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E344 Assignment 2

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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.

September 30, 2020

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D.H. von Eschwege Voorletters en van / *Initials and surname*	September 30, 2020 Datum / *Date*

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Nomenclature

Variables and functions

f_c	Cutoff frequency
V_{REF}	Reference voltage
A_v	Gain
β	Feedback fraction
V_{CM}	Common-mode voltage
V_{ID}	Differential-mode voltage

Acronyms and abbreviations

PWM	Pulse-width modulation
FFT	Fast Fourier transform
DC	Direct Current
Q-factor	Quality factor
BPM	Beats per minute
UTP	Upper trip point
LTP	Lower trip point

Chapter 1

System design

1.1. System overview

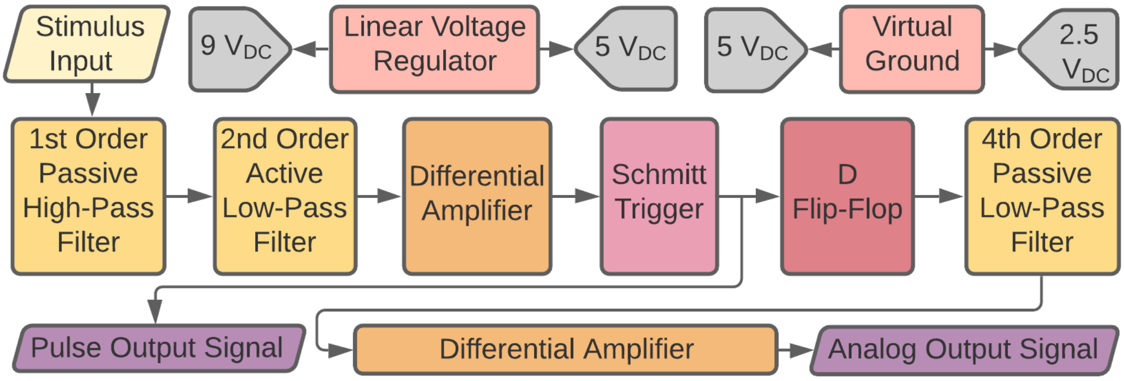


Figure 1.1: System Diagram

A heart-rate sensor is designed to receive an input signal from which pulses and analogue values are generated, corresponding to heart-beats and the heart-rate respectively. This report discusses voltage regulation, signal conditioning, pulse generation and conversion to analogue, as shown in figure 1.1. The circuit is powered by a voltage regulator [1] which can supply 100 mA, of which 12.8 mA is used for the temperature sensor - see E344 Assigment 1 [1] - leaving 87.2 mA at 5 V for Assignment 2 and 3, mandating design for conservative current use. The input signal has an amplitude of insufficient magnitude for conversion, and is subject to noise, necessitating signal conditioning: a first order passive high-pass filter and a second order active low-pass filter attenuate both high- and low-frequencies. Chosen with maximal simplicity in mind to reduce cost and complexity, the filters still performing adequately. Thereafter, a differential amplifier produces a signal with a large amplitude and little noise. A Schmitt Trigger then outputs a pulse signal with a frequency corresponding to the heart-rate. The Schmitt Trigger was chosen as it provides a noise margin via hysteresis. An analogue voltage output is required for the microcontroller - filtering and peak detection using diodes was considered but discarded, as non-linear diodes result in extremely slow simulation. Rather, the pulse output signal was converted to a pulse-width modulated signal, where the frequency of the former determines the duty cycle of the latter. This was done as PWM signals lend themselves to conversion-to-analogue by simple filtering. The PWM signal was obtained by using a D Flip-Flop and a RC-circuit (see section 2.2). A third order passive RC filter is then used - passive components reduce current usage and simulation time. The filter is of high order as to minimize noise, while meeting the settling time requirement. Finally, the signal was amplified to achieve the required range.

Chapter 2

Heart rate sensor

2.1. Introduction

Circuits pertaining to signal conditioning, pulse signal and analogue output generation will be discussed. Conditioning is done via filtering and amplification, thereafter thresholding and pulse-width modulation is used to generate outputs. Active filters provide high input and low output impedance, and a large Q-factor [2]. A differential amplifier is suitable for amplification, as the gain is referenced against a customizable voltage [3]. A Schmitt Trigger is well-suited for pulse generation, as it provides a noise margin [4]. PWM signals can be converted to analogue using filtering [5].

2.2. Design

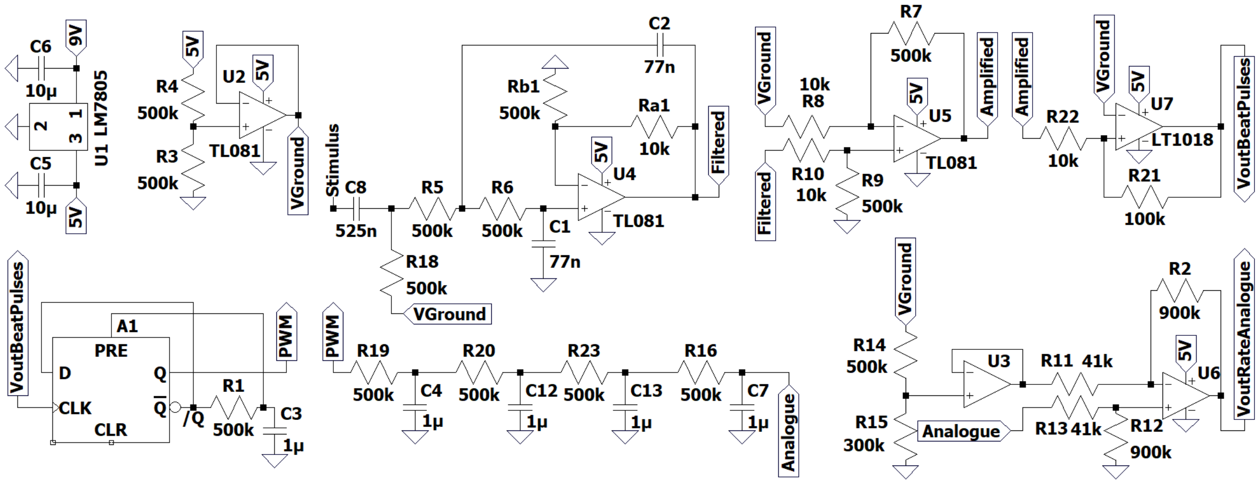


Figure 2.1: Complete Circuit

The complete circuit is shown upfront in figure 2.1 to aid explanation. The largest resistor in sub-circuits is always chosen to be 500 kΩ as to reduce current usage. The TLC2272 op-amp has V_{CM} of -0.3 to 4 V, $V_{ID} = \pm 16\ V$, $V_{in_{max}} = V_{DD+}$ and $V_{in_{min}} = V_{DD-} - 0.3$ [6]. The TL081 has V_{CM} of 1 to 4 V, $V_{ID} = \pm 30\ V$, $V_{in_{max}} = 3.5\ V$ and $V_{in_{min}} = 1.5\ V$ [7]. This information is given here as to avoid repetition - refer here when op-amp characteristics are discussed. The design process now follows. The FFT of the stimulus input in figure 2.2 shows information residing between 0.8 to 2.5 Hz, corresponding to 50 and 150 BPM respectively. The other peaks represent noise at 0.25 Hz, as well as at twice and three times the message signal, and higher. Noise distorts square wave output, necessitating filtering.

A first order passive high-pass filter, $f_c = 0.606\,\text{Hz}$, attenuates the low frequency noise at 0.25 Hz. R18 = 500 kΩ, thus C8 = 525 nF according to $f_c = \frac{1}{2\pi RC}$. A virtual ground centres the signal around 2.5 V, ensuring that the LPF input falls within the common-mode range of the TL081, which is used as it inexpensive [8]. A second order active low-pass filter, $f_c = 4.1\,\text{Hz}$, filters out high frequency noise. R5 = R6 = 500 kΩ, C1 = C2 = 77 nF - see aforementioned formula. Cutoff frequencies were selected to remove noise maximally while minimally affecting heart-rate data. The signal should reside slightly above 2.5 V to facilitate amplification (to be discussed) and to ensure common-mode input range compliance in the next stage. Thus, Rb1 = 500 kΩ and Ra1 = 10 kΩ since $A_v = 1 + \frac{R_A}{R_B}$ [9]. A filter output with DC offset slightly above 2.5 V allows for the use of a differential amplifier with the existing virtual ground connected to the negative input, removing the need for additional circuitry otherwise required. The signal is amplified according to $\mathrm{V_{OUT}} = \frac{\mathrm{R}_a}{\mathrm{R}_b}(\mathrm{V}_2 - \mathrm{V}_1)$ [3], where $\mathrm{R_a}$ corresponds to R7 and R9 and $\mathrm{R_b}$ to R8 and R10. The gain of 50 was selected to again provide a DC offset of 2.5 V, as it facilitates implementation of the comparator (to be discussed). Since the amplified signal has an amplitude of only 1.66 V, the inexpensive TL081 was chosen despite having a smaller output range. Next, the signal is fed into a Schmitt Trigger - the LT1018 comparator allows for output very close to 0 and 5V and has a high gain. 5 V is produced if the input exceeds the upper trip point (UTP) and 0 V if the input falls below the lower trip point (LTP) [4]. The hysteresis width (w_h = UTP - LTP) serves as a noise margin around the reference voltage, V_{REF} [4]. The DC offset of 2.5 V of the amplified signal requires $V_{REF} = 2.5\,\text{V}$, once again allowing for the use of the existing virtual ground. Pulse duration is designed for the highest frequency, as it will then also meet the requirement for lower frequencies. Since 150 BPM corresponds to a period of 0.4 s, w_h should be selected to ensure an output pulse width of at least 150 ms; 200 ms selected to account for noise. The amplified signal approximates a sinusoid:

$$y_1 = 1.66\sin(2\pi(2.5)t_1) + 2.5 \qquad (2.1) \qquad y_2 = 1.66\sin(2\pi(2.5)t_2) + 2.5 \qquad (2.2)$$

Subtracting 2.2 from 2.1 yields $y_1 - y_2 = w_h = 0.52\,\text{V}$ for t_1 =0.01 s and $t_2 = 0.21\,\text{s}$, resulting in a theoretical pulse width ranging from 0.2 to 0.625 ms, which is sufficient. w_h also is an order of magnitude larger than the highest input signal noise levels, increasing design fidelity. Now, UTP = 2.75 V, LTP = 2.25 V, UTP = $V_{REF} + \beta Vcc$ and LTP = $V_{REF} - \beta Vcc$, thus β = 0.05 [4]. Further, $\beta = \frac{\mathrm{R}_{22}}{\mathrm{R}_{22}+\mathrm{R}_{21}}$. Thus R21 = 190 kΩ and R22 = 10 kΩ. (R21 was adjusted to 100 kΩ to account for loading effects). A one-shot was considered to extend the pulses, but was omitted as it was superfluous. The pulse output frequency thus represents the heart-rate.

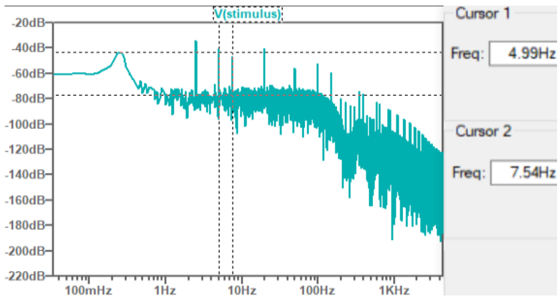


Figure 2.2: Fast fourier transform of stimulus

Obtaining an output suitable for a microcontroller presents itself as a problem of converting frequency to analogue. The decision was made to obtain a PWM signal from the FM pulse signal, as PWM signals readily lend themselves to conversion-to-analogue using a simple low-pass RC filter. Pulse-width modulation was achieved by means of a D flip-flop. The essence of the design is to firstly keep the output high continually, but to then drive the output low for a fixed time, once per period. Since the period of the clock signal varies against BPM, driving low for a fixed amount of time results in a different ratio of the pulse being low at different frequencies, producing a varying duty cycle. Specifically, the previously generated pulse signal is used as a clock signal for the D flip-flop, and the not-Q output is fed back into the data line. This would normally drive the output high for all time t; thus a RC circuit is added of which the capacitor is charged by the not-Q output, and connected to the SET-line of the flip-flop. Once the capacitor charges to 2.5 V, Q is pulled high and not-Q low until the next rising edge of the clock input inverts Q and not-Q. The result is a slightly delayed PWM signal, with a large duty cycle at high frequencies, and vice-versa. The result is a consistent PWM output, as long as the capacitor charge time is selected to be shorter than the narrowest pulse of the clock signal. See section 2.3, figure 2.7a for output. Calculations follow. R1 $= 500\,\mathrm{k\Omega}$ gives $C = 1\mu$, according to the capacitor charge/discharge exponential equations - the extended solution (via Matlab scripting) is given in Appendix C for the sake of brevity. Having obtained a PWM signal, a fourth order passive low-pass filter produces analogue values in the range of 0.95 to 1.2 V. A cutoff frequency of 0.5 Hz ensures maximal attenuation of noise while maintaining a 5% settling time of 10 seconds. Therefore R $= 500\,\mathrm{k\Omega}$ and C $= 1\,\mu$. Finally, the analogue output is amplified by means of a TLC2272 differential amplifier with a negative input at 0.938 V. A gain of 20 produces a sufficient output range. Resistor values were increased as to minimize loading effects - thus R2 = R12 = $900\,\mathrm{k\Omega}$ and R11 = R13 = $45\,\mathrm{k\Omega}$ according to $\mathrm{V_{OUT}} = \frac{\mathrm{R}_a}{\mathrm{R}_b}(\mathrm{V}_2 - \mathrm{V}_1)$ [3]. Loading effects still had a slight effect, and required adapting R11 and R13 to $41\,\mathrm{k\Omega}$. An analogue transducer is thus designed. For microcontroller integration, a log-linear scale can linearise the analogue output, and the calibration constant is calculated. The slope is $\frac{0.8-2.5}{4.7-0.5} = -0.404$. Thus, for 150 BPM, $f = 2.5 = -0.404V + C = -0.404(0.5) + C \quad \rightarrow \quad C = 2.702$.

Total current is calculated using 1.4 mA per TL081 [7], 2.2 mA per TLC2272 [6] and 110 μA per LT1018 [10]. Non-trivial resistors are all approximated using 5 V over, as to err on the side of caution: $4(1.4m) + (2.2m) + (110n) + 16\left(\frac{5}{500k}\right) + 3\left(\frac{5}{10k}\right) + 2\left(\frac{5}{41k}\right) + \left(\frac{5}{100k}\right) = 9.75\,\mathrm{mA}$. Voltage regulator current is considered in E344 Assignment 1 [1]. $100m - 9.75m - 11.39m$ leaves 78.86 mA for Assignment 3.

2.3. Results

The frequency response of the high- and low-pass filters gives $-3dB$ at 0.61 Hz and $-6dB$ at 4.3 Hz respectively - see figures 2.4a and 2.4b. This reflects design calculations near perfectly. Comparison of filter input and output in figure 2.3 shows a drastic noise reduction.

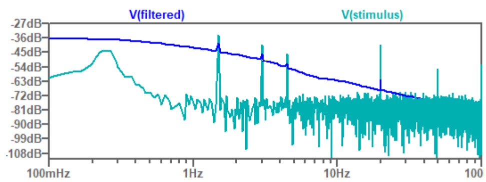
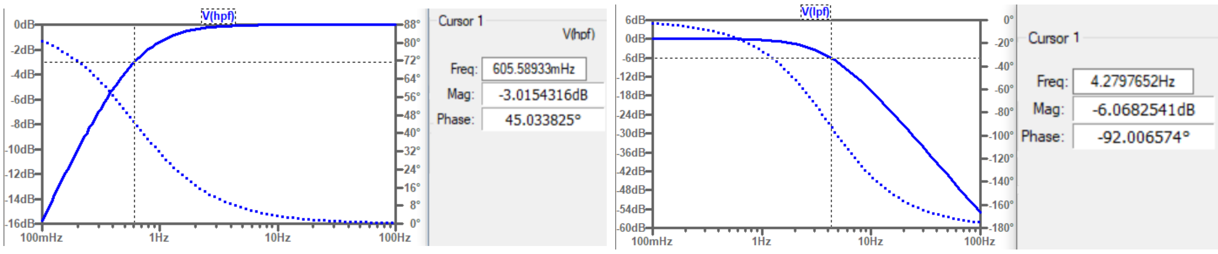


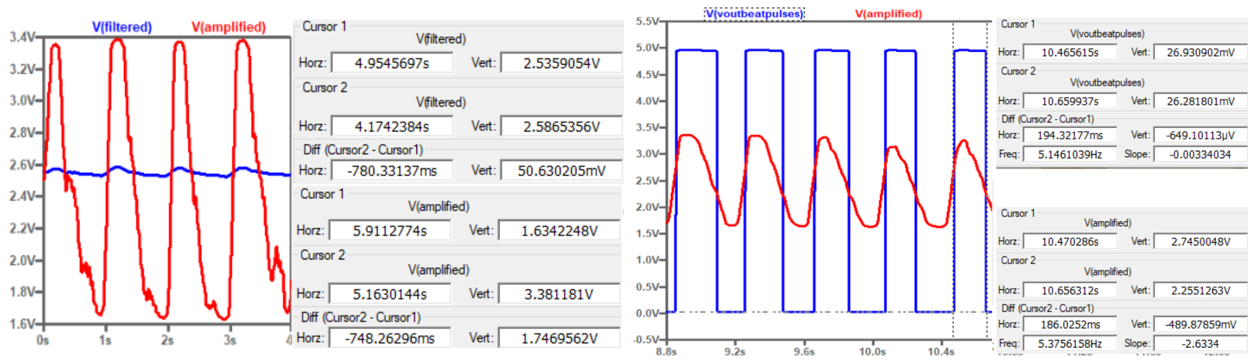
Figure 2.3: Filter input and output



(a) High-pass filter **(b)** Second-order low-pass filter

Figure 2.4: Frequency response of filters

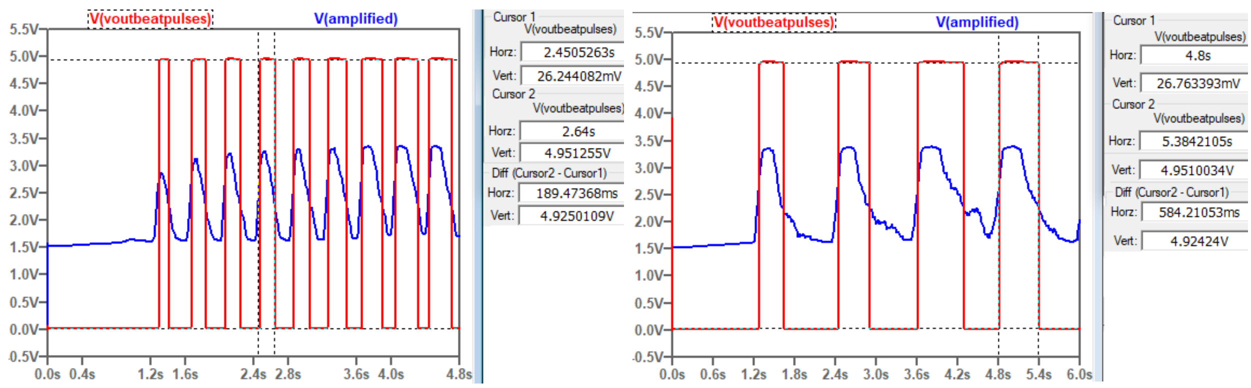
The filtered signal is passed through a differential amplifier; the input is shown in blue and the output in red in figure 2.5a. The filtered signal has a 50 mV amplitude (peak-to-peak), and the output 1.747 V. Therefore $A_v \approx 35$, which, although smaller than calculated, is to be expected due to non-linear behaviour for high amplification factors. Input and output of a LT1018 comparator, used for thresholding, is shown in figure 2.5b. See UTP = 2.75 V and LTP = 2.25V in figure 2.5b, giving a hysteresis width of 0.5 V, exactly as calculated.



(a) Amplification of filtered signal **(b)** Thresholding

Figure 2.5: Signal Conditioning

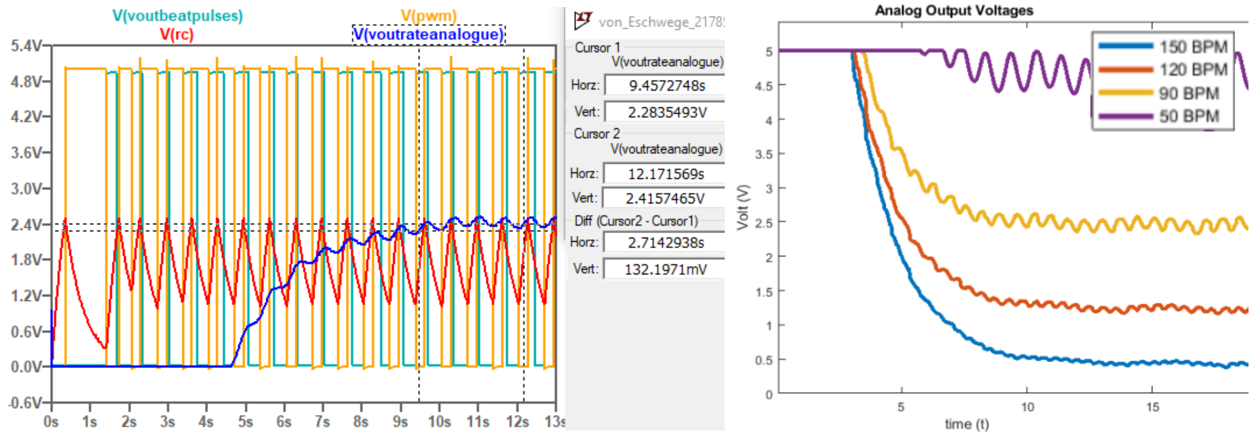
The output ranges from 0.026 to 4.95V, with a pulse duration in range 189 to 584 ms in figures 2.6a and 2.6b, meeting the 150 ms requirement. The design remains consistent for specification deviations of ±10% amplitude and a DC offset variation of ± 0.2 V, as the high-pass filter removes the DC component and the filter has a cutoff frequency low enough to greatly attenuate noise.



(a) 150 BPM **(b)** 50 BPM

Figure 2.6: Full input versus output range

For the analogue transducer, a PWM signal is obtained. Section 2.2 explains the interrelation, where `voutbeatpulses`, `pwm` and `rc` in figure 2.7a correspond to the clock signal, Q, and the charging capacitor voltage respectively. The 5% settling time is 9.45 s. The duty cycle is 73% at 60 BPM and decreases to 56% at 150 BPM. Finally, amplification and filtering of the PWM signal produces analogue output, range 4.2 V (2.7b).



(a) PWM and settling time

(b) Analogue voltage output

In figure 2.8, total current draw is measured through R_{sense}, averaging 13.0 mA, meeting the bonus requirement of 15 mA. When combined with Assignment 1, the combined current of 25.8 mA still is far less than the maximum current rating of 100 mA for the voltage regulator [1].

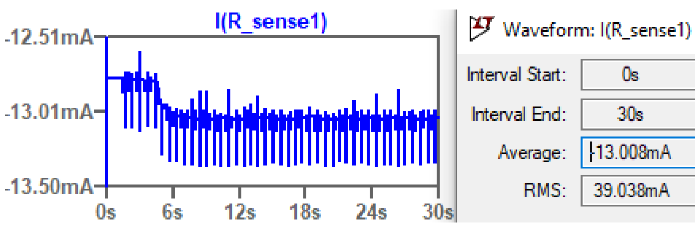


Figure 2.8: Total current usage

2.4. Summary

Concluding, it has been shown that the design performs as expected, meeting the all requirements. Output pulse width is broader than required, analogue output is generated with discrete, non-overlapping values exceeding the required range, while settling to a final value according to specification. Current draw remains sub-15 mA. Upon implementation of the microcontroller, be cognizant of the fact that analogue output values scale in a slightly non-linear fashion. This is not a problem, as it can be easily accounted for in software. By design, the system is limited to the 50 to 150 BPM range, but could be expanded using the existing design as a proof-of-concept.

Chapter 3

System and conclusion

3.1. System

The design of the heart-rate sensor goes to show that a noisy input signal could effectively be converted to a square wave output as well as an analogue output, enabling reliable interfacing of real-world measurements with digital systems. The heart-rate sensing circuit can now be integrated with the remainder of the health monitoring system by connecting the analogue output to the microcontroller input, to which the temperature sensor will also be connected - see E344 Assignment 1 [1]. The modular design of different parts of the health-monitoring system simplifies both design and debugging. Due to the second order low-pass filter, the heart-rate sensor is remarkably robust with respect to variations in noise, attenuating noise levels to below 6% of the signal amplitude before creating a pulse output. The design also performs well on a variety of the normalised input data sets. Calibration for microcontroller integration is done according to $f = -0.404V + 2.702$ (section 2.2, and the quantisation error is $f_{err} = \frac{2.5-0.8}{2^{10}}/(2.5-0.8) \times 100 = 0.1\%$. Noise is filtered to the extent that it has a negligible effect on the digitalized reading. The system utilizes less costly components where possible, is very power efficient, and adheres to all of the bonus requirements. The design objective has been met successfully.

3.2. Lessons learnt

1. LTSpice is a blessing, but can be an absolute nightmare with simulations. I've created a rough metric; it goes to show that 30% of my time was spent on circuit design, 10% on report writing, and 60% on debugging LTSpice simulation problems. I entered E-Design with the belief that simulation would greatly accelerate the pace of the subject, as soldering was eliminated, but I stand firm in my belief that I could have built a practical circuit much faster, as 'timestep too small' does not apply in the beauty of real-world continuity. However, maybe I just stand firm in this belief because I have not dealt with the problems arising from burnt-out components, messy soldering, melted PCB tracks and exploding diodes.
2. I believe to have greatly improved with regards to modularising and debugging not only circuits, but systems in general. Treating everything as a small problem to be solved furthered my conceptual understanding of component interaction.
3. I shouldn't have gone surfing for the entirety of the first week of the term. I said this in Assignment 1 as well, and lost marks for it, but it is still applicable, so I'll say it again.
4. If I could have it all over again, I wouldn't have texted my circuit; she did not reply.

Bibliography

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[7] *TL08xx: JFET-Input Operational Amplifiers*, TL081 Datasheet, Texas Instruments, May 2015.

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[10] *LT1018 Micropower Dual Comparator*, LT1018 Datasheet, Linear Technology.

Appendix A

Social contract

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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, Daniel Heinrich von Eschwege 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.

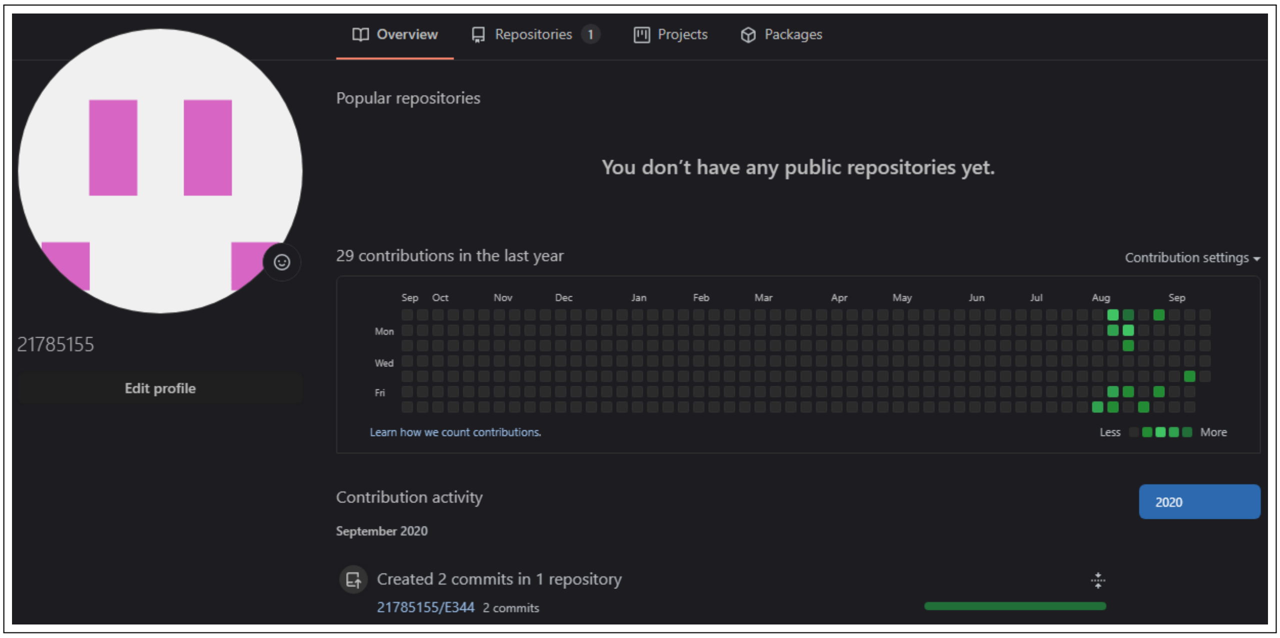
I acknowledge that E344 is an important part of my journey to becoming a professional engineer, and that my conduct should be reflective thereof. This includes doing and submitting my own work, working hard, starting on time, and assimilating as much information as possible. It also includes showing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

Signature: .... Date: 24/09/2020

1

Appendix B

GitHub Activity Heatmap



Appendix C

Stuff you want to include

R1 $= 500\,\mathrm{k}\Omega$. $\tau = RC$. The capacitor charge oscillates between V_L and V_H. $V_H = 2.5\,\mathrm{V}$. V_L is reached for the first time at t_{L_1} and V_H at t_{H_1}. V_L is then reached at t_{L_2}. For 150 BPM (or 2.5 Hz), the pulse drives high for 0.2 ms. Since the capacitor has to charge faster than 0.2 ms, a charge time of 0.16 ms was selected to add a 20% margin, accounting for noise. Thus $t_{H_1} - t_{L_1} = 0.16$ and $t_{L_2} - t_{H_1} = 0.4 - 0.16 = 0.24$ for the 150 BPM signal. Finally,

$$V_L = 5\left(1 - e^{\frac{t_{L_1}}{\tau}}\right)$$

$$V_H = 5\left(1 - e^{\frac{-t_{H_1}}{\tau}}\right)$$

$$V_L = V_H\left(e^{\frac{t_{L_1}}{\tau}}\right)$$

giving $C = 1\mu$, when solving using the following MatLab script:

```
syms t
syms tl1
syms th1
syms tl2
syms Vl
Vh = 2.5
R = 500000
A = Vl==5*(1-exp(-tl1/t))
B = Vh==5*(1-exp(-th1/t))
C = Vl==Vh*(exp(-(tl2-tl1)/t))
D = th1 - tl1 == 0.16
E = tl2 - th1 == 0.24
S = solve([A, B, C, D, E],[t, tl1, th1, tl2, Vl])
S.t
S.tl1
S.th1
S.tl2
Vl
C = R/t
```