IMPERIAL COLLEGE LONDON

BE3 – HBINST BIOMEDICAL INSTRUMENTATION

Coursework for Biomedical Instrumentation

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Contents

1 Part A - Analysis of Results and Discussion

1.1 Lab 1 - Passive Filters

- **Question 1.1**

The circuit of Fig.1 functions as a fourth-order band-pass filter, composed of a cascade of two low pass (LP) and two high pass (HP) filters with different cut-off frequencies. By subdiving the circuit into four isolated blocks, as shown in Fig.1, the transfer function of the i^{th} block ($H_i(s)$) is found. The transfer function of the circuit can be found by

$$H_{tot}(s) = H_1(s)H_2(s)H_3(s)H_4(s) \tag{1}$$

This is because the output of one block becomes the input to the next block. It should be noted that since block 1 and 2 have exactly the same components with the same connections, $H_1(s) = H_2(s)$. This also applies for block 3 and 4 so that $H_3(s) = H_4(s)$, and $H_1(s) \neq H_3(s)$. The transfer function of each block can be determined by utilising Kirchhoff's Current and Voltage Law, Ohm's Law and the expression for current across capacitor given by:

$$I(t) = C\frac{dV(t)}{dt} \tag{2}$$

Furthermore, the Laplace Transform of first derivative is required in the derivation of system's transfer function, given by

$$\frac{dV(t)}{dt} \overset{\mathcal{L}}{\leftrightarrow} sV(s) - V(0^-) \ , \ where \ V(0^-) \ is \ assumed \ to \ be \ 0. \tag{3}$$

To obtain the transfer function of the circuit, each block is analysed individually, assuming no current flow between the sub-circuits during calculations, as shown in Fig.2.

Blocks 1 and 2

Using KCL and KVL we can write the expression

$$\frac{V_{in} - V_{out}}{R_1} = C_1\frac{dV_{out}}{dt} \quad \overset{\mathcal{L}}{\rightarrow} \quad \frac{V_{in} - V_{out}}{R_1} = C_1 s V_{out}(s)$$

This can be re-arranged to give

$$V_{out}(s)(1 + R_1C_1s) = Vin(s) \ , \ so \ H_{1,2}(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{1}{1 + R_1C_1s}$$

This is the general equation for the transfer function of a first-order low-pass filter. The cut-off frequency, is the same as pole frequency for a first-order filter, thus it is given by $\omega_c = \frac{1}{R_1C_1} = 25.641 \ rad.s^{-1}$ or 4.08Hz, using the component values. This also means that when input frequency is greater than 4.08Hz, the output has a roll-off of -20dB/decade or -6dB/octave.

Blocks 3 and 4

Similarly, we can express

$$C_3\frac{dV_{in}}{dt} - C_3\frac{dV_{out}}{dt} = \frac{V_{out}}{R_3}Vin(s) \quad \overset{\mathcal{L}}{\rightarrow} \quad C_3s(V_{in}(s) - V_{out}(s)) = \frac{V_{out}}{R_3}$$

This can also be re-expressed as

$$C_3sV_{in}(s) = V_{out}(s)(\frac{1}{R_3} + C_3s) \ , \ so \ H_{3,4}(s) = \frac{s}{s + \frac{1}{C_3R_3}}$$

This is the general equation for the transfer function of a first-order high-pass filter. The cut-off frequency, is the same as pole frequency for a first-order filter, thus it is given by $\omega_c = \frac{1}{R_3C_3} = 0.5 \ rad.s^{-1}$ or 0.0796Hz, using the component values. This also means that when input frequency is below 0.0796Hz, the output has a roll-off of +20dB/decade or +6dB/octave.

Circuit 1 is essentially a cascade of two first-order low pass filters, forming a 2nd-order low-pass filter (roll-off -40dB/dec), with two first-order high-pass filters, forming a 2nd-order high-pass filter (roll-off

+40dB/dec). As a result, theoretically it is expected to have a resulting bandpass filter since the high-pass cut-off is lower than the low-pass cut-off frequency. The transfer function of Circuit 1 can be then found according to equation 1 to be:

$$H_{tot}(s) = \left(\frac{1}{1+R_1C_1s}\right)^2 \left(\frac{s}{s+\frac{1}{C_3R_3}}\right)^2 = \frac{s^2C_3^2R_3^2}{(1+R_1C_1s)^2(1+C_3R_3s)^2} \tag{4}$$

Note: The band-pass filter can also be considered as a cascade of a second-order low-pass and a second-order high-pass filter (two blocks instead of 4). In that case the transfer function of each block can be shown to be given by:

Low-Pass: $H(s) = \frac{\frac{1}{R_1^2C^2}}{s^2+s\left(\frac{3}{R_1C}\right)+\frac{1}{R_1^2C^2}}$, $f_{cutoff} = 4.08Hz$ (as before) and Q-factor = 1/3.

High-Pass: $H(s) = \frac{s^2}{s^2+s\left(\frac{3}{R_3C}\right)+\frac{1}{R_3^2C^2}}$, $f_{cutoff} = 0.0786Hz$ (as before) and Q-factor = 1/3.

These transfer functions can be multiplied together to give the overall transfer function of the band-pass filter:

$$H_{tot}(s) = \left(\frac{\frac{1}{R_1^2C^2}}{s^2+s\left(\frac{3}{R_1C}\right)+\frac{1}{R_1^2C^2}}\right)\left(\frac{s^2}{s^2+s\left(\frac{3}{R_3C}\right)+\frac{1}{R_3^2C^2}}\right) \tag{5}$$

- **Question 1.2** - Part a)

The frequency response from the measurements 1 and 2 were plotted on the same graph, as shown in Fig 6. Both plots show a low-pass filter response, with a cut-off frequency (in this case pole frequency) of around 2Hz (-3dB frequency) for measurement 1 and around 2.6Hz (-6dB frequency) for measurement 2. The results also show that the graph for measurement 2 has a steeper roll-off (-17.221dB/decade) than the graph of measurement 1 (-9.678dB/decade). In general, the roll-off is given by $-(20n)dB/decade$, where n is the filter order.

Measurement 1 plot is expected to give the magnitude plot of a first order passive low-pass filter (Block 1 in Circuit Fig.1). Although we expect to have a cutoff frequency of 4.08 Hz, we obtain -3dB at around 2 Hz. Furthermore, the roll-off of frequency response in measurement 1 in theory would be -20dB/decade (-6dB/Octave) but instead is computed to be -9.678dB/decade (-3.05dB/Octave).

Measurement 2, theoretically, corresponds to the output of a 2nd order low -pass filter created by the cascade of two first order low-pass filters. Since both first order blocks contain the same components such that $R_1C_1 = R_2C_2$, we expect that at the cut-off frequency (pole of each filter), the gain would be equal to $(\frac{1}{\sqrt{2}})^n$, where n is the order of the filter, so in this case this would be 1/2, i.e. -6.0206dB. Cut-off frequency is given by:

$$f_c = \frac{1}{2\pi\sqrt{R_1R_2C_1C_2}} \tag{6}$$

Therefore since $R_1 = R_2$ and $C_1 = C_2$, then we expect to have the same cut-off frequency as for individual first-order filters, i.e. cut-off of 4.08Hz. Additionally, the -3dB frequency for this second order filter is given by:

$$f_{-3dB} = f_c\sqrt{2^{\frac{1}{n}}-1} = 4.08\sqrt{2^{\frac{1}{2}}-1} = 2.63Hz \tag{7}$$

Furthermore, the roll-off measurement 2 in theory would be -40dB/decade (-12dB/Oct) for measurement 2, but it is obtained to be -17.221dB/decade (-5.58 dB/Octave).

The differences between the theoretical and experimental values for roll-off and cut-off frequencies can be explained by considering that during measurements the blocks in the circuit are not physically isolated. Due to signal input (from the signal generator) current flows to other parts of the circuit interfering with our results. The presence of this current outflow to high-pass blocks (3 and 4 in Fig.1), affects both the measurement for cut-off frequency of low-pass filter and roll-off after this frequency. The effect of this current outflow can be determined by either deriving the transfer function in the case that the filter blocks are not considered isolated or using PSpice simulations, which was done. Simulations for both measurements show that the practical roll-off increases to -19.24dB/decade for measurement 1 and -39.31dB/decade for measurement 2 after a frequency higher than 10Hz. This was not shown in the experimental graphs since we only considered frequency range up to 10 Hz in measurements. Finally, although the components of the two first order filters have the same nominal values, the true values might be different.

- **Question 1.2** - Part b)

By performing the modification shown in Fig.3, we will have a cascade of three first order filters with cut-off at 4.08Hz, since all components have the same nominal value in Block A. We are interesting in attenuation between 12Hz and 24Hz, which means that Block A(third-order low-pass filter) is the system causing this amplitude drop. If we assume three isolated first-order filter blocks, then we expect to have a roll-off of -60dB/dec in the stop-band region. Since 12 Hz ($> 10Hz$) is well within the stopband region ($\gg f_c$), we expect to have an attenuation of -60dB/decade which is equivalent to -18dB/Octave (3[-6dB/Octave]), so we will have an attenuation difference of -18dB from 12 Hz to 24Hz. This was also verified through simulations where the roll-off was found to be equal to -57.15dB/decade or -16.758dB/Oct as we expected.

- **Question 1.3**

The frequency response from the measurements 1 and 2 were plotted on the same graph, as shown in Fig 9. It should be noted that signal generator could not generate very low frequencies accurately, therefore we expect results for 0.05Hz to not be very reliable.
Both plots show high-pass filter characteristics, with a cut-off frequency (in this case pole frequency) of around 0.3Hz (-3dB frequency) for measurement 1 and around 0.3Hz (-6dB frequency) for measurement 2. The results also show that the graph for measurement 2 has a steeper roll-off (13.239dB/decade) than the graph of measurement 1 (7.657dB/decade).
Measurement 1 plot is expected to give the magnitude plot of a first order passive high-pass filter (Block 3 in Circuit Fig.1). Although we expect to have a cutoff frequency of 0.08Hz, we obtain -3dB at around 0.3 Hz. Furthermore, the roll-off of frequency response in measurement 1 in theory would be +20dB/decade (+6dB/Octave) but instead is computed to be 7.657dB/decade.
Measurement 2, theoretically, corresponds to the output of a 2nd order high-pass filter created by the cascade of two first order high-pass filters. According to justification in Question 1.2 a), the theoretical cut-off (pole) frequency is equal to $f_c = 0.0796Hz$ and -3dB frequency equal to 0.124Hz.
Differences between actual and theoretical values can be explained as well by the fact that the Low-pass and High-pass parts are not isolated in practice, cuasing current outflow to other parts of circuit as explained in Question 1.2 a). The input voltage set at the junction between LP and HP blocks, can create a current backwards towards the low-pass filter. Results were also verified using PSpice simulations.

- **Question 1.4**

The response across the entire circuit of Fig.1, is expected to have the characteristic of a band-pass filter, since it is essentially a cascade of two first-order low-pass and two first-order high pass filters, whose high-pass cut-off frequency is lower than high-pass cut-off frequency. The theoretical values of these cut-off frequencies were computed to be 0.0796Hz and 4.08Hz respectively. Therefore, if we vary frequency of input sinusoid voltage, we expect amplitude to increase with a +40dB/decade up to 0.0796Hz, then remain constant at 0dB (band-pass region) until 4.08Hz and then decrease with a roll-off of -40dB/decade.
In practice, the frequency response of the circuit was obtained by considering frequencies in the range

of 0.05 and 10 Hz, so that the pass-band and stop-band could be seen clearly. Fig.10 illustrates the results obtained. Since in practice the low-pass cut-off frequency is lower than the theoretical and high-pass cut-off frequency is higher than theoretical, this means that the actual filter's bandwidth is shorter.

- **Measurement required for Lab 2**

The frequency at which a 50mV peak-to-peak signal attenuates to 3mV peak-to-peak at the output was found to be 16.2 Hz. The method of obtaining this result, in the most accurate way due to resolution errors of oscilloscope, was to input a 5V input voltage and find the frequency when output reduces to 0.3V.

1.2 Lab 2 - Op-amp Amplifier

- **Question 2.1**

The circuit of Fig.11 is a single-supply, non-inverting amplifier circuit whose output varies between 0V and 3V (lower and upper op-amp saturation voltage levels or rails). The transfer function of this circuit can be derived by firstly considering that input and output voltages are measured with respect to ground and not with respect to 1.5V.

By assuming that op-amp is ideal with infinite input impedance and infinite (or very large) open-loop gain, then we can identify a virtual short between the two op-amp inputs, thus $V^- \approx V_{input}$. This means that non-inverting input is equal to inverting input. By using the simplification that $R_1 = 120\Omega$ and $R_2 = 120k\Omega = 1000R_1$, and KCL and KVL relations, we can obtain that:

$$\begin{aligned} \frac{1.5 - V_{input}}{R_1} &= \frac{V_{input} - V_{output}}{R_2} = \frac{V_{input} - V_{output}}{1000R_1} \\ 1000(1.5 - V_{input}) &= V_{input} - V_{output} \\ V_{output} &= V_{input} + 1000V_{input} - 1500 \\ V_{output} &\approx 1000(V_{input} - 1.5) \end{aligned} \tag{8}$$

We can set input and output voltages as $V_{input} = V'_{input} + 1.5$ and $V_{output} = V'_{out} + 1.5$, so that V'_{input} and V'_{output} are measured with respect to 1.5V (which is also the voltage mid-value of op-amp rails). In this case, the transfer function is given by

$$H(s) = \frac{V'_{out}(s)}{V'_{in}(s)} = 1001 \tag{9}$$

- **Question 2.2**

The effect of DC offset on the circuit operation was investigated by having as input signal a 1V peak-to-peak sinusoid with DC offset values of 0,1,1.5,2,2.5V, shown in Fig. 13-17.

Importance of DC offset for this particular circuit can be explained by considering the lower and upper bounds for input voltage total amplitude in order for amplifier amplitude to not be saturated (without "hitting" the rails). Therefore, DC offset can causes voltage output of the amplifier to be either saturated at V^-_{supply} or at V^+_{supply}, which means that system will be behaving in a non-linear manner (acts like a switch/trigger). The minimum amplitude of output voltage can be assumed to be (for an ideal op-amp) 0V, i.e. equal to the negative input supply voltage which is 0V (ground). In that case

$$\begin{aligned} 1001V_{in} - 1500 \geq 0 \\ V_{in} \geq \frac{1500}{1001}V = 1.4985V \end{aligned}$$

In other words, if $V_{in} \geq 1.4985$, this will ensure that amplifier output is not saturated. The maximum amplitude of output voltage can be assumed to be equal to the positive input supply voltage, i.e. 3V. This means that

$$1001V_{in} - 1500 \leq 3$$
$$V_{in} \leq \frac{1503}{1001}V = 1.5015V$$

Therefore, V_{in} must be less than or equal to 1.5015V to prevent saturation of output.
To sum up, saturation is prevented if

$$1.4985V \leq V_{in} \leq 1.5015V \tag{10}$$

Otherwise, if $V_{in} > 1.5015V$, output voltage will be 3V and if $V_{in} < 1.4985V$, then output will be 0V. In Fig.13, the maximum possible input voltage value is 1V, which means that the output voltage is expected to be saturated at 0V. In Fig. 14, the maximum possible input value is 1.5V and minimum is 0.5V. Therefore it is expected to not have saturation when input voltage approached 1.5V. However, the DC offset used was not exactly 1V but 0.992V, which means that even in the maximum possible input voltage, the output voltage will be 0V. Similarly, this is what happened in Fig. 16, where the DC offset was 1.99V instead of 2V, so output remained saturated at 3V. Interestingly, for a DC offset of 1.5V (Fig.16), the maximum input value is 2V and minimum is 1V, but since the range of voltage values to prevent saturation is very narrow, the output seems to change state from 0V to 3V. This is because a change of 3mV occurs almost instantaneously. Finally, for Fig. 17, the minimum possible input value was 2V, so at all times output will be 3V.
The device being implemented depends on this amplifier to be linear, so that with the correct DC offset (in this case 1.5V), small-amplitude signals (peak-to-peak 3mV or less) can be amplified (linear region of operation utilised).

- **Question 2.3** - Part a)

The outputs for measurements 5,6 and 7 of lab practical manual are shown in Figs. 18,(19-21) and (22, 23) respectively. The measurements were taken after transient period had passed.

- **Question 2.3** - Part b)

Measurement 5 (pg.21) in Lab practical instructions, corresponds to Fig. 18 in the results. In lab 1, we were asked to find the frequency at which 50mV peak-to-peak input gets attenuated to 3mV peak-to-peak in the output of the 4th order band-pass filter, which was found to be 16.2Hz. In Lab 2, measurement 5, a testing signal was used with 0V DC offset (above ground) with amplitude 50mV peak-to-peak and frequency 16.2Hz, as an input to the combined circuit of band-pass filter and non-inverting amplifier. In theory, this signal is expected to get attenuated to 3mV at the output of the filter and then become the input to the non-inverting amplifier. Since the maximum and minimum possible values of this signal are within the linear region of operation the amplifier (as explained in previous section), we expect to have an amplification of a factor of 1001. As a result, the output of the circuit would be theoretically 3.003V peak-to-peak.
Fig. 18 shows that the output of the circuit is 2.76V which is very close to the expected value, corresponding to a gain of 920 (in the same order of magnitude as true value). The deviations from the true values can be explained by the fact that input signal had peak-to-peak amplitude of 48mV, input frequency is 16.12Hz and from the fact that amplifier circuit is not ideal (affected also by component tolerances).

- **Question 2.3** - Part c)

Measurement 6 (pg.21) in Lab practical instructions, corresponds to Fig. 19-21 in the results. The three figures show that the output is exactly the same irrespective of the input DC offset. This is in

contrast with the results from Question 2.2, since in this case DC offset has no effect on the results. The difference in this case is that input signal first passes through the band-pass filter, which has a reference voltage of 1.5V, before entering the non-inverting amplifier. Due to presence of high-pass filter (2nd order) and 1.5V voltage reference (above ground), any DC offset values, above or below 1.5V, are are being significantly attenuated. Therefore the band-pass output signal will be shifted to a mean value of 1.5V as shown in the figures. So input signal to amplifier will contain only a DC offset of 1.5V, as shown by the mean value of blue trace in the figures.
Additionally, a transient (or start-up) period was observed and all the measurements were taken after the output signal became steady. This transient period is caused mainly by the capacitors within the high-pass filter which start to charge up to their steady state voltage through the 2MΩ resistors. The purpose of these capacitors is to charge up and remove any DC offset above or below 1.5V (capacitor is already positioned at 1.5V). This excess DC offset causes the amplifier to be saturated, so transient period can be though of as the period required for capacitors to bring offset to 1.5V and make amplifier non-saturated. Low-pass filter has less significant effect since its product of total resistance and capacitance is much smaller than for High-pass filter.

- **Question 2.3** - Part d)

Results for measurement 7, where input was 50mV peak-to-peak amplitude voltage, were recorded in Fig.22 and 23 for which frequency was first varied to 26.2Hz and then to 6.2Hz. These recordings can be explained by dividing the system into three sub-systems: low-pass filter (2nd order), high-pass filter (2nd order) and non-inverting amplifier, shown in Fig. 12.
The two 2nd order filters, form together a band-pass filter with lower cut-off frequency equal to 0.0796Hz and upper cut-off frequency equal to 4.08Hz. The band-pass filter has a reference voltage of 1.5V, meaning that any offset below or above 1.5V will be removed. The output of the band-pass filter will then become the input of the non-inverting amplifier. This will operate in linear region if input voltage is within the range of 1.4985V and 1.5015V, resulting in amplification with closed-loop gain of 1001. Output voltage will have an offset of 1.5V. In the case of non-linear operation there is saturation of output voltage. It should be noted that the output is expected to be shifted in time, since transfer function of band-pass filter suggests a non-uniform phase shift distribution.
We have seen before that for 0 DC offset and frequency of 16.2Hz, an input signal of amplitude 50mV peak-to-peak will attenuate to 3mV amplitude. Since 6.2Hz, 16.2Hz and 26.2Hz are above the low-pass cut-off frequency (4.08Hz), we expect a roll-off of -40dB/decade for all these frequencies. As a result, we expect higher attenuation for input signal frequency of 26.2Hz compared to 6.2Hz. For the case of 26Hz input signal, peak-to-peak voltage out is equal to 1.36V, with mean (DC offset 1.44V). In this case, no saturation is expected since amplifier input is 1.36V/1001=1.4mV amplitude, which is within the linear region of amplifier. However, for 6.2Hz, the attenuation is much smaller resulting in saturation of amplifier. This means that for this frequency the magnitude is above 3mV peak-to-peak. When its maximum value increases above 1.5015V, the output saturates to 3V and when minimum decreases below 1.4985V, there is saturation at 0V.

1.3 Lab 3 Part 1 - Non-linear Circuits

- **Question 3.1** - Part a)

The outputs of measurements 1 and measurement 2 are shown in Fig. shown in Figs. 24-29.

- **Question 3.1** - Part b)

According to the data sheet of Schottky Diode (BAT83S), which is a non-linear component, it allows a forward current (I_f) of 30mA when the forward voltage (V_f) is above 1V (Fig. 30). However, this

component conducts current for low threshold voltage values of 0.2-0.25V. This component was used in the implementation of the circuit shown in Fig. 31.
DC offset shifts the voltage value around which the input signal voltage varies, therefore, determines whether or not the threshold voltage is exceeded and thus diode's forward current is not zero. In the case of 0V DC offset, and 2V peak-to-peak amplitude of input signal, the circuit's voltage output follows the shape of input voltage only when the threshold voltage is exceeded (Fig.24). In this case, non-zero forward current creates a voltage drop across the diode and the parallel impedance of RC circuit. For the case of -1V DC offset (Fig.28), the input signal amplitude varies between 0V and -2V, therefore at all times this will be below the threshold value so Schottky diode is non-conductive (reverse biased) and output voltage is 0V. Finally, for the +1V DC offset (Fig.29), the amplitude of the input signal varies between 2V and 0V, thus for most of the time it is above the threshold value. It should be noted that in Fig.24 and 29, when input voltage decreases (at the negative slope region of its peak) the output voltage decays exponentially. This is due to the RC circuit, from which output voltage is measured from. When the voltage across diode falls below the voltage across the charged capacitor, then capacitor starts to discarge exponentially until disrupted by the next increase of input voltage above threshold.

- **Question 3.1** - Part c)

Input signal frequency affects the shape and peak-to-peak amplitude of the voltage output of circuit in Fig.31. Fig.24 and Fig.27-29, show that the higher the input signal frequency the lower the peak-to-peak amplitude of the output voltage, resulting in rectification of the input signal and thus becomes effectively an envelope detector. This is important since our input signal has frequency 32kHz, and this circuit can be used to detect envelope of amplitude-modulated signal. The effect of frequency on these records can be explained by first considering the parallel connection of Resistor (R) and Capacitor (C) in Fig.31. The total impedance of this part of the circuit is

$$Z_{parallel} = \frac{Z_R Z_C}{Z_R + Z_C} = \frac{R\frac{1}{j\omega C}}{R + \frac{1}{j\omega C}} = \frac{R}{1 + j\omega CR} \tag{11}$$

As frequency (ω) decreases towards 0, $Z_{parallel}$ approaches the value of R, which increases output voltage peak-to-peak amplitude. For the case that frequency (ω) increases towards infinity, $Z_{parallel}$ decreases to 0Ω thus output voltage decreases. This means that when forward bias voltage is above the threshold for the diode (i.e. forward current is non-zero), then the RC-circuit acts as a low-pass filter, since what we are measuring is the Voltage across this parallel impedance. In the case of sub-threshold input voltage, the diode acts as an open-circuit , so there is no current through to charge the capacitor.
It can be seen that the higher the frequency of the signal, the lower the capacitor charging time and discharging of capacitor occurs at earlier times, starting from higher voltage values. This is because the input signal decreases in a much higher rate than the rate at which capacitor discharges. Voltage across capacitor becomes higher than input voltage at earlier times, thus decay of output voltage is dominated by the discharging capacitor. Output voltage remains high since the input signal rises above the threshold value faster.
The time constant (τ) for the RC-circuit can be estimated to be $\tau = RC$, and since R=47kΩ and C=100nF, $\tau = 4.7ms$. This is equivalent to a frequency value of $\frac{1}{\tau}$, which gives 212.8Hz. The obtained frequency value, shows that above 212.8Hz, the amplitude of Vout will be significantly attenuated, as shown in Fig. 32, thus confirming the low-pass characteristics of the circuit.

1.4 Lab 3 Part 2 - Unguided Design

In this section, the component topology for implementing circuit in Fig.33 was designed together with connections to link it to the rest of the circuit. The design was approved by an invigilator and is shown in Fig.35.

textbullet **<u>Question 3.2</u>** - Part a)
The two 220pF capacitors used in the circuit, shown in Fig.33, are used as AC Coupling capacitors, to remove any DC offset above or below the reference voltage of 1.5V. In the first stage of circuit the RC-connection, between a 1μF capacitor and a 100kΩ resistor connected to 1.5V, aims to remove any excess offset bias due to noise in the circuit's input. However, additional DC offset may arise due to op-amp output or skin electrode connections (noise), which requires the two capacitors. These capacitors must be low in value in order to also attenuate noise frequencies of 50-60Hz, but large enough to not attenuate significantly the signals of interest.

- **<u>Question 3.2</u>** - Part b)

We will now consider what happens to current flow through the wires attached to the electrodes, when a square wave from 0V to 1V volt and of frequency 32 kHz is given as input to circuit shown in Fig.34. To begin with, the input voltage first passes through the RC-circuit prior to the non-inverting terminal on the op-amp. By assuming infinite input impedance of the op-amp, letting RC-circuit output being named V_{HP}, and using KCL:

$$\frac{V_{HP} - 1.5}{R} = C\frac{d(V_{in} - V_{HP})}{dt} \tag{12}$$

which is equivalent to the following, using Laplace transform

$$\begin{aligned} V_{HP} - 1.5 &= CRsV_{in} - CRsV_{HP} \\ V_{HP} &= \frac{CRsV_{in} + 1.5}{1 + CRs} \end{aligned} \tag{13}$$

Therefore, as $s \to 0$, $V_{HP} \to 1.5V$ and $s \to \infty$, $V_{HP} \to V_{in}$, which means it is a high-pass filter. This high-pass filter additionally brings any DC offset value to 1.5V, meaning that non-inverting input (V^+) of op-amp is a square wave of 32kHz frequency, 1.5V DC offset and 1V peak-to-peak (varies from 1V to 2V).
By assuming that the op-amp is ideal, operating in linear region with infinite gain, and by identifying the negative feedback connection, we can say that $V_{HP} = V^+ \approx V^-$, due to virtual short.
By also, assuming that the impedance across electrodes (total impedance of electrodes and skin) for 32kHz is around 1kΩ, then we can obtain the parallel resistance (Z_{par}) of 1kΩ resistor and impedance across electrodes to be 0.5kΩ. Using KCL, we can obtain an expression for op-amps output voltage:

$$\frac{V_{HP} - 1.5}{6.6k\Omega} = \frac{V_{out} - V_{HP}}{Z_{par}k\Omega}$$

This can be simplified to give

$$V_{out} \approx 1.076V_{HP} - 0.114 \tag{14}$$

Therefore, when V_{HP} is 1V in amplitude, then $V_{out} \approx 0.962V$ and when when V_{HP} is 2V in amplitude, then $V_{out} \approx 2.038V$. Using these values in Ohm's Law, the current flow (I) through wires attached to skin electrodes is given by

$$I = \frac{\Delta V}{R} = \frac{V_{out} - V_{HP}}{R} = \begin{cases} -0.038mA & , for \;\; V_{HP} = 1V \\ 0.038mA & , for \;\; V_{HP} = 2V \end{cases} \tag{15}$$

Therefore, injected current in the electrodes varies in a square manner between ±0.038mA, with frequency 32kHz. This is the current whose magnitude will be modulated by the varying respirative impedance, to generate a varying voltage component.

- **Question 3.2** - Part c)

The circuit of Fig.34, is therefore the current injecting (or Amplitude Modulation) stage of the impedance respirometric system. This stage is responsible for filtering first the input voltage to remove any DC offset not equal to 1.5V, and then generate the injected currents that vary with the same frequency as input voltage. The magnitude of these currents will be modulated by the varying respirative impedance of lung tissue, to generate a varying voltage component. Since this voltage component is measured and the injected current is known, the respirative impedance change can be obtained.

- **Question 3.3** - Part a)

The transfer function of circuit shown in Fig. 36 is derived. Firstly, we call the parallel impedance of the 33pF capacitor (C1) and 200kΩ resistor (R) as Z_{par}, whose expression was found in Question 3.2 to be $Z_{par} = \frac{R}{1+C_1Rs}$. Note that C_2 is the 220pF capacitor. By assuming infinite input impedance to op-amp terminals, we apply KCL to both inverting and non-inverting branch:

Inverting Branch - Where input voltage is called V_1

$$\begin{aligned} C_2\frac{d(V_1 - V^-)}{dt} &= \frac{V^- - V_{out}}{Z_{par}} \quad and\ using\ Laplace\ Transform\ , \\ C_2sV_1 - C_2sV^- &= \frac{V^- - V_{out}}{Z_{par}} \end{aligned} \tag{16}$$

Non-inverting Branch - Where input voltage is called V_2

$$\begin{aligned} C_2\frac{d(V_2 - V^+)}{dt} &= \frac{V^+ - 1.5}{Z_{par}} \quad and\ using\ Laplace\ Transform\ , \\ C_2sV_2 - C_2sV^+ &= \frac{V^+ - 1.5}{Z_{par}} \end{aligned} \tag{17}$$

Due to negative feedback, there is virtual short in input terminals and thus $V^+ \approx V^-$. Therefore, by subtracting equation 16 from 17, we obtain that

$$\begin{aligned} C_2s(V_2 - V_1) &= \frac{V_{out} - 1.5}{Z_{par}} \\ V_{out} &= Z_{par}C_2s(V_2 - V_1) + 1.5,\ \ ,and\ using\ the\ definition\ of\ Z_{par} \\ V_{out} &= \frac{RC_2s}{1 + C_1Rs}(V_2 - V_1) + 1.5 = A_d(V_2 - V_1) + 1.5 \end{aligned} \tag{18}$$

- **Question 3.3** - Part b)

The circuit of Fig. 36 is a subtractor circuit which amplifies the potential difference between non-inverting and inverting terminals by a differential gain A_d. This differential gain is not constant but is frequency-dependent. In fact, as frequency decreases to zero, differential gains becomes zero and output voltage is equal to 1.5V. On the other hand, as frequency increases, differential gain becomes equal to $\frac{C_2}{C_1} = \frac{220pF}{33pF} = 6.67$ so output voltage is 6.67 times the differential input with a DC offset of 1.5V. The behavior of the differential gain is a characteristic of a first order High-Pass filter, for which output is always non-negative and pass-band gain is 6.67 (16.5dB). This filter also was shown to have a theoretical cut-off frequency, i.e. frequency at which gain is 3dB below pass-band gain, of approximately 23.470kHz.

- **Question 3.3** - Part c)

Within the impedance respirometric system, the circuit allows amplification of the potential difference of the two skin electrodes and is therefore involved in the signal reading stage. This potential difference would then allow for calculation of respirative impedance. Since carrier current is at 32kHz, the circuit ensures amplification of the potential difference and attenuation of low frequency noise.

- **Question 3.4**

In this question, the common-mode attenuation (A_c) and differential gain (A_d) are measured in order to obtain the CMRR of the circuit in Fig.36. By calling the non-inverting input as V_2 and inverting input as V_1, the amplifier output voltage (V_{out}), for this specific circuit in Fig.36, is given by:

$$V_{out} = A_d V_{diff} + A_c V_{common} + 1.5 = A_d(V_2 - V_1) + A_c \left(\frac{V_2 + V_1}{2}\right) + 1.5 \tag{19}$$

- **Question 3.4** - Part a)

The purpose of this section was to measure the common mode attenuation of the circuit in Fig.36. To find A_c in practice, V_{diff} is set to zero by setting $V_2 = V_1 = V \neq 0$. Therefore, this means that

$$V_{out} = A_c V_{common} + 1.5 = A_c(V) + 1.5 \tag{20}$$

Therefore, by considering only the peak-to-peak values of V_{out} (to get rid of the 1.5 DC offset of output), then

$$A_c = \frac{V_{out}}{V_{common}} \tag{21}$$

For this measurement we used as input voltage (V_{common}) a 0.5V peak-to-peak sinusoid with 1.5V DC offset with varying frequency which was connected to both input terminals of the op-amp.

$$V_{common} = V = \frac{0.5}{2} sin(2\pi ft) + 1.5 \tag{22}$$

This means that we can write the following expression:

$$V_{out} = A_c V_{common} = A_c \left(\frac{0.5}{2} sin(2\pi ft)\right) + A_c(1.5) \approx A_c \left(\frac{0.5}{2} sin(2\pi ft)\right) \tag{23}$$

In the above equation we assumed that $A_c(1.5)$ is zero and this is due to the high-pass filter nature of the filter, which is shown by the plot of A_c against frequency is shown in Fig.37. Since we saw before that differential gain is frequency dependent, then we expect common mode attenuation (CMA) to also depends on frequency. As a result, expression for A_c , when considering peak-to-peak values at each frequency f:

$$A_c(f) = \frac{V_{out,p-t-p}(f)}{V_{common,p-t-p}(f)} = \frac{V_{out}(f)}{0.5} \tag{24}$$

For our measurements, frequencies in the range of 100Hz to 500kHz were used. Frequencies lower than 100 Hz were initially used, but due to high attenuation by the amplifier, they could not be distinguished from noise, so they were disregarded to make results more reliable.

- **Question 3.4** - Part b)

In this part of the question we were required to measure the differential gain of the amplifier circuit of Fig.36. Since the function generator in the lab could only produce one signal output, the non-inverting input (V_2) was set to 0.25V peak-to-peak with a DC offset of 1.5 V while the inverting input (V_1) was connected to 1.5V of the power supply. In other words

$$\begin{aligned} V_2 &= \frac{0.25}{2} sin(2\pi ft) + 1.5 \\ V_1 &= 1.5 \end{aligned} \tag{25}$$

Therefore this means that

$$V_{common} = \frac{V_2 + V_1}{2} = 0.5\left(\frac{0.25}{2}sin(2\pi ft) + 1.5 + 1.5\right) = \frac{0.125}{2}sin(2\pi ft) + 1.5$$
$$V_{diff} = V_2 - V_1 = \frac{0.25}{2}sin(2\pi ft) + 1.5 - 1.5 = \frac{0.25}{2}sin(2\pi ft) \tag{26}$$

To find A_d, once A_c is known, we use the expression for V_{out} for our amplifier circuit (equation 19), which is re-arranged to give:

$$A_d = \frac{V_{out} - A_c V_{common} - 1.5}{V_{diff}} = \frac{(V_{out} - 1.5) - A_c(\frac{0.125}{2}sin(2\pi ft) + 1.5)}{\frac{0.25}{2}sin(2\pi ft)} \tag{27}$$

For measurements, only the peak-to-peak values were used for each frequency, so the expression for A_d is simplified to:

$$A_d = \frac{V_{out,p-t-p}(f) - 0.125A_c}{0.25} \tag{28}$$

The measured values of A_d against frequency are shown in Fig. 38, for frequencies from 100Hz to 500kHz.

- **Question 3.4** - Part c)

Once the common-mode attenuation and differential gain are found, the common-mode rejection ratio (CMRR) is computed by:

$$CMRR(f) = 20log_{10}\left(\frac{A_d(f)}{A_c(f)}\right) = 20log_{10}\left(A_d(f)\right) - 20log_{10}\left(A_c(f)\right) \tag{29}$$

The CMRR values for different frequencies in the range of 100Hz to 500kHz are shown in Fig.39. The experimental CMRR plot was then compared to the CMRR plot in data sheet (Fig.40). The two plots are very different from each other and the main reason is that the one in data sheet for open-loop application of op-amp whereas in our case, the op-amp is used in the differential amplifier circuit. The common-mode attenuation of op-amp on its own is mainly due to transistor mismatches within the op-amp itself, whereas the common-mode attenuation of the differential amplifier (of which op-amp is part of) is caused mainly due to component tolerances, instead of transistor mismatches. It was shown that both the common mode attenuation and differential gain follow the shape of high pass filter, but after 100kHz, the op-amp itself malfunctions causing the CMRR to decrease instead of staying constant. This is probably due to decreasing open-loop gain of op-amp.

1.5 Final Lab

- **Question 4**

The electrical analogue model of the electrode-skin interface is shown in Fig. 41. To show that the impedance decreases with increasing injected frequency, Thevenin's theorem is used where only the Thevenin resistance is taken into consideration. From theory, to find Theverin Resistance (R_{thev}) between terminals A and B (labelled in the figure), all voltage sources are replaced by short circuits. Thevenin resistance in this circuit is composed of the parallel impedance of C_d and R_d (Z_{par1}), R_s, parallel impedance of C_e and R_e (Z_{par2}) that is in parallel with the parallel impedance of C_p and R_p (Z_{par3}), and R_u. These are annotated in Fig.41.

$$R_{thev} = Z_{par1} + R_s + (Z_{par2}||Z_{par3}) + R_u = Z_{par1} + R_s + Z_{par4} + R_u \tag{30}$$

The parallel resistances can be found by:

$$Z_{par1} = \frac{Z_{C_d}R_d}{Z_{C_d} + R_d} = \frac{\frac{R_d}{j\omega C_d}}{\frac{1}{j\omega C_d} + R_d} = \frac{R_d}{1 + j\omega R_d C_d}$$

Similarly,

$$Z_{par2} = \frac{R_e}{1 + j\omega R_e C_e} \quad and \quad Z_{par3} = \frac{R_p}{1 + j\omega R_p C_p}$$

Using the expressions for Z_{par2} and Z_{par3}, then Z_{par4} can be determined:

$$Z_{par4} = Z_{par2}||Z_{par3} = \frac{Z_{par2}Z_{par3}}{Z_{par2} + Z_{par3}} = \frac{\frac{R_e R_p}{(1+j\omega C_e R_e)(1+j\omega C_p R_p)}}{\frac{R_e}{1+j\omega C_e R_e} + \frac{R_p}{1+j\omega C_p R_p}} = \frac{R_e R_p}{R_e(1 + j\omega R_p C_p) + R_p(1 + j\omega R_e C_e)}$$

Therefore, the model's impedance (R_thev) is equal to:

$$R_{thev} = \frac{R_d}{1 + j\omega R_d C_d} + R_s + \frac{R_e R_p}{R_e(1 + j\omega R_p C_p) + R_p(1 + j\omega R_e C_e)} + R_u \tag{31}$$

Model's impedance decreases as injected frequency increases, which can be first seen from the fact that frequency (ω) is in the denominator of two out of four impedances composing R_{thev}. The frequency dependence of impedance can be seen by considering the limits of $\omega \to \infty$ and $\omega \to 0$.

$$R_{\omega\to 0} = \lim_{\omega\to 0} R_{thev} = R_d + R_s + \frac{R_3 R_d}{R_3 + R_d} + R_u \tag{32}$$

$$R_{\omega\to\infty} = \lim_{\omega\to\infty} R_{thev} = R_s + R_u + \frac{R_3 R_d}{R_3 + R_d} + R_u \tag{33}$$

Since $R_{\omega\to\infty} < R_{\omega\to 0}$, this shows that impedance of model decreases with increasing injected frequency.

- **Question 5**

The impedance respirometric system is composed of three main stages (or five sub-system blocks):

1. Current Injecting Stage
2. Signal Reading Stage
3. Analog Signal Processing Stage
   (a) Envelope Detector
   (b) Signal Filtering
   (c) Signal Amplification

As shown the Analog signal processing stage consists of three sub-stages. These five sub-bloxks, where each performs a distinct function, are labelled on the electrical circuit schematic (Fig.42) and actual implemented board (Fig.43). The function and crucial design aspects of each block are discussed below. Additionally, an important improvement for the entire design is the choice of components with improved tolerances. More improvements are discussed in each stage.

1. **Current Injection**
   This is the stage where the input carrier square wave of 1V peak-to-peak and frequency 32kHz is applied, that will be injected into the patient via the skin electrodes. The amplitude of this current will be modulated (Amplitude-modulation) by the varying respirative impedance of skin on chest to generate a varying voltage component. This system block is responsible to first filter the input voltage (through high-pass filtering) to remove any DC offset not equal to 1.5V with respect to ground. The op-amp used in this block then creates a small potential difference across the skin electrodes, which in turn causes a current flow square wave of 32kHz to be injected to the patient. This stage is the only one involving interface with patient therefore it must be designed in a way such that ensures safety of the patient. This block is connected to the signal reading stage through two 220pF capacitors, which ensure that any current pulled from the output of the op-amp in reading stage, is very low. This is crucial in order to maintain a low amplitude current (below 5mA) injected to patient (Higher amplitudes cause discomfort and even damage to patient).

Improvements to Current Injecting Stage:

- A 5mA fuse can be implemented between the circuit and skin electrodes to ensure that even in the case of system failure the patient would not be injected with more than 5mA current.
- The cut-off frequency of the high-filter part of the current injecting stage in the existing design is around 1.6Hz. Therefore, an improvement would be to increase this cut-off frequency to more than 1kHz, by changing the value of capacitor from 1μF to 1nF. Alternatively, a notch band-pass filter could be used to allow only frequencies close to carrier current frequency to pass. This is crucial because any frequency of injected oscillating current that is comparable to the cardiac rate (i.e. 0.5Hz $<$ f $<$ 3Hz) has the potential to cause arrhythmias and cardiac dysrhythmia and in extreme cases cardiac arrest.

2. **Signal Reading** This system block involves a subtractor circuit that allows amplification of the potential difference of the two skin electrodes in order to obtain a detectable varying voltage component that depends on the varying respirative impedance. Amplification is achieved by a frequency-dependent differential gain. In fact, as frequency decreases to zero, differential gains becomes zero and as frequency increases, differential gain becomes equal to 6.67 (non-unity gain High-pass filter with cut-off 23.47kHz). Since carrier current is at 32kHz, the circuit ensures amplification of the potential difference and attenuation of low frequency noise (50-60Hz). CMRR of system is at its maximum value at 32kHz (28dB). The two 220pF capacitors also remove any DC offset not equal to 1.5V.
   Improvements to Signal Reading Stage:

   - Choice of components with improved tolerances would increase the CMRR value, by reducing the common-mode error of the amplifier circuit. Additionally, op-amp must be chosen to have higher CMRR value on its own and maintain high open-loop gain for frequencies up to at least 50kHz.
   - Using the same number of components the block can be redesigned into two sub-blocks: a 2nd order high-pass filter and a differential amplifier (DA) circuit (involving only resistors). In this case, high-pass filter is more efficient. Also the DA circuit can be modified through the introduction of buffers in each input to increase input impedance , such that the sensor's output impedance is much less.

3. **Envelope Detector** The main function of this circuit block is to take the high-frequency amplitude modulated injected signal (output of signal reading stage) as input and to provide an output which is the envelope of the original signal. The capacitor in the circuit charges up on the rising edge of input signal and discharges slowly through the resistor when the input signal amplitude falls. The Schottky diode allows rectification of the signal. The time constant of this circuit is 4.7ms, which corresponds to 212.8Hz, thus when input frequency is 32kHz, the capacitor charges fast but does not discharge quickly, therefore its voltage follows closely to the peak values of the signal. It should be noted that a Schottky diode was used instead of a typical p-n junction diode, due to its low forward voltage and fast recovery time, allowing for high speed switching applications with lower noise levels.
   Improvements to Envelope Detector:

   - Due to high sensitivity of circuit block to DC offset, it would be ideal to introduce a high pass filter with voltage reference at 1.5V, before entering the envelope detector. Offsets not equal to 1.5V may cause signal to swing less than the diode forward voltage.

- Alternatively a precision rectifier (or super diode) composed of a diode and opamp can be used. This behaves like an ideal diode and rectifier. In this circuit, the forward voltage threshold value of the super diode is equal to the voltage threshold of the Schottky diode, divided by the gain of the op-amp.

4. **Signal Filtering** This sub-system acts a fourth-order unity-gain band pass filter, composed of a cascade of two first-order low pass and two first-order high-pass filters (or cascade of second-order low-pass with second-order high-pass filter) with cut-off frequencies of 0.0796Hz and 4.08Hz with a roll-off of $\pm 40dB/dec$. Filtering is required at this stage to remove any DC components of the original signal envelope, as well as any present noise components (50-60Hz). DC components of envelope need to be removed since these correspond to the baseline resistance of the skin and we are only interested in the changes due in resistance due to respiration. In practise, the cut-off frequencies were found to be 0.3Hz and 2Hz, while the gain in pass-band region was attenuated by 1.8dB. While these still satisfy the requirements for the device, it suggests that certain improvements can be made to achieve optimal operation.
Improvements to Signal Filtering:

   - In practice, cascading passive filters together to produce larger-order filters is not implemented accurately, since impedance of each filter affects its neighbouring network. An improvement to this is to use the corresponding active filters. Alternatively, voltage followers (buffer amplifiers) can be used to isolate the filter blocks. Another possible solution is to increase impedance of next filter block in the cascade by increasing Resistance value while decreasing Capacitance value (keeping product the same).
   - Instead of using cascade filters, ladder filters can also be used. These are less sensitive to small deviations from nominal component values, thus being more robust.

5. **Signal Amplification** This circuit is a non-inverting, single-supply amplifier, that allows amplification of the filtered signal envelope by a closed loop gain of 1001. The amplifier operates in its linear region for any input signal voltage in the range of $[1.4985, 1.5015]$V. If this condition is not satisfied, then the output voltage is saturated at 0V or +3V. The filtered signal envelope requires such high amplification, since respirative impedance variations are very small, causing small amplitude modulations of original signal.
Improvements to Signal Amplification:

   - To make the range of input signal voltage values for no saturation wider, the supply voltage of the op-amp can be increases to 0V and 5V. According to the OPA344 data sheet, the maximum supply voltage allowable is 5.5V. In order to make this modification, the reference voltage of the whole circuit must be set to 2.5V instead of 1.5V.
   - Another possible modification is to make the closed-loop gain of the signal amplification stage to be variable and not fixed. The gain of the circuit is given by $Gain = \frac{R_1+R_2}{R_1}$, where in original circuit $R_2 = 1000R_1$. Therefore, we can replace the resistor R_2, i.e. the 120kΩ resistor with a variable resistor.

1.6 Working Prototype Board - Measurements

The results of successful implementation and application of the impedance respirometric system are shown in Fig.44-48. The electronic board was tested using M98C Respirometric Simulatior with Variation=10Ω and Base impedance = 10kΩ, for different breathing rates.

2 Part B - Recordings and Graphs

2.1 Lab 1 - Passive Filters

Question 1.1

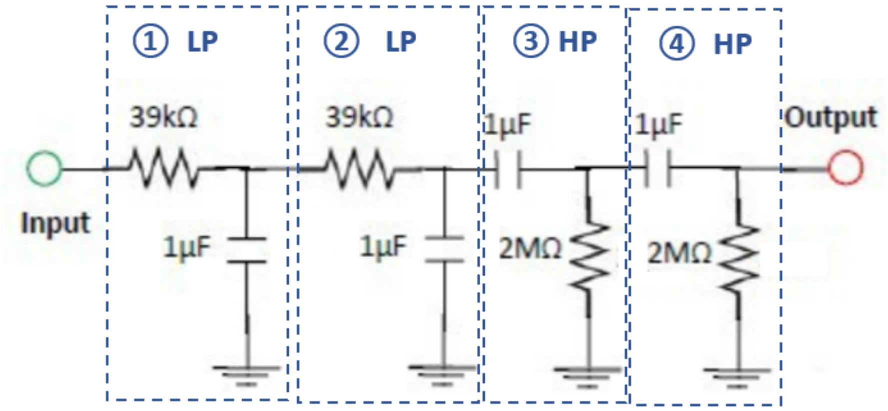
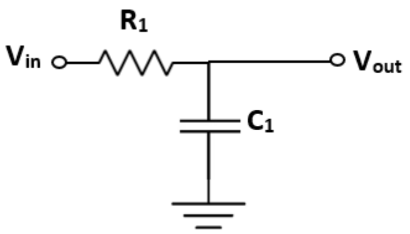
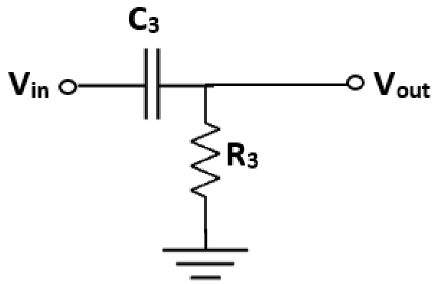


Figure 1: Circuit for Lab 1



(a) Block 1 (and 2)



(b) Block 3 (and 4)

Figure 2: Circuit blocks analysed

Question 1.2

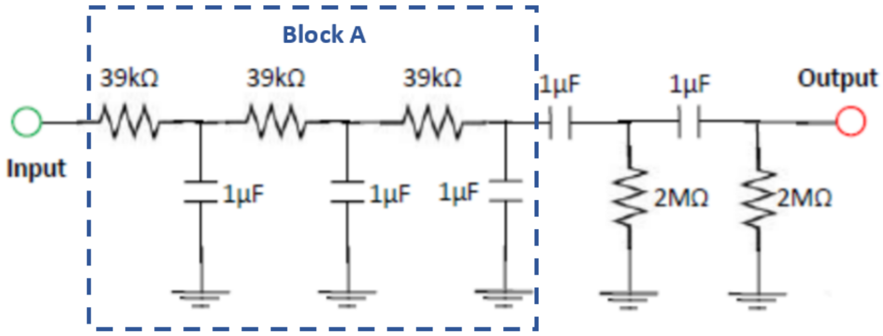


Figure 3: Extending circuit from Lab 1

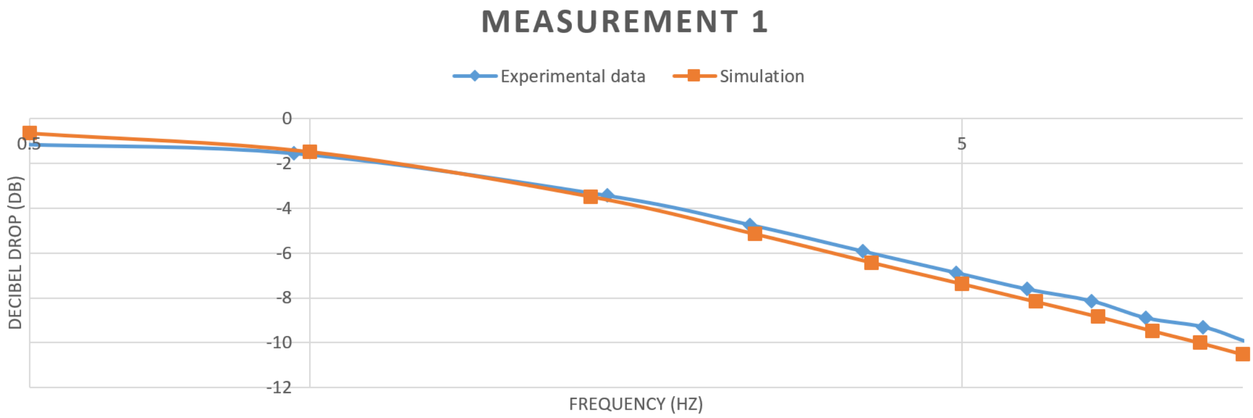


Figure 4: Measurement 1 - Measurements taken using a sinusoid of 1V peak-to-peak amplitude, in the frequency range of 0.5Hz to 10Hz. Resulting sinusoid amplitude drop was recorded in decibels (dB) on log-frequency axis. Circuit's behaviour was also simulated using PSpice, and simulation results were plotted for comparison.

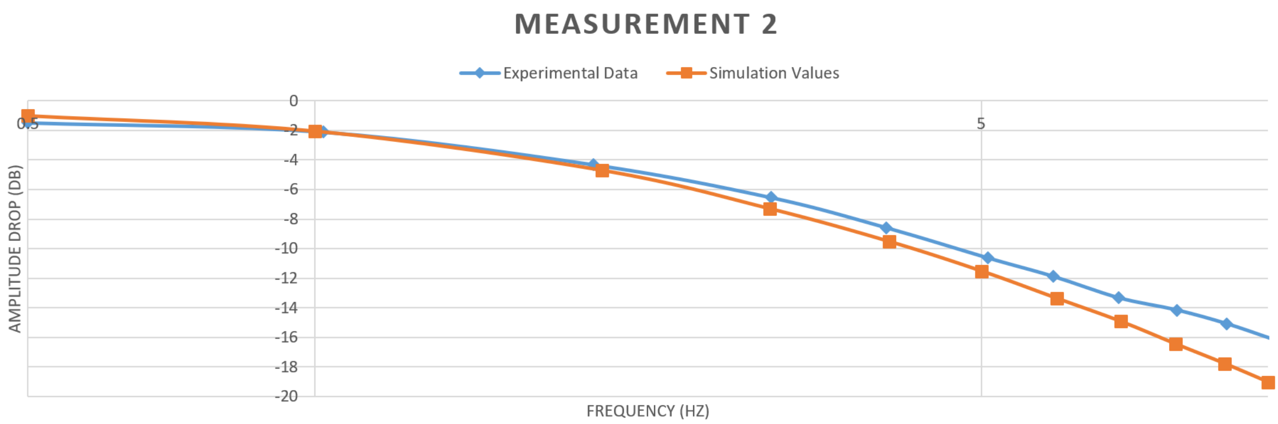


Figure 5: Measurement 2 - Measurements taken using a sinusoid of 1V peak-to-peak amplitude, in the frequency range of 0.5Hz to 10Hz. Resulting sinusoid amplitude drop was recorded in decibels (dB) on log-frequency axis. Circuit's behaviour was also simulated using PSpice, and simulation results were plotted for comparison.

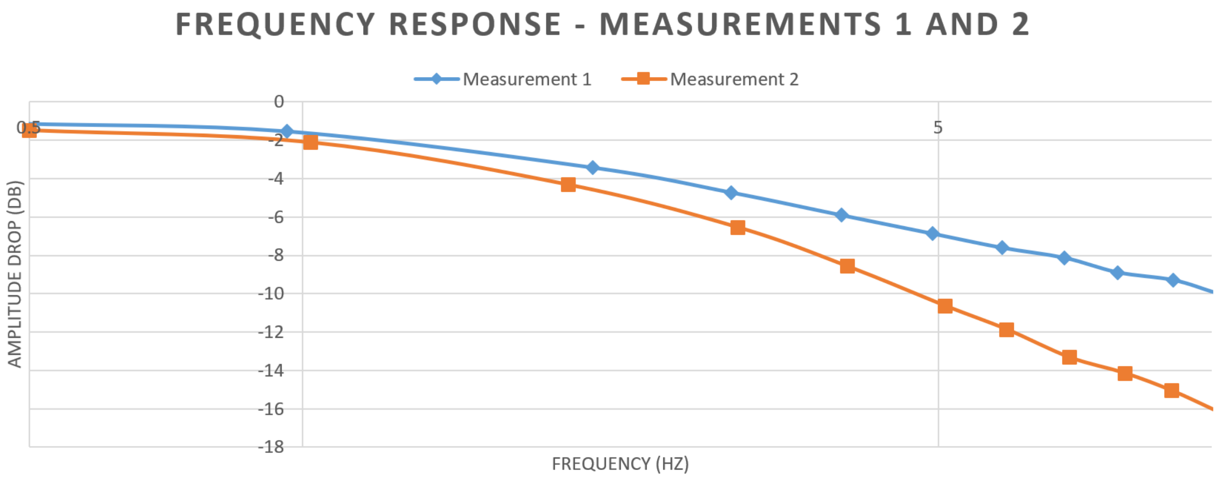


Figure 6: Experimental results of measurements 1 and 2 plotted on the same graph, in the frequency range of 0.5Hz to 10Hz.

Question 1.3

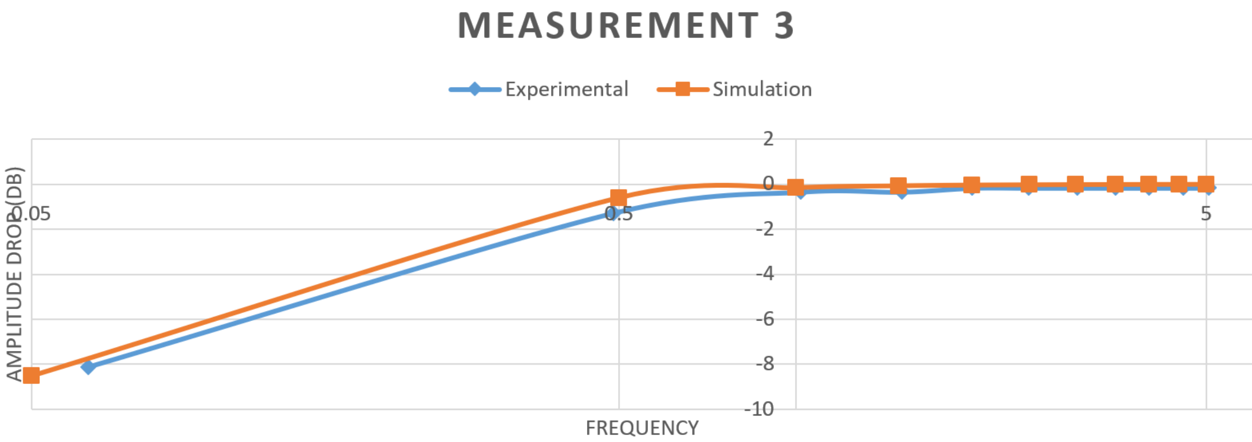


Figure 7: Measurement 3 - Measurements taken using a sinusoid of 1V peak-to-peak amplitude, in the frequency range of 0.05Hz to 5Hz. Resulting sinusoid amplitude drop was recorded in decibels (dB) on log-frequency axis. Circuit's behaviour was also simulated using PSpice, and simulation results were plotted for comparison.

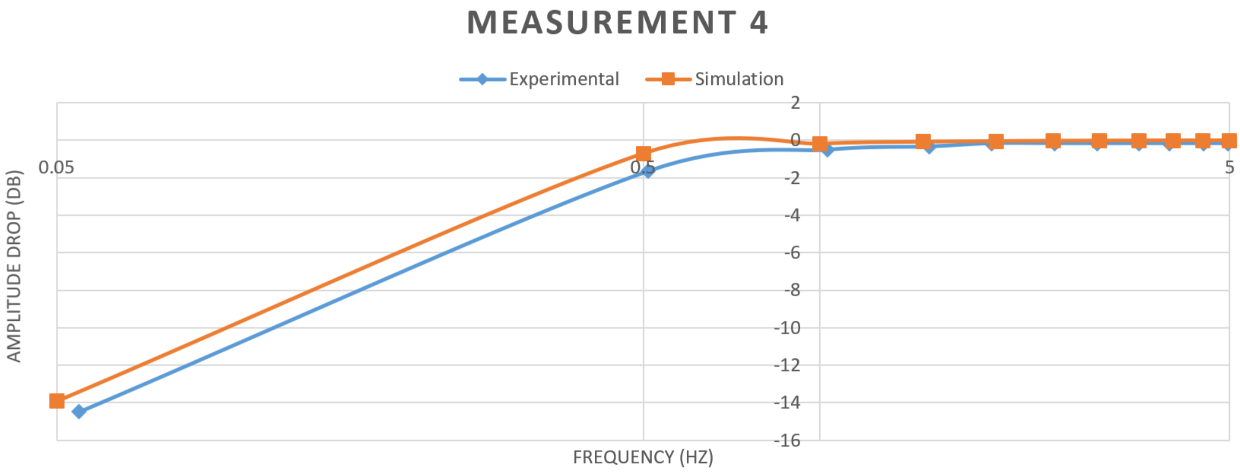


Figure 8: Measurement 4 - Measurements taken using a sinusoid of 1V peak-to-peak amplitude, in the frequency range of 0.05Hz to 5Hz. Resulting sinusoid amplitude drop was recorded in decibels (dB) on log-frequency axis. Circuit's behaviour was also simulated using PSpice, and simulation results were plotted for comparison.

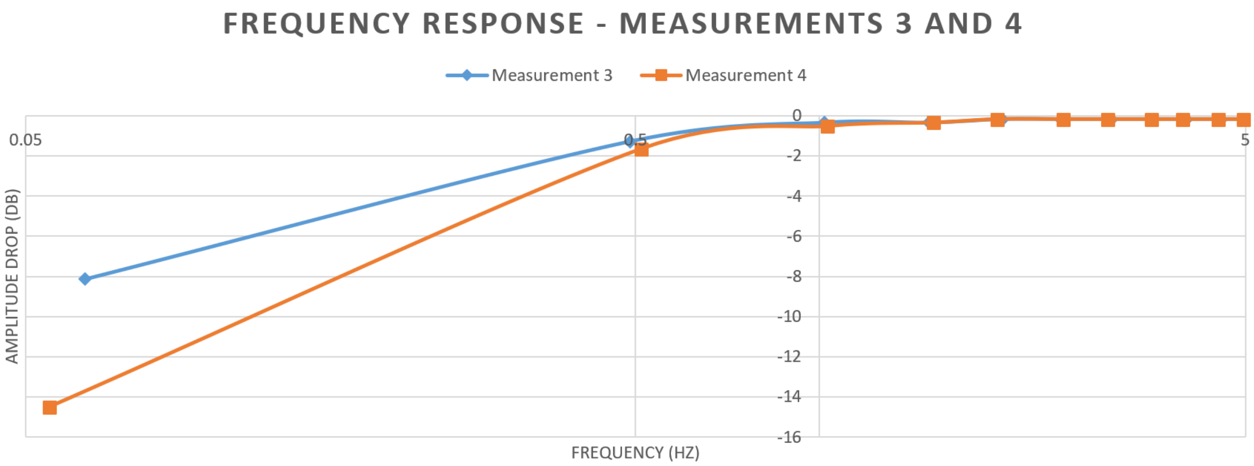


Figure 9: Experimental results of measurements 3 and 4 plotted on the same graph, in the frequency range of 0.05Hz to 5Hz.

Question 1.4

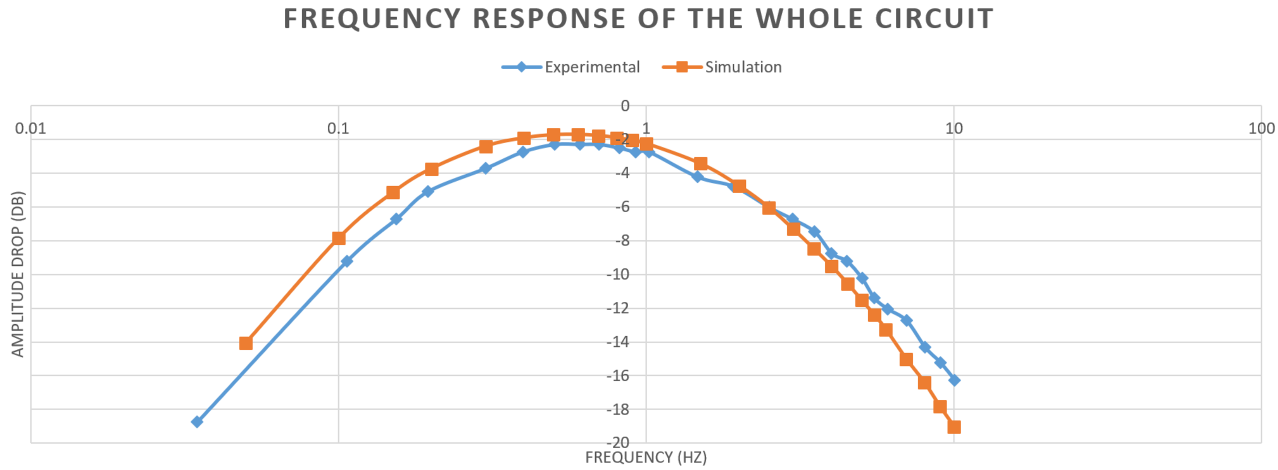


Figure 10: Frequency response the the entore circuit of Question 1.4. Circuit's behaviour was also simulated using PSpice, and simulation results were also plotted for comparison. Input signal was sinusoid of 1V amplitude peak-to-peak. Frequency range used was 0.05Hz to 10Hz.

2.2 Lab 2 - Op-amp Amplifier

Question 2.1

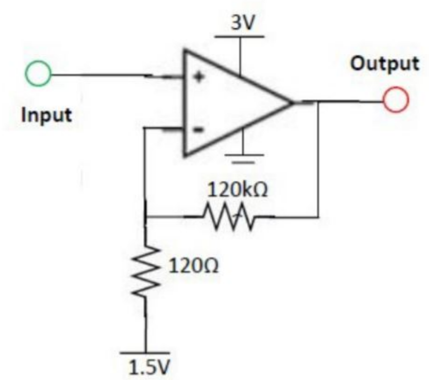


Figure 11: Circuit for Lab 2

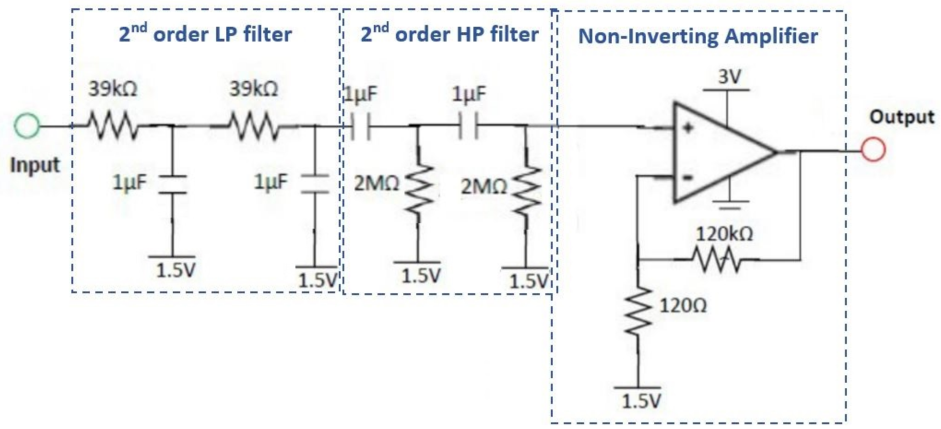


Figure 12: Sub-systems comprising the full circuit of Lab 2

Question 2.2

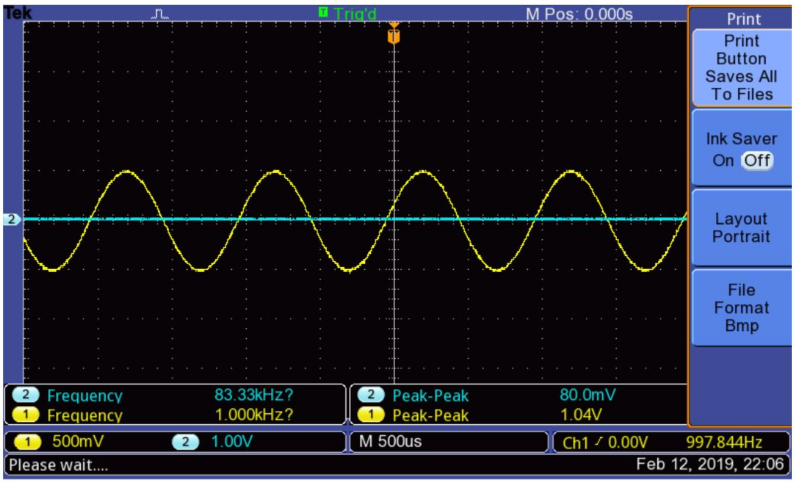


Figure 13: Testing Operation of single-supply non-inverting amplifier. Testing signal is 1V peak-to-peak sinusoid at 1kHz with zero DC offset

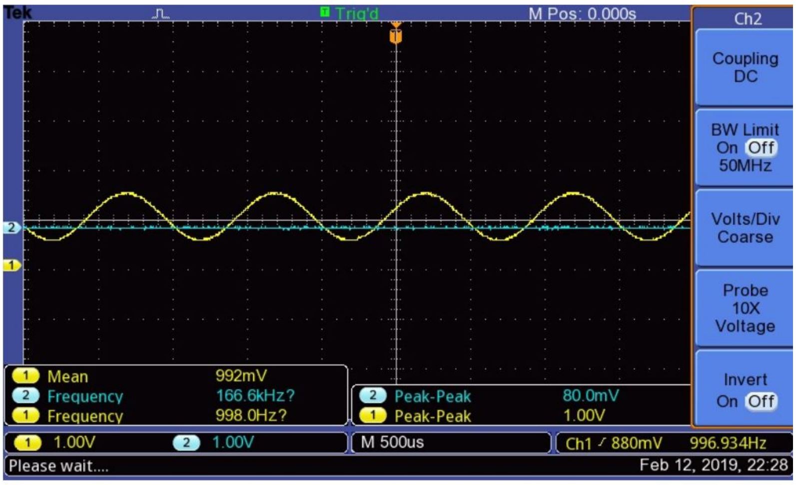


Figure 14: Testing Operation of single-supply non-inverting amplifier. Testing signal is 1V peak-to-peak sinusoid at 1kHz with 1V DC offset

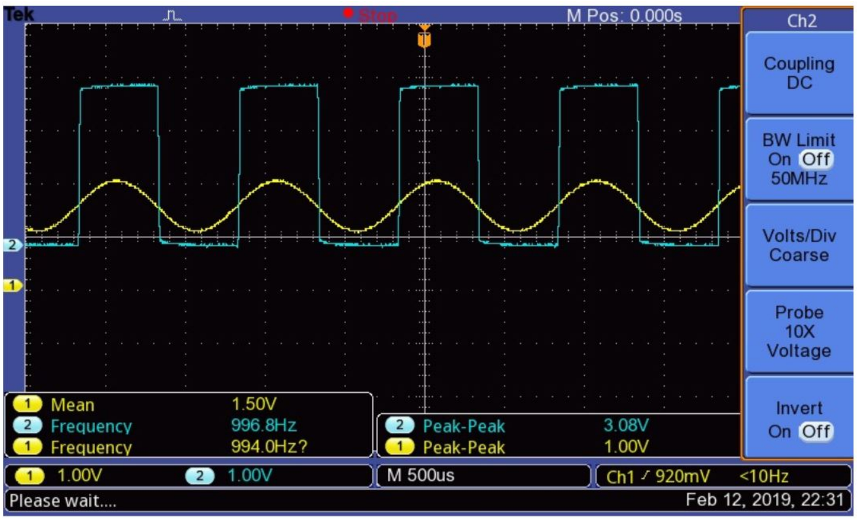


Figure 15: Testing Operation of single-supply non-inverting amplifier. Testing signal is 1V peak-to-peak sinusoid at 1kHz with 1.5V DC offset

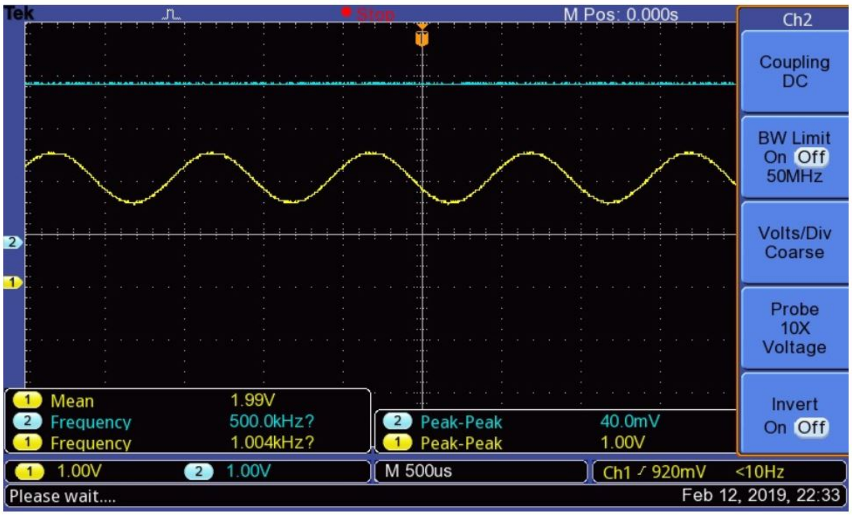


Figure 16: Testing Operation of single-supply non-inverting amplifier. Testing signal is 1V peak-to-peak sinusoid at 1kHz with 2V DC offset

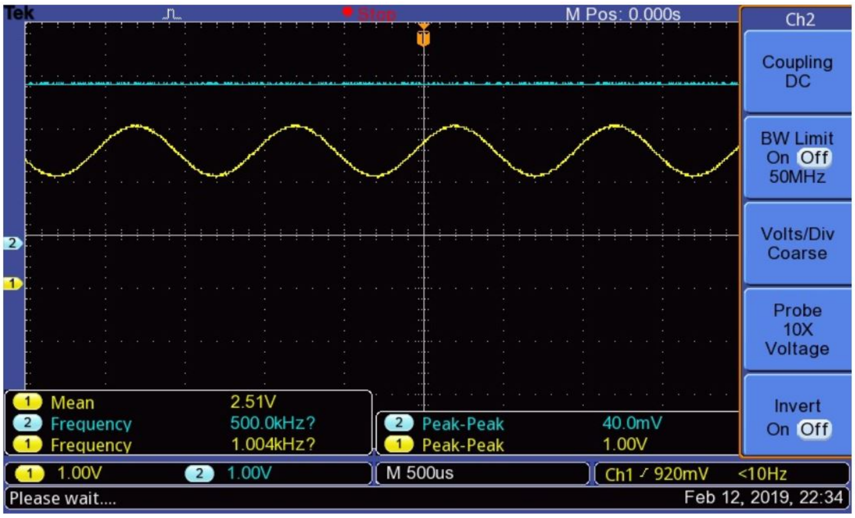


Figure 17: Testing Operation of single-supply non-inverting amplifier. Testing signal is 1V peak-to-peak sinusoid at 1kHz with 2.5V DC offset

Question 2.3

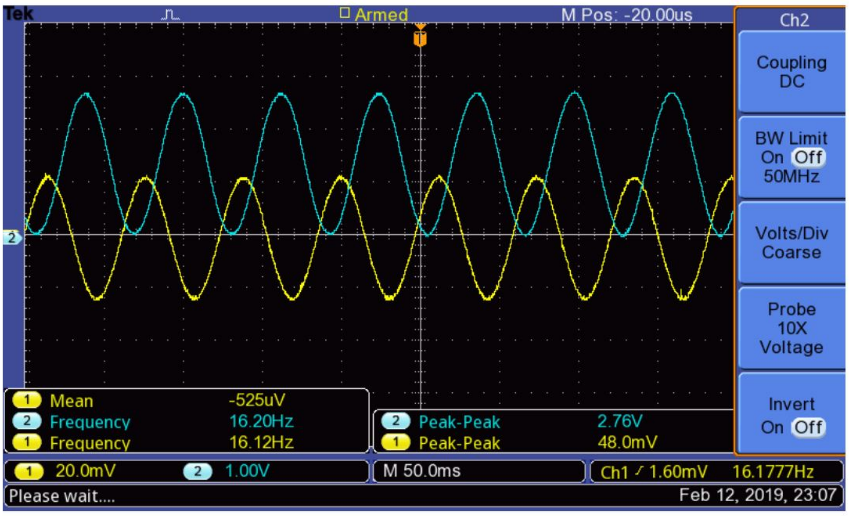


Figure 18: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 16.2kHz with no DC offset

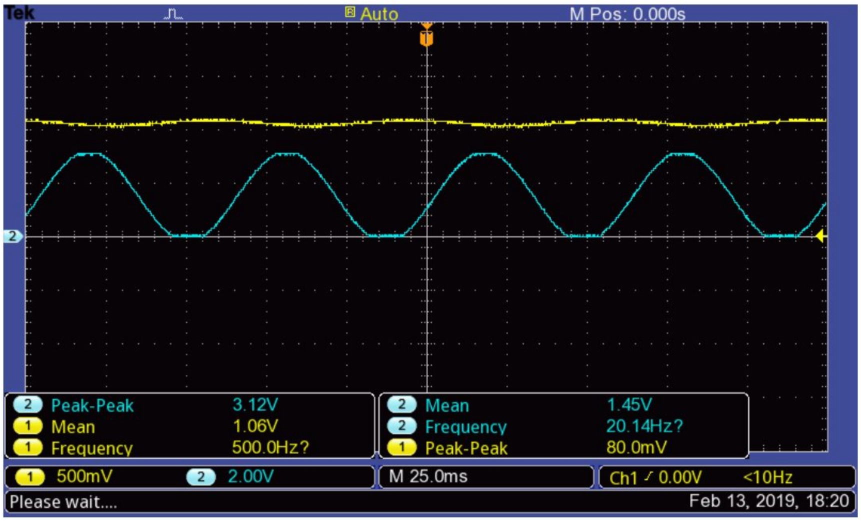


Figure 19: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 16.2kHz with 1V DC offset

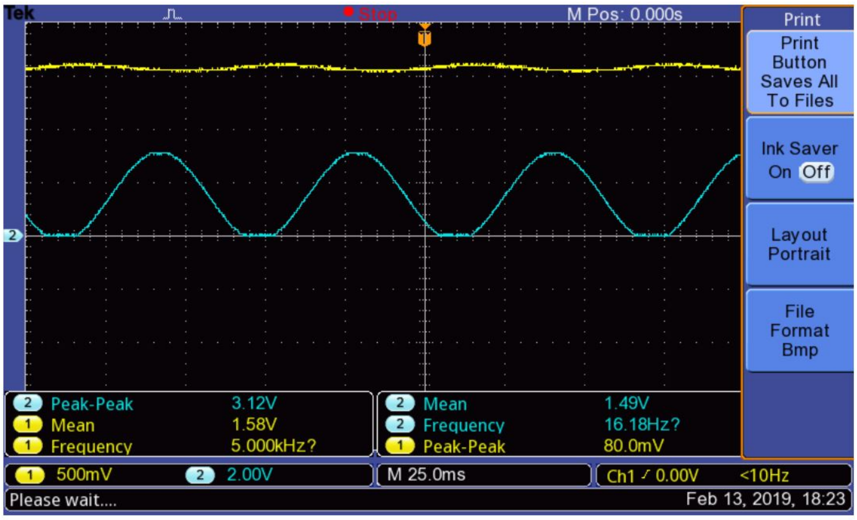


Figure 20: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 16.2kHz with 1.5V DC offset

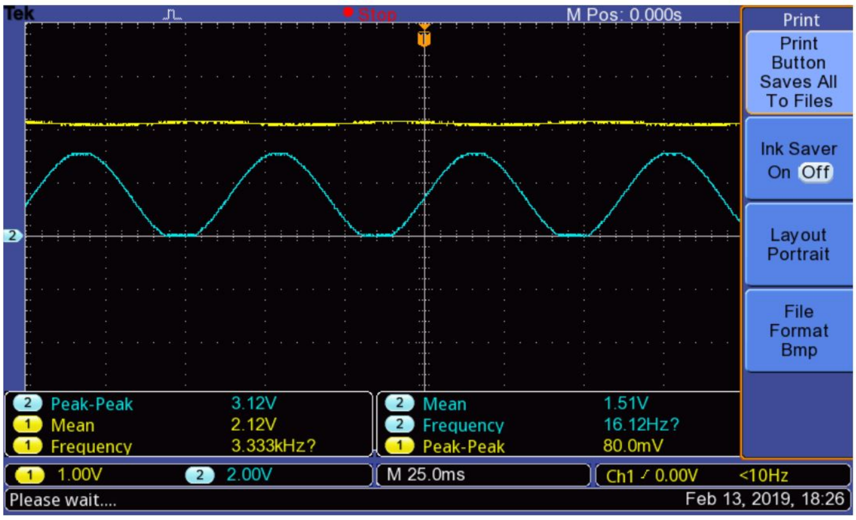


Figure 21: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 16.2Hz with 2V DC offset

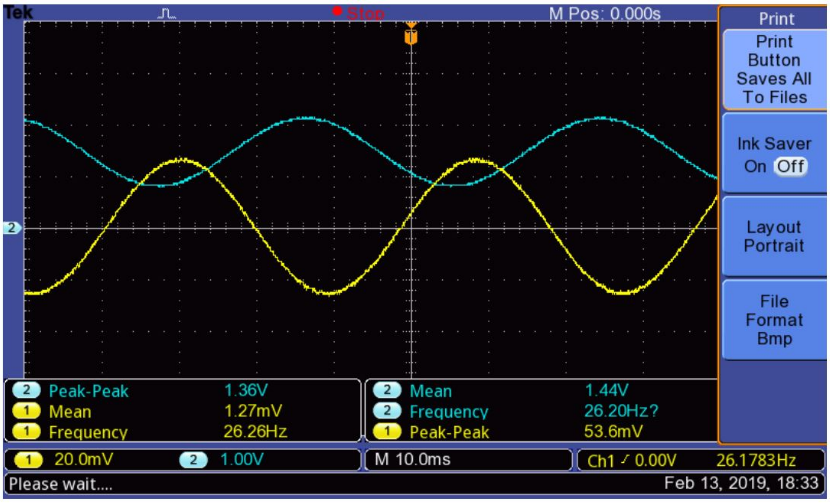


Figure 22: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 26.2Hz with no DC offset

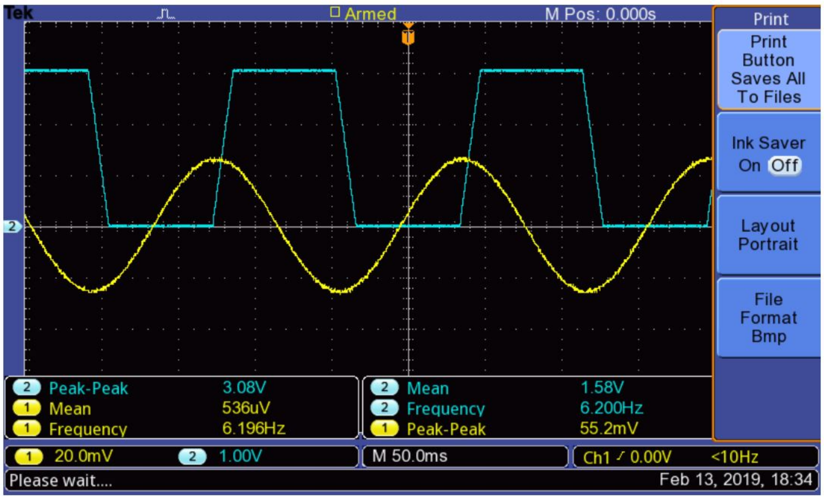


Figure 23: Testing Operation of the cascade of fourth order band-pass filter and non-inverting amplifier. Testing signal is 50mV peak-to-peak amplitude sinusoid at 6.2Hz with no DC offset

2.3 Lab 3 Part 1 - Non-linear circuits

<u>Question 3.1</u>

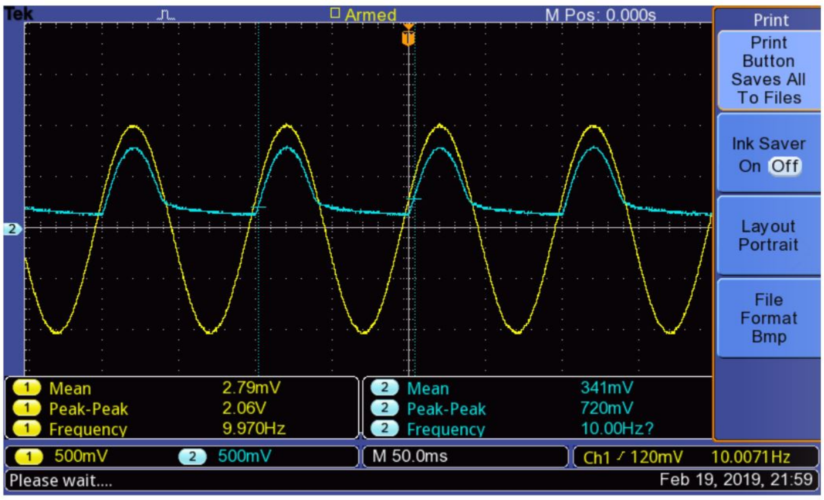


Figure 24: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 10Hz and 2V peak-to-peak and 0V DC offset

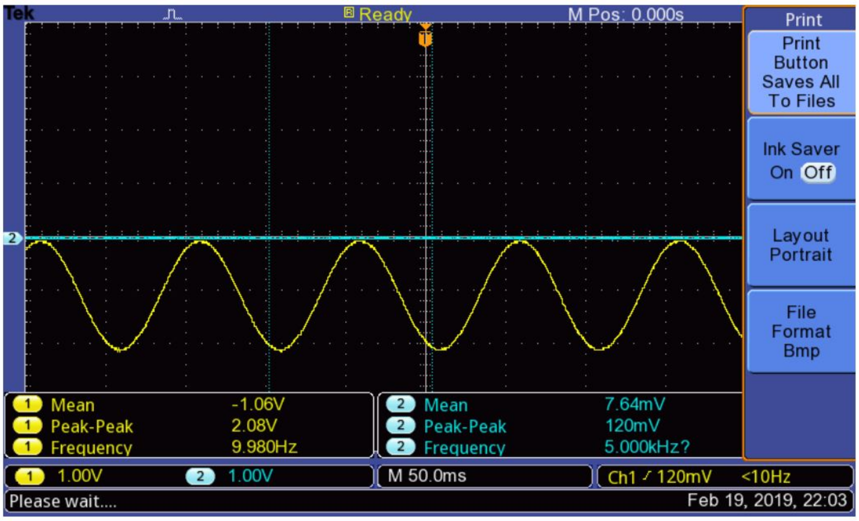


Figure 25: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 10Hz and 2V peak-to-peak and -1V DC offset

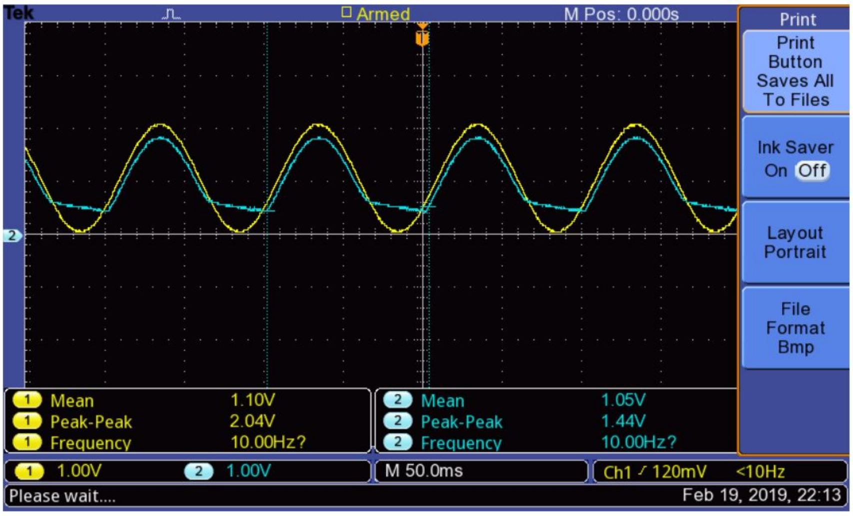


Figure 26: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 10Hz and 2V peak-to-peak and +1V DC offset

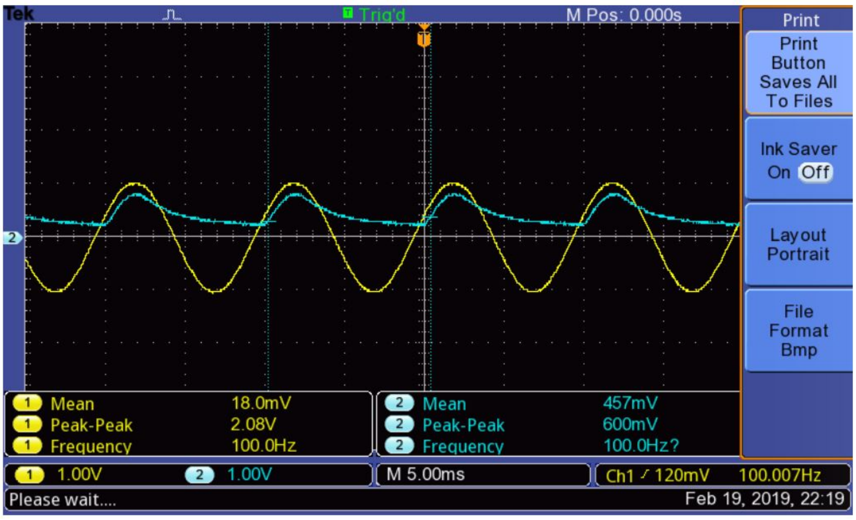


Figure 27: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 100Hz and 2V peak-to-peak and OV DC offset

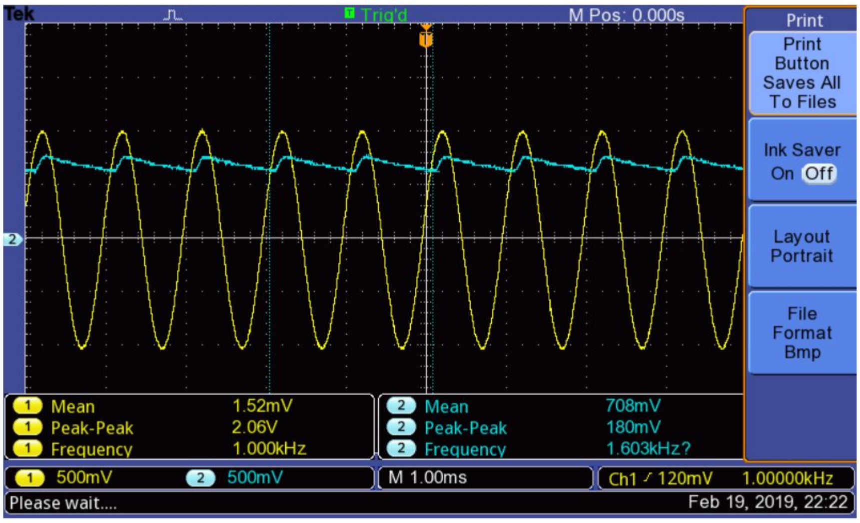


Figure 28: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 1kHz and 2V peak-to-peak and OV DC offset

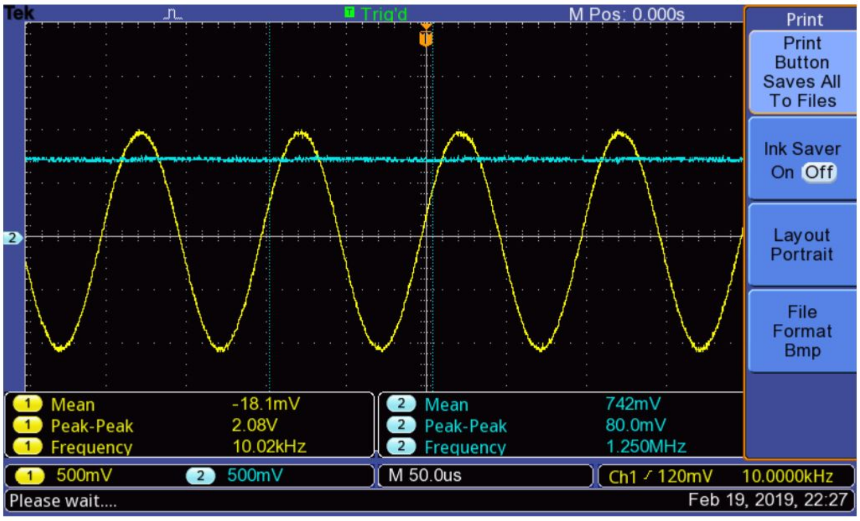


Figure 29: Testing the non-linear circuit of Lab 3 Part 1. Testing signal is a sinusoid with frequency of 10kHz and 2V peak-to-peak and OV DC offset

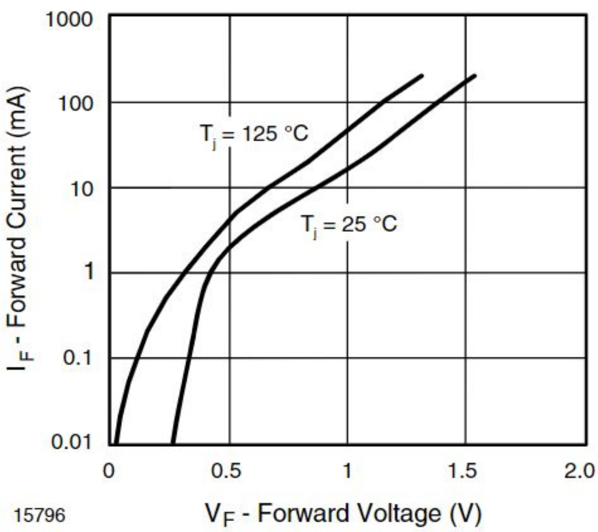


Figure 30: Forward Current vs. Forward Voltage plot for Schottky Diode BAT83S. This figure was obtained from the component's datasheet.

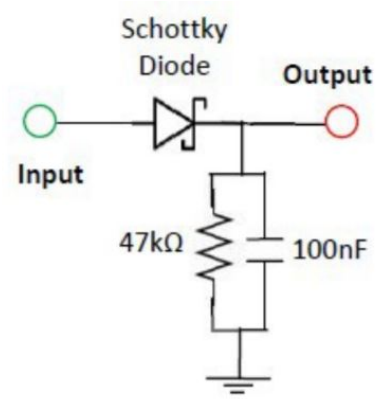


Figure 31: Circuit Diagram for Lab 3 Part 1, involving a Schottky diode which is a non-linear component.

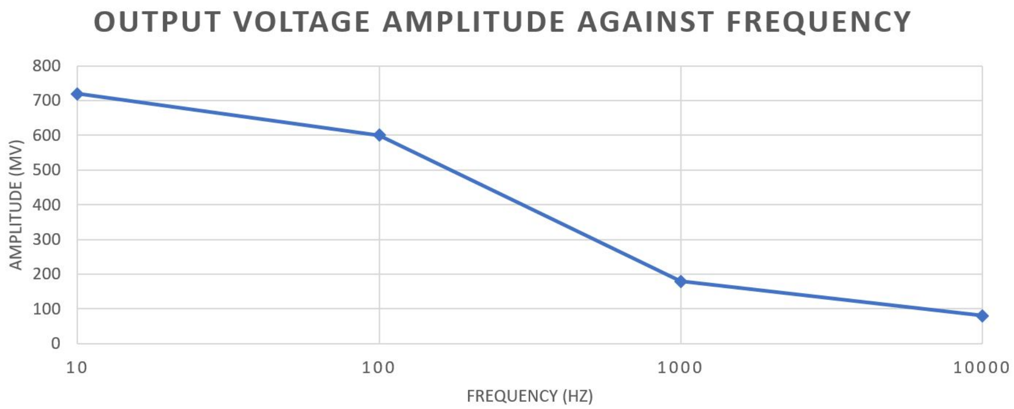


Figure 32: Plot of voltage output peak-to-peak values for Circuit Diagram of Lab 3 Part 1 against frequency. Values used were the peak-to-peak values from Fig.24 and 27-29. Voltage input was 2V peak-to-peak with 0V DC offset.

2.4 Lab 3 Part 2 - Unguided Design

Question 3.2

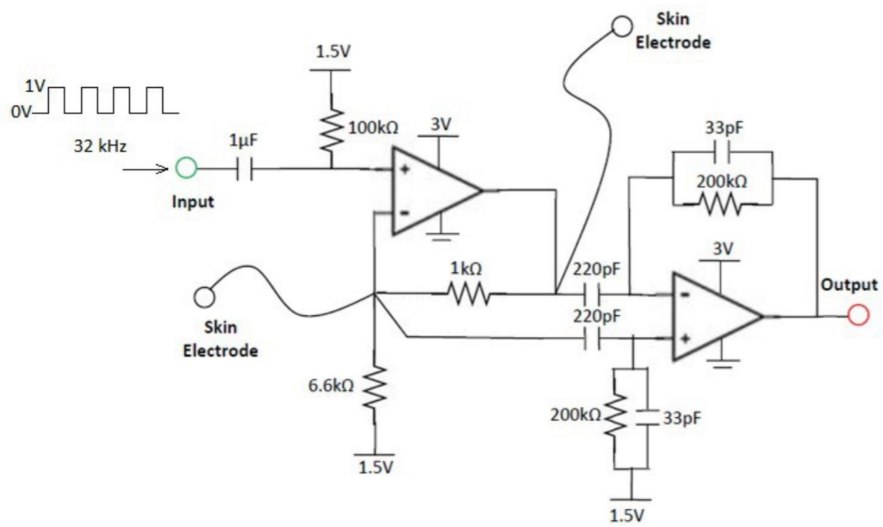


Figure 33: Circuit for Lab 3 Part 2

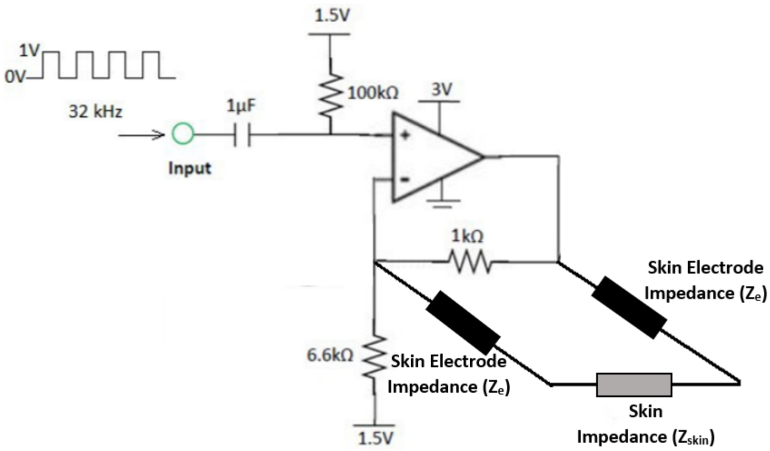


Figure 34: Circuit for Question 3.2 b). Alternative representation compared to the one given in Lb practical pg. 29.

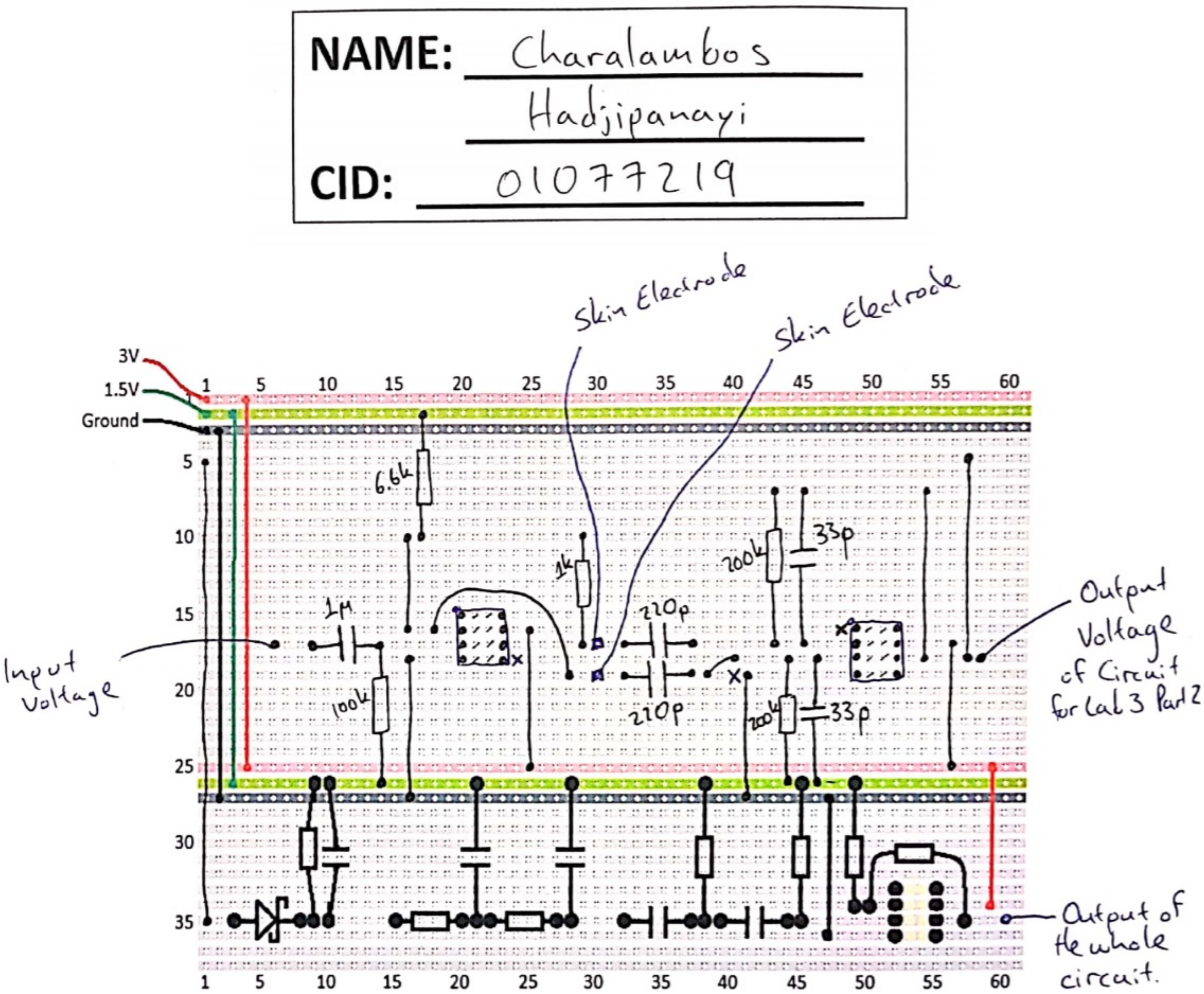


Figure 35: Approved Design of Circuit for Lab 3 Part 2, with connections to existing component topology

Question 3.3

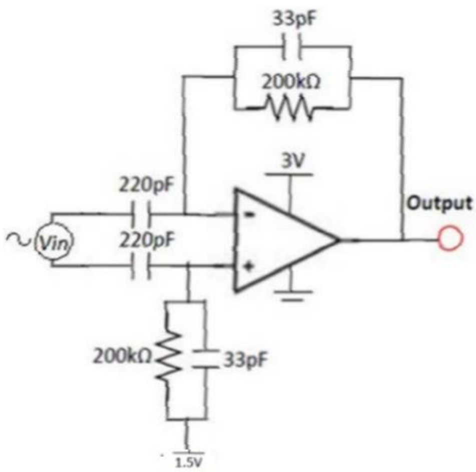


Figure 36: Circuit for Question 3.3 a)

Question 3.4

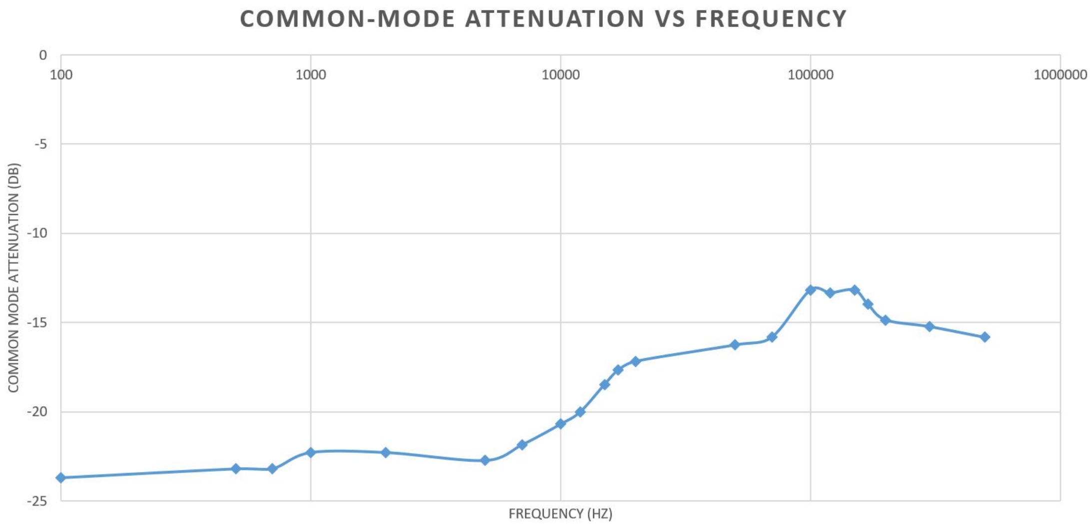


Figure 37: Variation of experimental common-mode attenuation values with frequency, in the range of 100Hz to 500kHz. The common-mode attenuation values are in Decibels and the frequency axis is logarithmic. Common signal used was sinusoid with 0.5V peak-to-peak amplitude and 1.5V DC offset.

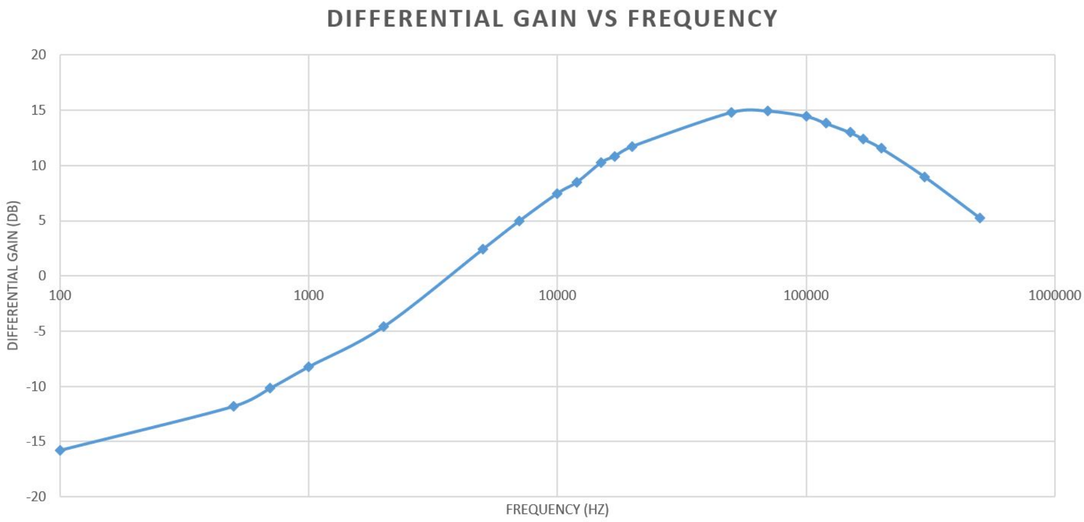


Figure 38: Variation of experimental differential gain values with frequency, in the range of 100Hz to 500kHz. The Differential gain values are in Decibels and the frequency axis is logarithmic. Description of input signals found in Part A, Question 3 Part b)

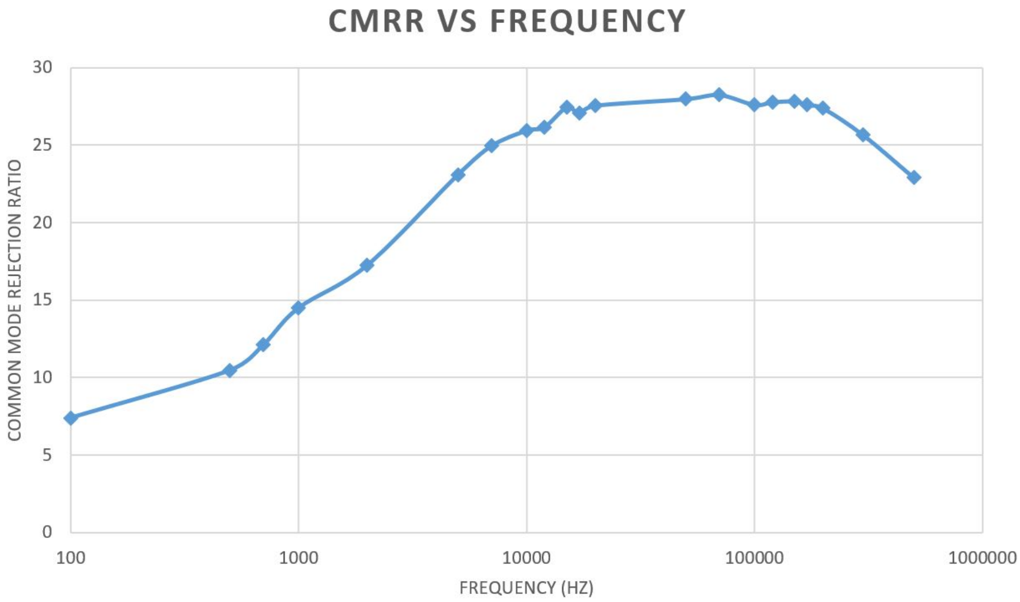


Figure 39: Plot of Common-Mode Rejection Ratio (CMRR) for frequencies in the range 100Hz to 500kHz, obtained using the method specified in Part A, Question 3 Part c)

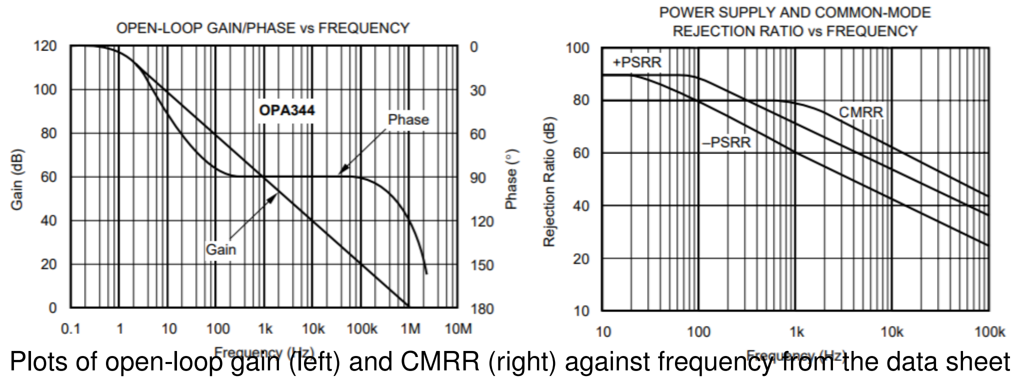


Figure 40: Plots of open-loop gain (left) and CMRR (right) against frequency from the data sheet of OPA344.

2.5 Final Lab

Question 4

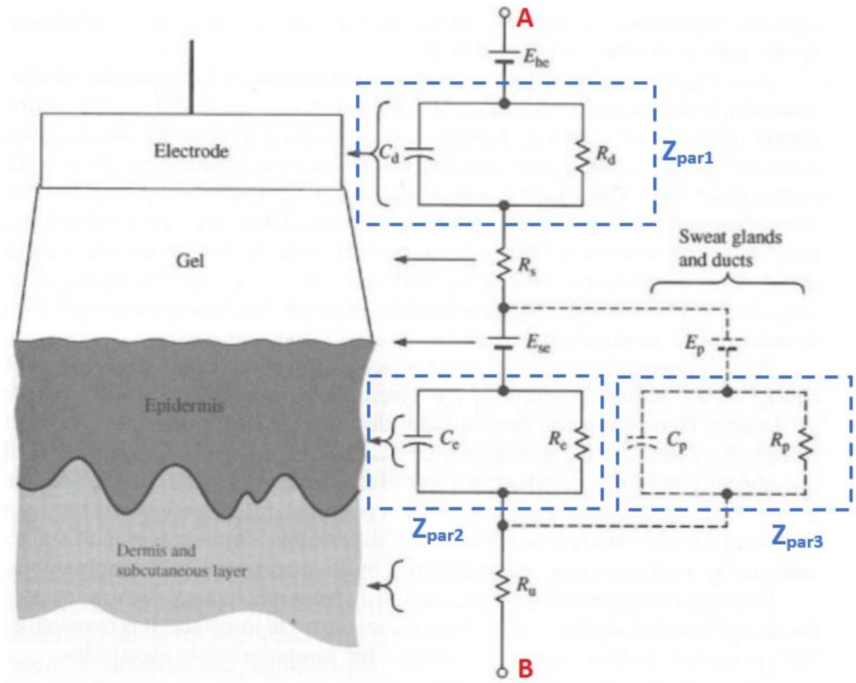


Figure 41: Electrical analogue model of the electrode-skin interface used for Question 4.

Question 5

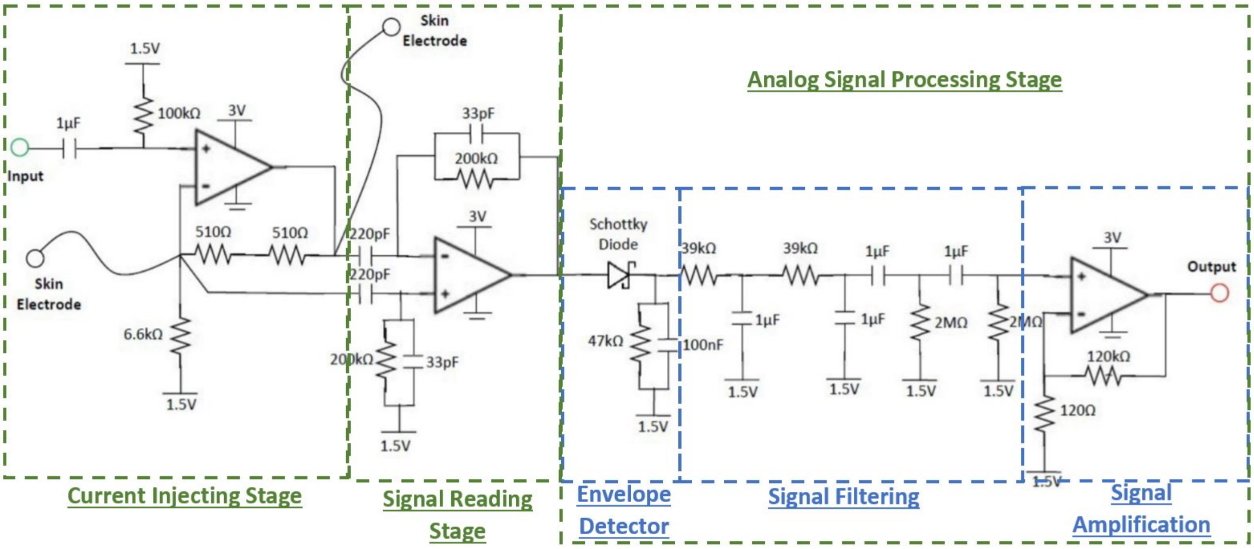


Figure 42: Full circuit of impedance respirometric system with identified system blocks with distinct functions.

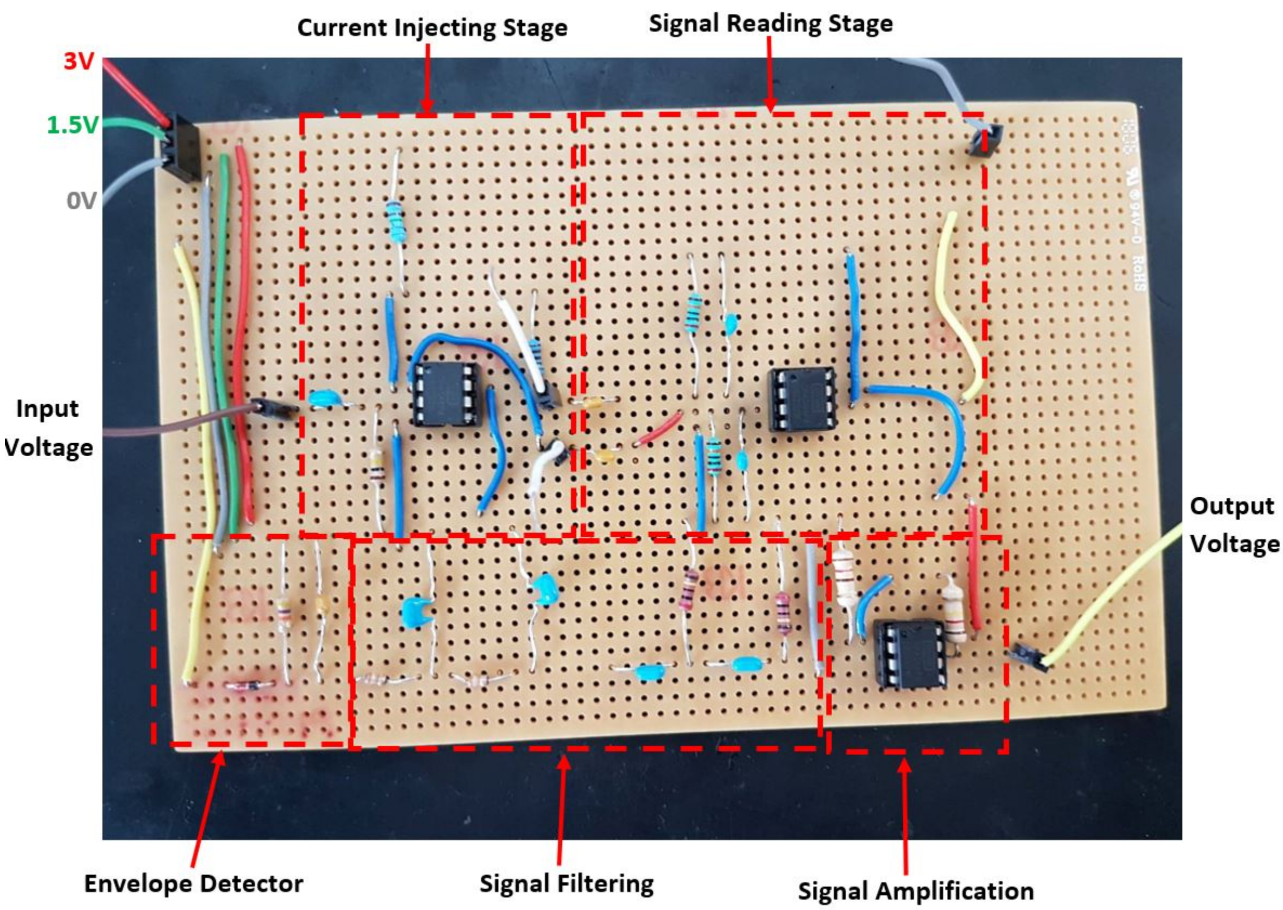


Figure 43: Actual Implemented Board with labelled system blocks.

2.6 Working Prototype Board - Measurements

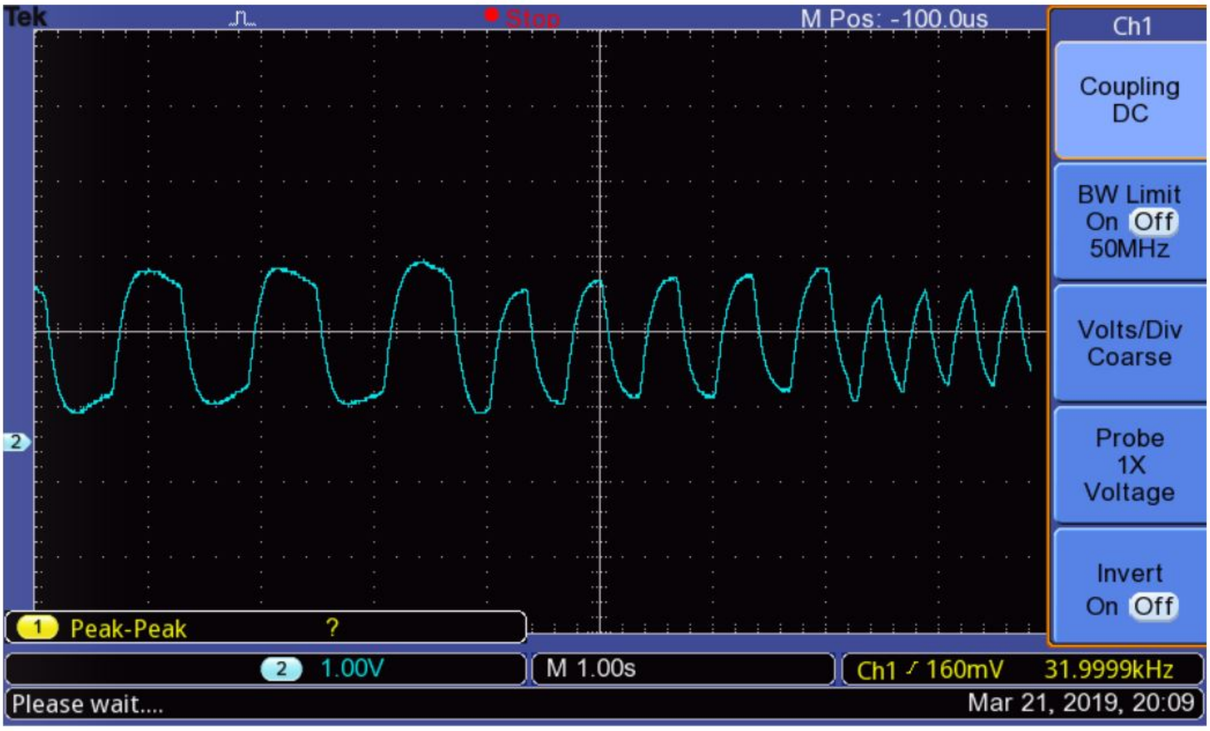


Figure 44: Output of entire impedance respirometric system using square carrier current of 1V peak-to-peak. Electrodes are connected to Respiration Simulator (M98C), with variation 10Ω and Base Impedance 1kΩ. The plot shows voltage output for breathing rates of 50,90,150 beats/min consecutively. DC coupling is used.

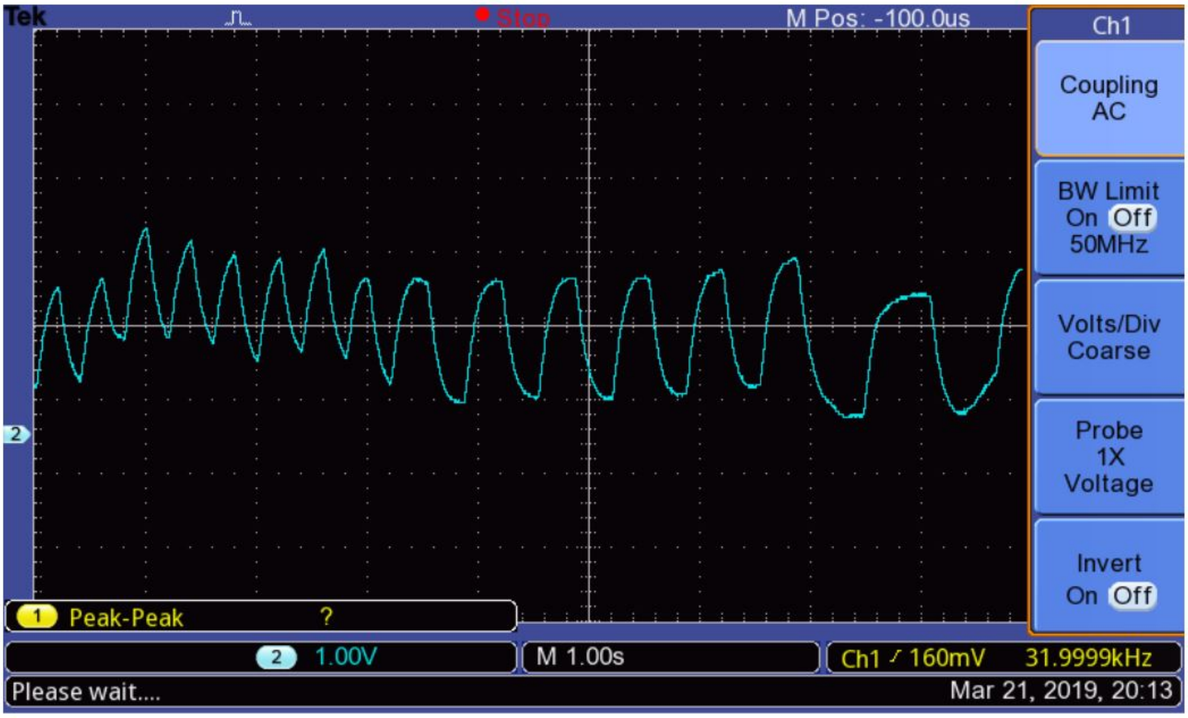


Figure 45: Output of entire impedance respirometric system using square carrier current of 1V peak-to-peak. Electrodes are connected to Respiration Simulator (M98C), with variation 10Ω and Base Impedance 1kΩ. The plot shows voltage output for breathing rates of 150,90,50 beats/min consecutively. AC coupling is used.

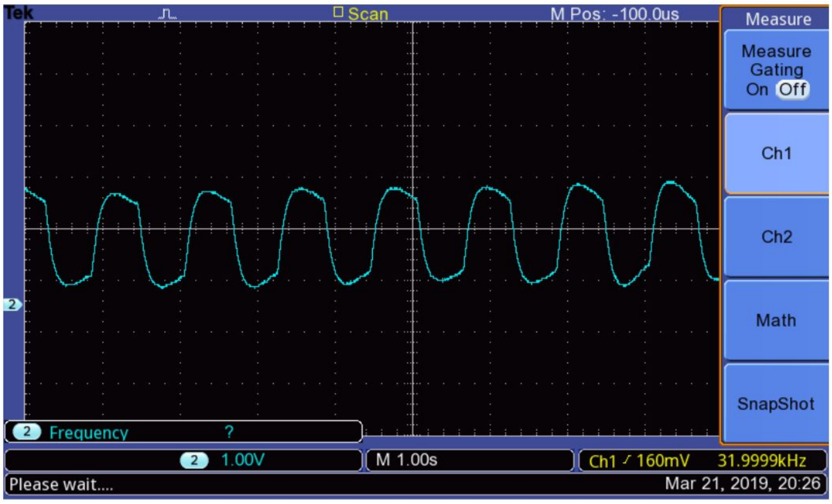


Figure 46: Output of entire impedance respirometric system using square carrier current of 1V peak-to-peak. Electrodes are connected to Respiration Simulator (M98C), with variation 10Ω and Base Impedance 1kΩ. The plot shows voltage output for breathing rate of 50 beats/min. DC coupling is used.

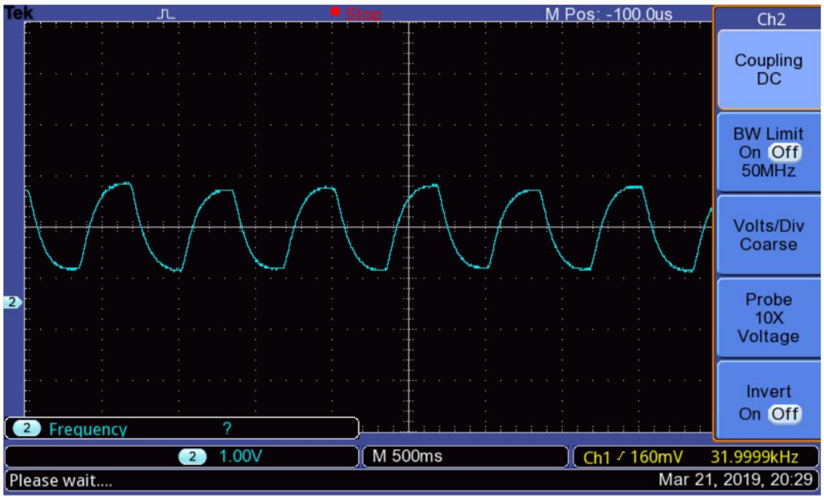


Figure 47: Output of entire impedance respirometric system using square carrier current of 1V peak-to-peak. Electrodes are connected to Respiration Simulator (M98C), with variation 10Ω and Base Impedance 1kΩ. The plot shows voltage output for breathing rate of 90 beats/min. DC coupling is used.

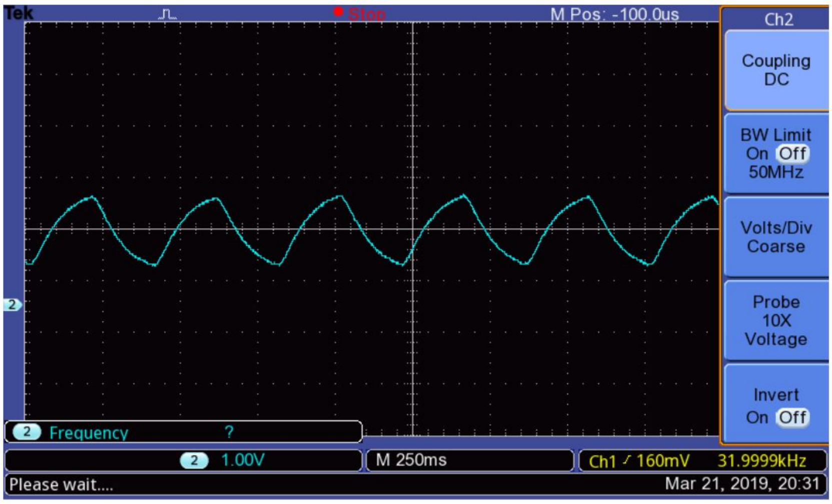


Figure 48: Output of entire impedance respirometric system using square carrier current of 1V peak-to-peak. Electrodes are connected to Respiration Simulator (M98C), with variation 10Ω and Base Impedance 1kΩ. The plot shows voltage output for breathing rate of 150 beats/min. DC coupling is used.

END OF COURSEWORK