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E344 Assignment 3

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Report submitted in partial fulfilment of the requirements of the module Design (E) 344 for the degree Baccalaureus in Engineering in the Department of Electrical and Electronic Engineering at Stellenbosch University.

October 25, 2020

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Nomenclature

Variables and functions

V^+	Operational amplifier non-inverting input
V^-	Operational amplifier inverting input
VIN_max	Operational amplifier maximum input voltage
VIN_min	Operational amplifier minimum input voltage
VIN_diff	Operational amplifier differential input voltage

Acronyms and abbreviations

Op Amp	Operational Amplifier
LPF	Low-pass Filter
HPF	High-pass Filter
FFT	Fast Fourier Transform

Chapter 1

System design

1.1. System overview

In this project, a health monitoring system comprising an analogue temperature sensor and an optic heart rate monitor are implemented. This report covers the implementation and design of both the temperature and heart rate sensors conditioning circuits. Figure 1.1 illustrates the full system diagram.

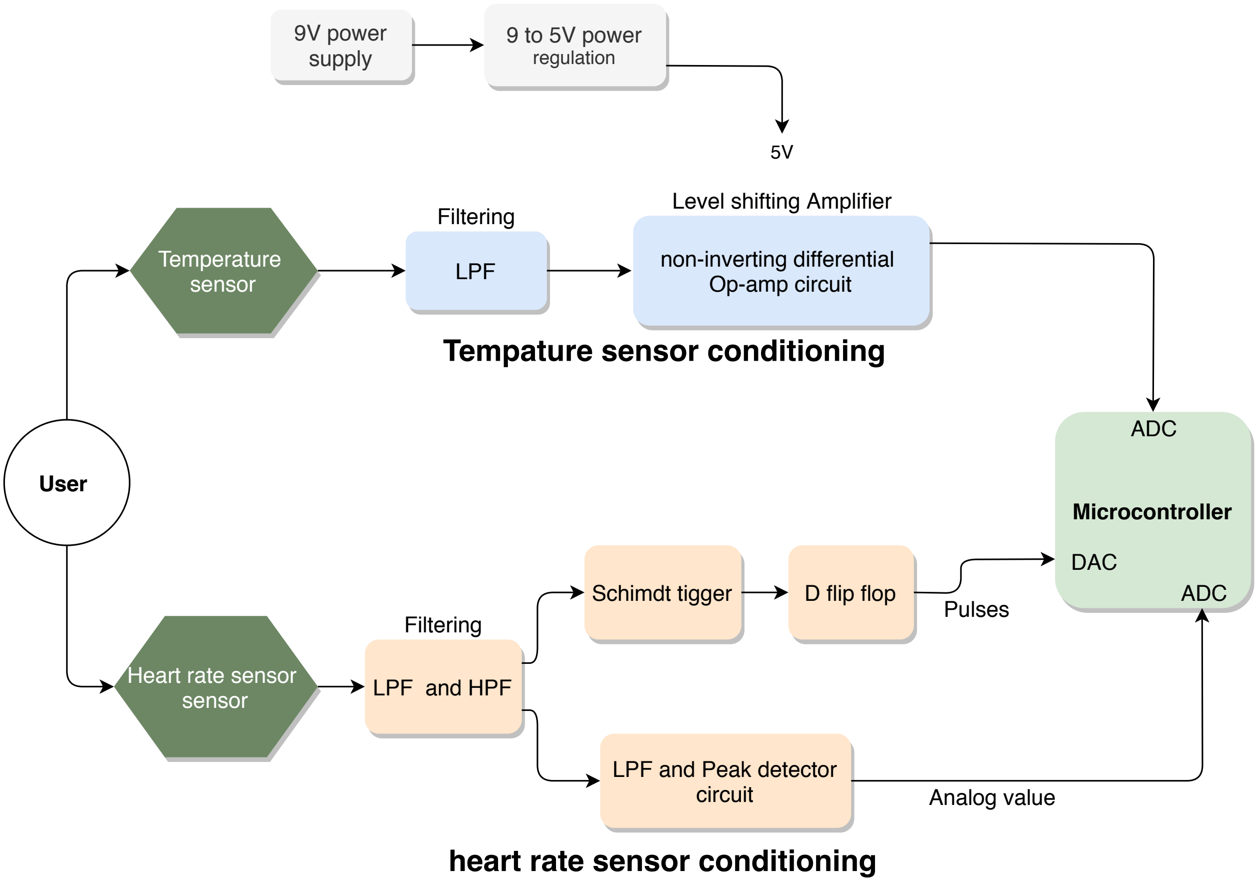


Figure 1.1: System diagram

- **Voltage regulation:** The voltage regulator shall provide a regulated 5VDC supply to the system from a nominal 9V power. Two alternative regulation modes are available, namely linear and switch mode regulation. The following chapter covers the rationale followed in the regulator choice.
- **Temperature sensor conditioning:** We wish to measure the users temperature via the temperature sensor and process the values such that it can be further analysed with

a microcontroller. This section comprises a filtering and level shifting amplifier modules. Before feeding the stimulus(sensor output) to the amplifier, it is first filtered using an **active low-pass filter** in the Butterworth configuration offering a fast transient response. The filter shall suppress noise from the stimulus by attenuating components at 50Hz and above. For amplification, an Op-amp circuit in a **differential amplifier** configuration is used: It provides a gain and removes a dc offset from the sensor output(stimulus) and thus, amplifies it from millivolts to volts while staying within the recommended output voltage range.

- **Heart rate sensor conditioning:** We wish to measure the user's heartbeats via an optical heart rate sensor and process the signal such that it can be further analysed with a microcontroller. This process requires first, filtering noise out of the heartbeats signal and applying a gain using both a Low-pass and a High-pass filters. The filtered signals are then fed into a Schmidtt trigger, which is a comparator circuit with both a high and low threshold voltages allowing to output square waves. However, the comparator output pulses have varying durations; Hence, a D-type flip-flop, in a monostable or one-shot configuration is used to normalise the pulses duration to about 200ms. In addition, a transducer is used to convert a given heart rate to a specific analog value. The transducer circuit comprises a LPF allowing different attenuations to the heartbeat signals depending on their frequencies(rate), a level shifting differential amplifier and a peak detector circuit.

- The full health monitoring system current drawing is to be limited to bellow 50mA with 25mA for the temperature and heart rate sensors conditioning circuits each. Large resistor values and power sensitive designs will therefore be used to ensure that we adhere to this requirement.

Chapter 2

Voltage regulation

2.1. Introduction

This section covers the power conditioning design of the System. Two alternatives were considered; Namely, a linear voltage regulator, the **LM7805** and a switching mode regulator, the **LM2595**. The working of a linear regulator involves varying a resistance according to the load, which results in a constant output voltage. Conversely, switching mode regulators rapidly switch a series of FETs on and off. These devices store the input energy temporarily and then release that energy to the output at a different voltage level. [2]
The following subsections document the rationale followed in the choice of the regulator.

2.2. Design

In order to pick one of the regulation modes in the design, both the linear and switching mode regulators were implemented with additional circuitry and their respective performances were compared.

- **Linear regulation**
  The circuit diagram is illustrated in figure 2.1 below. Capacitors C1 and C2 are used to smooth out noise on the 9V input supply and the 5V regulated output, respectively.

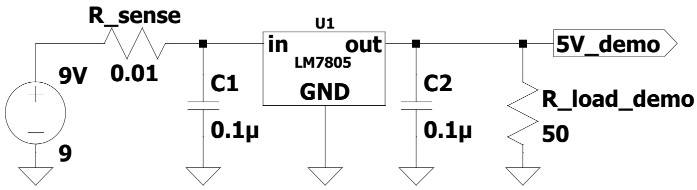


Figure 2.1: Linear regulator circuit diagram

  Capacitors C1 and C2 values, 0.1 µF each, were selected as suggested in the LM7805 data sheet specifications for typical applications circuits for a minimal noise linear regulation. [3]

- **Switch mode regulation**
  The configuration used is for an adjustable output voltage and it was preferred to a fixed output voltage for more flexibility as derived from the LM2595 data sheet [4]. The circuit diagram is illustrated in figure 2.2 below.

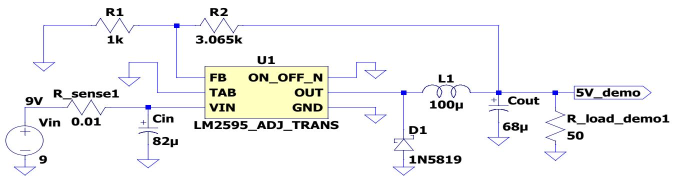


Figure 2.2: Switchmode regulator circuit diagram

In both the linear and switching regulator evaluations, a resistive load, $R_{\text{load\_demo}}$ was used to draw 100mA from the 5VDC side. Hence, $R_{\text{load\_demo}} = \frac{5V}{100mA} = 50\,\Omega$

• R_1 and R_2: $R_1 = 1\,\text{k}\Omega$, $R_2 = R_1\left(\frac{V_{\text{OUT}}}{V_{\text{REF}}} - 1\right)$, where $V_{\text{REF}} = 1.23\,\text{V}$.

$\therefore R_2 = 3,065k\Omega$

These two resistors set the output voltage value to the required 5V.

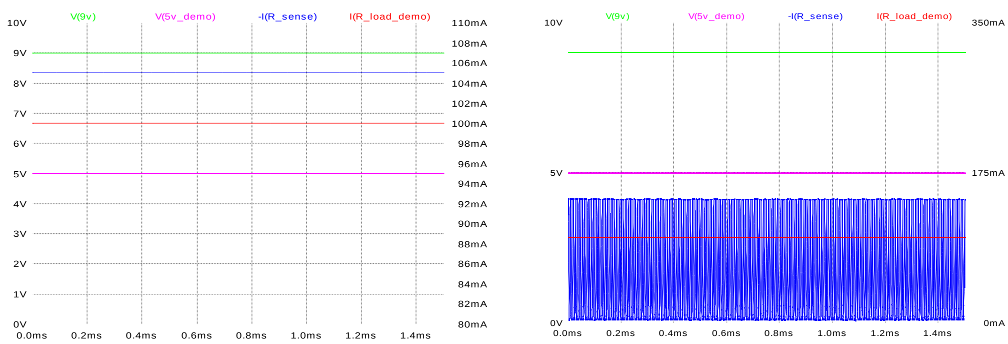
• Catch Diode D1: A Schottky diode, the 1N5819 was picked due to its fast switching speed and low forward voltage drop.

• Capacitors C_{in} and C_{out} values were selected following the *quick design table* for a 5VDC output from the **LM2595** datasheet [4].

• L1: from the given formula $\text{E} \cdot \text{T} = (\text{V}_{\text{IN}} - \text{V}_{\text{OUT}} - \text{V}_{\text{SAT}}) \cdot \frac{\text{V}_{\text{OUT}}+\text{V}_{\text{D}}}{\text{V}_{\text{IN}}-\text{V}_{\text{SAT}}+\text{V}_{\text{D}}} \cdot \frac{1000}{150\text{kHz}}(\text{V} \bullet \mu\text{s})$, where $\text{V}_{\text{SAT}} = 1\text{V}$ and table 30 in the datasheet [4], L1 value was chosen to be $100\,\mu\text{F}$.

2.3. Results

•Input and output voltage and current values: figure 2.3 below illustrates the voltage and current values at the input and output of the both the linear and switchmode regulators.



(a) **(b)**

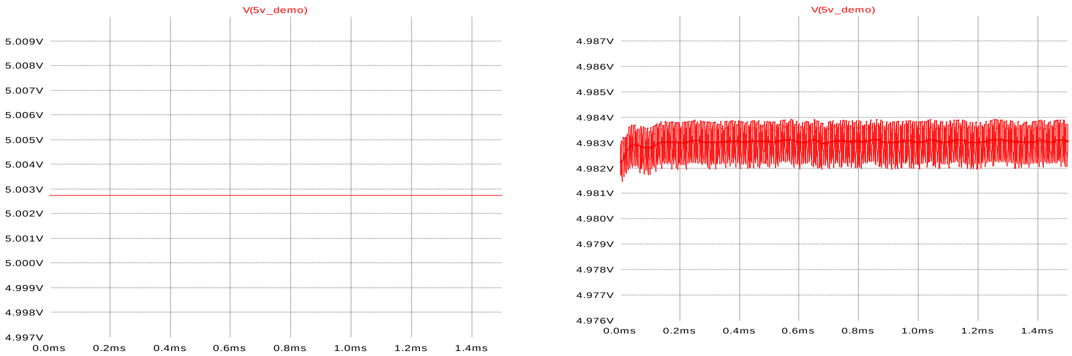
Figure 2.3: Linear and Switchmode regulators input and output voltages and currents (a)linear regulator (b)switchmode regultator

The values were recorded in table 2.1 below:

Regulator	**V(9v)**	**V(5v_demo): output V**	**-I(R_sense)**	**I(R_load_demo): output i**
linear	9	5.00	105.061mA	100.055mA
switchmode	9	5.00	78.135mA	99.674mA

Table 2.1: Linear and switchmode regulators input and output voltages and currents

•**Output noise levels:** figure 2.4 below illustrates the output noise levels for both regulation modes.



(a) **(b)**

Figure 2.4: Output noise levels: (a)linear regulator (b)switchmode regulator

•**Power calculations, efficiency and dropout voltages:**
Efficiency and power calculations as well as dropout voltages comparison are illustrated in table 2.2 bellow.

Regulator	Input Voltage	Input Current	Input Power	Output Voltage	Output Current	Output Power	**Efficiency**	**Dropout Voltage**
Linear	9VDC	105.06mA	945.547mW	5.003VDC	100mA	500.3mW	**52.91%**	**2V**
Switchmode	9VDC	60.23mA(rms)	560.03mW	5.00VDC	100.01mA	501.5mW	**89.55%**	**0.1V**

Table 2.2: Efficiency and power calculation

2.4. Summary

As expected, both the liner and switching mode regulators output 5VDC with 100 mA into the 50 Ω demo load. However, the switching mechanism in the switchmode regulator introduces lots of noise(approximately 1.61 mVpp) on its output despite the regulator's high efficiency(89.55%) and very low dropout voltage (0.1 V). Inversely, the linear regulator has very low output noise levels (practically none) but with low efficiency (52.91%) and relatively high dropout voltage (2 V), meaning the input voltage needs to be at least 7 V for the regulation to occur. **Given its low output noise levels, the liner regulator was preferred to the switchmode regulator.** In addition, despite the linear regulator low efficiency, the power loss is approximately 445.25 mW for the 100 mA output current. A thermal analysis in the above conditions resulted in a junction temperature of 53.94°C which is way smaller than the the 125°C maximum rated [3]. Keeping in mind that the full system will be designed to draw less than 50 mA, we can thus expect less power loss in the linear regulator.

Chapter 3

Temperature sensor conditioning circuit

3.1. Intro

This section documents the design of the of the temperature sensor conditioning circuit. It includes an **active low pass filter** and a **differential amplifier**.

3.2. Design

3.2.1. Low Pass Filter design

To suppress the high frequency noise form the temperature sensor output, a second order active LPF in the Butterworth configuration was selected. It features a fast step response as well a steep amplitude drop-of gradient after the corner (3dB) frequency. The circuit digram is illustrated in figure 3.1 below.

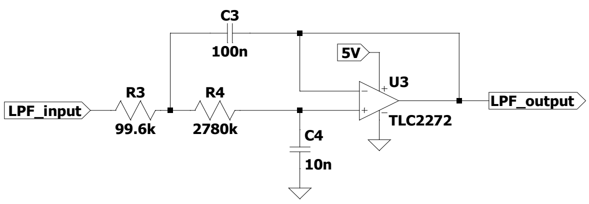


Figure 3.1: Low Pass Filter circuit diagram

Designing the above circuit involved selecting values for capacitors C3 and C4 and resistors R3 and R4 for a corner frequency of 5Hz. Given that it's a second order LPF, we can expect the noise signal at 50Hz (10 times the corner frequency) to be attenuated by -30 to -40dB. With a -30dB attenuation, the 15mV amplitude (corresponding to 1°C as per specifications) noise signal will be attenuated to 0.5mV amplitude. Given that the differential amplifier has a gain of 41V/V, this will ensure that the gain at the final output is at most $41 \times 0.5\text{mV} = 20.5\,\text{mVp}$ which is smaller than the required $80\,\text{mVp}$.

•Select $C4 = 10\,\text{nF}$; then choose $C3 = 10 \times C4 = 100\,\text{nF}$ for a faster step response.

•Select $R4 = 2780\,\text{k}\Omega$; With C4, C3 and R4 fixed, the only degree of freedom left is on R3, which is computed below for a 5Hz cut-off frequency:

$\mathrm{f}_c = \frac{1}{2\pi\cdot\sqrt{c_3 R_3 c_4 R_4}}$

$R_4 = \frac{1}{\mathrm{f}_c^2\cdot(2\pi)^2 c_3 R_3 c_4} = \frac{1}{5^2(2\pi)^2\cdot 100.10^{-9}\cdot 10\cdot 10^{-9}\cdot 2780.10^3} \therefore \quad R_4 = 99,6k\Omega$

Resulting Bandwidth: **5Hz** and an expected rise time of $\frac{1.8}{2\pi \mathrm{f}_c} = \frac{1.8}{2\pi\cdot 5} = \mathbf{57.30ms}$.

3.2.2. Differential Amplifier design

A non inverting differential amplifier configuration was used. The circuit features a reference voltage or virtual ground, used to perform level shifting on the amplifier output given that only a positive supply was used to power the OpAmp. This will allow a full input and output ranges for the respective signals. The circuit diagram is illustrated in figure 3.2 below:

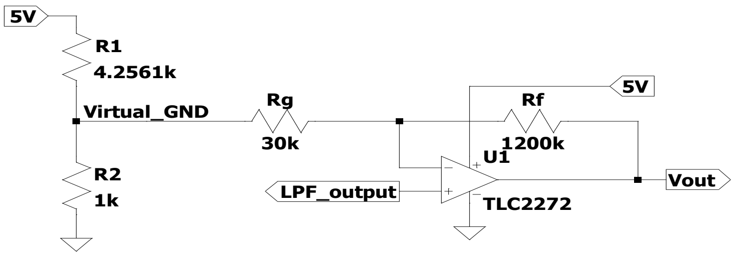


Figure 3.2: Differential Amplifier circuit diagram

•Rf and Rg: for **amplifier gain**

Input range: 930mV(34°C) to 1050mV(42°C) or a 120mVpp swing. For a targeted output range of 0.1 to 4.9V or a 4.8Vpp swing. Hence, gain $\alpha = \frac{4{,}8\text{V}}{120\text{mV}} = 40 = \frac{\text{R}_f}{\text{R}_G}$

Select Rf $= 1200\,\text{k}\Omega$ (large enough to limit current consumption). Hence, Rg $= \frac{1200\,\text{k}\Omega}{40}$

$\therefore \quad \text{Rg} = 30\,\text{k}\Omega$

•R1 and R2: for **virtual ground design**

V^+ and V^- symbols used bellow correspond to the inverting and non-inverting inputs of the Op Amp respectively.

KCL at V^-: $\frac{\text{V}^- - \text{V}_{\text{virtual\_GND}}}{R_G} + \frac{\text{V}^- - \text{V}_{\text{out}}}{R_f} = 0$

$\therefore \text{V}_{\text{out}} = \left(\frac{\text{R}_\text{f}}{\text{R}_\text{G}} + 1\right)\text{V} - \frac{\text{R}_\text{f}}{\text{R}_\text{G}} \cdot \text{V}_{\text{virtual\_GND}}$

$\text{V}^- = \text{V}^+ = \text{V}_{\text{LPF\_out}}$ (Ideal Op Amp)

$\therefore \text{V}_{\text{out}} = \left(\frac{\text{R}_\text{f}}{\text{R}_\text{G}} + 1\right) V_{\text{LPF\_out}} - \frac{\text{R}_\text{f}}{\text{R}_\text{G}} \cdot \text{V}_{\text{virtual\_GND}}$

The goal is to remove the DC component from the amplifier input, 990mV, corresponding to 38°C and introduce a 2.5VDC offset at the output for an optimal output swing.

$\therefore \left(\frac{\text{R}_\text{f}}{\text{R}_\text{G}} + 1\right) \cdot \text{V}_{\text{DC}} - \frac{\text{R}_\text{f}}{\text{R}_\text{G}} \cdot \text{V}_{\text{virtual\_GND}} = 2,5$

$\therefore \left(\frac{1200}{30} + 1\right) \cdot 0,990 - \frac{1200}{30} \cdot \text{V}_{\text{virual\_GND}} = 2,5$

$\therefore \text{V}_{\text{virual\_GND}} = 0.9525\,\text{V}$

Select $\text{R}_1 = 1\,\text{k}\Omega$

$\text{V}_{\text{virtual\_GND}} = \frac{\text{R}_1}{\text{R}_1 + \text{R}_2} \cdot 5$

$\therefore \quad 0,95225 = \frac{1000}{1000 + R_2} \cdot 5$

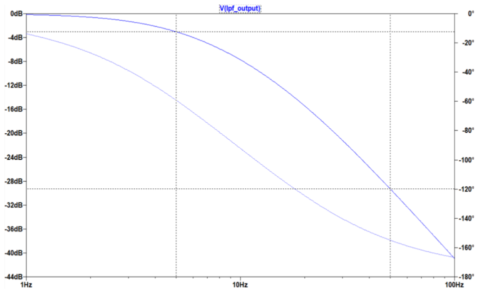
$\therefore \quad \text{R}_2 = 4,2561\text{k}\Omega$

In addition, with $\text{V}_{\text{virtual\_GND}} = 0.9525\,\text{V}$, the voltage at V^- will be at most be 0.9525 V and the voltage at V^+ will vary between 930 mV and 1052 mV, corresponding respectively to 34°C and 48°C plus noise. Hence, VIN_min = 930 mV and VIN_max = 1052 mV; This falls well within the TLC2272 common-mode input voltage range of -0.3 to 4V [5].

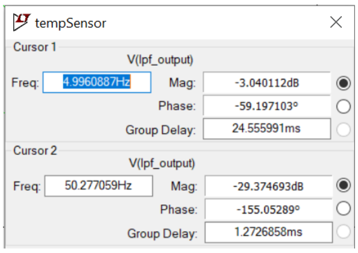
The total drawn current, considering both Op-amps and all resistors was computed as **10.21mA**.

3.3. Results

•LPF frequency response: figure 3.3 below illustrates the frequency response of the LPF.



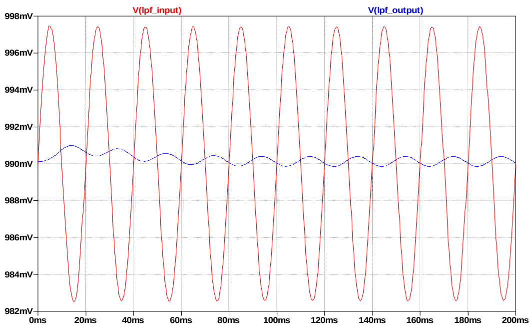
(a) LPF bode plot



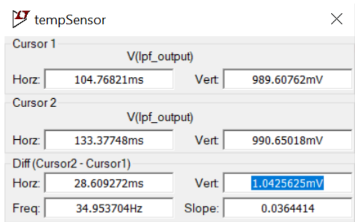
(b) corner frequency

Figure 3.3: Bode plot: corner frequency and amplitude response at 50Hz

•LPF noise suppression: figure 3.4 below illustrates the actual noise suppression performed by the LPF on a 15mVp, 50Hz signal.



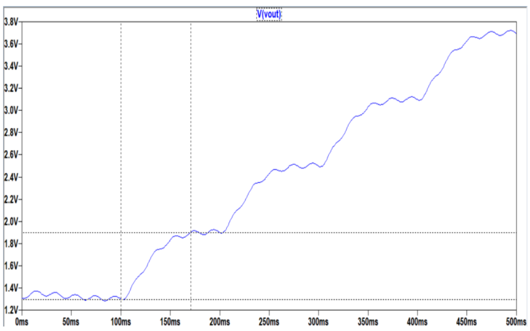
(a) LPF input and output



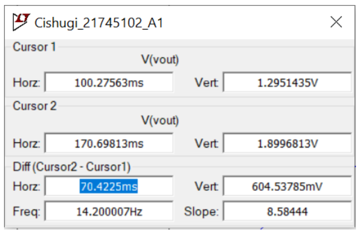
(b) output amplitude in mVpp

Figure 3.4: Low-pass filter noise suppression simulation results

•LPF rise time: figure 3.5 below illustrates the LPF rise time with the Op-amp circuitry included. The plot illustrates a 1°C step in temperature from 36°C to 37°C.



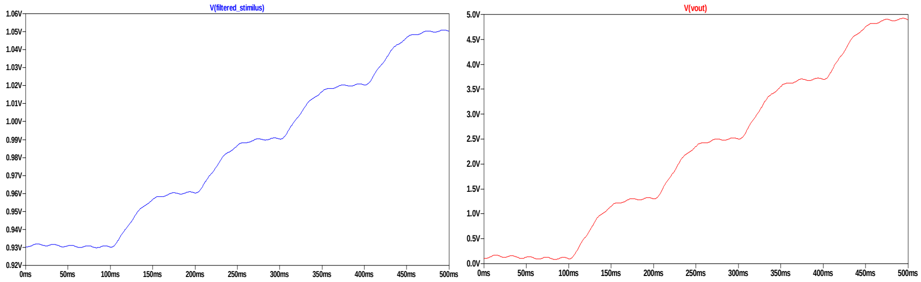
(a) final output volatge



(b) settling time in ms

Figure 3.5: Low-pass filter rise time simulation results with the differential amplifier included

•Op-amp input and output range: figure 3.6 below illustrates the full input range and output range of the differential amplifier circuit.



(a) Op Amp input range **(b)** Op Amp output range

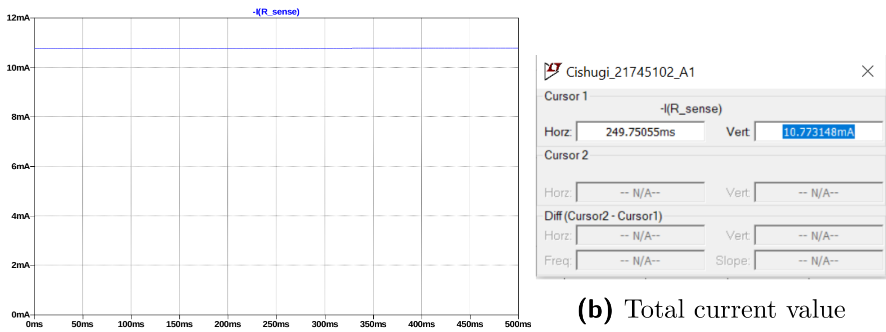
Figure 3.6: Full Op amp input and output range simulation results

Table 3.1 bellow records the measured full input and output ranges as well as the measured output voltages corresponding to temperatures between 34°C and 42°C.

T °C	34	35	36	37	38	39	40	41	42	**Full range**
Stimulus (mV)	930	945	960	975	990	1005	1020	1035	1050	**120 mV**
Vout (V)	0.1	0.7	1.3	1.9	2.5	3.1	3.7	4.3	4.9	**4.8V**

Table 3.1: measured full Op Amp input and output ranges

•**Total current drawn:** figure 4.5 below illustrates the total average current drawn by the temperature sensor conditioning circuit.



(a) Total current **(b)** Total current value

Figure 3.7: Total average current of temperature sensor circuit

3.4. Summary

•The LPF had satisfactory performances: the measured bandwidth(corresponding to f_c) is 4.996Hz(designed for 5Hz) and the attenuation at 50Hz is -29.37dB(close to the expected -30dB). Hence, a 15 mV amplitude, 50Hz noise signal (corresponding to 1°C) is attenuated to 1 mVpp. After amplification, with a gain of 41, the noise will thus be only 20.5 mVp, meeting the "bellow 80 mVp requirement". In addition, the measured rise time is 70.42ms, which, although higher than the 57.30ms designed for, it is smaller than the 100ms required.

•The differential Op-amp circuit had satisfactory performances: The input range was measured to be 120mVpp and the output range 4.8Vpp, greater than the required 3.5Vpp.

•The measured total average current used by the circuit is 10.77 mA (designed for 10.21 mA), which is smaller than the 25 mA required threshold.

Chapter 4

Heart rate sensor

4.1. Introduction

This section covers the design of the **heart rate sensor conditioning** circuit(with a filtering and thresholding modules) as well as a **transducer**.

4.2. Design

4.2.1. Filtering

To design the appropriate filters, the FFT of the heartbeat signals(50bpm, 90bpm and 150bpm) were analysed. The analysis results were recorded in table 4.1 below.

Heart rate	**main frequency component**	**Harmonics**	***low frequency components***	***high frequency components***
50bpm	0.833Hz	at 1.66 and 2,49Hz	between 0.2 and 0.3Hz	at 20, 50 and 60Hz and more
90bpm	1.5Hz	at 3 and 4.5Hz	between 0.2 and 0.3Hz	at 20, 50 and 60Hz and more
150bpm	2.4Hz	at 5 and 7.5Hz	between 0.2 and 0.3Hz	at 20, 50 and 60Hz and more

Table 4.1: Heartbeat signals frequency components from FFT analysis

From the above results, we can observe that unlike the main frequency components and harmonics, which are specific to each heart rate, the low and high frequency components are common to all signals. Additionally, in the time domain, the heartbeat signals have a sinusoidal variation in amplitude of frequency approximately 0.25Hz. Therefore, we can conclude that high frequency components correspond to the superimposed noise while the low frequency components are responsible for the sinusoidal amplitude variations. Hence, the useful frequency components are between 833mHz and 8Hz.

• **Low-pass filter design:** Figure 4.1 below illustrates the LPF design.

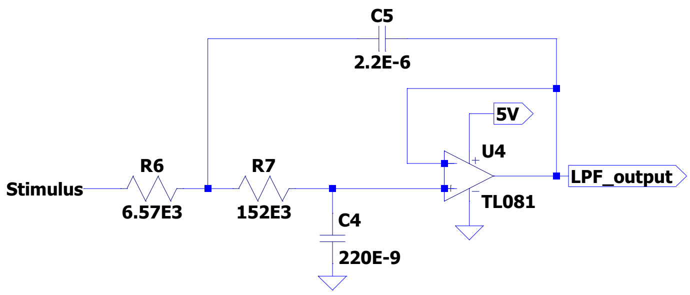


Figure 4.1: LPF circuit diagram

To suppress the high frequency noise, a second order active LPF is used with a corner frequency of 8Hz.
Resistors and capacitors values were computed using the formula: $f_C = \frac{1}{2\pi\sqrt{R_6 R_7 C_4 C_5}}$. The circuit has initially four unknowns. R7, C4 and C5 were picked as: $R7 = 152\,k\Omega$, $C4 = 220\,nF$ and $C5 = 10 \cdot C4 = 2.2\,\mu F$
$\therefore 8 = \frac{1}{2\pi\sqrt{R_6 \cdot 152\,k\Omega \cdot 220\,nF \cdot 2.2\,\mu F}}$, $\therefore R6 = 6.57\,k\Omega$.
• **High-pass filter design:** To suppress the sinusoidal amplitude variation, a second order active HPF is used with a corner frequency of 0,6Hz. Additionally, given that the heartbeat signals are 61.05mVp(without noise), a gain of 20V/V is introduced in the HPF bringing the amplitude to about 1.22Vp to facilitate thresholding. Figure 4.2 below illustrates the HPF design.

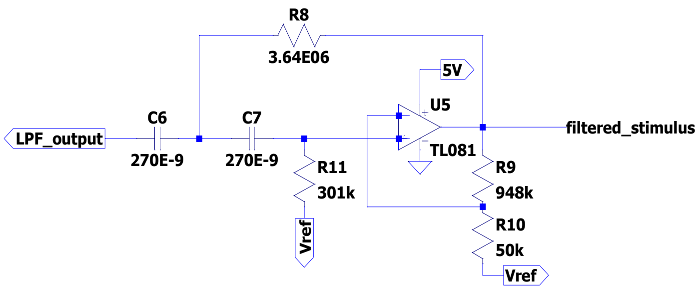


Figure 4.2: HPF circuit diagram

To get R9 and R10 values, the gain equation: $\text{Gain}(Av) = 1 + \frac{R9}{R10}$ is used.
We pick $R10 = 50\,k\Omega$
$\therefore 20 = 1 + \frac{R9}{50\,k\Omega}$; $\therefore R9 = 948\,k\Omega$
Similar to the LPF, to get C6, C7, R8 and R11 values, the formula: $f_C = \frac{1}{2\pi\sqrt{R_8 R_{11} C_6 C_7}}$ is used.
R11, C6 and C7 were picked as: $R11 = 301\,k\Omega$, $C6 = 270\,nF$ and $C6 = C7 = 270\,nF$.
$\therefore 0.6 = \frac{1}{2\pi\sqrt{R_8 \cdot 301\,k\Omega \cdot 270\,nF \cdot 270\,nF}}$, $\therefore R8 = 3.64\,M\Omega$.
To prevent the Op Amp from going into negative saturation, a virtual ground of 2,5V designed with a voltage divider form the 5V regulator output is used. This is labelled is Vref in the above circuit diagram.
In both the LPF and HPF circuits, the Op Amp used is the TL081. In the two cases, VIN_min and VIN_max to the Op Amps are both between 1.78 and 2.10; same thing for VIN_diff and the common-mode input voltages. These values fall well within the maximum ratings of the TL081 Op Amp for the supply used [6].

4.2.2. Thresholding

Thresholding happens in two steps: first using a Schimidt trigger to output square waves then the duration of the pulses is normalised to 200ms using a D flip-flop.
•**Schimdt trigger:** To develop appropriate thresholding, the filtered heart beat signals amplitudes were analysed in the time domain while considering the allowable spec deviations. Table 4.2 below records the analysis results:

Heart rate	DC-offset	min voltage	max voltage	range with spec deviation considered (+- 0.2Vdc and 10% noise)
60bpm	2.5V	1.8	3.4	1.6 to 3.6
150bpm	2.5V	1.65	3.4	1.45 to 3.6

Table 4.2: Thresholding

Hence, we can use 2.3V and 2.7V as low and high threshold voltages (V_{Th_low} and V_{Th_high}) respectively to detect all beats even with the specs deviations. Figure 4.3 below illustrates the Schmidt trigger design.

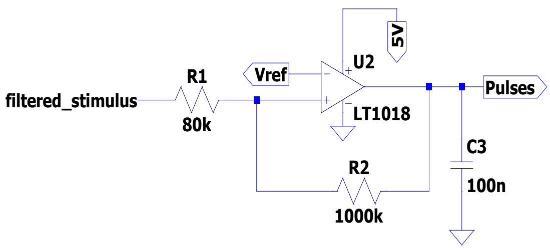


Figure 4.3: Schmidtt circuit diagram

R2 $= 1\,\mathrm{M\Omega}$ and R1 computed using the formula: $V_{in} = \frac{R_1}{R_2}\left(1 + \frac{R_2}{R_1}\right) \cdot V_{ref} - \frac{R_1}{R_2} \cdot V_{out}$
for $V_{in} = V_{Th\_high} = 2.7$, $V_{out} = 0V$
$\therefore 2.7 = \frac{R_1}{1000.60^3}\left(1 + \frac{1000.10^3}{R_1}\right) \cdot 2.5 - \frac{R_1}{1000.10^3} \cdot 0$
for $V_{in} = V_{Th\_low} = 2.3$, $V_{out} = 5V$
$\therefore 2.3 = \frac{R_1}{1000.60^3}\left(1 + \frac{1000.10^3}{R_1}\right) \cdot 2.5 - \frac{R_1}{1000.10^3} \cdot 5$
$\therefore$ R2 $= 80\,\mathrm{k\Omega}$, C3($100\,\mathrm{nF}$)is added to reduce noise on the trigger output.
•**One shot cicruit:** Implemented with a D flip-flop and an R-C network. The design is illustrated in figure 4.4 below.

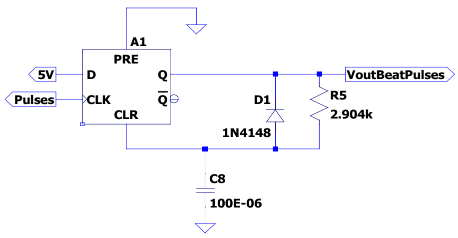
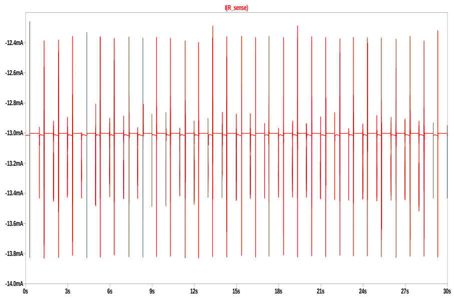


Figure 4.4: One-shot circuit diagram

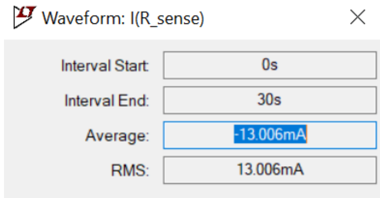
The RC network was designd to achieve a 200ms pulse duration.C8 was picked as C8 $= 100\,\mathrm{\mu F}$ and R5 computed using the formula: $R = -\frac{t}{C \cdot \ln(1 - V_{cap}/V_{final})} = -\frac{0{,}2}{100 \cdot 10^{-6} \cdot \ln\left(1 - \frac{2{,}489}{5}\right)} = 2.904\,\mathrm{k\Omega}$
In the Comparator circuit, the Op Amp used is the LT1018. VIN_min and VIN_max to the Op Amp are both between 1.6 and 3.6; same thing for VIN_diff and the common-mode input voltages. These values fall well within the maximum ratings of the LT1018 Op Amp for the supply used [7].
Transducer: The transducer was designed with a second order LPF with corner frequency 0.65Hz, a level shifting differential amplifier and a peak detector circuit. See Appendix C, figure C.1 for the circuit diagram and full design.

4.3. Results

•**Power supply:** figure 4.5 below illustrates the total current drawn.



(a) total current drawn plot



(b) Average value

Figure 4.5: Total current drawn

•**Filters frequency response:** figures 4.6 below illustrates frequency responses of the LPF and HPF.

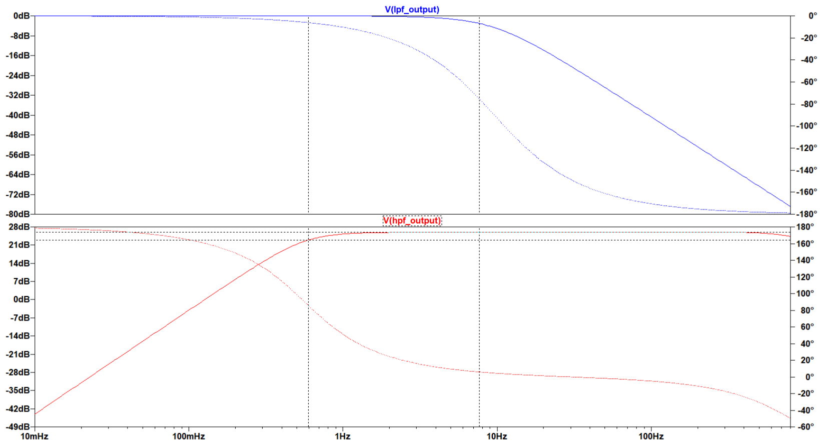


Figure 4.6: LPF and HPF frequency response

Table 4.3 records the measured cuff-off frequencies for the LPF and HPF.

filter	cutoff frequency
LPF	**7.688Hz**
HPF	**596.69mHz**

Table 4.3: LPF and HPF measured corner frequencies

•**Conditioning:** table 4.4 records the measured input and output ranges for 60bpm, 120bpm and 150bpm heartbeat signals. The screen-grabs of the spice simulations can be found in Appendix C in figure C.2.

Heart rate	**input range**	**output range**
60bpm	low:1.817 ;high:2.02	low: 0; high: 5V
120bpm	low:1.799 ;high:2.00	low: 0; high: 5V
150bpm	low:1.789 ;high:2.01	low: 0; high: 5V

Table 4.4: Input and output range measurements

•**Thresholding:** table 4.5 below records the measured high and low threshold voltages for a 60bpm and 150bpm heartbeat signals. The screen-grabs of the spice simulations can be found in Appendix C in figure C.3.

Heart rate	**measured low threshold**	**measured high threshold**	*Designed thresholds*
60bpm	2.356V	2.70V	low: 2.3V
150bpm	2.282V	2.769	high: 2.7V

Table 4.5: thresholding measurements

•**Pulses duration:** table 4.6 below records the measured output pulses duration for a 60bpm and 150bpm heartbeat signals. The screen-grabs of the spice simulations can be found in Appendix C in figure C.4.

Heart rate	**measured output pulse duration**
60bpm	196.1655ms
150bpm	187.8953ms

Table 4.6: Output pulses duration

•**Transducer:** table 4.7 records the measured range and settling time of the transducer. The screen-grabs of the spice simulations can be found in Appendix C in figure C.5.

Heart rate	**Transducer or analog output**	**Settling time**
50bpm	0.7855V	318ms
150bpm	4.6851V	655ms
achieved output range:	4.6851 - 0.7855 = 3.8996V	

Table 4.7: Transducer output range and settling time

4.4. Summary

From the above simulations results:
•The total current drawn is averaged at 13.00mA(smaller than the 15mA current requirement).
•The two filters had satisfactory performances: the measured the LPF corner frequency was 7.69Hz(designed for 8Hz) while the measured HPF corner frequency was 0.597Hz and a 25.96dB gain(designed for 0.6Hz and 26.02dB).
•The full heart rate sensor minimum to maximum output ranges from 0 to 5V while the input ranges from 1.817 to 2.02V.
•The measured low threshold, considering the allowable spec deviations varies between 2.82 and 2.356V (designed $V_{Th\_low} = 2{,}3V$) while the high threshold varies between 2.70 and 2.79 (designed $V_{Th\_high} = 2{,}7V$)
•The measured output pulses duration varies between 187.895ms(for 150bpm) and 196.166ms(for 60bpm), close to the 200ms design value and grater than the 150ms requirement.
•The measured transducer output range is 3.8995V and an average settling time smaller than 1s.
All these values meet the necessary requirements.

Chapter 5

Calibration and digitisation

This section covers the calibration and digitisation of the designed temperature and heart rate sensors conditioning circuits.

5.1. Temperature sensor

To perform the calibration, the conditioning circuit output voltages for temperatures ranging from 34 to 42°C were recorded. These, as well as the expected output voltages for each temperature are presented in table 5.1 below.

T°C	34	35	36	37	38	39	40	41	42
Sensor output[V]	0.93	0.945	0.960	0.975	0.990	1.005	1.020	1.035	1.050
Expected analog output [V]	0.1	0.7	1.3	1.9	2.5	3.1	3.7	4.3	4.9
Measured output [V]	0.10516	0.70412	1.30308	1.90204	2.50100	3.09996	3.69892	4.29788	4.89685

Table 5.1: Temperature Calibration

5.1.1. Analytical Design

Using the expected output voltages[V_{out}], the following empirical formula was derived:
$T\ °C = \frac{5}{3} \cdot V_{out} + \frac{203}{6} = 1.666666667 \cdot V_{out} + 33.83333333$

5.1.2. Empirical Design

Using the measured output voltages [V_{out}], we obtained the following calibration: $T\ °C = 1.669576668 \cdot V_{out} + 33.82431235$. This gives an updated formula to evaluate the temperature from the conditioning circuit output. Figure 5.1 below illustrates the evaluation of the performed calibration:

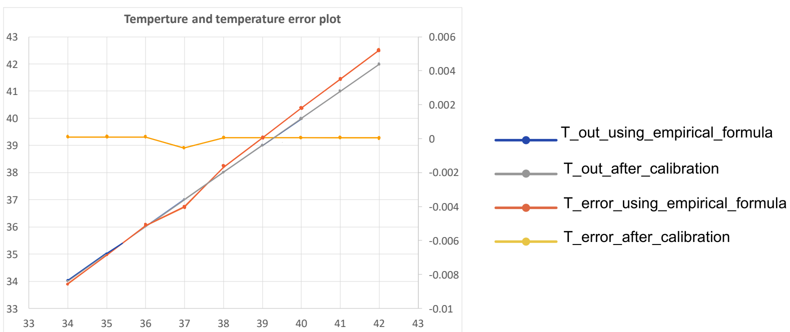


Figure 5.1: Temperature plots on primary axis, error plots on the secondary axis

As expected, given that the measured values are from a simulation and not actual physical measurements, the calibration and empirical formulae are very similar. However, the temperature errors remain under 0.001°C with the calibration formula.

5.1.3. Digitisation

•**Python digitisation**: the used algorithm flow diagram is presented in the list below
1.Get the average of the last 500 elements fo the output voltage array [Amplitude] : V_{out_avg}
2.Using the calibration formula:
$T\ °C = 1.669576668 \cdot V_{out\_avg} + 33.82431235$

• **5 volts, 10-bit ADC digitisation**: The ADC would read the conditioning circuit output $\mathbf{V_{out}}$ and outputs a digital value: **ADC_value**
$ADC\_value = V_{out} \cdot \frac{ADC\_resolution}{V_{ref}} = V_{out} \cdot \frac{2^{10}}{5} = 204.8 \cdot V_{out}$
To get the temperature value from the ADC_value, the following calibration formula will be used:
$T\ °C = 0.00815223 \cdot V_{out} + 33.82431235$

5.2. Heart rate sensor

Digitising the heart rate sensor conditioning circuit involves computing the heart rate captured by the sensor using the square wave signal output from the conditioning circuit. This was done with a python script. The used algorithm flow digram is presented in the list below:

- Starting from 2 s, read the conditioning circuit output voltage (from the 6 s simulation): Amplitude array.
- Check for steps of more than 4.8V in the array values : low-to-high transistions
- For each of those transitions, store the current time from the time array into another array: **transition_time_array**.
- Run into the **transition_time_array** and compute the period for each cycle : The difference between two consecutive "transition_times" corresponds to 1 period. And store the periods into a **"period_array"**
- The period of the signal is the average of the "period_array" elements: **average_period** in ms.
- The heart rate in bpm is thus, **BPM** $= \frac{60*1000}{average\_period}$

Chapter 6

System and conclusion

6.1. System

Overall, the designed Health monitoring system works as expected. The simulation results have proven satisfactory performances for both the temperature and heart rate conditioning circuits. In addition after calibration and digitisation, tests runs on both circuits for several heart rates and temperature yielded readings with a 100% accuracy.

•**Temperature sensor conditioning circuit:** Upon realisation of its design into a physical circuit, the conditioned analogue temperature signal, from the circuit output shall be connected to a microcontroller ADC where digitisation, as per developed calibration shall occur to produce a numeric temperature value. This section of the system drew 10.77mA on average.

•**Heart rate sensor conditioning circuit:** upon realisation of the heart rate sensor conditioning design into a physical circuit, its outputs will also be connected to a microcontroller. While the output pulses are best suited toward the microcontroller DAC input, the analog transducer outputs are designed for the microcontroller ADC input. This section of the system drew 13.00mA on average.

The total current drawn by the full designed health monitoring system is thus only
$10.77 + 13.00 = 23.77$mA.

6.2. Lessons learnt

•There are various ways to design a circuit given its specifications: for instance, in the transducer, while complex circuits and specialised components are advisable for such a task, a simple LPF, a level shifting differential amplifier and a peak detector were used and the given requirements were met.

•The earlier you start to work an a certain task, the more you enjoy doing it.

•How to select a certain filter configuration and design it given a bandwidth and rise time specifications.

•The full design of a differential amplifier with a virtual ground given specific requirements.

Bibliography

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[2] Renesas Electronics Corporation, "Linear vs. switching regulators," 2021. [Online]. Available: https://www.renesas.com/cn/en/products/power-management/linear-vs-switching-regulators.html

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[6] *General purpose JFET single operational amplifiers*, TL081 datasheet, ST Microcontroller, June 2008.

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Appendix A

Social contract

Sign and inlcude.

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E-design 344 Social Contract

2020

The purpose of this document is to establish commitment between the student and the organisers of E344. Beyond the commitment made here, it is not binding.

In the months preceeding the term, the lecturer (Thinus Booysen) and the Teaching Assistant (Michael Ritchie) spent countless hours to prepare for E344 to ensure that you get your money's worth and that you are enabled to learn from the module and demonstrate and be assessed on your skills. We commit to prepare for the module, to set the tests and assessments fairly, to be reasonably available, and to provide feedback and support as best and fast we can. We will work hard to give you the best opportunity to learn from and pass analogue electronic design E344.

Signature: .... Date: 13 July 2020 ....

....

I, Elijah Seth Cishugi .... have registered for E344 of my own volition with the intention to learn of and be assessed on the principals of analogue electronic design. Despite the potential publication of supplementary videos on specific topics, I acknowledge that I am expected to attend the lectures and lab sessions to make the most of these appointments and learning opportunities. Moreover, I realise I am expected to spend the additional requisite number of hours on E344 as specified in the yearbook.

I acknowledge that E344 is an important part of my journey to becoming a professional engineer, and that my conduct should be reflective thereof. This includes doing and submitting my own work, working hard, starting on time, and assimilating as much information as possible. It also includes showing respect towards the University's equipment, staff, and their time.

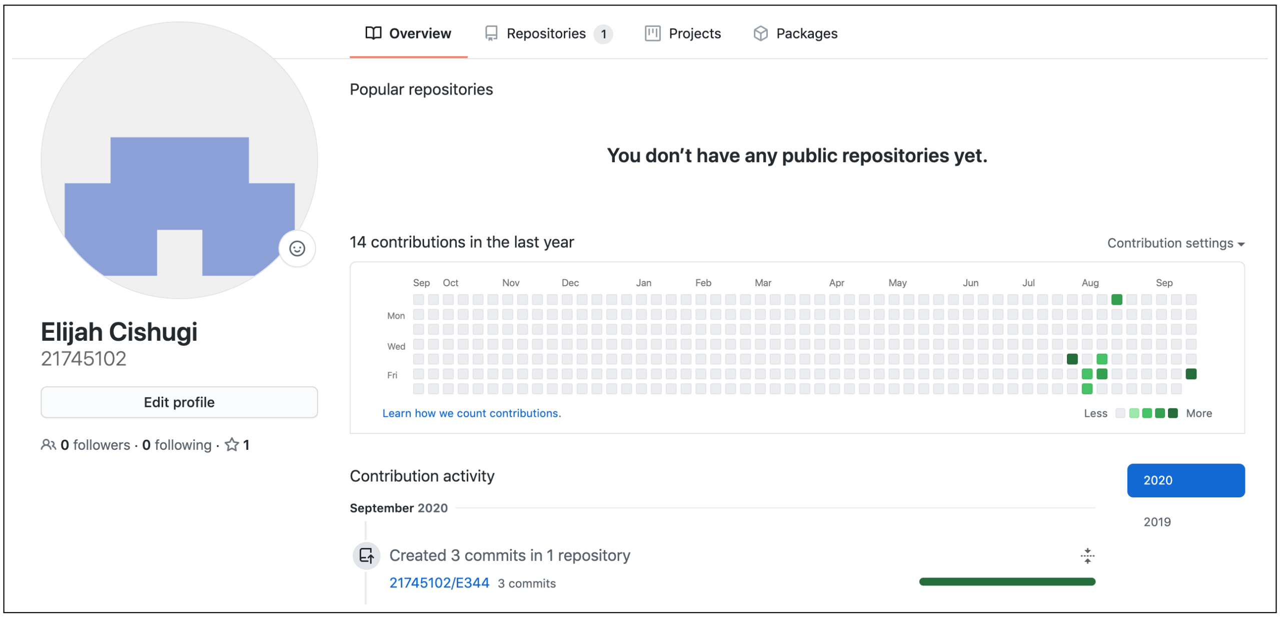
Signature: .... Date: 16th / August / 2020 ....

1

Appendix B

GitHub Activity Heatmap

Take a screenshot of your github version control activity heatmap and insert here.



Appendix C

Additional Circuit diagrams and Spice simulation results

1.Transducer circuit diagram:

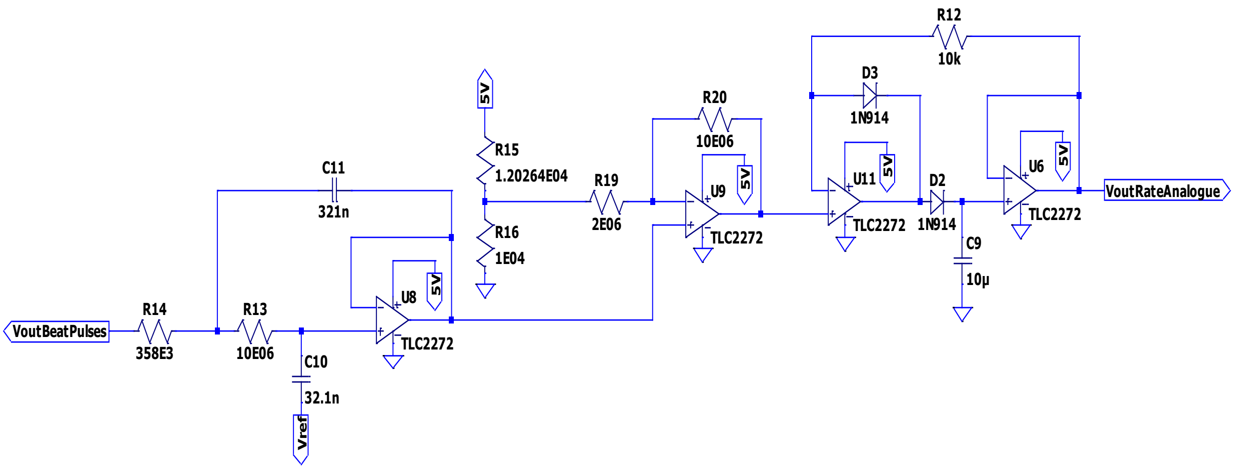


Figure C.1: Transducer circuit diagram

•LPF:

a LPF with corner frequency 0.65Hz is used to allow different attenuations to the pulses depending on the frequency.

Resistors and capacitors values were computed using the formula: $f_C = \frac{1}{2\pi\sqrt{R_{13}R_{14}C_{10}C_{11}}}$. The circuit has initially four unknowns. R13, C10 and C11 were picked as: R13 $= 10\,\text{M}\Omega$, C10 $=$ $32.1\,\text{nF}$ and C11 $= 10 \cdot \text{C10} = 321\,\text{nF}$

$\therefore 0.65 = \frac{1}{2\pi\sqrt{R_{14}\cdot 10\,\text{M}\Omega\cdot 32.1\,\text{nF}\cdot 321\,\text{nF}}}$, $\therefore$ R14 $= 358\,\text{k}\Omega$.

•Level shifting differential amplifier: Designed for a gain $A_v = 6$ and a 2.7V DC offset.

$A_v = 1 + \frac{R_{20}}{R_{19}}$, we pick R20 $= 10\,\text{M}\Omega$

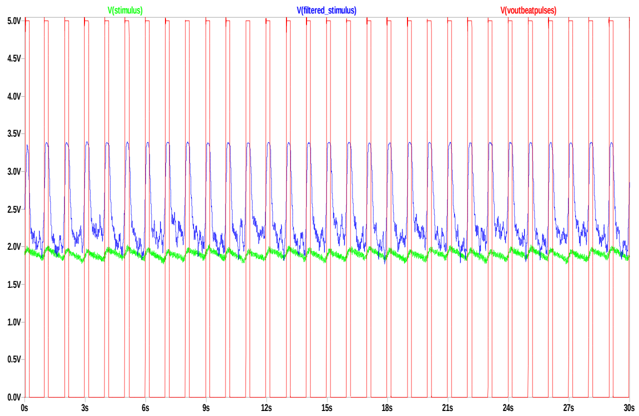
$\therefore$R10 $= 2\,\text{M}\Omega$

$V_{\text{off\_set}} = \frac{R_{16}}{R_{16}+R_{15}} \cdot 5$; We pick R16 $= 10\,\text{k}\Omega$

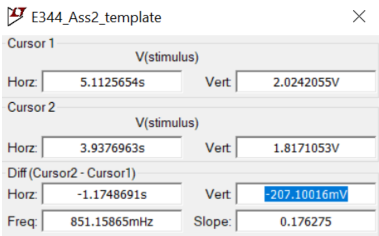
$\therefore$R15 $= 12.0264\,\text{k}\Omega$

•Peak detector: The configuration with two diodes and two Op Amp with a feedback resistor R12, was chosen to minimise noise and suppress any ripple at the transducer output. In the transducer circuit, the Op Amp used is the TLC2272. The minimum and maximum input voltages(VIN_min and VIN_max) to the Op Amp are both between 0 and 4.6; same thing for the differential(VIN_diff) and common-mode input voltages. These values fall well within the maximum ratings of the LT2272 Op Amp for the supply used [5].

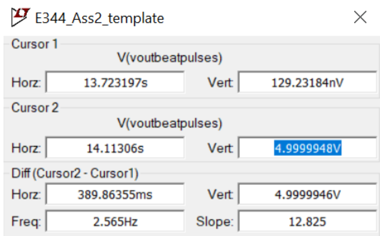
2.Input and output range measurements:



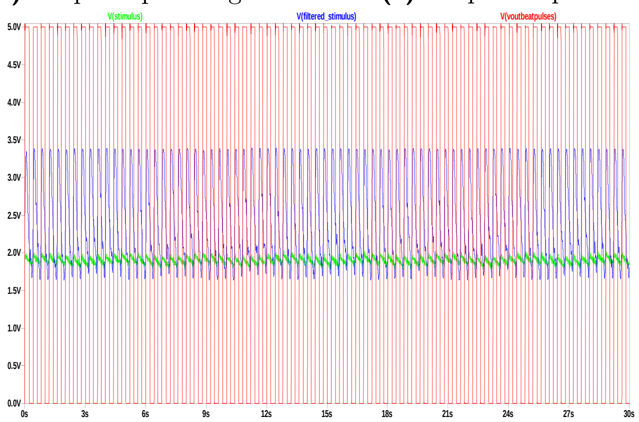
(a) 60bpm



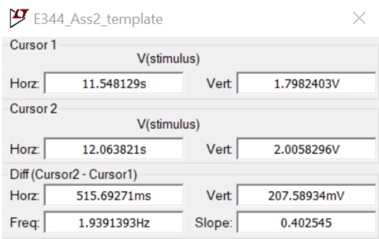
(b) 60bpm input range



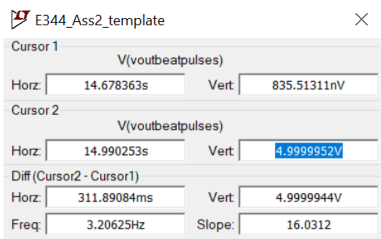
(c) 60bpm output range



(d) 150bpm



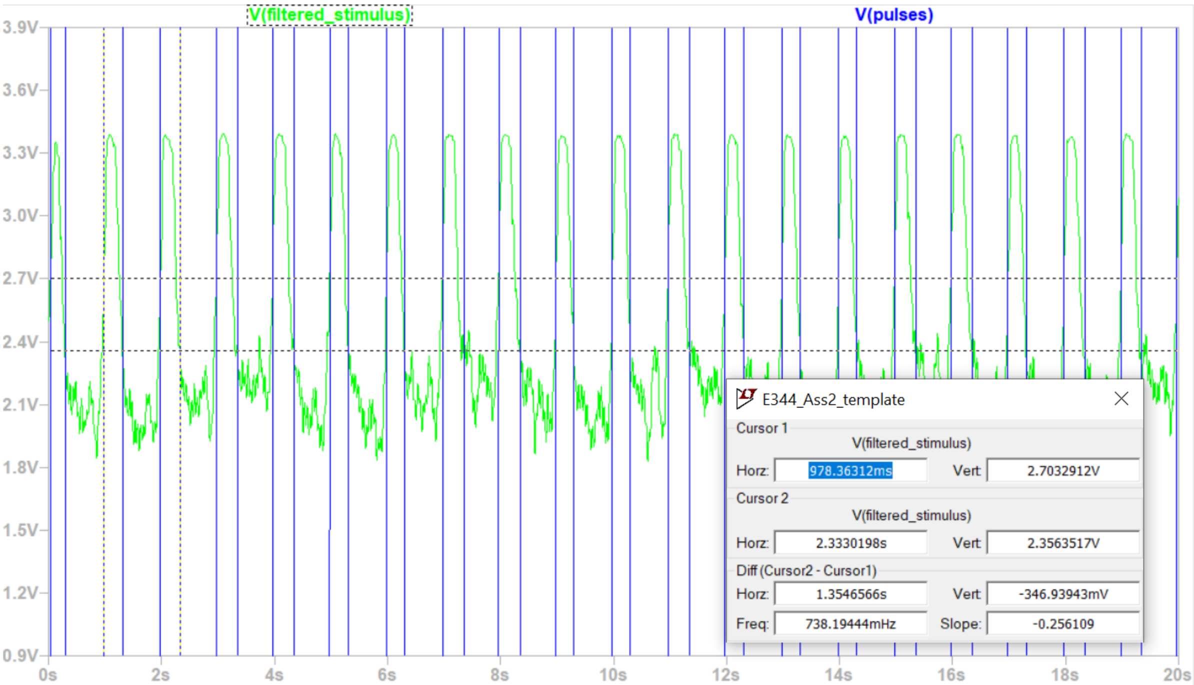
(e) 60bpm input range



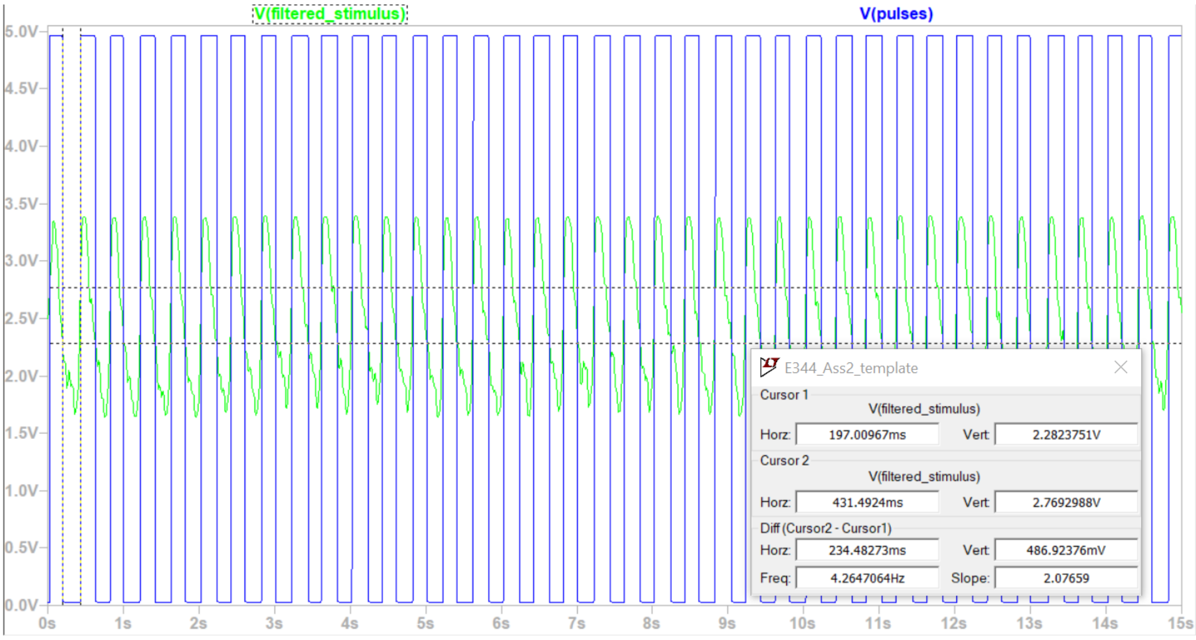
(f) 60bpm output range

Figure C.2: Input and Output ranges measurements

3.Thresholding measurements:



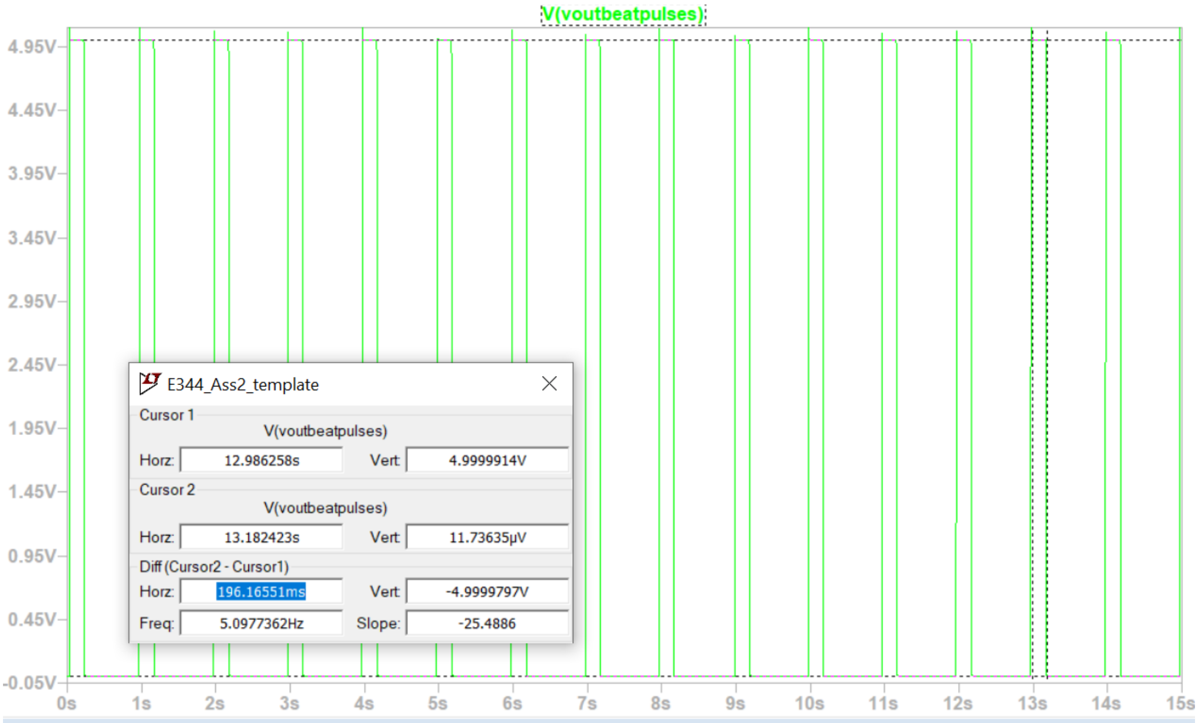
(a) 60bpm: cursor1 = measured high threshold, cursor 2 = measured low threshold



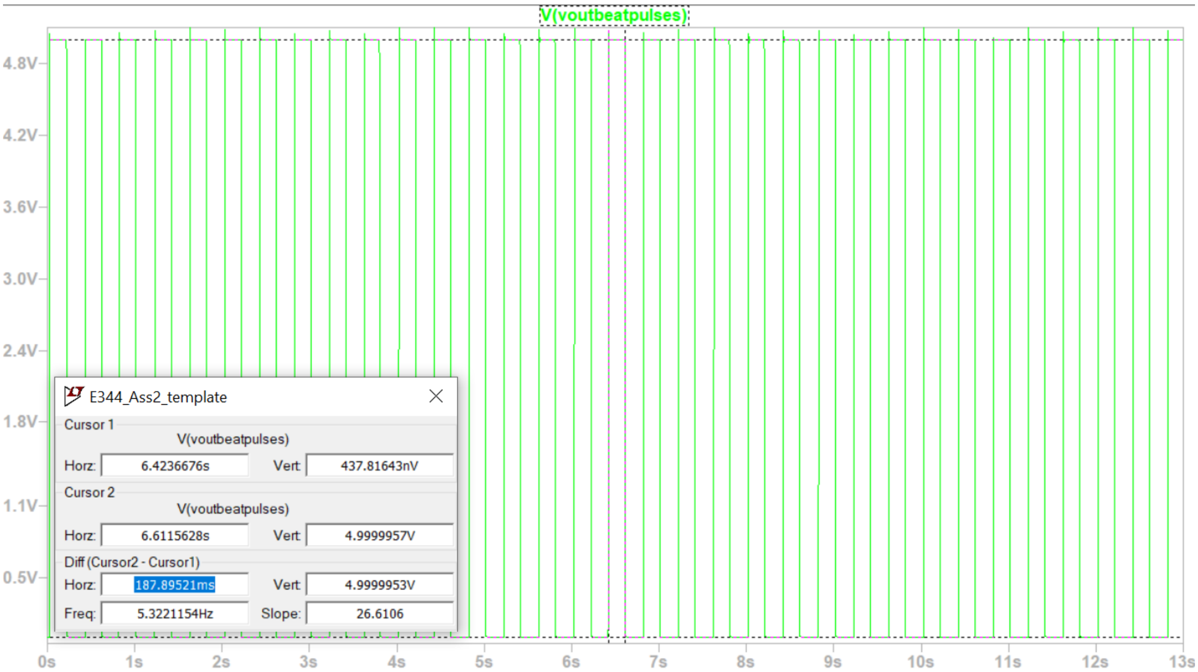
(b) 150bpm: cursor2 = measured high threshold, cursor 1 = measured low threshold

Figure C.3: Thresholding measurements

4.Pulses duration measurements:



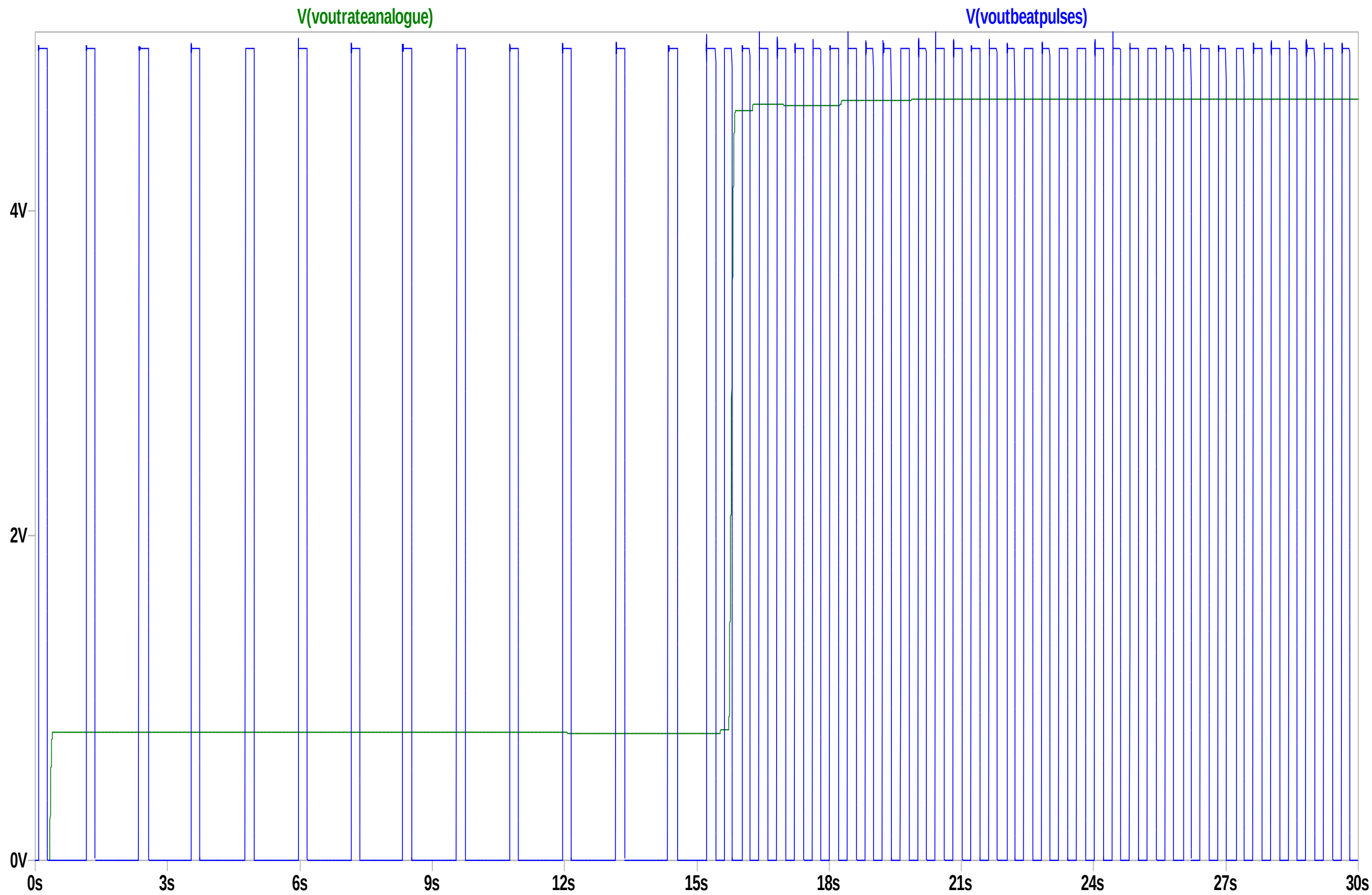
(a) 60bpm



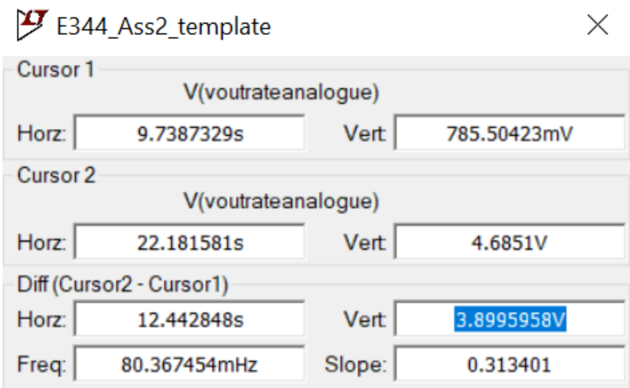
(b) 150bpm

Figure C.4: Pulses duration measurements

5.Transducer output range and settling time:



(a) 50bpm analog output: 0.7855V, settling time: 318ms; **150bpm** analog output: 4.6851V settling time: 655ms



(b) cursor reading

Figure C.5: Transducer output measurements