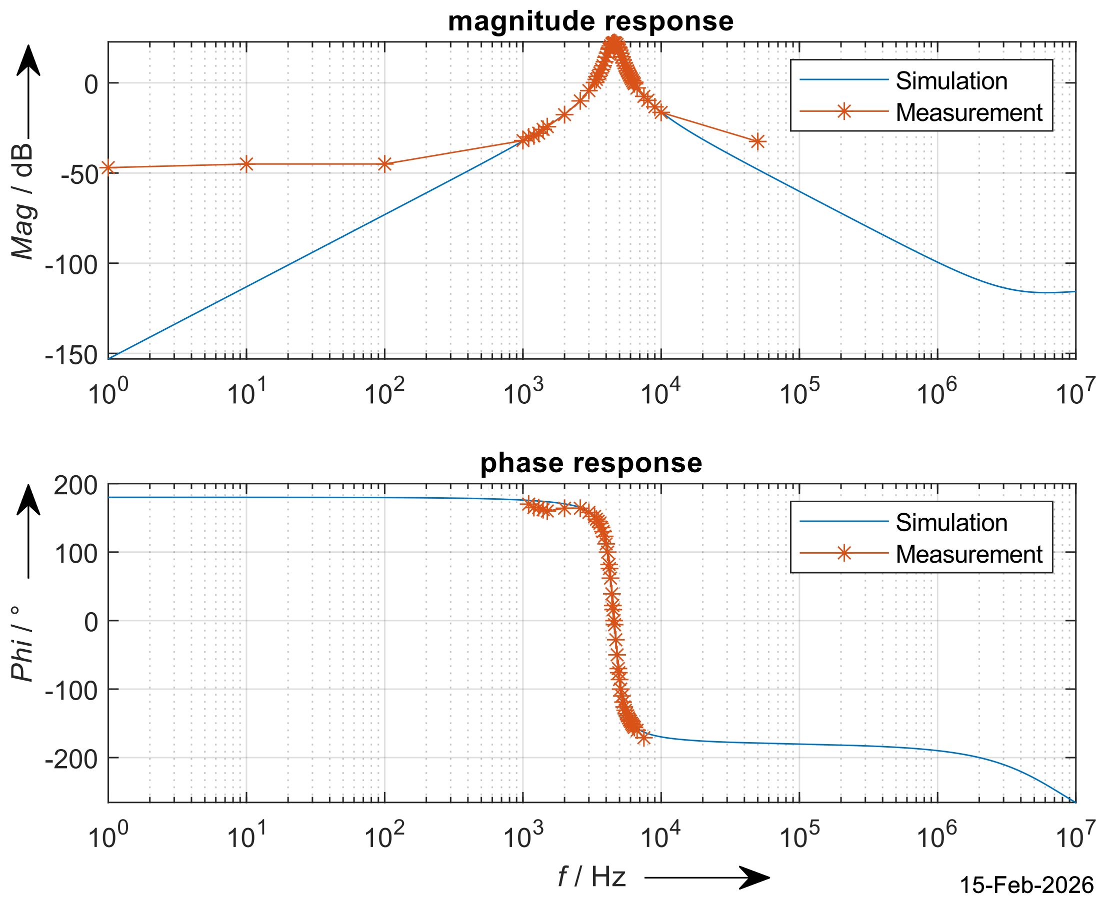


Figure 2.11: Bode plot displaying both simulated and measured values.

2.1.6 Discussion of Measurement Results

2.2 ADC-Driver And Anti-aliasing Filter

2.2.1 Task Description

A single-ended to differential driver for the AD7626 ADC was designed using the LM4562 dual operational amplifier. The circuit includes a second-order active low-pass filter (100 Hz – 40 kHz) with a total differential gain of 20 and a DC offset of 1 V at both outputs.

The circuit was simulated in PSpice (Bode plot, time-domain response and output noise), implemented on a prototype board, and experimentally characterized. Magnitude and phase response, step response, and differential output signals were measured and compared with the simulation results.

Finally, the SNR and ENOB were calculated for a full-scale ADC input signal.

C_1	150 pF
C_2	4,7 nF
f_{cl}	100 Hz
f_{cu}	40 kHz
A_0	20

Table 2.4: Given values.

2.2.2 Schematic

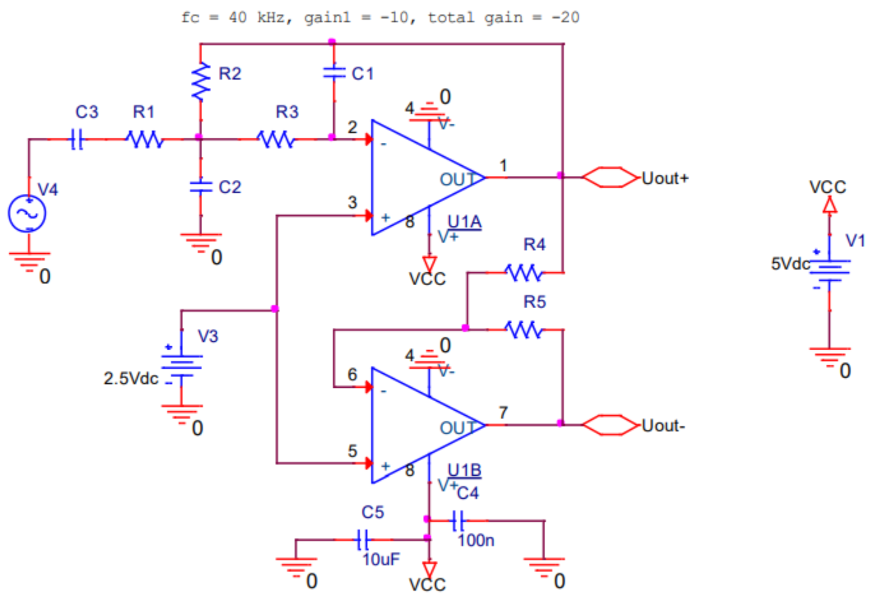


Figure 2.12: Schematic of the ADC driver with anti-aliasing filter

2.2.3 Formulas and Calculations

To calculate the component values the general transfer function for

$$H(s) = \frac{H_0}{1 + a \cdot s + b \cdot s^2} \quad (2.1)$$

$$H(P) = -\frac{\frac{R_2}{R_1}}{1 + \omega_c C_1 \left(R_2 + R_3 + \frac{R_2 R_3}{R_1} \right) P + \omega_c^2 C_1 C_2 R_2 R_3 P^2} \tag{2.2}$$

$$H_0 = -\frac{R_2}{R_1} \tag{2.3}$$

$$a = \omega_c C_1 \left(R_2 + R_3 + \frac{R_2 R_3}{R_1} \right) \tag{2.4}$$

$$b = \omega_c^2 C_1 C_2 R_2 R_3 \tag{2.5}$$

$$P = \frac{s}{\omega_c} \tag{2.6}$$

The decision was made to use a Butterworth characteristic for the second order multi-feedback filter. The values for a and b for a Butterworth filter are as follows:

$$a = \sqrt{2} \tag{2.7}$$

$$b = 1 \tag{2.8}$$

ω_c is calculated according to equation 2.9.

$$\omega_c = 2\pi f_{cu} = 2\pi \cdot 40\,\mathrm{kHz} = 251 \cdot 10^3 s^{-1} \tag{2.9}$$

SNR and ENOB Calculation of the ADC Driver System

The system consists of a single-ended to differential driver implemented with the LM4562 operational amplifier, followed by the AD7626 16-bit ADC. The total system performance is limited by both the ADC quantization noise and the analog front-end noise.

Ideal ADC SNR

For an ideal N-bit ADC with a full-scale sinusoidal input, the theoretical Signal-to-Noise Ratio (SNR) is given by:

$$SNR_{ideal} = 6.02N + 1.76 \; [dB] \tag{2.10}$$

For a 16-bit ADC:

$$SNR_{ideal} = 6.02 \cdot 16 + 1.76 = 98.08 \, dB \tag{2.11}$$

This value represents the quantization noise limit only.

Real ADC SNR (Datasheet Value)

According to the AD7626 datasheet, the typical SNR for a full-scale sine wave is:

$$SNR_{ADC} \approx 95 \, dB \tag{2.12}$$

This value includes internal non-idealities and thermal noise.

Analog Driver Signal Level

The input signal to the driver is:

$$V_{in,pp} = 100 \, mV \tag{2.13}$$

The differential gain of the driver is:

$$A_{diff} = 20 \tag{2.14}$$

Therefore, the peak-to-peak voltage at the ADC input is:

$$V_{out,pp} = A_{diff} \cdot V_{in,pp} = 20 \cdot 0.1 = 2 \, V \tag{2.15}$$

RMS Signal Value at the ADC Input

For a sinusoidal signal:

$$V_{signal,RMS} = \frac{V_{pp}}{2\sqrt{2}} = \frac{2}{2\sqrt{2}} = \frac{2}{2.828} = 0.707 \, V \tag{2.16}$$

RMS Noise Voltage of the Driver

From PSpice noise simulation over the bandwidth (100 Hz – 40 kHz):

$$V_{noise,RMS} = 40.7\,\mu V \tag{2.17}$$

Driver SNR

$$SNR_{driver} = \frac{V_{signal,RMS}}{V_{noise,RMS}} = \frac{0.707}{40.7 \times 10^{-6}} = 17375 \tag{2.18}$$

In decibels:

$$SNR_{driver,dB} = 20\log_{10}(17375) = 84.8\,dB \tag{2.19}$$

Total System SNR

Since ADC noise and driver noise are uncorrelated, they must be combined in linear scale:

$$\frac{1}{SNR_{total}} = \frac{1}{SNR_{ADC}} + \frac{1}{SNR_{driver}} \tag{2.20}$$

First, convert both SNR values to linear scale:

$$SNR_{ADC,lin} = 10^{95/10} = 3.16 \times 10^{9} \tag{2.21}$$

$$SNR_{driver,lin} = 10^{84.8/10} = 3.02 \times 10^{8} \tag{2.22}$$

Combine both contributions:

$$\frac{1}{SNR_{total,lin}} = \frac{1}{3.16 \times 10^{9}} + \frac{1}{3.02 \times 10^{8}} \tag{2.23}$$

$$SNR_{total,lin} \approx 2.76 \times 10^{8} \tag{2.24}$$

Convert back to decibels:

$$SNR_{total,dB} = 10\log_{10}(2.76 \times 10^{8}) = 84.4\,dB \tag{2.25}$$

Effective Number of Bits (System)

$$ENOB = \frac{SNR_{total,dB} - 1.76}{6.02} = \frac{84.4 - 1.76}{6.02} = \frac{82.64}{6.02} = 13.7\,bits \tag{2.26}$$

2.2.4 Table(s) with Measurement Results

Table 2.5: Component values for ADC driver and anti-aliasing filter

	Calculated	**Chosen**	**Measured**
R_1 in Ω	-	820	817
R_2 in kΩ	-	8.2	8.1602
R_3 in kΩ	-	2.7	2.6888
R_4 in kΩ	10	10	9.8642
R_5 in kΩ	10	10	9.8666
C_1 in pF	-	150	155
C_2 in nF	-	4.7	4.61
C_3 in µF	-	1.5	1.56
C_4 in nF	-	100	-
C_5 in µF	-	10	-

The measured resistor deviations are within the expected 1% tolerance range. The small deviations of the RC components slightly influence the cutoff frequency of the anti-aliasing filter, but remain within acceptable limits. The measured capacitance variation of C1 (150 pF → 155 pF) leads to a minor shift of the pole frequency.

Table 2.6: Measured frequency response

Frequency / Hz	**Gain / dB**	**Phase / Degree**
3	-6.80	80
7	0.40	85
10	3.40	87
20	9.30	80
40	15.00	73

Table 2.6: Measured frequency response (continuation)

Frequency / Hz	**Gain / dB**	**Phase / Degree**
50	16.80	69
60	18.10	66
70	19.20	61
80	20.10	59
90	20.90	56
100	21.50	54
110	22.00	51
120	22.40	48
130	22.80	46
135	23.00	45
140	23.10	44
150	23.40	42
160	23.70	40
170	23.90	38
180	24.10	36
190	24.20	35
200	24.40	34
210	24.50	33
220	24.60	31
230	24.70	30
240	24.80	29
260	25.00	27
280	25.10	25
300	25.20	24
330	25.40	22
350	25.40	20
400	25.60	18
450	25.70	16

Table 2.6: Measured frequency response (continuation)

Frequency / Hz	Gain / dB	Phase / Degree
500	25.80	14
600	25.80	10
750	25.90	8
825	25.90	7
1000	26.00	6
1250	26.00	4
1500	26.00	2
1750	26.00	0
2250	26.00	-1
2500	26.00	-2
2750	26.00	-3
3000	26.00	-4
3500	26.00	-5
6000	26.00	-11
10000	25.90	-18
11000	25.90	-21
12000	25.90	-23
13000	25.80	-28
14000	25.80	-30
15000	25.80	-31
16000	25.70	-34
17000	25.70	-35
18000	25.60	-37
19000	25.60	-40
20000	25.50	-42
22000	25.40	-48
24000	25.20	-54
26000	25.00	-56

Table 2.6: Measured frequency response (continuation)

Frequency / Hz	Gain / dB	Phase / Degree
28000	24.70	-60
30000	24.40	-66
32000	24.10	-72
33000	24.00	-75
34000	23.80	-75
35000	23.60	-78
37000	23.20	-80
38000	23.00	-83
39000	22.80	-84
40000	22.60	-86
45000	21.50	-93
50000	20.40	-106
55000	19.20	-109
60000	18.00	-114
65000	16.90	-120
70000	15.50	-122
75000	14.50	-125
100000	9.90	-130
120000	6.80	-140
150000	3.30	-156
190000	0.00	-180
200000	-1.00	-180

The measured response shows a constant gain of approximately 26 dB within the passband. The phase shift approaches $-180\,\mathrm{deg}$ at high frequencies, which confirms the expected second-order low-pass behavior. The cutoff frequency can be determined from the $-3\,\mathrm{dB}$ drop relative to the mid-band gain.

2.2.5 Curves & Diagrams

2.2.6 Discussion of Measurement Results

Although the AD7626 is a 16-bit ADC with an ideal SNR of 98.08 dB, the system performance is limited by the analog driver noise. For the selected input amplitude (2 Vpp at the ADC input), the overall system achieves:

- $SNR_{total} = 84.4\,dB$
- $ENOB = 13.7\,bits$

Therefore, the effective resolution is reduced due to analog front-end noise and the fact that the ADC full-scale range is not fully utilized.

2.3 Switched Capacitor Filter

2.3.1 Task Description

A 5th-order clock-tunable switched capacitor filter based on the LT1063 (dual $\pm$5V supply) was implemented. A non-inverting preamplifier stage using the LM4562 with a gain of 100 was placed in front of the filter.

The frequency response was measured for two clock frequencies (1 MHz and 500 kHz) and compared in a Bode plot. The filter behavior near the clock frequency was investigated in the time domain, including possible clock feedthrough effects.

The total output noise was measured using additional amplification stages, and the input-referred noise voltage was determined for different source resistances. Finally, the SNR and ENOB were calculated for a full-scale sine wave.

2.3.2 Schematic

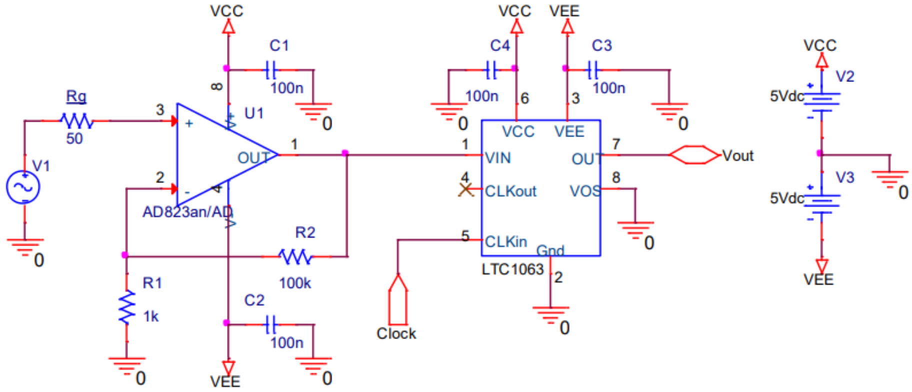


Figure 2.13: Schematic of the switched-capacitor filter with pre-amplifier stage.

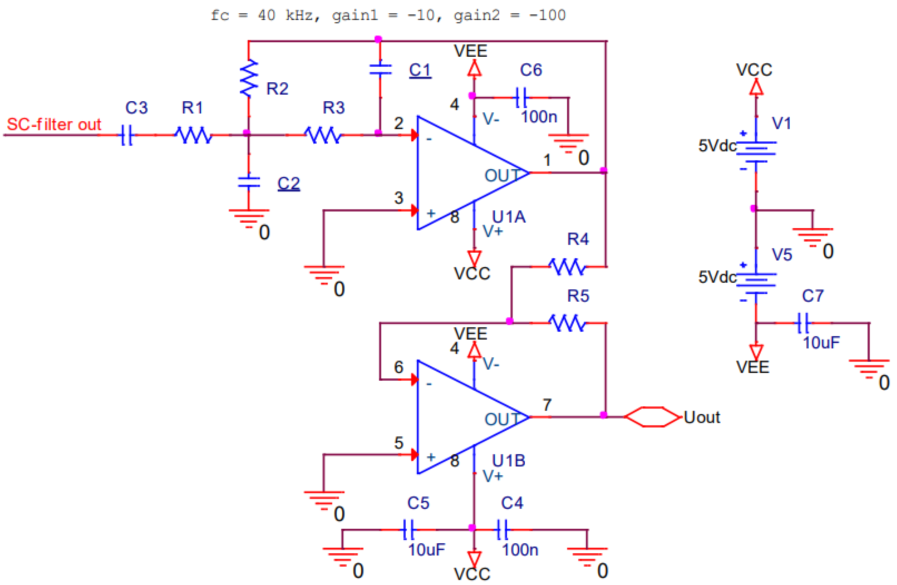


Figure 2.14: Schematic of the low pass filter and amplifier for noise measurement.

2.3.3 Formulas and Calculations

2.3.4 Table(s) with Measurement Results

Table 2.7: Component values for switched capacitor filter

	Calculated	Chosen	Measured
R_g in Ω	50	50	51.085
R_{g2} in kΩ	100	100	99.959
R_1 in Ω	-	1000	994.81
R_2 in kΩ	-	100	99.761
C_1 in nF	-	100	-
C_2 in nF	-	100	-
C_3 in nF	-	100	-
C_4 in nF	-	100	-

Table 2.8: Component values for low pass filter and noise measurement amplifier

	Calculated	Chosen	Measured
R_1 in Ω	-	820	817
R_2 in kΩ	-	8.2	8.1602
R_3 in kΩ	-	2.7	2.6888
R_4 in Ω	1000	1000	996.6
R_5 in kΩ	100	100	99.585
C_1 in pF	-	150	155
C_2 in nF	-	4.7	4.61
C_3 in µF	-	1.5	1.56
C_4 in nF	-	100	-
C_5 in µF	-	10	-
C_6 in nF	-	100	-
C_7 in µF	-	10	-

The measured resistor values confirm the 1% tolerance specification. Since the switched capacitor filter cutoff frequency is defined by the clock frequency rather than absolute capacitor values, the passive component tolerances have only minor influence on the filter characteristics.

Table 2.9: Measured frequency response of switched capacitor filter (CLK = 1 MHz)

Input Frequency / Hz	**Gain / dB**	**Phase / Degree**
1	40.2	1.33
10	40.2	0
50	40.2	0
100	40.2	0
300	40.2	-5
500	40.2	-10
1000	40.1	-18
2000	40.0	-30
3000	39.9	-50
4000	39.7	-70
5000	39.3	-90
6000	39.0	-100
7000	38.6	-120
8000	38.2	-135
9000	37.8	-160
10000	37.2	180
11000	36.6	160
12000	35.8	140
13000	35.0	120
14000	34.0	100
15000	33.0	84
16000	31.8	68
17000	30.7	50
18000	29.4	33
19000	28.1	18

Table 2.9: Measured frequency response of switched capacitor filter (CLK = 1 MHz, continuation)

Input Frequency / Hz	Gain / dB	Phase / Degree
20000	26.7	15
21000	25.3	-18
22000	23.6	-30
23000	21.7	-50
24000	19.7	-60
25000	17.5	-85
30000	6.7	-120
40000	-6.0	-150

Table 2.10: Measured frequency response of switched capacitor filter (CLK = 500 kHz)

Input Frequency / Hz	Gain / dB	Phase / Degree
1	40.2	1.4
10	40.2	0
50	40.2	0
100	40.2	0
300	40.2	-10
500	40.1	-18
700	40.1	-25
1000	40.0	-37
1500	39.9	-56
2000	39.6	-70
2500	39.3	-90
3000	39.0	-110
3500	38.6	-128
4000	38.2	-145
4500	37.7	-164
5000	37.2	180

Table 2.10: Measured frequency response of switched capacitor filter (CLK = 500 kHz, continuation)

Input Frequency / Hz	**Gain / dB**	**Phase / Degree**
5500	36.5	160
6000	35.8	140
6500	34.9	122
7000	34.0	100
7500	32.8	85
8000	31.8	70
9000	29.5	35
10000	26.7	0
11000	23.6	-33
12000	19.6	-70
13000	15.3	-90
14000	11.0	-115
15000	7.1	-136
16000	3.2	-150
17000	-0.4	-160

When reducing the clock frequency from 1 MHz to 500 kHz, the cutoff frequency of the switched capacitor filter decreases proportionally. This confirms the theoretical relationship:

$$f_c \propto f_{CLK}$$

The measured data shows approximately half the cutoff frequency compared to the 1 MHz case, which validates the internal switched capacitor principle. Phase wrapping at ±180° is again visible due to measurement representation.

2.3.5 Curves & Diagrams

2.3.6 Discussion of Measurement Results

Equipment

- Benchtop Multimeter
  Manufacturer: Agilent
  Model: 34450A
- Oscilloscope
  Manufacturer: Keysight
  Model: DSO-X 3014T
- Function Generator
  Manufacturer: Keysight
  Model: 33500B
- LCR-Meter
  Manufacturer: Keysight
  Model: U1733C

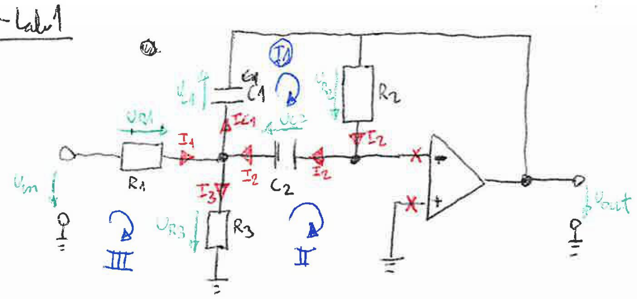
Graz, February 17, 2026 | Lorenz Buchinger | Agustin Sevil de Llobet

A Appendix

A.1 Calculations

A.1.1 Active Band-pass Filter

ACD-Lab1



$I_2 = \frac{U_{OUT}}{R_2}$ Ⓘ $U_{out} + U_{C2} + U_{C1} = \emptyset$

$U_{C2} = \frac{U_{out}}{s C_2 R_2}$ $U_{C1} = -U_{out} - U_{C2}$

$I_{C1} = \frac{U_{C1}}{\frac{1}{sC1}} = -sC_1\left(U_{out} + \frac{U_{out}}{s C_2 R_2}\right)$

Ⓘⓘ $U_{C2} + U_{R3} = \emptyset \rightarrow U_{R3} = -U_{C2}$ Ⓘⓘⓘ $U_{R1} + U_{R3} - U_{in} = \emptyset \rightarrow U_{R1} = U_{in} - U_{R3}$

$I_3 = -\frac{U_{C2}}{R_3} = -\frac{U_{out}}{s C_2 R_2 R_3}$ $I_1 = \frac{U_{R1}}{R_1} = \frac{U_{in}}{R_1} + \frac{U_{out}}{s C_2 R_2 R_1}$

$I_1 + I_2 = I_3 + I_{C1} \rightarrow \frac{U_{in}}{R1} + \frac{U_{out}}{s C_2 R1 R_2} + \frac{U_{out}}{R_2} = -\frac{U_{out}}{s C_2 R_2 R_3} - U_{out} \cdot SC1 \cancel{+} \frac{U_{out} \cdot C1}{C_2 R_2} \rightarrow$

$\rightarrow \frac{U_{in}}{R1} = -U_{out}\left(\frac{1}{R_2} + \frac{C1}{C_2 R_2} + s C_1 + \frac{1}{s C_2 R_1 R_2} + \frac{1}{s C_2 R_2 R_3}\right) \rightarrow$

$\rightarrow s\frac{U_{in}}{R_1} = -U_{out} \cdot \left(s\frac{C1+C2}{C_2 R_2} + s^2 C_1 + \frac{R_1+R_3}{C_2 R_1 R_2 R_3}\right) \rightarrow \frac{U_{out}}{U_{in}} = -\frac{s \cdot \frac{1}{R_1}}{\frac{R_1+R_3}{C_2 R_1 R_2 R_3} + s\frac{C1+C2}{C_2 \cdot R_2} + s^2 C_1} \rightarrow$

$$\rightarrow \boxed{H(s) = -\frac{\frac{R_2 \cdot R_3}{R_1+R_3} \cdot s C_2}{1 + s\frac{R_1 R_3 (C1+C2)}{R_1+R_3} + s^2 \frac{R_1 R_2 R_3 C_1 C_2}{R_1 R_3}}}$$

Lowpass to band Pass $\rightarrow$ 2.5.4 Rakam Analogschaltungsdesign. $P \Rightarrow \frac{1}{\Delta\Omega}\left(P + \frac{1}{P}\right)$ mith $\frac{1}{\Delta\Omega} = Q \rightarrow S \Rightarrow \frac{s^2 + \omega_0^2}{Bs}$

$\omega_0 = 2\pi f_0$

$B = \frac{\omega_0}{Q}$

$H_{LP}(s) = \frac{1}{s^2 + as + b} \xrightarrow{\text{substitution}} H_{BP}(s) = \frac{1}{\left(\frac{s^2+\omega_0^2}{Bs}\right)^2 + a\left(\frac{s^2+\omega_0^2}{Bs}\right) + b} \rightarrow \cdot B^2 s^2 \rightarrow$

$\rightarrow H_{BP}(s) = \frac{B^2 s^2}{(s^2+\omega_0^2)^2 + a Bs(s^2+\omega_0^2) + b B^2 s^2} \rightarrow \begin{cases} (s^2+\omega_0^2)^2 = s^4 + 2\omega_0^2 s^2 + \omega_0^4 \\ a Bs(s^2+\omega_0^2) = a Bs^3 + a B\omega_0^2 s \\ b B^2 s^2 \end{cases} \rightarrow$ Polinomio

$\rightarrow D(s) = s^4 + aBs^3 + (2\omega_0^2 + bB^2)s^2 + aB\omega_0^2 s + \omega_0^4 \rightarrow$ Substitute "B" $\rightarrow$

$\rightarrow D_{std}(s) = s^4 + \frac{\omega_0}{Q}s^3 + \left(2\omega_0^2 + \frac{\omega_0^2}{Q^2}\right)s^2 + \frac{\omega_0^3}{Q}s + \omega_0^4$

Comparation coefficients

Contruct Poli 8° $D_i(s) = s^4 + \frac{\omega_{0i}}{Q} s^3 + \left(2\omega_{0i}^2 + \frac{\omega_{0i}^2}{Q^2}\right)s^2 + \frac{\omega_{0i}^3}{Q} s + \omega_{0i}^4 \rightarrow \omega_{01} = \alpha\,\omega_0 \quad \omega_{02} = \frac{\omega_0}{\alpha}$

$$\left.\begin{aligned} D_1(s) &= s^4 + \frac{\alpha\omega_0}{Q}s^3 + \left(2\alpha^2\omega_0^2 + \frac{\alpha^2\omega_0^2}{Q^2}\right)s^2 + \frac{\alpha^3\omega_0^3}{Q}s + \alpha^4\omega_0^4 \\ D_2(s) &= s^4 + \frac{\omega_0}{\alpha Q}s^3 + \left(\frac{2\omega_0^2}{\alpha^2} + \frac{\omega_0^2}{\alpha^2 Q^2}\right)s^2 + \frac{\omega_0^3}{\alpha^3 Q}s + \frac{\omega_0^4}{\alpha^4} \end{aligned}\right\} D_8(s) = D_1(s)\cdot D_2(s) \rightarrow$$

$$\rightarrow D_8(s)\left(s^4 + A_3 s^3 + A_2 s^2 + A_1 s + A_0\right)\left(s^4 + B_3 s^3 + B_2 s^2 + B_1 s + B_0\right)$$

$f_L = 4000\,Hz \quad f_u = 5000\,Hz \rightarrow f_0 \overset{\text{central frequ.}}{=} \sqrt{f_L f_u} = 4472{,}14\,Hz \qquad \omega_0 = 2\pi\cdot f_0 = 2\pi\cdot 4472'14 = 28'099{,}3$

$BW = f_u - f_L = 1000\,Hz \quad Q = \frac{f_0}{BW} = 4'47 \quad \Delta\omega_n = \frac{BW}{f_0} = 0'2236 \qquad$ All $C = 10\,mF$

↳ normalitzat Bandwidth

---

sustitution

$$P \Rightarrow \frac{1}{\Delta\Omega}\left(P + \frac{1}{P}\right) \text{ with } \frac{1}{\Delta\Omega} = Q \rightarrow H(P) = \frac{H_0}{1 + a_1\left(\frac{1}{\Delta\Omega}\left(P + \frac{1}{P}\right)\right) + b_1\left(\frac{1}{\Delta\Omega}\left(P + \frac{1}{P}\right)\right)^2} \rightarrow \rightarrow$$

$$\rightarrow \frac{H_0}{1 + \frac{a_1 P}{\Delta\Omega} + \frac{a_1}{\Delta\Omega P} + \frac{b_1 P^2}{\Delta\Omega^2} + \frac{2b_1}{\Delta\Omega^2} + \frac{b_1}{\Delta\Omega^2 P^2}} \rightarrow \cdot \frac{\frac{\Delta\Omega^2 P^2}{1}}{\frac{\Delta\Omega^2 P^2}{1}} \rightarrow$$

$$\rightarrow \frac{H_0\,\Delta\Omega^2 P^2}{\Delta\Omega^2 P^2 + \frac{a_1\Delta\Omega^{\cancel{2}} P^3}{\cancel{\Delta\Omega}} + \frac{a_1\Delta\Omega^{\cancel{2}} \cancel{P^2}}{\cancel{\Delta\Omega P}} + \frac{b_1 P^4 \cancel{\Delta\Omega^2}}{\cancel{\Delta\Omega^2}} + \frac{2b_1\cancel{\Delta\Omega^2}P^2}{\cancel{\Delta\Omega^2}} + \frac{b_1\cancel{\Delta\Omega^2 P^2}}{\cancel{\Delta\Omega^2 P^2}}} \rightarrow$$

$$\rightarrow \frac{H_0\,\Delta\Omega^2 P^2}{\Delta\Omega^2 P^2 + a_1\Delta\Omega P^3 + a_1\Delta\Omega P + b_1 P^4 + 2b_1 P^2 + b_1} \rightarrow \cdot \frac{\frac{1}{b_1}}{\frac{1}{b_1}} \rightarrow \frac{\frac{H_0\Delta\Omega^2 P^2}{b_1}}{\frac{\Delta\Omega^2 P^2}{b_1} + \frac{a_1\Delta\Omega P^3}{b_1} + \frac{a_1\Delta\Omega P}{b_1} + P^4 + 2P^2 + 1} \rightarrow$$

$$\rightarrow \text{Normalization} \rightarrow = \frac{\frac{H_0\Delta\Omega^2 P^2}{b_1}}{1 + \frac{a_1}{b_1}\Delta\Omega P + \left[2 + \frac{\Delta\Omega^2}{b_1}\right]P^2 + \frac{a_1}{b_1}\Delta\Omega P^3 + P^4}$$

Script: 1st order lowpass:

$$H(P) = \frac{\frac{H_0}{b_1}\Delta\Omega^2 P^2}{\left[1 + \frac{\alpha}{Q_i}P + (\alpha P)^2\right]\left[1 + \frac{1}{Q_i}\left(\frac{P}{\alpha}\right) + \left(\frac{P}{\alpha}\right)^2\right]}$$

denominator:

$$\left[1 + \frac{1}{Q_i}\left(\frac{P}{\alpha}\right) + \left(\frac{P}{\alpha}\right)^2\right] + \left[\frac{\alpha}{Q_i}P + \frac{\cancel{\alpha}\cdot P^2}{Q_i^2\cancel{\alpha}} + \frac{P^3}{\alpha Q_i}\right] + \left[(\alpha P)^2 + \frac{\alpha P^3}{Q_i} + P^4\right] \rightarrow P \text{ common factor} \rightarrow$$

$$\Rightarrow = 1 + P\left[\frac{1}{Q_i\alpha} + \frac{\alpha}{Q_i}\right] + P^2\left[\frac{1}{\alpha^2} + \frac{1}{Q_i^2} + \alpha^2\right] + P^3\left[\frac{1}{\alpha Q_i} + \frac{\alpha}{Q_i}\right] + P^4$$

2

Escaneado con CamScanner

Figure A.2: Page 2 of the hand written calculations for the active band-pass filter.

My funcion denominator:

$$1 + P\left[\frac{a_1}{b_1}\Delta\Omega\right] + P^2\left[2 + \frac{\Delta\Omega^2}{b_1}\right] + P^3\left[\frac{a_1}{b_1}\Delta\Omega\right] + P^4$$

P^1:

$$\frac{1}{Q_i\alpha} + \frac{\alpha}{Q_i} = \frac{a_1}{b_1}\Delta\Omega \rightarrow \frac{1+\alpha^2}{Q_i\alpha} = \frac{a_1}{b_1}\Delta\Omega \rightarrow \cdot\frac{1}{(1+\alpha^2)} \rightarrow \frac{1}{Q_i\alpha} = \frac{a_1}{b_1+b_1\alpha^2}\Delta\Omega \rightarrow$$

$$\rightarrow (\;)^{-1} \rightarrow Q_i\alpha = \frac{b_1+b_1\alpha^2}{a_1\Delta\Omega} \rightarrow \cdot\frac{1}{\alpha} \rightarrow Q_i = \frac{b_1+b_1\alpha^2}{a_1\Delta\Omega\alpha} \quad \text{(I)}$$

P^2:

$$\frac{1}{\alpha^2} + \frac{1}{Q_i^2} + \alpha^2 = 2 + \frac{\Delta\Omega^2}{b_1} \rightarrow \text{Sustitution } Q_i \rightarrow \frac{1}{\alpha^2} + \frac{1}{\left(\frac{b_1(1+\alpha^2)}{a_1\Delta\Omega\alpha}\right)^2} + \alpha^2 = 2 + \frac{\Delta\Omega^2}{b_1}$$

Betterworth eq: $a_1 = \sqrt{2}$, $b_1 = 1$ } Table values

$$\Delta\Omega = \frac{f_{max} - f_{min}}{f_0} = \frac{1000}{4472,1} = 0'2236 \qquad f_{m1} = \frac{f_0}{\alpha} = 4'131{,}53$$

$$f_0 = \sqrt{f_{min}\, f_{max}} = \sqrt{4000 \cdot 5000} = 4472{,}1\ Hz \qquad f_{m2} = f_0\alpha = 4'840{,}82$$

Solve eq (I) with Ti-nspire CX II-T for $\alpha \rightarrow \alpha_{BW} = 1'08244 \rightarrow Q_{iBW} = \frac{b_1 + b_1\alpha^2}{a_1\Delta\Omega\alpha} = 6'34441$

$$B = \frac{f_{m1}}{Q_i} = \cancel{606'244}\ 651'208$$

$$C_1 = C_2 = 10\,mF \rightarrow C \qquad B = \frac{C_1 + C_2}{2\pi R_2 C_1 C_2} = \frac{2C}{2\pi R_2 C^2} = \frac{1}{\pi R_2 C} \rightarrow R_2 = \frac{1}{\pi C B} = 48'8799\ K\Omega$$

$$H_r = -\frac{CR_2}{2CR_1} = -\frac{R_2}{2R_1} \rightarrow |H_r| = \frac{R_2}{2R_1} \Rightarrow R_1 = \frac{R_2}{2H_r} = 5447{,}84\ \Omega$$

$$H_r = Q_i\Delta\Omega\sqrt{\frac{H_{mid}}{b_1}} = 6'34 \cdot 0'2236 \cdot \sqrt{\frac{10}{1}} = 4{,}48618$$

$$R_3 = \frac{R_1}{(R_1R_2 f_{m1}^2 - 1)(2\pi C)^2} = 303{,}59\ \Omega$$

BW2:

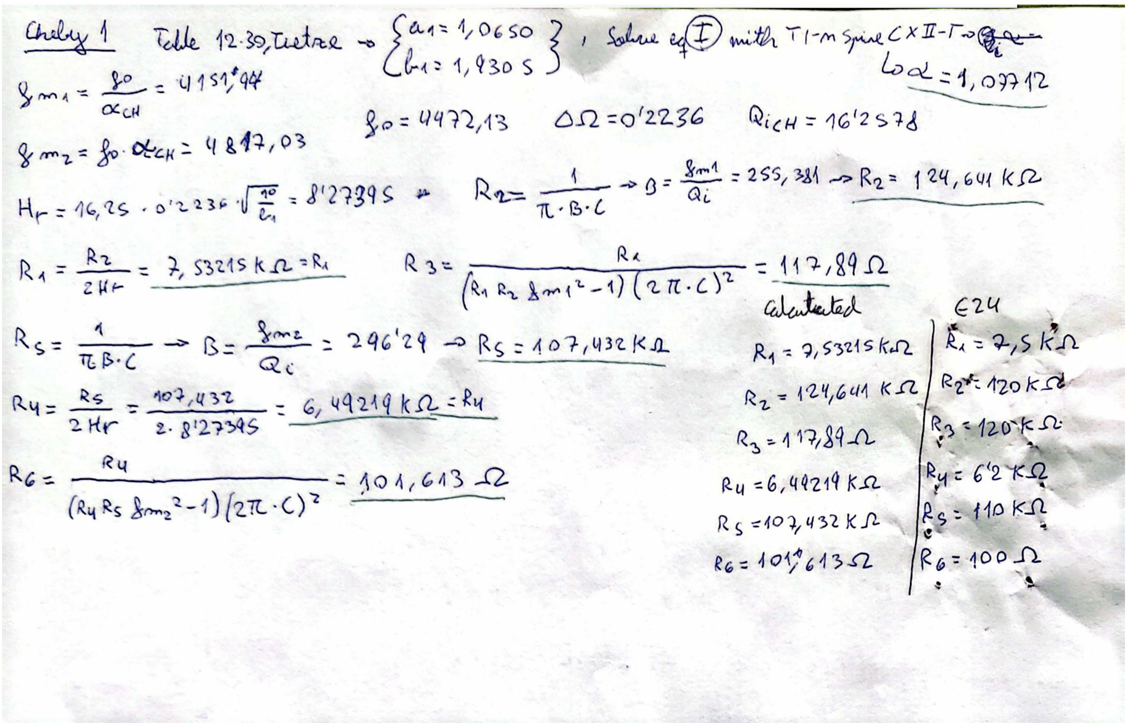
$$R_5 = \frac{1}{\pi C B} \rightarrow B = \frac{f_{m2}}{Q} = 763'005 \rightarrow R_5 = 41{,}7179\ K\Omega \qquad R_4 = \frac{R_5}{2H_r} = 4649{,}61\ \Omega$$

$$R_6 = \frac{R_4}{(R_4R_5 f_{m2}^2 - 1)(2\pi C)^2} = 259{,}107\ \Omega$$

Calculated	E24	
$R_1 = 5447{,}84\ \Omega$	$R_1 = 5'6\ K\Omega$	
$R_2 = 48879'9\ \Omega$	$R_2 = 47\ K\Omega$	
$R_3 = 303{,}59\ \Omega$	$R_3 = 300\ \Omega$	
$R_4 = 4649{,}61\ \Omega$	$R_4 = 4'7\ K\Omega$	$\rightarrow 4'6051\ K\Omega$
$R_5 = 41\,717{,}9\ \Omega$	$R_5 = 43\ K\Omega$	
$R_6 = 259{,}107\ \Omega$	$R_6 = 270\ \Omega$	

3

Figure A.3: Page 3 of the hand written calculations for the active band-pass filter.


Cheby 1 Table 12.39, Tietze → $\begin{cases} a_1 = 1{,}0650 \\ b_1 = 1{,}9305 \end{cases}$, Solve eq (I) with TI-nspire CX II-T → Lo $\alpha = 1{,}09712$

$f_{m1} = \frac{f_0}{\alpha_{CH}} = 4151{,}94$

$f_0 = 4472{,}13$ $\Delta\Omega = 0'2236$ $Q_{iCH} = 16'2578$

$f_{m2} = f_0 \cdot \alpha_{CH} = 4817{,}03$

$H_r = 16{,}25 \cdot 0'2236 \cdot \sqrt{\frac{10}{b_1}} = 8'27395$ $R_2 = \frac{1}{\pi \cdot B \cdot C} \to B = \frac{f_{m1}}{Q_i} = 255{,}381 \to R_2 = 124{,}641\ k\Omega$

$R_1 = \frac{R_2}{2H_r} = 7{,}53215\ k\Omega = R_1$ $R_3 = \frac{R_1}{(R_1 R_2 f_{m1}^2 - 1)(2\pi \cdot C)^2} = 117{,}89\ \Omega$

$R_5 = \frac{1}{\pi B \cdot C} \to B = \frac{f_{m2}}{Q_i} = 296'29 \to R_5 = 107{,}432\ k\Omega$

$R_4 = \frac{R_5}{2H_r} = \frac{107{,}432}{2 \cdot 8'27395} = 6{,}49219\ k\Omega = R_4$

$R_6 = \frac{R_4}{(R_4 R_5 f_{m2}^2 - 1)(2\pi \cdot C)^2} = 101{,}613\ \Omega$

calculated	E24
$R_1 = 7{,}53215\ k\Omega$	$R_1 = 7{,}5\ k\Omega$
$R_2 = 124{,}641\ k\Omega$	$R_2 = 120\ k\Omega$
$R_3 = 117{,}89\ \Omega$	$R_3 = 120\ k\Omega$
$R_4 = 6{,}49219\ k\Omega$	$R_4 = 6'2\ k\Omega$
$R_5 = 107{,}432\ k\Omega$	$R_5 = 110\ k\Omega$
$R_6 = 101{,}613\ \Omega$	$R_6 = 100\ \Omega$

Figure A.4: Page 4 of the hand written calculations for the active band-pass filter.

A.2 Matlab

A.3 Measurements

Table A.1: Measurements of ...

Frequency / Hz	Gain / dB	Phase / Degree
10	19,89	0,05
12,6	19,91	0,12
15,8	19,91	0,11
20	19,91	0,05
25,1	19,93	0,05
31,6	19,92	0
39,8	19,93	-0,05
50,1	19,88	-0,07

Table A.1: Measurements of ... (continuation)

Frequency / Hz	**Gain / dB**	**Phase / Degree**
63,1	19,88	-0,11
79,4	19,89	-0,18
100	19,89	-0,23
125,9	19,89	-0,36