A Self-Tuning 100 Watt Wireless Power Transfer System

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ABSTRACT

This thesis presents a new method of controlling wireless power transfer suitable for highly resonant magnetically coupled systems. An application of this system is unattended autonomous operation such as recharging of autonomous underwater vehicles or underwater sensor networks. Special attention is given to maximizing power transfer even when there may be spatial variations in transfer distance, which shifts the resonance peak frequency and hence requires automated control.

An automated system comprised of a 100 watt switching power amplifier coupled to a frequency controller is designed and implemented. The desired operating frequency is determined by quantification of the real-time AC power supplied to the resonant transmitter. The control system is preset to select operation at either of two selectable modes inherent to the resonant structure. The implemented system can operate underwater, requires only DC voltage inputs and operates over a range of distances while self-tuning to peak power transfer.
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1 Introduction

Wireless power transfer (WPT) has been an alluring application of electromagnetic principles since the formulation of Maxwell’s equations in the second half of the 19th century. Nikola Tesla first demonstrated wireless power transfer with his radio frequency resonant transformer in 1891 [1] and even attempted to realize long range power transmission via the Wardenclyffe tower, which due to mounting debts and loss of funding was never completed. Tesla’s works fell into relative obscurity after his death and it was not until many years later that microwave power resurrected the potential for practical wireless power transfer. In 1975 W.C. Brown and Richard Dickinson transmitted 30kW over a distance of one mile(!) with a 450kW input by means of a 2.388GHz microwave link [2]. As experiments continued all over the world, some took on a more practical nature; in 1987, a group at the Communication Research Centre (CRC) in Canada transmitted power to a model airplane flying at an altitude of more than 150 meters via a 10kW microwave beam [2].

Low power wireless transfer has also been of interest in the biomedical community for decades. A concrete and practical application of wireless power transfer, powering subdermal implants in medical patients, could see huge benefits from this technology. This use case is very important because it would solve the issue of requiring wires to run through the skin or repeated surgical procedures to replace drained battery packs but its importance is matched by the challenge in ensuring patient safety. As early as 1971 an inductively coupled (transformer) power transfer scheme was detailed [3] that delivered 1kW to a subdermal receiver implanted in the chest of a dog with less than 5°F temperature rise in the surrounding tissue. A few years later in 1977 a comprehensive procedure for designing implant wireless power transfer setups based on desired parameters such as frequency, distance, etc. was described [4]. Optimizing such an implementation is key to reducing losses and increasing safety and thus this is currently a very important topic in research.

With time the need for easy power transfer only grows. The arrival of the 21st century and the complete penetration of energy hungry gadgets into everyday life has given increased momentum to the efforts aimed at increasing the efficiency, safety and practicality of wireless power transfer. The technology has also finally begun its commercialization with the WiTricity project from MIT that paved the way for the eponymous company. Today they offer a number of wireless power transfer solutions aimed at handheld devices and claim to be working with the
automotive industry to bring wireless charging to the world’s growing fleet of electric vehicles [5]. Wireless charging pads in coffee shops are becoming ever more mainstream and cheap smartphone attachments for taking advantage of these spots are readily available. The improvement of wireless power transfer promises to bring both convenient charging to the consumer (gadgets and cars) and make energy transfer possible where it was previously exceedingly difficult to do so (through non-air media or inside the human body.)

Although this emergent technology is making waves and is alluring to the public due to its seemingly magical nature, the majority of commercial wireless power solutions are of a simple design that is intended to work at a fixed frequency and with very limited dynamic variation in transfer distance. This limits applications of the technology to simple setups and makes it more suitable as a show item than a practical and robust power transfer method. If wireless power transfer is to truly rival the wire it must become much more responsive, adaptive and intelligent.

The system described in this thesis is a step in the direction of smart and autonomous wireless power transfer. As will be shown in subsequent sections there is a clear relationship between the geometry of the wireless power link and the power transmission profile. The system described here is able to read this power profile and adapt its operation to maximize power transfer, all fully autonomously. An added strength of this system is that it is assembled from off-the-shelf components sourced from mainstream electronics suppliers and relies on basic principles of electromagnetism and energy conservation for its operation.

In the following section a brief overview of wireless power transfer schemes is presented and then notable existing research in the field is summarized. Subsequently the thesis moves on to present and explain in detail the system’s design and verify its operation. The thesis concludes with specifications and complete technical documentation necessary to replicate and operate this wireless power transfer system.


2 Overview of Wireless Power Transfer Schemes

A wireless power transfer system has three basic modules that bring power from the source supply to the load: a high frequency power amplifier, driver circuitry and the wireless link. The power amplifier must apply a purely AC voltage waveform to the drive side of the wireless link. The signal should not contain any DC because DC voltages will not be passed by the reactive components that make up the wireless link in accordance with the Maxwell-Faraday equation [6] which requires alternating magnetic fields. In addition, applying a DC voltage component to a system that uses inductive coupling would result in unbounded current growth through the magnetizing inductance of the driven coil. The drive, or primary, side of the wireless link thus accepts the AC current supplied by the source and induces an alternating voltage in the load side of the link. The induced voltage is accompanied by a current which together bring power to the load connected to the load side of the transfer link.

2.1 Magnetically coupled structures

There exist a number of different structures that can wirelessly transfer power. A few types that are related to the work in this thesis are summarized below.

2.1.1 Two coil non-resonant inductive coupling

The AC power source drives a source coil that creates alternating magnetic fields that are linked by the load coil (Figure 1.) The coils are not resonant externally and have reduces magnetic coupling and low efficiency. This setup is essentially a transformer with its ferromagnetic core removed.

Figure 1: Two coil non-resonant structure
2.1.2 Two coil discrete resonant inductive coupling

The source and load coils from the two coil non-resonant structure are connected in parallel with discrete capacitors which form parallel resonant LC tanks (Figure 2.) The AC power source is connected to the transmitter LC tank and the receiver LC tank supplies power to the load. The discrete parallel LC tank has a quality factor, $Q$, of $R \sqrt{\frac{C}{L}}$ which governs how efficiently the tank can store energy. Efficient storage yields efficient power transfer because the magnetic fields created during peak inductor energy storage (peak current) carry energy to the load side. The parallel configuration’s $Q$ factor is proportional to the parallel resistance and the main contributor to this parameter is the power source’s source impedance. Thus a low source impedance actually lowers the Q factor and hinders power transfer for this parallel circuit. This is shown in figure 3 — although a larger source impedance $R_s$ decreases the magnitude of the Norton equivalent current it simultaneously increases the $Q$ of the parallel RLC which is necessary for efficient power transfer.
2.1.3 Four coil inductive resonant coupling

This setup modifies the two coil discrete resonant structure by adding a separate “drive” coil magnetically coupled to the transmitter’s LC tank and driven by the power source. The receiver LC tank is likewise joined by a separate “source” coil that is connected to the load. The LC tanks makes no electrical connection to their respective drive/source coil but are tightly magnetically coupled to them. The tanks thus become series resonant circuits with a $Q$ factor of $\frac{1}{\pi \sqrt{\frac{L}{C}}}$ and will henceforth be assumed identical (equal L and C parameters.) The series resistance is composed of the combined ESR of the capacitive and inductive elements and can be made relatively small. Thus this setup achieves a high $Q$ factor that is mostly independent of the source and load parameters.

![Four coil inductive resonant structure](image)

*Figure 4: Four coil inductive resonant structure*

The four coil setup is an evolution of the basic two coil necessary to deal with the load and source impedances that prevent the resonant circuits from achieving a high quality factor. Since the resonant circuits are not electrically connected to the source and load they only see the source or load impedance transformed by the turns’ ratio and their $Q$ factor is influenced weakly by these external parameters. The drive and source coils are made from a single turn for minimal impedance and, since they are not self-resonant, form voltage transformers with the windings on the LC tanks. The LC tanks should have a large $Q$ factor for optimal energy exchange between them. Since the $Q$ factor is inversely proportional to the ESR of the resonant loop, Litz wire (low AC resistance) and ceramic capacitors should be used.

Thus the four coil resonant structure is in essence the same as the two coil resonant one with two important differences:

1. The LC tanks of the four coil resonant structure are weakly influenced by the impedance of the power amplifier and load. This allows for much higher $Q$ factor.
2. The tight coupling between the single turn loop and LC windings creates a voltage transformer that amplifies the voltage and reduces the current in the LC tanks which lowers Ohmic losses.

2.2 Double resonant structures

Each of the resonant systems described previously will exhibit the phenomenon of “frequency splitting” which occurs when there is weak coupling due to a small magnetic coupling coefficient $k$ between the two resonant coils. Given two coils with inductances $L_1$ and $L_2$ the coupling coefficient $k$ is defined as $\frac{M}{\sqrt{L_1L_2}}$ where $M$ is their mutual inductance. Since $M$ is inversely proportional to the distance $d$ between the coils \cite{7} it is evident that $k$ is likewise inversely proportional to $d$.

A class of circuits called the double tuned filter also exhibits the frequency splitting phenomenon \cite{8}. These circuits can be viewed as fourth order filters that possess two pass bands whose center frequencies are determined by the coupling coefficient $k$ between the two constituent resonators. A double resonant system (of which the four coil structure is an example) can thus be analyzed as exactly such a circuit. Given a coupling coefficient $k$ between the two tanks, the distance (in frequency) of each tank from the tanks’ resonance frequency $f_0$ can be given by

$$\delta = \pm \frac{1}{2} \sqrt{k^2 + \frac{1}{Q_1Q_2} - \frac{1}{2} \left( \frac{Q_1 + Q_2}{Q_1Q_2} \right)^2}$$

where $Q_1, Q_2$ are the quality factors of each resonant LC tank and $\delta$ is the normalized frequency deviation from self resonance $\frac{f-f_0}{f_0}$. Thus it is seen that although the LC tanks have their own resonance frequency $f_0$ (here for simplicity taken as identical for both tanks) the system as a whole possesses two resonant frequency modes \cite{9}, even and odd, which, under the condition of high $Q$, are related to $f_0$ by the coefficient $k$ as

$$f_{\text{even}} = \frac{f_0}{\sqrt{1-k}}, f_{\text{odd}} = \frac{f_0}{\sqrt{1+k}}$$

The above relationship shows that changing the separation distance between the coils shifts the two frequency modes (also referred to as peaks because they yield optimal transfer) closer to or farther from the circuits’ resonance frequency. When the two sides are close, $k$ is large
and the peaks are farthest apart. Increasing the separation distance decreases $k$ and shifts the peaks closer together. At a certain distance the peaks merge and become one that is centered on the individual circuits’ $f_0$. Increasing the distance further still will cause the peak to remain in place but decay until the receiver structure disappears from the transmitter’s view entirely. This is shown in figure 5; the tanks’ resonance frequency $f_0$ is around 10.6MHz and $\delta$, or the peaks’ normalized distance from $f_0$, falls with increasing distance and decreasing $k$.

The following set of graphs (Figure 5) from [10] illustrates the frequency profile for different separation distances for a four coil structure with coils of 400mm diameter. The peaks unite at around 500mm and then decay. Although the graphs show only efficiency the absolute power at the load follows a very similar profile. The graphs show also that tuning is absolutely necessary to maintain useful operation since the peaks move across a frequency range that is over 10% of the LC tanks’ resonance frequency. Keeping the system running at a single frequency cannot maintain good operation for all configurations because of the narrow peak width compared to the range across which the peaks move.

![Figure 5: Four-coil efficiency vs. frequency charts for varying separation distances. The peaks move closer, unite and decay as the separation distance (DIS) increases. Image credit: [10].](image)

Although the four coil structure is useful for high efficiency and high power transfer its advantageous LC tank presents its own challenges. A high $Q$ factor is necessary for efficient power transfer since all the energy that reaches the load must pass through the tanks. A high $Q$ factor, however, comes with a price of high component stresses. A series resonant tank having quality factor $Q$ driven by an input voltage $V_{in}$ will create a voltage of $QV_{in}$ across its capacitive
element at resonance. It is therefore quite easy to create voltage levels in the kilovolt range across the resonant capacitors and the components must be appropriate for such stresses. Other factors such as current ripple and series resistance must also be considered.

Another drawback of a frequency splitting system is the resonant tank’s behavior at large separations. As the frequency peaks merge into one at the resonance frequency of the LC tanks the tanks’ reactive components cancel leaving only its resistance to oppose current. The coupling to the receiver side is likewise too small to present the load to the primary side. The impedance seen by the power source rapidly falls and the transmitter’s LC tank’s resistance consumes real power which translates to large currents being demanded of the power source and significant heating of the tank while at the same time less power being delivered to the load. In practice this manifests itself as increasing power consumption (rapidly increasing current supplied by amplifier) with poorer overall efficiency since the transmitter LC tank dissipates rapidly increasing amounts of power. This highlights the importance of very low resistance resonant tanks but the problem cannot in practice be completely solved. It is therefore important to develop a system that can recognize such a condition and respond appropriately so that the amplifier and transmitter LC tank are not damaged.
3 Objectives and Motivation

The primary objective of this work is to design, build and validate a proof-of-concept wireless power transfer system that can continuously monitor its performance and adjust its operation to maintain the highest possible amount of power transfer under the given conditions. The system should be able to see how much power is being transferred to the load without a direct connection to it (no wires or communication of any kind between the load and drive sides.) Since the condition that is most readily and obviously changed in a wireless power transfer setup is the separation distance between the source and load coils, the system should be designed to notice and respond to changes in this parameter.

From the above analysis of the four coil system it is seen that the frequencies of optimal power transfer depend on the coupling coefficient \( k \) which is set by the separation distance between the receiver and transmitter coils. This means that every physical coil setup sets a value for \( k \) which then determines which two frequencies yield the highest amount of power transfer. As the distance and orientation between the two sides of the wireless link change, the frequency peaks move. The designed system must be able to notice these changes in sent power and adjust the frequency to stay on the peak as the physical setup changes. This is the dynamic peak hunting part that the system must be able to continuously execute.

As mentioned above, however, the system should not blindly chase only the peak because the peak can move past the point where useful power transfer is still occurring. The peak will then still exist but the power will be dissipated in the LC tank instead of going to the load. The system should thus be able to recognize this condition, shut continuous power transfer off and enter a kind of sleeping state in which no power is being transferred but the system periodically checks to see if the receiver has returned to resume charging.

3.1 Application of self-tuning technology

The motivation for this work comes from the fact that the environment in which wireless power transfer must occur is not always stable and predictable. An example application of this system is in charging an autonomous underwater vehicle (AUV) without necessitating that it surfaces or returns to port. Such a vehicle, performing autonomous exploration of the ocean floor, can broaden its range and increase the time spent underwater if it can wirelessly charge itself by approaching and floating next to small, portable underwater charging stations. Because
the vehicle works in water it does not have a capability to rest on hard surfaces and thus would not be landing on these platforms, but rather approaching and floating above or next to them during charging. The motion of the water would cause the vehicle to move around and thus continuously change the geometry of the wireless power transfer link. In this environment a conventional wireless power transfer system without feedback would fail to deliver maximum power because it would not be able to adapt to this changing geometry of the wireless link. The system described in this thesis, however, would continuously monitor its performance and maintain its frequency to stay on or very close to the peak power transfer possible at the current conditions. In addition the system should be able to determine when the vehicle has left useful charging range and automatically shut itself down (again without communicating with the vehicle) and likewise wake itself up and resume power transfer when the vehicle has returned. This smart system would be an important step forward in increasing electrical vehicle range and improving interaction between these two fully autonomous systems.
4 Previous Work

A sizeable amount of research and design has been completed in this field and serves as the basis for the project described in this thesis. A plethora of resources exist that describe progress relating both to the physical coil setup and the operating dynamics of a WPT system.

Some of the earliest publications regarding this topic have been produced by the biomedical community. In 1971 Schuder, Gold and Stephenson [3] showed an inductively coupled (basic two coil) wireless power transfer system to transfer 1kW from a transmitter mounted on the skin to a subdermal receiver implanted in the chest of a dog. This system is impressive for it managed this nontrivial power level with less than a 5°F temperature rise in the surrounding tissue which, over the short time intervals that the system would need to recharge the battery of an artificial heart, may be acceptable given the benefits. The key to minimizing tissue heating is reducing losses and this system demonstrates that it is possible to do so even at considerable power levels. The above quoted power level is also much higher than the perceived requirement of 35W for a patient during exercise and thus tissue heating would be an even smaller concern in practice. Short distance power transfer using the basic two coil inductively coupled structure is thus believed to be relatively safe to tissue and chronic exposure to 350kHz fields delivering 50W through the dog’s chest appeared to produce no ill effects.

In 1977 the design process of a two coil resonant wireless power transfer structure for low power subdermal implants was shown by Ko, Liang and Fung [4]. The procedure uses input and output variables, as well as transfer distance, to design the optimal wireless power transfer structure. Since for subdermal implant charging applications the transfer distance can be assumed constant no dynamic capabilities are needed and the design process can produce successful results with a fixed distance value. Using the design process a 3.5MHz structure achieving 18% total transfer efficiency over a distance of 2 inches was designed and verified. The high frequency used by the system allowed the system to be made of smaller sized passive components which is highly advantageous if the receiver must be sealed inside the human body. This paper is also one of the first to connect high $Q$ factor to increased transfer efficiency and this idea has become a central theme in coil construction.

In 2009 Low, Chinga, Tseng and Lin [11] presented a high power WPT system that transmitted up to 300W using a dual-channel class E amplifier powering an inductively coupled (two coil) setup operating at 134kHz. The system achieved 75% end to end efficiency at the maximum
power level with nearly adjacent receiver and transmitter coils.

Aside from introducing a viable hardware setup for transferring power over short distances this paper highlights some key characteristics of such a WPT system. One of these characteristics is the reduction of the real part of the apparent load impedance when the resistive load increases in magnitude. This means that larger resistive loads actually look smaller to the transmitter and thus draw more power from the source side. However, more of this power is dissipated across the receiver coil and thus the overall efficiency falls. This reduced efficiency yields the fact that the source must supply higher currents to power a larger resistance which taxes the amplifier more.

Sample, Meyer and Smith [12] detailed in 2011 a low power wireless power transfer system that performs frequency tuning for dynamic adaptation. The system employs four coil operation with an LC resonance frequency $f_0 \approx 7.65$MHz driven by an RF amplifier. In order to perform peak adaptation the system uses a directional coupler to measure the incident and reflected power from the wireless power transfer link. Using conservation of energy the system determines the point of maximum power transfer by comparing the incident and reflected power to make an assumption about the power provided to the load.

The capabilities of this system are considerate since it intelligently closes the loop by monitoring power throughput and adjusting accordingly to maintain best conditions. The system transfers around 30W and the power capability can be expanded using a commercial or custom amplifier. Unfortunately the nature and precise construction of the controller are not detailed. Nonetheless this system shows that dynamic frequency tracking is viable and necessary for a smarter WPT system. This capability should thus be embraced by designers of such systems and an implementation of this basic yet powerful idea is presented in this thesis.

A system capable of tuning its frequency and performing impedance matching was implemented by Sample, Waters, Wisdom and Smith [13] in 2013. One of the system’s capabilities involves tuning $\pi$-matching networks on both the receiver and transmitter side of the wireless link. The system monitors the incident and reflected power on the transmitter side and the output power on the load side and, using 2.4GHz radio communication between the two sides, tunes the $\pi$-matching networks on both sides of the wireless link for maximum efficiency. This scheme is advanced but requires communication between the two sides which may not be practical in certain environments (underwater.)
In addition to impedance matching the system can also tune in the frequency domain by performing incident and reflected power measurements and tuning to the frequency that optimizes the amount of power flowing into the transmitter side. The system performs the power measurements at a few operating points around the present one and selects the one with maximum power. It is also capable of changing the frequency step size as it nears the peak for more accurate peak tracking.

Dang and Qahouq performed an extensive comparison between the canonical two coil and four coil setups [10] in 2015. The efficiency advantages of the implemented four coil setup are shown and empirical data showing the “frequency splitting” phenomenon inherently present in these coil systems is presented.

[10] also presents a peak efficiency versus separation distance graph that shows the difference between their two and four coil systems. Their results show that their two coil system has a higher peak efficiency for separation distances less than approximately 85% of the coil diameter. Its superiority falls, however, past this mark and the four coil’s peak efficiency becomes larger and extends to 125% of the coil diameter. This reinforces the four coil system’s status as the more favorable choice for wireless power transfer, especially with relatively small coils (where the distance of one diameter is easily exceeded.)

Finally, [10] shows the effects of moving one coil set off-axis relative to the other. As the distance between the axes of the coil sets increases the efficiency versus frequency curves behave identically to the scenario of increasing separation distance. Increasing misalignment shifts the frequency peaks closer together until they unite at the tanks’ resonance frequency and decay. The WPT system sees the same effect because both actions decrease the coupling coefficient between the two coil bundles. This also means that the controller can respond to both scenarios since, from its point of view, they are the same.

It should be noted that this article uses a somewhat unconventional design in that their single turn and multi-turn tank windings are of very different sizes. This is somewhat unfavorable for coupling because the fields emitted by the single turn drive element can loop back and produce a net field through the LC winding that is smaller than if the coils were the same size and wound snugly on a single coil former.

An interesting system that can wirelessly power multiple smaller receivers while performing dynamic tuning was presented by Waters, Fidelman, Raines and Smith [16] in 2015. This system
is very applicable as a commercial product sold for the home environment where multiple small
gadgets can be simultaneously charged by being placed on a single base station. The system
also utilizes a directional power coupler to measure incident and reflected power from the power
amplifier to the transmitter coil and seeks the frequency that maximizes their difference for
maximum power transfer. In addition to this approach the system also possesses OOB radio
capabilities that can be used for getting information about power transfer directly from the
target devices. Finally, every external device has independent impedance matching capabilities
and can do its own tuning to increase or decrease its received power.

Cao, Dang, Qahouq and Phillips [14] implemented in 2016 a strategy to dynamically adapt
to operating conditions by building a setup with multiple drive and source loops that can be se-
lected from in real-time to achieve the best performance. Their setup can adapt to both distance
and alignment variations by using two arrays of switches that select the transmitter and receiver
loops based on the power going into and coming out of the wireless link. During operation
the controller uses the “perturb and observe” [15] scheme to determine whether it is necessary
to change the drive or sense loop combination. This system achieves dynamic adaptation to
optimize efficiency but its operating points are limited by the fixed set of coils that are sup-
plied. In addition the system requires monitoring of the output which means that an additional
communication link would have to be established between the receiver and transmitter.

Blakiewicz, Jakusz, Jendernalik and Szczepański [17] showed in 2016 an application that
fuses the dynamic tuning and biomedical aspects in the form of a system capable of powering
a miniature endoscopic capsule that is traveling inside the patient. The challenges presented
by this use case are considerate in that the receiver is forced to be very small (1cm diameter
receiver coil) and the separation distances between receiver and transmitter are relatively large
and vary over a large range. The system works by transmitting at a fixed frequency from outside
the body from which the receiver capsule can extract variable amounts of power by selecting
one of a number of capacitances placed in parallel with the receiver coil. The resulting receiver
waveform is then rectified to power the load. The system selects the parallel capacitance to
use based on how much power it needs at the present moment. Too little power and it selects
the capacitance which brings its resonance frequency closer to the drive frequency to increase
induced voltage and thus power. Too much power and the resonance frequency is moved further
from the drive frequency for lower voltage and power.

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This system is quite unique in that it uses a smart receiver and a constant frequency transmitter instead of the conventional dynamic transmitter with a passive receiver. This is clearly necessary in this case, however, because the receiver’s performance is critically dependent on the amount of power received and thus the goal isn’t simply to maximize power to the fullest extent. This system also functions on a different time scale than those described above because its objective is not maximum power at all times but rather a steady supply of power that is appropriate for its activity level.

Thus it is well documented that wireless power transfer can be realized at good efficiencies and power levels but a passive coil structure is not enough. Different applications require different power transfer profiles and tuning is always a beneficial feature regardless of the application details. One implementation of a system that realizes this feature is presented in this thesis.


5 System Overview

The system proposed in this thesis consists of six main hardware modules (shown in figure 6.) The last section, Underwater capability, describes a set of additional hardware and requirements for underwater operation and does not fundamentally alter the design or behaviour of the first six modules.

1. Microcontroller
2. Power amplifier
3. Analog measurement module
4. Frequency generator module
5. Wireless power transfer (WPT) structure
6. Load
7. Underwater capability

The first four modules form a feedback loop that controls the amount of power being transferred to the WPT module and act to determine and set the frequency at which the maximum amount of power is being transferred. In the following description the load is assumed to be fixed and does not influence the control scheme.

Figure 6: Complete system diagram
5.1 Control loop

The control loop begins with the microcontroller which receives information about the amount of power that is currently being sent over the wireless link to the load and sets the best operating frequency under present operation conditions. The power amplifier receives the frequency and drives the WPT module with a 50% duty cycle AC square wave voltage waveform at the commanded frequency. The power amplifier is connected to the drive coil and the resultant voltage and current applied to this coil are measured by the analog measurement module. The analog measurement module processes the waveforms and produces two DC voltage values that the microcontroller uses to characterize the operating point. The first is a voltage that is proportional to the AC power that the amplifier supplies to the WPT structure and the second is produced by a peak detector that measures the peak of the AC current supplied by the amplifier to the WPT structure. The controller uses these two values to make decisions about the system’s operation.

In order to make the system function without direct communication with the load an assumption must be made that all power consumed by the drive side of the WPT structure will be sent to the load. Although this is not the case experiments have shown that maximizing the power supplied to the drive side will come very close to maximizing the power delivered to the load within a defined range of WPT geometries. This system will therefore aim to deliver maximum power, as opposed to maximum efficiency, possible in every configuration.

5.1.1 Analog measurements

At the core of this system is the ability to measure large voltage and current waveforms and convert them into small DC voltage levels that can be sampled by the controller. This is achieved by the analog measurement module (module 3) which quantifies the voltage and current waveforms applied to the WPT module and performs analog multiplication of these waveforms in real time. The result is then low pass filtered (time-averaged) which gives the real, or DC, power supplied to the wireless link \[18\]. All processing is done in the analog format away from the microcontroller because multiplying and low pass filtering two analog waveforms automatically preserves their relative phase information which is crucial to calculating the amount of power delivered by these two waveforms.
5.1.2 Frequency search

The microcontroller searches for the power peak over the frequency range by sampling coil power and current derived by the analog measurement module. The microcontroller does this search by dynamically stepping toward the optimal operating point via a hill climbing algorithm that uses the “perturb and observe” strategy. Once it has found the peak the controller remembers the frequency and peak coil power and then periodically compares the present level of power to the found maximum. If the power level changes by a certain amount (which is scaled with the absolute amount of power transferred) the microcontroller assumes that the peak has shifted and sets out to find it again.

The controller should not, however, simply follow the peak. There is a range of distances over which a peak exists but the receiver is too far (meaning that the magnetic coupling is too weak) for useful power transfer. The controller should be able to detect this condition and shut the system down to avoid wasted power and unnecessary component stresses and it does this by checking the coil current. A current larger than a preset value indicates that the receiver has left the acceptable range for charging and the amplifier is wasting energy by heating the LC tank.

It is also important to note that there are two different control conditions — one for initial system startup (“startup sequence”) and another for the remainder of time after the system has been turned on (“in service” loop or post-startup phase.) The initial startup condition is different from the post-startup because as part of the “in-service” loop the microcontroller checks present values against past ones. Since during initial turn on there are no past values the microcontroller must perform a default action that will allow it to quickly gauge the state of the world it just entered. The controller defaults to a preset frequency and uses information gathered at that frequency to find the relative location of the power peak — this allows it to quickly determine if the receiver is too far or close enough for effective power transfer.

5.1.3 Frequency synthesis

To change the system operating frequency the controller communicates with the frequency generator via an I2C interface. The frequency generator creates a 50% duty cycle waveform using an astable 555 timer circuit tuned by a digital potentiometer, the value of which the controller sets. Although any modern microcontroller is capable of producing a PWM waveform to
control the amplifier having a separate module is advantageous in terms of safety and frequency resolution. The frequency module’s output waveform connects to the amplifier and closes the feedback loop.

5.1.4 Control constraints

The system is subject to the following important constraints:

- The microcontroller cannot make any assumptions about the location of the receiver and the wireless link geometry (magnetic coupling) at any time and must make all decisions based only on present and past information provided to it by the analog measurement module and the user-programmed presets.

- The microcontroller must be able to detect when the receiver has moved outside the charging range and as a result shut itself down, both during initial turn on and during the subsequent long-term monitoring stage. It should likewise realize when the receiver has returned to within active charging distance and resume power transfer.
• The microcontroller should perform actions on a timeframe over which the outside world does not change “too much,” in that it should act on measurements as soon as possible after acquiring them. If too slow it may end up tracking a peak that has long since moved from its perceived location. For example, an initial idea for a tracking sequence was to perform a complete frequency sweep across the frequency module’s entire range and then lock onto the higher power peak. Although this works when the receiver is stationary it would fail if it is in motion since such a sweep would take a few seconds to complete and the frequency profile would be changing as data was collected. This would result in the controller getting stale data points and potentially making wrong decisions. Thus it is necessary for the system to always function rapidly “one step at a time” and avoid delays between acquiring data and acting upon it.

5.2 Wireless power transfer

Power is transferred wirelessly when the amplifier, which serves as a simple power gain of the drive signal from the frequency generator, applies an AC voltage waveform to the single turn drive coil of the transmitter side of the WPT structure. The energy is magnetically coupled (as in a transformer) to the transmitter LC tank which resonates and exchanges the energy with the receiver LC tank. The LC tank likewise magnetically couples with the receiver single turn coil which is connected to the load. The power amplifier has essentially instant response and provides a voltage amplification of the drive signal with sufficiently low output impedance to supply required current for the drive coil impedance.

5.3 Load

The wireless power transfer system’s load is the main power sink. For the purpose of design and verification of this system low resistance power resistors are used but the system will eventually be implemented in the field for the purpose of battery charging. Hence, for completeness, the receiver’s output AC voltage waveform is rectified using a diode bridge and large parallel capacitance so that a DC voltage is applied to the load resistance. This output rectification to DC allows the system to easily accept other DC loads such as the aforementioned battery.
5.4 Underwater capability

The system should also be able to perform wireless power transfer with tuning underwater because it is envisioned that such a system could be placed on the bottom of the ocean or river for the purpose of charging AUV’s tasked with exploring the body of water. Although the developed electronics and software can be used in water exactly as in air they must be properly protected from the water around them. For this purpose the electronics and wireless power transfer structure are enclosed in non-conductive watertight containers with the system connected to shore only via DC power cables. The electronics and WPT structure reside in two separate boxes (not counting the receiver, or AUV, which will be its own device) which are rated IP68 or NEMA6P \[25\] for prolonged immersion. Connections are made from shore to the electronics enclosure and between the electronics and transmitter enclosure using proper undersea cables and underwater connectors.

5.4.1 Underwater constraints

- **Water pressure** — For underwater testing the system will be submerged no deeper than a few feet which, while shallow, still exerts pressure on containers. The chosen containers must withstand this water pressure for extended periods (days) so that the system can be adequately tested.

- **Heat dissipation** — Since water application forces the electronics into a sealed container heat dissipation becomes a much bigger concern than in the air. The power amplifier must have a dedicated heat sink mounted flush against the inside of the container.

- **Electrical connections** — Connections to electronics inside the enclosure must pass through the enclosure. For this purpose special underwater connectors are required. In addition, holes must be carefully created in the enclosures so that the seal between each connector and the surrounding enclosure material is watertight.

- **Non-conductivity of materials** — The materials used in the enclosure containing the coils must be non-conductive to avoid induced eddy currents and losses.
6 System Design

The following section goes into qualitative detail about each of the modules listed in the previous section. For schematics and complete system specifications please see the appendix.

6.1 Microcontroller

6.1.1 Introduction

The microcontroller chosen for this system is the PSoC 5LP (Programmable System-on-Chip) from Cypress Semiconductor. The PSoC is a powerful chip that combines the traditional microcontroller hardware bundle (CPU, RAM/ROM etc.) with mixed-signal arrays of configurable digital and analog peripherals. The PSoC’s responsibility is to monitor the two analog voltages provided to it by the analog measurement module and make adjustments to keep the system operating at the peak when possible. The two analog DC voltages provided by the analog measurement module will henceforth be referred to as DCVP (DC voltage corresponding to AC coil power) and DCVI (DC voltage corresponding to AC peak coil current).

The PSoC 5LP was chosen for this system because of its advantages over regular microcontrollers. The PSoC possesses all the characteristics of a normal microcontroller but also has a wide range of analog and mixed-signal hardware that can be optionally programmed into the IC from the development software. For example, a fundamental building block of this project is a pair of ADCs (one each for measuring DCVP and DCVI.) The power ADC has 9 bits of resolution and the current ADC uses 8. Instead of resorting to sourcing (at best) a separate chip with two ADC modules (one of which has a non-standard number of bits) the PSoC can be programmed to include both ADCs in its on-board hardware which can be immediately connected to selectable analog I/O pins. Analog voltages can then be directly applied to the selected pins and convert them to integers that are ready for use. This powerful capability allows the system to save space and integrate these functions into a single module.

The PSoC microcontroller is also a very fast chip. A base 24MHz out-of-the-box clock allows dozens of ADC sampling cycles every iteration of the “in-service” loop to get accurate measured values from the DC voltage supplied by the analog measurement module. Because the system moves one step at a time the controller must perform many steps to find the peak and a fast clock allows redundancy in ADC sampling without noticeably compromising overall system response
6.1.2 PSoC hardware — Internal

This section details the internal, or “virtual,” hardware blocks that are programmed into the PSoC to serve various functions. Because these features emulate hardware components and are not part of the controller’s program code they are described here rather than the subsequent “Software” section.

The following hardware blocks are initialized and configured for this project.

- **9 bit Delta-Sigma ADC.** This ADC block is fed the DCVP analog voltage signal from the analog measurement module. The ADC runs in single sample conversion mode during which it waits for a software prompt to perform an analog to digital conversion.

- **8 bit SAR ADC.** This ADC block is fed the DCVI analog voltage signal from the analog measurement module and also runs in single sample conversion mode.

- **Master I2C Communication.** This block connects externally to the frequency generator module via the two signal lines SDA (data) and SCL (clock) to set the frequency of operation.

- **Variable clock source.** This clock module uses a low frequency base clock running at 1kHz to generate the interrupts necessary for long term monitoring of the system. It contains a divider register that can be filled with an integer value \( n \) for an output frequency...
of $\frac{1kHz}{n}$. The divider register can be manipulated by software in real time to change the frequency of the interrupts during program execution.

6.1.3 PSoC hardware — Pinout

This section documents the external hardware connections between the microcontroller chip and the rest of the system. Each entry corresponds to a pin connected to some internal and/or external hardware component. For actual pin numbers please see Appendix A.1.

- **ADC_P**. This analog pin has a hardware connection to the input of the Delta-Sigma ADC. Externally it is connected to the analog measurement module and brings DCVP into the microcontroller.

- **ADC_I**. This analog pin has a hardware connection to the input of the SAR-ADC. Externally it is connected to the analog measurement module and brings DCVI into the microcontroller.

- **AMP_IO**. This digital output pin is externally connected to the amplifier and is software-driven HIGH/LOW to turn the amplifier ON/OFF.

- **SCL_1**. This digital pin is connected internally to the I2C block and externally to the frequency generator module. It carries the clock signal that must be provided by the I2C master to the slave devices (digital potentiometer) on the channel. The pin is configured for open drain, drive low operation meaning that it must be externally connected through a pull-up resistor to $V_{cc}$.

- **SDA_1**. This digital pin is connected internally to the I2C block and externally to the frequency generator module. It carries the serial data signal between the master and slave devices on the channel. Like the above SCL_1 this pin operates as open drain, drive low.

6.1.4 Software — Overview

The PSoC’s responsibility is to monitor the DC voltage values produced by the analog measurement module and adjust the drive frequency to maintain operation at the power peak. Its operation is divided into two main phases. The first phase (startup) is the initial setup right after system turn on during which the microcontroller’s hardware is initialized with presets. The
controller makes initial measurements to determine if it will enter the “sleep” or “power” state. After this phase, if the power state is entered, the controller periodically checks operation and makes necessary frequency adjustments. If the sleep state is entered then the controller shuts the amplifier off and checks (relatively infrequently) whether it should enter the power state and transfer power.

The following software section abstracts away the hardware used for interacting with the world. For example, “measure DCVP” is written with the understanding that to do this the microcontroller must prompt the Delta-Sigma ADC, wait for the conversion to finish and read the result from its register.

6.1.5 Software — Startup

During the first phase, after initial system startup and initialization of the required system hardware, the controller must gauge the state of the world it just entered. Specifically, it must understand where the largest power peak is and whether it should tune to it (or if it’s too far away and the system should enter sleep state.) Experiments have shown that the second (i.e. higher frequency) peak gives higher power transfer at a lower efficiency and, since the goal is to maximize power transfer from the power source, this is the peak that the system will track. The controller does this by setting its operating point to a pre-determined frequency point $f_p$. 

![Figure 9: Controller decision tree during initial startup](image)
and finding the direction of increasing power transfer by comparing the values of DCVP at two frequency points adjacent to \( f_p \). This is defined as the local slope and is positive if DCVP is strictly increasing on the interval \( \{ f_{p-1}, f_p, f_{p+1} \} \), negative if strictly decreasing and undefined otherwise. Figure 10 shows an example DCVP versus frequency profile that is positive.

![Power vs. Frequency](image.png)

**Figure 10:** Example DCVP versus frequency curve. In this example the local slope is positive.

The predetermined frequency \( f_p \) at which the controller starts is a preset that is manually programmed by the user based on the parameters of the LC tank. As described earlier, the system’s power versus frequency profile consists of two peaks that converge or separate around a fixed center frequency when the transmission distance increases or decreases. This means that the frequency of the desirable second peak decreases as the transmission distances increases (receiver leaves.) Using an experimentally determined power versus frequency curve for a given LC tank, \( f_p \) is chosen so that the frequency location of the second peak \( f_s \) is larger than \( f_p \) when the receiver is “close enough” for good power transfer and smaller than \( f_p \) when the receiver is “too far” for good power transfer. This choice of \( f_p \) makes it easy for the controller to quickly find the transfer distance at initial turn on by measuring the local slope. If the local slope is negative then the controller notes that the receiver is too far and enters sleep state. If the slope is positive (or if the preset frequency \( f_p \) is at the peak) then the controller notes that the receiver is near and the peak should be found (or maintained if \( f_p \) is the peak frequency) by taking the system into the power state. After taking one of these two courses of action the controller notes its decision and enters the post-startup idling state.
6.1.6 Software — Post-startup operation — Overview

During the post-startup monitoring phase the microcontroller does not execute any main program code. Instead it idles and services a time-based interrupt, configured to trigger at specific time intervals, and functions as a state machine of the following variables:

- **max_val** — the value of DCVP recorded the last time the system found the second power peak
- **max_val_ix** — the index of the current operating point (i.e. the value sent to the digital potentiometer in the frequency generator)
- **hunting** — a Boolean value that is set to 1 if the system is not operating at the peak and 0 otherwise

Every instance of the interrupt service routine (ISR) consists of checking the current status of the system, making necessary adjustments and updating the above state variables.

6.1.7 Software — Post-startup operation — Power state

Suppose that the controller is awakened by the interrupt and starts the ISR and, checking the value of **hunting**, sees that it is 0 — at the conclusion of the previous ISR instance the system was operating at the peak. Time has passed, however, and the situation may have changed.
The controller thus samples \texttt{DCVP} corresponding to the power level (provided by the analog measurement module.) With a constant drive voltage and fixed load the only scenario that would change the power throughput is a changing distance between the receiver and transmitter. If the controller checks the power and it is the same as the value stored in \texttt{max_val} (with some allowed difference margin due to noise, etc.) then the controller assumes that the geometry is still the same and the system is still at the peak. The values of the state variables remain unchanged and the ISR ends.

It may, however, happen that the measured power differs from the saved by more than the allowed margin. If this happens the controller assumes that the peak has shifted due to changing transmission distance and must be re-found. It responds by setting the value of \texttt{hunting} to 1. At this point the ISR ends – the controller makes no attempt to find the peak immediately. The change has been noted and appropriate action will be taken during subsequent ISR instances.
When, 50ms later, the interrupt is triggered again the controller knows exactly what to do. It checks **hunting** and sees that it is 1 – the current power transfer is not optimal and the peak must be re-found. The controller thus sets out to find the new peak. To do this, it uses a dynamic hill-climbing scheme in which it makes steps (one per ISR instance) toward the new peak. The controller starts by finding the local slope in the same manner as when it did the initial check at $f_p$ — it sets the operating frequency incrementally to the two adjacent points and samples **DCVP** at each one. With a total of three power measurements the controller knows that it will get closer to the peak if it shifts its operation to the adjacent point with a higher **DCVP** than the present.

Before making a step in the direction of the peak, however, the controller also samples **DCVI** at the two adjacent points above. Maintaining peak operation at higher separations results in increasing amplifier output current with lower overall efficiency. Thus if the controller sees that the peak current value is at or above some threshold then it understands that the receiver is reaching the distance threshold for useful power transfer. The threshold current value is determined experimentally and usually corresponds to the current supplied by the amplifier during on-peak operation at a separation distance of slightly less than a coil diameter (the spatial cutoff point.)

Now the controller has all the information it needs to make a decision about its next operating point. The controller’s mission is to move the system to higher power operating points as long as the amplifier current permits. Thus its next operating point will be located at the adjacent point that brings higher power transmission (**DCVP**) at an allowed peak current (**DCVI**). If one of the two adjacent points satisfies these criteria the controller signals the frequency module to shift to that point. It then updates the state variable **max.val_ix** to contain the index of this new operating point. At this time the ISR is over. Since the controller made a frequency adjustment during this iteration of the ISR, **hunting** remains 1 and controller waits until the next interrupt trigger to repeat the process.

The controller repeats the above steps as long as the peak has not been found and the current remains within the allowed limit. At some point the controller may reach an operating point at which the 3-point local slope is discontinuous (a local maximum or minimum.) If the point is indeed a local maximum then the controller understands that it has found the new peak (if it is a minimum then the controller increments the operating point by 1 and **hunting** stays 1.) At
that point the value of `hunting` is changed to 0, the values of `max_val` and `max_val_ix` are updated to contain the present power level and operating index, and the controller resumes its normal monitoring cycle. Now, the next time the ISR is triggered, the controller will perform just the `max_val` comparison routine and then set `hunting` to 1 again to adjust the frequency if the peaks have moved.

6.1.8 Software — Post-startup operation — Sleep state

It may, however, happen that as the controller is stepping along to find the new peak it reaches a point at which `DCVI` exceeds the preset limit. This will happen if the receiver is pulling away and the peaks are merging together – the higher frequency peak shifts to lower frequencies while nearing the resonance frequency of the LC tank which results in a falling impedance as seen by the power amplifier. Since more and more power is wasted in the LC tank’s resistance and not sent to the load the system should be shut off. Thus when the controller sees that the current at an upcoming operating point is above the threshold it does not shift to that operating point but rather assumes that the receiver has left useful charging range and stops power transfer by shutting the amplifier off. It should be noted that the shutoff will happen at different frequencies depending on the input voltage to the amplifier — larger input voltages will reach the threshold current at a closer separation. Likewise a lower input voltage will allow the system to keep transmitting closer to the resonance frequency of the LC tanks (and thus at a larger separation.)

Having detected a departed receiver the controller does not change the operating point from the last valid one. The system remains at the operating point that is at the “edge” of the useful charging range so that power transfer can continue as soon as the receiver returns.

At this point the system is not transferring power but the controller still needs to be able to detect the return of the receiver so that it can resume normal power transfer when it happens. To check the status of the world the controller uses the same interrupt as during normal operation but with a modified clock for less frequent triggering (since the receiver presumably moves slowly and travels far there is no need to check for it too often.) When the receiver leaves the controller changes the interrupt’s clock’s divider to a larger number so that it now triggers once every few seconds. Otherwise the interrupt’s behavior remains completely unchanged. The controller still checks `DCVP` at two neighboring points and extrapolates the location of the peak. It then
checks DCVI at the point with the greatest DCVP and if the receiver has returned this value will be acceptable, allowing the controller to shift to that point. If the value of DCVI is still above the threshold the controller turns the amplifier back off.

Thus a tree of conditional statements (figure 12) programmed into a single interrupt can adequately control such a system in the long term. The state machine implementation allows the system to perform the same set of tasks with variations depending on the state. This keeps the program simple and allows for easier modifications.

From figure 12 it is evident that certain steps may be redundant in certain cases. For example, the controller turns the amplifier on and resets the interrupt frequency to fast at the start of every ISR instance. This is done because the controller does not actually know whether the receiver is too far and must start every ISR instance treating it as close. Then, if the measurements tell it otherwise, it turns the amplifier back off and sets the frequency to slow. This approach was chosen for simplicity and to avoid needing another state variable which would store the receiver’s status (close or far) as well as make further conditional statements unnecessary. Thus even if the receiver is close and the system is at the peak the function to turn on the amplifier will still be executed every time. This, however, has no negative impact and can safely be included.

6.2 Power amplifier

The power amplifier is the backbone of the power transfer capability of this system. Powered by a DC voltage level the amplifier is driven by the frequency generator and applies an AC voltage waveform to the WPT drive coil. The basic amplifier topology consists of four transistors in the H-bridge configuration [19] working as on-off switches which allows the transmitter to receive a full-wave AC voltage signal with no DC component.

The H-bridge works by turning on pairs of diagonally opposed switches while keeping the other pair off. The transistors function as switches, meaning that they are ideally either fully on or fully off with no time in between. Using Figure 13, switches A turn on and B off to bring \( V_{cc} \) to the transmitter during the first half of the cycle. During the second half, switches B turn on and A off such that the voltage across the transmitter is now \(-V_{cc}\). Making sure that switches A and B are on for equal time periods is important to prevent application of DC voltages to the drive coil.
Figure 13: H-bridge. Switches A and B alternate being ON to bring $+/– V_{cc}$ to the transmitter

The role of the switches is played by n-channel MOSFET transistors. The transistors must have good AC performance (low capacitances) so that they can be quickly switched to avoid transition losses. They should also have low on resistance although this characteristic is of secondary importance to the switching performance. Finally, the transistors must have a high enough drain-source voltage rating $V_{DS}$ because in the off state they must withstand the full input voltage $V_{cc}$.

Since the transistors must act as on/off switches they must be very quickly turned on and off and the transition time should be minimized. For this purpose dedicated gate drivers are used for every MOSFET. Two types of gate drivers are required for this circuit — high and low side. The high side gate drivers connect to the high side transistors (drain terminal connected to $V_{cc}$) and the low side drivers to the low side transistors (source terminal connected to ground.) The low side drivers are relatively straightforward and are simply amplifiers that can supply a large current for short durations to quickly charge and discharge the gate capacitance of their transistor. The high side drivers are more complex because in the on state they must hold a constant 12V across their gate-source capacitance which itself may be at a high potential above ground. The supply voltage $V_{cc}$ is thus constrained by the high side drivers because, when the high side MOSFET is on, its corresponding gate driver is directly connected to $V_{cc}$ through the open transistor as well. The current design’s choice of high-side gate drivers limits $V_{cc}$ to 80V.

Although the H-bridge is controlled by the frequency generator the signal cannot be directly connected to the gate drivers. The output TTL waveform of the generator drives two pairs
of transistors and must thus be split into two waveforms shifted by \( \pi \) radians relative to each other so that only one pair receives a HIGH at any time. In addition, because transistors do not switch instantly, a “dead-time” must be built in so that one transistor pair is securely off before the second pair turns on. Failure to do this can result in a “shoot-through” condition, or a short from \( V_{cc} \) to ground.

The splitting and shifting of the frequency generator signal, as well as dead-time introduction, are accomplished by a simple circuit utilizing a Schmitt trigger IC with some resistors, capacitors and diodes [20]. The IC contains a number of individual Schmitt trigger gates and thus a single chip is needed. The circuit is shown in Figure 14(a).

![Figure 14: (a) Circuit for splitting outputs and creating dead-time (b) Frequency generator output and the resulting switch signals from the output of (a)](image)

The key is to ensure that a LOW-HIGH transition of signal A occurs after a HIGH-LOW transition of B and vice-versa. This is achieved by the RCD combination that retards rising edges but passes falling edges unchanged. Since the capacitors control the voltage input to the final inverter stages of signals A and B the diodes must be included such that the capacitors charge instantly (so that the inverter output falls instantly) while slowing the discharge by means of the resistor (yielding slow rise of the inverter output.) The dead time is then simply the time delay between the fall of one signal and the rise of the other. The resistances necessary to achieve equal dead-times vary between inverter channels and trimmer potentiometers should be used for best results. The approximate capacitance and resistance values can be found by taking the desired dead-time and using the Schmitt trigger’s transition threshold (input) to calculate how
many RC time constants must pass for the capacitor to charge through the threshold. The $\pi$ phase shift between the two outputs is guaranteed by the two channels having an odd and even number of cascaded inverter stages.

Finally, the issue of amplifier shutdown must be addressed. Splitting the frequency generator’s output into two phase shifted signals means that both the HIGH and LOW state of the original signal correspond to one transistor pair being on and no input state can turn off all four. Thus the two split signals must be passed through AND logic gates so that the output signals to the drivers can be set to LOW (effectively turning off the amplifier) by the controller independently of the switching signal from the frequency generator. Each signal A and B is fed into its own AND gate with the other input terminal receiving a HIGH/LOW signal from the controller. To turn the amplifier off the controller must only apply a LOW and the AND gate’s output stays low regardless of the periodic drive signal. To turn the amplifier back on the controller applies a HIGH and the drive signal is passed through as is.

![Figure 15: Amplifier system diagram. The gate drivers are shown as triangles with the “DRV” label](image)

As in figure 13 the transmitter coil is connected across the amplifier outputs (O1 and O2 above.) This configuration applies a purely AC voltage waveform to the transmitter and permits power transfer.

The amplifier also has an LC low pass input filter. This filter is necessary to prevent high frequency components from polluting the DC supply that powers the H-bridge. The filter’s cutoff frequency should be several orders of magnitude below the switching frequency of the H-bridge so that only the DC component of the current passes through the filter and the supply does not have to source any high frequency AC components.
6.3 Analog measurement module

If the controller is the brain of the system, the analog measurement module (AMM) is the five senses. In order to track the peak the controller must measure the power provided by the amplifier but, being a fragile low voltage IC, cannot do so without extensive processing of these waveforms. The AMM performs the analog signal processing and outputs a low voltage DC signal that is proportional to the amplifier’s power output (DCVP) as well as a low voltage DC signal corresponding to the amplifier’s output current (DCVI). These low voltage signals can easily be sampled and understood by the controller which can then make decisions about the system’s operation.

6.3.1 Power measurement - DCVP

To measure the power and produce DCVP the AMM invokes the definition of power. The module replicates (i.e. produces an attenuated copy of) the amplifier output voltage and current waveforms, multiplies them in real time and produces the DC component of the result. This approach is optimal because not only are the magnitudes of the voltage and current components important but the relative phase is also. The analog multiplication and low pass filtering automatically includes the phase in the power calculation and paints a complete picture of the amplifier’s power output.

To replicate the amplifier output voltage the AMM scales the voltage levels present at outputs O1 and O2 of the amplifier and takes their difference using an instrumentation amplifier (INA).
The signals are brought from the power amplifier to the pre-INA resistor divider via 50Ω coax cables. Since the resistor divider loads the power amplifier’s outputs the resistances must be large enough to create a negligible effect compared to the parallel WPT structure. The INA’s output is then a scaled version of the voltage waveform across the drive coil connected to the power amplifier.

The AMM must also produce a waveform which is a replica (i.e. scaled copy) of the output current waveform of the power amplifier. This waveform is produced via a Rogowski coil \[21\] cascaded with an op-amp integrator. The coil is mounted around one of the two output conductors leaving the power amplifier (between points O1 or O2 and the transmitter coil.) It is made by winding a few turns around a toroidal core with the conductor whose current is to be measured running through the core’s center hole. Alternating currents through the main conductor induce alternating magnetic fields which in turn induce a voltage in the coil wound on the toroidal core. This voltage waveform is the derivative of the main conductor’s current and must then be integrated using an integrator op-amp circuit (self-integrating coils are also possible but they are less tunable once built which is valuable if the phase of the signal must be changed.)

![Figure 17: Schematic of Rogowski coil and integrator combination](image)

The integrator is a delicate circuit for which trimmer potentiometers are used for tuning after assembly. R2 must be included because otherwise even the most miniscule common mode signal between the op-amp’s inputs will boundlessly integrate and saturate the amplifier. R2 should then be adjusted so that the low pass filter formed by C1 and R2 has a cutoff frequency that
is much smaller than the frequencies applied to the integrator. This ensures that the integrator performs the integration function and doesn’t simply do a DC amplification as would happen if the applied frequencies were in the low pass filter’s pass band. Reducing the cutoff frequency has the effect of “pushing” any given higher frequency further into the attenuation band and making the integrator “better” by bring the phase difference between the input and output closer to $\frac{\pi}{2}$ but also attenuates the output more. Thus R2 should be adjusted until the input and output have close to $\frac{\pi}{2}$ phase difference but the magnitude of the output is still large enough to be useful. Increasing R2 decreases the C1-R2 low pass filter cutoff frequency.

Increasing R2, however, also has the effect of increasing the DC gain of the amplifier, which is $\frac{R_2}{R_1}$. Increasing R2 can thus lead to clipping when the output’s DC offset plus the AC voltage swing exceeds the op-amp’s rail voltage. Since the AC voltage swing is the actual current replica that is needed this scenario is undesirable. Thus it is evident that a delicate balance must be struck between the passive components of the integrator and trimmer potentiometers are necessary to create a good output after assembly.

Since the integrator’s output is an AC voltage with a DC offset the result should be high pass filtered to remove the unneeded DC component. Since filtering introduces a phase shift the potentiometers in the integrator should be adjusted so that the final, AC only, output has the desired phase. Once this is achieved the two AC voltage waveforms (current and voltage replicas) can be fed into an analog multiplier IC which will produce their real-time product. It should be noted that, at this point in the signal processing chain, the phase difference between each replica signal and the original is not as important as the replicas’ relative phase difference. As long as the phase difference between the replica waveforms is the same as between the original signals the correct power measurement can be made even if each replica waveform is significantly lagging behind the original.

Once computed by the analog multiplier the product can be low pass filtered to leave only the DC voltage which is proportional to the real power. The resulting DC voltage should then be scaled again so that the DC voltage given to the controller is within the controller’s voltage limit for all geometries. The signal may need to be attenuated or amplified so that the controller does not receive too much voltage but also so that most of its entire input voltage range is used for more precise tracking. The output from this final scaling stage is the voltage $\text{DCVP}$ and this is sampled by the controller to gauge the amount of power being transferred.
6.3.2 Current measurement - DCVI

The AMM must also create a DC voltage level corresponding to the current consumed by drive coil, DCVI. This is achieved by a simple peak detector. The output current waveform replica created by the above Rogowski coil and integrator combination is connected to a buffer amplifier to avoid loading the original integrator circuit. The buffer’s output is then connected to a half-wave rectifier and the output is heavily low pass filtered. The output DC voltage level is DCVI. It corresponds to the current supplied by the amplifier and, although not linearly proportional to the current, can be used by the controller to determine when the input current has exceeded the desired threshold which is a preset based on experimental data.

6.4 Frequency generator module

![System diagram of the frequency generator module](image)

The frequency generator module is responsible for receiving serial commands from the microcontroller and producing a 50% duty cycle TTL level square waveform for driving the amplifier. Although the microcontroller is perfectly capable of generating PWM waveforms, having a separate module is advantageous for several reasons:

- Frequency resolution: Although the PSoC’s maximum clock speed of 80MHz [22] is more than enough to produce the high frequencies necessary for the system the frequency resolution is not adequate. This is because microcontrollers produce arbitrary frequencies by dividing the chosen clock by an integer number. To generate 1MHz, for example, the 80MHz clock must be divided by 80 which means that a 50% duty cycle 1MHz waveform can be produced by making a level transition every 40 cycles of the main clock. The next available frequency, however, is \(\frac{80}{79} = 1.012\text{MHz}\). This is too big a step for acceptable tracking. In addition, odd divider values cannot produce 50% duty cycle output.
Revised version:

- Reliability: As with any transformer DC voltage cannot be applied to the drive coil and the amplifier can be damaged by overcurrent if the drive signal fails to switch. If the controller were solely responsible for the amplifier’s drive signal then any situation that would make the controller momentary unable to switch the waveform (such as a user initiated or system mandated reset) could result in damage to the amplifier. Thus it is advantageous to have a simpler and more robust subsystem that can be guaranteed to always work regardless of the status of the more complex microcontroller.

Thus a separate module is used in this system to generate the drive waveform and the microcontroller is almost completely removed from this process. The basic circuit used is the canonical astable 555 timer circuit. This circuit uses a 555 timer chip and a few passive components to generate an output waveform with the frequency calculated by

\[
 f = \frac{1.44}{(R_A+2R_B)C_1}
\]

\[23\].

The 555 timer generates the variable frequency waveform by charging and discharging the C1 capacitor and feeding its voltage into a hysteretic inverter which controls the charge/discharge. When the output (pin 3) is LOW the DISCH (pin 7) terminal is shorted to ground through a transistor and C1 discharges through \( R_B \). As soon as the voltage across C1 falls to \( \frac{V_{cc}}{3} \) the inverter output switches to HIGH and the transistor on pin 7 turns off. This makes C1 charge to \( V_{cc} \) through \( R_B \) and \( R_A \). When C1 charges to \( \frac{2V_{cc}}{3} \) the output once again switches to LOW and C1 drains through the \( R_B \) and the DISCH pin. This simple yet powerful circuit makes it easy
to generate variable frequency waveforms with only a DC input voltage. A TTL rectangular wave, the output of the hysteretic inverter, is present on pin 3 and can be directly connected as needed.

In order to generate a variable frequency waveform the circuit must have a variable element. For this purpose the resistance $R_A$ is built from a resistor combination that includes a digital potentiometer, the 5144A by Analog Devices. This IC contains four individually programmable potentiometer modules whose resistance is set by serial commands over the I2C interface. The position of each potentiometer’s “wiper” can be set linearly by writing to the appropriate register 8-bit values in the range $[0,255]$. For this project a variable resistance in the range $[0, 20k\Omega]$ is desired and thus two $10k\Omega$ potentiometer channels are connected in series.

Each time the controller needs to change the frequency of the system it must send five bytes over the I2C channel to the 5144A in the following order:

1. **7-bit address byte.** This is a feature of I2C whereby multiple devices can be connected to the same data and clock lines (the controller’s SDA_1 and SCL_1 pins). Before the master (controller) device can communicate to a slave (5144A) it must first transmit its 7-bit address byte. The slave whose address byte is sent will listen to the subsequent message and the rest will ignore it. The address byte is given in the device datasheet and can be one of several values depending on the voltage levels connected to two pins, ADDR0 and ADDR1, on the potentiometer chip. The address byte must be transmitted at all times, even when there is only one slave device on the channel.

2. **Control byte for potentiometer 1.** This byte tells the 5144A that the subsequent byte must be written to the appropriate register to change the wiper position of potentiometer 1.

3. **Data byte for potentiometer 1.** This is the 8-bit byte in the range $[0,255]$ that sets the wiper position of potentiometer 1.

4. **Control byte for potentiometer 2.** This byte tells the 5144A that the subsequent byte must be written to the appropriate register to change the wiper position of potentiometer 2.

5. **Data byte for potentiometer 2.** This is the 8-bit byte in the range $[0,255]$ that sets
the wiper position of potentiometer 2.

When the above information is received the 5144A makes the necessary changes and the output of the 555 timer changes accordingly.

Although the 555 timer circuit is sufficient to generate the required frequency it cannot in conjunction set the proper duty cycle. This is because the charge and discharge time constants are governed by the $R_A + R_B$ and $R_B$ resistances and thus cannot be made equal unless $R_A = 0$ which can damage the 555 chip. It is necessary to use the waveform created by the 555 to ultimately make a 50% duty cycle TTL waveform and this can be done via a D-type flip-flop [24]. The flip-flop accepts an input TTL waveform of frequency $f$ with an arbitrary duty cycle and outputs a TTL waveform of frequency $\frac{f}{2}$ with a duty cycle of 50%. It does this by changing the output value only on the rising edge of the input. Thus the 555 timer must generate a waveform of double the required frequency which the flip-flop can then cut in half to output the desired signal. The output drive signal is then fed to the amplifier.

6.5 Wireless power transfer structure

The wireless power transfer structure is the combination of four coils that permits wireless power transfer through air and water. Power is applied to the transmitter side and extracted from the receiver side.

The receiver and transmitter side are identical in their construction. For the transmitter, a single turn coil is wound on a coil former made of non-conducting material and connected across the amplifier outputs (O1 and O2 in figure 15.) On the same coil former another multi-turn coil is wound and connected to a parallel capacitance. This multi-turn coil and capacitance combination is the LC tank whose resonance frequency $f$ is $\frac{1}{\sqrt{LC}}$ and $Q$ factor is $\frac{1}{R} \sqrt{\frac{L}{C}}$. The LC tank is electrically isolated from the driven single turn coil but is strongly magnetically coupled to it by virtue of it being tightly wound on the same coil former.

The LC tank is a very high-stress environment for the parallel capacitance. If the $N$-turn LC tank with $Q$ factor $Q_{LC}$ is driven by the H-bridge amplifier with a supply voltage $V_A$ then the voltage across the capacitance at resonance is $\frac{4}{\pi} N Q_{LC} V_A$. Although the system never operates exactly at the LC resonance frequency this voltage can still reach thousands of volts and appropriate components must be chosen to withstand these voltages. The structures must also be made as lossless as possible because a high $Q$ factor, which is inversely proportional to
the resistance of the LC loop, is necessary for efficient power transfer. The coils are wound from Litz wire, which reduces the skin effect, and the chosen capacitors should be at least ceramics.

The receiver side is identical to the transmitter side. The single turn coil on the receiver side is connected across the load electronics. Driving the single turn drive coil couples magnetic fields to the transmitter LC tank which efficiently exchanges energy with the receiver LC tank and couples it to the single turn receiver coil. Since the transfer distances are relatively small the system works in the near-field regime and the amplifier is effectively loaded by the receiver side (just as in the everyday transformer.) This permits the amplifier’s output power to be dictated by the system geometry and allows transmitter input power to be a predictor of load output power.

6.6 Load

The load utilized in developing and testing this system is a straightforward power resistor. In the early stages of development low-inductance (TO-247-2 through hole package) power resistors with resistive value of 5Ω, 10Ω or 20Ω were used and connected directly across the single turn source coil.

For completeness, however, a diode bridge is added which rectifies the incoming voltage waveform. This voltage is then low pass filtered by a large aluminum electrolytic capacitor across the diode bridge’s output terminals and the load resistor is connected in parallel with this capacitor. This yields a purely DC voltage across the load power resistor.

The diodes for the diode bridge must be carefully chosen. They must be able to conduct amps of current as well as switch fast enough to rectify voltages of frequencies up to and exceeding 1MHz. They should also have a low forward voltage drop to minimize losses since the bridge always conducts current through two diodes in series with the load.

The filter capacitor should also be chosen so that the cutoff frequency presented by the resulting RC low pass filter is much lower than the frequency of power transfer. Since the load resistances are small the capacitor should be relatively large and have large ripple current capability.
6.7 Underwater capability

The system has the ability to function while fully submerged in both fresh and salt water and this is the intended use case in order to charge autonomous underwater vehicles. In this scenario the electronics and WPT structure are placed in watertight enclosures on the ocean floor or riverbed. The only connections permitted from the shore to the electronics are DC power cables and a cable carrying a reset signal for the microcontroller. The transmitter side of the wireless link sits in its own enclosure which is connected to the main electronics container through an underwater power cable. The transmitter must be far enough from the electronics enclosure to prevent the created magnetic fields from interfering with the circuits. The enclosures containing WPT parts must be made of non-conductive material which will resist water pressure.

6.7.1 Underwater enclosures

The enclosures in which the electronics and transmitter are placed are rated IP68 or NEMA6P which allows them to protect their contents during continuous submersion in water of limited
The electronics enclosure consists of two parts: the box and a lid. The box and lid form a watertight seal using a rubber gasket which should be periodically treated with a lubricant to prevent drying. The lid attaches to the box with screws that are tightened into threaded holes in the box.

The enclosures for the transmitter and receiver are smaller rectangular boxes that house the parts of the WPT system and load. These enclosures must also have the appropriate rating and be made of non-conductive materials. The enclosures chosen for this purpose have hinged lids that compress a rubber gasket which should also be lubricated and provide complete protection during immersion in water.

6.7.2 Underwater electrical connections

Electrical connections to the electronics and transmitter coils are realized via underwater connectors of the SubConn series from MacArtney. The connectors are inserted into holes drilled through the sides of the enclosures and mate with matching underwater cables which contain conductors wrapped in a thick rubber sheath. The cables can be connected and disconnected wet (immediately after immersion without drying) and are held together by friction.

The DC connection from the on-shore power supply to the electronics enclosure (figure 21) is made via two underwater cables. The first is a 2-conductor cable that carries the DC power to the H-bridge. The second cable has 8 conductors (1 unused) that carry power to the rest of the system and a reset line to the microcontroller. It contains the following connections:

- (+12VDC, ground) to amplifier — 2 conductors
- (±15VDC, ground) to analog measurement module — 3 conductors
- reset line for microcontroller — 2 conductors

The transmitter (TX) enclosure is connected to the electronics enclosure through a 3-conductor (1 unused) underwater cable.

6.7.3 Amplifier heat sinking

Since the amplifier is sealed in an air- and watertight box a heat sink is used to interface between the power transistors and the inside of the electronics enclosure so that heat can be
conducted through the enclosure and into the surrounding water. The heat sink is a single aluminum piece that is used for all four transistors.

The heatsink is located underneath the amplifier board and has four slots into which the power transistors are inserted. Each transistor’s drain tab (of the TO-220 package) interfaces with the inside of its slot through a thin mica insulator (the drain tabs must be insulated because the heatsink is shared.) The transistors are fastened to the heatsink using nylon screws through the holes in their drain tabs (see Appendix A.9.2.)

The aluminum heat sink has a flat bottom face which is mounted to press tightly against the bottom of the enclosure. Thermal paste is used to increase the heat conductivity of this interface.
7 System Verification

This section presents measurements and data that show individual modules’ performance as well as overall system performance. The system is composed of two printed circuit boards (PCB’s):

- Amplifier board — this board contains the amplifier and the necessary circuits for driving the four power transistors. The board is powered by a 12V DC line from which 5V are created using a linear regulator. The transistors and gate drivers require 12V whereas the smaller logic chips for processing the drive signal require 5.

- Measurement (“small”) board — this board contains the microcontroller, analog measurement module and the frequency generator. This board is powered by a ±15V DC connection that is used by some of the op-amps on the board. A pair of linear regulators convert the incoming voltage to ±5V DC for the microcontroller and IC’s that require this lower voltage.

The presented graphs frequently refer to transfer distance as the independent variable — this is defined as the distance between centers of the coil formers when traveling along the vector normal to the coil plane. All tests are done using laterally aligned coils.

7.1 Microcontroller

The microcontroller must continuously sample the drive coil’s input power and adjust the frequency as necessary. To do this effectively the controller must be fast enough to follow the peak as the geometry (and coupling) of the wireless link changes. The microcontroller’s default interrupt period in the power state is 50ms and the controller must be fast enough to perform all its sampling and decision making in this time frame so that the execution of the subsequent ISR is not delayed.

Figure 23 shows the time it takes an ISR instances to complete. A microcontroller pin is used for the timing measurements — the first line of code in the ISR sets the pin HIGH and the last line turns it back LOW and the amount of time spent in the HIGH state corresponds to the time it took the ISR to execute. When the controller is not searching for the peak (figure 23 (a)) the ISR takes less than 4ms to finish. When the controller is searching for the peak, however,
the ISR takes around 26ms (figure 23 (b)). The controller has plenty of time to do everything it needs between interrupts and the interrupt frequency can even be increased for faster peak tracking if desired.

7.2 Amplifier

The amplifier can deliver up to 140W to a resistive load at efficiencies exceeding 82% at frequencies up to 1.1MHz (figure 25.) To test the amplifier as a standalone module it is loaded with low-inductance power resistors in the TO-247 package. The amplifier is tested with 5, 10 and 20 Ohm loads.
All of the amplifier’s circuits reside on one large board (figure 24.) The signal circuits along with their power sources are located on one end of the board and the input filter and H-bridge on the other. The power transistors are mounted up-side down under the board.

The drive signals for the amplifier are created by the pre-drive circuitry. The delay circuit potentiometers can be adjusted to change the dead-time between the gate drive signals.

Figure 26 shows the drive signal dead-time which is the time delay between the falling and rising edges of adjacent CH3 and CH4 waveforms. The two dead-time intervals are invidually controllable but should be made equal for even performance.
Figure 26: Frequency generator output (top, CH2) and amplifier gate drive signals (bottom, CH3 and CH4).

Figure 27: (a) Amplifier current (CH1) and voltage (CH2) powering a resistive load (b) Amplifier current, voltage and real time power (MATH)

7.3 Analog measurement module

The analog measurement module resides on the same PCB as the microcontroller and the frequency generator. It receives the amplifier current and voltage information through BNC cables (voltage) and Rogowski coil (current) and processes the waveforms to get a picture of the real power coming out of the amplifier.

Figure 29 shows the current (figure (a) CH3)) and voltage (figure(b) CH3) replicas that are created by the analog measurement module. The analog multiplier multiplies the two waveforms and low pass filters the result. The signal is then scaled and sampled by the microcontroller.

Figure 30 shows the performance of the current replicator and peak detector blocks during one particular transmission scenario. Waveform G3 corresponds to values of DCVI for different frequencies. The amplifier is loaded with the WPT structure which is connected to a 5Ω load.
Figure 28: (a) The analog measurement module’s Rogowski coil around one of the H-bridge’s output conductors (b) Analog measurement module components on the measurement board. A: Rogowski coil input and integrator B: Voltage inputs and processing C: Analog multiplier, LPF and scaler

Figure 29: (a) Actual (CH1) and replicated (CH3) amplifier output current (b) Actual (CH2) and replicated (CH3) amplifier output voltage.

The transmission distance is 2 inches and the H-bridge is powered by 12VDC.

Figure 31 shows the performance of the entire analog measurement module in producing the DC voltage $\text{DCVP}$, the values of which are plotted by curve G3 in both graphs. G1 and G2 correspond to amplifier output power and load power, respectively. It should be noted that the performance of the module is better for the second peak which is the one that the system focuses on. This better performance may be due to the low pass filtering performed by the integrator.
Figure 30: Amplifier output current (G1), replicated current (G2) and peak detector output (G3) versus frequency in air. Amplifier is loaded with full WPT structure with 5 Ohm load across 2" distance with 12V input H-bridge input.

Figure 31: (a) Actual and replicated power values under real operating conditions in air (distance = 2", Load = 5Ω, H-bridge \( V_{in} = 12V \)) (b) Same conditions as (a) but with \( V_{in} = 26V \) whereby higher frequency signals are integrated with a phase shift that is closer to the desired \( \frac{\pi}{2} \).

7.4 Frequency generator

The frequency generator resides on the measurement board near the microcontroller.

The microcontroller sets the amplifier drive frequency by sending the frequency generator module two 8-bit integers in the range \([0,255]\). The two integers set the resistances of two individual potentiometers in the digital potentiometer IC and, since the two resistances are in
Figure 32: Frequency generator on the measurement board. A: Digital potentiometer B: 555 timer C: 
D-type flip-flop

Figure 33: Frequency generator output frequency versus potentiometer index series, the combined resistance corresponds to the sum of the two integers. Smaller integer 
values set a higher resistance and thus lower overall frequency, and vice-versa. Figure 33 plots 
the resulting drive frequency versus the sum of the two integers sent to the potentiometer IC. 
The useful frequency range for this system is approximately \([770, 1100]\) kHz and this corresponds 
to the sum in the range \([265, 500]\).

Figure 34 shows the output waveform of the 555 timer (CH4) and final generator output 
waveform (CH3.) The final output has a frequency of half the 555 timer’s output since the 
D-type flip-flop only transitions its output on a rising edge of its input. Thus the 555 timer’s
Figure 34: Example 555 timer output waveform (CH4) and final frequency generator output waveform (CH3)

The wireless power transfer structure consists of the four coils wound on two coil formers that exchange energy wirelessly. In figure 35 the two alligator clips connected to the right coil bundle bring power from the amplifier to the drive coil. The drive coil, which is just the single turn of light wire in the right bundle, is tightly coupled to its neighboring black winding, the transmitter (TX) coil. The TX coil resonates and through resonant coupling exchanges energy with the receiver (RX) coil in the left coil bundle across the gap. The RX coil is likewise tightly coupled to its single turn source coil which delivers an AC voltage (which is purely sinusoidal at the drive frequency) to the load through the two alligator connections on the bottom left. Driving the coils at resonance creates very large voltage levels across the resonant capacitor banks which are shown in rectangles in figure 35.

A high $Q$ factor is necessary for efficient power transfer between the two resonant tanks. However, a high $Q$ leads to a low loss system which creates very large voltages across the resonant capacitances when the system is driven at the peak frequencies. Figure 36 shows the RMS voltage across the TX resonant capacitance and the voltage at the load across a range of frequencies for a 2.25" separation distance with a 12V H-bridge input and a 5Ω load. The figures show that even with a small drive voltage (12V) the resonant capacitance is subject to
hundreds of volts RMS.

Figure 35: WPT structure. The coils’ names (top) omit the word “coils” for brevity.

Figure 36: (a) Resonant capacitance and load voltages with H-bridge $V_{IN} = 12V$ (b) Resonant capacitance and load voltages with H-bridge $V_{IN} = 26V$. Data taken in air.

Figure 37(a) shows that increasing the distance of wireless transmission decreases the frequency at which the second peak is located. If the system stays on the peak as the distance increases the amplifier must supply larger amounts of power because it sees a falling impedance.
as the peak frequency nears the resonance frequency of the resonant tank. Because the amplifier is dumping more energy into the LC tank the resonant capacitance’s peak voltage grows with increasing distance. At smaller distances this is beneficial because the load voltage also increases and thus the largest amount of power can be transmitted not at the smallest distance but farther away. Past this point, however, the peak load voltage drops with distance but the resonant capacitance voltage continues to increase. This is shown in figure 37(b).

![Figure 37](image)

*Figure 37: (a) Frequency of second peak versus distance (b) Load and TX resonant capacitance voltages at 2nd peak versus distance. Data taken in air.*

### 7.6 Underwater capability

To make the system submersible in water additional hardware is used. The amplifier PCB is assembled with the power transistors mounted underneath the board (opposite side from other components) so that they can be directly interfaced to the heatsink underneath.

The heatsink is mounted to the drain tabs of the power transistors (TO-220 package) and has a vertical thickness (figure 38) that is equal to the height of the drain tabs. This creates a space between the heatsink and the amplifier that is equal to the length of the transistor leads between the bottom of the drain tab and the PCB. The heatsink is mounted in the enclosure by inserting thin aluminum beams (cross-beams in figure 39) into this space and bolting them down into the mounting beams.

The measurement board sits above the amplifier’s logic circuitry on two metal posts (figure 40.)
Figure 38: Heatsink mounted under amplifier PCB

Figure 39: Amplifier mounted in enclosure
The entire electronics assembly sits in the electronics enclosure. Electrical connections to the outside are made through underwater connectors (figure 41.)

Figure 40: Measurement board and amplifier assembly

Figure 41: Electronics assembly inside enclosure. Note the power connectors installed in the side of the enclosure.
Figure 42: (a) Sealed electronics box (b) Sealed transmitter box

The enclosures are connected with undersea cables and submerged. The wireless power transfer system can now function exactly as it did in the air.

Figure 43: Wireless power transfer system submerged inside fish tank

7.7 System

This section shows the performance of the system as a whole.

Figure 44 shows the performance of the wireless power link at various distances around the second peak with H-bridge $V_{IN} = 12$V and a $5\Omega$ load. This figure plots the input power, or power coming out of the DC supply to the H-bridge and load, or DC output, power measured at the resistive load — thus this represents wall plug-to-load performance and includes losses in the amplifier, WPT structure and load rectifier.
Figure 44: Input and output power around the 2nd peak versus frequency for varying distances

Figure 45: Input and output power around the 2nd peak versus frequency for varying load resistances

Figure 45 shows the input and output power curves around the second peak for three different values of the load resistance for a 12V H-bridge input and a 3” separation distance. Larger load resistances absorb more power but at poorer efficiencies and the graph shows that with a 25Ω load and an H-bridge supply voltage of only 12V the amplifier outputs over 100W at the second peak. Thus it is advantageous to test and operate the system with the low resistance value of 5Ω since the power can always be increased by increasing the H-bridge supply voltage.

Figure 46 confirms that increasing the load decreases the efficiency and these losses come
primarily from the amplifier. Larger load resistances, when transformed by the WPT structure, appear smaller to the amplifier and its losses increase which is shown by the larger disparity between the total (dashed) and WPT (solid) efficiency curves for the larger loads. The graph also shows that peak power operation does not yield best efficiency and operating at frequencies lower than peak is more efficient but with much lower power. In addition, efficiency falls rapidly with increasing frequency past the peak and this is due in part to increasing losses in the amplifier because of the increasingly large phase shift between the amplifier output current and voltage.

Figure 47 shows that at the peak the WPT structure and load look almost perfectly resistive to the amplifier (at the fundamental frequency component) and the phase difference between the current and voltage is very small. At frequencies above the peak, however, the system looks more and more inductive and this results in an increasing phase shift between current and voltage which forces the amplifier’s transistors to switch while carrying a significant current thereby increasing switching losses.

Figure 48 shows the performance of the auto-tuning system at varying power levels with the 5Ω load. The graphs show how well the system can auto-tune and, importantly, how much power the system could transmit if left to perform the desired tuning autonomously. The independent variable is distance and the graphs show the auto-tuned and actual 2nd peak frequencies as well as the peak load power at the frequency to which the system auto-tuned for every separation distance. The range of distance decreases with increasing H-bridge supply voltage (shown in the upper left corner of every graph) because higher input voltages cause the current to reach the
allowed limit sooner causing the system to shut off.

It should be noted that the data was collected by letting the system tune to the stationary peak from a reset, turning the system off, changing the amplifier drive from the frequency generator to a user-controller signal generator and then, with the system on, manually finding the peak. Thus there are a few points at which the same frequency point was found by both means but the power values slightly differ — this is due to the fact that the system was turned off and on between measurements and was therefore not completely consistent. Likewise in some cases the system “found” slightly more power than present at the manually found peak — this can be similarly explained.

Figure 49 shows that the system can transmit over 90% of the possible peak power when left to its own devices without user input. Predictably the system performs poorer for the lowest power setting (12V) because the sampled DC voltages have smaller values and smaller differences at and around the peak which makes it more difficult to tune accurately. At the higher power levels the system can perform at over 95% found power which means the auto-tuning can accurately keep the system running at the proper frequencies for close to maximum possible power transmission at distances where the amplifier output current is within limits. As
Figure 48: Auto-tuning performance at the second peak in air. For varying distances the auto-tuned and actual peak frequencies and load power levels are shown.

in the previous figure some points show over 100% which is due to the slight differences between power levels recorded when the system was power cycled between measurements.

Figure 50 shows the performance of the system when fully submerged. The graph plots the second peak frequency versus separation distance with the adjacent number corresponding to the load power. Because of the difficulty of controlling the system from the outside when submerged the system was allowed to auto-tune at every separation distance. The load power peaks at a significant distance and then decreases when the receiver moves further away. The H-bridge input voltage is 12V and the load resistance is 5Ω. The graph terminates at approximately 5 inches where the controller shuts the system off.
Figure 49: Percentage of actual peak load power transmitted when system auto-tuned in air

Figure 50: Second peak frequency versus separation distance with corresponding load power in water. Distance includes enclosures.
8 System Scaling

The wireless power transfer system has many variables, most of which are held constant in the development of the system in this thesis. Power levels, resonant tank resonance frequency and mode of operation (maximum power or maximum efficiency) can be changed while maintaining the same principle of operation. Although the system’s current hardware is selected for maximum power (second peak) operation with resonant tanks of resonance frequency $f_0 \approx 900$kHz, changes can be made to accommodate other operation conditions.

8.1 First peak tuning

The current system’s objective is to tune its operation to the second frequency peak because the current coil setup provides maximum power at the second peak. Some coil setups, however, may benefit from operation at the first power peak. The system can be adapted to operate at this first peak with only a few software modifications.

The main difference in behavior between the two peaks has to do with the fact that the first peak’s location increases (in frequency) when the receiver moves away whereas the second peak’s location decreases. Thus the controller must execute the same code as before but with some slight changes to take this into account.

8.1.1 Startup

The microcontroller performs a specific routine when first turned on to find the location of the peak that it must tune to. If the controller must tune to the first instead of second peak the code responsible for these actions must be altered in the following places. Each item in the list corresponds to the new line of code inserted in place of the old at the stated line or lines. Please refer to Appendix B.3 for original code.

- Line 37:

  if(c_H_val>c_val && c_val>c_L_val){

  This prevents the controller from following the positive slope at the preset frequency during startup because if the receiver is out of range then the first peak is at a higher
frequency than the preset. This is in contrast to the present case where the system is not allowed to proceed if the 3-point slope at the preset frequency is negative.

- Line 51:

```
max_val_ix=CUTOFF-4;
```

This sets the starting frequency to a value lower than the preset cutoff frequency value. This allows the system to start at a lower frequency than the preset when the receiver is within range upon startup so that the system can reach the peak slightly faster. This is in contrast to the present code where the index is set to a few points higher than the preset cutoff.

### 8.1.2 Post-startup operation

During the post-startup phase the controller must execute the same peak hunting code except now it must check for the overcurrent condition (too high DCVI) when attempting to shift its operation to higher frequencies because the first peak shifts higher when the transmission distance increases. The decision tree code (lines 78-118 in Appendix 10.4 must be replaced with the following altered decision tree:

```
//----------DECISION TREE LOGIC--------------------------------------
if(SAMPLES==3){ //how many sample points (3 in thesis)
  if(diffs[0]>0){
    if(diffs[1]<0){
      hunting = 0;
      //if both adjacent points lower -> local max ->
      //found peak
    }else if(diffs[1]>=0){ //local slope is positive or
        //close to peak
      if(currents[2]<C_CUTOFF){ //current OK?
        max_val_ix++; //move up one frequency point
      }else{ //current at next highest frequency is too high
        Amp_IO_Write(0);//turn off amplifier
        hunting=1; //still haven’t found peak
        Clock_1_SetDividerRegister(AWAY_WAIT*1000,RESTART);
      }
    }
  }else if(diffs[0]<0){ //is the local slope negative?
    max_val_ix--; //decrease frequency because
    //peak is to the left
    Clock_1_SetDividerRegister(ISR_CLK_DIV,RESTART);
  }else if(diffs[0]==0){
```
if(diffs[1]>0) { //power increasing to the right
    if(currents[2]<C_CUTOFF)//current OK?
        max_val_ix++; //follow increasing power
    else{
        Amp_IO_Write(0); //turn off amplifier
        hunting=1; //still haven’t found peak
        Clock_1_SetDividerRegister(AWAY_WAIT*1000,RESTART);
    }
} else if(diffs[1]<0) {
    max_val_ix--; //decrease frequency
} else{
    hunting = 0; //plateau = assume found peak
}

In addition, the preset cutoff frequency point at which the controller begins its operation must be set to a new point $f_p$ which is higher than $f_s$ (the location of the first peak) when the receiver is close and lower when it is too far away. To do this, the preset must be changed in Appendix B.2 by changing line 1 to

```c
#define CUTOFF 425
```

The above preset is experimentally determined. The key is that the preset corresponds to the location of the peak when the receiver is at the edge of useful charging distance — any larger separation and the peak moves higher than the cutoff and the controller can quickly see this by measuring the local slope. This preset is usually only a few percentage points away from the resonant $f_0$ of the tanks so that the system can charge over a wide range.

### 8.2 Different LC resonance frequency $f_0$

The current system implementation is designed for operating with a WPT structure whose resonant LC tanks have a resonance frequency $f_0$ of around 900kHz. The system is versatile, however, and can operate at other frequencies to accommodate other resonant tanks. In order to operate at another frequency some of the passive circuit components must be resized. The list below lists the necessary hardware changes when the resonant tanks’ resonance frequency changes from $f_0$ to $f_n$. Appendix A presents the locations of modifications.
8.2.1 Microcontroller

No changes to the microcontroller are necessary since its interface to the power transmission is a DC voltage level and it does not directly see the operating frequency. Although the controller sets the frequency of operation by programming the frequency generator changing the frequency can be done by modifying the generator’s circuit only.

8.2.2 Power amplifier

The H-bridge itself can operate at essentially any frequency with the upper bound being determined by the gate drivers’ ability to provide the fully-ON charge to the transistors in time and the transistors’ full turn on and turn off times. In addition, larger frequencies have more on/off transitions and thus larger losses which must be offset by adequate heat sinking.

The delay and dead-time creation circuits may have to be modified if the switching frequency is to drastically increase because the present dead-time may be unnecessarily large if faster transistors are used for higher frequencies. To decrease the dead-time by a factor of \( n \) the RC time constant of the circuit in figure 14(a) must be decreased by a factor of \( n \) and this is best done by changing the resistances of potentiometers R1 and R8 by \( n \). If the desired dead-time is outside the potentiometers’ range then the capacitances C5 and C6 can be scaled by \( n \) as well (figure A.9 left.)

In addition the H-bridge input filter must be changed, especially if \( f_n < f_0 \). Since the cutoff frequency of the filter is \( \frac{1}{\sqrt{ LC}} \) changing the switching frequency to \( f_n \) requires scaling the LC product by \( \left( \frac{f_0}{f_n} \right)^2 \). The affected components L1 and the C20-C26 capacitor bank are shown in figure A.10 bottom.

8.2.3 Analog measurement module

To prepare the analog measurement module for the new operating frequency \( f_n \) the cutoff frequencies of its filters should be scaled accordingly by scaling the following capacitances.

- Current replicator (Figure A.4 Top) — Scale the cutoff frequency of the RC filter formed by C12 & R30 by \( \frac{f_n}{f_0} \) by scaling the C12 capacitance by \( \frac{f_0}{f_n} \).

- Analog multiplier output low pass filter (Figure A.5 Top) — Scale the cutoff frequency of the RC filter formed by C24 & R16 by \( \frac{f_n}{f_0} \) by scaling the C24 capacitance by \( \frac{f_0}{f_n} \).
• Analog multiplier input high pass filters (Figure A.5 Top) — Scale the cutoff frequency of the RC filters formed by (C13 & R5) and (C25 & R14) by \( \frac{f_n}{f_0} \) by scaling the C13,C25 capacitances by \( \frac{f_0}{f_n} \).

• Coil current peak detector (Figure A.5 Bottom) — Scale the cutoff frequency of the RC filter formed by C54 & R15 by \( \frac{f_n}{f_0} \) by scaling the C54 capacitance by \( \frac{f_0}{f_n} \).

8.2.4 Frequency generator

To prepare the analog measurement module for the new operating frequencies in the range of \( f_n \) the frequencies created by the 555 timer must be set to around \( 2f_n \). Although a simple way to change the frequency of the 555 timer chip is to scale the C47 capacitance by \( \frac{f_0}{f_n} \) there is another option. The 555 timer’s pin 5 accepts a DC voltage level which also to an extent controls the output frequency. In practice, increasing the voltage applied to this pin decreases frequency and visa-versa. If it is necessary to increase the operating frequencies then changing this control voltage is the simplest method which is better than decreasing the already small C47 capacitor. The 555 timer chip currently used has a nominal limit of 3MHz which limits the frequency module’s output to 1.5MHz. Frequencies above this limit will need to explore other options such as voltage to frequency converters. In addition, increasing the frequency of operation worsens frequency resolution.

If the new frequency \( f_n \) is lower than \( f_0 \) then increasing the C47 capacitance is the most straightforward option. For \( f_n < f_0 \) the C47 capacitance must only be scaled by \( \frac{f_n}{f_0} \).

8.2.5 WPT structure

The WPT structure resonance frequency is calculated by \( \frac{1}{\sqrt{LC}} \). The most straightforward way to change the resonance frequency of the LC tanks is by scaling the parallel capacitance. If the frequency must be changed from \( f_0 \) to \( f_n \) then the capacitance must be scaled by \( \left( \frac{f_0}{f_n} \right)^2 \). This is necessary if the structure needs to be operated at a different frequency without physically changing the coils. Altering the coils by increasing their diameter, etc. will change their inductance which will also change the resonance frequency and the parallel capacitance must be recalculated to set the resonance frequency to the desired.

The L and C values, however, set more than just the resonance frequency. The \( Q \) factor, which is necessary for efficiency power transfer, is also determined by their values via \( Q = \frac{1}{\pi \sqrt{LC}} \).
Decreasing the inductance while at the same time maintaining the same resonance frequency by increasing C will be detrimental to the Q factor and care should be taken that it does not fall so low that the advantages of the four coil structure disappear.

8.2.6 Load

The load side must be changed to accommodate drastic increases in frequency. Most importantly, the filter capacitor, which is currently an aluminum electrolytic, may be unsuitable for high frequency filtering. If the resonance frequency increases to $f_n$, the cutoff frequency of the RC filter formed by the capacitor and load resistance can also be scaled by $\frac{f_n}{f_0}$ by scaling the filtering capacitor by $\frac{f_n}{f_0}$. This is beneficial because increasing the cutoff point permits using smaller sized capacitors with better high frequency properties. The diode bridge does not need to be altered unless the frequencies increase so much that the diodes’ switching speed significantly increases their power consumption and bridge efficiency.

8.2.7 Underwater capability

The underwater hardware does not depend on the switching frequency with the exception of the heatsink which will have to accommodate larger heat dissipation from higher switching rates.
9 Conclusion

This thesis presents the motivation behind and design details of a wireless power transfer system capable of auto-tuning based on externally imposed changes as well as underwater power transfer. The system can completely autonomously transfer power, detect changes in the power versus frequency profile and adjust its frequency of operation to respond to these changes with the goal of always maintaining the highest level of power transfer possible. In addition, the system can detect when the wireless power transfer target (the receiver) has departed and shut itself off to prevent wasted power. All this is achieved without any kind of communication with the wireless power receiver.

9.1 Summary of operation

The system performs the auto-tuning function using a microcontroller executing a dynamic hill-climbing algorithm that seeks to maximize the input power to the wireless power transfer link. The microcontroller performs frequency adjustments and measures the input power by sampling a DC voltage level derived by the analog measurement module. This DC voltage level is calculated by multiplying the waveforms corresponding to the input current and voltage and taking their time average which is the textbook definition of DC, or real, power. The microcontroller then seeks to maximize the input power by adjusting frequency in the direction of increasing power until it has found the peak. It then stays on this peak until the operating conditions have changed and the peak has moved to a different frequency at which point it resumes the peak tracking process.

9.2 System strengths

The system has two main advantages over traditional fixed-frequency wireless power transfer systems.

1. The system’s ability to adjust its frequency of operation allows it to have a constantly high power throughput in use cases where the receiver is not able to remain stationary relative to the transmitter. This is useful in environments such as underwater where it may not be possible to “park” the receiving vehicle and it must remain floating (and in slight motion) around the transmitter during charging. Even if the receiver is able to park and
remain stationary, however, its parking orientation may be inconsistent between charging instances and every time a different frequency is optimal. The system will produce more power throughput and faster charging than a fixed-frequency system in both cases.

2. The system performs its decision-making process regarding frequency without directly communicating with the load in any way. This is important because it can thus operate in environments that are difficult for communication, such as underwater. This also lets it work with legacy vehicles for which it may not be possible or practical to implement other communication capabilities. With the system using only input power characteristics it is possible to charge any target with only a receiver coil structure implanted within.

9.3 Further work

The current system is a fully working prototype capable of bringing over 100W to a load while performing auto-tuning. From here there exist a number of topic extensions that are of interest.

9.3.1 Increasing power throughput

The current system is a low power prototype the goal of which is to demonstrate the feasibility of auto-tuning using input-side information only. There is a lot of interest in vastly increasing the amount of power that such a system would deliver into the kilowatt range and beyond. Using the current hardware it is possible to increase the amount of power and this can be done by implementing MOSFETs with higher voltage and current ratings. In order to accommodate these MOSFETs it may be necessary to lower the transfer frequencies which is possible with the H-bridge inverter amplifier.

9.3.2 Reducing losses

The current system transfers power with total efficiencies exceeding 50% over distances less than 70% of the coil diameter. Losses can be mitigated by implementing certain hardware changes.

- **Improving $Q$ factor** — The transfer efficiency of the four-coil system is critically dependent on the $Q$ factor of the resonant tanks coupled to the drive and source coils. The
current resonant tanks, although possessing a nominal $Q$ factor of over 2000 (using DC resistance measurement,) can be greatly improved. This can be achieved using larger diameter wire with more insulated conductors and also by winding coil turns in parallel as opposed to series. The current LC tanks get noticeably warm when operating which makes it clear that they exhibit non-trivial losses.

- **Reducing MOSFET impedance** — The current MOSFETs used in the power amplifier exhibit relatively high conduction losses which only increase when the switches transfer power and warm up. These switches were chosen specifically because they, although possessing a high on resistance, have good high speed switching capabilities. Reducing the operating frequency would allow somewhat slower switches to be used which can then be selected to have much lower impedances.

### 9.3.3 Implementing battery charging

As it stands the system is capable of providing DC power to charge a high capacity AUV battery. Thus a straightforward extension of the current work effort is to load the wireless power transfer structure with a battery charging structure to evaluate its performance when doing this practical task.

### 9.3.4 Inclusion of metamaterials

A metamaterial is a man-made assembly of like elements that exhibits properties not found in natural materials [26]. Although much beyond the scope of this thesis experiments have shown that a particular metamaterial structure can improve wireless power transfer efficiencies at low power levels [27]. The application of this technology can be extended to higher power levels and combining it with a smart wireless power transfer system may yield further improvements in performance. Because the metamaterial acts as a selective frequency filter of its own its inclusion would impose additional frequency control constraints that would have to be taken into account.

### 9.4 Final remarks

The wireless power transfer system discussed in this document is a rudimentary yet promising example of smart wireless power transfer that is cost effective and simple to implement. Using the knowledge gained in its development it is possible to develop scaled systems that will
deliver higher amounts of power at greater efficiencies while using the same basic principles of energy conservation and electromagnetism. It is the hope and intent of the author and all those involved in the system’s development that this small first step can pave the way for much more powerful and capable systems that can radically transform the way we power our increasingly technologically dependent world.
10 Bibliography


http://www.modularcircuits.com/blog/articles/h-bridge-secrets/h-bridge-control/.


A Appendix — Hardware

This section contains all technical documentation and specifications regarding the hardware of the wireless power transfer system. Schematics are divided into submodules that are enclosed in rectangles. Connections to other modules and submodules are denoted by solid black rectangles with accompanying text labels.

The microcontroller, frequency generator and analog measurement module all reside on a single PCB and share the onboard power supply which is shown in Appendix A.4. The schematics for the microcontroller, frequency generator and analog measurement module are located in their respective appendices but the bill of materials, being shared by the three modules, appears in its own Appendix A.5. The power amplifier has its own board whose bill of materials is presented in Appendix A.6.

The dimensions of all technical drawings are in inches.
A.1 PSoC microcontroller

The software and mixed signal layout of the PSoC is done in PSoC Creator which can be downloaded from the Cypress Semiconductor website at http://www.cypress.com/products/psoc-creator-integrated-design-environment-ide. The IC is physically programmed using the CY8CKIT-059 Prototyping kit via USB from within PSoC Creator. Once implemented in a PCB the PSoC microcontroller is connected to the programming kit via connector J6.

Figure A.1: Microcontroller schematic
Figure A.2: PSoC programmed hardware blocks as created in the PSoC Creator Top Design tool

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Figure A.3: PSoC programmed hardware block specifications
A.2 Analog measurement module (AMM)

Figure A.4: Schematic of circuits for measuring and replicating coil voltage and current waveforms
Figure A.5: Schematic of the analog multiplier, subsequent scaling circuit and coil current peak detector
A.3 Frequency generator

Figure A.6: Schematic of frequency generator module
A.4 Measurement board power supply

The following is a schematic of the power supply circuit that powers the microcontroller, analog measurement module and frequency generator. Power is brought from a ±15V power supply. In addition to these levels the board also uses ±5V which is created by the circuit below.

Figure A.7: Schematic of the ±15V to ±5V converter
A.5 Measurement board bill of materials

The following is the bill of materials for the measurement board. All components with the exception of the Rogowski coil sourced from Digikey.

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Figure A.8: Measurement board bill of materials
A.6 Power amplifier

The power amplifier (H-bridge and associated circuitry) resides on a separate PCB from the other circuits.

Figure A.9: Schematic of power amplifier circuit that splits the input drive signal, creates dead-time and allows on/off capability

Figure A.10: Schematic of 12V to 5V converter and H-bridge input filter
Figure A.11: Schematic of H-Bridge outputs A and B
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<td>J1 J2 J3 J4 J5 J6</td>
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</tr>
<tr>
<td>BNC-PCB connector</td>
<td>A32246-ND</td>
<td>J7 J8 J9</td>
<td>3</td>
</tr>
<tr>
<td>Inductor</td>
<td>732-2168-1-ND</td>
<td>L1</td>
<td>1</td>
</tr>
<tr>
<td>Power transistor</td>
<td>497-7522-5-ND</td>
<td>Q1 Q2 Q3 Q4</td>
<td>4</td>
</tr>
<tr>
<td>Potentiometer</td>
<td>987-1710-ND</td>
<td>R1 R8</td>
<td>2</td>
</tr>
<tr>
<td>Resistor 2512</td>
<td>A121282CT-ND</td>
<td>R4 R5 R6 R7</td>
<td>4</td>
</tr>
<tr>
<td>Keystone 5021 Test Point</td>
<td>36-5021-ND</td>
<td>TP1 TP2 TP3 TP4 TP5 TP6 TP10 TP11 TP12 TP13 TP14 TP15</td>
<td>12</td>
</tr>
<tr>
<td>Linear regulator 5V</td>
<td>LM7805CT-ND</td>
<td>U2</td>
<td>1</td>
</tr>
<tr>
<td>Low side gate driver</td>
<td>CLA363-ND</td>
<td>U5 U8</td>
<td>2</td>
</tr>
<tr>
<td>High side gate driver</td>
<td>LTC4440EMS8E-5#PBF-ND</td>
<td>U4 U6</td>
<td>2</td>
</tr>
<tr>
<td>Schmitt inverter 6-pack</td>
<td>296-1177-1-ND</td>
<td>U1</td>
<td>1</td>
</tr>
<tr>
<td>And gate</td>
<td>568-2472-5-ND</td>
<td>U7</td>
<td>1</td>
</tr>
</tbody>
</table>

*Figure A.12: Amplifier bill of materials*
A.7 WPT structure

Figure A.13: (a) Coil former half technical drawing (b) Coil former connector technical drawing

Figure A.14: Computer rendering of assembled coil former
Figure A.15: Cross section of coil former with windings. The Litz wire turns (inner circles) are enclosed in shrink tubing (outer circles) for added insulation (LC tank coils only.) D/S: drive/source coil.

Figure A.16: WPT coil specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Drive/source coil # turns</td>
<td>1</td>
</tr>
<tr>
<td>LC tank coil # turns</td>
<td>10</td>
</tr>
<tr>
<td>LC tank coil inductance</td>
<td>29uH</td>
</tr>
<tr>
<td>LC tank coil ESR</td>
<td>.048Ω</td>
</tr>
<tr>
<td>LC tank parallel capacitance</td>
<td>1.25nF</td>
</tr>
<tr>
<td>Wire type</td>
<td>Litz, 42 strands of 44AWG conductor,</td>
</tr>
<tr>
<td></td>
<td>individually insulated, .088” total diameter</td>
</tr>
<tr>
<td>Shrink tubing</td>
<td>1/8” initial diameter, .116” final diameter</td>
</tr>
</tbody>
</table>
A.8 Load

The load side is flexible and a number of loads may be accommodated. All data collected in this thesis is taken with purely real loads with resistances between 5 and 30 ohms.

![Load schematic](image.png)

*Figure A.17: Load schematic*

<table>
<thead>
<tr>
<th>Component</th>
<th>Distributor Part No</th>
<th>Ref Name</th>
<th>Qty</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power diode</td>
<td>497-7577-5-ND</td>
<td>D1 D2 D3 D4</td>
<td>4</td>
</tr>
<tr>
<td>Electrolytic capacitor</td>
<td>495-6057-ND</td>
<td>C1</td>
<td>1</td>
</tr>
<tr>
<td>Load resistor 10 Ohms</td>
<td>TEH100M10R0FE-ND</td>
<td>R1</td>
<td></td>
</tr>
</tbody>
</table>

*Figure A.18: Load bill of materials*
A.9 Underwater capability

This appendix contains the technical specifications and assembly instructions of the hardware used in building the underwater system.

A.9.1 Amplifier heatsink

Figure A.19: Isometric bottom view of amplifier heatsink. Rectangular slots A-D penetrate through the entire height of the sink, slots E and F do not (See figure A.20 for dimensions.) Holes 1-4 penetrate the walls between the slots parallel to the dashed arrows.
Figure A.20: (Top) Side view of heatsink (Bottom) Bottom view of heatsink
A.9.2 PCB and heatsink assembly

The amplifier and measurement PCB’s are assembled by mounting the measurement PCB above the amplifier PCB on two mounting posts. The posts are cut from a $\frac{1}{4}''$ aluminum dowel.

![Figure A.21: Technical drawing of vertical mounting post](image1)

![Figure A.22: Attachment of heatsink to transistors underneath the amplifier board](image2)

The four nylon nuts N2 are inserted into the two middle slots. The four nylon screws S3 are drawn original size and must be trimmed length-wise to fit snugly into the holes and tighten...
against the metal transistor drain tabs. The transistor drain tabs must be insulated from the heatsink using mica insulating pads (.002” thickness.)
A.9.3 Electronics enclosure assembly

Figure A.24: Technical drawing of bottom mounting beam component

Figure A.25: Technical drawing of top mounting beam component
Figure A.26: Assembly of mounting beam from top and bottom component. The left (MRL) and right (MBR) mounting beams are mirror images of each other and assemble from identical top and bottom components.

Figure A.27: Technical drawing of cross beam
Figure A.28: Assembly of complete structure. CB = cross beam, MBL = left mounting beam, MBR = right bounding beam. The four screws S1 screw into mounting beam holes threaded with 6-32 tap.
Figure A.29: Mounting of assembly inside electronics enclosure
<table>
<thead>
<tr>
<th>Item</th>
<th>Qty</th>
<th>Source</th>
<th>PN #</th>
</tr>
</thead>
<tbody>
<tr>
<td>Main electronics enclosure with lid</td>
<td>1</td>
<td>Polycase</td>
<td>YQ-100806-14</td>
</tr>
<tr>
<td>Mounting plate with (S2) screws</td>
<td>1</td>
<td>Polycase</td>
<td>YX-1008K</td>
</tr>
<tr>
<td>Coil enclosure</td>
<td>2</td>
<td>Automation Direct</td>
<td>P9082C</td>
</tr>
<tr>
<td>8 contact undersea connector + O-ring</td>
<td>1</td>
<td>MacArtney</td>
<td>Subconn Micro Series MCBH8F</td>
</tr>
<tr>
<td>2 contact undersea connector + O-ring</td>
<td>1</td>
<td>MacArtney</td>
<td>Subconn Power Series BHB2F</td>
</tr>
<tr>
<td>3 contact undersea connector + O-ring</td>
<td>2</td>
<td>MacArtney</td>
<td>Subconn Power Series BHB3F</td>
</tr>
<tr>
<td>8 contact undersea cable</td>
<td>1</td>
<td>MacArtney</td>
<td>Subconn Micro Series MCIL8M</td>
</tr>
<tr>
<td>2 contact undersea cable</td>
<td>1</td>
<td>MacArtney</td>
<td>Subconn Power Series ILB2M</td>
</tr>
<tr>
<td>3 contact undersea cable</td>
<td>2</td>
<td>MacArtney</td>
<td>Subconn Power Series ILB3M</td>
</tr>
<tr>
<td>5/8 - 18 Armor coat hex nut</td>
<td>3</td>
<td>McMaster-Carr</td>
<td>93827A255</td>
</tr>
<tr>
<td>7/16 - 20 Armor coat hex nut</td>
<td>1</td>
<td>McMaster-Carr</td>
<td>93827A241</td>
</tr>
<tr>
<td>5/8 washer for power connectors</td>
<td>3</td>
<td>McMaster-Carr</td>
<td>92140A122</td>
</tr>
<tr>
<td>1/2 washer for 8-pin connector</td>
<td>1</td>
<td>McMaster-Carr</td>
<td>90107A032</td>
</tr>
<tr>
<td>(S1) 6-32 9/16” screw</td>
<td>14</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(N1) 6-32 nut</td>
<td>6</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(S3) 6-32 1” Nylon screw</td>
<td>4</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(N2) 6-32 Nylon nut</td>
<td>4</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure A.30: Underwater hardware bill of materials
B Appendix — Software

This appendix contains the program code that the microcontroller executes during operation. Code from four files is included:

- **main.c** — Contains the function **main** which is executed every time the microcontroller is powered on. This function contains the necessary hardware initializations and the wake-up routine that the controller goes through to find the second peak the first time.

- **isr_1.c** — This file contains the functions that control the time-based interrupt used during the post-startup phase. Only user-created code, including variable definitions and code for the interrupt service routine (ISR) is included.

- **wpt.c** — Source file containing functions written for use by the main function and the ISR.

- **wpt.h** — Header file associated with **wpt.c**.

Code from **wpt.c** is presented first because it includes function definitions used extensively by the **main.c** and **isr_1.c** source files. Any function not defined in **wpt.c** can be assumed to be a standard PSoC function provided by the generated source files for their respective hardware blocks.

B.1 **wpt.c**

This subsection includes functions necessary to program the frequency generator and sample and process DCVP and DCVI.

```c
#include "wpt.h"
#include "project.h"

#define SAMPLES 10 // how many successive samples of
    // DCVP and DCVI to take

int max_val; // global maximum value
int max_val_ix; // global maximum value
int hunting; // global state of the system

// Function to write to potentiometer
uint8 dev_addr = 42; // 7 bit slave address
uint8 wrt_bufftotal[4] = {16,0,18,0}; // four bytes sent to the
    // potentiometer;
    // bytes 0 and 2 are command bytes
    // (where to write to and what to do)
    // bytes 1 and 3 are the wiper values that the chip should set
    // they can be in the range [0,255]
uint8 *wrt_bufftotal_ptr = wrt_bufftotal; // need pointer to pass to function
void send2pots(int n){
    // split n into two numbers that will fill byte 1 and 3 above
    if(n<256){
        wrt_bufftotal[1]=n;
        wrt_bufftotal[3]=0;
```

107
else{
    wrt_bufftotal[1]=255;
}
I2C_1_MasterWriteBuf(dev_addr,wrt_bufftotal_ptr,4,I2C_1_MODE_COMPLETE_XFER);
    //transmit the values over the I2C channel
    return;
}

//Function to get ADC value
int j;
int ADC_result;  //temporary result after every conversion
int ADC_sum;    //sum of all results (will be used for averaging)
int get_avg_val(void) { //perform a dummy conversion to clear the buffer
for(j=0;j<SAMPLES;j++){ //perform SAMPLES subsequent conversions
    RMS_DC_ADC_StartConvert(); //start conversion
    while(!RMS_DC_ADC_IsEndConversion(RMS_DC_ADC_RETURN_STATUS));
    //wait for conversion to finish
    ADC_result = RMS_DC_ADC_GetResult16(); //get result from ADC's buffer
    if(ADC_result>511){ //make sure the result is within range
        ADC_result = 511;
    }else if(ADC_result < 0){
        ADC_result = 0;
    }
    ADC_sum += ADC_result; //add to running sum
}
    ADC_sum /= (SAMPLES);   //divide by the number of samples -> average
    return ADC_sum;
}

//Function to perform ADC sampling twice
int total_sample(void){
    get_avg_val();
    return get_avg_val();
}

//Function to return the square of number
int sq(int i){
    return i*i;
}

//Function to get value of current peak detector
int get_current_value(void){
    for(j=0;j<SAMPLES;j++){
        Current_ADC_StartConvert(); //start instance of DCVI conversion
        while(!Current_ADC_IsEndConversion(Current_ADC_RETURN_STATUS));
        //wait to finish
        ADC_result = Current_ADC_GetResult16(); //get result from buffer
        if(ADC_result>255){ //make sure the result is within range
            ADC_result = 255;
        }
    }
    return ADC_result;
}
else if(ADC_result < 0) {
    ADC_result = 0;
}
ADC_sum += ADC_result;   //add to running sum
ADC_sum /= (SAMPLES);  //divide to average
return ADC_sum;

//Function to perform current sampling twice
int total_sample_current(void){
    get_current_value();
    return get_current_value();
}

/* [] END OF FILE */

B.2 wpt.h
#define CUTOFF 440   //cutoff frequency
#define AWAY_WAIT 3
extern int max_val_ix;
extern int max_val;
extern int hunting;
void send2pots(int);
int get_avg_val(void);
int total_sample(void);
int sq(int);
int get_current_value(void);
int total_sample_current(void);
/* [] END OF FILE */

B.3 main.c
#include <stdio.h>
#include <project.h>
#include <wpt.h>

int main(void)
{
    Amp_IO_Write(0);   //turn off amplifier
    T_P_Write(0);      //test pin write low
    CyGlobalIntDisable;  //disable interrupts during startup
    RMS_DC_ADC_Start();  //start DCVP converter
    Current_ADC_Start();   //start DCVI converter
    CyGlobalIntEnable;   //enable global interrupts
    isr_1_Stop();        //do not run hunting interrupt yet
    I2C_1_Start();      //start I2C communications module
    int adc_val;       //value of power currently found

int ix_max; //index at which maximum value occurs
int res_max = 0; //maximum found value
Amp_IO_Write(1); //turn on amp
send2pots(CUTOFF); //set operating point to cutoff frequency
CyDelay(5); //Wait 5 ms
int c_val=total_sample(); //sample power, save into c_val

//same procedure as above for adjacent lower frequency
send2pots(CUTOFF-1);
CyDelay(5);
int c_L_val=total_sample();

//same procedure as above for adjacent higher frequency
send2pots(CUTOFF+1);
CyDelay(5);
int c_H_val=total_sample();

//boolean decision tree
if(c_H_val<c_val && c_val<c_L_val){ //negative slope
    hunting=1; //peak not yet found
    max_val_ix=CUTOFF; //stay at f_p
    Amp_IO_Write(0); //turn off amplifier
    Clock_1_SetDividerRegister(AWAY_WAIT*1000,1);
    //set interrupt frequency to slow
} else if(c_H_val<c_val && c_L_val<c_val){
    //local maximum
    max_val=c_val; //the maximum power is current power
    hunting=0; //found peak already
    max_val_ix=CUTOFF; //current location = f_p
} else{
    hunting=1; //not at peak, must hunt
    max_val=0; //no maximum in memory yet
    max_val_ix=CUTOFF+4; //start at a few points
    //above f_p
}
send2pots(max_val_ix); //program frequency generator
isr_1_Start(); //enable interrupt
while(1){}

B.4 isr_1.c

Below is the custom definition and ISR code written in the isr_1.c file. It only includes sections explicitly written for the system and does not include default code generated by PSoC Creator containing default functions for manipulating the interrupt such as isr_1_start(), etc.

#include <project.h>
#include <wpt.h>
#define SAMPLES 3 //how many samples taken when change detected
    //must be odd number with middle
//being the current frequency
#define STEP_MUL 1 //how big is the frequency step for peak hunting
#define C_CUTOFF 145//current cutoff point
#define ISR_CLK_DIV 50//set divider on normal ISR monitor clock
   //(50 -> 20Hz interrupt)
#define RESTART 1 //restart clock when changing divider?

int res_c; //current result from ADC
int threshold; //this is set by the controller depending on how
 //much power is coming through (more power -> need
 //larger threshold for stability)
int diff_sq; //difference between current result and previously
 //established maximum
int indices[SAMPLES]; //array of indices (at which the below values
 //are taken)
int vals[SAMPLES]; //array of values
int currents[SAMPLES]; //array of currents from peak detector circuit
signed int diffs[SAMPLES-1]; //array of differences between the
 //values above
int k; //index for iterations
int two[3]; //max value and current value for debugging

CY_ISR(isr_1_Interrupt)//this function is referred to as
   //the ISR in the thesis
   //this function is generated empty and must be filled
   //with the following code
{
    #ifdef isr_1_INTERRUPT_INTERRUPT_CALLBACK
    isr_1_Interrupt_InterruptCallback();
    #endif /* isr_1_INTERRUPT_INTERRUPT_CALLBACK */

    /* Place your Interrupt code here. */
    /* '#START isr_1_Interrupt' */
    
    T_P_Write(1); //flip digital pin (timing measurement)
isr_1_Stop();
    CyGlobalIntEnable;//global interrupts must be enabled to use I2C

    if(hunting == 0){ //if operating at the peak
        Amp_IO_Write(1); //turn on amplifier (no effect if amplifier
        //already on)
        Clock_1_SetDividerRegister(ISR_CLK_DIV, RESTART);
        //set interrupt frequency to fast (no effect
        //if already fast)
        res_c = total_sample(); //measure power
        if(res_c<100){ //this sets adaptive threshold (larger power
        //value -> larger threshold
            threshold = 1;
        }else{
            threshold = (res_c/100)*2;
        }
        diff_sq = sq(res_c-max_val); //calculate how much power changed
        if(diff_sq>sq(threshold)){
    

59    hunting = 1;    // hunt if changed more than threshold
60 }
61 }else if(hunting == 1){ // currently looking for peak?
62    Amp_IO_Write(1); // turn amplifier on (no effect if already on)
63    Clock_1_SetDividerRegister(ISR_CLK_DIV,RESTART);
64    // set interrupt frequency to fast (no effect
65    // if already fast)
66    for(k=0;k<SAMPLES;k++) // this loop sets the frequencies to adjacent,
67       // measures power and gets the local slope
68       indices[k]=max_val_ix*STEP_MUL*(k-(SAMPLES-1)/2);
69       send2pots(indices[k]);    // set the frequency
70       CyDelay(5);    // allow system to settle
71       vals[k]=total_sample();   // get value at that frequency
72       currents[k]=total_sample_current(); // get currents from
73       // peak detector circuit
74       if(k>0){ // this loop calculates differences in power
75          // between the adjacent points
76          diffs[k-1]=vals[k]-vals[k-1];
77       }
78  }
79  // ----------- DECISION TREE LOGIC ---------------------------------
80  if(SAMPLES==3){   // how many sample points (3 in thesis)
81     if(diffs[0]>0){ // is the slope positive?
82        if(diffs[1]<0){
83           hunting = 0;
84           // if both adjacent points lower -> local max ->
85           // found peak
86        }else if(diffs[1]>=0){ // local slope is positive or
87           // close to peak
88           max_val_ix++;
89           // move up one frequency point
90           Clock_1_SetDividerRegister(ISR_CLK_DIV,RESTART);
91           // set interrupt frequency to fast
92        }
93     }
94     }else if(diffs[0]<0){ // is the slope negative?
95        if(currents[0]<C_CUTOFF){  
96           max_val_ix--;  // decrease frequency only if
97           // the current permits
98        }else{     // if current too high
99           Amp_IO_Write(0); // turn off amplifier
100          hunting=1; // still haven’t found peak
101          Clock_1_SetDividerRegister(AWAY_WAIT*1000,RESTART);
102           // set interrupt frequency to slow
103     }
104     }else if(diffs[0]==0){
105        if(diffs[1]>0){ // power increasing to the right
106           max_val_ix++;
107           Clock_1_SetDividerRegister(ISR_CLK_DIV,RESTART);
108           // set interrupt frequency to fast
109        }else if(diffs[1]<0){
110           if(currents[0]<C_CUTOFF){
111             max_val_ix--; // decrease frequency if current permits
112          }else{
113             Amp_IO_Write(0); // turn off amplifier
114          }
115     }
116  }
117  
118
hunting=1; //not yet found peak
Clock_1_SetDividerRegister(AWAY_WAIT*1000,RESTART);
    //slow down interrupts
}
}else{
    hunting = 0; //found peak
}
}
} //----------------------------------------------
send2pots(max_val_ix); //program frequency generator with
    //new index
CyDelay(5);
if(!hunting){
    max_val = total_sample(); //record and save new present
    //power value
}
isr_1_Start(); //re-enable the ISR
T_P_Write(0); //flip pin low (timing)
/* '#END' */