Wireless Power Transfer and Optogenetic Stimulation of Freely Moving Rodents

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Wireless Power Transfer and Optogenetic Stimulation of Freely Moving Rodents

THESIS FOR THE DEGREE OF MASTER OF SCIENCE

in

BIOMEDICAL ENGINEERING

Biomedical Electronics

by

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Abstract:

Animal studies are commonly used to test the feasibility and effectiveness of promising novel neuroscience research ideas. One such new technique is optogenetic stimulation, a state-of-the-art brain stimulation technique. In optogenetics, genetic techniques are used to create light-sensitive proteins within the neuron membrane, thus allowing the affected region to become sensitive to light stimulation, for example through an inserted LED.

Current optogenetic stimulation methods use tethered setups and, typically, the animal-under-study is put into a fixed position. This introduces stress, which, besides an obvious reduction in animal welfare, may also influence the experimental results. Hence, an untethered setup is highly desirable. Therefore, in this study, we propose a wireless optogenetic stimulation setup, which allows for full freedom of movement of multiple rodents-under-study in a 40 cm $\times$ 40 cm $\times$ 20 cm environment.

We investigate a variety of wireless power transfer methods, which results in the choice for wireless power transfer through inductive coupling, as this allows for efficient power transfer over short range and has the least side-effects, making it the most suitable approach for this particular environment. The efficiency of inductive coupling is highly susceptible to vertical, lateral and angular misalignment of the coils. The wireless link is, therefore, designed to maximize the link efficiency and minimize the misalignments between the coils. In order to maximize the inductive power transfer link, we look into all the aspects that have an influence on the link efficiency, including coil shape and coil material.

The implementation of the wireless optogenetics setup is divided into three parts:

**Transmitter Coil:** The design of a transmitter coil capable of providing sufficient link efficiency throughout the entire 40 cm $\times$ 40 cm $\times$ 20 cm region of interest, in order to be able to power the optogenetic stimulation receiver, regardless of lateral and vertical misalignments.

**Receiver Module:** The design of a receiver module that resides on the animal and, as such, is severely restricted in both size and weight. The complete module with receiver coil, rectifying and regulating electronics and microcontroller can occupy at most 1 cm $\times$ 1 cm $\times$ 1 cm and weighs below 1 g. Moreover, half of allowable volume of the receiver module is kept unused for the future assembly including the wireless ECoG recording electronics.

**Optogenetics Optrodes:** The creation of optogenetics optrodes using a novel $\mu$LLED mounting technique, which allows for the $\mu$LLED array with multiple LEDs to be directly inserted into the brain. The use of a $\mu$LLED array greatly improves the power efficiency, as the traditional LED-to-optical-fiber coupling is accompanied by large losses in light-intensity. Moreover, a single optrode is able to replace a number of optical fibers, resulting in a less-invasive procedure if multiple stimulation sites are required.

For an input current of 0.5 A into the primary coil, an average inductive link efficiency of 0.28 %, and an angular misalignment of 45 degrees, the final setup is capable of delivering at least 8.5 mW of light power into the brain.

**Keywords:** Brain Stimulation, Epilepsy, Inductive Coupling, Optogenetics, Optrodes, Wireless Power Transfer
ACKNOWLEDGMENTS

This Master thesis presents months of work on my master final project in Biomedical Engineering, direction Biomedical Electronics at Delft University of Technology. The project is the result of the collaboration between the Bioelectronics department of TU Delft and the Department of Neuroscience of Erasmus Medical Center.

I would like to very much thank my supervisors Prof. Wouter Serdijn and Dr. Freek Hoebeek for creating the opportunity of this project for me. I feel very lucky for being a part of this work, for not only have I learned much during this period, I have also met amazing people at the Bioelectronics group of TU Delft, as well as at the Neuroscience group of Erasmus MC.

Furthermore, I would like to sincerely thank our great technician, Wil Straver, for supervising me throughout the work of this thesis. I am very glad I had the opportunity of learning from you. Thank you so much for your help, valuable input and creativity, for teaching me all I know now about the work in the electronics lab, and for the ‘gezelligheid’.

Farnaz Nassirinia
Delft, The Netherlands
November 24, 2016
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INTRODUCTION

The brain is the mystery of the human body. Neurons, as primary units of the nervous system, are joined together into a complicated biological interconnected network. A conventional method to manipulate the neural performance within this network is to use drugs that alter the chemical balance of the brain. However, a crucial aspect of the nervous system is the electrical signalling between the neurons.

Bioelectronics has advanced the neural modulation techniques beyond the conventional methods by developing electrical brain stimulation tools. Electrical brain stimulation is truly beneficial in helping to understand the mechanisms underlying neural behaviour, and to develop novel therapeutic methods. Optogenetics is another breakthrough method in neural stimulation techniques, which has opened up entirely new avenues of research opportunities in the fields of neuroscience and bioelectronics.

The structural and functional complexity of a neural network creates countless biological, diagnostic and therapeutic unknowns. Epilepsy is one such neural disorder, which, despite being widespread, is not entirely understood. Epilepsy exposes itself in the form of seizures. The joint effort of the Bioelectronics group at TU Delft and the Neuroscience department at Erasmus MC has resulted in the development of a closed-loop system to detect and suppress epileptic seizures.

In order to stop epileptic seizures, there are two main challenges to overcome: first, finding an early and accurate method for detection of the seizure; and second, developing an efficient and safe method for suppression of the seizure. The closed-loop system offers detection and suppression of certain types of epileptic attacks. Its necessity of using wires, however, limits the range of its applicability. A tethered setup prevents the experiments from running on freely moving objects and, as such, limits the types and settings in which experiments can be performed. Therefore, developing a wireless neural recording and stimulation system is required to circumvent these issues. This idea is illustrated in Figure 1.1, where multiple rodents can freely move in an environment that provides power for recording and stimulation to the receiver module residing on their body.

![Figure 1.1: Visualisation of the wireless neural recording and stimulation setup.](image)

Development of a wireless neural recording and stimulation setup is divided into: the design of recording electrodes, establishing a wireless data transmission link, developing a wireless power transfer link and building stimulation optrodes. MSc students Mattijs Weskin and Ide Swager have dedicated their final projects to the recording electronics and wireless communication for transmission of neural recorded data from multiple mice. This thesis presents the design of the wireless power transfer setup, and the development of optogenetics stimulation optrodes. It also introduces two algorithms for detection of the seizure. The combined achievements of these bioelectronics projects aim to create a complete multi-channel, multi-object, closed-loop wireless neural recording and stimulation module. The main building blocks of the complete module is shown in Figure 1.2, in which the green blocks are developed during the work of this thesis.
The remainder of this thesis is organized as follows. Chapter two introduces two algorithms for the detection of epileptic seizures. In Chapter three, different neural stimulation methods are discussed and our choice for optogenetic stimulation is motivated. In chapter four, our choice for the wireless power transfer method is made. Inductive coupling is found to be most effective here. Chapter four also includes the parameters that affect the coil design, as well as the design of the transmitter coil. Chapter five includes the design of the receiver coil, the power module, optical module, the control module, and the assembly of the head-mounted receiver module. The overall wireless link efficiency of power transfer and power conversion is investigated in Chapter six. This thesis ends with the conclusions and future work as explained in Chapter seven.
Epilepsy

Epilepsy is a chronic disorder, characterised by epileptic seizures, which affects one percent of the world’s population. During a seizure, cells signal abnormally and cause symptoms such as loss of consciousness and muscle rigidity. Hence, it is crucial to be able to detect epileptic seizures and stop them at an early stage. In this chapter, we investigate two detection methods for a severe type of epileptic seizures, called grand-mal seizures.

2.1 Grand-Mal Seizures

In a healthy brain, neurons fire with low probability of synchronicity. Synchronicity in this context means firing at the same time instant. For someone with epilepsy, however, sometimes a group of neurons fires in an abnormally excessive and hyper-synchronous manner. This results in an abnormal electrical activity in the brain, which often appears as ictal sessions called seizures.

Seizures are classified into two main groups, based on the area of the brain they encounter. The first group of seizures, called partial seizures, arise from a localized region in the brain. Generalized seizures, in contrast, target the entire brain and arise from both hemispheres. The partial and generalised seizures include many different subtypes, based on the duration, symptoms and wave-characteristics of the seizure [15]. The focus of this thesis is only on two types of generalised seizures: petit-mal seizures (also known as absence seizures), and grand-mal seizures (also known as tonic-clonic seizures).

Grand-mal seizures often coincide with repetitive, jerking movements, convulsions, loss of consciousness and loss of memory. Repeated occurrence of seizures can result in serious brain injuries. Furthermore, seizures may result in life-threatening situations, for example, when the loss of balance causes someone to fall resulting in head trauma. Therefore, it is imperative to develop a system to recognise and stop these seizures at an early stage.

In order to stop epileptic seizures, first we must find an early and accurate method for detection of the seizure. A closed-loop system, developed by TU Delft, is in use at the Neuroscience lab of Erasmus MC [37]. The closed-loop provides detection and suppression of seizures. On the detection side, a low-cost single-board ARM-based computer called BeagleBone is programmed with a wavelet-based filter, suitable for real-time detection of absence seizures. To test this, a setup has been created to perform studies on mice that are afflicted with epileptic attacks. These mice are genetically modified to include light-sensitive proteins. As a result, their brain can be stimulated through optogenetics.

The wavelet-based algorithm for the detection of absence seizures is not capable of detecting the other type of generalised seizure, the grand-mal seizure. Grand-mal seizures, or generalised tonic-clonic seizures, include, as the name suggests, a tonic phase and a clonic phase (inset a in Figure 2.1). The tonic phase of grand-mal seizures are characterised by tonic spike and wave discharges (inset b in Figure 2.1). Behaviourally, this appears mainly as loss of consciousness and muscle-movement stiffness. The clonic phase, however, is accompanied by very savage muscular contractions. This manifests itself in the ECoG as a high power, high frequency firing signal (inset c in Figure 2.1).
In this thesis, two algorithms for the detection of grand-mal seizures are developed, one of which was found to be also accurate for the detection of absence seizures. These algorithms are discussed in the following section.

### 2.2 Detection of Grand-Mal Seizures

The sudden abnormal signalling in the brain appears as seizures. Hence, by observing the behaviour of a mouse during the recording, the on-set time of the seizure is defined. The brain abnormalities during a seizure session can also be observed by analyzing the recordings of the electrical activity of the brain. An electrocorticography, or ECoG, is a recording method in which multiple electrodes are placed on the exposed surface of the brain and record the electrical activity from the cerebral cortex. The ECoG during a grand-mal or tonic-clonic seizure arises with high frequency and high power signals compared to the normal ECoG. Figure 2.2 illustrates the ECoG recorded from three electrodes. At time stamp $8.229 \times 10^6$ s, the seizure begins. The grand-mal seizures are also generalized seizures that affect a large area of the brain. Therefore, they can be present in the ECoG recordings from different sites of the cortex. Each of these recording sites is called a channel. During a tonic-clonic session, the electrical activity recorded in different channels starts to become phase-synchronised. This is the basic idea used for the development of the first algorithm. In the same figure, the synchronised windows between the three channels is shown. We can see that at the onset of the seizure, the cyclic signals in the three channels tend to oscillate.

The first algorithm is based on the Montez synchronization-likelihood with explicit time-frequency priors [23]. Please see Figure 2.3a and 2.3b. In the figure, the two recording channels are named Ch.A and Ch.B. This method is able to detect for each channel if there is some pattern that repeats itself. Then, it checks if these repetitions (of potentially different patterns) occur at the same time for different channels. This is called generalized synchronization. It works by creating a state vector $X_i$ containing a number of samples within a small time window $W_1$. Then, this reference vector is compared to the other state vectors within window $W_2$. Similar vectors that are within a critical distance are called recurrences. In the figure, these are colored white. These calculations are performed for each channel separately. Then, the synchronization likelihood is calculated as being the number of simultaneous repetitions between channels divided by the total number of recurrences within the channels.

The critical distance is also visualized in Figure 2.3c. State vectors within the critical distance fall within the white ellipse. The critical distance is calculated dynamically, per channel, and is chosen in
2.2. Detection of Grand-Mal Seizures

Figure 2.2: An ECoG recorded from three locations on the cortex. The red and green windows illustrate the synchronised phases due to onset of the seizure (recorded at the neuroscience department of Erasmus MC).

such a manner to ensure that a fixed number of recurrences are included. It can be observed that the critical distance for Channel B is slightly smaller than the critical distance for Channel A.

Figure 2.3: The Montez Synchronization-likelihood method (image courtesy of [23]). On the left, the time windows $W_1$ and $W_2$ is shown. On the right, the critical distance is illustrated.

In such a way, the algorithm calculates the synchronicity-likelihood between the channels and plots it versus time. When this probability breaches the threshold (obtained empirically by analysing many hours of ECoG data) for at least two seconds, a seizure is detected. Figure 2.4 shows the ECoG of two channels during a seizure. Figure 2.5 illustrates the corresponding output of the algorithm, where the synchronicity-likelihood is remarkably increased at the seizure. When we set the threshold of synchronicity to 0.4, we are able to detect the seizures within a maximum of four seconds of delay (within the tonic phase) and with a 100 percent Average-Detection-Rate (ADR) [34].

This algorithm has been applied to a random selection of twenty seizures. The result in the form of a heat plot is shown in Figure 2.6. Each horizontal line determines a window of the ECoG, which contains a grand-mal seizure. The onset of the seizures are determined by observation of the mice behaviour. The onset then is set to 55, so that values below 55 on the horizontal line determine the pre-ictal level of synchronisation. The warmer the colour of the heat bars, the higher the synchronisation likelihood between the two ECoG channels.

As we can see in Figure 2.6, this algorithm shows synchronised activity before the onset of the seizure in 70 percent of the ECoGs. On the one hand, using this result, we can even be able to detect the grand-mal seizures before they actually occur. On the other hand, this pre-ictal synchronisation was
2. Epilepsy

Figure 2.4: An ECoG recording from two channels during the seizure (recorded at the neuroscience department of Erasmus MC).

Figure 2.5: Synchronicity likelihood obtained from the algorithm output corresponding to the seizure of Figure 2.4.

Figure 2.6: Synchronicity likelihood obtained from the algorithm output for twenty seizures. The blue line determines the onset of the seizures, empirically determined by the observation of mice behaviour.

not observed in six cases as can be seen from Figure 2.6. Therefore, this method, while providing a 100 percent specificity (Specificity = True negative/(True negative + False positive) \times 100\ percent), will not offer a full sensitivity (Sensitivity = True positive/(True positive + False negative) \times 100\ percent), resulting in a low ADR. The combination of the two detection methods, however, using pre-ictal synchronisation and ictal synchronisation, could potentially be an interesting seizure detection method. In this case, the preference is on the detection of the seizure before the onset of it, and if not detected,
2.2. Detection of Grand-Mal Seizures

to use the ictal synchronisation information. The synchronisation studies of generalised seizures can provide much information about this type of seizures, which requires more in-depth analysis that is beyond the scope of this thesis. We can also observe in Figure 2.6 that empirical observation of mice behaviour is a poor method for determination the onset of the seizures. In most of the analysed seizures, the high synchronicity occurs before the observed onset of the seizure.

The second algorithm is based on the already-existing Morlet wavelet-based algorithm, which is in use for the detection of absence seizures [37]. The idea behind this method was to adapt the already-existing algorithm so that it will detect both absence and grand-mal seizures. Figure 2.7 shows the incapability of the wavelet filter for detection of the grand-mal seizures, as the detection delay is too high. The algorithm is applied to a random selection of twenty seizures. It detects the seizures with 100 percent sensitivity with an average delay of 51 seconds. This delay is too large, since the seizure is already at the clonic phase. If we lower the threshold, the detection delay decreases. However, the specificity drops as well (see Figure 2.9 inset a,b).

![Figure 2.7: ECoG and the data output from the Morlet wavelet-based algorithm. The sampling frequency is 500 Hz.](image)

Therefore, another condition is added to this algorithm to create a multiple event detection mechanism, in order to improve its specificity. For improved seizure detection, the threshold method is augmented by a state-machine with three states (refer to Figure 2.8). This state machine improves the seizure detection by requiring that a number of threshold breaches occur within a small time interval before detecting an actual seizure, thus guarding against sporadic spikes. Initially, we are in the 'No Seizure Detected' state. When a threshold breach occurs, we move to the 'Potential Seizure Detected' state. If no additional threshold breaches occurs within a certain time interval, called MaxInterval, we return to the 'No Seizure Detected' state. Otherwise, if additional threshold breaches do occur, the threshold breach count is increased, until at least N threshold breaches have occurred. Then, we move to the 'Seizure Detected' state. We remain in this state until no threshold breaches occur for at least the MaxInterval time interval.

The optimum N value, MaxInterval and threshold level are set to 10, 100, 2e7, respectively. These values were obtained by training the algorithm on a series of a hundred seizures. The addition of the multiple event detection mechanism resulted in greatly improved detection capability of the algorithm. As there is always a trade-off to be made between detection delay and detection capability, for a five seconds detection delay, the algorithm is able to detect grand-mal seizures with a 100 percent
Figure 2.8: The state machine that was added to the Morlet-based algorithm.

Figure 2.9: Applying the Morlet-based algorithm to grand-mal seizures. The red lines indicate the algorithm has detected a seizure. In this figure: a) ECoG data, b) output of Morlet-wavelet filter, c) count of threshold breaches, d) output of improved-wavelet algorithm.

ADR (see Figure 2.9c,d). Moreover, this algorithm can also be used to detect both absence and grand-mal seizures with 100 percent ADR. Achieving this, however, increases the delay time for both types of detection. Absence seizure can be detected within 0.5 seconds and grand-mal seizures within 7 seconds. The best-fit N value, Maxinterval and threshold in this case are set to 6, 130, 1e7 respectively.

In the real-time suppression of epilepsy, there are two factors of importance: the actual detection latency (after how many samples can the algorithm determine if a seizure is happening), and the computational complexity of the chosen algorithm. As we are dealing with real-time detection and suppression, the available computational resources are limited. For example, although the synchronization likelihood-based algorithm yields superior detection and provides much insight into the nature of the generalized seizures, its computational complexity makes it less suited for application in a real-time environment. The current synchronization likelihood algorithm requires about three hours to compute the synchronicity for a two-channel recording of two hours. Optimizations could potentially greatly improve the performance of the algorithm, however, this would require further development.
In this chapter, two methods for automatic detection of grand-mal seizures were proposed, as empirical observations of the mice behaviour is a time-consuming and error-prone method for determining the onset and end of the seizures. There is a trade-off to be made between detection sensitivity and detection delay. The computational complexity is as important as a low detection delay when creating a real-time seizure detection algorithm, as the run-time of the algorithm should be feasible for the constrained environments of a real-time system. Once a seizure is detected, it needs to be brought to a halt, which can be achieved through stimulation of the brain. This is the topic of the next chapter.
Modulation of Neural Network

The nervous system, as a complex network of nerves and cells, receives sensory information at its central element, the brain, and sends messages from the center towards the rest of the body. The information through this neural network is chemically and electrically transferred. In an unhealthy brain, the neural signalling mechanism malfunctions. In order to enhance the performance of a malfunctioning nervous system, one must modulate the neural network. Manipulation of the neural signaling within the neural network can occur through direct distortion of the chemical and electrical activity of the brain. In order to influence the chemical performance of the neural signalling, drugs can be applied to the brain. This method, however, lacks brain target specificity, and does not always yield sufficient results.

Electrical stimulation is another direct method of neural manipulation, which is able to distort the electrical balance of the brain. Electrical brain stimulation occurs by applying electrical voltage across the tissue (voltage stimulation), or by releasing electrical charges into the tissue (current stimulation). Electrical stimulation of the brain can also occur indirectly. Indirect approach first affects the non-electrical properties of the neural tissue, which next results in the change in the electrical balance of the neurons. These properties give neurons a completely different role, such as: neural tissue as electro-magnetic receiver, neurons as temperature sensors, neurons as optic sensors or neurons as acoustic receivers. In this chapter, the methods for brain stimulation are discussed in more detail.

3.1 Electrical Brain Stimulation

Deep brain stimulation (DBS) is the most widely used electrical stimulation technique in research and in clinical neurobiology. Deep brain stimulation involves implanting electrodes within certain areas of the brain. DBS manipulates the electrical performance of the neural network by applying a voltage difference over the neural tissue (voltage stimulation), or by releasing electrical charges into the tissue (current stimulation). There are risks associated with voltage-based stimulation, making it potentially unsafe, as the tissue impedance in the stimulation pathway is unknown. Hence, the applied voltage does not indicate the amount of charge released into the tissue. Therefore, current-based stimulation is more favored.

The excitation strength is proportional to the polarity of the stimulation, to the tissue conductivity and to the charge density. Charge density refers to the amount of electrical charge released into the tissue per unit time and per unit cross-section of the stimulator electrode. Cathodic stimulation depolarises neurons easier than anodic stimulation [19]. Axons with larger diameter excite easier than the thinner axons due to higher conductivity [3]. At the electrode-tissue interface, the electrical current is converted into ionic current. This interface has a non-linear impedance characteristic. The bigger the cross-section area of the electrodes, the lower the interface impedance, the larger the electrode capacitance, and the slower its polarization. The stimulation, therefore, requires more current. DBS does not offer cell-specific neural manipulation. It, however, provides local brain stimulation that can reach deep targets in the brain. Electrical brain stimulation can also be applied using electrodes,
which are placed over the scalp. This type of electrical stimulation is called transcranial direct-current stimulation or tDCS. TDCS application induces much pain in the patient and, therefore, it is rapidly becoming obsolete in research and clinical use[30].

### 3.2 Transcranial Magnetic Stimulation

Transcranial magnetic stimulation (TMS) is a non-invasive procedure for brain stimulation. TMS uses magnetostatic fields to stimulate nerve cells in the brain. An electromagnetic (EM) coil is placed near the area of the brain that is intended to be stimulated. The alternating current through the transmitter creates an alternating EM field. The magnetic field lines travel into the tissue and generate closed circles of an electric field $E$, which has the same shape as the transmitter coil. The tissue acts as a conductor. Hence, the field $E$ causes the flow of local current $i$ ($i = \rho \times E$) within the tissue. The tissue is a non-uniform conductor and, therefore, its impedance characteristics change along the pathway of the current flow. This causes the induction of electrostatic potential difference, which changes the trans-membrane potential. This change, if significant enough, will result in the local activation of the neurons [1]. The basic principle is very much the same for tDCS, where, instead of the application of a magnetic field, an electric field is applied close to the desired area of activation.

Both TMS and tDCS suffer from poor spatial resolution. Many techniques, such as field-shaping using multiple coils have not yet been successful enough to produce a local focused beam of stimulation. Furthermore, many research efforts have been dedicated to use TMS for stimulation of the deep brain structures. The idea here is to create a strong electrical field in the deep brain, while it remains a weak field in the more superficial layers [30]. Unfortunately, this was also not achievable for the following reasons. On the one hand, the head is an enclosed volume-conductor. At quasi-static frequencies, the electrical field is stronger on the boundary than in the inside of a volume-conductor with constant conductivity. Even if one considers the brain as a conductive volume with non-constant conductivity, the field is still stronger on the boundaries, since it is a spherically-symmetric conductor. If we consider the brain a non-symmetric, non-constant conductor, then it can indeed be possible to create an electrical field stronger in the inside of the volume (low-conductivity region) than on the surface.

However, despite the many research efforts involved, this assumption does not provide focal stimulation in deep tissue. The reason is that since such regions can pin the locus of the field maximum, smooth changing of the site of neuronal excitation is not possible [10]. Despite all of this, coil design capable of generating 3D TMS focusing to reach deep brain structures remains a great challenge in this field, as well as an interesting research topic.
3.3 Magneto-Thermal Stimulation

Magneto-thermal brain stimulation is another indirect method of brain stimulation that relies on the magnetic, thermal and genetics techniques. First, the cells are transfected with heat-sensitive proteins. Then, after the cells express the heat-sensitive receptors TRPV1, the magnetic nano-particles (MNPs) are inserted into those specific areas that are required to be stimulated. These particles, since they are magnetic, go through hysteresis in response to the applied alternating magnetic field, and, thereby, dissipate heat. When an alternating magnetic field (100 kHz to 1 MHz) is applied to the tissue, the MNPs generate heat and cause the tissue’s local temperature to increase. This activates the TRPV1 receptors. Hence, the trans-membrane potential increases and, therefore, the neuron fires [4]. The mechanism of the magneto-thermal stimulation is summarized in figure 3.1, assuming the cells are already expressing TRPV1.

![Figure 3.1: The magneto-thermal stimulation procedure.](image)

3.4 Optogenetics

Optogenetics is the combined technology of optics and genetics. Optogenetics controls the activity of the brain through application of light. The working principle is based on the creation of light-sensitive proteins within the neuron membrane. A piece of DNA that encodes the light-sensitive protein is placed into the neuron. The neuron unpacks the DNA and builds the proteins within its membrane. The activity of these neurons can now be controlled using pulses of light. Upon such pulses, the light-sensitive protein channels open and gate positively charged ions across the membrane of the cell. This results in an increase of the membrane potential and causes this one type of neuron to activate. The process is summarized in Figure 3.2.

![Figure 3.2: The optogenetics procedure.](image)

For example, Channelrhodopsin, or ChR2, light-gated channels are good receptors for wavelengths around 480 nm (peak reception), and, in response to the emission of light in this band, cause the neuron to fire. They, however, are neutral to the emission of light out of the visible blue spectrum. Figure 3.3 shows the stimulation of the ChR2-expressing neuron with blue light.
Another example is the light-sensitive receptor called Halorhodopsin. Halorhodopsins are Chloride pumps, which, upon emission of yellow light (peak absorbance of 570 nm), open up and hyper-polarise the cell, thus inhibiting the neural excitation. Hence, it is clear that the light-sensitive channels are wavelength and function-specific. Furthermore, optogenetics can be cell-specific. This means that upon emission of the light, only those cells that are transfected with the light-sensitive channels will become stimulated. This is a major difference as compared to electrical stimulation, where all the cells lying in the electrical stimulation pathway are affected. However, optogenetics, like electrical stimulation, is invasive and requires the insertion of the light source into the stimulation targeted brain area. Still, optogenetics remains a widely-used cell-specific brain stimulation technique. However, for now its applications are mostly on the experimental level and not much for clinical procedures.

Optogenetics-driven research has provided insight into many neurological and psychiatric disorders and applications, such as Parkinson’s disease, autism, schizophrenia, drug abuse, anxiety, and depression [39], [17], [33]. Another key application of optogenetics is in the suppression of epilepsy [37], the topic of this thesis.

3.5 Transcranial Pulsed Ultrasound Stimulation

Ultrasound refers to the sound waves at frequencies above 20 kHz [7], [35]. Transcranial pulsed ultrasound stimulation, or, in short, TPUS, is a non-invasive neuro-modulation method, which can stimulate the cortex and hippo-campus of the brain, without any need of an external agent, using ultra-sound waves. Tetrodotoxin, or TTX, is an effective neuro-toxin that binds to the voltage-gated membrane channels of the cell and inhibits the cell from firing action potentials. Research has shown that ultrasound can actually trigger the TTX-sensitive neural activity [36]. This occurs at low intensity ultrasound (< 500mW/cm²), where the stimulation is equithermal and any variation in the electrical activity of the brain is only due to the mechanical bio-effects. High intensity ultrasound (> 1W/cm²), however, has a more dominant effect on the temperature-related bio-effects than on the mechanical ones. In any case, ultrasound stimulation does not confer local and focused stimulation.
3.6. Conclusions

In this chapter, different neuromodulation techniques were studied. Table 3.1 summarizes the material explained in this chapter. The cell-specific, deep brain targeting capability is a crucial aspect for the neuroscience research topics that are planned to be investigated with the final wireless system. Furthermore, the risks associated with the Magneto-thermal stimulation techniques are not well-studied yet, in particular its effect of heating up tissue. Considering this, the implementation of the wireless stimulator in this project is based on the optogenetics method.

<table>
<thead>
<tr>
<th>Stimulation method</th>
<th>Cell Specific</th>
<th>Target Deep drain</th>
<th>Non-invasive implant</th>
<th>Safety</th>
<th>No exogenous Factors</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transcranial Magnetic Stimulation</td>
<td>-</td>
<td>-</td>
<td>+</td>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>Deep Brain Stimulation Current/Voltage</td>
<td>-</td>
<td>+</td>
<td>-</td>
<td>+/-</td>
<td>+</td>
</tr>
<tr>
<td>Optogenetics Stimulation</td>
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<td>+</td>
<td>-</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Magneto-thermal Stimulation</td>
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<td>+</td>
<td>+</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Acoustic Stimulation</td>
<td>-</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>+</td>
</tr>
</tbody>
</table>

Table 3.1: Review of stimulation methods and their characteristics.
Wireless Energy Transfer

In this chapter, we take a look at various methods for energy transfer. These methods are compared to one another, and their advantages and disadvantages are discussed, so that we are able to choose the optimal approach for the design of the energy transfer link of this project. A key design element arises from the fact that our application domain deals with animal studies. A tethered setup impacts an animal’s freedom of movement, which, in turn, causes distress. This may even influence an experiment’s results. Therefore, a wireless approach is highly preferred.

We conclude that the wireless inductive link is the best method for power transfer. In essence, the inductive link can be regarded as the transfer of energy from a transmitter coil to a receiver coil. Therefore, the coil design is an important aspect to consider in order to achieve the best link performance. A number of design aspects, such as material, shape, and more, are considered and the best choice for each of these aspects is determined. Finally, the transmitter coil is designed to oscillate at the resonance frequency, and provide a 40 cm × 40 cm × 20 cm environment for powering the receiver coil with low impact of lateral and vertical misalignments between the transmitter and the receiver coil.

4.1 Overview of Energy Transfer Methods

There are three fundamental methods for energy transfer: through a conduction link (e.g., a wire), through a propagating wave (e.g., a sound wave), and through a quasi-stationary field (e.g., inductive coupling). The traditional method for energy transfer is through the use of wires, allowing for convenient transport of energy through a point-to-point link. This wired link can be a plain conductive wire that transmits energy in the form of an electrical charge. This is called a conduction link. Such a link can also serve as a platform for the transportation of different forms of energy: for example, optical fibers in case of a wired optical link. Here, the energy is transported in the form of light.

In contrast to tethered energy transfer, wireless energy transfer (or WET), as its name suggests, is the transmission of power from a transmitter to a receiver without any physical conducting link between the two. Both energy transfer through a propagating wave and energy transfer through a quasi-stationary field fall into this category. Wave propagation energy transfer can be subdivided into three main types: the optical link, the acoustic link, and the radio-frequency (or RF) link.

Energy transportation through an optical link relies on the propagation of light waves, with frequencies ranging from 300 GHz to $10^{17}$ Hz. Any light source, like the beautiful sun, or even a lamp or an LED, can be considered in a way as an optical energy transmitter. Other, more practical examples include the infrared optical link of a TV remote control. In many technical applications, the light source is implemented as a laser, which emits a coherent beam of light [27]. Lasers are highly directional sources and, thus, are able to provide efficient power transfer capabilities over very long distances. For example, an orbital solar array could send back its harvested energy through a high-powered laser beam over thousands of kilometers. Any optical link, however, requires direct line-of-sight between transmitter and receiver.
Energy transport by means of an ultrasound link uses acoustic waves to transmit energy at frequencies beyond the audible range for humans, or above 22 kHz [11]. However, when compared to humans, mice and many other rodents have a more sensitive auditory system and are able to perceive frequencies far exceeding the human hearing range: mice are able to hear sounds with a frequency as high as 70 kHz. Unfortunately, ultrasound waves cause distress in mice and other rodents sensitive to high pitches. Moreover, ultrasound waves with frequencies higher than 40 kHz are even able to cause tissue damage [38].

Finally, power can be transmitted through a radio frequency link, with a propagating electromagnetic field. The oscillating electric and magnetic fields are related to each other by the famous Maxwell series of equations:

\[
\vec{\nabla} \times \vec{E} = -\frac{\delta \vec{B}}{\delta t} \tag{4.1}
\]
\[
\vec{\nabla} \times \vec{H} = -\frac{\delta \vec{D}}{\delta t} + \vec{J} \tag{4.2}
\]
\[
\vec{\nabla} \cdot \vec{D} = \rho \tag{4.3}
\]
\[
\vec{\nabla} \cdot \vec{B} = 0 \tag{4.4}
\]

Where \( E \) is the electric field strength, \( H \) the magnetic field strength, \( B \) magnetic flux density, \( D \) is the electrical displacement and \( \rho \) is the electric charge density. These formulae show the contribution of the oscillating \( H \) and \( E \) fields in the electromagnetic link. Hence, a magnetic disturbance is not possible without an electric one and vice versa. For a short wire carrying a current \( i = I \times \sin(\omega t) \), where the distance \( r \) from the wire is much larger than the carrier wave wavelength \( \lambda \), based on Maxwell equations in spherical coordinates \((r, \theta, \phi)\) we obtain:

\[
E_r = -\sqrt{\frac{\mu I l \cos \theta}{r \lambda}} \left[ -\frac{1}{4\pi^2} \frac{\lambda^2}{r^2} \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) + \frac{1}{2\pi} \frac{\lambda}{r} \sin \left( \frac{2\pi r}{\lambda} - \omega t \right) \right] \tag{4.5}
\]
\[
E_\theta = \sqrt{\frac{\mu I l \sin \theta}{2\pi r \lambda}} \left[ -\frac{1}{4\pi^2} \frac{\lambda^2}{r^2} \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) - \frac{1}{2\pi} \frac{\lambda}{r} \sin \left( \frac{2\pi r}{\lambda} - \omega t \right) \right. \\
+ \left. \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) \right] \tag{4.6}
\]
\[
E_\phi = H_r = H_\theta = 0 \tag{4.7}
\]
\[
H_\phi = \frac{I l \sin \theta}{2\pi r \lambda} \left[ -\frac{1}{2\pi} \frac{\lambda}{r} \sin \left( \frac{2\pi r}{\lambda} - \omega t \right) + \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) \right] \tag{4.8}
\]

Where \( \mu \) and \( \varepsilon \) are respectively the magnetic permeability and electric permittivity of the medium surrounding the wire. These equations describe an electric field and a magnetic field that propagate away from the wire. This is called the radiative field. The RF energy transfer link offers long range for power transfer. Now, if we consider the situation where the distance from the RF transmitter \( r \) is much smaller than \( \lambda \), we obtain the following formulae:

\[
E_r = E_\phi = H_r = H_\theta = 0 \tag{4.9}
\]
\[
E_\theta = \sqrt{\frac{\mu I l \sin \theta}{2\pi r \lambda}} \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) \tag{4.10}
\]
\[
H_\phi = \frac{I l \sin \theta}{2\pi r \lambda} \cos \left( \frac{2\pi r}{\lambda} - \omega t \right) \tag{4.11}
\]
In other words, the equations are reduced to quasi-stationary expressions for both the electrical and the magnetic fields. These fields are non-radiating fields called quasi-stationary electromagnetic fields. Through these non-radiating, quasi-stationary electromagnetic fields, energy can be transferred by two main methods: capacitive coupling and inductive coupling.

Energy transport through a capacitive link can, in the most elementary case, be realized using a capacitor, where one capacitive plate is located at the energy transmitting side, and the other plate at the energy receiving side. This is illustrated in Figure 4.1. In the figure, $V_{TX}$ applies a voltage difference to the base unit plates of capacitors $C_1$ and $C_2$, which together act as the transmitter. The capacitor plates in the remote unit behave as the receiver. This behavior can be described by the following equation:

$$V_{RX} = V_{TX} \frac{Z_{RX}}{Z_{RX} + \frac{1}{j\omega} \left( \frac{1}{C_1} + \frac{1}{C_2} \right)}$$

where

- $V_{RX}$: voltage over the remote unit (load)
- $V_{TX}$: supplied voltage at the base unit
- $Z_{RX}$: receiver load

![Figure 4.1: A simple capacitive link: the voltage source of the base unit on the transmitting side $V_{TX}$ applies a voltage difference of $V_{TX} - V_{RX}$ over capacitive plates $C_1$ and $C_2$. The capacitive plates at the remote unit act as a receiver.](image)

We can deduce from the equation that in order for a capacitive link to have a high efficiency ($\eta = \frac{V_{RX}}{V_{TX}}$), high capacitance values are required, as well as high carrier signal frequency. The capacitance value is directly proportional to its surface area, and inversely proportional to the plate distances ($C = \frac{\varepsilon A}{d}$). Therefore, in order to obtain large values for $C_1$ and $C_2$, we must have large plates $A$ combined with a small distance $d$ between them. For the wireless application of this project, the receiver area is limited to a maximum of $1 \text{ cm} \times 1 \text{ cm}$. Moreover, we have estimated the distance from the receiver to the transmitter plate to be on average $5 \text{ cm}$. Hence, capacitive coupling does not seem to be a suitable method for the wireless power transfer of this application, as the efficiency would be too low [14]. Finally, capacitive coupling can occur accidentally in a closely wired circuit, since any two adjacent conductors can start behaving as a capacitor. Such a capacitor is called stray or parasitic capacitance.

The second method of wireless power transfer through electromagnetic fields is through inductive coupling, which uses a quasi-stationary portion of the electromagnetic field. This method is illustrated in Figure 4.2. Here, an alternating current is applied to the primary coil. This current creates an alternating magnetic field. The magnetic field flux of this field affects the secondary coil, where it induces an electromotive force (EMF) in the secondary side of the link. In order to obtain a high link efficiency, the two magnetically coupled coils should resonate at the same frequency, called resonance frequency.
Figure 4.2: An inductive link: the alternating current in the primary coil creates an alternating magnetic field. This alternating magnetic field induces an electromotive force in the secondary coil.

Figure 4.3: Power transfer capabilities, data rates and complexity of the transmission barrier for different types of wireless links [32].

Figure 4.4 shows the maximum achievable efficiency of the power transfer as a function of distance between transmitter and the receiver (i.e., the link distance) for both inductive coupling and ultrasound power transfer. In both cases, the receivers have the same dimensions (1 cm diameter). We can see that for a link distance of less than 3 cm, the inductive link outperforms the ultrasonic link. This, however, is not the case for larger link distances. Unfortunately, as explained before, ultrasound can distress animals. Furthermore, the acoustic link works well if there is no reflection by a solid surface. This is not the case for implantable receiver, which encounters reflections at the air-tissue interface. So, taking the above into account, the preferred method for the power transfer is still the inductive link. However, as shown in Figure 4.4, the efficiency of inductive link drops exponentially with the source-receiver distance. Hence, it is important to design the link in such a way that the distance to the transmitter always remains within a few centimeters.
4.2 Inductive Resonance Coupling

Inductive resonance coupling refers to the wireless transfer of electrical power between two magnetically coupled coils, which are tuned in such a way as to resonate at exactly the same frequency. As a result, these two coils have the property that one coil at a particular frequency is able to drive another coil to oscillate at the same frequency. This frequency is called resonance frequency. The phenomenon of one coil driving another is called magnetic induction. In order to explain magnetic induction, we will briefly revisit a history of relevant theorems.

By the early 19th century, the French mathematician and physicist, André-Marie Ampère stated that when an electric current flows through a wire, a magnetic field $B$ is established around that wire. See Figure 4.5 and Equations 4.13. Ampere’s law relates this magnetic field and this current to one another through the magnetic permeability $\mu$ of the space surrounding the wire. This permeability expresses how willing a material is to form a magnetic field within itself.

\[
\oint \vec{B} \cdot d\vec{l} = \mu_0 \cdot i \quad (4.13)
\]

\[
dB = \frac{\mu_0}{4\pi} \frac{i \cdot dl \cdot \sin \theta}{r^2} \quad (4.14)
\]

\[
\Phi_B = \int_A \int \vec{B} \cdot d\vec{A} \quad (4.15)
\]
Whereas Ampere’s law describes the relationship between a magnetic field and a current flow, the Biort-Savat law allows us to calculate the magnetic flux density $B$ at any point $P$ around the wire (Equation 4.14). This law also gives a definition for the magnetic flux, which flows through an arbitrary enclosed surface $A$ (Equation 4.15).

In short, Ampere and Biort-Savat explain that a flow of current causes the creation of a magnetic field in the surrounding material of the wire, and that the field strength for an arbitrary point is a value that can be calculated. In the years 1831-1832, Michael Faraday and Josef Henry both discovered the law of induction independently:

"The induced electromotive force (EMF) in any closed circuit is equal to the negative of the time rate of change of the magnetic flux enclosed by the circuit." - Faraday 1831

$$\oint \vec{E} \cdot d\vec{l} = \frac{d\Phi_B}{dt}$$ (4.16)

However, as Faraday was the first to publish this discovery, we presently know this theorem as Faraday’s law of induction. After that, Heinrich Lenz in 1834 added a very important negative sign to the induction formula. This means, according to Lenz’s law:

"The direction of current induced in a conductor by a changing magnetic field due to Faraday’s law of induction will be such that it will create a field that opposes the change that produced it." - Lenz 1834.

Thus, finally, after years of theorems that build onto one another, the phenomenon of inductive coupling power transfer was discovered!

### 4.2.1 Mutually Coupled Coils

A current $i$ through a coil $L_1$ can establish a magnetic field $B$ and magnetic flux $\phi$ (Ampere’s law). Now, if this flux changes, and a second coil $L_2$ is placed near $L_1$, a portion of the field lines pass through the second coil. Based on Faraday’s law, this creates an electromotive force within the second coil (refer to Figure 4.6). In such a case, we call these coils mutually coupled.

![Figure 4.6: Mutually coupled coils [38].](image)

The current in $L_2$ projects the same effect onto $L_1$ and, thus, generates the electromotive force $V_1$. $M$ is the mutual inductance between the two coils. In a similar way, in a single current-carrying wire,
4.2. Inductive Resonance Coupling

If the current changes, a voltage can be induced that has opposite polarity as the external voltage applied to the coil. This mutual coupling of a single coil with itself is called self-induction. Hence, the voltage induced on a coil in an inductive link is partially due to coil self-induction, and partially due to the mutual induction. Therefore, the main inductive link equations become:

\[ v_1(t) = L_1 \frac{di_1(t)}{dt} + M \frac{di_2(t)}{dt} \]  \hspace{1cm} (4.17)

\[ v_2(t) = M \frac{di_1(t)}{dt} + L_2 \frac{di_2(t)}{dt} \]  \hspace{1cm} (4.18)

Considering the fact that only the mutual inductance portion of a coil inductance contributes to the power transfer, we can introduce the coupling factor or coupling coefficient. The coupling factor \( k \) is the portion of the flux generated by the primary coil, that flows into the secondary coil. The parameter \( n \) is the square root of inductance ratios. The voltage and the current gain in an inductive link can be modeled using the \( k \) and \( n \) factors:

\[ k \equiv \frac{M}{\sqrt{L_1 \cdot L_2}} \]  \hspace{1cm} (4.19)

\[ n \equiv \sqrt{\frac{L_2}{L_1}} \]  \hspace{1cm} (4.20)

The wireless inductive link can be modeled as in Figure 4.7. This circuit shows the equivalent electrical circuit model of the network.

![Figure 4.7](image)

**Figure 4.7:** Equivalent model for the inductive coupling, with current-dependant current source link (left side), and voltage dependant voltage source (right side).

The secondary leak inductance \( L_2(1 - k^2) \), does not contribute to the mutual inductance. At high frequencies and low coupling, this inductance is much larger than the useful load. Hence, in order to induce enough voltage at the secondary side (\( k.n.v_1 \)), the primary voltage must be increased. The increase in the primary voltage requires increase in the primary current and hence, more losses. Therefore, since the leak inductance contributes to losses, it is beneficial to cancel this out. This is achieved by capacitively loading the coil to form a resonance circuit or an LC tank. This LC tank has an equivalent reactance of \( X(\omega) = \frac{2\pi f C}{L} \). At a certain frequency, called the resonance frequency, the capacitive and inductive reactances are equal in magnitude but opposite in influence, and hence, they cancel each other. The primary coil resonance can also improve the power transfer efficiency. The primary coil is connected to a waveform amplifier or directly to an RF function generator with a finite output resistance. In order to obtain maximum power from a source, the resistance of the transmitter coil must equal the resistance of the source. Hence, the coil leak inductance of the primary coil must be cancelled out by a capacitance, and the coil resistance must match the one from the source. If both coils are tuned to work at one resonance frequency, the link is called a resonance inductive coupling.

An ideal coil does not include any energy losses. Real coils, however, have losses due to their winding wire resistance. This resistance is modeled as a series resistance with the inductance of the coil and is
a combination of the DC resistance and the AC resistance of the coil. The DC resistance is only due
to the wire geometry and the material ($\rho \ast L / A$). The AC resistance, however, is frequency-dependent
and is caused by two undesirable effects: the skin effect and the proximity effect. These effects are
discussed in more details in the next chapter. For now, it is important to realize that the closer the
ideal coil is approximated, or, in other words, the higher quality a coil becomes, the less the series
losses become. The coil quality is indicated by the quality factor $Q$, the ratio of the coil’s reactance
to its resistance at a certain frequency ($Q = \frac{\omega L}{R}$). Now, if we consider the coil’s series resistances
$R_s$ and model the load as $R_{load}$, we can formulate the link efficiency. Thus, we obtain the following
formula for the calculation of the link efficiency, considering both coils are working at the resonance
frequency [38]:

$$\eta_{\text{link}} = \left(\frac{k}{n}\right)^2 \frac{R_{\text{load}} \omega^2 L_1^4 k^4}{(R + R_{S_1}) \omega^2 L_1^4 k^4 + R^2 R_{S_1}}$$  \hspace{1cm} (4.21)

$$R = \left(\frac{k}{n}\right)^2 (R_{\text{load}} + R_{S_2})$$  \hspace{1cm} (4.22)

Expressing this in terms of quality factors, we obtain the maximum link efficiency achievable in an
inductive link:

$$\eta_{\text{link},\text{max}} = \frac{k^2 Q_{L_1} Q_{L_2}}{2 + k^2 Q_{L_1} Q_{L_2} + 2 \sqrt{1 + Q_{L_2}^2 + k^2 Q_{L_1} Q_{L_2}}}$$  \hspace{1cm} (4.23)

In the above formula, $Q_{L_1}$ refers to the quality factor of the transmitter coil, and $Q_{L_2}$ represents the
quality factor of the receiver coil. See Figure 4.8.

**Figure 4.8:** Maximum link efficiency vs coupling factor and the quality factors of the coils [38].

From the figure, it is clear that the efficiency of the inductive link is only dependant on two parameters:
$k^2 Q_{L_1}$ and $Q_{L_2}$. Therefore, maximizing these parameters maximizes the link efficiency as well.
4.3. Coil Design

In the last chapter, we found that the efficiency of an inductive link only depends on parameters $k^2Q_{LS1}$ and $Q_{LS2}$. Therefore, in order to maximize the link efficiency, we must increase the coils’ quality factors and coupling.

- The quality factor of the coils is defined as: $Q = \frac{L}{R}$. The inductance of the coil is proportional to $L = \mu N^2 A$, where $A$ is the surface area of the coil ($\pi (\frac{D}{2})^2$). See Figure 4.9. $\omega$ refers to the angular frequency (set to $2\pi \times 13.56$ MHz), the value of $\mu$ depends on the core material and $N$ refers to the number of turns. The resistance of the coil is proportional to $\rho \frac{L}{A}$, with $L$ the length of the wire. Therefore, in order to improve the quality factor of the coils, we must look into the material and the structure of the individual coils.

![Figure 4.9: Coil dimension parameters. In this figure: p is the pith, D the coil diameter, d wire diameter and l the coil length](image)

- The coupling factor, however, does not depend on the parameters of the coils individually. With a fixed transmitter or receiver dimension, a change in the number of turns of the coils does not affect the coupling factor. The number of turns in a coil, at a specific current, generates a magnetic field. A change in the number of turns influences the strength of this field, and not the fraction of the flux captured by the secondary coil. The coupling factor, therefore, only depends on the relative geometry and the orientation of the coils. Hence, in order to improve the coupling factor, we must look into the relative vertical, lateral and angular misalignment of the coils.

For this, first, we list the application specifications. Then, we investigate the different coil parameters that have an influence on the link efficiency. Finally, a design for the coils is decided upon.

4.3.1 Application Specifications

In order to design the coils in such a way that they are optimal for the application, we must list the application requirements and specifications. The wireless cage is designed for the optogenetic stimulation of multiple mice. Therefore, the design specifications require a minimum area of 40 cm $\times$ 40 cm, in order to provide enough space for multiple mice to freely walk around. The receiver coil and electronics, therefore, will be mounted on the body of the mice. Hence, we first investigate the body dimensions of a mouse, in order to understand what would be the optimal placement for implantation of the receiver module. For this, the physical bodily dimensions of a medium size, 20-day old mouse are measured. Please refer to Figure 4.10 for the details. Also, a 3-D model of the head of the mouse was created for more accurate measurement of the head-ring and nose length, and to examine the applicability of different shapes of receiver. This results in the following measurements: height: 4 cm, head ring diameter: 1.4 cm, between the ears: 1 cm, neck ring: 5 cm, back length: 6 cm, nose length: 2 cm.
We also know that the stimulation will be performed in the brain. To avoid having wires all around
the body, the closer the implantation resides to the place of implantation, the better. Hence, it should
be located close to the brain. The receiver coil also is included in the receiver module. The larger the
plane surface of the receiver coil, the higher the coupling factor. Considering these two facts, and
looking at the space available in Figure 4.10, the area near the head ring is chosen as the implantation
site of the receiver. Therefore, the available area for the head-board module becomes 1.13 cm². Thus,
the receiver coil can have either a ring shape with a diameter of 1.2 cm, or a rectangular shape with a
diameter of 1 cm (the same effective area). The maximum allowable height of the head-board module
is set to 1 cm.

Another important fact related to this project, which might seem obvious, is the fact that mice are
almost continuously moving. This implies that the receiver coil will be moving the entire time as
well. This movement causes lateral, vertical and angular misalignment between the receiver and the
transmitter coil. Misalignment dramatically drops the link efficiency and, hence, must be minimized
as much as possible in the design of the coils. The lateral and vertical misalignments are unavoidable
due to the movement of the mice. The angular misalignment of 90 degrees, however, only occurs at
the seizure. This conclusion is obtained based on the observation of mice behaviour directly and by
hours of movie.

![Figure 4.10: Mouse model and dimensions.](image)

![Figure 4.11: ISM frequency bands.](image)
4.3. Coil Design

<table>
<thead>
<tr>
<th>Conductor</th>
<th>Conductivity (S/m) at 20 °C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminium</td>
<td>$36.9 \times 10^7$</td>
</tr>
<tr>
<td>Copper</td>
<td>$58.6 \times 10^7$</td>
</tr>
<tr>
<td>Gold</td>
<td>$44.2 \times 10^7$</td>
</tr>
<tr>
<td>Silver</td>
<td>$62.1 \times 10^7$</td>
</tr>
</tbody>
</table>

| Table 4.1: Skin depth for various conductors. |

Next, we must specify the resonance frequency of the link. This is achieved by looking at the allowable bands according to the industrial, scientific, and medical radio band (ISM band). ISM is a range of radio bands that are internationally reserved for the use of radio frequency (RF) energy in scientific, medical and industrial applications. Please see Figure 4.11. The higher the resonance frequency, the smaller the receiver antenna can become. However, before picking just any frequency band, we must also look into the electromagnetic influence of this frequency on the neural tissue.

Biological tissues have frequency-dependant electromagnetic properties. At frequencies below 100 MHz, the cell membrane acts as a high capacity insulating layer with high dielectric constant $\varepsilon$ and low conductivity $\sigma$. Research shows that the cell membrane becomes conductive for frequencies above 100 MHz [31, 24]. Therefore, the 13.56 MHz band is chosen for the operating of the inductive link, as this is the highest allowable and safe frequency for stimulation. In short:

- The available volume for mounting the headboard module including the receiver coil is 1 cm$^3$.
- The design specifications ask for a minimum area of 40 cm $\times$ 40 cm for the mice to freely walk around in.
- Due to the application’s particularities, the receiver and the transmitter coil experience frequent lateral, vertical and angular misalignment.
- The resonance frequency of the inductive link is chosen to be 13.56 MHz.

4.3.2 Coil and Core Material

To build the transmitter coil and the receiver coil, a variety of wire types can be used. When choosing the proper wire material, many aspects must be considered, including the resistance of the conducting material, its resilience against corrosion, the bending flexibility and the cost. Here, we consider a number of commonly available materials: aluminium, copper, gold and silver.

Table 4.1 includes the conductivity of the materials. Given the above considerations, silver and copper seem to be the best conducting materials. That is, they seem to have the lowest electrical resistivity. For an application that uses DC current, this would be the most important characteristic to consider. However, for applications dealing with high frequency AC current, DC resistivity can no longer be the sole determining parameter for the conductivity of the wire. At high frequencies, the resistivity becomes frequency-dependent. This is caused by the so-called skin-effect.

The alternating electric current creates an alternating magnetic field not just around the conductor, but also inside the conductor. The change in the current intensity induces changes in the magnetic field, which results in the creation of circular electric field. This electric field within the conductor establishes circulating currents, called Eddy currents (Faraday’s law). To visualise this better, please refer to Figure 4.12.

Eddy currents resist the change in the magnitude of the current flow. This is called ‘back electromotive force’, or back-EMF. Back-EMF is stronger towards the center of the wire and weaker towards the skin. In other words, Eddy currents cancel the electric flow in the center of the conductor and reinforce it to the surface. Hence, the charge carriers are moved towards the surface. This phenomenon, known
as skin effect, alters the effective conducting surface of the conductor and, consequently, increases the effective resistance of the wire. The skin depth indicates the depth in which the current density is lower than \( \frac{1}{e} \). From the formula below, it is clear that the skin depth is inversely proportional to the square root of frequency. Accordingly, at high frequencies, the skin effect will prove to be a major conducting loss that must be taken into account.

\[
\delta = \sqrt{\frac{2\rho}{\omega\mu}} \sqrt{\sqrt{1 + (\rho\omega\epsilon)^2} + \rho\omega\epsilon}
\]  

(4.24)

Table 4.2 shows the conductivity of the above mentioned metals at 1 GHz. From this table, it is obvious that silver and copper have slightly lower skin depth and, therefore, lower conductivity at high frequencies.

We can see that copper appears to have slightly lower conductivity at higher frequencies as compared to gold and aluminium. Nonetheless, it has many advantages over the other conductors mentioned here, such as being low-cost, light-weight, flexible to form, easy to solder and to make joints, and long-lasting. Silver is very rigid and corrodes very easily. This means that it is highly receptive to form a high-resistance oxide on its surface. It is very hard to solder aluminium. And finally, gold has a much higher cost per gram as compared to copper, making it an unsuitable choice for the design. One drawback of copper is that it corrodes over time. To reduce the impact of its tendency to corrode, copper wires are coated. Therefore, copper is chosen as the wire material for the coil.

The coil wire material is not the only parameter that affects the quality factor of the link. The inductance of the coil is directly proportional to the permeability of the core material. The permeability of a material is expressed as relative to the permeability of free space, or \( \mu = \mu_r \mu_0 \). For example, aluminium is a non-magnetic conductor and, therefore, has a relative permeability of 1. This means placing any piece of aluminium in a magnetic field will not affect this field. The material surrounding the wire can have a very high relative permeability. For example, ferromagnets are built from very tiny permanent magnetic particles. Hence, the induced field by the current causes strong magnetization of the material, even after the external magnetic field is no longer present. This is called the ferromagnetic effect. These materials, due to their high permeability, have low magnetic
resistivity. Since the magnetic field lines will find the route that has the lowest permeability, more flux will be captured by the receiver. Hence, using ferromagnets as core material for the receiver coil can dramatically improve the link performance. This depends, however, on the final design of the coil shape. Ferromagnetic materials have a high mass density, and thus, they are not considered in the initial design of the receiver coil. If the receiver module design and weight allows, they will be included.

### 4.3.3 Coil Wire Type

Once the *material* of the wire has been decided upon, the next step is to choose the *type* of wire. Besides the skin effect, another conduction loss that becomes more apparent for higher frequency ranges, is the proximity effect. This effect, in contrast to the self-induced skin effect, is caused by external magnetic forces. When two wires are placed in close proximity of one another, such as in a coil, the nearby wires produce a magnetic field that induces eddy currents in the adjacent conductor. This causes the electrical charges to be compressed into a smaller regions furthest away from the adjacent conductor. Similar to the skin effect, this effect also becomes more prominent with increased frequency, and increases the effective resistance of the wire [40].

One way to reduce both the skin effect and proximity effect is to use the so-called Litz-wire. A Litz-wire consists of many individually isolated conductors, which are bundled together. Hence, for a Litz-wire, the current density is distributed over the entire cross-section of the wire. This effect, however, only works for frequencies below 1 MHz. Above this frequency, the single wires do show higher quality factor. Please refer to Figure 4.13 to see the electrical performance of solid wire versus Litz wire for various frequencies.

![Figure 4.13](image)

*Figure 4.13:* Relation between frequency and inductance, and frequency and quality factor for a solid wire, and for a Litz wire. The blue arrow points to the corresponding y-axis for the graphs.

Another conductor design approach aimed at minimizing the skin and proximity effect is copper tubing. Such tube conductors have a hollow core and, therefore, have a much lower AC resistance than conventional copper wires. Unfortunately, these wires are much more difficult to be wound than regular wires. When viewing from a circuit-level, these conductors have a larger outer surface, which causes the parasitic capacitance of the windings to increase.

From the above, we can conclude that standard copper wire is preferred for those applications where frequencies above 1 MHz are used. For the design of the transmitter coil, enameled round copper wire
of 18 AWG (American wire gauge) has been chosen, since it offers a large cross-section and, therefore, lower coil resistance, and at the same is thin enough to be flexibly formed into the coil shape. Copper wires with lower gauge number are not possible to be wound without mechanical tools.

### 4.3.4 Coil Shape

The coupling factor between the coils only depends on their relative geometry, and on the relative angular, vertical and lateral positions of the coils. For studying the efficiency of power transfer, different shapes for the transmitter and the receiver coil have been investigated, as well as a number of combinations thereof. The design requirements impose strict limitations onto the design of the transmitter and the receiver coils. For both coils, the maximum base area is known. Moreover, the receiver coil can only be of a maximum height. For the transmitter coil, however, no height limit is specified. To find the optimal shape for designing the transmitter coil, we have first looked into the fields generated by various types of coil through the use of HFSS simulations. Figure 4.14 shows the different magnetic fields created by a solenoid, by two types of toroid and by a flat spiral coil.

![Magnetic field generated by various coil shapes: a) solenoid b) toroid c) toroid d) flat spiral.](image)

It would be ideal to have a coil shape that offers the biggest volume with an almost uniform field, so that the vertical and lateral misalignment will not affect link efficiency. As we can see in the figure, the solenoid coil surrounds an area inside itself, which has an almost uniform magnetic field. The flat spiral coil also appears to be a good option. The toroidal coils do not seem to provide the required qualities. This narrows down the potential options to choose from the solenoid and flat spiral coils.

As we are designing an inductive link, looking at the field uniformity of only the transmitter coil is not the only concern. We must also consider the performance of these coils in the inductive link. For this analysis, we need to investigate the two main limitations that the application poses onto the link design. One of these limitations is the always moving location of the head of the mouse. Hence, vertical and lateral misalignment between the transmitter and the receiver coil are likely to occur frequently. Second, the angular movements of the head of the mouse, which causes the angular misalignment between the transmitter and the receiver coil, is also to be expected. We must investigate each of these two types of misalignment in isolation of one another, in order to ensure
that any conclusions drawn from the comparison of the two types of coils for the lateral misalignment is only due to the lateral misalignment, and not caused by the angular misalignment, and vice versa. Therefore, two sets of coils for two inductive links were designed. First, a 40 cm $\times$ 40 cm flat spiral coil with a single loop receiver. Second, a 40 cm $\times$ 40 cm rectangular helical coil with the same receiver (see Figure 4.15).

The receiver is chosen to be a simple loop (diameter of 1.2 cm) for both cases, since at this point in time, we are only investigating the transmitter coil performance. In order to investigate the impact of the lateral misalignment on the two different types of transmitter coils, we calculate the inductive link efficiency for:

- all the points where the receiver loop is placed on the top of the flat spiral coil, with no vertical or angular misalignment.
- all the points where the receiver loop is placed on a plane located at the middle of helical coil, with no vertical or angular misalignment.

Please refer to Figure 4.16 for a visualization of the lateral and angular misalignments. The parameters shown in the figure are transformed into a set of calculations and then implemented in Matlab. These calculations are taken from [8]. The result is shown in Figure 4.17.

This figure visualizes the link efficiency when lateral misalignment is introduced. The peak of each graph is for the situation where no misalignment occurs. Lateral misalignment in the x-dimension and/or y-dimension reduces the link efficiency substantially. We can see from these figures that in a flat spiral coil, lateral misalignment dramatically reduces the link efficiency for all the situations where the receiver coil is located at the edges of the transmitter coil. In this case, for a band of 5 cm in width all around the coil, the link efficiency drops to almost zero. This is about 43 percent of the overall area of the transmitter coil. Hence, a flat spiral coil as a transmitter has a very poor performance when the application involves lateral misalignment. The helical coil, however, only reached low link efficiency for the situation where the receiver coil is located at the edges of the transmitter coil. This area takes up less than 7 percent of the transmitter area. Hence, it follows that a flat spiral coil is more prone to lateral misalignment than a helical coil.

There are a few ways that can be used to overcome the problem of lateral misalignment with flat spiral coils. One example is to create a multi-layer mesh field of smaller flat spiral coils. Using a double-layer mesh field, as opposed to a single large flat spiral coil, reduces the sensitivity of the transition lines to lateral misalignment. The mesh field still does not compensate for the vertical misalignment, however. The smaller the mesh spirals become, the less sensitive the link efficiency becomes to the lateral misalignment. But, at the same time, the field produced by smaller spiral coil has a reduced vertical range. As a result, the transmitter coil is less capable of supplying a field for the vertically misplaced receiver coil. In other words, the smaller the spiral coils become in a

![Figure 4.15: (left) a 40 cm $\times$ 40 cm flat spiral coil; (right) a 40 cm $\times$ 40 cm rectangular helical coil.](image)
spiral mesh, the less sensitive the link efficiency becomes for lateral misalignment, but more sensitive to vertical misalignment. A helical coil, on the other hand, can offer a magnetic field over a larger volume. This is important, since it would be desirable to design this setup not only for mice, but also for rats, which are taller. It would also be interesting to design the setup for different experimental conditions, such as for rodents that are using a treadmill.

The effect of angular misalignment is the same for both the helical and the spiral coil and is shown in Figure 4.18. From this figure, it is clear that both transmitters exhibit poor performances when an angular misalignment of 90 degrees occurs. In this case, the efficiency of the power transfer drops to zero. As discussed in the design specification section, the mice head ring is almost never perpendicular to the ground. During seizures, however, this is prone to happen. As the main focus of the optogenetics stimulator is to stimulate the seizure at the onset, this case should not occur. However, even in the unlikely event that such an event does take place, a battery-type storage element could be used to ensure that sufficient energy is available for stimulation. This battery could be charged when the induction power transfer operates at high efficiency, and supply power in low-efficiency conditions.

Based on the above, a helical shape is chosen for the transmitter coil. In order to obtain the maximum area possible while still abiding by the design specification, and also due to the convenience of designing the box in a rectangular shape, the helical coil shape is decided to be of rectangular shape.
4.4. Coils at Resonance

Based on the results described in the previous sections, both receiver and transistor coils are designed as a rectangular solid-copper solenoid. In an electrical circuit, a simple circuit equivalent of a coil includes an inductor in series with a resistor and in parallel with a capacitor. Please see Figure 4.19.

![Figure 4.19: Simple equivalent circuit of an inductor (coil).](image)

In this LC circuit, the inductor L stores the energy in the magnetic field, and the capacitor C stores the energy in the electrical field. The reactance of this capacitor and inductor cancel each other out at the resonance frequency. A coil can operate at the resonance frequency in three ways:

- By loading a coil with an external parallel capacitance. An inductive link with parallel secondary resonance behaves as a voltage source.
- By loading a coil with an external series capacitance. An inductive link with a series secondary resonance behaves as a current source.
- By designing the coil such that the reactance of the self-inductance and the parasitic capacitance of the coil will cancel each other out and cause the coil to work at resonance without the need of any external component.

On the one hand, the transmitter coil is designed in such a way that a self resonance at 13.56 MHz occurs. The receiver coil, on the other hand, is loaded with a parallel capacitor. This is due to the fact that the electronics used at the receiver side have a higher operating efficiency for a higher input voltage to current ratio. In order to achieve these resonances, we consider two different cases:
• The transmitter coil is a distributed system of losses and parasitic capacitance over the entire length of the coil and, therefore, since it is highly complex, it cannot be modeled using an equivalent circuit as simple as the one shown in Figure 4.19. Hence, in order to find the optimum operating point of the transmitter coil, the industry-standard engineering software package called High Frequency Electromagnetic Field Simulation (HFSS) is used.

• Since the receiver coil has small dimensions, it can be modeled as a localized coil. An equivalent electrical circuit model is used to model this coil, its losses and its parasitic capacitance. Furthermore, Matlab simulations are performed to find the optimum operating point for this coil. The receiver coil is connected to the electronics at the receiver side. These electronics have a lossy and non-linear characteristic. Therefore, they affect the performance of the receiver coil, as well as the resonance frequency. Hence, the design of the receiver coil is included in the next chapter together with the electronics on the head-mounted module.

4.4.1 Transmitter Coil at Resonance

For the design of the transmitter coil, first we look at the inductance for such a coil. This can be calculated as follows:

\[
L_{rec} = \frac{N^2 \mu_0 \mu_r}{\pi} \left[ -2(w + h) + 2\sqrt{h^2 + w^2} + temp \right]
\]

(4.25)

where

\[
temp = -h \ln \left( \frac{h + \sqrt{h^2 + w^2}}{w} \right) - w \ln \left( \frac{w + \sqrt{h^2 + w^2}}{h} \right) + h \ln \left( \frac{2h}{r} \right) + w \ln \left( \frac{2w}{r} \right)
\]

N indicates the number of turns, h the height of the coil, w the width of the coil, and r the wire radius. Looking at the formula and considering the design specification, we realize that the cross-section area of the coil is fixed to an area of 40 cm × 40 cm. Therefore, the only parameters that we can modify in the design are: the wire diameter, the coil height and the number of turns. An 18 AWG copper wire is chosen, since it has the largest cross-section while still being flexible. The reason for this is that the larger the wire cross-section, the lower the losses in the coil will be. If for whatever reason the design procedure does not provide a good solution for this wire type, we can look into using wires of different sizes.

Based on the above, we make a coil design in HFSS. The coil performance is shown by the S\(_{11}\) vs frequency plot, as well as by the Smith chart. S\(_{11}\) is a scattering parameter, which describes the transmitting capability of the coil. The value of S\(_{11}\) indicates how much of the power is reflected from the coil. S\(_{11}\) is expressed in dB and can also be translated into the power ratio with the following relation: \(\text{Ratio}_{\text{dB}} = 10 \log_{10} \left( \frac{\text{reflected power}}{\text{incident power}} \right)\). For example, an S\(_{11}\) of 0 dB means that no power is radiated from the coil and all the power is being reflected. An S\(_{11}\) of -10 dB means that 10 percent of the power is reflected and 90 percent is transmitted or absorbed as losses in the coil. A key factor for the design is to ensure that the majority of the power delivered to the coil is radiated. Therefore, the lower the S\(_{11}\), the better the performance of the power transmitting coil. The maximum acceptable S\(_{11}\) for a transmitting coil is set to -10 dB.

We first simulate a coil with only 5 turns to investigate the resonance point and the magnetic field generated by the coil (refer to Figure 4.20 a). We see in this figure that for the simulations the coil is placed in an air box and excited through the red arrow. For this coil we have a very shallow resonance with an S\(_{11}\) of -5 dB at frequencies much higher than the resonance. The inductance increases with the number of turns squared. Therefore, we increase the number of the turns until we reach a frequency close to the desired resonance frequency. The inductance is inversely proportional
to the pitch. Hence, for further fine tuning of the frequency, we increase the pitch. Finally, at a spacing of 0.8 cm between the turns, and with a number of turns of 30, we obtain a resonance at the frequency of 13.56 MHz with a scattering parameter of -20 dB. Please refer to Figure 4.20 b.

We can also see this from the corresponding Smith charts in Figure 4.21. The mid horizontal line of the Smith chart shows all the points where the coil is purely resistive. The closer we are to the center of the Smith chart, the more matched our coil impedance is to 50 \( \Omega \) and, therefore, the less fraction of power from a 50-\( \Omega \) source is being reflected. This is due to the fact that impedance mismatch between the coil and the source results in the standing wave in the transmission line. Voltage Standing Wave Ratio (VSWR) is an indication of the AC voltage due to a standing waves along the feed line. In other words, VSWR shows how well the coil is matched to the source impedance of usually 50 \( \Omega \). A VSWR of 1:1 indicates no AC voltage due to standing waves, and is only obtained when the coil impedance is perfectly matched to the source impedance. For the semi-circle above the mid-line, the coil is more inductive than capacitive. For the lower semi-circle, however, the coil behaves more capacitive than inductive. The center point of the Smith chart belongs to the impedance of 50 \( \Omega \). After fine-tuning the coil parameters (number of turns, wire spacing and pitch) exactly for the frequency of 13.56 MHz, we have a purely resistive coil that at the same time is matched to a 50 \( \Omega \) impedance. The flow of the Smith chart for the coil impedance shown in Figure 4.21 demonstrate how tuning the parameter of the coil bring its operating frequency to the resonance.

Next, the magnetic field generated by the coil is investigated. The basic idea of choosing such a coil as a transmitter was to obtain an almost uniform field for a big volume inside the coil. As we explained before, the reasoning behind this idea was to eliminate the impact of the lateral and vertical misalignment on the link efficiency. To achieve this, while changing the coil parameters to obtain resonance, we also looked at the magnetic fields generated by the coil. The magnetic field for the transmitter coil with 30 turns is depicted in Figure 4.22.

The bottom 4 cm and upper 4 cm of the coil include weak magnetic field. This is where the field

---

**Figure 4.20:** Scattering parameter vs frequency. The s parameter is shown on the vertical axis with each segment equal to 1 dB.
36 4. Wireless Energy Transfer

<table>
<thead>
<tr>
<th>N</th>
<th>Wire radius (mm)</th>
<th>Coil width (mm)</th>
<th>Pitch (mm)</th>
<th>L (µH)</th>
<th>Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>30</td>
<td>0.5</td>
<td>400</td>
<td>1</td>
<td>367</td>
<td>113</td>
</tr>
</tbody>
</table>

Table 4.3: Calculated coil parameters

lines begin to deflect from being perpendicular to the cage bottom. Hence, we define the available height as the middle 20 cm of the coil. From here on, when we mention the button of the coil, we mean the red line shown in Figure 4.22. The transmitter coil is finally built over a transparent box made of Polymethylmethacrylate or PMMA. Copper wires are fixed in position around the box using screw terminal. The built prototype is shown in Figure 4.23. Table 4.3 includes the building and measurement parameters of the coil.

4.5 Conclusions

In this chapter, different wireless power transfer methods were briefly discussed. Wireless power transfer through an inductive link represents the best trade-off between requirements for the application

Figure 4.21: Smith chart for a) 5-turn transmitter coil, b) 10-turns transmitter coil, c) 20-turn transmitter coil, d) 30-turn transmitter coil.
of this project. Furthermore, various parameters that have an influence on the wireless link efficiency were investigated. These parameters include: operating frequency, coil material, wire type, shape, vertical misalignment, lateral misalignment and angular misalignment. For all these parameters we have made choices that offer the best fit for maximizing the link efficiency. Finally, a solid-copper wire rectangular solenoid coil for both transmitter and the receiver was selected. Using High Frequency Electromagnetic Field Simulation Software (HFSS), the design of the transmitter coil is finalized in such a way as to resonate at the frequency of 13.56 MHz and provide enough area with a uniform magnetic field, in order to reduce the effects of lateral and vertical misalignments.
Wireless Optogenetics Receiver

The wireless power harvested by the receiver coil cannot be directly used to supply the optical load: first of all, the harvested power generates AC, whereas our load requires DC. Second, it cannot always be guaranteed that the received power conforms to the power requirements of the load. Therefore, the harvested power needs to be rectified, regulated and controlled.

The overall power transfer efficiency of the entire wireless system not only depends on the inductive link efficiency, but it also heavily depends on the receiver module efficiency itself. We call this the line efficiency. The receiver module efficiency, or line efficiency, is basically the efficiency of the power conversion from the harvested AC power to the power supplied to the optical load. Therefore, in order to maximize the total wireless system efficiency, we must choose the most efficient, but at the same time also most suitable design for the sub-modules below:

- Optical module
- Controller module
- Regulator
- Rectifier
- Power module

A holistic view of the entire system is essential for achieving optimal results. This is reflected by the design approach taken, which not only considers the best optimal choice for picking the individual blocks, but also the interaction with their neighbouring blocks. The design of the receiver sub-modules are explored in this chapter.

5.1 Optic Module

The first step in designing the receiver module is to have a clear understanding of the load of the system. As discussed in the previous chapters, the stimulation method that has been selected in this project is optogenetic stimulation, which requires the emission of light. This light can be generated through the use of an LED or a laser diode. In this section, first we compare the suitability and performance of lasers and of LEDs for optogenetics. Then, we define the design specifications for the optrodes. Finally, we explain the fabrication procedure of the optrodes developed in this project.
5.1.1 LED vs Laser

The laser most commonly used in optogenetics is the diode-pumped solid state (DPSS) laser. This laser’s working principle is based on optical pump radiation, which increases the efficiency of emission compared to competing laser technologies. DPSS lasers also provide a highly focused beam. A drawback of this is that this highly focused radiation generates unwanted reflections from the sample, so-called self-interference, as the transmitted beam scatters when projected onto the sample. These scatter beams have the same frequency as the emitted light, but differ in their phase and amplitude. Hence, due to subtraction and addition, random local radiation of various strengths is generated. These variations are called speckle noise and are highly undesirable for optogenetics, since during an experiment we would like to know the exact stimulation intensity, as this often has a large influence on the experiment result. Moreover, high light intensities are harmful to the tissue [28]. Another problem with lasers is that the electromagnetic energy reflects against the walls of the laser cavity, thus producing standing waves. This effect produces a cavity resonator, which causes different modes of radiation, resulting in power and wavelength instability. This phenomenon is called mode hopping, which makes lasers an unsuitable power source for optogenetic stimulation.

A variety of different stimulating patterns are required for optogenetic stimulation. These patterns can vary in their output power levels, as well as in their different temporal patterns and frequencies. The latter implies that the optical stimulation should be able to switch on and off with high precision. The optogenetic stimulation can, for example, occur with a frequency of 50 Hz [2]. Hence, the laser switch needs to be able to reach this stimulation rate and have fast rise and fall times in order to provide a clean stimulating pattern. As DPSSs themselves do not provide a high-speed transistor-transistor logic (TTL) modulation, an additional switch must be added to the circuit. Hereof, mechanical switches (i.e., shutters) are used in combination with the laser. These switches are fast enough to provide precise, high-speed switching capability to the laser. These switches, however, are acoustically noisy and bulky. Despite the above mentioned disadvantages of lasers for optogenetic stimulation, lasers do have one great benefit over LEDs: lasers, due to their ability of emitting very focused beams, can be very efficiently coupled to narrow openings, such as an optic fiber with 200 \( \mu \text{m} \) cross-section.

In contrast to lasers, LEDs do not have a resonant cavity and, therefore, do not produce mode hopping. LEDs have a nearly linear relation between electrical current and optical power output. Hence, they offer the ability to generate very precise temporal stimulation patterns. LEDs provide a homogeneous, stable and, therefore, safer and more accurate stimulation compared to lasers. One disadvantage of LEDs is their incapability of emitting focal radiations. This leads to the main disadvantage of LEDs: LEDs can not provide focused beams. Therefore, when an LED is coupled to a smaller opening, most of the light is lost. A good example of this is the conventional Thorelabs optical fibers (CFMLC12L05) that are used for optogenetic stimulation. These fibers are coupled to an LED source. At the coupling between the fiber and LED, at least 99% percent of the power is lost. This is a big drawback for this project, since in a wireless power transfer link, power-efficiency is a key design parameter.

Table 5.1 contains a summary of the above comparison between LEDs and lasers. Taking the above into account, LEDs are the better choice to use as light source in this project, if we can find a way to eliminate the coupling losses that occur when connecting the laser to the optic fibers. Therefore, a contribution of this work has been the idea to directly implant micro-scale LEDs, thereby avoiding the above explained coupling altogether. A novel optogenetic \( \mu \)LED implant solution has been developed during the work of this thesis. The fabrication, assembly and mounting techniques thereof are explained in the next section.
5.1. Optic Module

### Table 5.1: LED vs laser.

<table>
<thead>
<tr>
<th>Optogenetic Stimulation Requirements</th>
<th>LED</th>
<th>Laser</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stability of wavelength</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Low noise</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Illumination intensity control</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Well-defined temporal control</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Well-defined spatial homogeneity</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Focus beam to very small points</td>
<td>-</td>
<td>+</td>
</tr>
<tr>
<td>High coupling efficiency</td>
<td>-</td>
<td>+</td>
</tr>
</tbody>
</table>

5.1.2 Optrode Design Specifications

The initial stimulation target for the design of the LED is a section of the brain called the thalamus. As can be seen in Figure 5.1, which shows the horizontal cross-section of a mouse brain, the thalamus is located at a depth of between 2.5 mm and 3.8 mm from the surface of the brain. Therefore, this is the depth at which the optogenetic stimulation must take place. Light with a wavelength of 475 nm, transmitted from a 5 mW source (with a 200 µm diameter), will lose around 75% of its irradiation power within a radius of 0.2 mm in the brain tissue. A source with the same dimensions that transmits light with a 570 nm wavelength will lose around 70% of its irradiation power within a distance of 0.2 mm in the brain tissue (for the calculation, refer to [25]). Therefore, to achieve effective optogenetic stimulation, and to avoid heating up the tissue, the LEDs must be placed in close proximity of the stimulation site in the thalamus.

![Horizontal cross-section of the mouse brain. The dimensions are specified in mm [26].](image)

Many parameters are considered when choosing between the available LEDs, such as the LED chip size, output power and wavelength to be used for stimulation. Products from a number of LED manufacturers have been investigated: Samsung, Cree Semiconductor, Rohm Semiconductor, and Plessey Semiconductors. Of all the evaluated LEDs, the most suitable ones that comply with the power, wavelength and size requirements are:

**Blue LED at 470 nm:** Cree Semiconductors Razer Thin Gen III LEDs C470UT200 bare die. 470 nm is the peak wavelength for activation of light-gated membrane channels, Channelrhodopsins.

**Yellow LED at 570 nm:** Cree semiconductors C570UT170 (low height). 570 nm is the peak wavelength for activation of light-gated membrane channels, Halorhodopsin.

Figure 5.2 shows the top view, cross section and the bottom view of these µLED chips. In the experiments that have been performed with a wired setup, the light intensity at the tip of the implanted fiber was $550 \pm 50 \mu \text{W/mm}^2$ [18]. With a forward voltage of 2.7 V and a current of 5 mA, both 470 nm and 570 nm µLEDs emit more than 6 mW/m$^2$ of output power [5]. This is more than sufficient for optogenetic stimulation. The stimulation can require more than one µLED:
• First use of multiple $\mu$LEDs is to obtain focal stimulation effect, using $\mu$LEDs with the same wavelength. This requires multiple $\mu$LEDs, which upon emission, cause optogenetic stimulate on the portion of the neural tissue, which is not covered by the emission of the other $\mu$LEDs. For example, using three beam-focused blue $\mu$LEDs, we can compare the effect of stimulation in only upper thalamus, only lower thalamus or both.

• Second use of multiple $\mu$LEDs is to obtain different stimulation effects, using $\mu$LEDs with different wavelengths. For example, stimulating by an optrode with an blue $\mu$LED and an orange $\mu$LED, we can cause neural excitation and neural inhibition on the affected neural tissue.

5.1.3 Design of Rigid Needle Optrodes

For the design of rigid optrodes, we use an injecting medical syringe as a platform for mounting the $\mu$LEDs. The cavity inside the needle allows for the anode wires to pass through and reach the $\mu$LEDs. This needle is made of aluminium. Hence, we are able to use the structure of the needle as the ground plane of the $\mu$LEDs. This way, we reduce the number of required wires and, in turn, the size of the implant. In Figure 5.3a, it can be seen that each individually-controllable LED requires its own anode wire. Hence, the larger the number of $\mu$LEDs in the LED array, the thicker the needle needs to be, as it requires a separate wire for each of the LEDs in the array. In contrast, if the needles do not require to be individually controllable, they can placed in parallel, as is depicted in Figure 5.3b. In this case, additional $\mu$LEDs only increase the length of the optrode, and not its thickness.

The design procedure is explained in Figure 5.4. After creating an LED array according to this procedure, it is now ready to be tested. Testing of the optrodes is performed in two stages:

Bonding test: This first test is used to evaluate the functionality of the $\mu$LED after bonding (refer to Figure 5.5 a).

Isolation test: This test is performed after the optic glue coverage of the optrodes. The optrode is tested in isolation and also in water and gelatin (Figure 5.5 b), to ensure that the optrode is humidity isolated and also that it is mechanically strong. A long duration test is performed on the $\mu$LEDs with the actual stimulation program (this program runs on a nano-Arduino).
The design of the optrodes has evolved through a number of different shapes. During an initial test on gelatin, it was discovered that the tip of the optrode in the first model caused a large angle opening in the tissue (refer to Figure 5.6a). Hence, after testing a variety of different shapes and their effect on the tissue, the final tip shape of the µLED optrodes was decided to be a V-shape, as can be seen in Figure 5.6b. Using this shape as needle tip also allows for the placement of the µLEDs closer to the needle tip. Image c shows how close the µLED can be placed to the tip of the needle.

Figure 5.4: Needle optrode fabrication procedure: a) and b) Shape the edge of the needle to the flatter surface using the scrub roller machine that rubs the surface with a rolling stone (image 1). c) Remove the plastic connector. d) Remove the isolation from the end of the wire and edge button of the needle. Apply conductive silver glue to the needle. e) Wrap the other end around the back-end of the needle and heat it up. f) Insert one, two or three anode wires through the needle cavity. The wires are fixed in position and the isolation of the middle part of the wire is removed. The needle tube is filled with non-conducting fast curing glue. Glue is heated up to 170 degrees Celsius to quickly dry out within a minute. g) Apply integrated circuit silver paste on the flatten surface (image 2). h) Place the µLEDs on the paste using the pick-and-place machine (image 3), heating up the needle. Then, perform gold bonding on the µLEDs, using a bonding machine (image 4). i) Cover the µLEDs with the UV-curing optic glue (image 5) and fix the plastic end to the wires by using heat-shrink tubes. The needle is then placed under a UV lamp for 15 minutes. Now, the needle is ready for testing.

In order to reduce the fabrication time of the needle optrodes, a printed circuit board (PCB) has been designed that simplifies some of the manual design steps. Please refer to Appendix A, Figure A.2 for the PCB layout.
5.1.4 Design of Flexible Optrodes

In the rigid optrodes, the needle body was used as a platform to keep the rigid shape of the optrode, as well as to provide the ground plane. For the design of the flexible optrodes, we must eliminate the needle in order to obtain flexibility. However, then we unfortunately also lose the ground plane. The grounding can be replaced by either using another flexible conductor, or by designing the optrodes without a separate ground line. The second option allows us to keep the number of wires equal to the number of individually-controllable μLEDs. This is done by placing two μLEDs flipped in parallel, as in Figure 5.7a.

In this approach, the anode of one μLED faces the cathode of the other μLED. In theory, this parallel configuration should be possible. In practice, however, this was not feasible when using the two μLEDs of Figure 5.2, as these μLEDs have the same polarities. This means that the ground metalization of both μLEDs acts as the cathode, and the gold plate, where the gold bonding wire sits, acts as the anode. When using these μLEDs in the configuration described in Figure 5.7a, this
5.1. Optic Module

Approach results in mounting one $\mu$LED's ground metalization onto the mounting platform, and the other $\mu$LED's gold plate onto the mounting platform. Therefore, as light on these $\mu$LEDs shines around the gold plate, we block the emission from one of the $\mu$LEDs. Hence, for this configuration, we use $\mu$LEDs that have the same packaging, but different polarities. Then, for one $\mu$LED, the ground metalization is the anode, and for the other one, the cathode. This way, both gold plate sides of the $\mu$LEDs face upwards. This is shown in Figure 5.7b.

Figure 5.8: Semi flexible FFC optrode. a) prototype: two sets of jointly controlled $\mu$LEDs, red and blue. The inverse polarity allows for the configuration as shown in part b), b) $\mu$LEDs configuration.

In the design of the semi-flexible optrodes, the $\mu$LEDs are mounted on the tip of a flat flexible cable (FFC). The FFC platform consists of multiple rows of flat wires that are attached together and insulated using a thick layer of polyester and flame retardant adhesive. The minimum width of the FFC optrode is determined by the number of wires required to mount individually controllable $\mu$LEDs. These wires are placed next to each other, with a small distance in-between them to isolate them from each other. Therefore, the more individually controllable $\mu$LEDs an array includes, the wider the size of the implant becomes. We can see in Figure 5.8a and corresponding $\mu$LEDs configuration in Figure 5.8b, that addition of $\mu$LEDs with joint control increases the length of the optrode, whereas addition of individually controllable $\mu$LEDs increases the width of the optrode.

Figure 5.9: 18 $\mu$m thick wires sitting on 160 $\mu$m wire.

Polyester is a semi-flexible material and, hence, not a suitable platform material in order to build fully flexible implants. The most flexible platform for the implant would actually be to have no platform. This thought initiated the design of the fully flexible implants, which consists of $\mu$LEDs and wires only. The main idea is to use different size wires for mounting the $\mu$LEDs and to bond them. A thicker wire by itself can then act as a wide platform for smaller modules. Please see Figure 5.9, which illustrates the idea of using thick wires as a $\mu$LED mounting platform.

In order to build the fully-flexible optrodes, a tool called the hoek tool has been developed by ing. Wil Straver. The hoek tool is used to keep the wires in place around the needle’s tip and to twist
the wires. Twisting the wires is an essential step during building the flexible optrodes, and when dealing with wires as thin as 80 µm diameter, it is an impossible task without the hock tool (refer to Figure 5.10). The fabrication procedure of fully-flexible optrodes is described in Figure 5.11. The final optrode is 400 µm wide (the width of two µLEDs) at the tip and 200 µm diameter thick at the fully-flexible optrode body.

![Figure 5.10: The hock tool.](image)

<table>
<thead>
<tr>
<th>Step</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Cut 20 cm piece of wire (80 µm, 100 µm or 160 µm in diameter)</td>
</tr>
<tr>
<td>2.</td>
<td>Make a tiny loop of wire inside the needle</td>
</tr>
<tr>
<td>3.</td>
<td>Put the loop inside the “hock tool”, twist the wires</td>
</tr>
<tr>
<td>4.</td>
<td>Heat up to 170 °C for a few seconds</td>
</tr>
<tr>
<td>5.</td>
<td>Apply a micro-drop of two-component glue on the twisted knot</td>
</tr>
<tr>
<td>6.</td>
<td>Remove wires from the “hock tool”</td>
</tr>
<tr>
<td>7.</td>
<td>Check the stiffness of the glue. Glue color turns from gray to brown</td>
</tr>
<tr>
<td>8.</td>
<td>Cut the loop open close to the lower wire and shape it flat</td>
</tr>
<tr>
<td>9.</td>
<td>Cover the open loop with two-component glue</td>
</tr>
<tr>
<td>10.</td>
<td>Heat up to 170 °C for a few seconds</td>
</tr>
<tr>
<td>11.</td>
<td>Clean up with air pressure</td>
</tr>
<tr>
<td>12.</td>
<td>Apply integrated circuit glue</td>
</tr>
<tr>
<td>13.</td>
<td>Mount the bare dies of µLED on the lower step</td>
</tr>
<tr>
<td>14.</td>
<td>Gold-bond the µLEDs gold plate to the copper dot</td>
</tr>
<tr>
<td>15.</td>
<td>Clean the heat-generated oxidation of the copper dot by LPKF</td>
</tr>
<tr>
<td>16.</td>
<td>Heat up to 170 °C for a few seconds</td>
</tr>
<tr>
<td>17.</td>
<td>Test the µLEDs blinking</td>
</tr>
<tr>
<td>18.</td>
<td>Cover the top of the optrode with UV-curing glue</td>
</tr>
<tr>
<td>19.</td>
<td>Place the optrode under UV lamp for 15 minutes</td>
</tr>
</tbody>
</table>

![Figure 5.11: Flexible optrode fabrication procedure.](image)

On the left you can see the fabrication flow of the flexible optrodes. Some steps of the procedure correspond to an image to the right side. In image j you can see the finished prototype next to a needle with a 400 µm cross section. Image k shows a flexible optrode with a blue and a red LED being active at the same time, resulting in purple light being emitted. It is clear the flexible implant wire is able to take any shape. Note that the µLED chips are bare dies. Hence, it is important to use an electrostatic armband during the entire manufacturing process.
5.2 Control Module

The control module drives the \( \mu \text{LEDs} \). The control module is responsible for the individual control of the stimulation frequency for each of the \( \mu \text{LEDs} \), as well as for each LED’s individual adjustment of its illumination intensity. Since the light intensity of a LED is directly proportional to its forward current, the control unit must be able to output different current levels. To be able to optimally fulfill all the above requirements, following four different configurations have been considered:

1. Directly connecting a programmable LED driver to the \( \mu \text{LEDs} \).
2. Using a microcontroller to provide the stimulating frequency, and using a LED driver to set the output intensity.
3. Using a microcontroller together with a switch to modulate the power in order to set the stimulation frequency and the light intensity.
4. Connecting a microcontroller unit (MCU) directly to the \( \mu \text{LEDs} \).

Since we are strictly limited in the size of the receiver module, an approach that uses the minimum amount of components and space is preferred. Hence, our preference is for either option 1) the direct connection of a LED driver or 4) the direct connection of an MCU to the LEDs. However, after some research into available LED drivers, we realized that a large drawback of option 1) is that there are no programmable LED drivers available that include a non-volatile memory. Therefore, direct connection of an LED driver to the LEDs is not feasible, as the LED driver itself would require programming whenever it is restarted. On the other hand, one condition of option 4) is that a direct connection between the MCU and the LEDs is only possible if the MCU is able to source enough current to or sink enough current from the \( \mu \text{LED array} \).

Figure 5.12: MSP430G2553 typical high-level output current vs high-level output voltage [12].

As discussed in Section 5.1.2, the input current required for the stimulation at maximum intensity is 5 mA at 2.75 V. Therefore, to implement the control module, we make use of a Texas Instruments MSP430G2553 microcontroller [12], a microcontroller intended for use in ultra-low-power environments. This MCU is capable of providing 5 mA at a high-level output voltage of 2.75 V, for an input voltage of 3 V (refer to Figure 5.12). The MSP430G2553 has a number of low-power modes to achieve low power consumption. Also, this microcontroller integrates a number of peripherals, such as, of course,
the 16-bit CPU itself, registers and memory, hardware timers and an oscillator. This integration of components allows us to save valuable space on the wireless module by not having to include additional discrete components.

In our application, the microcontroller acts as an integrated LED driver and is programmed with a stimulation pattern that can be customized for a particular experiment. The code, which is included for reference in Appendix B, has been optimized for low-power operation, operating in low-power mode whenever possible. The MSP430G2553 is $4.85 \times 4.85$ mm in the QFN32 packaging. It includes non-volatile memory and, thus, retains its programming even if the device is restarted. This functionality is critical for an environment such as ours, where we cannot guarantee continuous power to the MCU. First of all, after implantation, the mouse will not be always residing in the wireless cage to power. Second, even when it is in the cage, the inductive link may not yield sufficient power at all times. If not for the non-volatile memory, a microcontroller restart would imply the need for reprogramming.

For the implemented stimulation program, the microcontroller controls two $\mu$LEDs, LED0 and LED1, which blink in an alternating fashion. A number of parameters are user-definable:

- The number of times each LED blinks before switching to the other LED.
- The waiting time between blinking LED0 and LED1, and between LED1 and LED0.
- The intensity with which an LED blinks, implemented using a PWM duty cycle.

For the implementation of these functions, the MSP430 launch pad is used, in combination with Texas Instruments Code Composer Studio, an integrated development environment for MSP-class devices. This combination allows for rapid prototyping of the embedded software, using the high-level C/C++ programming language for programming of the device. This board is used to:

- Program the MCU for testing and as a proof-of-principle. Here, for ease-of-use, a larger bulky plastic dual in-line packaging format ($20 \text{ mm} \times 7 \text{ mm}$) of the MCU is used. The evaluation board with the mounted MCU is shown in Figure 5.13a.

- To program the MCU for the actual assembly on the head-mounted module, we use the smallest packaging available, which is a QFN32 ($4.85 \text{ mm} \times 4.85 \text{ mm}$) packaging format. In order to adapt the QFN packaging to the DIP inputs of the launch pad, we use a QFN to DIP programming adapter. This is shown in Figure 5.13b.

![Figure 5.13: a) MSP430 launch pad with a bulky DIP packaged MSP430G2553, b) QFN32 to DIP20 adaptor.](image-url)
The stimulation program has been written to place the microcontroller into a low power mode as often as possible. Hence, the waiting routines are implemented using the hardware counters, during which the CPU can be put into a low power state. When the counter reaches a certain value, an interrupt is generated, waking up the CPU by putting it in an active state. By connecting the LEDs directly to a pin that is connected to the hardware counter, even the PWM duty cycle was implemented using these hardware counters. A final optimization is to run the microcontroller at a low clock speed of 1 MHz, which is still more than sufficient for our purposes, which further saves power. Again, refer to Appendix B for the actual source code. The programmed MCU consumes 0.99 mW in active mode and 66 µW in standby mode. The MCU is further tested with the actual µLEDs.

5.3 Regulator

In order to ensure that the microcontroller receives a stable DC input voltage, regardless of the coupling and load variations, a DC-DC converter is required in combination with storage element. In short, the DC-DC regulator must perform the following:

- Boost the voltage when the input voltage is at a lower level than the load voltage, by using a boost converter.
- Suppress the voltage level when the load demands a lower voltage than what is harvested, by using a buck converter.
- Store the energy when a surplus is available and supplement the harvested energy with stored energy when needed, by using either a conventional capacitor, a super capacitor, or a rechargeable battery.
- Match the voltage-to-current ratio of the input power to the level that is required by the storage element using a maximum power point tracker (MPPT).

![Figure 5.14: BQ25570 typical solar application circuit [13].](image)

To fulfill all the above requirements, an ultra-low-power boost charger and buck converter device, the BQ25570, is utilised. The BQ25570 chip is designed to efficiently extract even µwatts of power
that are generated by, typically, solar cells. The data sheet of the module can be found at [13]. This chip is 3.5 mm × 3.5 mm and requires external resistors to form a voltage divider that sets the output voltage level (refer to Figure 5.14). Since extra functions of this chip require extra external components, in order to save space, the unnecessary functions of the chip are not used. Hence, not all the pins and their external components are utilized. The external components are determined in such a way as to provide a constant output of 3 V to the microcontroller unit. The component values are calculated based on the data sheet instructions and are listed in Table 5.2.

![Figure 5.15: The BQ25570 evaluation module: BQ25570EVM-206. The added potentiometer can be used to tune the circuit for an output of 3 V.](image)

To evaluate this board and measure the precision of the calculations, the evaluation module BQ25570EVM-206 is used. In order to adjust the already existing components on the board to the ones calculated, an external potentiometer is added to the circuitry of the evaluation module. Figure 5.15 shows the addition of the potentiometers to the evaluation board. This potentiometer creates a voltage divider at the R9 to R10 junction and allows for tuning of the circuit. The circuit diagram of the evaluation module can be seen in Appendix A, Figure A.1. By tuning the potentiometer and monitoring the output signal, we can confirm that the calculated values are indeed correct and that they result in a constant 3 V output to the microcontroller unit.

The maximum threshold for the input voltage of the BQ25570 is 5.5 V. Therefore, in order to ensure that any peaks with a higher voltage level will not damage the chip, we use a Zener diode (PLVA653A) in parallel with the DC input. We chose a Zener diode with the closest Zener voltage equal or lower than 5.5 V, which turns out to be 5.3 Volt. Therefore, when the voltage across the inputs of the BQ25570 exceeds 5.3 V, the Zener diode shorts any surplus current to the ground, so that it will not reach the converter. Table 5.2 shows the component values that are used in order to adapt the BQ25570 to this project.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value (µF)</th>
<th>Component</th>
<th>Value (µH)</th>
<th>Component</th>
<th>Value (MΩ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>CIN</td>
<td>4.7</td>
<td>L1</td>
<td>22</td>
<td>R_{out2}</td>
<td>7.76</td>
</tr>
<tr>
<td>CREF</td>
<td>0.1</td>
<td>L2</td>
<td>10</td>
<td>R_{ov1}</td>
<td>5.62</td>
</tr>
<tr>
<td>CSTOR</td>
<td>4.7</td>
<td>COUT</td>
<td>22</td>
<td>R_{ov2}</td>
<td>7.32</td>
</tr>
<tr>
<td>CBYP</td>
<td>0.1</td>
<td>R_{out1}</td>
<td>5.24</td>
<td>CBAT</td>
<td>440</td>
</tr>
</tbody>
</table>

**Table 5.2:** Component values for the BQ25570.

The efficiency of the BQ25570 depends on the input current, input voltage, storage element voltage (VSTORE) and the temperature. At a fixed VSTORE, this efficiency drops dramatically at low input
voltages $V_{in}$ (refer to Figure 5.16). However, if the voltage is above 1.2 V, even with an input current as low as $100 \mu A$, the DC-DC conversion efficiency is above 90 percent.

![Figure 5.16: The DC-DC conversion efficiency of BQ25570 vs $V_{in}$][13].

### 5.4 Rectifier

The alternating magnetic field induces an AC voltage in the secondary coil. In order to make this AC voltage ready to be applied to the BQ25570 DC-DC converter, we must convert the AC voltage into a DC voltage. This conversion is performed using rectifiers. Rectifiers appear in different configurations, depending on application demands. For this project, we are mostly interested in that particular rectifier that not only matches our power requirements, but also contains as few components as possible, due to the stringent restrictions in size we are dealing with. Hence, we look into some simple configurations for our rectifier, which utilise the minimum number of components. The rectifier is supplying DC power to the buck-boost converter BQ25570 (discussed in the previous sub-section). The efficiency of BQ25570 increases with higher DC voltage. Hence, we choose for a rectifier that outputs a higher DC voltage to current ratio. Therefore, we select the full-wave rectifier with double voltage output (see Figure 5.17). This is also called a voltage-doubler rectifier. $R_{load}$ is the equivalent AC load resistor and equals to: $\frac{R_{DC}}{8} \times (1 + \frac{2V_{load}}{V_{DC}})$.

![Figure 5.17: The voltage-doubler rectifier.][1]

Since the power received at the secondary coil is relatively low and the transmission frequency is high, Schottky diodes are chosen. Schottky diodes offer a very low knee voltage and switching frequencies much higher than 13.56 MHz. The stabiliser capacitor must be large enough so that its time constant is much larger than the intervals between the consecutive AC peaks.
The load is connected to the rectifier output and in parallel with C1 and C2. The output ripple voltage equals: \( V_{\text{ripple}} = \frac{V_p}{(R_{\text{load}} \cdot f_{\text{ripple}} \cdot C)} \). \( V_p \) is the peak output voltage of the rectifier, \( R_{\text{load}} \) is the input impedance of the BQ25570, \( f_{\text{ripple}} \) is the ripple frequency, which equals twice the AC input frequency and \( C \) is the rectifier output capacitance. The minimum input impedance of the regulator module is 0.91 Ω. Hence, in order to keep the voltage ripple within 1% of the output voltage, we must have a capacitor larger than 0.41 μF. We choose for a larger capacitor value of 4.7 μF, since a larger value capacitance means a larger time constant, a lower ripple and hence a higher average output voltage. Please refer to Appendix A, Figure A.3 for the simulation result of the maximum output ripple of the rectifier when having an input voltage of 5 V. Table 5.3 contains the component type and values for the rectifier. The diodes have an average reverse capacitance of 0.7 pF for a voltage between 500 mV to 5 V.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1</td>
<td>B00340</td>
<td>C1</td>
<td>4.7 μF</td>
</tr>
<tr>
<td>D2</td>
<td>B00340</td>
<td>C2</td>
<td>4.7 μF</td>
</tr>
</tbody>
</table>

Table 5.3: Component type and values for the rectifier.

The wireless stimulation can be programmed for different light intensities and different stimulation frequencies. These changes cause variations in the load and, hence, the DC voltage level. The rectifier efficiency versus DC voltage is plotted in Figure 5.18.

![Figure 5.18: The AC-DC conversion efficiency of the voltage-doubler rectifier vs \( V_{\text{DC}} \).](image)

### 5.5 Power Module

Inductive links with a series resonance secondary and a parallel resonance secondary have respectively a current output and voltage output. As explained in the previous sections, the efficiency of the rectifier and the regulator is higher for high input voltages. Therefore, we choose a parallel resonance for the secondary LC tank. The design of the parallel LC resonance circuit (i.e., \( L_2 \) in parallel with \( C_2 \)) must fulfill the following conditions:

**Condition 1:** The total impedance of the parallel LC tank must equal infinity at 13.56 MHz.

**Condition 2:** The wireless link voltage gain at the secondary input is proportional to \( n = \frac{L_2}{L_1} \). Hence, a fixed value for \( L_1 \) implies that the voltage gain increases when \( L_2 \) increases. At resonance, it
holds that \( L_2 \times C_2 = \frac{1}{w_0^2} = 137.8e^{-18} \). Hence, it is more desirable to have a high value for \( L_2 \) and a low value for \( C_2 \).

**Condition 3:** The coil must have low losses and, hence, a high quality factor.

**Condition 4:** The coil must not cause a significant increase in weight and volume of the headboard module.

### 5.5.1 Parasitic Capacitance of the Coil

The coil does not only include positive reactance. Due to its non-ideality it also includes resistive losses and negative reactance. The negative reactance in the coil is called parasitic or stray capacitance. The parasitic capacitance in the coil includes the coil-to-core capacitance, the turn-to-turn capacitance and the winding-to-ground capacitance. In case of an air-cored coil, the coil-to-core capacitance can be considered negligible.

In the literature (e.g., [21], [9], [20]), the stray capacitance is calculated based on the turn-to-turn, winding-to-core and turn-to-ground capacitance of conductors with round cross-section. These calculations are thus based on the medium permittivity, winding pitch, wire cross-sectional shape, wire insulation material and the wire diameter. However, despite what is stated in the literature, my experience suggests that these parameters are insufficient to fully predict turn-to-turn capacitance, since the turn-to-turn capacitance should not have any significant contribution to the parasitic capacitance of the coil. And even if it does, its effect not only depends on the parameters considered in the literature, but it also depends strongly on the electric field strength in-between the turns. The formula \( C = \varepsilon \frac{A}{d} \) is valid in the specific case that there is a uniform electric field between the parallel plates of the capacitor for the entire area of \( A \). This, however, is not the case for a coil such as our receiver coil.

Figure 5.19: The magnetic field and electric field lines of two adjacent wires. Left: cross-section of wires carrying currents in opposite directions. Right: cross-section of wires carrying current in the same direction.

Figure 5.19 shows the electric field lines generated by two adjacent wires that are carrying opposite directional current, and two adjacent wires carrying current in the same direction. The turn-to-turn wires of our coil are similar to the situation in the image on the right: almost no electric field lines exist that lead directly from one wire to the other. If there is no electric field, there is no capacitance either, no matter how close together the wires are and no matter how large their surface area might be. This is true in the case of our receiver coil, as the difference in the distance from the feeding points of the coil is much smaller than the wavelength. Hence, winding the wires into a coil with zero pitch or higher will only affect its inductance, and not its parasitic capacitance.

In order to calculate the parasitic capacitance, we need to accurately attribute the turn-to-turn electric field to the stray capacitance. The only study that I found on this topic [16], formulates this attribution empirically, which results in the parasitic capacitance as follows:
\[ C_s = (4\varepsilon_0\varepsilon_{rx}/\pi)(1 + k_c(1 + \varepsilon_{ri}/\varepsilon_{rx}))/2]/\cos^2\psi, \text{ where:} \]

\[ k_c = 0.717439(D/l) + 0.933048(D/l)^{3/2} + 0.106(D/l)^2 \]

\[ p = \text{winding pitch} \]
\[ r = \text{wire radius} \]
\[ D = \text{rectangle diagonal} \]
\[ l = \text{coil height} \]
\[ \psi = \text{Arctan}[p/(2\pi r)] = \text{pitch angle} \]
\[ \varepsilon_{ri} = \text{relative permittivity inside the solenoid} \]
\[ \varepsilon_{rx} = \text{relative permittivity outside the solenoid} \]

5.5.2 Coil at Resonance

The equivalent circuit of the receiver coil is shown in Figure 5.20. The parasitic capacitance \( C_s \) and the coil losses \( R \) (explained and calculated in the next subsection) have an effect on the resonance frequency, since \((2\pi f_0)^2 = L^{-1}/(C_0C_s)\). The parasitic capacitance is affected by the coil shape, by the coil material, by the wire shape and material, by the coil core, and also by the environmental effects of nearby materials with a dielectric constant other than 1.

![Figure 5.20: Equivalent circuit of the receiver coil.](image)

For achieving the highest inductance-to-stray-capacitance ratio, it would be best to design the coil with a minimum parasitic capacitance, and construct the inductor in such a way so as to resonate at 13.56 MHz with its parallel parasitic capacitance. This would be ideal for the situation in which the receiver coil is in a well-defined, stable environment. In practice, however, the receiver coil assembly and environmental condition is not fixed. The dielectric constant of the medium outside the coil varies, based on the location of the receiver, as it could be near the transmitter windings, near another receiver of another mouse, and so on. The internal dielectric constant of the coil also varies with different assemblies of the electronics inside the coil. These variations change the parasitic capacitance of the coil and, therefore, affect the resonance frequency. For this reason, we would like to ensure that the parasitic capacitance is not dominant, through the addition of a capacitance that is at least a hundred times larger than the coil stray capacitance. This way, the coil resonates at a frequency at least ten times lower than its self-resonance frequency. This represents a trade-off between lower voltage gain and the addition of an extra component, and a more reliable resonator.

We choose a 31 AWG copper wire, since it is light-weight, but at the same time sufficiently robust. The inductance value is calculated as in Formula 4.25. In this calculation, we set the pitch to 0.4 mm (wire diameter (0.22 mm) plus twice isolation layer (0.08 mm) and 0.1 mm spacing due to practical
assembly error) and only vary the number of turns. This results in a different inductance value for each number of turns, and also in a change in coil losses. These variations affect the parallel capacitance value as the following equation shows:

$$C_2 = -C_s + \frac{L}{(2\pi f_0)^2 L^2 + R^2}$$

For each number of turns, R is calculated based on the calculations in the next subsection.

The change in number of turns also affects the parasitic capacitance as described in Formula 5.1. As we stated that C2 should be at least a hundred times bigger than C_s, we plot both C2 and 100 x C_s for an increasing number of turns, the result is shown in Figure 5.21. For a number of turns lower than 14, the coil will resonate at 13.56 MHz with its parallel capacitance C2, whereas its parallel capacitance is at least hundred times larger than its parasitic capacitance C_s. The parallel capacitance also includes the two times diode capacitance at the reverse voltage. Based on this, we design the receiver coil with 11 turns, which results in the coil to resonate at a frequency 12 times lower than its self-resonance frequency. The coil design parameters are listed in Table 5.4. Using the coil loss calculation from the next section, we can see that the frequency sweep simulation of the receiver coil input impedance in Figure 5.22 confirms the correctness of the calculations for the coil to resonate at 13.56 MHz. The coil is loaded with two capacitors of 47 pF, which due to inaccuracy resulted in a 92 pF.

As explained, this 92 pF parallel capacitor, in parallel with two diode reverse capacitance and the coil parasitic capacitance, brings the coil at 13.56 MHz resonance. The diode capacitance varies vs its inverse voltage. We have chosen for a low capacitance diode and in the calculations, have considered an average capacitance of 0.7 pF. However, it is important to know the effect of the diode reverse
voltage variations on the resonance frequency. The resonance frequency vs diode reverse voltage is plotted in Figure 5.23.

\[ V_{\text{reverse}} \text{ effect on the resonance frequency, arrow shows the working frequency 13.56 MHz.} \]

### 5.5.3 Losses in the Coil

In order to better understand the losses in our receiver coil, it helps to think of our coil as follows: 'A piece of conducting wire that conducts a high frequency, alternating current and is wound into a helical shape.' Since the coil is a conducting wire, due to its resistivity it experiences DC losses. The coil conducts a high frequency, alternating current, therefore, losses are introduced due to the skin effect. Finally, since the coil is wound into a helical shape, it suffers from the proximity effect.

Based on the different wire materials, the DC losses can be calculated as follows:

\[ R_{\text{DC}} = \rho l / A, A = \pi r^2 \]  \hspace{1cm} \text{(5.3)}

\( \rho \) is the resistivity in [\( \Omega \cdot \text{m} \)]

At high frequency, due to the skin effect, the area is reduced to the following:
5.6. Assembly and Measurement

\[ A_{\text{eff}} = \pi(2r\delta - \delta^2), \]
\[ \delta = \sqrt{\frac{\rho}{\pi f \mu}}, \]
\[ \mu = \mu_0 \mu_r, \text{where: } \mu_0 = 4\pi \times 10^{-7}. \]

Parameter \( \delta \) is the skin depth in meters, the depth at which the current density is reduced by a factor \( e \) (\( e = 2.71 \), better known as Euler’s number). The skin depth for copper wire at 13.56 MHz equals 17.7 \( \mu \text{m} \). Hence, the effective area for 31 AWG wire is \( 2.35 \times 10^{-8} \text{ m}^2 \). This results in an AC resistance of 3.15 \( \Omega \).

The proximity effect also increases with the AC resistance. It is explained by the proximity factor \( \psi \), which is defined as the ratio between the coil AC resistance and the AC resistance of the same piece of wire, but then not wound into a helical coil shape. It is extremely complex to describe the electromagnetic effects associated with the proximity effect. Hence, the value of \( \psi \) is most often obtained empirically. A study in 1947 [22] has defined the proximity factor for a set of inductors categorized by their length to width (\( l/D \)) and pitch to wire diameter ratio (\( p/d \)). Based on these empirical measurements (see Table A.4 in Appendix A), we can deduce that for our coil, the proximity factor equals 1.86. Therefore, the total losses in the coil equal: 5.86 \( \Omega \). This results in a quality factor of \( Q = 20.94 \), and a BW of 648 kHz.

5.6  Assembly and Measurement

The circuit schematic of the receiver module components is shown in Figure 5.24. In order to test

![Figure 5.24: Receiver module circuit.](image)

the performance of the complete receiver module, all the sub-modules are connected together with their evaluation boards using a breadboard (see Figure A.5 in Appendix A). For the assembly of the actual head-mounted module, a cross-shaped PCB is designed (Figure 5.25a). The PCB offers twice the area that is required. The other half is reserved for the wireless recording electronics. The PCB is further folded into a cube that acts as the platform for winding the coil (Figure 5.25b and c). In order to evaluate the calculations and optimize the coil design, we measure the impedance of the coil using a vector network analyser (VNA). The measurement results are listed in Table 5.5.

Based on the measurements, the coil impedance is modeled as: \( Z = 1589 - 1213j \). The input impedance of the RLC circuit of the resonator equals:
Table 5.5: Measured coil parameters

<table>
<thead>
<tr>
<th>f (MHz)</th>
<th>R (Ω)</th>
<th>X (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>13.56</td>
<td>1589</td>
<td>-1213</td>
</tr>
</tbody>
</table>

\[
Z = \frac{R}{((1 - (w^2)LC)^2 + (w^2)(C^2)(R^2))} + j \frac{(w(L(1 - (w^2)LC) - C(R^2)))}{((1 - (w^2)LC)^2 + (w^2)(C^2)(R^2))} \quad (5.5)
\]

Using this formula, we can deduce that the value of R is equal to 6 Ω and that C is equal to 99 pF. We observe that the measured parameters are very close to the calculations. However, the receiver coil at 13.56 MHz, has a non-zero reactance and is capacitive. The coil resonates at around 13.31 MHz. This could be due to fact that the PCB conductor plates are in close proximity of to the windings. In the calculations, we considered an air-cored coil. However, the conducting plate of the PCB can add a winding-to-core capacitance. In order to improve upon this, we use conductor etching liquid to remove the conducting layer of the outer surface of the PCB (Figure 5.25d). This new module now resonates at 13.55 MHz. The quality factor Q of the coil, based on the measurement results, equals 22.72. The bandwidth BW equals 597 KHz. The receiver box including all the components in Figure 5.24 and a thin layer of epoxy over the box, weights 1.1 gram.

Figure 5.25: Receiver cross-shape PCB.

### 5.7 Conclusions

In this chapter, the receiver module was designed. The receiver module consists of multiple main blocks that are individually designed with regard to their input requirements from the previous module and the output criteria to the next module. Finally, the whole system is assembled and optimized to operate at the resonance frequency. The power measurements of the module are covered in the next chapter.
Wireless Efficiency

The efficiency of the entire wireless system depends on the link efficiency, on the link-to-line efficiency and on the line efficiency. These can be defined as follows. The link efficiency $\eta_{\text{link}}$ is defined as the ratio between the power that is harvested at the secondary circuit and the power put into the inductive link at the primary coil. The link-to-line efficiency $\eta_{\text{link-to-line}}$ is the ratio between the useful power that arrives at the load and the power harvested at the secondary coil. Finally, the line efficiency $\eta_{\text{line}}$ is the ratio between the power delivered to the $\mu$LEDs and the useful power transferred to the load. The calculations and measurements of these three efficiencies are the topic of this chapter.

One might wonder why efficiency is important for an application which requires only milliWatts of power. However, the above mentioned efficiencies all result in quite severe losses, which, as they are stacked on top of each other, can easily result in very high input power requirements. Therefore, careful selection and optimization of each of these components results in a system that requires orders-of-magnitude less power than a naive approach.

6.1 Efficiency Measurement

Link Efficiency

The mutual coupling $M$ and coupling factor $k$ between the transmitter coil and receiver coil are calculated as:

$$M = \frac{\mu N_2 N_1 \pi (r_2^2)(r_1^2)}{(2(d^2 + r_1^2)^{3/2})}$$

with $r = \frac{a}{\sqrt{\pi}}$

$$k = \frac{M}{\sqrt{L_1 L_2}}$$

In the above formula, $a$ is the rectangle width in meters, $N$ indicates the number of turns, $d$ is the on-axis distance between the coils, and $k$ is the coupling factor between the coils. As explained in the previous chapter, due to the nature of the application, the coupling factor varies, depending on the particular location and orientation of the receiver coil. With no angular misalignment, the coupling factor equals 0.31%. The vertical and lateral misalignment within the 40 cm $\times$ 40 cm $\times$ 20 cm volume of the transmitter coil do not reduce the coupling factor.

The wireless link efficiency is affected by both the coupling factor and the load. The secondary load value has a damping effect on the link efficiency. The load as seen by the receiver LC tank is modelled as $R_{\text{load}}$ in parallel with C2. Figure 6.1 shows how the link efficiency behaves for a varying load, considering a maximum coupling of 0.31%. For a higher load (which means a low value for $R_{\text{load}}$), the efficiency drops dramatically. However, for an unloaded receiver (which means a high output resistance) the link efficiency is maximized.
Link-to-Line Efficiency

Not all the power harvested at the secondary side is usefully dissipated in the load. The harvested power at the secondary side is partially dissipated in the series resistance of the coil. This introduces, what we call here the link-to-line efficiency. Figure 6.2 shows the impact of the receiver coil loss on the total wireless link efficiency for various loads.

Total Inductive Link Efficiency

The total link efficiency $\eta_{\text{inductive\ link}}$ at a fixed coupling of 0.31 % obtained from the simulation is shown in Figure 6.3. For an $R_{\text{load}}$ of 430 $\Omega$ to 14.4 k$\Omega$, the link efficiency is above 0.28 %, which is half the peak efficiency.

On the one hand, there was no effect of lateral and vertical misalignment observed on the link efficiency. This is as expected, since transmitter coil has been designed to be able to provide an area of 40 cm $\times$ 40 cm $\times$ 20 cm irregardless of the vertical and lateral misalignment (refer to Chapter 4 for details). On the other hand, the angular misalignment between the transmitter and receiver coil has an obvious effect on the inductive link efficiency, as can be seen in the measurements shown in Figure 6.4. Only a limited number of angles has been measured, and for certain angles, it was hard to fix the receiver at that exact angle. These are shown by a line interval in the figure. Hence, we can conclude that the angular misalignment is the only misalignment that has a major effect on the link efficiency. This effect is notated as $\eta_\theta$. Therefore, the inductive link efficiency for any $R_{\text{load}}$ at any position and orientation of the receiver coil equals:

$$\eta_{\text{total\ link}} = \eta_{\text{link}} \times \eta_{\text{link-to-line}} \times \eta_\theta$$ (6.3)
6.1. Efficiency Measurement

Figure 6.3: Effect of $R_{\text{load}}$ on overall inductive link efficiency at fixed coupling of 0.31%. The horizontal green line indicates the load at which half of the peak efficiency is provided.

Figure 6.4: The measurement result of angular misalignment on the inductive link efficiency ($\eta_\theta$).

Line Efficiency

The power efficiency further decreases due to the losses in the rectifier, in the regulator, and in the microcontroller blocks. Figure 6.5 shows the circuit diagram for the efficiency measurement of the individual blocks. Figure 6.6 includes the measurement results of line efficiency for these blocks for varying input power, as well as the total line efficiency. We observe that for an input power above 4 mW, the line efficiency is above 50%.

Figure 6.5: Circuit diagram for the measurement of the efficiency of rectifier block, the regulator block and the microcontroller block.

Finally, the total wireless system power efficiency for any $R_{\text{load}}$ at any position and orientation of the receiver coil equals:

$$\eta_{\text{total}} = \eta_{\text{total link}} \times \eta_{\text{line}}$$  \hspace{1cm} (6.4)
Figure 6.6: Line efficiency versus input power. In this figure: blue: rectifier efficiency, green: regulator efficiency, red: MCU efficiency and black: total line efficiency. The colors match the corresponding blocks in Figure 6.5.

Figure 6.7: Magnetic field strength $H$ (A/m) within the transmitter coil box, along the width of the transmitter coil, for different RMS values of current (A) fed into the transmitter.

Thus, the power delivered to the optogenetics $\mu$LEDs equals: $P_{secondary} = \eta_{total} \times P_{primary}$. Besides increasing the wireless link efficiency, in order to harvest more power at the receiver, the input power power into the transmitter coil can also be increased. The increase of the power to the transmitter coil results in an increase in the EM field, as the magnetic field strength $H$ of a coil is directly proportional to the input current through the coil. Figure 6.7 demonstrates the magnetic field strength $H$ for different input currents (root mean square RMS) along the axis of the transmitter coil. The figure shows that even for an input current of 10 A, the magnetic field strength is orders of magnitude below what would be considered harmful, for example through neural excitability or by driving neural activity [29].

With an input current of 0.5 A into the primary coil, an average inductive link efficiency of 0.28 %, and an angular misalignment of 45 degrees, 8.5 mW of power can be delivered to the $\mu$LEDs.
6.2 Conclusions

This chapter included the simulation and measurement results for the power efficiency of the wireless optogenetics system. The amount of useful power harvested at the secondary coil depends on the inductive link efficiency, as well as on the link-to-line efficiency, since not all the power that arrives at the secondary side can be usefully dissipated by the load. The inductive link efficiency and link-to-line efficiency were plotted for different loads. The power efficiency of the rectifier, the regulator, and the MCU, together called the line efficiency, is also measured for various input power levels. Furthermore, measurements confirm the effect of the angular misalignment on the total efficiency. Finally, for various input currents fed into the transmitter coil, the magnetic field strength within the transmitter coil was shown.
CONCLUSION AND RECOMMENDATIONS

To overcome the drawbacks that a tethered neurological research setup suffers from, such as limiting the types of studies that can be performed and reducing the animal’s welfare, in this thesis, we have designed a wireless power transfer for optogenetic stimulation of freely moving rodents. This thesis report explains the rationale and methods used to develop this wireless power harvesting system.

Contributions

During the work of this thesis, the initial project was broken down into a number of subcomponents: the wireless power transfer link, the head-mounted module, and the optrode implants. In the design of these components, a number of contributions were made:

- One of the initial use cases of the optogenetics stimulation setup will be for the detection and suppression of epileptic seizures. Two methods for the automatic detection of grand-mal seizures were proposed. This automates a tedious task for the researchers, as empirical observations of the mice’s behaviour is a time-consuming and, moreover, error-prone method for determining the onset and end of the seizures. It also makes it possible to achieve real-time detection of the seizure.

- A variety of wireless power transfer methods were investigated for their suitability to this application. In the end, power transfer through an inductive link was chosen, as it seems to best fulfill the requirements of our application. The transmitter coil and the receiver coil were designed in such a way so as to conform to the application requirements, including the choice for parameters such as coil material, wire type, and shape. Especially the receiver module has stringent requirements on size and weight, as it is to be mounted on the animal’s head. The end result of the design and optimization of this inductive link, is a link that is able to provide power throughout an area of 40 cm × 40 cm × 20 cm, unaffected by lateral or vertical misalignments. The link is, however, susceptible to angular misalignment.

- The receiver module is made of a foldable PCB of 1 cm × 1 cm × 1 cm. The receiver coil is wound around the PCB box that includes the head-mounted electronics. It consists of a number of parts: an optical module, a controller module, a regulator, a rectifier, and a power module. The combined function of these modules is to guarantee a stable power supply to the optogenetics optrodes.

- Finally, a significant contribution is the creation of the optogenetic optrodes. Novel μLED mounting techniques were used. The use of a μLED array greatly improves the power efficiency, as the traditional LED-to-optical-fiber coupling is accompanied by large losses in light intensity. Moreover, a single μLED array is able to replace a number of optical fibers, resulting in a less-invasive procedure if multiple stimulation sites are required.
7. Conclusion and Recommendations

**Recommendation**

There are certain recommendations that would allow the current wireless setup performance to be improved even further. These suggestions are listed below in no specific order:

- Self-resonance makes the coil more sensitive to the environmental conditions. Therefore, it is suggested to choose a lower frequency band (below 130 KHz) for resonance. This way, when we lump the coils with large capacitors, the coil becomes much less sensitive to the environment impacts on coil parasitics. Hence, a more stable transformer performance, and even a more stable receiver coil is obtained. Designing the coils for a lower frequency range requires changes in the design choices in order to obtain the maximum efficiency possible. For example, for a lower frequency, Litz wire has a superior performance compared to solid wires.

- The transmitter coil includes thirty turns of enameled copper wire, which is fixed in place using screw terminals on the edges of the box. In order to make the wires straight, the wires were stretched and then screwed in the terminals. However, after three months, the copper wires became longer and looser. This caused the coil not to function at resonance any more. Temperature change and mechanical stretch could be the reasons for this undesirable effect. To overcome this effect, a better fixation of wires is required. For example, where the wires can fit in hollow spaces and glued in between two layers of PMMA.

- The current topology of the transmitter coil is superior to the flat spiral coils, as it is less sensitive to the lateral and vertical misalignments. This topology, however, constrains the experimental environment to this cage only. If one requires the wireless power to be available inside different cages or under various experiment setups, a flat spiral coil would be preferred. An interesting idea is to develop puzzle of smaller flat coils that can easily be connected and provide a wireless area in any shape, and as big as required.

- The developed wireless optogenetic stimulation setup in this thesis can be used for long term stimulation on freely moving rodents in animal experiments. In order to use this stimulation on demand, or in other words, to reach a closed loop recording and stimulation setup, this project must be combined with the wireless recording from the head-mounted module to outside the cage (developed by ir. Ide Swager), as well as wireless stimulation data transfer from outside of the cage to the head-mounted stimulation module.

- The optical and the temperature-related effects of the implanted µLEDs can be studied further in order to understand their effect on the neural tissue.

- To obtain focal stimulation in the optrodes including multi-µLEDs, µ-scaled lenses can be constructed on the µLEDs, which focus the light beam.


Figure A.1: The evaluation module BQ25570EVM-206.
Figure A.2: The Gerber file of the PCB layout of the needle optrodes. In this figure: red is the conducting layer; blue indicates holes; white is the PCB non-conducting body. a) The single optrode mounted on the PCB. The needle is attached to the optrode body, where the connections to ground and three anodes sit. In order to adjust the needle to the right length and in order to place the μLEDs in the right position, we have used the conducting layer to pattern a ruler on the PCB. The first red arrow distance to the optrode body is 4 mm. The distance between each two successive arrows is 1 mm. This distance is further divided into 200 µm sections by the shark teeth pattern. b) The 12x10 cm PCB board.

Figure A.3: Voltage doubler rectifier maximum output ripple for an input of 5 V_{ac}. 
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**Figure A.4:** Proximity factor $\psi$ [22].

**Figure A.5:** Receiver testing module on breadboard. Blue circle: rectifier, red circle: the dc-dc converter electronics, black: control unit and green: $\mu$LED.s
//***********************************
// LED blinking example
// - repeatedly blinks two LEDs (or ports) after each other
// - use the defines to modify behavior:
// LED_PWM_CYCLE_INTERVAL duty cycle = LEDx_PWM_DUTY_CYCLE/LED_PWM_CYCLE_INTERVAL
// LEDx_PWM_DUTY_CYCLE length of one duty cycle interval for LEDx
// LEDx_BLINK_COUNT how many times to blink LEDx
// WAITx_COUNT how long to wait between blinking LEDs
// WAIT_CYCLES number of cycles before interrupt is thrown
//
// Note:
// - the MSP430 only allows specific pins to be connected to
// TimerA0.1 (see data sheet). Pins that can be used are P1.2,
// P1.6. Therefore, the red LED on the Launchpad (P1.0) is not
// affected by the duty cycle setting. The green LED (P1.6),
// however, is.
// - the main timer run at 32kHz, an interrupt is generated every
// WAIT_CYCLES cycles
//
//***********************************

//***** Includes ***************
#include <msp430g2553.h>

//***** Prototypes ***************
void initClocksTimers(void);
void initGPIO(void);
void disablePWMonPin(int pin_out);
void enablePWMonPin(int pin_out, int duty_cycle);
void waitForInterrupt(void);

int timer_multiplier;

//***** Defines *******************
#define LED0_PWM_DUTY_CYCLE 100
#define LED1_PWM_DUTY_CYCLE 100
#define LED_PWM_CYCLE_INTERVAL 100
#define LED0_BLINK_COUNT 1
#define LED1_BLINK_COUNT 1
// how long to wait between blinking LED0 and LED1
#define WAIT0_COUNT 0
// how long to wait between blinking LED1 and LED0
#define WAIT1_COUNT 0
// the number of cycles before an interrupt is thrown
#define WAIT_CYCLES 60000
#define TIME_MULTIPLIER 20
// define to which pins the LEDs are connected
#define LED0 (BIT0+BIT2)
// PIN1.2 and PIN1.6 should be used (refer to data sheet)
#define LED1 BIT6
// use low power mode (1=enable, 0=disable)
#define USE_LPM3 0

//***** Global Variables*************/

//***** Functions *******************
void main(void)
{
    // Initialize GPIO
    initGPIO();

    // Initialize clocks and interrupts
    initClocksTimers();

    // Declare variables
    int i;

    while(1) {
        // Blink LED0 for LED0_BLINK_COUNT times
        for (i=LED0_BLINK_COUNT; i>0; i--) {
            // Set LED0 (use specified duty cycle)
            enablePWMonPin(LED0, LED0_PWM_DUTY_CYCLE);
            waitForInterrupt();
            // Reset LED0
            disablePWMonPin(LED0);
            waitForInterrupt();
        }

        // Wait for WAIT1_COUNT times
        for (i=WAIT1_COUNT; i>0; i--) {
            waitForInterrupt();
            waitForInterrupt();
        }

        // Blink LED1 for LED1_BLINK_COUNT times
        for (i=LED1_BLINK_COUNT; i>0; i--) {
            // Set LED1 (use specified duty cycle)
            enablePWMonPin(LED1, LED1_PWM_DUTY_CYCLE);
            waitForInterrupt();
            // Reset LED1
            disablePWMonPin(LED1);
            waitForInterrupt();
        }
    }
}
// Wait for WAIT0_COUNT times
for (i=WAIT0_COUNT; i>0; i--) {
  waitForInterrupt();
  waitForInterrupt();
}

// Repeat
}

// Timer A0 interrupt service routine
// wake up the processor after an interrupt
#pragma vector=TIMER1_A0_VECTOR
__interrupt void Timer_A (void) {
  __disable_interrupt();
  timer_multiplier++;
  if (timer_multiplier > TIME_MULTIPLIER) {
    #if (USE_LPM3==1)
      // wake up from low power mode
      _BIC_SR(LPM3_EXIT);  
    #else
      // wake up from low power mode
      _BIC_SR(LPM1_EXIT); 
    #endif
    timer_multiplier=0;
  }
}

// turn pin (pin_out) off
void disablePWMonPin(int pin_out) {
  // disconnect from timerA0
  P1SEL &= ~(pin_out);
}

// turn pin (pin_out) on with duty cycle of (duty_cycle/LED_PWM_CYCLE_INTERVAL)
void enablePWMonPin(int pin_out, int duty_cycle) {
  // CCR1 PWM duty cycle
  TA0CCR1 = duty_cycle;
  // connect to timerA0 output
  P1SEL |= (pin_out);
}

// Set pins to output direction and initialize low
void initGPIO(void) {
  P1DIR = 0xFF;  // All P1.x outputs
  P1OUT = 0;    // All P1.x reset
  P2DIR = 0xFF;  // All P2.x outputs
  P2OUT = 0;    // All P2.x reset
  P3DIR = 0xFF;  // All P3.x outputs
  P3OUT = 0;    // All P3.x reset
}
// initialize clocks: stop watchdog and initialize
timer A0, timer A1
void initClocksTimers(void) {
    timer_multiplier=0;
    // Stop WDT
    WDTCTL = WDTPW + WDTHOLD;
    // Tell the 430 to run at a calibrated
    BCSCTL1 = CALBC1_1MHZ;
    // Clk speed of 1MHz
    DCOCTL = CALDCO_1MHZ;
    // timer A0 is used for the PWM:
    // just creates a continuous PWM pulse
    // PWM Period
    TA0CCR0 = LED_PWM_CYCLE_INTERVAL-1;
    // CCR1 reset/set
    TA0CCTL1 = OUTMOD_7;
    #if (USE_LPM3==1)
        // Set the timerA0 to ACLK, up mode
        TA0CTL = TASSEL_1 + MC_1;
    #else
        // Set the timerA0 to SMCLK, up mode
        TA0CTL = TASSEL_2 + MC_1;
    #endif
    // timer A1 is used to time the waiting interval
    TA1CCR0 = WAIT_CYCLES;
    // generate an interrupt every (WAIT_CYCLES/32k) seconds
    TA1CCTL0 = CCIE;
    #if (USE_LPM3==1)
        // Set the timerA1 to ACLK, up mode
        TA1CTL = TASSEL_1 + MC_1;
    #else
        // Set the timerA1 to SMCLK, up mode
        TA1CTL = TASSEL_2 + ID_2 + MC_1;
    #endif
}

// go into low power mode and wait for interrupt
void waitForInterrupt(void) {
    __enable_interrupt();
    #if (USE_LPM3==1)
        // Enter LPM3
        __bis_SR_register(LPM3_bits + GIE);
    #else
        // Enter LPM1
        __bis_SR_register(LPM1_bits + GIE);
    #endif
}