E344 Assignment 1

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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.

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Nomenclature

Variables and functions

$max(A_1, A_2, \ldots, A_n)$	Returns the maximum value from the list of given numbers f_c
Corner Frequency	
G	Gain
τ	Time Constant of a changing signal (10% to 90% of final value)

Acronyms and abbreviations

op amp	Operational Amplifier
AC	Alternating Current
DC	Direct Current
NL	Not Listed
US	Ultrasonic Sensor
DAC	Digital to Analogue Converter
PWM	Pulse Width Modulation
ADC	Analogue to Digital Converter
RC	Resistor Capacitor
X_{pp}	peak to peak
X_{rms}	root mean square
NPN	Negative-Positive-Negative
BJT	Bipolar Junction Transistor
RC Car	Remote Contolled Car
RC Circuit	Resistor Capacitor circuit
DIP	Dual In-line Package
MOSFET	Metal-Oxide-Semiconductor Field-Effect Transistor
X_{batt}	Variable of the battery

Chapter 1

Literature survey

To fully design the individual parts of our RC Car system that make the whole, a full understanding of each part needs to be reached. These concepts are explored here to make decisions for the best outcome.

1.1. Operational amplifiers

Operational amplifiers: limitations and considerations

The operation amplifier that we will be using is the **MCP6242**. Thus, to ensure that the limitations of the given op amp are not exceeded.

From the datasheet [8], the notable limitations are listed in Table 1.1

Rating	Min	Max
$V_{DD} - V_{SS}$	-	7.0V
V_{DD}	1.8V	5.5V
$DifferenceInputVoltage$	-	$\|V_{DD} - V_{SS}\|$
V_{CMR}	$V_{SS} - 0.3V$	$V_{DD} + 0.3V$
V_{Output}	$V_{SS} + 35mV$	$V_{DD} - 35mV$
$I_{quiescent}$	$30\mu A$	$70\mu A$

Table 1.1: Notable limitations of the MCP6242 Operational Amplifier IC

Operational amplifier configurations

Firstly, op amps work on a couple of basic assumptions and assuming that they are ideal. The assumptions made are as follows:

1. No current flowing in and out of the input terminals of the op-amp (infinite input impedance of op-amp).

2. If the output is not in saturation, the voltage between the inverting and non-inverting input terminals is zero.

There exist many operational amplifier configurations, but for the purpose of amplifying signals, a configuration with a gain is needed.

Thus, the choice is between an inverting operational amplifier , a non-inverting operational amplifier and a differential amplifier.

Inverting	Non-inverting	Differential
$V_{out} = V_{in}(1 + \frac{R_2}{R_1})$	$V_{out} = -V_{in}\frac{R_2}{R_1}$	$V_{out} = \frac{R_2}{R_1}(V_{in2} - V_{in1})$

Table 1.2: Types of Operational Amplifiers and their respective equations

In Table 1.2, each op amp's output equations are listed.

The **non-inverting op amp** is powerful when the output signal needs to be positive and no negative input is expected, but has the disadvantage of always having a gain greater than one. This type of op amp only magnifies a source relative to ground.

The **inverting op amp** functions the same as the inverting op amp, but does not enforce the gain to be greater than one. It does however invert the input, meaning that if a positive output signal is required, the op amp's positive and negative terminals have to be switched. This type of op amp also only magnifies a source relative to ground.

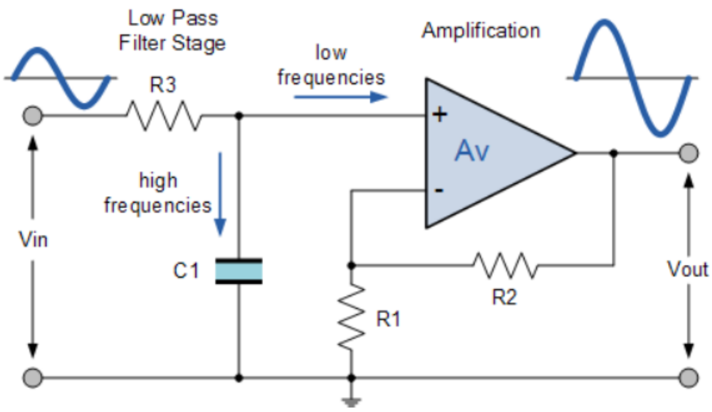


Figure 1.1: Non inverting ap amp circuit design (*Note.* Adapted from *Active Lowpass Filter*, by Electronics Tutorials, 2022) [1]

Finally, the **differential op amp** acts like the non-inverting op amp, but compares two inputs, instead of one. Thus, it does not magnify relative to ground, but instead the voltage at the negative terminal relative to the positive terminal.

1.2. Current sensing

The basic principle of current sensing uses Ohm's law

$$V = IR$$

to create a voltage that can be measured. There exists a couple of ways that a circuit can be designed to measure this voltage, ranging from **invasive** and **non-invasive** sensing, **high-side** and **low-side** sensing, and **AC** and **DC** sensing. Non-invasive sensing for AC signals is the process of measuring the current through a conductor by placing an inductor around that conductor. The magnetic field produced by the conductor then induces a current through the inductor and thus the current can be measured.

For DC signals, a hall sensor is used instead, with the same principal.

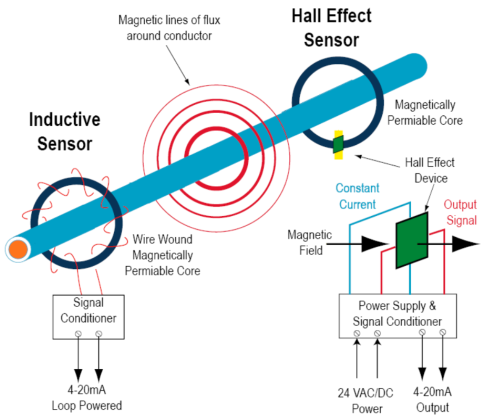


Figure 1.2: Current sensing with either a Hall sensor or an inductor (*Note.* Adapted from *Current Sensing Theory*, by NK Technologies, 2022) [2]

Invasive current sensing is the process of measuring the current through a conductor by placing an impedance in the circuit and measuring the voltage drop over it. This, in conjunction with Ohm's law can then be used to determine the current flowing through that impedance. [9]

High-side current sensing is the process of measuring the current through a load with a sensing resistor coupled in series, before the load impedance. The voltage over the load resistor would be inputted into an operational amplifier. Advantages of this method is that the load is grounded, and thus cannot be activated by accidental short at the power connection. A disadvantage of this approach is that the measured voltage could potentially be very high, and be out of the limit ranges of your operational amplifier.

Low-side current sensing is the exact same as high-side current sensing, but with the advantage of measuring low voltages, yet with the disadvantage of not having your load grounded. [10]

When measuring current invasively, measuring the current is as straightforward as described above. However, if measuring non-invasively, either a hall sensor or an inductive current sensor is used. With DC, a hall sensor is required, as there is no change in electric field to induce the inductor. With AC, an inductor can be used, as it will pick up the change in electric field. [2]

The power that a resistor dissipates, in the case that current is sensed with a invasive resistor, can be calculated as

$$P = I^2 * R$$

1.3. Interfacing with and using the ultrasonic range sensor

The model of US that will be implemented is the *HC-SR04* ultrasonic sensor.

This sensor has 4 pins, namely V_{cc} (5V), **Trigger Pulse Input**, **Echo Pulse Output** and **Ground**.

To get the HC-SR04 to output pulses, a short $10\mu s$ high pulse needs to be connected to the **Trigger Pulse Output** pin. This will cause the HC-SR04 to send out an 8 cycle burst of ultrasound at 40 kHz and raise its echo to h. Once the 40kHz signal is received, **Echo Pulse Output** goes low again. The datasheet [11] recommends repeating this cycle every 60ms (16.67Hz). This ensures that an echo signal doesn't return after another cycle has started and a false distance is reported.

A realistic timing diagram can be seen in Figure 1.3.

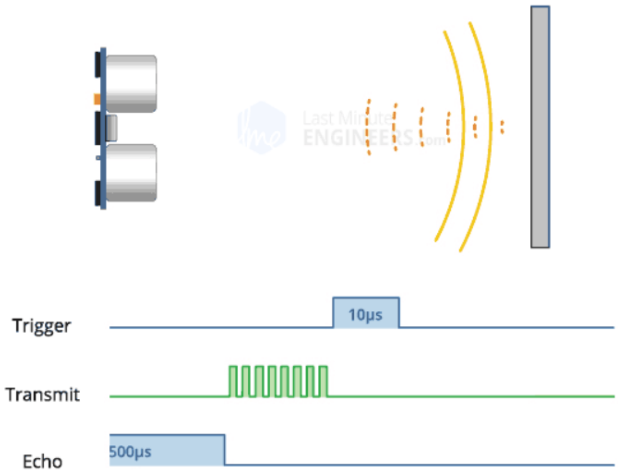


Figure 1.3: A timing diagram of the sensor working, with an obstacle in its way. *Note*: Adapted from *How HC-SR04 Ultrasonic Sensor Works & Interface It With Arduino* by lastminuteengineers.com, 2022 [3]

It is also important to know that the HC-SR04 will automatically switch **Echo Pulse Output** to low after 38ms if no echo is received back. This timeout is implemented as not to get the sensor stuck waiting for a signal that is never going to return. [3]

The PWM duty cycle thus directly relates to the distance that an object is away from the sensor, using the equation:

$$2d = v * t$$

Where d represents the distance the echo has traveled, v the speed of the media that the sound waves are propagating through, and t, the time elapsed since the pulse was sent out (time that **Echo Pulse Output** is high). The wave must travel double the distance (to the object and back), so this must be accomodated for in out equation.

1.4. Converting PWM signals to analogue equivalents

Converting PWM signals to analogue can be as simple as implementing an RC filter and setting the corner frequency to the frequency of your PWM signal. Depending on the way your PWM signal is constructed, and the way in which you interpret that result to your analogue signal, you can design for either a low-pass or high-pass filter.

The duty cycle of the PWM signal directly correlates to an analogue voltage level. The higher the duty cycle, the higher the voltage out, and vice versa.

The corner frequency needs to be set lower than the frequency of the PWM signal. If it is selected to be higher than the PWM frequency, then the harmonics of the system will start to appear. This is illustrated below in Figure 1.4.

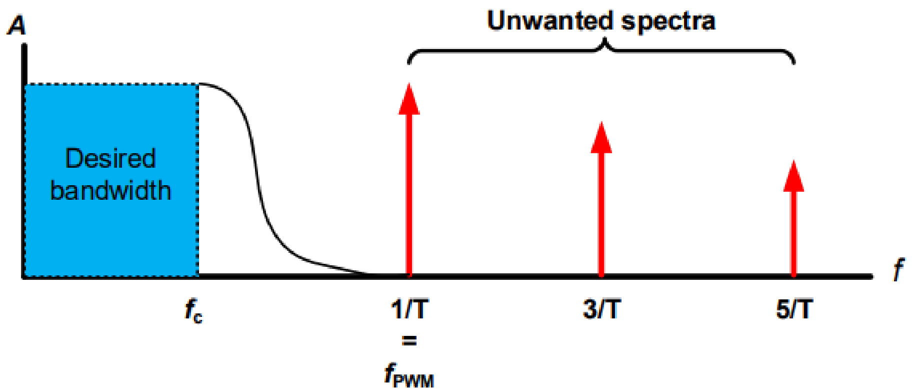


Figure 1.4: The desired bandwidth and potential harmonic positions. *Note:* Adapted from *Using PWM to Generate an Analog Output* by Microchip Technology Inc, 2020 [4]

”When the PWM frequency is close to the cut-off frequency, the filter responds quickly, but produces a high amount of ripple in the output signal. As the distance between the cut-off frequency and PWM frequency increases, the response time decreases, but the ripple in the output signal also decreases.” [4]

1.5. Ultrasonic range finder principles of operation

The ultrasonic sensor uses the principal of soundwaves propagating through a medium to determine the distance between itself and a reference.

A high frequency (above human hearing) sound pulse will be sent out the US will wait for a returning sound wave (after bouncing back from an object), at the same frequency. The sensor then uses the relationship between the speed of a sound wave in a respective medium to determine the distance, using the formula

$$d = v * t$$

where **d** represents the total distance traveled, **v** the velocity of the media in which the sound wave is propagating, as well as **t** being the time difference between the time the initial wave was generated and the time the wave was received. Thus, the distance between the US and the object measured is $d/2$. [12]

Sound travels faster in denser media, and thus the temperature, especially in gasses, plays a role. In Figure 1.5, this can be seen quite well. Note: this is only important for interpreting the signal, not generating it.

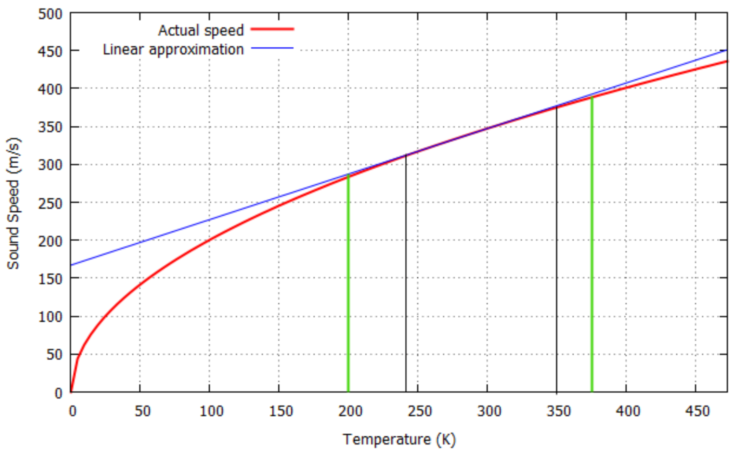


Figure 1.5: The speed of sound in air at varying temperature. *Note*: Adapted from *Speed of Sound*, by Wikipedia, 2022 [**?**].

Once a trigger pulse (from a micro-controller or any sort of switch on circuit) is detected, the US sends out a few pulses, listens for them, and will indicate the time waited for a return signal on one of its pulses, wither with a PWM signal, or by sending the time difference via a data transmission line. PMW is the easiest, most cost effective and least complicated course to follow.

1.6. Converting digital values to analogue equivalents

For conversion of a digital logic signal to analogue, a summing amplifier can be used with varying resistances (in scale) to produce the expected output of a digital to analogue conversion circuit. An example of such a circuit can be seen in Figure [5]

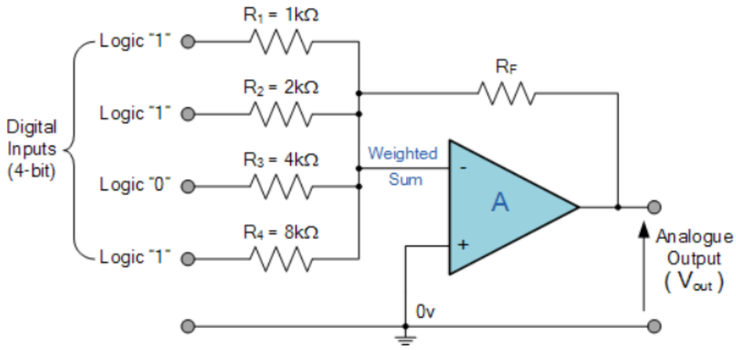


Figure 1.6: An example of a digital to analogue conversion circuit. *Note*: Adapted from *The Summing Amplfier*, by Electronics Tutorials, 2022 [5]

This configuration inverts, but other configurations work to include gain as well as being non-inverting. Inverting can be an issue if the op amp's V_{SS} pin is connected to ground, and thus the non-inverting configuration is very helpful.

One thing to take into consideration when designing these type of op amp circuits, is what the output and input impedances are. The effect of designing with too low input resistances (R_1 through to R_4), is that the current draw from the voltage sources can be very high. If something like a micro-controller is used, having a high current draw can be detrimental. Therefor, relatively large resistances should be used.

The output resistance should also be taken into consideration, as a low output resistance would increase the current draw from the voltage source (V_{DD} of the op amp). Therefor, when considering the circuit that follows the DAC, the total resistance that is "seen" by the op amp is very important, as to keep the current draw of the complete circuit low.

1.7. Lead-acid battery voltages and currents

For charging lead-acid batteries, there are multiple ways to do so. These methods are **Constant Voltage**, **Constant Current**, **Taper Current** and **Two-step Constant Voltage** [6].

In this section, constant-voltage-constant-current will be discussed.

At a fully charged state, the battery had a lower than nominal voltage. Thus, connecting it to a high voltage (or charge voltage), will allow current to flow (due to the voltage difference). This method of charging requires the current into the battery to remain constant, as well as the charging voltage.

Once the battery reaches the desired voltage, the voltage difference is much lower and thus current will struggle to flow. As such, the current can no longer remain constant and thus needs to decrease in a logarithmic fashion.

Once the circuit detects that a very low amount of current is flowing, the battery is fully charged.

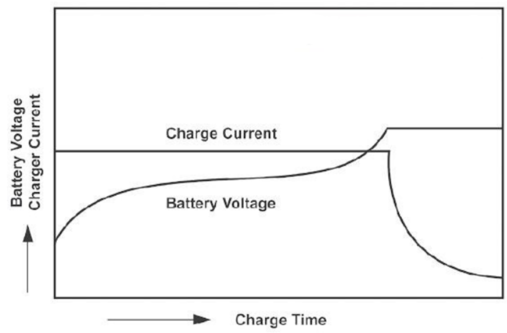


Figure 1.7: Constant-Voltage-Constant-Current Voltage and Current lines during charging (*Note*: Adapted from *How to charge a lead acid battery*, by Power Sonic, 2022 [6]).

The voltage and current values can be seen in Figure 1.7.

Care should be taken when designing for a charging voltage, as overcharging the battery can lead to lowered service life of the battery itself, and undercharging can lead to not using the full capacity of the battery.

It is important to note that each cell inside of the battery is also $2.2V$ nominal [6].

Batteries can also be charged constantly, or be charged cyclically. Depending on the use case, different charging methods should be used to ensure that the battery does not get damaged. Constant-voltage-constant-current is applicable to charge for both cases.

Chapter 2

Detail design

2.1. Current sensor

The current sensor needs to conform to the following specifications set out in [13]:

1. The current sensor needs to operate with a 5V rail and ground
2. The sense resistor must be on the low-side of the DC motor
3. The maximum voltage over the DC motor and the sense resistor must be 6V
4. The circuit will give an output range that is less than 0.1 V for 0 A and greater than 3 V for maximum current.
5. The circuit shall respond to a step change from minimum to maximum current to reach 90% of the output voltage in less than 100 ms.
6. The circuit shall filter the signal such that a 10 mV signal at 1 kHz or higher results in less than 250 mV on the output.
7. Circuit shall use less than 150 μA

From testing, the maximum current through the motor at 6V would be 700mA when the motor is stalled, 160mA with no load and just under 300mA when running close to stall. There will be designed for 300mA and limiting the output to 3.3V after the amplifier as per the ESP32's max voltage input [14]. Thus, to get the maximum voltage over R_{sense}:

$$V = I * R = max(0.3, 0.16) * 0.01 = 3mV$$

It is also necessary to calculate the maximum power consumed by R_{sense} to ensure that a resistor with a high enough power rating is implemented. Thus:

$$P = I^2 * R = 1.5^2 * 0.01 = 22.5mW$$

Thus, to calculate gain, using 5.0V as maximum:

$$Gain = \frac{5.0V}{0.003V} = 1666$$

Referring to [1] to then design for R_2, choose R_1 as 400Ω:

$$Gain = 1 + \frac{R_2}{R_1} \text{ therefore: } (Gain - 1) * R_1 = R_2 = 680k\Omega$$

These resistors will be tested in LTSpice to ensure that the current draw limitation is adhered to.
Finally, a voltage divider will be implemented after the amplifier to limit the output to 3.3V, using a $25k\Omega$ and $120k\Omega$ resistor.
The limitations of the op amp will be taken into consideration as they are rated in Table 1.1. Due to noise on the current sense line, a RC filter will also be implemented to negate this. The requirement set out in specification 5 and 6 needs to be taken into account here.

To calculate the τ needed, the initial transfer function was drawn, and the maximum gain in dB was calculated to be 50.39. To reach a gain of 25 (requirement 6), a dB gain of 27.95 is needed. In a broad scope, the process that was followed was plotting a -20dB line to cross through the 250mV threshold at 1kHz, as well as a straight line to represent the gain in the transfer function. The intersection of these two points will give the corner frequency needed to meet requirement 6. This process can graphically be followed in Figure 2.1.

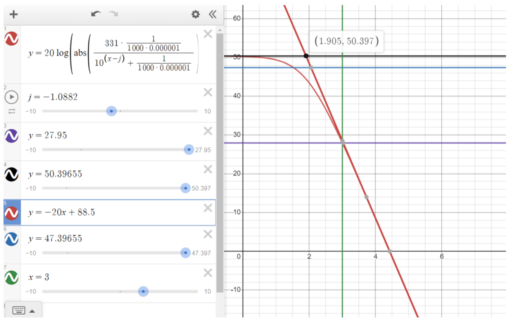


Figure 2.1: Gain Analysis of RC filter

The final corner frequency comes out to

$$f_{cutoff} = 10^{1.905} = 80.35Hz \text{ or from [1]: } Gain = \frac{A_f}{\sqrt{1+(\frac{f_c}{f})^2}}, \text{ thus } f_c = 75\text{Hz}$$

Taking 80.35Hz as f_c, and choosing the capacitor value as 47nF, solving for resistance:

$$R = \frac{1}{2\pi * 47 \cdot 10^{-9} * 80} = 42320\Omega \text{ therefore, } \tau = R * C = 198\mu s$$

The full, built circuit as well as the final values of the resistors and capacitor can be seen in Figure 2.2

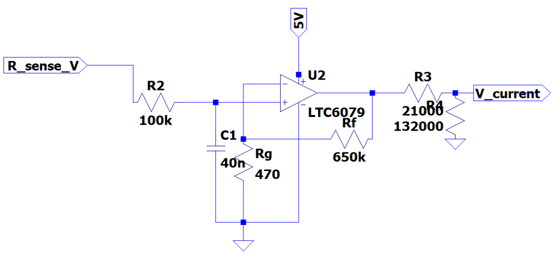


Figure 2.2: Circuit design of RC Filter and Non-inverting Operational Amplifier setup

2.2. Voltage regulator

The voltage regulators that are available are the **LD1117V**, **LM317T** and the **AP7384-50V-A** voltage regulators [7]. The specifications relevant to the system main power voltage regulator are listed below in Table 2.1. The specifications listed take into consideration that $V_{out} = 5V$ and $T_{amb} = 25°C$.

Regulator Name	V_{in-min}	V_{in-max}	P_{tot}	V_{drop}	$I_{load-max}$
AP7384 [15]	3.3V	40.0V	NL[1]	500mV @ V_{out} = 3.3V	50mA
LM317T [16]	NL[1]	NL[1]	20W	2.5V	2.2A
LD1117 [17]	-	15.0V	12W	1.1V - 1.2V @ I_O=800mA	1.3A

Note 1: Data not listed on datasheet

Table 2.1: Voltage regulator options and relevant ratings

Due to the battery possible having a maximum value of 7.2V and a minimum value of 6V, we need to take the dropout voltage, V_{in-max}, V_{in-min} and the maximum load current into consideration. Due to the low load current of the **AP7384**, the other two voltage regulators are viable options. Because of the high drop off voltage of the **LM317T**, this will not accommodate the battery being at low voltages very well. Thus, the clear option, and the regulator that will be implemented, is the **LD1117** voltage regulator.

The main voltage regulator will need to power multiple segments of the full project, including the ESP32, and most of the circuits designed in Chapter 2. A diagram of this can be seen in Figure 2.3.

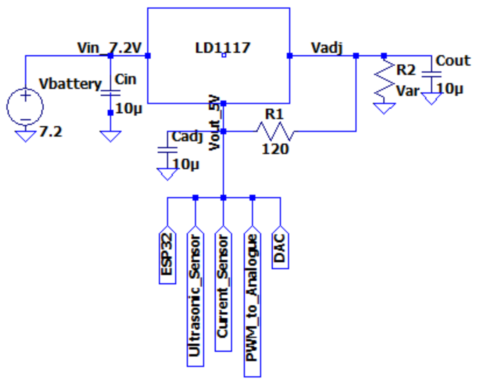


Figure 2.3: Diagram of the voltage regulator and segments that it would supply 5V to

The voltage regulator will be implemented with three $10\mu F$ capacitors as per the data sheet [17]. This also specifies that for R_1, a value of a couple $k\Omega$ is applicable, but a variable resistor will be implemented to allow for adjustments..

2.3. Ultrasonic sensor PMW to analogue signal

The design details that need to be adhered to can be found in [18]. These limitations are:

1. Filter PWM-type output from distance of 1m down to 5cm, such that dist $\geq$ 1m gives output $>$ 3V, and distance $\leq$ 5cm gives output $<$ 0.3V.
2. Response time must be such that if PWM changes (step change) from 5cm to 1m, the output must reach 90% of final within 1.5s. Note: a step change in PWM is not a step change in voltage.
3. The noise on the analogue signal must be less than 70 mVpk$-$pk.
4. The whole circuit must use less than 750μA

First, a 3nd order RC filter will be implemented because a single order wont be sufficient enough to get a clean analogue output from the PWM signal. A relatively high τ will need to be selected to ensure that the sharp edges of the PWM signal converted into slew which accurately represents an analogue voltage. Therefor, a $\tau = 0.047s$ will be selected. Picking $R = 1k\Omega$, and solving for C:

$$C = \frac{\tau}{R} = \frac{0.047}{1000} = 47\mu F$$

R_1 will be implemented in the form of a potentiometer for fine tuning.

To ensure that most noise is attenuated in the end signal, the second and third RC filter will have a slightly larger τ, namely $0.0705s$ $(R = 1.5k\Omega)$ and $0.094s$ $(R = 2k\Omega)$. For all three the RC filters, a capacitor of $47\mu F$ will be used.

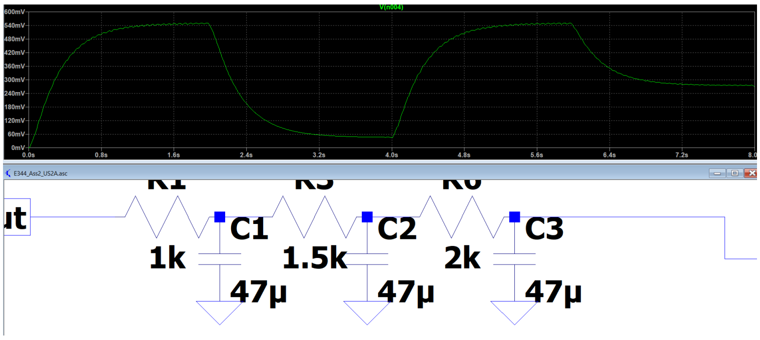


Figure 2.4: Built and simulated 3rd rder RC filter

In Figure 2.4, a built 3rd Order RC filter can be seen. The input is a simulated expected input from the US.

This signal needs to be amplified to feed into the ESP32, and thus at close proximity, the signal output needs to be $>$ 3V (see specification 1). Thus, a gain will need to be implemented, using a inverting op amp, discussed in Section 1.1.

$$Gain = 1 + \frac{R_2}{R_1} \; Gain = \frac{3.3}{0.545} = 6 \; R_2 = (Gain - 1) * R_1 = 5 * R_1$$

Selecting $R_1 = 20k\Omega$, and solving for R_2:

$$R_2 = 5 * R_1 = 100k\Omega$$

R_1 will be implemented by a potentiometer to adjust the gain, in the case that the attenuated PWM signal delivers lower or higher voltages than expected.

To calculate the current draw for the op amp, the maximum voltage out needs to be known. It has been designed to output at max 3.3V. With a $100k\Omega$ pull down resistor on the ESP32, we can fully calculate for I_{out}. Thus, I_{out} is:

$$I_{out} = \frac{V_{out-max}}{R_{total}} = \frac{3.3}{(R_1+R_2)||R_{pull-down}} = \frac{3.3}{54.54k\Omega} = 60\mu A$$

Then the total current draw will be:

$$I_{circuit} = I_{quiescent} + I_{out} = 50\mu A + 60\mu A = 110\mu A$$

which satisfies specification 4.

It is important to also note that the US works with 5V and draws 15mA of current, although this will not be included as part of the current draw in our calculation.

The fully implemented PMW to analogue converting circuit can be seen in Figure 2.5.

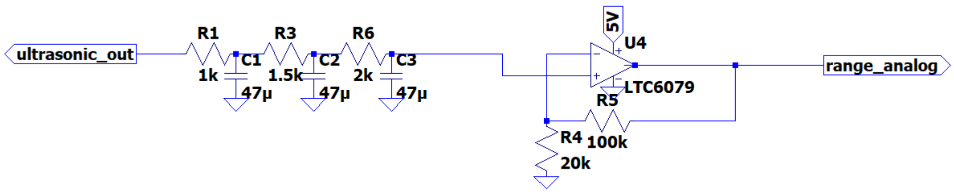


Figure 2.5: Fully designed PWM to analgue signal converter

2.4. Digital to analogue conversion circuit

The requirements set out in [19] are stated below:

1. The digital inputs are 3.3 V TTL signals.
2. The output is to be inverted - a low digital number results in a high analogue value, and vice versa. **b3** is the **msb** in the LTSpice simulations.
3. The analogue output ranges from 0 V to 3.3 V.
4. The digital 0000 results in an analogue signal greater than 3 V
5. The digital 1111 results in a signal analogue signal smaller than 0.5 V.

Thus, to implement this circuit, refering to Table 1.1, and [5] was used as reference. For the relative voltages, B3 was to be the MSB. Therefor, $R = 10k\Omega$, and thus

Bit Number	Resistor Name	Resistance (Ω)
B3	R_1	$10k$
B2	R_2	$20k$
B1	R_3	$40k$
B0	R_4	$80k$

Table 2.2: Summing amp resistance values

Instead of using a summing amplifier, a normal resistor ladder is used. This ensures the same usability as a summing amplifier, but without an amplifier. As this section of the circuit doesn't need to be amplified, a gain of 1 will be used. Thus $R_1 = R_2 = 100k\Omega$.

Finally, to invert the values (due to **0000** being 3.3V output and **1111** being 0V output), a differential amplifier will be implemented. V_2 will be used as a reference voltage. Using the 5V voltage rail as reference, this needs to be voltage divided into 3.3V. Thus, $R_2 = 17k\Omega$ and $R_4 = 33k\Omega$. As R_1 and R_3 determine the gain, these will be selected such that $R_1 = R_3 = 100k\Omega$ for unity gain as the previous op amp already factors in the required gain.

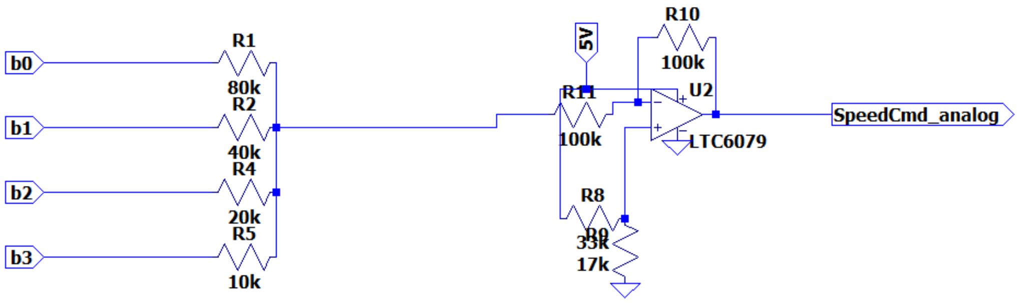


Figure 2.6: Fully designed DAC circuit

2.5. Motor Control circuit

This design section we will be implementing the circuits designed in Section 2.1, 2.3 and 2.4 to control the voltage that is being given to the motor.

The requirements that need to be met are given in [20]:

1. Except for the DIP switch that uses 3.3 V (if used), the whole circuit must be powered from a DC power supply, set to 7.2 V.
2. The voltage applied over the motor shall range from a 0 V to to at least 6.2 V when a power supply of 7.2 V is used.
3. The voltage over the DC motor must be low when an object is near.
4. If an object is far from the proximity sensor, the motor voltage must be ’low’ when the torque command is zero (0000 binary), and ’high’ when the torque command is high (1111 in binary).
5. A ‘low’ voltage over the motor shall be below 0.5 V.
6. A ‘high’ voltage over the motor shall satify $V_{out} > (V_{batt} - 1)$, where V_{batt} will be represented by a DC power supply, which can be as high as 7.2 V.
7. The current used from 5 V regulator shall be less than 500 μA
8. Current used from battery (7.2 V supply) shall be less than 1.5 A.
9. Provision shall be made for motor back-emf.

2.5.1. Differential Amplifier

Thus, for the voltage control, a differential op amp will be used with a gain to ensure that the output voltage is 6.2V maximum (when the US detects far away and the output of the DAC is **1111**).

The TLC2722 op amp will be used in this implementation.

Thus we can start design. Following [21], we first pick $R_1 = R_3$ and $R_2 = R_4$. Since the output is then equal to:

$$V_{out} = \frac{R_3}{R_1}(V_{US} - V_{DAC}) \text{ where } G = \frac{R_3}{R_1} = \frac{6.2V}{2.5V} = 2.48$$

Note: 2.5V is used as the maximum voltage difference due to imperfect design on the DAC

Selecting R_1 as $100k\Omega$, R_3 becomes $248k\Omega$. The current draw on this section of the circuit will then be (assuming a high input impedance - $100k\Omega$ - on the next section of the circuit):

$$I_{max} = I_{quiescent} + I_{output} = 50\mu A + \frac{6.2V}{100k\Omega} = 2462\mu A$$

Note: From [22], the $I_{quiescent}$ can be read as 2.4mA

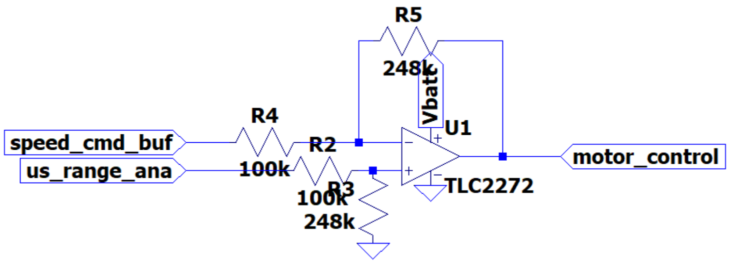


Figure 2.7: The designed motor control circuit using a differential amplifier

2.5.2. Emitter Follower

Due to the low current output of TLC2272 used in the design of Section 2.5.1, a TIP31C NPN BJT will be used. This will be set up in an emitter follower configuration, and the current supplied by the TLC2272 will be amplified to feed the motor with at least 200mA. The collector side of the transistor will be directly connected to the battery, the base will be connected to the output of the differential amplifier set up in Section 2.5.1, and the emitter will be connected to the motor.

A flyback diode will be implemented (reversed polarity with regards to the motor) to combat run off current when the motor is switched off.

The built and designed circuit can be seen in Figure 2.8.

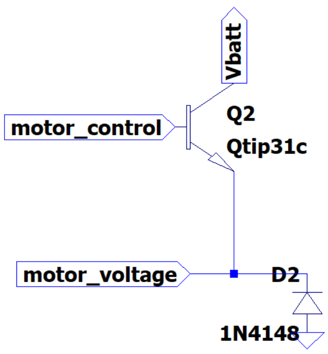


Figure 2.8: Designed emitter follower circuit

2.6. Left wheel low-side switch and current sensor

For this section a MOSFET will be driving the left wheel as a switch, being switched by the ESP32. The current drawn from the motor also be measured.

Note: The current sensing circuit will be exactly the same as the current sensor in Section 2.1.

For the selection of the MOSFET, the options and important details are listed in Table 2.3:

Reference	Name	Max Current	Max Voltage	Max Gate Threshold Voltage
[23]	FQD13N06L	11A	60V	2.5V
[24]	2N27000	400mA DC	60V	3V
[25]	IRF530	17A	100V	4V
[26]	IRF9Z24N	11A	60V	4V

Table 2.3: Available MOSFETs as specified in [7]

For the MOSFET switching circuit, a few things need to be kept in mind when designing. Firstly, the motor that will be driven needs a max current of about 800mA at stall, otherwise the voltage will drop, so the selected transistor needs to have a maximum current of $> 1,5A$, per design selection. Further, the MOSFET will be driven by the ESP32, so a maximum gate threshold voltage lower than $3.3V$ is required. From Table 2.3, the only MOSFET that matches these specifications is the **FQD13N06L**. Thus this is the one that will be implemented.

The ESP32 will be connected to the gate of the MOSFET, and the voltage directly after the motor will be connected to the drain. Finally, a current sensor will be implemented. The design for this has already been done in section 2.1. Some of the resistors will be implemented with variable resistors to negate tolerance differences from the resistors and op amp.

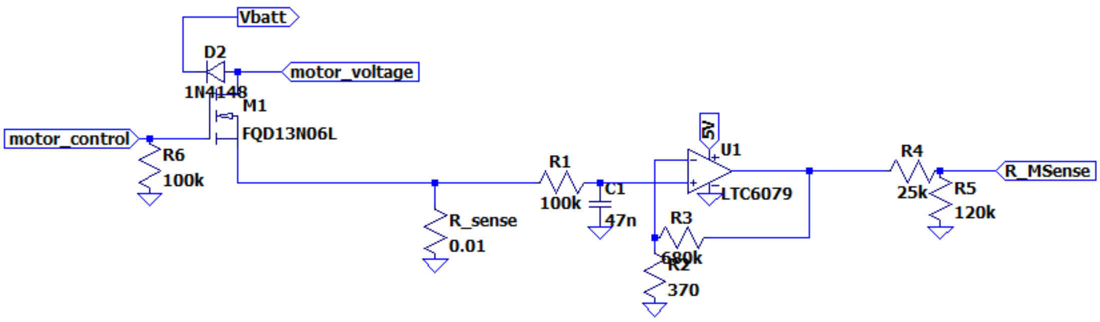


Figure 2.9: The full circuit with MOSFET and current sensor (from 2.1

A diode will also be implemented between the output of the motor and the battery, so that runaway current from the motor has somewhere to go when suddenly being disconnected from ground.

2.7. Range sensor

Most of the design for this section is software related, but something to keep in mind is that the ultrasonic sensor outputs a 5V echo signal, and thus this will need to be voltage divided to ensure that the 3.3V input pin of the ESP32 does not get damaged. Choosing $R_2 = 33k$:

$$R_1 = \frac{V_{in} * R_2}{V_in} - R2 = \frac{33k\Omega * 5V}{3.3V} - 33k\Omega = 17k\Omega$$

Figure 2.10 shows the implementation of this circuit.

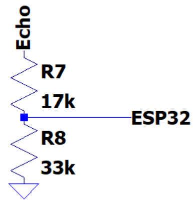


Figure 2.10: Ultrasonic Sensor 5V voltage divider circuit

Further, on the software side, the US will need to be triggered once every 62.4ms, and then the duty cycle must be calculated to determine what distance the US is detecting. From this distance, a PWM signal will be generated, enabling current to flow through the motor.

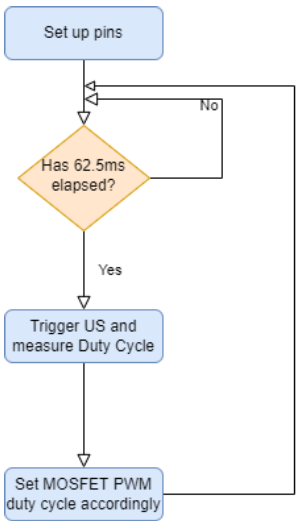


Figure 2.11: Flowchart detailing the process the software follows to set motor PWM duty cycle

2.8. Battery charger, under-voltage protection, battery voltage signal conditioning

2.8.1. Battery Charger

As our battery is a 4000mAh capacity, the charging current will need to be 400mA, and from the datasheet needs to be charged to a voltage of 7.3V [27]. Thus, we will model the battery as a constant load. At 400mA, the Schottky Diode has a forward voltage drop of $\pm 350mV$.

$$R_{batt} = \frac{7.3V+0.35}{400mA} = 19.125\Omega$$

From Table 2.3, the gate threshold voltage of the **IRF9Z24N** is 4V. Due to the resistor connecting the drain and the gate of this transistor, the gate voltage will be read and the transistor will behave like a switch. The total current coming from the regulator will then be

$$I_o = \frac{V_o}{R_{batt}} + \frac{V_o}{R_{enable-drain}} = 400mA + \frac{7.65V}{1000} = 0.4765A$$

Because of the $V_o - V_{adj} = 1.25V$, and knowing the output pin should be 7.65V, we know that $V_{adj} = 7.65V - 1.25V = 6.4V$. From [28], $I_{adj} = 100\mu A$. To reach 6.4V, the total current to ground, when selecting $R_4 = 500\Omega$, must be $12.8mA$. Thus, $I_{o-to-adj} = 12.8mA - 0.1mA = 12.7mA$. Given that

$$R_{o-to-adj} = \frac{V_{o-to-adj}}{I_{o-to-adj}} = \frac{1.25V}{12.7mA} = 98\Omega$$

$R_{o_to_adj}$ will be rounded to 100Ω. Finally, A small sense resistor of 0.4Ω will be added directly after the voltage regulator for current setting. A pull down resistor will also be added to the charging on transistor, with a value of 250Ω. The final circuit can be seen in Figure 2.12.

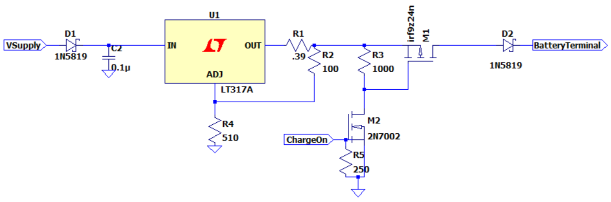


Figure 2.12: The fully designed battery charge circuit

2.8.2. Under-Voltage Protection

For the under-voltage protections, a Schmitt Trigger will be implemented using an op amp.The requirements of this section are to switch off all non-essential circuits once the voltage of the battery reaches 6.0V, and only to switch them on again once the voltage reaches 6.1V.

Firstly, the 5V regulated line will be used as a reference, as this will still be active once the battery. Since we expect $V_{batt} = 6V\ 6.1V$, the 5V line will be voltage divided to 3.025V. This will be implemented with a $47k\Omega$ variable resistor and a $15k\Omega$ resistor.

The unregulated V_{batt} will be voltage divided in harf to 3.025V as well.

Finally, the sensitivity of the forward feedback will need to be very sensitive, as the switch will need to function withing a 0.1V range. Thus $82k\Omega$ and 820Ω will be used to create a 100:1 ratio.

The fully implemented Schmitt Trigger can be seen in Figure 2.13.

The **MCP6242** op amp will be used to implement this circuit so the limitations in Table 1.1 need to be adhered to.

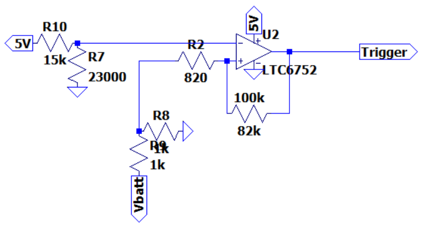


Figure 2.13: Fully designed Schmitt Trigger

This confirms that the limitations in Table 1.1 have been adhered to.

One thing to note about the design of this section is that, due to the lowresistor values used in the V_{batt} voltage dividing section, the current draw will be quite high. These values were played around with in SPICE and the conclusion was that they can not be any higher or distortion of the output signal will occur.

2.8.3. Battery Voltage Signal Conditioning

For this section, a battery voltage from 5V to 7.2V needs to be converted to give a 0.1V-3.3V output signal, depending on the voltage level of the battery.

For this implementation, a differential amplifier will be used.

From Section 1.1, and setting $R_1 = 10k\Omega$:

$$V_o = \frac{R_3}{R_1}(V_2 - V_1) \text{ thus}$$
$$R_3 = \frac{\Delta V_o * R_1}{V_2 - V_1} = \frac{3.3V * 10k\Omega}{7.2V - 5V} = 15000\Omega$$

The **MCP6242** op amp will be used to implement this circuit so the limitations in Table 1.1 need to be adhered to.

The fully designed circuit can be seen in Figure 2.14.

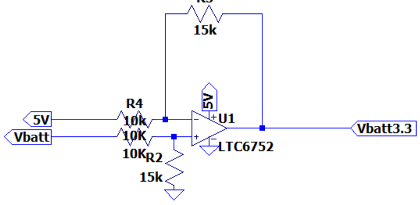


Figure 2.14: Fully designed battery voltage signal circuit

This confirms that the limitations in Table 1.1 have been adhered to.

Chapter 3

Results

Simulations and measurements will be used to ensure that the designed circuits from Chapter 2 satisfy design specifications and meet project requirements.

3.1. Built voltage regulator

This section of the full project does not have hard specifications, but from the requirements of the other circuits it is clear that the input can be as high as 7.2V, and as low as 6V. The output of this circuit also has to be a constant 5V, with as little as possible noise.

With a DC power supply supplying 7.2V to the voltage regulator circuit, the output is as can be seen in Figure 3.1.

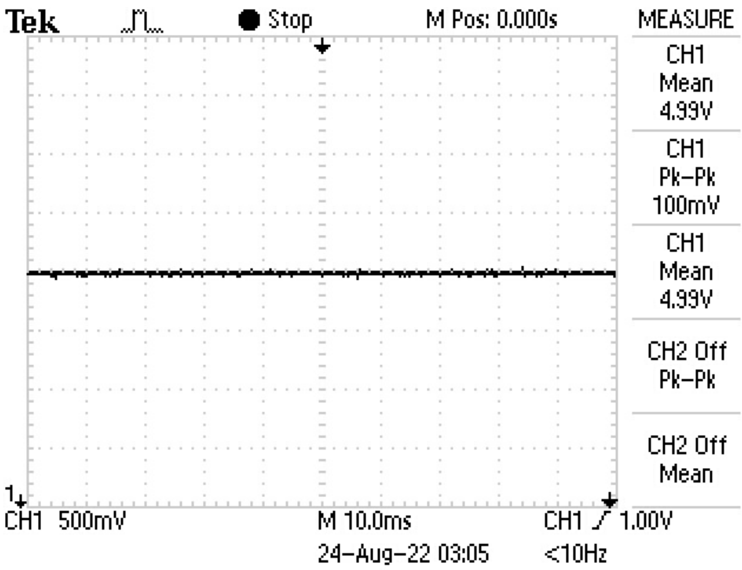


Figure 3.1: The output of the voltage regulator circuit with 7.2V input

From 3.1, it can be seen that there is little to no noise on the line and thus this circuit is highly applicable for powering the project as a whole.

3.2. Simulated current sensor

The simulation circuit was built in LTSpice as can be seen in Figure 2.2. The requirements that the circuit needs to follow are assessed later in this section.

The output, input and current draw of the circuit can be seen in Figure 3.2

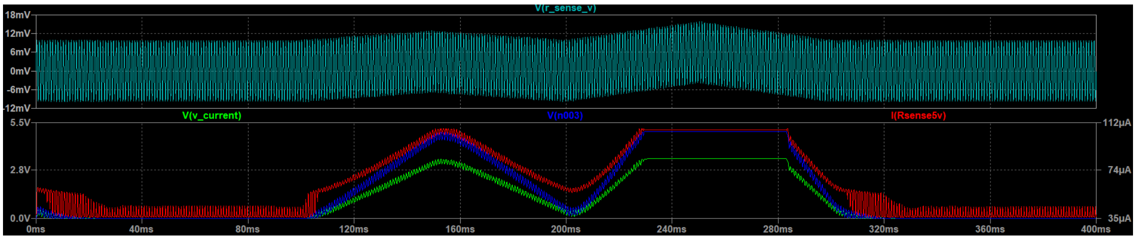


Figure 3.2: Input (red), output (blue) and current draw (red) of the current sensor built circuit in 2.2

For this circuit to meet the gain requirements, the voltage over R_{sense} needs to be amplified to higher than 3V. From the simulation results in Figure 2.1, it is clear that the circuit amplifies the voltage over R_{sense} to 3.3V, which satisfies the requirement. The maximum current applied over R_{sense} is 800mA (from testing) and the slopes in the simulation is under semi-load conditions.

For this circuit to meet the noise attenuation requirement, noise signals of 10mV need to be smaller than 250mV. From the figure in 2.1, it is clear that the noise present never exceeds 250mV amplitude, and reaches a maximum of 165mV amplitude, which satisfies the requirement.

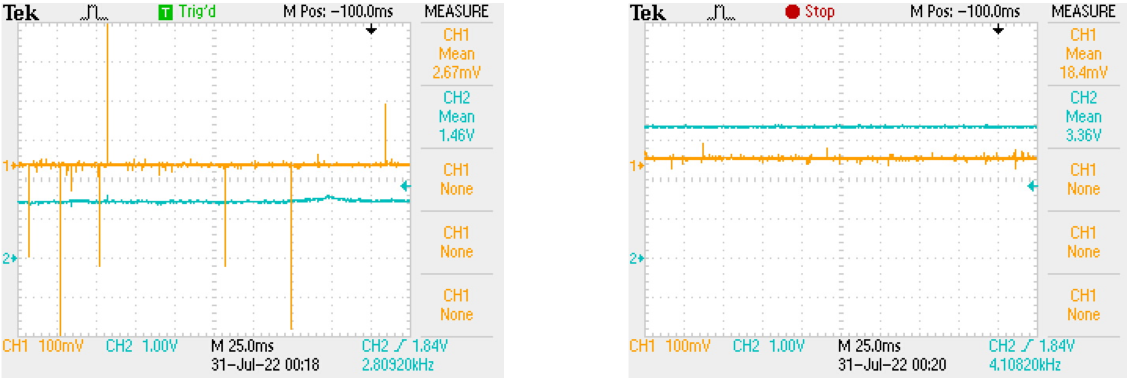
For this circuit to meet the rise time requirements, a change in 90% of maximum voltage level should be should be reflected in less than 100ms. From the simulation results in Figure 2.1, it is clear that the rise time is $105ms - 101ms = 4ms$, which satisfies the requirement.

For this circuit to meet the current draw requirement, a maximum current draw of 150μA must be adhered to. From the simulation results in Figure 2.1, it is clear that a maximum current draw of 105μ is reached, which satisfies this requirement.

Therefore, this circuit meets all the requirements specified in section 2.1 and will thus be implemented.

3.3. Built current sensor

The current sensor circuit was built and can be seen in Figure 4.1. Small adjustments were made to ensure it functions as closely as possible to the simulated circuit. These changes are discussed in Section 4.1. In Figure 3.3a, the circuit's behaviour can be seen when there is



(a) Circuit running with no resistance on motor

(b) Circuit running with motor stalled

Figure 3.3: The built current sensor running (a) with no resistance and (b) at stall. Channel 1 (orange) is the voltage over the current sensing resistor and Channel 2 (blue) is the voltage output of the circuit.

no load on the motor. The designed gain is reached as expected and the voltage divider as well. From the power bench, it can be seen that the motor draws 110mA when running in this state. Thus,

$$V_{expected} = \frac{110mA}{300mA} * 3.3V = 1.21V$$

Thus, the circuit is not perfect. This could be due to tolerances in the resistors, the op amp, the current sensing resistor, and also due to the added resistance on the input line. If it becomes clear that these issues will hinder performance, they will be reviewed.

The circuit running with the motor stalled, as can be seen in Figure 3.3b, shows that the maximum output of the circuit is 3.36V, which is an acceptable voltage for the ESP32 to receive, and thus meets the requirements.

In either case (Figure 3.3a and Figure 3.3b, the noise on the input line can be clearly seen. However, in the output, there is no noise, and thus this requirement for the circuit is satisfied as well.

3.4. Simulated ultrasonic sensor circuit

The simulation circuit was built in LTSpice as can be seen in Figure 2.5. The requirements (set out in Section 2.3) that the circuit needs to follow are assessed later in this section. The output, input and current draw of the circuit can be seen in Figure 3.4.

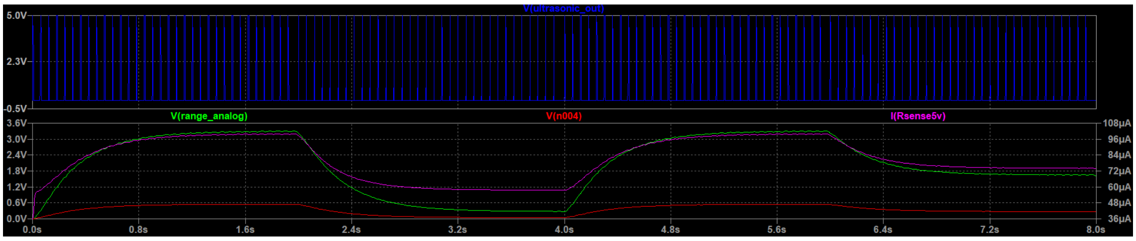


Figure 3.4: Input (blue), output after RC filter (red), output after amplifier (green) and current draw (pink) of the ultrasonic sensor PWM to Analogue voltage conversion circuit

For this circuit to meet the output requirements, it needs to have a maximum output of 3.3V when an object is in close proximity, and an output of < 0.3V when an object is far away. From the simulated results, it can be seen that this holds true, and thus this specification is met.

For the circuit to meet the time requirements, a step change needs to happen in less than 1.5s. From the simulated results, it can be seen that a change of 90% happens within 0.75s, which satisfies the requirements.

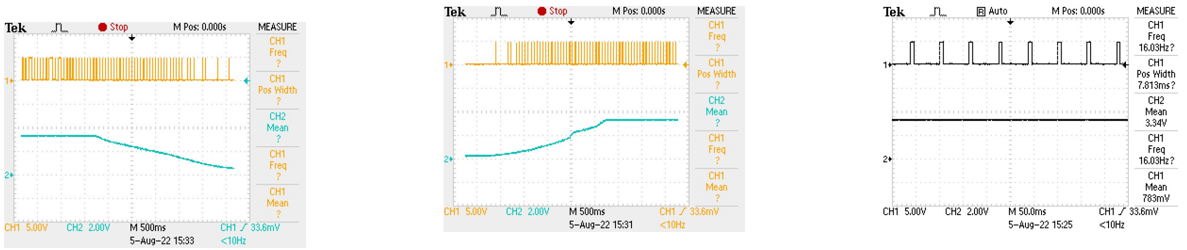
The noise on the output line must be lower than $70mV_{pp}$ to meet the design requirements. Since the output's noise can be measured at $40mV_{pp}$, this design requirement is also met.

Finally, from the simulations, the current draw of the whole circuit is reported at $100\mu A$ maximum, which matches the hand calculated current draw quite well and also satisfies the requirement of using less than $750\mu A$.

In conclusion, this circuit meets all the design requirements and can thus be built and implemented.

3.5. Built ultrasonic sensor circuit

The ultrasonic sensor circuit was built and can be seen in Figure 4.3. Two variable resistors were implemented instead of the $160k\Omega$ (to change the gain (which correlates to maximum measure distance) and the $11k\Omega$ resistors (to ensure that the maximum input voltage of the ESP32 was not exceeded).



(a) PMW Echo from US (top) and analogue of circuit (bottom) whilst moving closer to sensor

(b) PMW Echo from US (top) and analogue output of circuit (bottom) whilst moving away from sensor

(c) PMW Echo from US (top) and analogue output of circuit (bottom) at maximum distance from the sensor

Figure 3.5: Output response from the built circuit when moving closer (a) and further (b) from the US

(a) PMW Echo from US (top) and analogue output of circuit (bottom) whilst close range

(b) PMW Echo from US (top) and analogue output of circuit (bottom) with big step changes (close to far proximity quickly)

Figure 3.6: Output response from the built circuit showing step change as well as close range measurements.

From Figure 3.6a it can be seen that the circuit has a noise level of $60V_{pp}$, and at < 1cm, the output is 262mV. Both of these measurements meet the requirements. The rise and fall time (demonstrated in Figure 3.6b), can be seen as $\pm 750ms$. It can also be seen in Figure 3.5c that the maximum voltage the circuit reaches is $3.34V$. Thus all the requirements for this circuit are met.

3.6. Simulated digital to analogue conversion circuit

The simulation circuit was built in LTSpice as can be seen in Figure 2.6. The requirements (set out in Section 2.4) that the circuit needs to follow are assessed later in this section. The output, input and current draw of the circuit can be seen in Figure 3.7.

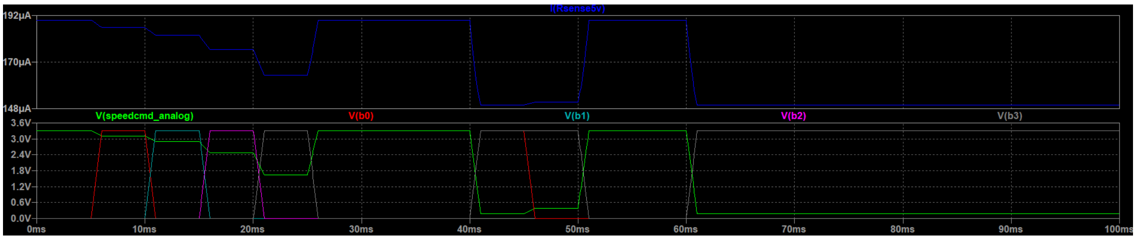
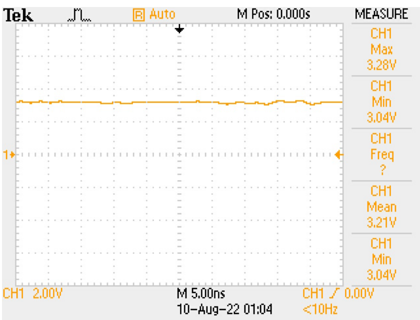


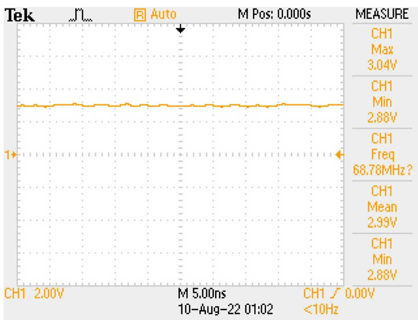
Figure 3.7: The inputs (B0 in red, B1 in turquoise, B2 in pink and B3 in grey), output (light green) and current draw (blue) of the simulated digital to analogue conversion circuit.

3.7. Built digital to analogue conversion circuit

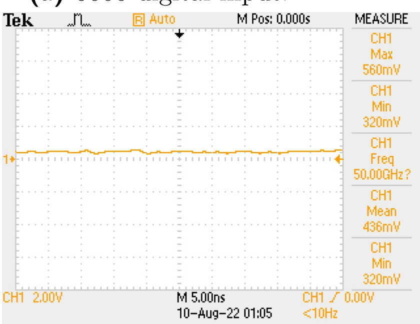
The physical implementation of this circuit will be built as can be seen in Figure 2.6. This real world implementation will use variable resistors for R_1 through to R_4, so that output voltages can be tweaked if need be.



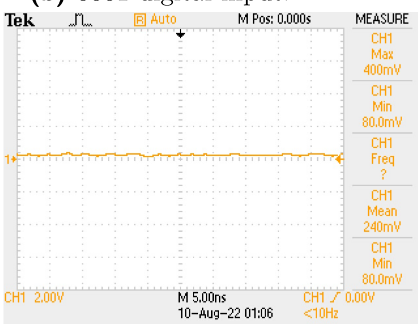
(a) 0000 digital input.



(b) 0001 digital input.



(c) 1110 digital input.



(d) 1111 digital input.

Figure 3.8: The output of the DAC at values **0000, 0001, 1110, 1111**.

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3.8. Simulated motor control circuit

The simulated circuit needs to meet the requirements set out in Section 2.5

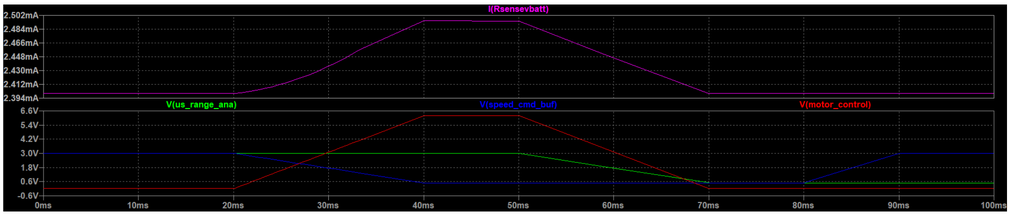


Figure 3.9: The differential amplifier section of the full circuit

From Figure 3.9 it is clear that the output will be above 6V when the DAC is outputting 0.5V and the analogue sensor output is 3V. Furthermore, all of the specifications relevant to this section set out in Section 2.5 are met.

The current draw from this section of the circuit is quite high, at around $2.5mA$. This is due to the high current draw of the TLC2272 op amp and thus can not be lowered.

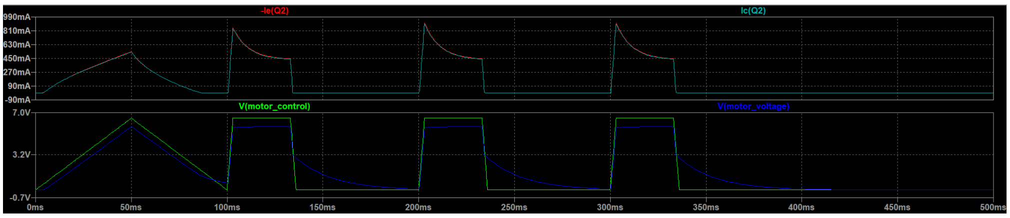
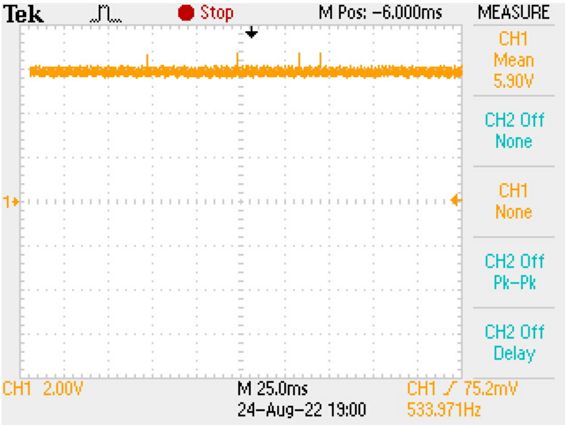


Figure 3.10: The emitter follower of the full circuit

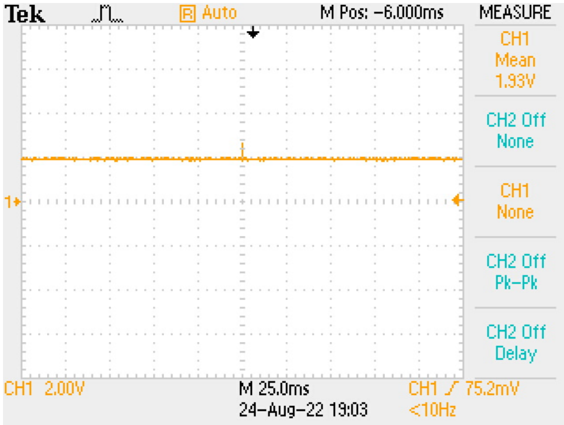
From Figure 3.10 it can be seen that the motor is receiving the necessary current to function at full power and the current gain from the TIP31C is behaving as expected.

Thus, these two circuits circuits in series with one another will meet all of the specifications and can thus be implemented.

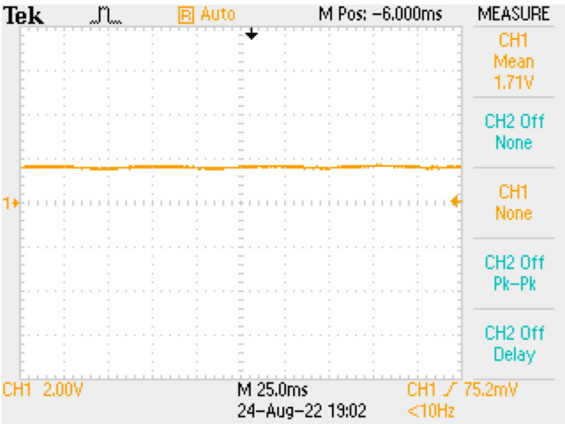
3.9. Built motor control circuit



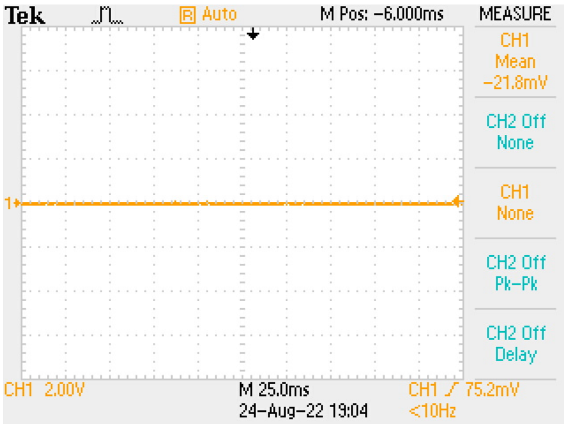
(a) Motor control output with DAC set to **1111** and **no object** in front of US



(b) Motor control output with DAC set to **1111** and **object** in front of US ($\pm 25cm$)



(c) Motor control output with DAC set to **1100** and **no object** in front of US



(d) Motor control output with DAC set to **1100** and **object** in front of US ($\pm 25cm$)

Figure 3.11: The four main cases that the output of the motor control can handle

From Figure 3.11, the results can be seen quite clearly. These measurements are also neatly tabled in Table 3.1. From these measurements, it is clear that this circuit is behaving as expected and meets all of the requirements as set out in Section 2.5.

Figure	DAC Input	Object Proximity	Output voltage
3.11a	1111	Far	5.9V
3.11b	1111	Near	1.99V
3.11c	1100	Far	1.71V
3.11d	1100	Near	0.0V

Table 3.1: Tabled results from Figure 3.11

3.10. Simulated PWM low-side switch and current sensor

The circuit detailed in Section 2.5 was built.

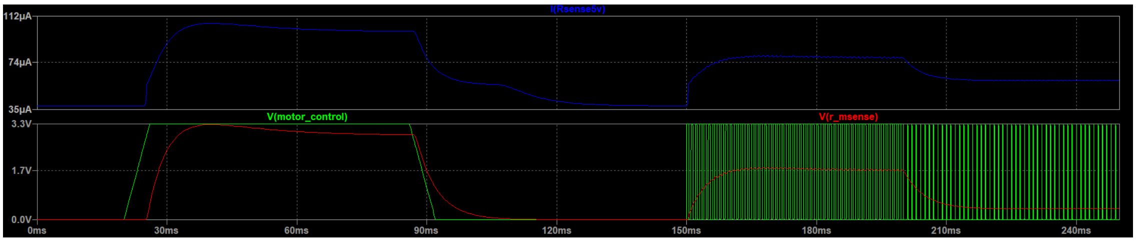
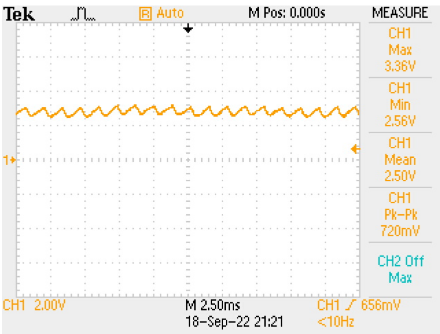


Figure 3.12: Simulated left wheel low-side control and current sense circuit Input in green, output in red and current draw in blue).

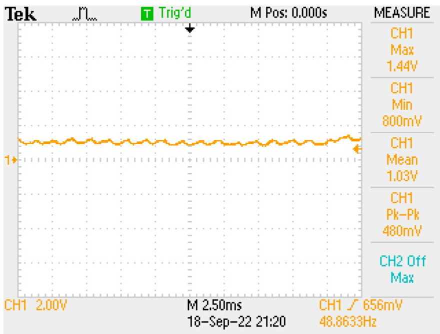
From 3.12 it is clear that this circuit meets the expected requirements. The maximum output from the current sensor does not exceed 3.3V, which is ideal since this is the maximum voltage that the ESP32 can handle. Further, the response time is quite fast, and the motor follows well when a high frequency PWM signal is applied to the MOSFET. Thus, this circuit is ready to be implemented.

The circuit also draws a maximum of $110\mu A$, which is ideal.

3.11. Built PWM low-side switch and current sensor



(a) Left wheel current sensor output under **load**



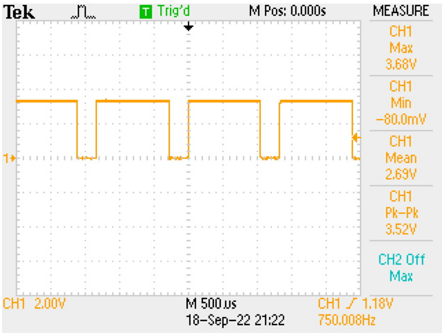
(b) Current sensor circuit output under **no load**

Figure 3.13: Left wheel current sensor outputs

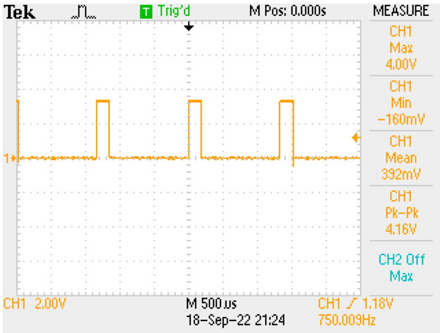
From Figure 3.13b and 3.13a it can be seen that the current sensor circuit outputs expected values. It also has low noise. From **3.13a** the transient noise is 480mV, which under 500mV and thus satisfied the requirements set out in Section 2.1.

From Figures 3.14a to 3.14d it can be seen that the PWM duty cycle matches the expected outputs.

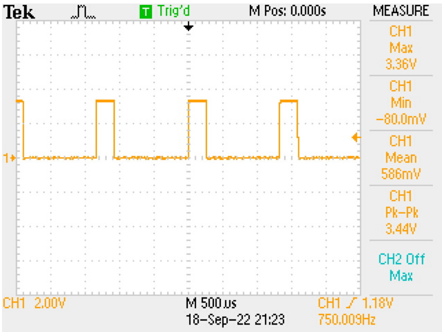
Thus, this section of the project can be controlled via proximity as well as via the DIP (which will later be implemented to be the ESP controlled via bluetooth). Thus, the requirements have been met and this section can be integrated into the rest of the build.



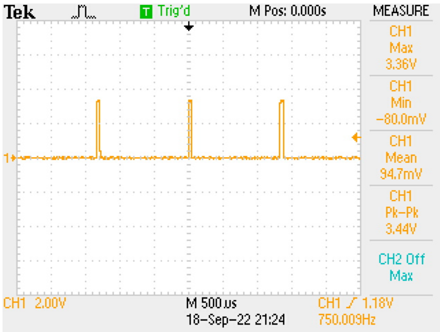
(a) Motor control PWM output with DAC set to **1111** and **no object** in front of US



(b) Motor control PWM output with DAC set to **1111** and **object** in front of US ($\pm 10cm$)



(c) Motor control PWM output with DAC set to **1100** and **no object** in front of US



(d) Motor control PWM output with DAC set to **1100** and **object** in front of US ($\pm 10cm$)

Figure 3.14: The four main cases that the output of the motor control can handle

3.12. Simulated battery charger, under-voltage protection, battery voltage signal conditioning

3.12.1. Simulated battery charger

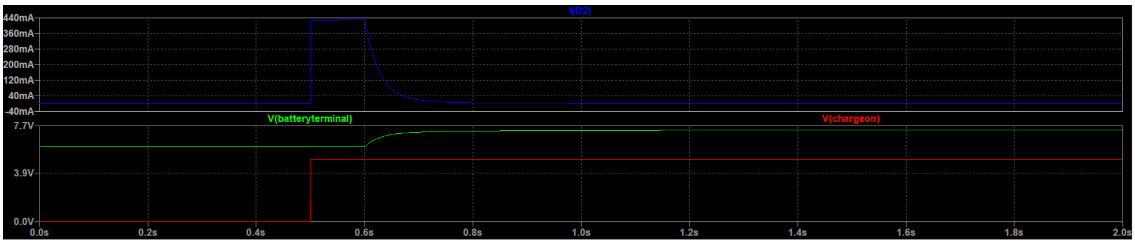


Figure 3.15: The input (red), output (green) and current flow into the battery (blue) of the battery charging circuit

As can be seen in Figure 3.15, the circuit outputs a constant 430mA into the battery, until the simulated battery reaches a charged state. Once this point is reached, the typical natural decay graph can be seen, as is expected, which decays to 0mA.

The charger was designed to output a voltage of between 7.3V and 7.4V, and as can be seen from the figure, reaches a voltage of 7.36V after a bit of charging, once the enable output has been switched on.

Thus this circuit satisfies expectations and can be implemented physically.

3.12.2. Simulated under-voltage protection

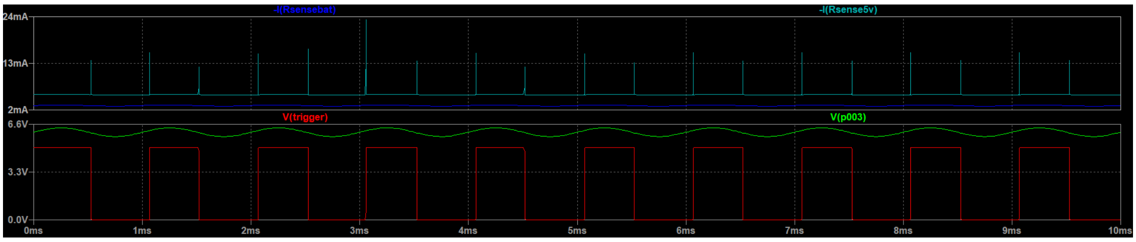


Figure 3.16: Input (green), output (red), 5V current draw (cyan) and V_{batt} current draw (blue) of the Schmitt Trigger

As can be seen in Figure 3.16, the output is as we expect and can be used to trigger a MOSFET to drive the main circuits. As mentioned in Section 2.8.2, the current draw of both V_{batt} and the 5V line is quite high. In future versions of this project, this is definitely a section of the total system that can be improved upon.

3.12.3. Simulated battery voltage signal conditioning

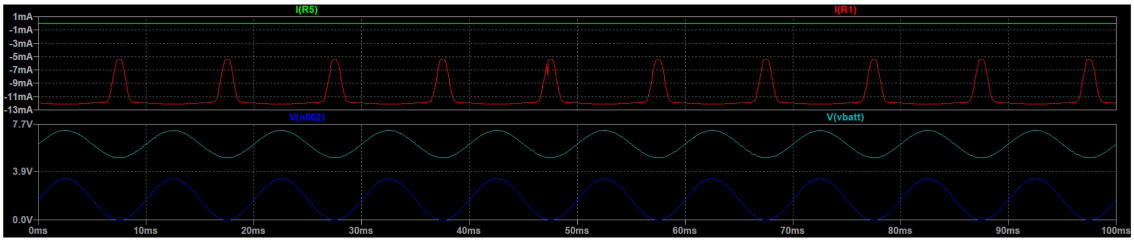


Figure 3.17: Input (cyan), output (blue), 5V current draw (red) and V_{batt} current draw (green) of the battery voltage signal conditioning circuit

As can be seen from Figure 3.17, as the input voltage varies, the output follows closely, never exceeding 3.3V. This simulated output proves that this circuit is ready to be implemented.

One thing to note is that the current draw from the 5V line is quite high. The cause of this can not be found, but is speculated to be the current drawn from the op amp in this configuration.

Future implementations of this circuit should aim to decrease this current draw.

Chapter 4

Physical Implementation

4.1. Built current sensor

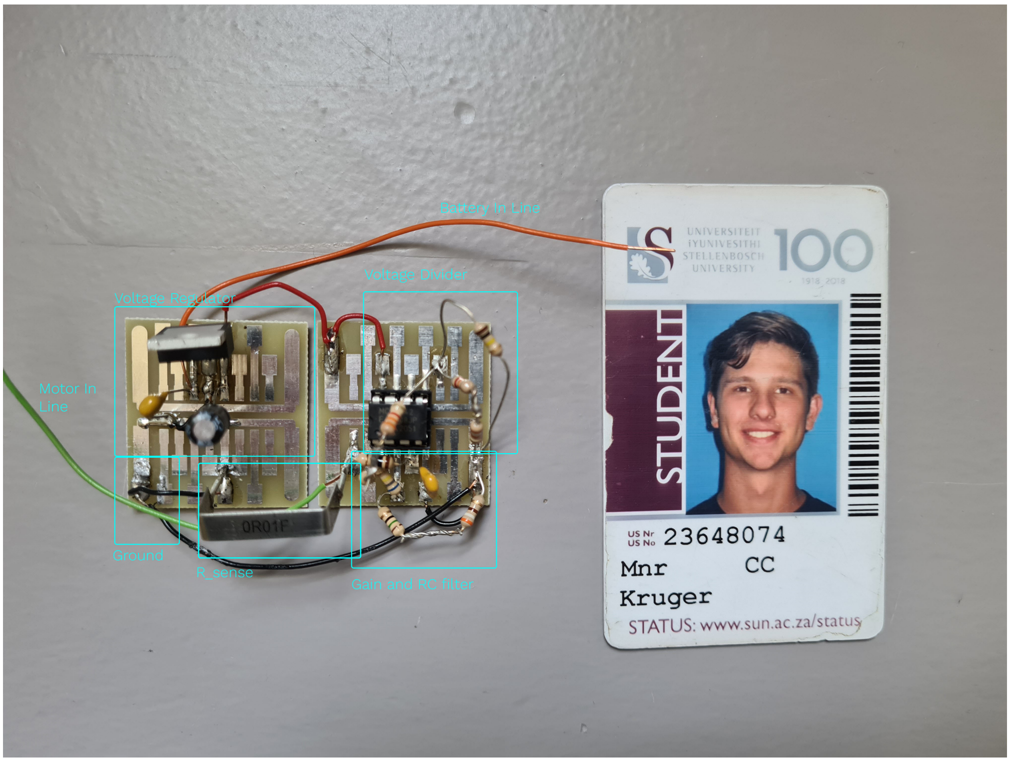


Figure 4.1: Built Current Sensing and Voltage Regulator Circuit

Above, in Figure 4.1, the built circuit can be seen. The circuit was tested and some design adaptations were made to ensure that it meets all the requirements. To ensure that the circuit functions the same as the simulated current sense circuit, a $100k\Omega$ resistor was added in parallel with the $120k\Omega$ resistor. This is to mimic a pull down resistor from the ESP32.

The op amp was also installed on a pad instead of being directly soldered onto the circuit board in case it gets burnt out. This makes it easy to replace.

The op amp is powered from the voltage regulator provided - due to the chosen voltage regulator not being available at the time. In testing, the *Motor In Line* was used to connect the motor to the sensing circuit, however, due to added resistance on the line, this connection will have to be shortened to avoid interference with the circuit in the future. The circuit was connected to a test bench with 7.2V via the *Battery In Line*.

4.2. Built voltage regulator

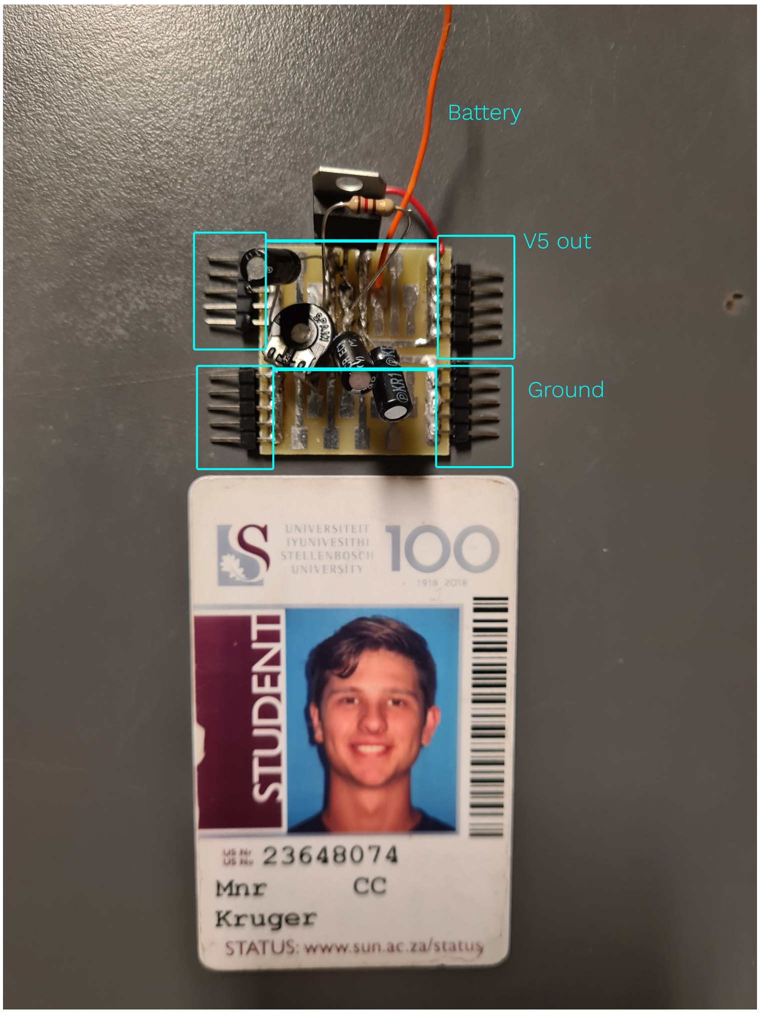


Figure 4.2: The built voltage regulator circuit

The built voltage regulator circuit can be seen in Figure 4.2. It has been implemented with a variable resistor to allow for tuning of the 5V output, and an additional noise capacitor was added to negate even more noise.

From Section 2.2 it is clear that this circuit behaves as expected and the project can be developed further.

4.3. Built ultrasonic sensor conditioning circuit

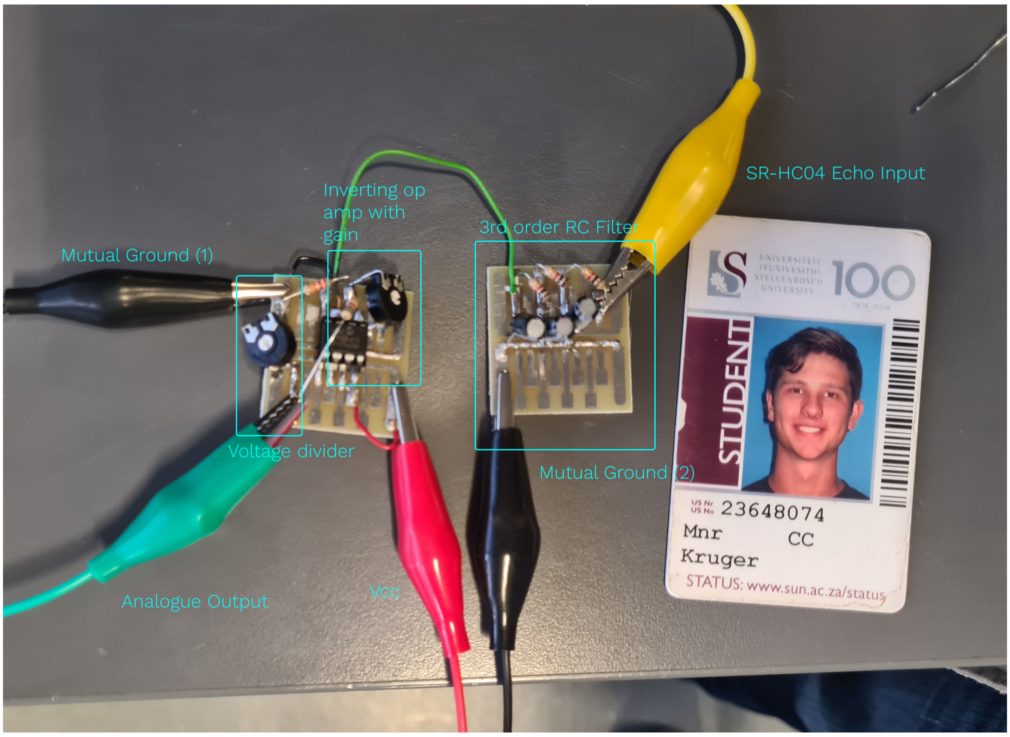


Figure 4.3: Built Current Sensing and Voltage Regulator Circuit

Above, in Figure 4.3, the built PWM to analogue conversion signal can be seen. The circuit was tested and a few adaptations were made to ensure correct working as well as calibration methods being put in place to fine tune in the case that the circuit is measuring slightly below or over specification. The addition of the variable resistor in the voltage divider gives the option of accounting for the $100k\Omega$ pull down resistor in place on the ESP32. The full circuit is powered from the voltage regulator selected in Section 2.2. The full circuit is at this stage connected to the voltage regulator via crocodile clips to ensure that future implementations to the project don't overcrowd the soldered boards. Finally, the two boards are grounded to each other as well as to the ESP32, the voltage supply powering the voltage regulator, and the grounds thereof.

4.4. Built digital to analogue conversion circuit

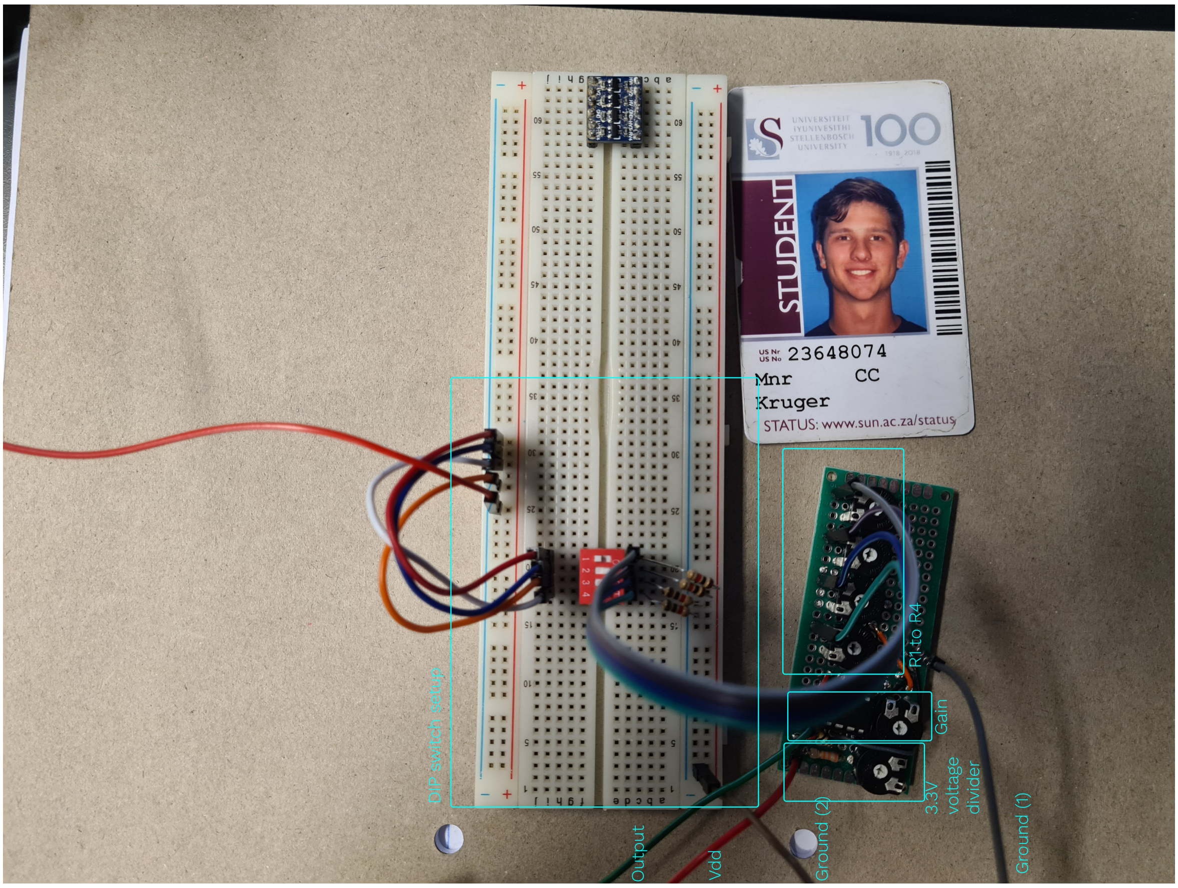


Figure 4.4: The built and tuned DAC circuit

The built DAC circuit can be seen in Figure 4.4.

Due to multiple variable resistors being in the circuit, thy all have to be individually tuned to ensure that the output at any given point is what is expected.

The DIP switch is implemented with a $1k\Omega$ pull-down resistor. This ensures that when the switch is flipped, the circuit receives a 0V input instead of a floating pin. If the pin was to be floating, the output would not at all be what is expected. The pull-down resistors and DIP switch will be replaced by the ESP32 at a later stage in the development of this project.

4.5. Built motor control circuit

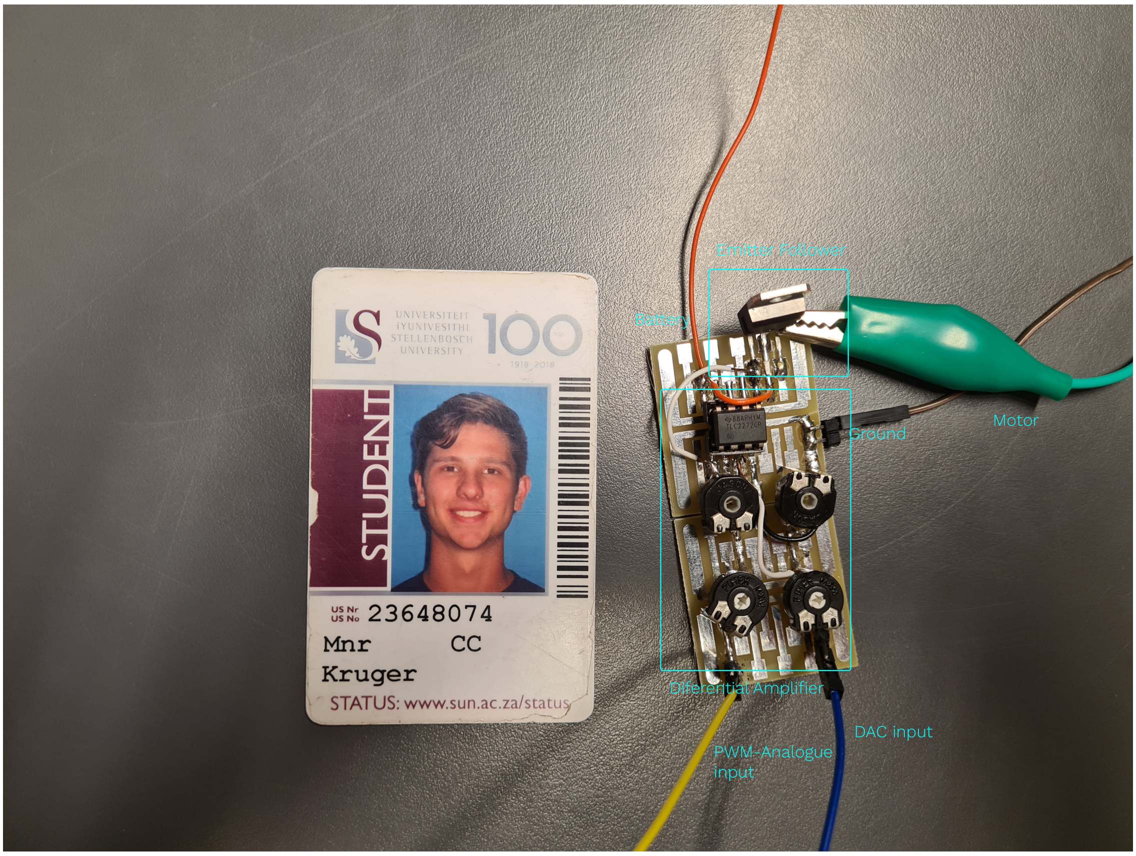


Figure 4.5: The built motor control circuit

The built motor control circuit can be seen in Figure 4.5. Four variable resistors are implemented to ensure that the output can be adjusted if need be. This was quite difficult to work with as precise values were needed.

The circuit was tested via initially supplying 3.3V to both pins until the output was as low as possible. Then the DAC simulated output was set to 0V to ensure that the respective variable resistors was implemented correctly. Final adjustments were made to fine tune the output.

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Appendix A

Social contract

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E-design 344 Social Contract

2022

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 a few paid helpers (Rita van der Walt, Keegan Hull, and Michael Ritchie) spent countless hours to prepare for E344 to ensure that you get your money's worth, that you are enabled to learn from the module, and demonstrate and be assessed on your skills. We commit to prepare the assignments, to set the 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.

I, Christo Kruger .... 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 online of supplementary videos on specific topics, I acknowledge that I am expected to attend the scheduled lectures 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.

Prof. MJ (Thinus) Booysen

Signature: MJ Booysen Digitally signed by MJ Booysen Date: 2022.07.02 13:22:09 +02'00'

Date: 01 July 2022

Student number: 23648074

Signature: ....

Date: 19 July 2022

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Appendix B

GitHub Activity Heatmap