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Department of Electrical Engineering

Examensarbete

Design and evaluation of a capacitively coupled sensor readout circuit, toward contact-less ECG and EEG

Examensarbete utfört i elektroniska komponenter
vid Tekniska högskolan i Linköping
av

Daniel Svärd

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INSTITUTE OF TECHNOLOGY

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Abstract

In modern medicine, the measurement of electrophysiological signals play a key role in health monitoring and diagnostics. Electrical activity originating from our nerve and muscle cells conveys real-time information about our current health state. The two most common and actively used techniques for measuring such signals are electrocardiography (ECG) and electroencephalography (EEG).

These signals are very weak, reaching from a few millivolts down to tens of microvolts in amplitude, and have the majority of the power located at very low frequencies, from below 1 Hz up to 40 Hz. These characteristics sets very tough requirements on the electrical circuit designs used to measure them. Usually, measurement is performed by attaching electrodes with direct contact to the skin using an adhesive, conductive gel to fixate them. This method requires a clinical environment and is time consuming, tedious and may cause the patient discomfort.

This thesis investigates another method for such measurements; by using a non-contact, capacitively coupled sensor, many of these shortcomings can be overcome. While this method relieves some problems, it also introduces several design difficulties such as: circuit noise, extremely high input impedance and interference. A capacitively coupled sensor was created using the bottom layer of a printed circuit board (PCB) as a capacitor plate and placing it against the signal source, that acts as the opposite capacitor plate. The PCB solder mask layer and any air in between the two acts as the insulator to create a full capacitor. The signal picked up by this sensor was then amplified by 60 dB with a high input impedance amplifier circuit and further conditioned through filtering.

Two measurements were made of the same circuit, but with different input impedances; one with 10 M Ω and one with 10 G Ω input impedance. Additional filtering was designed to combat interference from the main power lines at 50 Hz and 150 Hz that was discovered during initial measurements. The circuits were characterized with their transfer functions, and the ability to amplify a very low-level, low frequency input signal. The results of these measurements show that high input impedance is of critical importance for the functionality of the sensor and that an input impedance of 10 G Ω is sufficient to produce a signal-to-noise ratio (SNR) of 9.7 dB after digital filtering with an input signal of 25 μ V at 10 Hz.

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Acronyms

ADC analog-to-digital converter

ECG electrocardiography

EEG electroencephalography

FFT fast Fourier transform

LSB least significant bit

PCB printed circuit board

RMS root mean square

SNR signal-to-noise ratio

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Chapter 1

Introduction

1.1 Background and Motivation

In modern medicine, the measurement of electrophysiological signals play a key role in health monitoring and diagnostics. Electrical activity originating from our nerve and muscle cells conveys real-time information about our current health state. The two most common and actively used techniques for measuring such signals are electrocardiography (ECG) and electroencephalography (EEG). ECG is the measurement of the electrical signals generated by the heart muscle as it is beating; and has applications ranging from diagnosis of arrhythmia and infarction to simple heart rate monitoring. Similarly, EEG is the measurement of the electrical signals originating from the brain; and is used to monitor and diagnose diseases and disorders such as epilepsy, sleep disorders, Alzheimer and Parkinson. In addition, EEG could be used to help develop advanced prosthesis that could be controlled by the patient's thoughts. Another, non-clinical application might be detection of drowsiness in car drivers to avoid accidents related to driver fatigue [1,2].

ECG and EEG signals are extremely weak, with EEG signals reaching down to the order of tens of microvolts [3]. Additionally, these signals are very slow with frequencies reaching down to between 0 and a few Hz. These two characteristics combined makes them extremely difficult to measure and sets strict requirements on the signal-to-noise ratio and linearity of the interfacing analog circuitry. The most critical part of the measurement is a robust and accurate sensor interface circuit to amplify these slowly changing, low-level signals in the presence of noise sources and other interference. The design of such electronic circuits is a major challenge.

Today's circuits for measuring EEG largely relies on direct contact with the subject's skin and uses conductive coupling to measure these signals. This is done by fixating electrodes – such as silver chloride or gold electrodes – to the naked skin with a conductive gel. This gel acts as a glue to fix the electrode to the skin and together they form an electrical transducer to transform ionic current at the surface of the skin to electron current that can be amplified with a high-precision, low-noise electrical readout circuit.

Figure 1.1. Silver chloride (AuCl) EEG electrodes for use with a conductive gel and skin preparation pads for cleaning of the skin.

To use this kind of direct contact electrodes, preparation to the measuring surface is needed by means of shaving of hair and possibly even abrasion of the skin. Moreover, the chemical makeup of the gel risks irritating the skin, making it uncomfortable for the patient. For long-period measurements, the gel dries out, hardens and loses its conductive ability as well as its ability to hold the electrode in place. This means that the skin has to be cleaned and the whole process of attaching the electrodes has to be remade – a very time consuming task. All of these issues makes it very desirable to develop a measurement technique without the need for direct electrical contact with the skin.

Such a technique would eliminate patient discomforts due to the use of a conductive gel as well as increase the speed and efficiency of the measurement. In extension, such a technique could enable measurement at greater distances, for instance through clothing, chairs, mattresses, telephones (while talking to your physician perhaps) or any other sufficiently closely located object. This eliminates the need for presence at a medical clinic during measurement; one could for instance have a simple ECG monitoring device at home that automatically sends measurement data to one's physician over the Internet. This would result in a low-cost, simple, remote monitoring system that improves the patient care, reduces the overall health care cost for society and would enable a significant improvement in treating diseases and disabilities.

1.2 Goal of This Thesis

As a part of a long-term objective of designing and manufacturing a non-contact, capacitively coupled sensor system for measuring ECG, EEG and other electrophysiological signals; this thesis investigates the feasibility of using a small printed circuit board (PCB) with a low-noise amplifying circuit – where the PCB itself works as a capacitive sensor – to measure input signals of very low amplitude and frequency.

1.3 Organization of the Thesis

This thesis consists of five chapters, a glossary of terms and a number of appendices.

Chapter 1 provides the background, motivation and describes the goal of this thesis. This chapter also describes and characterizes ECG and EEG signals.

Chapter 2 deals with the design choices made for the circuit. It also includes an analysis of the complete circuit by deriving the transfer functions of each part of the circuit. This allows for a better, more quantitative understanding of the circuit functions.

Chapter 3 describes the layout of the PCB manufactured to support circuit measurements.

Chapter 4 presents the data acquired from measurements of the circuit and contrasts different measurements to find the most suitable circuit configuration.

Chapter 5 summarizes the report and presents the conclusions of the results with regards to the thesis goal. Moreover, this chapter also presents the author's ideas of what can be improved and what would benefit from further investigation.

The glossary introduces acronyms and terms that are used throughout this report and that are thought to be new to the reader.

Appendices include material that the author feels is necessary to repeat the work done for the thesis and that are too long to include within the report itself; such as: circuit schematics, detailed layout figures and source code for computations.

1.4 Electrophysiological Signals

Electrophysiological signals are electrical signals originating from cells in the human body. There are many different types of cells, but the ones of consequence for ECG and EEG are muscle cells and nerve cells. The electrical signals are created in the same way for both types of cells; where the cell membranes produce electrochemical impulses as a result of cell excitation. These impulses are conducted over the membranes of the cells. With a collection of a large number of such cells, an electrical signal strong enough to be measured can be generated [3].

1.4.1 Electrocardiography

Electrocardiography is the process of recording electrical impulses generated by the heart as it is beating. The heart is a complex system and it is outside the scope of this thesis to explain exactly how it works and how ECG signals are generated (more information can be found in [3]). However, a simple, qualitative explanation is made below.

The heart consists of four different blood-filled chambers; the left and right atria and the left and right ventricles. When blood arrives to the heart through the blood vessels from the body, it first arrives to the atria. The blood is then transferred to the ventricles before it is pumped back out into the body. It is the

Figure 1.2. Graph of a typical ECG signal with its respective parts marked out.

activation of the muscle cells in these chambers by means of muscle contraction that produces the ECG signal.

A signal of a normal ECG cycle in the time domain is shown in Figure 1.2; the different parts of the signal marked by letters P, Q, R, S, T and U. The P-wave is a result of atrial depolarization (activation of the muscle cells). The part consisting of letters Q, R and S is commonly referred to as the QRS-complex and appears at the depolarization of the ventricles; also during this time, the electrical repolarization (that happens during resting of the muscle cells) of the atria happens. The T-wave appears during the repolarization of the ventricles. The U-wave is very weak and is not always observed. While there are different theories, a generally accepted explanation to the origin of the U-wave is yet to be found.

Part	Amplitude	Duration
P-wave	100-150 μV	100-120 ms
P-R segment	—	100 ms
QRS complex	1-1.5 mV	80-100 ms
S-T segment	—	140-160 ms
T-wave	150-200 μV	16-180 ms
U-wave	50 μV	40 ms

Table 1.1. Summary of the parts of an ECG signal as shown in Figure 1.2, and their approximate amplitude and duration as measured on the surface of the skin [3].

Table 1.1 shows approximate values of the amplitudes and durations of the different parts the ECG cycle that is shown in Figure 1.2. These numbers show that the ECG signal is very weak and slow, making it a challenging signal to measure. It has been shown that almost all the power in a typical QRS-complex is contained in frequencies below 30 Hz and that the peak power is in the range 4-12 Hz [4] – information that is very useful when designing a sensor readout

circuit.

1.4.2 Electroencephalography

Electroencephalography is the process of recording electrical impulses generated from nerve cells in the brain. The many nerve cells in the human brain creates electric fields when excited. These fields can be measured at the scalp of the head or in the brain tissue. The origin and meaning of these electrical signals is very complex and not fully understood to this day, hence an explanation similar to that of the ECG above is difficult to make. Therefore, the study of EEG is largely based on statistical and empirical data.

Observing the time domain signal of an EEG is rarely of interest, and as such most studies observe the information contained in the frequency domain of the signal. Most of the power of the EEG is contained in frequencies below 50 Hz, with special attention often given to alpha (α) waves at around 10 Hz [3, 5–7]. The reason for this is that the alpha waves are easily recreated and measured at the back of the head when the subject's eyes are closed. The other information carrying frequencies of the EEG and where they can be observed is listed in Table 1.2.

Wave		Observed	Frequencies
Alpha	(α)	At the back of the head with eyes closed	8-13 Hz
Beta	(β)	At the top and front of the head	13-30 Hz
Delta	(δ)	In infants and sleeping adults	0.5-4 Hz
Theta	(θ)	In children and sleeping adults	4-8 Hz

Table 1.2. Summary of the different EEG waves: where they are measured and their frequency bands.

When measured at the brain tissue, EEG signals have an amplitude of about 100 mV, however the level is significantly lower than this at the scalp; reaching down to 10-100 μ V [3]. Thus, EEG signals are weaker than ECG signals and puts a stricter requirement on the signal-to-noise ratio of the sensor and readout circuitry that is used to measure them.

Chapter 2

Circuit Design

The design of the circuit in this thesis is based on the papers by Sullivan, Deiss and Cauwenberghs [5] and Harland, Clark and Prance [7]. The design makes use of a low-noise sensor readout circuit that uses the PCB itself as the sensor. One metal layer on the PCB acts to capacitively couple charge from the skin of the subject to the amplifying circuitry. Because of the size of the circuit, it was split up into two different PCBs intended to be placed on top of each other in order to reduce the area of the sensor. A reduced area would mean that more sensors could be placed on the same surface – an important property for EEG measurement where it is beneficial to have a large number of sensors [8].

The idea here is that the bottom layer of the sensor board is filled with metal and acts as an antenna to pick up electromagnetic fields in its vicinity. The board is put flat onto the subject to create a capacitor with its metal layer as one plate and the skin of the subject as the other. Isolation between the plates are created by the solder mask, air and possibly the hair and/or the clothing of the subject. A metal shield layer is placed between the sensor layer and the rest of the circuitry to reduce the amount of interference caused by the circuit. The signal picked up by this capacitor is then amplified 50 times by a very low noise and low input bias current (high input impedance) instrumentation amplifier; and then further filtered and amplified 20 times by a low noise operational amplifier. The total amplification achieved is thus 1000.

Because the sensor capacitor effectively cuts off the DC connection to ground at the input of the instrumentation amplifier, the charge supplied by its input bias currents integrates onto the capacitor to generate an increasing offset voltage. Eventually this voltage would be large enough to go outside of the amplifier's common mode input range and at that point, the amplifier would stop operating properly and the output would be undefined. To prevent this, there is a need to supply the input bias current with a return path to ground. In this thesis, a resistor is used to provide this path to ground. This resistor, together with the sensor capacitance, constitutes a high-pass filter and thus the resistance needs to be very high (because the sensor capacitance is low) to not filter out the low frequency ECG and EEG signals.

The goal is to see if it is possible to get a usable signal with this simple circuit (as opposed to the more complex bias circuit described in [5]), and to investigate the possibilities and limitations of such a circuit.

2.1 System Overview

The system consists of four main parts; the capacitor plate that acts as the actual sensor; a first amplification stage consisting of an instrumentation amplifier with high input impedance; two notch filters for filtering out interfering signals; and a second amplification stage providing additional gain and filtering. The complete system is shown in Figure 2.1.

Figure 2.1. Circuit diagram of the complete system including component values, but excluding decoupling capacitors for the amplifiers.

The sensor consists of a metal layer on the PCB that acts as a capacitor plate and is insulated from its opposing plate (thought to be the skin of the subject) by the solder mask and anything else that is placed in between the two plates. This capacitor is shielded from the rest of the circuit by a separate metal layer on the PCB and connected to the positive input of the instrumentation amplifier.

The instrumentation amplifier is set to amplify the signal about 50 times and is connected in a feedback loop. The loop filter acts as a low-pass filter and makes the negative input slowly track the positive input. The last part of the circuit further amplifies the signal approximately 20 times and provides band-pass filtering over the wanted signal band. This creates a theoretical maximum gain of 1000 or 60 dB in the circuit passband.

2.1.1 The Capacitive Sensor

The sensor capacitor is built up as a parallel-plate capacitor, i.e. two metal plates separated by an insulator material. In this case the two plates are made up by the bottom metal layer on the PCB and the skin of the person whose signal is to be measured. This setup is described in Figure 2.2. Assuming that the area of the plates are much larger than the distance between them, the fringing capacitance at the edges can be neglected and the capacitance of the parallel-plate capacitor is given by the following relation [9]:

$$C = \epsilon_0 \epsilon_r \frac{A}{d}, \quad (2.1)$$

where A is the area of the two plates that overlap; d is the distance between the two plates; and ϵ_0 and ϵ_r are the electric constant and the relative dielectric constant of the insulating material, respectively.

Figure 2.2. A two-dimensional cross-section of the sensor PCB. This shows how the different variables in equation (2.1) are determined physically.

In this case, two of these variables are varying with the position of the sensor on the subject's body, the distance d and the dielectric constant ϵ_r . This makes it difficult to get an exact figure on the capacitance of the sensor. The total dielectric constant is made up a combination of the constants of the solder mask, the air trapped between the plates and possibly even hair and/or clothing that is caught between the two capacitor plates.

As will be seen later, in section 2.2, the sensor capacitance creates a high-pass filter together with the input resistance of the first amplifier. Lowering the capacitance by, for example, moving the sensor further away from the body, will thus shift the high-pass filter cutoff frequency upwards. Because of this, it is necessary to have as high an input resistance as possible to still enable the low frequencies that are to be measured through. Moreover, the higher the input resistance, the less impact the capacitance change will have on the cutoff frequency of the filter.

2.1.2 Instrumentation Amplifier

The instrumentation amplifier used is the INA116 from Texas Instruments. The reason this amplifier is used is that it has a very high input impedance and thus

very low input bias currents. Moreover, it has a very low current noise, which makes it particularly suitable for this sort of low noise application.

Other features are buffered guard output pins for each input, that follows the input but has a low output impedance [10]. These are ideal to use as the shielding of the sensor, because if the potential of the shield follows that of the sensor plate, the stray capacitance between the two metal layers are minimized and thus reduces leakage.

The amplifier is of a standard instrumentation amplifier type with three operational amplifiers where, the first two amplifies the signal and the third acts as a difference amplifier subtracting the two amplified signals. The analysis of this internal circuit is shown in section 2.2.2.

2.1.3 Notch Filters

A notch filter is a very narrow band-stop filter, i.e. it filters out a specific frequency from the input signal. Such filters are commonly used to suppress interferers in radio circuits, and to filter out hum from the main power lines in audio frequency circuits.

Upon first measurement of the circuit, large signal amplitudes at 50 Hz and 150 Hz were measured at the input of the instrumentation amplifier. These were large enough to saturate the last stage of the circuit and thus needed to be removed. To suppress these interferers, two notch filters were inserted between the two stages; one at each of these frequencies. The notch filters are of the active twin-T type and are analyzed in section 2.2.5.

Two operational amplifiers were needed to realise each notch filter, so a total of four amplifiers were used. The amplifier used for these filters were the OPA27 from Texas Instruments. This is a low noise precision amplifier [11] that was selected because it would not contribute significantly to the circuit noise.

2.1.4 Second Gain Stage and Band Pass Filtering

This circuit provides additional gain and filtering to the signal with the help of an operational amplifier. The operational amplifier used for this circuit is the LT6010 from Linear Technology. It is a precision amplifier that can be run off a single ended 3 V supply [12]. It is set in a band-pass configuration with a gain of 20.5 in the passband. Immediately following it is an RC low-pass filter that limits the upper band of the signal and reduces the aliasing during a subsequent sampling of the signal. The transfer function for this part of the circuit is derived in section 2.2.4.

The first part of the circuit is a high-pass filter that removes any DC voltage from the output of the instrumentation amplifier. It also serves to shift the common mode level at the input of the operational amplifier to the center of its supply voltage. The filter is analyzed in section 2.2.3.

2.2 Circuit Analysis

The rest of this chapter is devoted to the analysis of the different parts of the circuit. Here, the transfer functions of the whole system are derived analytically. The Laplace transform is used throughout to express the frequency dependency of the circuits.

2.2.1 Capacitive Sensor

This part of the circuit is the actual sensor. It is meant to create a capacitor between the circuit board and the human skin to pick up the weak signals that are generated by the human body. The capacitor creates a very high impedance to DC voltages and together with the biasing resistors for the input instrumentation amplifier, it constitutes a filter with a high-pass characteristic.

The capacitance value of the sensor is difficult to estimate since it depends on a number of factors:

- The area of the sensor,
- The distance between the skin and the sensor,
- The dielectric constant of the materials in between the sensor metal layer and the skin of the body, such as the PCB solder mask, any body hair and air that is caught in between.

Figure 2.3. Small signal model of the capacitive sensor. C_{sensor} is the capacitance made up by the PCB and the skin; R_{bias} is the resistor biasing the input of the instrumentation amplifier; and C_{in} is the input capacitance of the instrumentation amplifier.

The circuit is shown in Figure 2.3 where C_{in} is the capacitance at the input of the instrumentation amplifier. A simple AC analysis reveals its high-pass filter nature. Assuming that the input resistance of the instrumentation amplifier is much higher than the bias resistor, the voltage at the output is

$$v_{out}(s) = \frac{R_{bias} \parallel \frac{1}{sC_{in}}}{\frac{1}{sC_{sensor}} + \left(R_{bias} \parallel \frac{1}{sC_{in}}\right)} \cdot v_{skin}(s). \quad (2.2)$$

After some algebra the transfer function $H(s)$ is found as

$$H(s) = \frac{v_{out}(s)}{v_{skin}(s)} = \frac{C_{sensor}}{C_{sensor} + C_{in}} \cdot \frac{s}{s + \omega_0}, \quad (2.3)$$

where

$$\omega_0 = \frac{1}{R_{bias}(C_{sensor} + C_{in})} . \quad (2.4)$$

The high-pass characteristic of the filter is visible in (2.3) and the cutoff frequency is given by (2.4). Another noticeable thing in (2.3) is the constant loss of $\frac{C_{sensor}}{C_{sensor} + C_{in}}$ in the passband. Depending on the capacitance value of the sensor, this loss can be more or less significant. If the capacitance drops to within one order of magnitude of the input capacitance, the loss will be more than 10 % of the amplitude.

Figure 2.4. Transfer function of the filter at the input. The -3 dB cutoff frequency is indicated with the dashed line and is 0.12 Hz.

Figure 2.4 shows the transfer function as given by (2.3) with $R_{bias} = 10 \text{ G}\Omega$, $C_{in} = 7 \text{ pF}$ as per the amplifier datasheet [10] and $C_{sensor} = 125 \text{ pF}$ (from measurements described later).

2.2.2 Input Amplification Stage

This is an analysis of the internals of the INA116 instrumentation amplifier. The amplifier is made up of three operational amplifiers as shown in Figure 2.5. The first two amplifiers provides gain to both the inputs. The gain is set by choosing the appropriate value of R_G , which is an external resistor. The final amplifier then outputs the difference between these two amplified signals. The INA116 also includes a buffer amplifier at each of its inputs [10], however since they ideally should not affect the signal, they are left out of the following analysis. A feedback network that is external to the amplifier is connected to the negative input. The input signal is connected to the positive amplifier input.

To simplify the analysis, it is assumed that all operational amplifiers are ideal, i.e. they have infinite gain, infinite input resistance and zero output resistance.

Figure 2.5. Small signal model of the instrumentation amplifier with the interesting nodes marked out by A, B, C, D, E, F. R_1 , R_2 and R_f are internal to the amplifier; R_G sets the amplification; and R_{ext} and C_{ext} is an external feedback loop.

From Figure 2.5 the following nodal equations can be identified:

$$\text{Node A: } -v_a(s) \cdot sC_{ext} + \frac{v_{out}(s) - v_a(s)}{R_{ext}} = 0, \quad (2.5)$$

$$\text{Node B: } \frac{v_a(s) - v_{in}(s)}{R_G} + \frac{v_d(s) - v_{in}(s)}{R_f} = 0, \quad (2.6)$$

$$\text{Node C: } \frac{v_{in}(s) - v_a(s)}{R_G} + \frac{v_e(s) - v_a(s)}{R_f} = 0, \quad (2.7)$$

$$\text{Node F: } \frac{v_d(s) - v_f(s)}{R_1} - \frac{v_f(s)}{R_2} = 0, \quad (2.8)$$

$$\frac{v_e(s) - v_f(s)}{R_1} + \frac{v_{out}(s) - v_f(s)}{R_2} = 0. \quad (2.9)$$

The above equations can be rewritten as

$$v_a(s) = v_{out}(s) \cdot \frac{1}{1 + sR_{ext}C_{ext}}, \quad (2.10)$$

$$v_d(s) = \frac{R_f}{R_G} \left(v_{in}(s) - v_a(s) \right) + v_{in}(s), \quad (2.11)$$

$$v_e(s) = -\frac{R_f}{R_G} \left(v_{in}(s) - v_a(s) \right) + v_a(s), \quad (2.12)$$

$$v_{out}(s) = \frac{R_2}{R_1} \left(v_d(s) - v_e(s) \right). \quad (2.13)$$

Inserting (2.11) and (2.12) into (2.13) and substituting $v_a(s)$ according to (2.10) yields

$$v_{out}(s) = \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right) \left(v_{in}(s) - v_{out}(s) \cdot \frac{1}{1 + sR_{ext}C_{ext}} \right). \quad (2.14)$$

It then follows that the transfer function is given by

$$H(s) = \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right) \cdot \frac{s + \omega_z}{s + \omega_p}, \quad (2.15)$$

where

$$\omega_z = \frac{1}{R_{ext}C_{ext}} \quad \text{and} \quad \omega_p = \frac{1 + \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right)}{R_{ext}C_{ext}}. \quad (2.16)$$

Inspection of (2.15) and (2.16) shows that the low-pass filter in the feedback loop yields one pole, ω_p , and one zero, ω_z , in the transfer function. It is also evident that (provided that the gain is larger than 1) $\omega_p \gg \omega_z$, which indicates a high-pass characteristic.

At low frequencies, the gain of the circuit is

$$|H(0)| = \frac{\frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right)}{1 + \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right)} \approx 1 \quad \text{if} \quad \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right) \gg 1.$$

At higher frequencies, the gain of the circuit becomes

$$|H(s \rightarrow \infty)| \approx \frac{R_2}{R_1} \left(1 + \frac{2R_f}{R_G} \right).$$

Another factor not taken into account here is the upper bandwidth limit of the circuit. This limit is dependent on the instrumentation amplifier itself and varies depending on the gain of the amplifier. At a gain setting of 100 (40 dB) – this thesis configures the circuit to a gain of 51, however 100 is the closest setting to 51 specified in the datasheet – the -3 dB bandwidth of the amplifier is 70 kHz [10], which is well above the requirement for this system.

Figure 2.6 shows the plotted transfer function as given in (2.15). R_f , R_1 and R_2 have values 25 k Ω , 60 k Ω and 60 k Ω respectively, as specified in the datasheet [10]. From (2.15) the gain resistor R_G is 1 k Ω to give 51 times amplification. Finally, the external feedback network is set to give a closed-loop pole at 1.62 Hz with $R_{ext} = 510$ k Ω and $C_{ext} = 10$ μ F as given by (2.16).

2.2.3 Interstage Filter and Level Shifter

This circuit is a simple passive high-pass filter with a built in level shift. The level shift is done to bring the DC bias of the signal to the middle of the voltage rails of the second amplifier that uses a single ended supply voltage. The circuit is shown in Figure 2.7.

Figure 2.6. Transfer function of the input amplification stage. It is a high-pass filter and its -3 dB cutoff frequency is indicated by the dashed line at 1.62 Hz.

Figure 2.7. Small signal model of the interstage high-pass filter and level shifter.

Starting with a DC analysis, the capacitor is replaced by an open circuit. Since node V_{out} is floating, there is then no current flowing through the resistor and the output voltage must be equal to the voltage of the DC voltage source, i.e. there is a level shift.

$$V_{out} = V_{ref} \quad (2.17)$$

For the small signal AC analysis the DC voltage source is short circuited to ground. The current flowing through the circuit is then

$$i_{out}(s) = \frac{v_{in}(s)}{R + \frac{1}{sC}}. \quad (2.18)$$

The output voltage is then the voltage drop over the resistor caused by i_{out} and is given by:

$$v_{out}(s) = \frac{sRC}{1 + sRC} \cdot v_{in}(s). \quad (2.19)$$

Rewriting (2.19) gives the transfer function of the circuit as

$$H(s) = \frac{v_{out}(s)}{v_{in}(s)} = \frac{s}{s + \omega_0}, \quad \text{where } \omega_0 = \frac{1}{RC}. \quad (2.20)$$

Equation (2.20) shows that this is a high-pass filter with a cutoff frequency of $\omega_0 = \frac{1}{RC}$.

Figure 2.8. Transfer function of the interstage high-pass filter. The -3 dB cutoff frequency is marked by the dashed line and is at 0.80 Hz.

Figure 2.8 shows the plotted transfer function of the filter. With component values $R = 20 \text{ k}\Omega$ and $C = 10 \text{ }\mu\text{F}$, the cutoff frequency is at 0.8 Hz.

2.2.4 Second Amplification Stage

The second amplification stage provides more gain to the signal as well as additional filtering. It consists of an active band-pass filter followed by a passive low-pass filter. The circuit is shown in Figure 2.9.

For the small signal AC analysis of this circuit we assume first that the operational amplifier is ideal and then we short circuit the DC voltage source V_{ref} to ground. Looking at Figure 2.9 we get the following nodal equations:

$$\text{Node A: } \frac{-v_{in}(s)}{R_1 + \frac{1}{sC_1}} + \frac{v_b(s) - v_{in}(s)}{R_2 \parallel \frac{1}{sC_2}} = 0, \quad (2.21)$$

$$\text{Node C: } \frac{v_b(s) - v_{out}(s)}{R_3} - v_{out}(s) \cdot sC_3 = 0. \quad (2.22)$$

Solving (2.21) and (2.22) for $v_b(s)$ and equating them gives the output voltage as

$$v_{out}(s) = v_{in}(s) \left(1 + \frac{R_2 \parallel \frac{1}{sC_2}}{R_1 + \frac{1}{sC_1}} \right) \cdot \frac{1}{1 + sR_3C_3}. \quad (2.23)$$

Figure 2.9. Small signal model of the second amplification stage and output filter with nodes used for analysis marked out with A, B, C. R_1 , C_1 and R_2 , C_2 creates a band-pass filter; and R_3 , C_3 is a low-pass filter.

This results in the transfer function

$$H(s) = \left(1 + \frac{R_2}{R_1} \cdot \frac{s}{s + \omega_{HP}} \cdot \frac{\omega_{LP1}}{s + \omega_{LP1}} \right) \cdot \frac{\omega_{LP2}}{s + \omega_{LP2}}, \quad (2.24)$$

where

$$\omega_{HP} = \frac{1}{R_1 C_1}, \quad \omega_{LP1} = \frac{1}{R_2 C_2}, \quad \omega_{LP2} = \frac{1}{R_3 C_3}. \quad (2.25)$$

If $\omega_{LP1} \approx \omega_{LP2} \gg \omega_{HP}$, we see from (2.24) that this stage forms a band-pass filter with a DC gain of 1 and a passband gain of $1 + \frac{R_2}{R_1}$.

Figure 2.10. Transfer function of the second amplification stage. The -3 dB cutoff frequencies are marked by the dashed lines and are 0.78 Hz and 67.50 Hz.

Figure 2.10 shows the transfer function with the following component values:

$$R_1 = 20 \text{ k}\Omega$$

$$C_1 = 10 \text{ }\mu\text{F}$$

$$R_2 = 390 \text{ k}\Omega$$

$$C_2 = 3.9 \text{ nF}$$

$$R_3 = 10.5 \text{ k}\Omega$$

$$C_3 = 150 \text{ nF}$$

The passband of the filter in the figure is between 0.78 Hz and 67.5 Hz.

2.2.5 Notch Filter

A notch filter removes a single frequency component from a signal. The effectiveness of the filter depends on its quality factor (Q). A higher value of Q generates a deeper notch and a narrower band of attenuation, while the opposite is true for a low value of Q .

There are a few different types of circuits that can be used to generate a notch filter; some of them are listed in [13]. The circuit chosen was an active filter based on the common passive twin-T notch filter as shown in Figure 2.11.

Figure 2.11. Circuit diagram of an active twin-T notch filter circuit.

As will be shown below, the original passive filter can at best give $Q = \frac{1}{4}$, while the active feedback makes it possible to achieve much higher values for Q . This circuit topology, however, makes Q quite sensitive to component matching. Poor matching between elements will reduce the depth of the notch, but it is a simple circuit requiring few elements and is simple to realize.

The transfer function for a notch filter is given below [13]:

$$H(s) = \frac{s^2 + \omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2} \quad (2.26)$$

This analysis assumes that the relationships between the passive elements in the filter are as follows:

$$R = R_1 = R_2 = 2R_3 \quad \text{and} \quad C = C_1 = C_2 = \frac{C_3}{2}. \quad (2.27)$$

And the frequency of the notch (f_0) is then given by

$$2\pi f_0 = \omega_0 = \frac{1}{RC}. \quad (2.28)$$

Figure 2.12 shows a simplified circuit of the notch filter with important nodes marked out. The operational amplifier buffers are assumed to be ideal and thus removed, and the voltage divider at the output substituted for a voltage dependent voltage source.

Figure 2.12. Small signal model of the notch filter with nodes used for analysis marked out by A, B, X and Y.

Looking at Figure 2.12, we get the nodal equations

$$\text{Node X: } v_x(s) = \frac{R_4}{R_4 + R_5} \cdot v_{out}(s) = K \cdot v_{out}(s), \quad (2.29)$$

$$\text{Node A: } \frac{v_{in}(s) - v_a(s)}{R} + \frac{v_x(s) - v_a(s)}{1/s2C} + \frac{v_{out}(s) - v_a(s)}{R} = 0, \quad (2.30)$$

$$\text{Node B: } \frac{v_{in}(s) - v_b(s)}{1/sC} + \frac{v_x(s) - v_b(s)}{R/2} + \frac{v_{out}(s) - v_b(s)}{1/sC} = 0, \quad (2.31)$$

$$\text{Node Y: } \frac{v_a(s) - v_{out}(s)}{R} + \frac{v_b(s) - v_{out}(s)}{1/sC} = 0. \quad (2.32)$$

Rewriting equations (2.30), (2.31) and (2.32) and substituting for $v_x(s)$ as given

by (2.29) gives

$$v_a(s) = \frac{v_{in}(s)}{2 + 2 \cdot sRC} + v_{out}(s) \cdot \frac{1 + 2K \cdot sRC}{2 + 2 \cdot sRC}, \quad (2.33)$$

$$v_b(s) = v_{in}(s) \cdot \frac{sRC}{2 + 2 \cdot sRC} + v_{out}(s) \cdot \frac{2K + sRC}{2 + 2 \cdot sRC}, \quad (2.34)$$

$$v_{out}(s) = \frac{v_a(s)}{1 + sRC} + v_b(s) \cdot \frac{sRC}{1 + sRC}. \quad (2.35)$$

After inserting (2.33) and (2.34) into (2.35) and using (2.28), some algebraic re-ordering results in the transfer function of the circuit as

$$H(s) = \frac{v_{out}(s)}{v_{in}(s)} = \frac{s^2 + \omega_0^2}{s^2 + 4(1 - K)\omega_0 \cdot s + \omega_0^2}. \quad (2.36)$$

Figure 2.13. Transfer functions of notch filters with different values of Q . The number next to each line is the value of Q as given by equation (2.37).

The value of Q can now easily be obtained by identification of equation (2.36) with equation (2.26) and substituting for K as given by equation (2.29).

$$Q = \frac{R_4 + R_5}{4R_5} = \frac{1}{4} \left(1 + \frac{R_4}{R_5} \right) \quad (2.37)$$

This means that Q can easily be controlled by the ratio of R_4 to R_5 and varied between $\frac{1}{4}$ and infinity. In reality, Q can never reach infinity because of circuit non-idealities and component mismatches, but it can get significantly higher than in the passive case¹. A potentiometer could be used to tune the quality factor in the actual filter implementation.

Figure 2.13 shows the difference in width of the notches depending on Q . Here, the ratio $\frac{R_4}{R_5}$ of the three curves is 0, 4 and 49.

¹In the passive twin-T notch filter, the common node in both branches is connected to ground, thus equivalent with setting R_4 to 0 in Figure 2.11. This yields a Q of $\frac{1}{4}$ as per (2.37).

2.2.6 The Complete System

Putting all the separate transfer functions for these sub-circuits together by multiplication results in the complete transfer function for the whole system and is shown in Figure 2.14. All the different sub-circuits were included once in the transfer function except for the notch filter, for which two copies were included – one at 50 Hz and one at 150 Hz, both with a Q of 5.

Figure 2.14. Transfer function of the complete system including two notch filters at 50 Hz and 150 Hz.

A comparison to the transfer function without any notch filters (Figure 2.15) reveals that the 50 Hz notch limits the upper bound on the bandwidth of the circuit somewhat – from 69.2 Hz to 43.4 Hz. This does not pose a big problem, since the EEG and ECG signals have most of their power in frequencies below 40 Hz.

Figure 2.15. Transfer function of the complete system without any notch filters.

Chapter 3

Printed Circuit Board Layout

This chapter describes the construction of the two different PCBs that was created for this thesis. It also describes how the two notch filters were constructed on a piece of stripboard.

Schematic and PCB design software tools were used to create the necessary files for fabrication of both boards and are described in the next section. For consistency, the schematics created with this software were redrawn to look the same as the rest of the document. However, the schematics for the software tools are included in Appendix B and likewise the PCB layouts are included in Appendix C.

3.1 Software Tools

For the layout of the PCBs, an open source layout software called *PCB* from the *gEDA project*¹ was used. The *gEDA project* works to provide a set of tools for electronic design automation that are licensed under the GNU General Public License. This includes tools for schematic capture, netlisting, analog and digital circuit simulation and PCB layout, among other things.

PCB supports exporting its layout to the industry standard RS-274-X format and thus it is compatible with most PCB fabricators.

To support the layout development, schematic capture was first done in another *gEDA project* tool called *gschem* and netlisting from *gschem* to *PCB* was done with the *gnetlist* tool.

3.2 Sensor and Instrumentation Amplifier Board

This PCB contains the sensor and the INA116 instrumentation amplifier part of the system. It is a four layer board with the bottom layer constituting the sensor

¹<http://www.gpleda.org/>

capacitor plate; the layer above it is the shield; and the two top layers are circuit traces.

The schematic used to create this board is shown in Figure 3.1 and the supply decoupling is shown in Figure 3.2. The INA116 runs off a $\pm 5\text{ V}$ supply voltage and its reference pin is connected to ground. Figure 3.1 also shows a pair of bipolar transistors and circuitry related to them; this will be further explained in section 3.2.2.

Figure 3.1. Schematic for sensor and instrumentation amplifier stage including component values. Pin numbers are shown for the amplifier and the bipolar transistor packages.

Figure 3.2. Decoupling for the instrumentation amplifier.

3.2.1 Layout Considerations

Since this PCB was to be used to sense the very low-level input signals, it was important to try to layout the board as to reduce the noise and disturbances as much as possible. In effect, this meant to take certain precautions with the shielding of the sensor, the grounding scheme and guarding of the sensitive input signal.

Guarding of Inputs

The inputs of the INA116 amplifier are internally buffered before being amplified. The output of these buffers are not only connected to the rest of the instrumentation amplifier, but they are also connected to two guard pins on each side of the input pins of the amplifier package as shown in Figure 3.3. This means that there is access to a low impedance source that follows the input.

To reduce coupled interference on the input trace, another trace surrounding the input trace (and preferably also the input source) should be connected in a closed loop to these two guard pins [10]. In the case of this circuit, it was not possible to surround the complete sensor, as it was made up by a separate layer that only constituted one half of a capacitor, however, the input traces was still enclosed as much as possible with these guard rings.

Figure 3.3. Guard pins of the INA116 and recommended layout practice. The right hand side shows the internal buffers of the amplifier package and where the guard outputs; and the left hand side shows the physical package and how the input traces are surrounded by its corresponding guard.

Shielding

The shield metal layer was inserted to reduce the amount of charge coupled capacitively from the circuitry to the sensor metal layer. With the shield connected to a low impedance source, any charge coupled to it from the other circuitry would be absorbed into the source.

However, because the shield covered a whole layer of the PCB and with its proximity to the sensor layer, they would act as a parallel-plate capacitor and there would be a significant stray capacitance between these two layers as well. If the shield is assumed to be connected to the low impedance ground, then any

sudden change in potential at the sensor layer would couple some charge to the shield thus reducing the amplitude of the sensed signal.

To reduce the capacitive coupling between these two layers, the shield was connected to the guard pin of the input instead of to ground. This means that the shield follows the input and that the difference in potential between the two will be close to zero.

Grounding

In a low frequency, low noise circuit, it is important to control the return current paths. As opposed to the case of high frequency circuit – where the return current takes the path with the lowest impedance – at low frequencies it takes the closest path back. This means that there is a large risk that current paths will cross and interfere with each other if using a ground plane; instead a star grounding scheme was used, where each ground pin has a separate path to a common ground point on the board.

3.2.2 Unused Circuitry

Also included on this PCB was circuitry for resetting the input common mode levels with bipolar transistors that are controlled with a separate input signal. This is the technique used to bias the instrumentation amplifier in [5]. The footprints were included on the PCB to make it possible to investigate this method further, but due to time limitations and lack of information, this option was not pursued.

Moreover, the bias resistor of the negative amplifier input was found unnecessary – the bias current has a path to the output of the amplifier through the feedback resistor – and was thus not used.

3.2.3 Layout

The layout consists of a 25x25 mm, four layer PCB. The top layer houses the components and traces; the second layer contains additional traces; the third layer is the metal shield layer; and the bottom layer is the sensor plate. The first two layers are shown in Figure 3.4(a) – notice the guard rings on the right side of the instrumentation amplifier. Figure 3.4(b) shows a photo-realistic rendering of how the final fabricated PCB looks.

One problem due to limitations with the software is that all vias go through all layers. This means that both the sensor layer and shield are perforated with a number of vias, and this quite possibly contributes to interference on the sensor. Ideally, all vias except the one to the bottom layer would be blind vias, i.e. end inside the card. Unfortunately this was not possible with this PCB design software.

All components on this board are surface mount components. The different components are listed in Table 3.1 together with their values, tolerances and packages. Unused components are also marked in Table 3.1.

As will be seen in the measurements later, two different biasing configurations were measured; one with a 10 M Ω resistor and one with a 10 G Ω resistor. The

Figure 3.4. Instrumentation amplifier and sensor PCB in scale 2:1. Shield and sensor metal layers are not shown in (a) to better show the signal traces.

10 M Ω resistor is of the size 1206 with a tolerance of 1 %, whereas the 10 G Ω resistor is of size 0805 with a tolerance of 30 %.

Name	Value	Tol.	Pkg.	Note	Name	Value	Tol.	Pkg.	Note
R101	510 k Ω	1 %	0805		C102	1 nF	5 %	0603	
R102	1 k Ω	0.1 %	0805		C103	1 μ F	10 %	0805	
R103	60 k Ω		0805	Unused	C104	1 nF	5 %	0603	
R104	20 k Ω		0805	Unused	C105	1 μ F	10 %	0805	
R105	10 M Ω		1206	Unused	U101			SO16W	INA116
R106	10 M Ω	1 %	1206		Q101			SOT23	Unused
C101	10 μ F	10 %	0805		Q102			SOT23	Unused

Table 3.1. Components used on the sensor PCB including values, tolerances and packages.

3.3 Signal Conditioning Board

The second board contains a signal conditioning circuit and additional amplification. This is a two-layer board where the bottom layer is completely covered by a ground plane, and the top layer contains components and traces. Since the signal has already received substantial amplification by the first stage, the requirements on low noise are relaxed for this stage.

The schematic for this PCB is shown in Figure 3.5 and includes component values and supply decoupling for the operational amplifier.

This stage uses the LT6010 operational amplifier in a band-pass configuration running at a 3 V single-ended supply voltage. Since the first stage runs at a dual 5 V supply, the input needs to be shifted up to half of the supply voltage (1.5 V) which is done with the REF input. The LT6010 also has a shutdown pin that, when pulled up, effectively shuts off parts of the circuit to minimize its current

Figure 3.5. Schematic for second amplification stage including component values and supply decoupling. Pin numbers are shown for the operational amplifier package.

consumption [12]. However, this application is time-continuous so it was tied to the negative supply, i.e. ground.

3.3.1 Layout

The layout consists of a 21x21 mm board with two layers. The top layer holds all the components, signal traces and ground, while the bottom layer is completely covered with a ground plane. Figure 3.6(a) shows a screenshot from the PCB design tool of the top layer of the board; the bottom layer is not shown as it is just filled with metal. A photo-realistic rendering of the fabricated PCB is shown in Figure 3.6(b).

(a) The top layer of the board shows the component placement and signal routing.

(b) Photo-realistic rendering of the complete PCB seen from the top.

Figure 3.6. Signal conditioning PCB in scale 2:1. The bottom layer is not shown in (a) since it is only a metal ground plane.

All components used are surface mounted components. The different components are listed in Table 3.2 together with their values, tolerances and packages.

Name	Value	Tol.	Pkg.	Note	Name	Value	Tol.	Pkg.	Note
R201	20 k Ω	1 %	0805		C203	3.9 nF	10 %	0805	
R202	20 k Ω	1 %	0805		C204	150 nF	10 %	0805	
R203	390 k Ω	1 %	0805		C205	1 nF	5 %	0603	
R204	10.5 k Ω	0.1 %	0805		C206	1 μ F	10 %	0805	
C201	10 μ F	10 %	0805		U201			SO8	LT6010
C202	10 μ F	10 %	0805						

Table 3.2. Components used on the second stage PCB including values, tolerances and packages.

3.4 Notch Filters on Stripboard

Because the need for notch filters in the system was discovered after initial measurements, these circuits were not included on the original PCBs. Instead, they were designed afterwards and point-to-point soldered with through-hole components on a stripboard. This board was then connected in between the other two PCBs to complete the circuit. The schematic used for the design is shown in Figure 3.7. The figure also includes component values as well as supply decoupling. Resistor values are for both the 50 Hz notch and the 150 Hz notch (the latter within parentheses).

Figure 3.7. Schematic for notch filter including component values for a 50 Hz notch and supply decoupling. Values in parentheses are for a 150 Hz notch.

The operational amplifiers were both the OPA27 from Texas Instruments running from a ± 5 V supply. Both supply rails were decoupled to ground with a 10 μ F electrolytic and a 100 nF ceramic capacitor. The twin-T connection was made with 4.7 nF polypropylene capacitors with a tolerance of 2.5 %; and metal

film resistors of $680\text{ k}\Omega$ (for the 50 Hz filter) and $226\text{ k}\Omega$ (for the 150 Hz filter) with a tolerance of 1 %. The Q of the filter was adjusted with a $10\text{ k}\Omega$, 12-turn cermet potentiometer.

Figure 3.8. Stripboard complete with the two notch filters. Circuit Q s can be adjusted with the potentiometers.

Figure 3.8 shows the assembled stripboard with both notch filters placed side-by-side. Looking at one of the filters; the bottom part of the circuit is the twin-T connection; above that are the two operational amplifiers with the potentiometer in between; and at the top the electrolytic and ceramic supply decoupling capacitors are seen.

Chapter 4

Measurement

Measurement of the designed circuit was done in a regular electrical engineering lab. No special measures were taken to try to reduce the amount of electronic waves in the room.

Because of difficulties estimating the input capacitance of the circuit board – and thus the cut off frequency of the high pass filter it creates together with the input bias resistance – two different versions of the circuit was measured; one with a $10\text{ M}\Omega$ bias resistor at the input of the instrumentation amplifier; and one with a $10\text{ G}\Omega$ resistor as bias.

Figure 4.1. This photograph shows both PCBs connected together and with a Swedish one krona coin for scale. The right-most board is the sensor board, it is a bit larger than the signal conditioning board to the left of it.

Figure 4.1 shows the two PCBs connected together, ready to be measured. To get a feeling of the sizes of the two boards, the photograph was taken together

with a Swedish one krona coin. The larger of the two boards is 24x24 mm and the smaller is 21x21 mm.

4.1 Measurement Setup

For the circuit to function properly, there was a need to generate four different DC voltages and one single ended sine wave input signal. Measuring of the output signals were done on a real time digital oscilloscope. The following equipment was used for the measurement:

- Agilent InfiniiVision MSO7104A Mixed Signal Oscilloscope
- Agilent U8002A Single Output DC Power Supply
- Hewlett Packard 66312A Dynamic Measurement DC Source
- Thurlby Thandar Instruments TG120 20 MHz Function Generator

In order to have a deterministic input signal to the system for these measurements, two copies of the sensor board was used; one as the sensor and the other connected to the signal source. These two boards were then fixed to each other back to back with tape to create a capacitor. Care was taken so that there would be no overlap of the vias on the two boards in order to eliminate the risk for conductive coupling between them. Figure 4.2 shows how the two PCBs were attached to each other.

Figure 4.2. This photograph shows how the PCBs were fixed to each other to create the capacitor from the metal layers on each side of the boards.

The input signal board was connected directly to the function generator when large input signals were needed, and through an attenuator circuit, when low

input signals were needed. The sensor side board was populated with its intended components and connected to the second stage board. All the necessary power supply was connected to these two boards as well as the output of the second stage to the oscilloscope. All grounds from the circuit boards and the equipment were connected together in a star grounding scheme.

4.1.1 Oscilloscope Resolution

The Agilent MSO7104A oscilloscope was the most suitable oscilloscope available for these measurements. Since the measured signals are of very low amplitude, the resolution of the oscilloscope needs to be very high in order to accurately sample the signals. This oscilloscope has an 8-bit analog-to-digital converter (ADC). 8 bits means that there are 256 different voltage levels available; with the highest vertical resolution setting of 2 mV/div (full range of 20 mV), this translates into a least significant bit (LSB) of 78.125 μV . The quantization noise contributed to the signal would then be $\frac{1}{\sqrt{12}}\text{LSB} \approx 22.55 \mu\text{V}^1$. When comparing this value to the noise of the instrumentation amplifier ($28 \text{ nV}/\sqrt{\text{Hz}}$ [10]), it is clear that this resolution is too low to be able to characterize the circuit to a satisfactory degree.

However, the oscilloscope specification [15] states that it has a high resolution mode, in which the sequential signal data points are serially filtered before being mapped to the screen. Supposedly, this will allow for 12 bits of resolution when a time scale of $> 10 \mu\text{s}/\text{div}$ is used. A 12-bit conversion with the same vertical resolution would result in an LSB of $\sim 4.883 \mu\text{V}$. This translates to $\sim 1.41 \mu\text{V}$ of quantization noise. While this is still not a very low number, it is considerably better.

4.1.2 Function Generator Signal Level

The function generator was only able to generate voltages with amplitudes down to 2 mV. This was not low enough to measure the circuit's response to really low signals. Therefore – for most measurements – the output of the function generator was divided down by a simple voltage divider.

The voltage divider was designed to match the signal source output impedance at 50 Ω and to produce a 100 (40 dB) reduction of the signal. However, due to component tolerances the actual measured loss of the divider was 71.94 (37.14 dB). Because of the limit on the smallest possible measurable signal amplitude of the oscilloscope, when signals smaller than this were to be measured, they were taken directly from the function generator output. The real input signal to the circuit was then computed with the help of this value of the divider loss.

4.1.3 Measurement Methodology

Measurements were made by first setting up the circuit correctly for that particular measurement. The functionality of the circuit was then investigated by turning

¹The quantization noise is a random variable with a uniform distribution and its power is $\frac{\Delta^2}{12}$, where Δ is the smallest quantization step [14].

on all supply voltages and then observing the output signals from each stage and confirming that they appeared reasonable. To get precise values on the readings required for a measurement, the needed signals were sampled by the oscilloscope, saved to disk and then further processed in MATLAB®; for example to generate the frequency spectrum.

For the transfer function measurements, both the input and output signals were saved at different input frequencies. For each frequency point, the fast Fourier transform (FFT) was generated for both the input – after subtracting the attenuator loss if necessary – and the output. Then the frequency and root mean square (RMS) value of the input signal was identified in its frequency spectrum. Using this frequency, the RMS value of the output signal was found in the same way. With both RMS values found, the gain of the circuit at this frequency was easily computed by dividing them both. The gain was then plotted at each of these frequency points to create the transfer function.

4.2 Problem with Main Power Line Noise

Upon connection of the instruments and start up of the circuit, it was rather quickly realized that the capacitor plate created by the PCB picked up relatively high levels of noise from the ambient air. When inspecting this noise closer (see Figure 4.3), it seem to not be simple random noise, but rather a very regular pattern.

Figure 4.3. Plot of the noise present at the input of the instrumentation amplifier. This clearly shows large amounts of 50 Hz noise.

Performing an FFT (Figure 4.4) on the noise revealed that it was indeed high level noise at 50 Hz and additional noise at 150 Hz. The 50 Hz noise is quite obvious that it stems from the main power lines where large currents drawn by the many university electrical devices flow. The 150 Hz noise was more unexpected, but most likely it is the third harmonic of the 50 Hz signal on the mains. Interestingly, the level of this third harmonic is quite high compared to its fun-

damental (approximately 10 dB lower) – a characteristic of a square wave². My own speculation is that this might be attributed to the fact that most electrical equipment, e.g. computers, operate on switched power supplies, drawing more non-linear currents.

Figure 4.4. The Fourier transform of the noise at the input. The difference in voltage between the 50 Hz fundamental and the 150 Hz harmonic is 9.55 dB.

Regardless of the sources of these noise components, they lie in band (or slightly outside in the case of the 150 Hz noise) with the signals intended to be measured, and with their amplitude they would ruin the output signal and even overload the last amplification stage. Thus, they need to be filtered out as much as possible – at least to a degree that will allow the final operational amplifier to operate without saturating its output. As a solution to this, two notch filters were inserted between the first and second stage of the circuit; one to filter out the 50 Hz noise and one to filter out the noise at 150 Hz.

4.2.1 Notch Filters

The notch filters were of the active twin-T type with adjustable Q as was analyzed in section 2.2.5, and were put in series between the instrumentation amplifier and the intermediate high pass filter. Since there was no time to manufacture a separate PCB for these filters, they were built with through-hole components on an experimental board. Four operational amplifiers were required for the two filters; they were all the OPA27 low noise operational amplifier from Texas Instruments.

This filtering was enough to get an output signal that did not saturate the outputs of the last stage. However, noise was still very much present on the output signal, particularly at low input amplitudes, so additional filtering was needed in order to get a working signal.

²The Fourier series expansion of a square wave $x_{sq}(t) = \frac{4}{\pi} \left(\sin(\omega t) + \frac{1}{3} \sin(3\omega t) + \dots \right)$ shows that the third harmonic is $\frac{1}{3}$ (i.e. -9.54 dB) that of the fundamental.

4.2.2 Filtering in MATLAB®

While the analog notch filters removed enough noise to keep the signal from saturating the operational amplifier outputs, mismatches in component values reduce the depth and width of the notches so that they are far from ideal, and thereby passing some of the noise anyway. At low amplitudes the wanted signal can still be hidden by these noise components, as will be seen later.

The signal is meant to be digitized by an A/D converter, which means that digital filtering becomes available and that makes it easier to filter out the remaining noise. MATLAB® was used to emulate this digital filtering and steep notch filters for 50 and 150 Hz as well as a steeper low pass filter at 100 Hz was created to suppress higher frequency noise. The code to generate these filters are listed in Appendix A.

4.3 Measurements with a 10 MΩ Bias Resistor

The first set of measurements of the circuit was made with a high impedance 10 MΩ resistor for input biasing. Three measurements were made; one at the input of the instrumentation amplifier in order to characterize the transfer function of the input filter created by the bias resistor and the sensor capacitor; the second was a measurement of the whole system transfer function; and finally a measurement with a low level input signal at 10 Hz.

4.3.1 Input Filter

For this measurement, the input signal was connected directly to one of the capacitor plates, without connecting the intermediate attenuator. This was done to have an input signal large enough not to be significantly affected by the 50 Hz noise. The output signal was measured from the guard pin of the instrumentation amplifier. The oscilloscope probe only had a impedance of 1 MΩ, but since this pin outputs a buffered version of the input signal, it was possible to measure it without loading the high impedance input node, which would have destroyed the input signal.

Figure 4.5 shows the measured transfer function at the input of the instrumentation amplifier. The dashed line indicates the -3 dB frequency of the filter as 127.79 Hz, which is very high. Seeing as the signals this circuit aims to measure are normally well below 100 Hz, this is indeed unsatisfactory.

The constant loss at the pass band is most likely produced by voltage division between the sensor capacitance and the input capacitance of the instrumentation amplifier; and possibly leakage currents through the resistor.

Another result of this measurement is the possibility to put an estimate on the capacitance made up by the two PCBs connected together. The pole of a first order RC high pass filter is the same as the -3 dB frequency of the transfer function: $f_{-3dB} = \frac{1}{2\pi RC}$. With a resistor tolerance of 1 %, this means that the capacitance is between 123-126 pF, depending on the exact value of the resistor.

Figure 4.5. Measured transfer function of the input filter with a 10 M Ω bias resistor. The dashed line shows the -3 dB point at an input frequency of 128 Hz.

4.3.2 Transfer Characteristic

This measurement was done to try to generate the transfer function for the entire system, from the capacitively coupled sensor input to the filter at the output of the second stage operational amplifier. The measurement was made with a large amplitude input signal, but small enough not to saturate the output of any of the amplifiers.

An important part of this measurement was the notch filters work correctly. Without them, the output signal would not have been possible to measure satisfactorily, because the output of the second stage amplifier would simply swing between its minimum and maximum output voltage at 50 Hz.

Figure 4.6. Transfer function of the complete circuit with a 10 M Ω bias resistor. The -3 dB cut off frequencies are shown by the dashed lines. The maximum passband gain is 49.3 dB.

Figure 4.6 shows the measured transfer function of the whole circuit. Maximum gain of the circuit is around 50 dB which is significantly lower than the 60 dB it was designed to produce. This is due to the high cut off frequency of the input filter that suppresses these low frequency signals. Signals at 10 Hz are only amplified about 36 dB, and at 1 Hz, signals are not even amplified, but suppressed with over 10 dB.

A more positive observation is that the notch filters are working as expected. They are correct in frequency and they have quite deep and narrow notches. Exact measurement of the depths of the notches was not possible because of the very limited ability to fine tune the frequency of the function generator, particularly around lower frequencies, e.g. 50 Hz.

4.3.3 Low Level Signal Response

One important ability of the circuit is to amplify low amplitude signals at low frequencies. This measurement shows how the circuit handles an input signal at 10 Hz with an amplitude of $25\ \mu\text{V}$, the smallest signal possible to generate with the function generator and the attenuator.

Figure 4.7. Output of the circuit when a low level input signal is present.

As would be expected with regards to the input filter measurement, Figure 4.7(a) shows virtually no sign of the 10 Hz input signal. What is visible is instead a 50 Hz square wave. The FFT in Figure 4.7(b) shows a weak spur at 10 Hz, but compared to the first and third harmonics of the 50 Hz noise, it is unnoticeable. Other odd harmonics – although suppressed by circuit filtering – is also clearly visible in the FFT, indicating a square-like waveform at the output.

Not even the additional filtering in MATLAB® gives any usable signal. The result of the additional filtering is shown in Figure 4.8. Despite the FFT in Figure 4.8(b) showing that the 50 Hz and 150 Hz tones have properly been filtered out,

Figure 4.8. Output of low level signal after digital filtering in MATLAB®.

it is not possible – not even with a wild imagination – to find the 10 Hz signal in the time domain plot shown in Figure 4.8(a).

The power consumption for this measurement is summarized for the different circuit parts in Table 4.1. The table shows that the largest current consumers are the notch filters. Because they were not included on any PCB and were instead designed afterwards with through-hole components, they were not designed to minimize power consumption. Moreover, the two filters consists of two operational amplifiers each, for a total of four amplifiers. Assuming they consume a roughly equal amount of current, this means that each operational amplifier consumes approximately 26 mW of power. Contrasting this with the other amplifier (the LT6010 on the signal conditioning board) that consumes 0.38 mW, there is definitely room for future improvement here.

Circuit part	Voltage	Current	Power
Instrumentation amplifier	10 V	1.10 mA	11 mW
Notch filters	10 V	10.524 mA	105.24 mW
Operational amplifier	3.145 V	0.123 mA	0.38 mW
Total			116.62 mW

Table 4.1. Supply voltage, current and power consumption of the different parts of the sensor readout circuit with a 10 M Ω resistor. The notch filters entry here includes both filters and thus four operational amplifiers, which is why the number is so large.

4.4 Measurements with a 10 G Ω Bias Resistor

In the second set of measurements, the 10 M Ω biasing resistor at the input of the instrumentation amplifier was substituted for a higher resistance 10 G Ω resistor. This was done to try to lower the pole in the first high pass filter generated by the sensor capacitance and this input bias resistance. Two implications of using resistors with resistances this high are:

1. High thermal noise voltage, as given by the resistor thermal voltage noise equation: $v_{noise}^2 = 4kT\Delta fR$,
2. Even a very small current generates a significant voltage drop over the resistor.

The second point does not have any effect in this circuit because the DC offset is filtered in the feedback loop of the amplifier, and because of the low input bias currents of the amplifier – nominally at 3 fA [10] – the offset is still low enough to keep inside the common mode input range of the amplifier.

The first point, however, may limit the dynamic range of the sensor. The theoretical thermal noise of a 10 G Ω resistor at room temperature is 12.65 $\mu\text{V}/\sqrt{\text{Hz}}$, which is very high considering that the input referred voltage noise of the instrumentation amplifier is 28 nV/ $\sqrt{\text{Hz}}$ [10].

4.4.1 Input Filter

To characterize the input filter, the same setup was used as in the measurement with the 10 M Ω input resistor in section 4.3.1. The output was measured at the instrumentation amplifier guard pin and then used to compute and plot the transfer function.

Figure 4.9. Transfer function of the high pass filter at the input of the instrumentation amplifier. This shows a -3 dB cut off frequency around 0.1 Hz.

The measured transfer function is shown in Figure 4.9. It was difficult to measure the signal at very low frequencies; because of the time needed for sampling

the signal as well as the function generator not being able to generate low frequency signals. Despite this, through simple extrapolation, one can conclude from the figure that the -3 dB frequency of the filter is around 0.1 Hz. This cut off frequency is low enough to let through the signal we are interested in.

Also noticeable in Figure 4.9 is a constant loss of about 0.4 dB ($\approx 4\%$ of the original signal). However, this is not a significant amount, and is most likely due to voltage division between the sensor capacitance and the input capacitance of the instrumentation amplifier.

An attempt to estimate the input capacitance in the same way as in the 10 M Ω measurement is difficult; mainly because the tolerance of the 10 G Ω resistor was $\pm 30\%$. Assuming a cut off frequency of 0.1 Hz, the capacitance would then be somewhere in the range of 122-227 pF. The lower part of this range fits well with the value computed in section 4.3.1; thus it is fairly safe to say that the capacitance lies in that area.

4.4.2 Transfer Characteristic

The setup for this measurement was the same as for the one in section 4.3.2. The transfer function resulting from the measurement is seen in Figure 4.10. An immediate difference when compared to Figure 4.6 is the depth of the notch at 50 Hz. This is just a result of it being difficult to adjust the input signal to an exact frequency, the notch filter circuit is the same, and as such; the actual depth of the notch should be the same for both measurements.

Figure 4.10. Transfer function of the complete circuit with a 10 G Ω bias resistor. The -3 dB cut-off frequencies are shown by the dashed lines. The maximum passband gain is 58.3 dB.

In Figure 4.10 the dashed line represents the cut off frequency of the circuit. The cut-off frequency is at 3.1 Hz, ten times lower than for the one with the 10 M Ω bias resistor, however, that circuit did not reach the same gain at that frequency. Moreover, the figure shows that the mid-band gain reaches almost up to the 60 dB that the circuit is configured to produce. With amplification of

over 40 dB at 1 Hz, this circuit configuration shows promise for usefulness – the previous circuit had an amplification of -10 dB at this frequency.

4.4.3 Low Level Signal Response

In the same manner as the measurement in section 4.3.3, a low level input signal measurement was made with this circuit configuration as well. To be able to better compare the two circuits, the same input signal was used – 25 μV at 10 Hz, the lowest amplitude possible to generate with the function generator and attenuator connected together.

Figure 4.11. Output when a low level input signal is present.

Figure 4.11 shows the signal as it was output by the circuit, Figure 4.11(a), and its frequency spectrum, Figure 4.11(b). The wanted signal is almost completely covered in the 50 Hz noise – as can be seen in the frequency spectrum, the 50 Hz spur is significantly higher than the 10 Hz one – and is quite unusable in this state. In the time domain it is possible to distinguish the wanted signal as amplitude modulated onto the 50 Hz signal. While in the frequency domain, the signal is clearly visible, the visual time domain analysis needs a better signal.

After digital filtering, the 50 Hz and 150 Hz noise is removed and the results are shown in Figure 4.12. Comparing the frequency spectrum in Figure 4.12(b) with the one in Figure 4.8(b) shows that the amplitude of the output signal is now approximately 20 dB higher. This gain in amplitude makes the 10 Hz signal clearly visible in Figure 4.12(a). Despite this, there is however still a significant amount of noise left that distorts the sinusoidal signal. The signal-to-noise ratio (SNR) of the signal was computed to 9.7 dB.

The power consumption for this measurement is shown in Table 4.2. The same thing applies here as for the previous measurement: there is a large room for improvement in consumed power by redesigning the notch filters to utilize lower

Figure 4.12. Output of low level signal after digital filtering in MATLAB®.

Circuit part	Voltage	Current	Power
Instrumentation amplifier	10 V	1.02 mA	10.3 mW
Notch filters	10 V	10.524 mA	105.24 mW
Operational amplifier	3.145 V	0.125 mA	0.39 mW
Total			115.83 mW

Table 4.2. Supply voltage, current and power consumption of the different parts of the sensor readout circuit with a 10 G Ω resistor. The notch filters entry here includes both filters and thus four operational amplifiers, which is why the number is so large.

power operational amplifiers as well as using amplifiers that can run on lower supply voltage.

4.5 Conclusions

From these measurements, a few conclusions can be made. First of all, a very high input impedance of the circuit is necessary. Without this, low frequency signals will be suppressed to a point at which they can not be detected in the output. This impedance needs to be at least 10 G Ω ; while the 10 G Ω impedance worked out quite well in this measurement, one has to remember that this was with two capacitor plates taped together, separated only by their two layers of solder mask. In an application, the distance between the plates may be greater and the dielectric coefficient of the isolating material(s) may differ. Taking this into account, the cut-off frequency at the input will most likely rise.

Secondly, the size of the capacitance made up by the two PCBs is estimated to be around 125 pF. Increasing the input impedance further would mean that this capacitance could be lowered and still be able to pick up frequencies low enough

for the application. Being able to have a lower capacitance means that the area of the capacitor plate could be made smaller – a significant advantage especially for ECG where high spatial resolution is desirable.

After digital filtering, the circuit achieves an SNR of 9.7 dB at an input amplitude of $25\ \mu\text{V}$. However, the noise here consists of not only the circuit noise; but also the noise produced by the oscilloscope, the quantization noise as well as any aliased noise from the analog-to-digital conversion. Therefore, the SNR could be improved by including a high resolution ADC directly on the PCB as well as a more aggressive anti-aliasing filter before it.

The power consumption for both circuits is very high because of the urgently needed design of notch filters to remove unexpected interference. Including the notch filters on one of the PCBs as well as designing them for low power with the use of different operational amplifiers, would surely reduce the power consumption significantly.

Chapter 5

Summary

The goal of this thesis was to design and evaluate a capacitively coupled sensor and readout circuit that could be used to measure very low-frequency, low-level signals as required by electrocardiography (ECG) and electroencephalography (EEG) measurements without needing direct contact to the skin.

The properties of these signals, that put the requirements on the circuit are listed in Table 5.1. These numbers show that measuring such signals requires a very low noise circuit, and at those frequencies, it is a great design challenge.

Characteristic	ECG	EEG
Bandwidth	0–30 Hz	0–40 Hz
Signal level	0.05–1.5 mV	from 10 μ V

Table 5.1. Signal levels and bandwidths of ECG and EEG signals.

To measure these signals, two main components had to be designed; one sensor that senses the signals; and a readout circuit that amplifies and conditions the signals.

A sensor was created by using the bottom layer of a printed circuit board (PCB) as a capacitor plate and placing it against the signal source, which acts as the opposite capacitor plate. The solder mask of the PCB and the air between the two acts as an insulator and completes the capacitive sensor.

The readout circuitry was placed on the other side of this PCB as well as on another PCB – due to the size of the circuit. This created a compact sensor with the area of 25x25mm.

The readout circuit consisted of two different stages. The first stage provided 50 times amplification, and was realized with an instrumentation amplifier. Due to the capacitance at the input created by the sensor capacitor, the input bias resistor of the instrumentation amplifier had to be very high in order not to suppress the low frequency signals sought to be measured. The second stage provided 20 times gain and band-pass filtering of the signal, limiting it to the interesting bandwidth.

Later, during measurement, it was discovered that additional filtering was

needed in order to suppress interfering noise at specific frequencies. These were designed on provisional stripboards and were inserted in between the two PCBs.

Measurements were made with two different circuit configurations. The difference between these configurations was the size of the bias resistor at the input of the first stage instrumentation amplifier; sizes 10 M Ω and 10 G Ω were used for this purpose.

During initial measurements, high-level noise in the air at 50 Hz and 150 Hz originating from the main power lines was observed. This noise was in band and saturated the readout circuit output, and was therefore removed by notch filtering at these specific frequencies on a separate circuit board as mentioned above. Moreover, further filtering at these frequencies and additional low-pass filtering to reduce high frequency noise was made digitally with a computer after the collection of measurement data. The results from the measurements are summarized in Table 5.2.

Metric	10 M Ω bias	10 G Ω bias
Input filter -3 dB cutoff	128 Hz	$\approx$ 0.1 Hz
System bandwidth (-3 dB)	31–115 Hz	3.1–43 Hz
System passband gain	49 dB	58 dB
SNR at 10 Hz and 25 μ V input	—	9.7 dB
Power consumption	117 mW	116 mW

Table 5.2. Measurement results and performance of the circuit.

As can be seen from the results, with 10 M Ω biasing resistance the circuit transfer function is not satisfactory. The lower end of the bandwidth is 31 Hz, which is above the highest frequencies of the wanted signals. This of course means that these signals will not be amplified enough and that the difference in amplification will be very large between different frequency components in these signals. Furthermore, the maximum gain is significantly lower than the gain that the circuit was designed for, 49 dB instead of 60 dB, and the peak gain is at 70–80 Hz – well above the desired signal bandwidth.

On the other hand, with a 10 G Ω biasing resistor, the circuit performance is greatly increased. With a bandwidth of approximately 3–40 Hz, it will certainly be able to amplify both ECG and EEG signals (compare with Table 5.1). The gain also reaches a value of 58 dB, which is very close to the 60 dB the circuit was designed for. For a low-level, low-frequency input signal at 25 μ V and 10 Hz, the circuit achieved a signal-to-noise ratio (SNR) of almost 10 dB – after digital filtering. This number could possibly be increased with better a/d-conversion and more aggressive anti-alias filtering.

The conclusion of this investigative thesis is that it is possible to measure the low-level, low-frequency signals required for non-contact ECG and EEG monitoring capacitively with a simple PCB sensor construction. However, it is not without its problems. The two most significant issues of the techniques used in this thesis are, 1) that the input resistor of the readout circuit needs to be very high to be able to sense low frequency signals; and 2) that large amounts of noise from the

main power lines are picked up by the sensor and amplified by the circuit.

5.1 Future Work

This section lists things that the author feels could benefit from further investigation to improve the performance of the circuit.

- On-board a/d-conversion. Because of the limited resolution of the oscilloscope used for measurement, a significant amount of quantization noise is introduced into the output signal. With an on-board, high resolution a/d-converter, this noise could be reduced, producing an output signal with higher SNRs.
- Blind/buried vias on the PCB. Because of limitations in the PCB layout software, all vias of the sensor PCB punched holes through all the layers of the board, effectively perforating the sensor metal layer as well as the shield layer. These might cause extra interference on the sensor plate and the signal could be improved by stopping unnecessary vias from going through all layers.
- On-board notch filters. Since the need for the two notch filters was realized after the manufacturing of the printed circuit boards, they were not included on them and had to be built on a separate prototyping stripboard. This made the whole system, very bulky and makes it difficult to properly fix the sensor to the body in the case of an ECG or EEG measurement. With a redesign of the circuit, this should be taken into account, and the notch filters should be included on one of the PCBs – as a suggestion, on the back side of the second board.
- Other input biasing techniques. As have been seen in the measurements, the higher the input impedance of the instrumentation amplifier the better the circuit works. Raising the impedance further would create opportunities to reduce the area of the capacitive sensor while maintaining signal quality. However, simply using a higher value resistor might not be preferable since it increases the thermal noise of the resistor. Instead, other techniques for high impedance biasing, such as that in [5], should be investigated.
- Substitute the instrumentation amplifier for an operational amplifier. Instrumentation amplifiers are designed to amplify differential signals. In this case it is used with a single ended signal and the negative input is connected to the output through a low-pass filter, making it track the positive input. Since the instrumentation amplifier is built up of three operational amplifiers and two buffers and the same functionality could be implemented with a single operational amplifier, it is reasonable to believe that such an arrangement would reduce the amount of noise contributed by this first stage.

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Appendix A

MATLAB® Code Listings

A.1 Notch Filter

```
1 function Hd = notchfilter(sampleFreq, f0)
% NOTCHFILTER Returns a discrete-time filter object.

%
5 % M-File generated by MATLAB(R) 7.7 and the Signal Processing
% Toolbox 6.10.
% Modified to support input arguments for variable notch frequencies.
%
% Generated on: 14-Jan-2010 10:48:22
10 % Butterworth Bandstop filter designed using the BUTTER function.

% All frequency values are in Hz.
Fs = sampleFreq; % Sampling Frequency
15 Fpass1 = f0 - 10; % First Passband Frequency
Fstop1 = f0 - 1; % First Stopband Frequency
Fstop2 = f0 + 1; % Second Stopband Frequency
Fpass2 = f0 + 10; % Second Passband Frequency
20 Apass1 = 0.5; % First Passband Ripple (dB)
Astop = 30; % Stopband Attenuation (dB)
Apass2 = 1; % Second Passband Ripple (dB)

% Calculate the order from the parameters using BUTTORD.
25 [N,Fc] = buttord([Fpass1 Fpass2]/(Fs/2), [Fstop1 Fstop2]/(Fs/2), ...
min(Apass1, Apass2), Astop);

% Calculate the zpk values using the BUTTER function.
[z,p,k] = butter(N, Fc, 'stop');
30 % To avoid round-off errors, do not use the transfer function.
% Instead get the zpk representation and convert it to
% second-order sections.
[sos_var,g] = zp2sos(z, p, k);
35 Hd = dfilt.df2sos(sos_var, g);
% [EOF]
```

A.2 Low Pass Filter

```
1  function Hd = lpfilter(sampleFreq)
   % LPFILTER Returns a discrete-time filter object.

   %
5  % M-File generated by MATLAB(R) 7.7 and the Signal Processing
   % Toolbox 6.10.
   % Modified to support variable sampling frequencies.
   %
   % Generated on: 14-Jan-2010 11:08:31
10 %

   % Butterworth Lowpass filter designed using the BUTTER function.

   % All frequency values are in Hz.
15 Fs = sampleFreq; % Sampling Frequency

   Fpass = 100;      % Passband Frequency
   Fstop = 400;      % Stopband Frequency
   Apass = 1;        % Passband Ripple (dB)
20 Astop = 60;       % Stopband Attenuation (dB)

   % Calculate the order from the parameters using BUTTORD.
   [N,Fc] = buttord(Fpass/(Fs/2), Fstop/(Fs/2), Apass, Astop);

25 % Calculate the zpk values using the BUTTER function.
   [z,p,k] = butter(N, Fc);

   % To avoid round-off errors, do not use the transfer function.
   % Instead get the zpk representation and convert it to
30 % second-order sections.
   [sos_var,g] = zp2sos(z, p, k);
   Hd          = dfilt.df2sos(sos_var, g);
   % [EOF]
```

Appendix B

Schematics

This appendix shows the schematics created with the schematic capture software *gschem* to create the PCBs.

B.1 Schematics

Appendix C

PCBs

This appendix shows masks for the different PCB layers at scale 1:1. Only masks with relevant information are shown; empty masks such as the back layer solder mask and silkscreen have been left out since they are empty and full, respectively.

C.1 PCB Layers

C.1.1 Sensor PCB

Front Layer

Metal

Silkscreen

Solder mask

Buried Layers

Layer 2 metal

Layer 3 metal

Back Layer

Metal

Plated drill

C.1.2 Signal Conditioning PCB

Front Layer

Metal

Silkscreen

Solder mask

Back Layer

Metal

Plated drill