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Benjamin Waters

Adaptive Wireless Power for Ventricular Assist Devices

Benjamin Waters

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Reading Committee:

Joshua R. Smith, Chair

Matthew S. Reynolds

Shwetak N. Patel

Program Authorized to Offer Degree:
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University of Washington

Abstract

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Benjamin Waters

Chair of the Supervisory Committee:
Associate Professor Joshua R. Smith
Computer Science & Engineering and Electrical Engineering

Every device needs power to function. Whether that device is a cell-phone, laptop, wearable device, or even a robot, if the power source dies then the device is unusable until charged up again. In our world today, this is something we are all too familiar with: the battery dies and the device is off until it is physically plugged in again.

As irritating as this can be for our consumer devices today, what if our lives depended on that device staying on? Consider an implanted medical device for example. The battery powering that device simply cannot die. If it does, the consequences are far more severe than being unable to use a cell-phone for a few hours. How can we make sure that all devices – industrial, consumer, and especially medical devices – always have a reliable power supply in any environment - in our homes, outside, underwater, in a factory or even inside the body?

The challenging aspect about reliability is that it depends highly on the environment in which the device is used. If a device is solar powered, it will work great in Australia, but not so well in cloudy Seattle. Or if a device is powered by harvesting ambient energy, that might work great in a big city, but not so well in a remote town. For a device implanted inside the body, how does one ensure that the power supply is always reliable? These power questions are among the most important questions

to ask when designing any device. Finding answers to these questions has driven my research in wireless power systems.

The wireless power system presented in this dissertation is adaptive. Adaptive wireless power means that the distance and alignment between the transmitter and the receiver can change without impacting the efficiency of the system. Existing wireless charging solutions require the device to be placed directly on a charging pad. If the device is not perfectly aligned or if it moves at all, the wireless charging stops.

An application that not only benefits from having wireless charging, but inherently needs adaptive wireless charging is a ventricular assist device (VAD). VADs are implanted medical devices for end-stage heart failure patients that pump blood throughout the body when the heart is unable to do so. For patients with end-stage heart failure, VADs are the only option for treatment other than a full heart transplant. However, heart transplants are not available to all patients, and there is a long wait list for a transplant: the number of heart failure patients is increasing while the number of available donor hearts continues to decline.

VADs are unique implanted devices because they require a lot of power to pump blood through the body. To power these devices, large batteries are worn on a belt outside the body. A thick power cable protrudes through the patients stomach and powers the implanted VAD. This power cable can become infected at the site where it penetrates through the skin. Which is why patients with this device are told that they can never take a full shower. Imagine not being able to shower for the rest of one's life! So far, that's what every one of the nearly 20,000 VAD patients have been told.

This dissertation presents an adaptive wireless power system for ventricular assist devices. This system is called the Free-range Resonant Electrical Energy Delivery (FREE-D) system. A thorough analysis shows the techniques used to design

each component in the FREE-D system, including the coils used for wireless power transfer, the transmitter circuit, and the receiver circuit. Next, three different techniques used for adaptive tuning are presented including frequency tuning, adaptive impedance matching, and power tracking. A control algorithm implements all three of these techniques simultaneously, for maximum flexibility and optimal efficiency at any distance and orientation between the coils. *in-vivo* experiments are presented that demonstrate the viability of the FREE-D system in eight separate acute animal trials. Finally, additional capabilities of the adaptive wireless power system are shown using transmit coil arrays and phased-array wireless power transfer systems. These capabilities demonstrate that the FREE-D system may be able to deliver efficient wireless power to a device anywhere inside an entire room.

TABLE OF CONTENTS

	Page
List of Figures	iii
List of Tables	vii
Chapter 1: Introduction	1
1.1 Research Scope	3
1.2 Dissertation Organization	4
Chapter 2: Wireless Power: Past, Present and Future	5
2.1 Successes and Failures of Wirelessly Powered Cardiovascular Devices .	8
2.2 Safety and Regulatory Considerations	12
Chapter 3: The FREE-D System	17
3.1 Overview of Magnetically Coupled Resonators	20
3.2 Coil Design	29
3.3 Transmitter Design	50
3.4 Receiver Design	52
Chapter 4: Adaptive Tuning Techniques	57
4.1 Frequency Tuning	57
4.2 Adaptive Impedance Matching	60
4.3 Power Tracking	70
4.4 Dynamic VAD Speed Control	73
Chapter 5: In-Vivo Animal Trials	76
5.1 Overview	76
5.2 Experimental Results: Wireless Power	81
5.3 Experimental Results: Continuous and Pulsatile Flow	86

Chapter 6: Working Towards Room-Size Wireless Power Range for VADs . . .	90
6.1 Overview of Wide Area WPT	90
6.2 Wide Area WPT Using Relay Coils and Adaptive Tuning	91
6.3 Wide Area WPT Using A Phased Array Wireless Power System . . .	100
Chapter 7: Conclusion	125
7.1 Dissertation Summary	125
7.2 Future Work	127
7.3 Future Outlook for Wireless Power	129
Bibliography	134

LIST OF FIGURES

Figure Number	Page
1.1 Gartner Hype Cycle 2014 [22].	2
2.1 Wireless power timeline.	5
2.2 Thoratec HeartMate II LVAD system	10
3.1 FREE-D system with vest-worn transmitter and implanted receiver .	18
3.2 FREE-D system with vest-worn transmitter and implanted receiver for patient outside home	18
3.3 Long-range FREE-D system with transmit resonators installed through- out the patient's home.	19
3.4 Long-range FREE-D system with transmit resonators installed around the patient's desk.	19
3.5 System-level block diagram of FREE-D system.	20
3.6 Diagram of two-coil inductive power transfer system.	21
3.7 Equivalent circuit model of an inductor with series resistance and par- asitic capacitance.	22
3.8 Equivalent circuit diagram of multi-transmitter WPT system with two Tx coils and one Rx coil.	23
3.9 Theoretical analysis showing $ S_{21} ^2$ for 2, 3, 4 coil wireless power systems.	27
3.10 Front side of Tx and Rx coils used for the FREE-D system.	29
3.11 Back side of Tx and Rx coils used for the FREE-D System.	29
3.12 Cross-sectional view of flat spiral coil.	32
3.13 Coils used for validation of parasitic component calculations.	36
3.14 R and L vs. p for the eight test coils from Fig. 3.13. The first R calculation result accounts for both proximity effects and skin effects [23] while the second calculation accounts for only the skin effect and DC resistance (3.23).	37
3.15 $ S_{21} $ as a function of Tx to Rx coil size ratio ($\frac{D_{O,TX}}{D_{O,RX}}$) and distance between the coils.	39

3.16	Sets of Tx coils and Rx coils used for experimental measurements. . .	41
3.17	Plot of measured and simulated $ S_{21} $ for the coils from Fig. 3.16. . .	42
3.18	Simulated plot of S_{21} as a function of coupling between the Tx and Rx coils for various Rx coil Q values.	43
3.19	Measured results for $ S_{21} ^2$ versus distance and frequency for 2, 3, and 4 coil wireless power systems.	44
3.20	Comparisons of efficiency ($ S_{21} ^2$) versus distance for 2, 3, and 4 coil wireless power systems using single frequency operation at 13.56MHz (A), frequency tracking (B) and adaptive impedance matching (C). .	44
3.21	Rx test coils with Rx loop for temperature evaluation.	47
3.22	Rx test coils for temperature evaluation.	47
3.23	Temperature plots for Rx loop and coil configurations.	47
3.24	Temperature plots for Rx coil-only configurations.	47
3.25	Imaginary part of S_{11} for various coil environments.	49
3.26	FREE-D coil in air	49
3.27	Coil in PDMS	49
3.28	FREE-D coil in ham	49
3.29	Block diagram of FREE-D system transmitter.	51
3.30	FREE-D system transmitter.	52
3.31	Schematic diagram of class E RF amplifier.	53
3.32	Block diagram of FREE-D system receiver.	54
3.33	Front side of FREE-D system receiver.	55
3.34	Back side of FREE-D system receiver.	55
3.35	FREE-D receiver and Rx coil mounted on HeartWare HVAD.	56
3.36	FREE-D receiver, Rx coil and two-cell backup battery next to Thoratec HeartMate II LVAD.	56
4.1	Measured scattering parameters for the FREE-D coils at a distance of 2.5 cm along the axis of coil. Panel A shows the S_{21} linear magnitude of the system. Clearly visible are the two resonant modes caused by the coupling of the two high-Q resonators. Panel B shows a Smith chart of the input impedance (S_{11}) as function of frequency. This panel shows that there are two frequencies where the input impedance to the coils is well matched to the 50Ω source and maximum power transfer can occur.	58

4.2	Algorithm diagram for adaptive local maximum frequency tracking. .	59
4.3	Calculated ratio ($\frac{R_{in}}{R_s}$) of coil to source impedance where $R_{in} = \text{real}(Z_{in})$ and $R_s = 50\Omega$ for the calculated input impedance of a 2, 3, and 4-coil wireless power system as a function of the coupling coefficient (k_{COIL}) between the two high-Q coils for each configuration.	62
4.4	Measured $ S_{21} $ for the FREE-D coils with π -match networks at both the Tx and Rx sides at a distance of 2.5cm along the axis of the coil. Clearly visible are the four resonant modes caused by the coupling of the two high-Q resonators and the impedance matching networks. . .	63
4.5	$ S_{21} $ for a 4 coil WPT system without impedance matching (A) and with adaptive impedance matching (B) as measured by a VNA. . . .	64
4.6	Schematic diagram of the adaptive impedance matching networks. . .	65
4.7	Flow chart of the process used for the adaptive impedance matching component selection algorithm to extract the shunt capacitor values in the π -match network.	66
4.8	Measured S_{21} for the FREE-D coils with and without auto-tuning. . .	67
4.9	Tx and Rx source and load capacitor values required for adaptive impedance matching.	67
4.10	Schematic of adaptive impedance matching switch on Tx side.	69
4.11	Schematic of adaptive impedance matching switch on Rx side.	69
4.12	Algorithm diagram for adaptive impedance matching.	70
4.13	Algorithm diagram for power tracking.	72
4.14	FREE-D wireless power and pump control GUI.	74
5.1	Equivalent circuit diagram of multi-transmitter WPT system with two Tx coils and one Rx coil.	76
5.2	FREE-D <i>in-vivo</i> vest coil configuration.	78
5.3	FREE-D <i>in-vivo</i> long-range coil configuration.	78
5.4	In-vivo Rx coil placement	81
5.5	Long-range <i>in-vivo</i> animal trial results from June 4, 2014.	84
5.6	Vest-coil <i>in-vivo</i> animal trial results from July 15, 2014.	85
5.7	In-vivo animal trials results from July 15, 2014 at the Yale School of Medicine.	87

5.8	Effects of VAD pulsatility on pump flow. The top panels show the pump flow (red line) and the ECG pattern (black line) generated by the HVAD device. The bottom panels show the left ventricular pressure (blue line) and the aortic pressure (orange line), which represent the inlet and outlet pressures of the VAD respectively.	88
6.1	FREE-D home mobility with Tx coils placed throughout the patients home.	90
6.2	Schematic diagram of the wide-area wireless power system using relay coils.	93
6.3	Transmit coil array powering five receivers simultaneously	95
6.4	Charge current for five wirelessly powered battery chargers.	98
6.5	Rectified voltage for 1-5 receivers using adaptive tuning techniques.	99
6.6	Rectified voltage for 1-5 receivers without adaptive tuning techniques.	99
6.7	Sequential charging using time division multiplexing.	100
6.8	Equivalent circuit diagram of a WPT system with multiple transmitters and receivers.	105
6.9	V_o for a phased-array WPT system with $k_{12} = 0$ and $\alpha = 1$. Hashed areas represent regions where a single Tx coil achieves greater V_o than two Tx coils.	108
6.10	V_o for a range of k_{12} , k_{13} , k_{23} , α and ϕ for a three-coil phased-array system.	110
6.11	V_o of one Rx coil powered by (A) two, (B) three and (C) four Tx coils.	112
6.12	System level block diagram (top), and experimental configuration using two Tx coils and one Rx coil (bottom).	113
6.13	Experimental and simulation results for received power corresponding to three different coil configurations and various Tx power levels.	114
6.14	Experimental coil configurations: (A) adjacent, (B) overlapping, (C) opposing, and (D) single Tx coil configurations.	117
6.15	System DC-RF efficiency for three different Tx coil configurations.	119
6.16	FOM measured at three separate locations: behind Tx ₁ , behind Tx ₂ and behind the Rx coil for an 8cm separation between the Tx coils and the Rx coil.	120
6.17	Receiver block diagram (left) and photo of receiver circuit (right).	121
6.18	Efficiency versus distance comparisons with auto-tuning enabled.	123
7.1	Results of 90-day duration bench top testing of the FREE-D system.	133

LIST OF TABLES

Table Number	Page
2.1 SUMMARY OF FCC OET65C LIMITATIONS ON EMI AND MPE FOR PART 15 AND PART 18 WIRELESS COMMUNICATION DE- VICES	14
3.1 FREE-D SYSTEM COILS PARAMETERS.	30
5.1 SUMMARY OF <i>in-vivo</i> ANIMAL TRIAL RESULTS	82

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DEDICATION

to my wife and family, with all my love.

Chapter 1

INTRODUCTION

Wireless power transfer was listed on the 2014 Gartner Technology Hype Cycle [22] as a technology in its “peak of inflated expectations.”. This status defines a technology in which “early publicity produces a number of success stories often accompanied by scores of failures. Some companies take action; many do not.” Although the initial discovery of wireless power transfer began over 100 years ago with the work of Nikola Tesla, there are a number of reasons why wireless power transfer has only just reached its peak of inflated expectations.

Tesla envisioned that wireless power would be available everywhere to everyone when he designed the Wardencllyffe Plant in 1897 [77]. However, history has shown that market-changing technologies take time to develop and for the public to accept. Perhaps Tesla’s vision will be realized sometime in our future; however his discovery had to wait nearly 100 years before electronic toothbrushes commercialized the technology for consumer devices in the early 1990s.

For electronic toothbrushes, wireless power provided convenience, ease of use, but most importantly the unmet need of safely charging these devices given the constant presence of water. Thirty years later, in 2003, MIT introduced a new phase of wireless power transfer that achieves longer range power transfer to conveniently power devices through several feet of air. Since 2003, the expectations of wireless power transfer for charging modern-day consumer devices including cell-phones, laptops, and electric vehicles have grown considerably. These market-driving applications present a very wide range of power levels, ranging from milli-watts to kilo-watts of power transfer at radio frequencies ranging from 100kHz to 26MHz. The hype from these initial

wirelessly powered devices contributed to wireless power's name on the Gartner Hype Cycle (Figure 1.1). But where will the technology go from here?

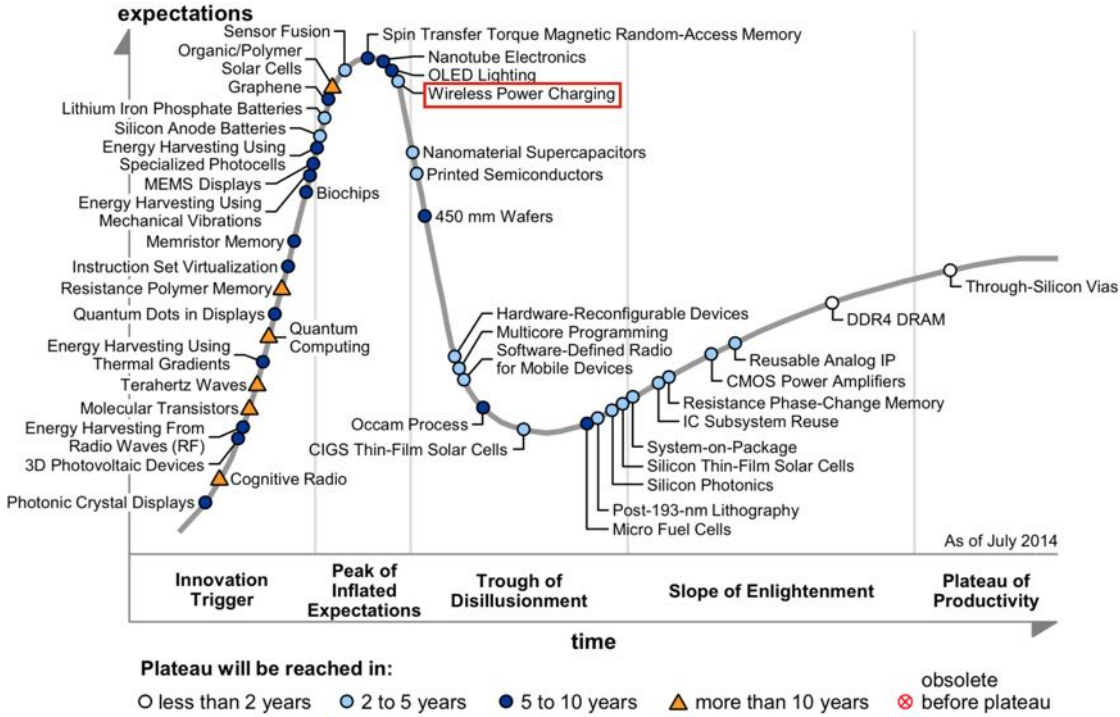


Figure 1.1: Gartner Hype Cycle 2014 [22].

By Gartner's definition, in the next 2-5 years, wireless power transfer should theoretically diminish in its technical capabilities to accommodate the consumer market requirements for regulatory compliance and implementation on a larger scale. From a research point of view, this presents a very exciting opportunity because in 5-10 years, wireless power transfer will reach the slope of enlightenment. When it reaches that stage, the true capabilities of the technology will crystallize and an increasing number of enterprises will accept the technology because it will have already been proven for the devices that are the driving applications today. It will finally reach the stage where the more challenging wirelessly powered devices can accommodate the technology. As a researcher, it is these future devices that present the most opportunistic,

and also most challenging problems to begin solving now.

1.1 Research Scope

In my opinion, the most exciting applications for wireless power transfer are those that inherently require a wireless solution for the device to reach its full potential in either a research or consumer environment. An application that certainly meets this requirement and the application that will be the focus of this dissertation is a ventricular assist device (VAD). VADs are implanted devices for end-stage heart failure patients that use axial or centrifugal motors to pump blood throughout the body when the heart is unable to do so. VADs have become increasingly popular as a bridge to transplant for patients eligible for a full heart transplant, but waiting on the constantly-growing wait-list for a donor heart. VADs also present the only alternative for destination therapy patients: patients ineligible for a heart transplant who must rely on VAD support for the rest of their lives.

The most challenging technical aspect with VADs is that they require anywhere from 5-25 watts to sufficiently power the motor and pump blood throughout the body. Compared to pacemakers and neuro-stimulators in the milli-watt power range, the power consumption of VADs is the highest for any implanted device on the market today. Therefore, VADs have used percutaneous drivelines that protrude out from the patient's stomach and connect to an external controller for power and communication with the implanted device. However, the driveline is one of the most restricting aspects of VAD technology because it is a source of infection, re-hospitalization and diminished quality of life for VAD patients.

My goal has been to design a fully implantable, wirelessly powered VAD controller that will eliminate the driveline and safely transfer wireless power through the skin to the implanted unit. The motivating criteria for my research is to design and implement a reliable, safe and practical wireless power transfer system.

At the highest level, reliability implies that the system should always deliver suf-

ficient power to the VAD such that the device never shuts off. Safety implies that the device operates within all of the regulatory requirements governing temperature of implanted devices, radio-frequency (RF) field strength requirements, induced current limitations on human body tissues, and specific absorption rate (SAR) limitations. Practicality implies that the implanted device be miniaturized, convenient to implant, enables freedom of movement for the patient, and ultimately improves the quality of life for VAD patients.

These three requirements consist of several subcategories that will be discussed throughout this dissertation. Ultimately, the goal of this dissertation and my research in general consists of optimizing all of these features.

1.2 Dissertation Organization

This dissertation is organized as follows: Chapter 2 begins with a detailed background of wireless power transfer technology, including a timeline highlighting notable milestones for wireless power transfer in corporate and academic research labs around the world. Section 2.1 outlines the successes and failures of previous wirelessly powered cardiovascular devices. Section 2.2 discusses the constantly evolving regulatory scope for wirelessly powered devices.

Chapter 3 describes the design and implementation of the Free-Range Resonant Electrical Energy Delivery (FREE-D) System for VADs. This Chapter highlights the design considerations of the hardware and software for the FREE-D wireless power transmitter and receiver. Chapter 4 presents the adaptive tuning capabilities including automatic frequency tracking, adaptive impedance matching, power tracking, and dynamic VAD speed control. Chapter 4 presents the *in-vivo* results of the FREE-D system. Finally Chapter 6 outlines the additional capabilities of the FREE-D system, including the ability to implement a long-range FREE-D system across half-meter distance ranges using relay resonators and a phased-array wireless power transfer system.

Chapter 2

WIRELESS POWER: PAST, PRESENT AND FUTURE

A timeline of power transfer levels versus time for several notable wireless power innovators is shown in Figure 2.1.

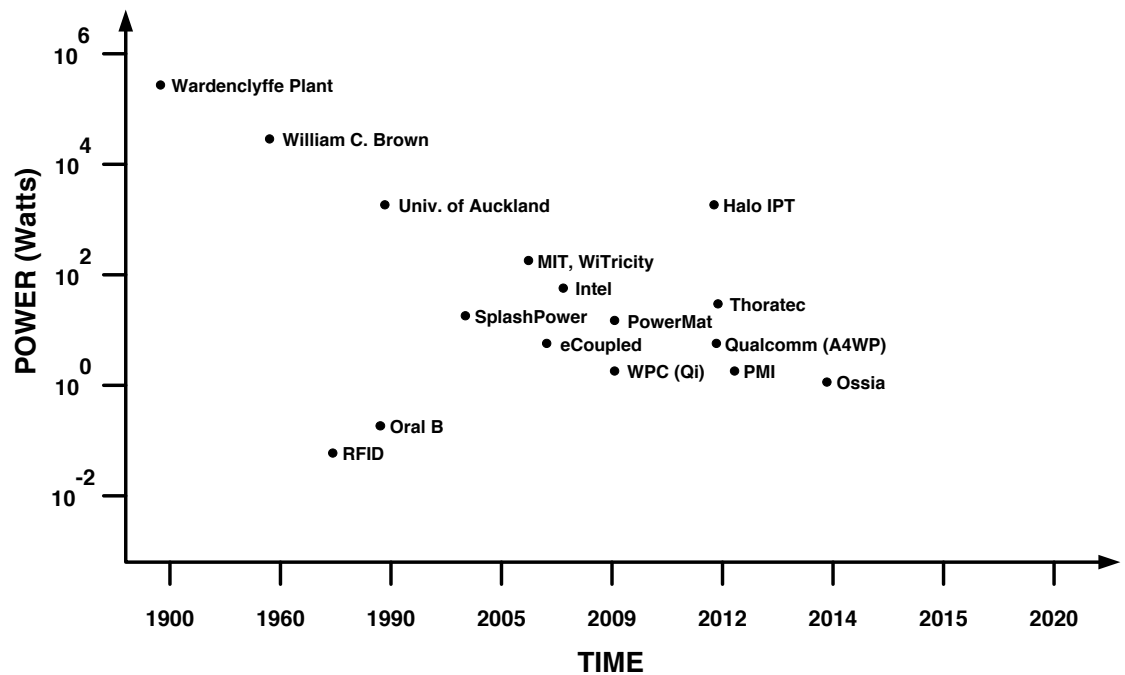


Figure 2.1: Wireless power timeline.

Nikola Tesla introduced the concept of wirelessly transmitting electrical power in 1891 when he claimed “the wireless art offers greater possibilities than any invention or discovery heretofore made” [75, 76]. However Tesla’s discovery was perhaps 100 years too early for mass acceptance because there were no end devices, like cell-phones or electric cars that could ease the technology into the market. Although Tesla never

completed the 200kW wireless generator at the Warendclyffe Plant – a massive helical resonator with potential to transmit wireless power globally – a recent resurgence of wireless power technologies could eliminate the need for wired and battery-powered devices.

In 1971, William C. Brown demonstrated 30kW of microwave power transfer over 1.6km to a helicopter at 84% efficiency [7, 8]. Like Tesla, Brown also aspired to transmit power wirelessly from one home to another, but implementing the technology that could wirelessly transfer nearly 500kW of power at 3GHz never came to fruition. The high power requirements, safety concerns and massive infrastructure costs of implementing both Tesla and Brown’s ambitious wireless power systems surely contributed to the lack of mass implementation for these technologies.

Around the same time as Brown, near-field wireless power transfer finally found an application that was suitable for public acceptance: RFID. Although many dedicate the invention of RFID to Leon Theremin in 1945, it was Mario Cardullo in 1973 who successfully implemented passive wirelessly powered RFID tags for electronic tolling systems in New York [9]. A major reason why Cardullo succeeded was because the infrastructure for tolling systems (bridges, highways, etc.) was already in place, thus there was a clear application that could ease the technology into the marketplace. RFID has expanded significantly since then to numerous other applications.

Today, after the emergence of cell-phones, laptops and electric vehicles, there is a new realm of potential applications for wireless power transfer using near-field power transfer techniques. With the device infrastructure already in place and well-established, now is finally the time when wireless power transfer will catch on. The startup company Halo IPT from the University of Auckland (1990s) and the spin-off from Professor Marin Soljacic at MIT called WiTricity (2006) re-introduced the concept of magnetically coupled resonators for electric cars, cell-phones and televisions [73, 43].

Since then, large corporations are driving the standardization of modern-day wire-

less power transfer. The three leading standards include the Qi standard from the Wireless Power Consortium, the Alliance for Wireless Power (A4WP) using Qualcomm's Rezence Technology, and the Power Matters Alliance (PMA). Each standard proposes various power levels, operating frequencies, and communication methods all with hopes of becoming the primary technology distributor for all future wirelessly powered devices. A4WP and PMA announced interoperability between their standards in February 2014. The advantage of standardization for application developers will be that if they use one of the accepted standards, their device will inherently be compliant with any federal regulatory requirements for wireless power transfer. For end users and electronic device consumers, standardization will allow for a single transmitter to power any of their devices that they might purchase from different manufacturers. This will become a critical piece to wireless charging systems as the number of battery-powered devices per person continues to increase [74]. Therefore, charging multiple devices simultaneously from one or more transmitters is an important research topic today.

In the realm of medical devices, wireless power transfer enables improved quality of life, reduced hospitalization for battery replacement and reduced infection for transcutaneously powered implanted devices. However, the current generation of nearly all implanted medical devices rely on battery power from either implanted batteries that cannot be recharged or external batteries that require a transcutaneous driveline to power the implant. Although several Universities and corporations are researching and developing systems for wireless power transfer to implanted medical devices, the lengthy experimental trial periods and regulatory approval processes will likely delay the time when wireless power will become a commercially viable means for charging implanted medical devices.

Other wireless power technologies have recently emerged. In 2009, LaserMotive presented a technique that accurately reflects high power lasers off a series of dynamically controlled mirrors to a distance of over 1km at 1kW power level to a robotic

climber. Far-field wireless power systems such as Cota and WattUp developed by Ossia and Energous respectively utilize a series of phased microwave array emitters to beam form 1W across nearly 10m. A drastic improvement in range from inductive and resonant coupling, microwave power transfer seems promising for in-home consumer device charging. However concerns about safety, regulatory compliance and cost of these devices still must be overcome by these new methods for wireless power transfer.

2.1 Successes and Failures of Wirelessly Powered Cardiovascular Devices

Heart failure is a terminal disease with a very poor prognosis and constitutes Medicare's greatest area of spending with close to 35 billion spent each year [17]. Pacemakers and VADs are two very different devices for heart failure patients that have unique challenges in adopting wireless power for implanted battery charging.

At first glance, implanted medical devices seem like an obvious application for wireless power transfer. Pacemakers for example have a very clear problem that must be solved: implanted batteries cannot recharge and require patient hospitalization for device replacement. Additionally, pacemakers require very little power since they are idling most of the time, aside from the 20mW pulses for stimulation. Therefore the power requirements of a wireless charging system would be relatively low and potentially facilitate the regulatory approval process. However the low power level also works against the case for wirelessly powered pacemakers: the 5-10 year life cycle for the batteries in these devices has been deemed acceptable by patients, surgeons and manufacturers. Also from the business perspective, more devices are sold when the device needs to be replaced every 5-10 years. For the patient, the pacemaker is completely implanted, requiring no external components aside from the wireless controller, which is only required periodically. Wireless power would add an external transmitter that would need to be nearby or worn by the patient to power or recharge

the implant. For these reasons, the business model for wirelessly powered pacemakers in their current form is not practical.

However, an additional problem has recently surfaced for pacemakers: the leads from the implanted device to the electrodes directly embedded in the heart can become damaged, and few cases have shown these leads can short and shock patients. An alternative wirelessly powered device that is small enough to be placed directly at the location of the electrode inside the tissue of the heart has been developed by Ho et al. at Stanford University [32]. Perhaps this device, which poses to solve all three problems of battery-power, requiring replacement, and eliminating internal leads will ultimately lead to innovation for pacemakers after nearly 50 years.

For end-stage heart failure patients, a full heart transplant remains the most desirable treatment. However, only a limited number (approximately 2000 per year) of patients can truly benefit from transplants due to donor shortages and high costs. An alternative which has become increasingly popular is mechanical circulatory assistance with ventricular assist devices in which a pump with a central rotor accelerates blood throughout the body [31, 78]. The first generation VADs were approved for use in the United States by the FDA in October, 1994. The pumps require a transcutaneous drive line, meaning a biocompatible cable protrudes from the body to connect the VAD to a power source and the external system controller. The system controller communicates with the heart pump to modify pump conditions and simultaneously provide power to the VAD from either a battery for portable use or a central power supply unit. Figure 2.2 shows the HeartMate II left ventricular assist device (LVAD) system from Thoratec.

VAD technology has significantly improved in the past 15 years. Initially, VADs were a temporary alternative solution to heart transplants: supporting patients for only a few months. Now, VADs can survive patients for upwards of five years [51, 60]. As a result of the extended lifetime of the VAD, the most common cause for patient readmission to the hospital and patient death is no longer the technical failure of the

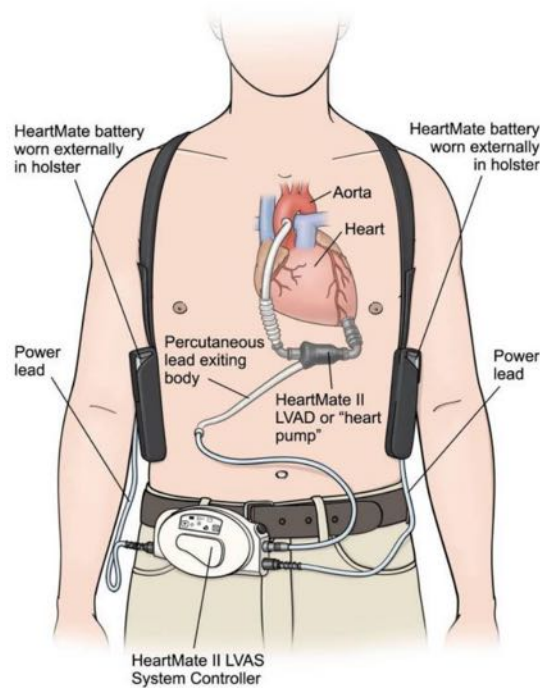


Figure 2.2: Thoratec HeartMate II LVAD system

VAD, but rather the exit site infection (ESI) from the percutaneous driveline. The increasing risk of ESI hampers the patient's quality of life and can lead to repeated hospitalizations for antibiotic treatment, surgical interventions or even a costly VAD replacement [52].

Medical research has demonstrated the relationship between ESI, pump pocket infection – infection in the abdominal pocket where the VADs are implanted – and subsequent sepsis – bacterial growth in the bloodstream [33, 65, 86, 3]. 70% of VAD patients' first readmission to the hospital is due to ESI [52]. Patients who develop ESI spend more time in the hospital and have 10 times as many readmissions as the patients without ESI [79]. Patients are also required to clean the exit site on a daily basis, and cannot shower for the duration of the implantation. The net result of these effects from ESI is reduced survival and increased cost negating the intended benefit of

VAD therapy. Implementing wireless power to the VAD or to recharge an implanted battery will eliminate the need for the driveline, and consequently eliminate ESI.

The greatest challenge for VADs adopting wireless power is that VADs require 5-25W of power – a huge amount for an implanted device. Previous attempts have been made to wirelessly power VADs using Transcutaneous Energy Transfer Systems (TETS) [19, 50, 59]. TETS uses inductive coupling techniques to transfer power between coils on the inner and outer surfaces of the skin.

In 2003, LionHeart presented the first fully implantable VAD system using TETS [19]. The overall size of the implanted controller was 4x3.75x1.125 inches and weighed over 1100 grams (2.5lbs)! Even with the massive implant and extremely invasive surgery required for implantation, TETS successfully powered an implanted LVAD in humans. However, the clinical and laboratory experiments demonstrated several drawbacks with the current TETS technology. Restrictions on misalignment between the transmitting and receiving coils and the necessity for a close separation distance between the coils limit the practicality of TETS. The system also incurs significant wireless energy loss beyond 10 mm separation. The proximity limitation requires that the receiving coil be implanted just under the skin and the external transmitting coil be secured in a single position on the skin surface with adhesives. For angular misalignments or excessive separation between the coils, the transmitter will attempt to supply more power to account for the reduced efficiency. This effect proved to cause skin irritability and thermal injury from the increase in coil temperature due to greater power transmission, which resulted in burns and led to infection on the exterior surface of the skin. The LionHeart TETS was removed from the market shortly after these human trials, and VADs have continued to be powered by transcutaneous drivelines ever since.

In 2010, Dissanayake et al. from the University of Auckland presented a new, miniaturized TET system for VADs [18]. The system was used in-vivo in sheep and their results emphasized low temperature of the implanted receive coil after several

hours of operation. However, the University of Auckland system was limited to 20mm of separation between the transmit and receive coils. Considering VAD patients vary drastically in size and body type, 20mm may not be a sufficient range because the distance between the coils could exceed 20mm for some patients.

Although both of these previous TETS applications have successfully powered VADs in-vivo, the inflexible range and misalignment between the coils diminishes the practicality of the system. If the range of the wireless power systems can be extended out to tens of centimeters or even half-meter ranges, then significant improvements to patient quality of life can be realized. Most importantly, with flexible range capabilities, the wireless power system for VADs will be suitable for all patients, even those with implant depths beyond 20mm. The patient may also be able to completely remove the external transmitter and sleep throughout the night with a transmitter placed in the patients bed. Additionally, the patient may be able to shower without worrying about the internal battery running low, since transmit coils placed around the shower could power the implanted unit. Therefore optimizing the efficiency and range of the wirelessly powered VAD system while minimizing the temperature, size and thickness of the implanted components are the primary design considerations for the system proposed in this report.

2.2 Safety and Regulatory Considerations

The growing number of medical devices utilizing wireless power and communication introduces additional regulatory challenges associated with the health and safety of the technology. Since the human body is more conductive than air, wireless RF energy can be absorbed by the body and dissipated as heat, causing localized heating of bodily tissues.

In the United States and similarly for other countries, federal regulating bodies ultimately define and regulate the standards based on scientific experiments and simulation published by international scientific bodies including ICNIRP, NCRP, CISPR,

IEEE-SA, etc. Any wireless device in the US must demonstrate Federal Communication Commission (FCC) compliance. Additionally, any medical device must receive approval from the Food and Drug Administration (FDA).

In the US, the FCC OET65C governs the two primary metrics that any wireless applications must follow: electromagnetic interference (EMI) and maximum permissible exposure (MPE) levels [49]. EMI is restricted by the strength of electric and magnetic fields across specific frequency bands at certain distances from the transmit antenna. MPE is characterized by both specific absorption rate (SAR) and induced and contact currents in the human body. The induced and contact current limits protect against shock and burn hazards. The maximum induced and contact current limits for continuous sinusoidal waveforms are reached when the induced current flowing between the body and a grounded object cause either electro-stimulation or tissue heating. SAR is the measurement of the energy absorbed by the human body per unit mass when it is exposed to RF electromagnetic fields. SAR can be characterized in terms of electric field strength per unit mass (2.1) where σ is the frequency-dependent conductivity of the tissue and ρ is the mass density of the tissue. SAR is equivalently defined in terms of temperature change over time (2.2) where C is the specific heat of the tissue. The FCC cites the IEEE SCC-34/SC-2 in P1528 for the tissue parameters and calculation techniques for SAR [35].

$$SAR = \frac{|E_{RMS}^2|\sigma}{\rho} \quad (2.1)$$

$$SAR = C \frac{dT}{dt} \quad (2.2)$$

Currently the FCC categorizes telecommunication devices into two sections that have different EMI and MPE levels. FCC Part 15 for Radio Frequency Devices (47CFR15) allocates frequency bands with limitations on EMI and MPE [14]. FCC Part 18 for Industrial, Scientific, and Medical (ISM) Equipment (47CFR18) allows

for unlimited radiated energy in specific ISM frequency bands; however, the device must receive the specialized classification of ISM equipment (i.e. magnetic resonance imaging (MRI) and induction heating systems) [15]. In both Parts there are also different requirements for intentional and unintentional radiators; however, wireless power transmitters are all classified as intentional radiators. The current requirements outlined in the FCC OET65C for EMI and MPE are summarized in Table 2.1. Additional summaries of the relevant ICNIRP, CISPR, and IEEE-SA MPE limitations are available in [12].

Table 2.1: SUMMARY OF FCC OET65C LIMITATIONS ON EMI AND MPE FOR PART 15 AND PART 18 WIRELESS COMMUNICATION DEVICES

Frequency (MHz)	Bandwidth	Usage	Part 15 Field Strength	Part 18 Field Strength
0.510-1.705	1195 kHz	RF Devices	$15\frac{\mu V}{m}$ at $\frac{\lambda}{2\pi}$ m	n/a
6.765-6.795	30 kHz	RF Devices/ISM	$100\frac{\mu V}{m}$ at 30m	unlimited
13.410- 13.553	14 kHz	RF Devices	$334\frac{\mu V}{m}$ at 30m	n/a
13.553-13.567	14 kHz	RF Devices/ISM	$15848\frac{\mu V}{m}$ at 30m	unlimited
13.567-13.710	14 kHz	RF Devices	$334\frac{\mu V}{m}$ at 30m	n/a
26.957-27.283	326 kHz	RF Devices/ISM	$10000\frac{\mu V}{m}$ at 3m	unlimited
40.660-40.700	40 kHz	RF Devices/ISM	$1000\frac{\mu V}{m}$ at 3m	unlimited
174-216	42 MHz	Biomedical Telemetry	$1500\frac{\mu V}{m}$ at 3m	n/a
401-406	5 MHz	MedRadio	n/a	n/a
433.050-434.790	1.84 MHz	ISM	$11000\frac{\mu V}{m}$ at 3m	unlimited
608-614	6 MHz	WMTS	n/a	n/a
902-928	26 MHz	RF Devices/ISM	$500\frac{\mu V}{m}$ at 3m	unlimited
1395-1400	5 MHz	WMTS	n/a	n/a
1427-1432	5 MHz	WMTS	n/a	n/a
2400-2500	100 MHz	RF Devices/ISM	$500\frac{\mu V}{m}$ at 3m	unlimited
5725-5875	150 MHz	RF Devices/ISM	$500\frac{\mu V}{m}$ at 3m	unlimited

For wireless charging specifically, the FCC has published KDB 680106 entitled “RF Exposure Considerations for Wireless Charging Applications” [21]. This document

clarifies the usage conditions for wireless charging applications with data communication. In order to be classified as a Part 18 device, the wireless transmitter must "generate and use the RF energy locally" which means the RF power is contained within a very close proximity or in an enclosure. The benefit is that these products may operate with higher RF power and they are excluded from routine RF exposure evaluation to show compliance, but evaluation may be requested. Also required for Part 18 classification is that the radiator must not be a telecommunication device. Transmitter modulation only means the transmitter is not communication device; however, load modulation for data telemetry would classify the system as a Part 15 communication device where the EMI and MPE levels are strictly defined.

The wireless power applications that might receive Part 18 classification for the local use of RF energy include charging pads where the receiver is placed directly on top of the transmitter and magnetic coupling materials are used to constrain the magnetic fields to the device and minimize leakage fields. However, RF energy transmitted over a short distance through air or through human tissue would require additional evaluation by these current requirements.

Another obstacle for demonstrating FCC compliance of wireless power applications is that below 30MHz dielectric parameters for tissue-equivalent simulating models are not yet acknowledged by the FCC. The FCC currently requires "a combination of analytical analysis, field strength, radiated and conducted power measurements, and limited numerical modeling to assess compliance" in the form an analysis report to document RF exposure [21].

For a wireless power system designer, the quantities that are constrained due to the FCC EMI and MPE levels are the maximum allowable transmit power, the maximum receive power to the device, the distance at which the device can be powered, the amount of time required to power the device, data communication with the receive device, and the operating frequency. For high power systems that must adapt to variations in distance and misalignment between the transmit and receive coils such

as wirelessly powered VADs that, demonstrating FCC compliance will be very difficult and will most likely require special accommodation. Additionally, researchers from the Foundation for Research on Information Technologies in Society (IT'IS) have shown that the existing EMI and MPE levels are not considered safe for users with implanted medical devices due to the conductive nature of these devices inside the body, and it is likely that special testing and accommodations must be in place for wirelessly powered implanted medical devices [44].

Medical devices also must seek approval from the FDA. The FDA categorizes devices into three Classes (I, II, and III). The higher the class, the greater the risk the device presents to the patient, and consequently the regulatory control increases. Class III devices typically support and sustain human life. VADs (and any accessories to the VAD, such as a wireless power system) are considered Class III devices. These devices require Pre-Market Approval (PMA), which contributes to the 5-10 year approval process for Class III devices that can hinder businesses from advancing technology in these fields of use.

The FDA typically defers to the FCC criteria for EMI and MPE regulations if the medical device utilizes a wireless power or communication system [55]. However elaborate reports including information on the design, validation, verification and changes to the system must be submitted to the FDA [57, 56]. One additional metric that the FDA defines is the whole body or local temperature increase for a medical device inside the body caused by RF heating. In the FDA guidelines for MRI devices, RF heating may not cause a core body temperature increase greater than 1°C or a localized temperature increase of 1°C in the head, 2°C in the trunk or 3°C in the extremities [25].

Seeking FCC and FDA approval for the wirelessly powered VAD system is outside the scope of this report. However, relevant metrics have been evaluated in bench top and in-vivo experiments, and ultimately these results could be used in a compliance report submitted to either regulatory body for this technology.

Chapter 3

THE FREE-D SYSTEM

The Free-range Resonant Electrical Energy Delivery (FREE-D) system provides wireless power to a VAD using magnetically coupled resonators. FREE-D affords seamless energy supply without compromising mobility or requiring direct contact between the individual and the energy source as in a TETS system. The transmit (Tx) and receive (Rx) resonators are coils of wire that are tuned to resonate at a specific frequency.

The key feature that distinguishes FREE-D from prior inductive schemes is the use of high quality factor (Q) resonators combined with an automatic tuning scheme that keeps the system operating at maximum efficiency. The FREE-D system is able to adapt to variations in Tx-Rx separation distances, Tx-Rx orientations, and power requirements of the load. By automatically controlling the frequency and power level of the transmitted RF signal, as well as the impedance matching networks at both the Tx and Rx coils, the FREE-D system can achieve high power transfer efficiency for nearly any angular orientation over a range of separation distances, as will be demonstrated in this Section.

There are two specific use cases that have the potential to significantly improve VAD patients' quality of life, which can be achieved using the FREE-D System. The first scenario is to allow patients to wear a vest that has a Tx coil built into the vest. The Tx coil will be driven by a battery-powered power amplifier that can be placed in a pack. An Rx coil connected to the wireless power receiver will be implanted along with the VAD motor controller, and a backup battery. The Tx coil will be situated above the implanted Rx coil, but does not need to be adhered to the patients

skin and will have much greater range and flexibility than TETS. Additionally, the patient can completely remove the vest for limited periods of time while the implanted battery can power the VAD so that the patient may shower. When the implanted battery is running low, the vest-coil system will both power the VAD and recharge the implanted battery. A sketch of this vest-coil scenario is shown in Figures 3.1 and 3.2.

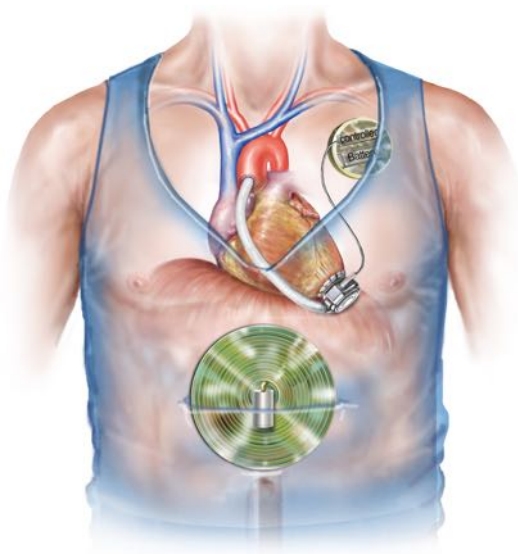


Figure 3.1: FREE-D system with vest-worn transmitter and implanted receiver



Figure 3.2: FREE-D system with vest-worn transmitter and implanted receiver for patient outside home

The second targeted use case for the FREE-D System consists of installing multiple Tx coils throughout the household that can ultimately be hidden inside walls, floors, couches, tables and beds. In this long-range FREE-D System scenario, the vest-coil acts like a relay coil and can extend the range of the wireless power from the Tx coils throughout the home to the implanted Rx coil. Although it may not be practical for the patient to completely remove the vest everywhere inside of the household because of technological limitations in transmitting power from a large Tx coil to a small, implanted Rx coil, there are specific scenarios where the long-range FREE-D System

can significantly improve VAD patients' quality of life. For example, with multiple Tx coils placed beneath the patient's bed, the patient may be able to remove the vest-coil and vest-worn transmitter while sleeping at night – something that no VAD patient has ever been able to do. Two other examples of the long-range FREE-D System are shown in the sketches in Figures 3.3 and 3.4.



Figure 3.3: Long-range FREE-D system with transmit resonators installed throughout the patient's home.



Figure 3.4: Long-range FREE-D system with transmit resonators installed around the patient's desk.

In order to accommodate the flexible range and orientation between the Tx and Rx coils, the FREE-D system requires dynamic wireless power delivery to maintain high efficiency over larger distances. High efficiency is extremely important for regulatory compliance of wirelessly powered implanted medical devices. Higher efficiency implies lower transmit power levels for the same amount of power delivered to the load, which means lower field strengths and safer operating conditions for humans around the wireless power system.

The system-level block diagram of the FREE-D System is shown in Figure 3.5. In the following sections, the design considerations for both the hardware and software associated with each element in this block diagram will be outlined.

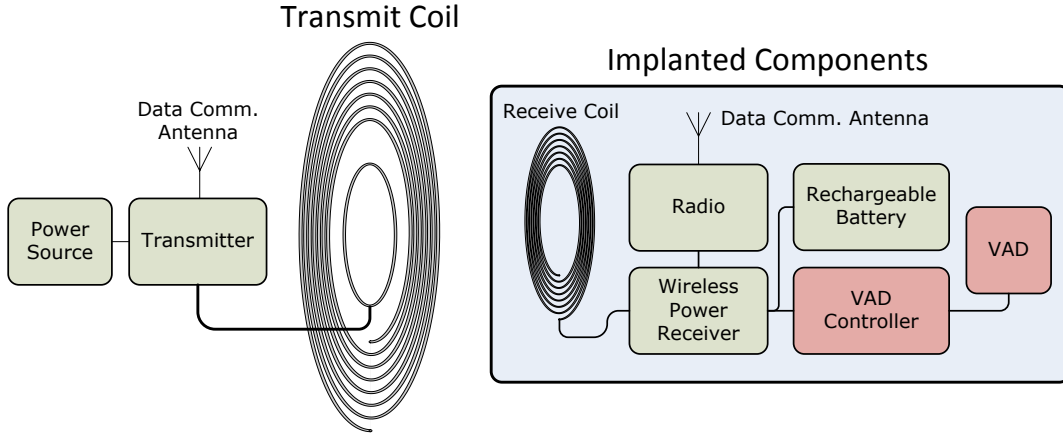


Figure 3.5: System-level block diagram of FREE-D system.

3.1 Overview of Magnetically Coupled Resonators

This Section analyzes the unique properties of the magnetically coupled resonators (MCRs) used in the FREE-D system to enable seamless wireless power delivery to implanted VADs.

Ampere’s Law states that electric current and changes in electric field are proportional to the magnetic field circulating in the same area of a conductor. In other words, current flowing through a loop of wire generates a time-varying magnetic field. When the magnetic field changes, Faraday’s Law of induction states that the wire acquires an electromotive force. It is this fundamental concept that makes WPT possible.

The most simplistic wireless power system consists of one transmit coil and one receiver coil (Figure 3.6). Current flowing through a transmit coil generates an oscillating magnetic field. When a receiver coil comes within range of that magnetic field, electric current is induced in the receiver coil. The amount of current induced in the receiver coil is dependent on many different aspects of the wireless power system,

but primarily depends on the coupling coefficient k that effectively quantifies how much magnetic flux from the transmit coil is induced upon the receiver coil. The magnetic coupling coefficient k_{ij} depends on the inductance of the primary inductor L_i , the secondary inductor L_j , and the mutual inductance M_{ij} between them (3.1). The mutual inductance is a function of the geometry of the coils and the distance between them. The key insight is that the Tx and Rx coils form a single system of coupled resonators, which can transfer energy back and forth.

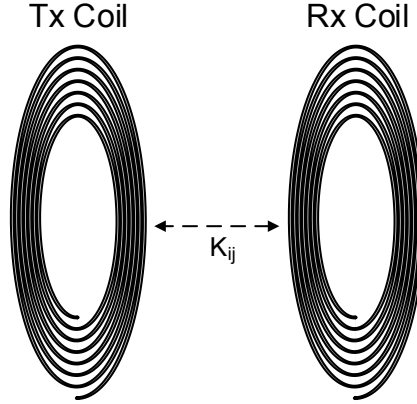


Figure 3.6: Diagram of two-coil inductive power transfer system.

$$k_{ij} = \frac{M_{ij}}{\sqrt{L_i L_j}} \quad (3.1)$$

3.1.1 Circuit Analysis

In order to account for the losses associated with a practical inductive power transfer system, a more accurate model of the inductors can be used. A common model for inductors shown in Figure 3.7 consists of an equivalent series resistance r_s and a parasitic capacitance c_p . A two coil inductive power transfer system can be analyzed using this equivalent circuit model.

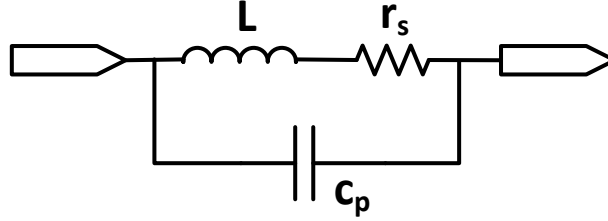


Figure 3.7: Equivalent circuit model of an inductor with series resistance and parasitic capacitance.

Rather than continue with the analysis of an inductive power transfer system, the focus of this dissertation is on a resonant power transfer system. The difference between an inductive system and a resonant system is that the inductors in a resonant system are intentionally tuned to be resonant at the exact frequency used for WPT. Typically, an additional tuning capacitor C_T is placed either in series or in parallel with the inductor to ensure the system is resonant at the intended resonant frequency ω_O (3.17). The complete schematic diagram of a two-coil resonant power transfer system, driven by a voltage source V_S with a source resistance R_S and terminated by a load resistance R_L is shown in Figure 3.8. The series tuning capacitance and the parasitic capacitance of each inductor are lumped together in this model and are shown as C_2 and C_3 on the transmit and receive coils respectively.

The current flowing through each coil can be calculated using the impedance matrix in (3.2). The parameters Z_{1-2} in (3.3) represent the impedance transformations of each individual coil. The general solution for the current flowing through any coil can be found using (3.5) and (3.6). The output voltage V_O across R_L can be determined by solving for I_2 as in (3.4). Finally, the scattering parameter S_{21} can be calculated from V_O using (3.7). $|S_{21}|$ represents the linear gain magnitude. $|S_{21}|^2$ represents the efficiency from the transmit coil to the receiver coil when the source and load impedance are matched.

This solution can be used to accurately model the efficiency of a two-coil resonant

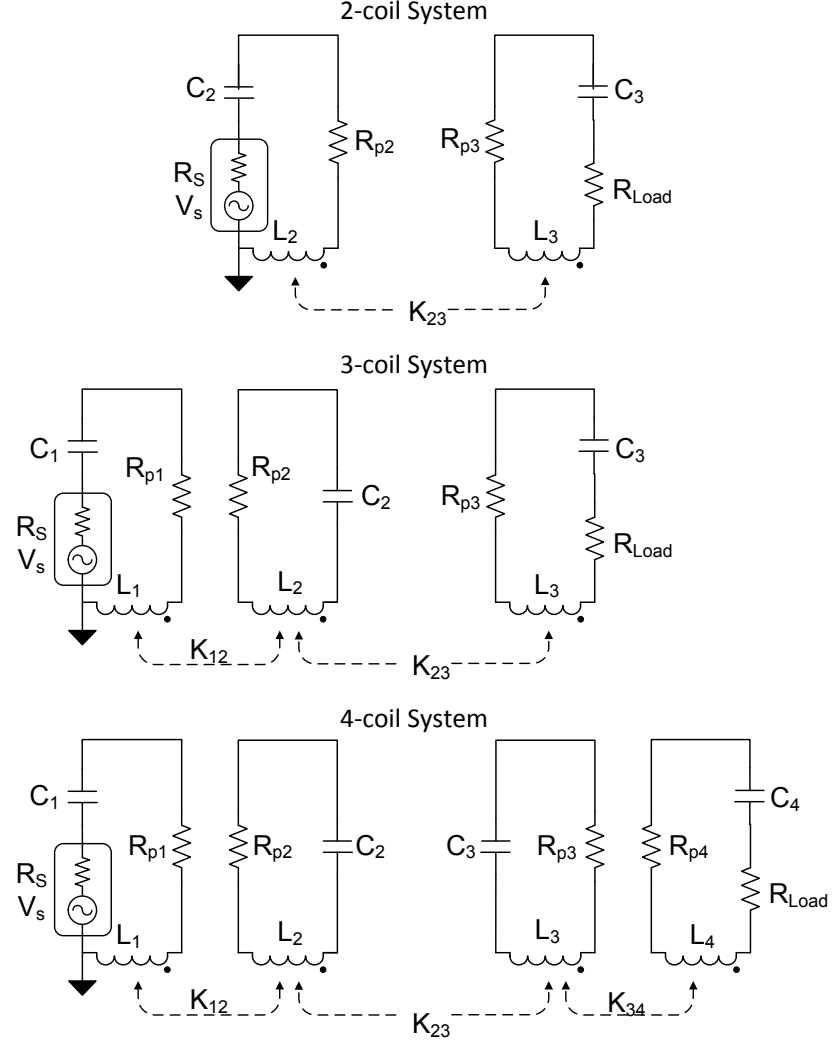


Figure 3.8: Equivalent circuit diagram of multi-transmitter WPT system with two Tx coils and one Rx coil.

wireless power system. But before considering those results, it is important to go through this analysis for wireless power systems that use more than two coils. The equivalent circuit diagrams of 3 and 4 coil wireless power systems are shown in Figure 3.8. 3 and 4 coil systems utilize additional coils (typically referred to as loops) on the

Tx and/or the Rx sides of the primary and secondary coils. These loops are coupled to the coils, and act as an impedance matching component to improve the efficiency of the wireless power system at various distances (or coupling coefficients) between the Tx and Rx coils. The loops also physically isolate the Tx coil from the power amplifier on the transmitter circuit and the Rx coil from the rectifier on the receiver circuit, which improves the quality factor Q of the resonant system. A high coil Q means that more energy can be stored on the coil, which also results in greater magnetic flux density at a given point in space. Ultimately, 2, 3, and 4 coil systems are all identical if proper impedance matching techniques are applied [66]; however, there are other effects of using the various systems that must be considered for FREE-D. These considerations for the coils used in the FREE-D system are discussed in Section 3.2.

For the 3 coil system, a drive loop is added to the Tx coil and for the 4 coil system, loops are added to both the Tx and Rx coils. When the transmitter drives current through the drive loop at the system's resonant frequency, the resulting oscillating magnetic field excites the Tx coil. The coil stores energy in the same manner as a discrete LCR tank. This results in a large oscillating magnetic field in the vicinity of the transmit coil.

Configuring the coupling coefficient between the loops and coils is extremely important to maintain high efficiency for 3 and 4 coil systems. Optimizing the loop-coil coupling coefficient is thoroughly discussed in [67]. Typically, increasing the loop-coil coupling coefficient increases the maximum achievable efficiency, but reduces the range at which that maximum efficiency can be achieved. Alternatively, decreasing the loop-coil coupling coefficient decreases the maximum achievable efficiency, but increases the range at which that peak efficiency can be achieved. The quality factor of the loop also plays an important role in optimizing the efficiency of the coils [45].

The analysis of 3 and 4 coil systems follows the same procedure as the 2 coil system analysis. However there are additional mutual inductances. M_{12} and M_{34} represent

the loop-coil mutual inductances. M_{13} , M_{14} , and M_{24} represent the cross-coupling mutual inductances. Typically these can be neglected because they are much smaller than the loop-coil and coil-coil mutual inductances. For all 2, 3, 4 coil systems, M_{23} represents the coil-coil mutual inductance. This term is effected by the physical distance and alignment between the transmit and receive coils. The impedance matrix for 3 and 4 coil systems are shown in (3.8) and (3.11) respectively. Following the same procedure as the 2 coil system, the output voltages across the load resistor are shown in (3.10) and (3.14). These expressions have been simplified by approximating the cross coupling terms M_{13} , M_{14} , and M_{24} to zero.

$$\mathbf{V} = \begin{bmatrix} V_S \\ 0 \end{bmatrix}, \mathbf{Z} = \begin{bmatrix} Z_2 & j\omega M_{23} \\ j\omega M_{23} & Z_3 \end{bmatrix}, \mathbf{I} = \begin{bmatrix} I_2 \\ I_3 \end{bmatrix} \quad (3.2)$$

$$\begin{aligned} Z_2 &= R_S + R_{P2} + j\omega L_2 + \frac{1}{j\omega C_2} \\ Z_3 &= R_L + R_{P3} + j\omega L_3 + \frac{1}{j\omega C_3} \end{aligned} \quad (3.3)$$

$$|V_O| = I_3 \times R_L = \frac{\omega M_{23} R_L}{Z_2 Z_3 + \omega^2 M_{23}^2} \quad (3.4)$$

$$\mathbf{V} = \mathbf{Z}\mathbf{I} \quad (3.5)$$

$$I_i = \frac{\det(\mathbf{Z}_i)}{\det(\mathbf{Z})} : i = 1, 2, 3 \dots n \quad (3.6)$$

$$S_{21} = 2 \frac{V_O}{V_S} \sqrt{\frac{R_S}{R_L}} \quad (3.7)$$

To convey the fundamental tradeoffs between range and efficiency for 2, 3, and 4 coil wireless power systems, each of the expressions for S_{21} have been evaluated in the following theoretical analysis. This analysis assumes identical Tx and Rx coils. The

$$\mathbf{V} = \begin{bmatrix} V_S \\ 0 \\ 0 \end{bmatrix}, \mathbf{Z} = \begin{bmatrix} Z_1 & j\omega M_{12} & j\omega M_{13} \\ j\omega M_{12} & Z_2 & j\omega M_{23} \\ j\omega M_{13} & j\omega M_{23} & Z_3 \end{bmatrix}, \mathbf{I} = \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} \quad (3.8)$$

$$\begin{aligned} Z_1 &= R_S + R_{P1} + j\omega L_1 + \frac{1}{j\omega C_1} \\ Z_2 &= R_{P2} + j\omega L_2 + \frac{1}{j\omega C_2} \\ Z_3 &= R_L + R_{P3} + j\omega L_3 + \frac{1}{j\omega C_3} \end{aligned} \quad (3.9)$$

$$|V_O|_{M_{13}=0} = I_3 \times R_L = \frac{\omega^2 M_{12} M_{23} R_L}{Z_1 Z_2 Z_3 + \omega^2 (Z_3 M_{12}^2 + Z_1 M_{23}^2)} \quad (3.10)$$

$$\mathbf{V} = \begin{bmatrix} V_S \\ 0 \\ 0 \\ 0 \end{bmatrix}, \mathbf{Z} = \begin{bmatrix} Z_1 & j\omega M_{12} & j\omega M_{13} & j\omega M_{14} \\ j\omega M_{12} & Z_2 & j\omega M_{23} & j\omega M_{24} \\ j\omega M_{13} & j\omega M_{23} & Z_3 & j\omega M_{34} \\ j\omega M_{14} & j\omega M_{24} & j\omega M_{34} & Z_4 \end{bmatrix}, \mathbf{I} = \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (3.11)$$

$$\begin{aligned} Z_1 &= R_S + R_{P1} + j\omega L_1 + \frac{1}{j\omega C_1} \\ Z_2 &= R_{P2} + j\omega L_2 + \frac{1}{j\omega C_2} \\ Z_3 &= R_{P3} + j\omega L_3 + \frac{1}{j\omega C_3} \\ Z_4 &= R_L + R_{P4} + j\omega L_4 + \frac{1}{j\omega C_4} \end{aligned} \quad (3.12)$$

$$|V_O|_{M_{13}=M_{24}=M_{14}=0} = I_4 \times R_L \quad (3.13)$$

$$|V_O| = \frac{\omega^3 M_{12} M_{23} M_{34} R_L}{Z_1 Z_2 Z_3 Z_4 + \omega^2 (Z_3 Z_4 M_{12}^2 + Z_1 Z_4 M_{23}^2 + Z_1 Z_2 M_{34}^2) + \omega^4 M_{12}^2 M_{34}^2} \quad (3.14)$$

circuit parameters used for this theoretical analysis include $L_1 = L_4 = 0.1\mu H$, $L_2 = L_3 = 3.6\mu H$, $C_1 = C_4 = 700pF$, $C_2 = C_3 = 38pF$, $R_1 = R_4 = 0.1\Omega$, $R_2 = R_3 = 1.0\Omega$. S_{21} has been evaluated for a range of coupling coefficients from 0–0.3 and for a range of operating frequencies from 5MHz–25MHz. The resonant frequency of each loop and coil is 13.56MHz. The results for each 2, 3, 4 coil wireless power system are shown in Figure 3.9. The coupling coefficient is inversely proportional to the distance between the coils: larger coupling coefficient correlates to shorter distance between the coils.

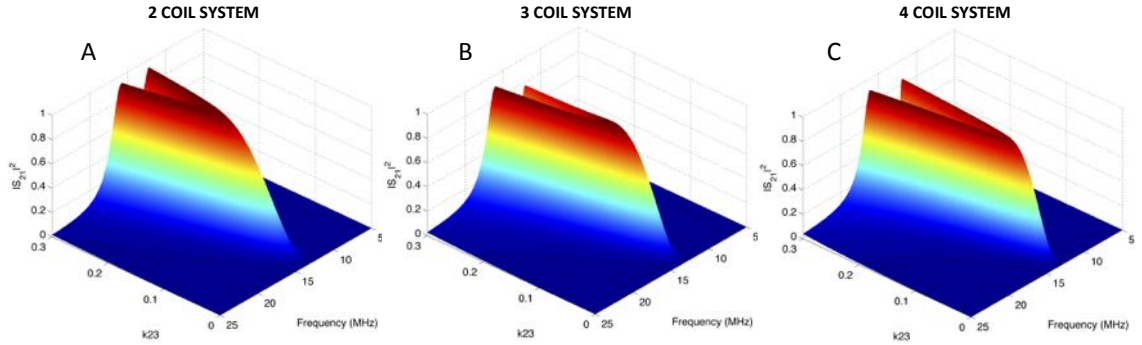


Figure 3.9: Theoretical analysis showing $|S_{21}|^2$ for 2, 3, 4 coil wireless power systems.

All three systems have a large region at which high $|S_{21}|^2$ can be achieved. As these ridges converge, high $|S_{21}|^2$ occurs only at one very specific frequency. Beyond this peak, S_{21} declines rapidly. For a near field wireless power system, this rate of decline is approximately proportional to $\frac{1}{d^3}$. The region where high $|S_{21}|^2$ can be achieved correlates to strong coupling between the coils, or short distances between the coils. In this region, there are two frequencies at which the peak $|S_{21}|$ can be achieved. This is a very important concept and is referred to as the frequency splitting (or frequency bifurcation) effect.

The region at which two peaks occur is referred to as the overcoupled region. The 2, 3 and 4 coil systems all have an overcoupled region; however, the length

of the overcoupled region varies for the different systems. The overcoupled region is unique because S_{21} (which correlates to the coil efficiency) is held constant even as the coupling coefficient changes. This overcoupled region is extremely powerful and counter-intuitive to the typical assumptions about a wireless system. According to Friis transmission equation, as the distance between a transmitter and receiver in a wireless system increases, the amount of power delivered to the receiver should decrease because of free-space path loss. However in the overcoupled region for a near-field wireless power system, the efficiency is at its maximum and does not decrease even as the distance between the coils increases! The challenge, however, is that the wireless power system will need to dynamically track those optimal frequencies or use another means to ensure that maximum efficiency can be achieved in the overcoupled region. This adaptive tuning capability is the focus of this dissertation, and will be discussed thoroughly in Section 4.

As the two peaks in the overcoupled region converge, the region where the peak occurs only at one frequency is called the under-coupled regime. In the under-coupled regime, the shared flux falls below a critical point. The result is that maximum efficiency cannot be achieved. Critical coupling is the point of transition between these two regimes and corresponds to the greatest range at which maximum efficiency can still be achieved. The under-coupled regime is still capable of WPT, but efficiency decreases rapidly as distance increases.

The 4 coil wireless power system has the largest overcoupled region. For the 2, 3, and 4 coil systems, critical coupling occurs at $k = 0.165$, $k = 0.91$ and $k = 0.085$ respectively. Since the 4 coil system has the smallest critical coupling point, it can achieve high $|S_{21}|^2$ across the largest range. The potential consequence of this larger overcoupled region is that the valley between the two peaks is deeper. Therefore adaptive tuning becomes critically important to maintain high efficiency throughout the longer overcoupled region in a 4 coil system.

3.1.2 *FREE-D System Coil Overview*

Section 3.2 thoroughly discusses additional considerations for the coil design tradeoffs. For the FREE-D system in particular, there are several medically imposed design considerations that constrain the size, shape, frequency, and operating temperature of the coils. Before discussing how the FREE-D coils were designed, Figure 3.10 shows a prototype of the final coils used in the FREE-D system. A 3 coil system was selected for the FREE-D system. The full system diagram can be seen in Figure 3.5.

The Tx resonator consists of a single turn drive loop and a multi-turn, spiral coil. The receiver coil consists of a multi-turn spiral coil, which is connected to the load device. These coils will be referenced throughout this report and a detailed summary of their design can be found in Section 3.2 and their properties can be found in Table 3.1.

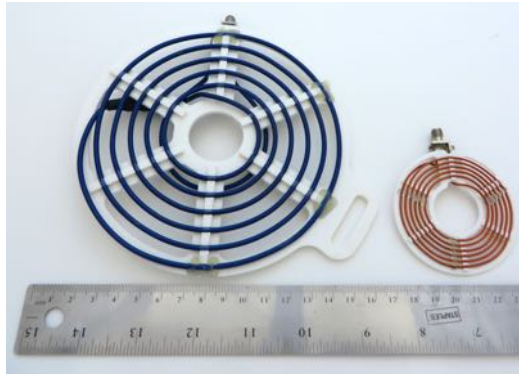


Figure 3.10: Front side of Tx and Rx coils used for the FREE-D system. Figure 3.11: Back side of Tx and Rx coils used for the FREE-D System.

3.2 *Coil Design*

There are three primary design decisions that must be made regarding the coils used for WPT: the size and geometry of the coils, the frequency used for WPT, and the number of coils used in the wireless power system. All three of these parameters

Table 3.1: FREE-D SYSTEM COILS PARAMETERS.

Parameter	Tx Loop	Tx Coil	Rx Coil
L (μH)	0.24	4.17	2.12
R (Ω)	0.21	0.99	0.58
C_T (pF)	575	33	65
Q	90	355	313
f_o (MHz)	13.56	13.56	13.57
D_o (cm)	7	14	6.5
N	1	6	6
pitch (mm)	n/a	9	2.5
w_d (AWG)	14	14	18
mass (g)	9.4	71.9	19.2

directly affect the coil Q, operating temperature of the coil, and range of the of the wireless power system. Q can be defined in many ways, but typically Q is defined by a ratio of the energy stored in a reactive component to the energy dissipated in a resistive component (3.15). For a series inductor (L), resistor (R) and capacitor (C), Q can be expressed as in (3.16) taking into account the energy stored in L or C, power dissipated in R and the resonant frequency ω_o (3.17). For wireless power systems, high Q implies lower loss, narrower bandwidth and higher efficiency. Therefore optimizing Q is often the goal for a coil designer.

For the FREE-D system, our collaborator at the Yale School of Medicine, Dr. Pramod Bonde, MD helped define the maximum Rx coil size of 7cm for implantation capabilities, and a maximum vest-worn Tx coil size of 15cm. The target WPT range for FREE-D is 1 – 7cm, where 1cm corresponds to the scenario where the Tx coil is placed directly on the skin and 7cm corresponds to the case where the Tx coil is worn in a vest that sits approximately 6cm off the skin. The Rx coil is typically implanted

approximately 1cm beneath the surface of the skin. These medically imposed design constraints will provide physical limitations on the coil size and geometry in the proceeding analysis.

$$Q = \omega \frac{\text{energy stored}}{\text{average power dissipated}} \quad (3.15)$$

$$Q_s = \frac{1}{R} \sqrt{\frac{L}{C}} = \frac{\omega_o L}{R} = \frac{1}{\omega_o RC} \quad (3.16)$$

$$\omega_o = \frac{1}{2\pi\sqrt{LC}} \quad (3.17)$$

3.2.1 Coil Size and Geometry

The first criteria to consider is the size and geometry of the coils. In general, the larger the coils the greater the WPT range, assuming a large coil does not deteriorate Q . It is a challenge to design and accurately simulate these near-field resonators because expressions for the parasitic L , R and C of a circular or square-shaped winding are not accurate for a wide range of coil geometries and resonant frequencies. Additionally, the size of the Tx coil relative to the Rx coil impacts the amount of coupling between the two coils. For asymmetrical coil sizes, the maximum achievable coupling coefficient is less than that for two symmetrical coils; however, the coupling coefficient drops off faster with increasing distance between symmetrical coils.

Extensive prior work has provided expressions for the inductance L , resistance R , and capacitance C for lumped modeling of flat spiral coils in terms of the coil geometry [23], [29], [28]. Optimization techniques have also been outlined for coil design using Litz wire to minimize the skin effect [69] and for biomedical applications [63]. However, these expressions and techniques are typically complex, computationally intensive, and can be inaccurate for a wide range of coil sizes and operating frequencies. Other work has shown how to calculate the coupling coefficient between two flat

spiral coils [6]. However few articles demonstrate the effects that the coil size has on both Q and k , and the overall impact on WPT range and efficiency.

To identify the optimal coil size and geometry for both the Tx and Rx coils, an accurate coil design procedure has been developed [83]. Given the medically imposed system constraints, the design tool exhaustively analyzes all possible Tx and Rx coil geometries given the maximum coil size requirements, calculates the parasitic L , R , C and k for each coil, and identifies the optimal size and shape (outer diameter, number of turns, turn-to-turn pitch, and wire diameter) of the Tx and Rx coils to achieve both maximum Q for each coil as well as the optimal efficiency at the targeted WPT range for the intended application.

Expressions for L , R , C , Q , and k of flat spiral coils are solved in terms of the outer diameter D_o , number of turns N , spacing between each turn p and wire diameter w of the coil. Although D_o , N , p and w fully define a flat, spiral coil, all geometric parameters are shown for completeness in Fig. 3.12. The inner diameter D_i , total wire length l , winding radius a , and radial depth of the winding c are defined in (3.18) and (3.19). All units of length are in meters.

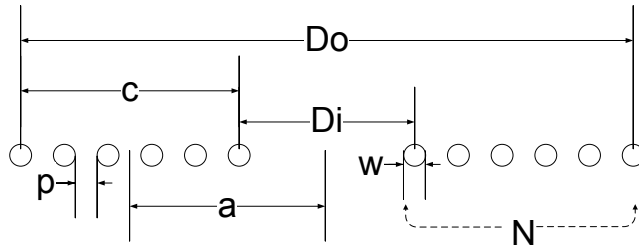


Figure 3.12: Cross-sectional view of flat spiral coil.

$$D_i = D_o - 2N(w + p), \quad l = \frac{1}{2}N\pi(D_o + D_i) \quad (3.18)$$

$$a = \frac{1}{4}(D_o + D_i), \quad c = \frac{1}{2}(D_o - D_i) \quad (3.19)$$

Inductance

The self inductance for flat, spiral coils is shown in (3.20). L is derived from a modification of Wheeler's formula for a single-layer helical coil, while accounting for the conversion from inches to meters ($39.37 \frac{\text{in.}}{\text{m}}$) and μH to H (10^{-6}) [29].

$$L(H) = \frac{N^2(D_o - N(w + p))^2}{16D_o + 28N(w + p)} \times \frac{39.37}{10^6} \quad (3.20)$$

The inductance expression is validated in prior work for a wide variety of coils and found to be accurate for most geometries except when the coil has very few turns, when the pitch is very large relative to the wire diameter ($p \gg w$), and when $\frac{c}{a} < 0.2$ [29]. Since these boundary conditions typically correspond to low inductance (i.e. low Q for a series resonator) and the goal for WPT is high Q , the conditions that would lead to inaccuracy in (3.20) are typically inapplicable when designing WPT coils.

Capacitance

The self capacitance of a spiral coil depends on the relative permittivity of the conductor, the diameter of each turn, the number of turns, and the pitch. As the number of turns increases, the self capacitance becomes increasingly difficult to calculate accurately due to the nonlinear adjacent winding capacitance [28]. Typically, the self capacitance is on the order of a few pF , and is small compared to the required tuning capacitor for resonance at 13.56MHz. Therefore, the required tuning capacitance of the coil is defined in terms of the inductance and resonant frequency f as in (3.21) and the parasitic self capacitance is neglected for this simplified expression.

$$C(F) = \frac{1}{(2\pi f)^2 L} \quad (3.21)$$

Resistance

Power loss in a spiral coil consists of radiation and conduction losses. Typical WPT coils are relatively small compared to the operating wavelength ($\sim 22\text{m}$ at 13.56MHz). Thus conduction loss is the dominate loss mechanism, while radiation loss is typically negligible. Conduction loss is dependent on the skin effect and proximity effect. Both effects confine current flow to smaller cross-sectional areas through the conductor, which increases the effective resistance of the conductor [23].

For spiral coils, the parameter that contributes most to fluctuations in total resistance R is the pitch p . Because of the proximity effect, R is inversely proportional to p , and this effect is nonlinear. For tightly wound coils, accurate expressions for R that account for the proximity effect are complex and can be difficult to calculate [23]. For loosely wound coils, the proximity effect can be negligible, so R consists of only the DC resistance R_{DC} and the skin depth δ , which can be approximated with a simpler expression shown in (3.23) where μ_0 is the permeability of free space and σ is the conductivity of the conductor ($\sigma = 59.6 \times 10^6 \frac{\text{S}}{\text{m}}$ for copper). At 13.56MHz , $\delta = 17.7\mu\text{m}$ and is much smaller than the wire diameter ($\simeq 1\text{mm}$) for the coils investigated in this analysis. So (3.23) has been derived from the very high frequency model of AC resistance from Kaiser [40].

$$R_{DC} = \frac{l}{\sigma \pi (w/2)^2}, \quad \delta = \frac{1}{\sqrt{\pi f \sigma \mu_o}} \quad (3.22)$$

$$R = R_{DC} \frac{w}{4\delta} = \sqrt{\frac{f \pi \mu_o}{\sigma}} \frac{N(D_o - N(w + p))}{w} \quad (3.23)$$

Quality Factor

Using the L , C and R expressions from (3.20), (3.21), and (3.23) respectively, Q is defined as follows for the series resonant flat spiral coil:

$$Q = \frac{1}{R} \sqrt{\frac{L}{C}} = \frac{39.37}{10^6} \sqrt{\frac{f \pi \sigma}{\mu_o} \frac{w N (D_o - N(w + p))}{8 D_o + 14 N(w + p)}} \quad (3.24)$$

This expression can be used to optimize coil design for high Q and ensure resonance at the desired operating frequency.

Coupling Coefficient

The amount of magnetic flux generated by the Tx coil which passes through the Rx coil determines k . Typically k depends on the geometries of each coil, the distance between the coils, and the relative orientation of the coils. The full expression for k between two single-turn loops can be simplified by assuming the loops are always oriented with zero angular misalignment between them [6]. Using this simplified expression for k and summing over the i^{th} turn from the Tx coil and the j^{th} turn from the Rx coil, the mutual inductance M between two multi-turn spiral coils is simplified to

$$M = \sum_{i=1}^{N_{TX}} \sum_{j=1}^{N_{RX}} \mu_o R_i R_j \int_0^\pi \frac{\cos(\Theta) d\Theta}{\sqrt{R_i^2 + R_j^2 + d^2 - 2 R_i R_j \cos(\Theta)}} \quad (3.25)$$

where d is the separation distance between the Tx and Rx coils, R_i and R_j are the corresponding radii of each turn for the Tx and Rx coils respectively, and k is related to M by (3.1).

Evaluating (3.25) shows that maximum k occurs for two coils of equal size. However, k diminishes faster with increasing distance between symmetrical coils. Since most WPT applications consist of asymmetrical coil sizes, it is necessary to optimize Tx and Rx coil geometries to achieve maximum Q without diminishing k .

3.2.2 Validation of Coil Parameters

The expressions for L and R shown in (3.20) and (3.23) respectively are validated against measured results for eight different test coils shown in Fig. 3.13. Each test coil has the same $D_o = 8\text{cm}$, $D_i = 0$, $w = 1.024\text{mm}$, $f = 13.56\text{MHz}$; however, the pitch and number of turns for each coil are different and are listed in Fig. 3.13. The measured values of L and R for each test coil were evaluated by extracting S_{11} from an HP8753 vector network analyzer (VNA) and using a best-fit analysis to match L , R , and C to the equivalent circuit model for a series resonant coil with input impedance Z_{IN} shown in (3.26).

$$Z_{IN} = R + j\omega L + \frac{1}{j\omega C}, \quad S_{11} = \frac{Z_{IN} - 50\Omega}{Z_{IN} + 50\Omega} \quad (3.26)$$

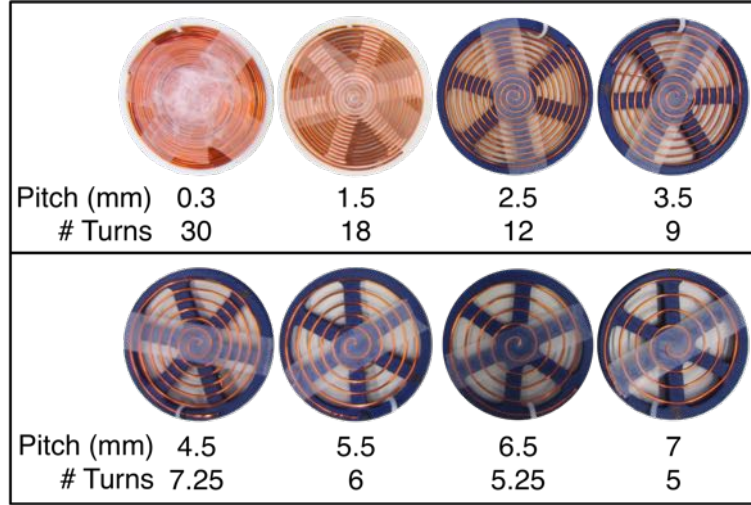


Figure 3.13: Coils used for validation of parasitic component calculations.

The results for the measured and calculated L and R are plotted in Fig. 3.14. The inductance calculation is accurate for both tightly and loosely wound coils. For resistance, the proximity effect model using the AC resistance calculation from [23] is more accurate for tightly wound coils, but over-approximates R for larger p . The skin

effect model from (3.23) underestimates R for tightly wound coils, but is more accurate for $p > 2.5\text{mm}$ (approximately 2.5 times greater than the wire diameter of 1.024mm). Since R increases exponentially for tightly wound coils, and high Q requires low R , it is preferable to design coils that are not tightly wound for WPT applications. Therefore (3.23) will be used for the remainder of this analysis to calculate R with the constraint that $p > 2.5w$.

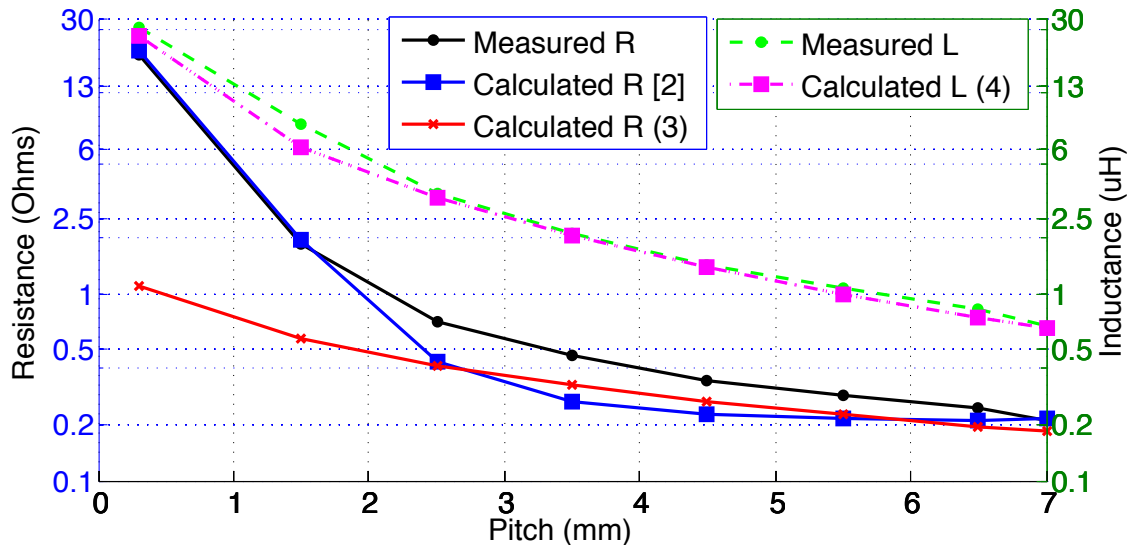


Figure 3.14: R and L vs. p for the eight test coils from Fig. 3.13. The first R calculation result accounts for both proximity effects and skin effects [23] while the second calculation accounts for only the skin effect and DC resistance (3.23).

3.2.3 Coil Design Procedure

The analysis for Q and k will be used to show how the Tx and Rx coil geometries impact both range and efficiency for a 4-element WPT system. Although a design example using a 4-element WPT system is outlined here, the following analysis can be generalized to a 2, 3, or n -element WPT system by using the transfer function for the desired WPT system. The results from this design example are plotted in Fig. 3.15. However, the following procedure can be used for any Tx and Rx coil size,

provided the expressions for L , R , and C of the coils are accurate for the specified range of coil geometries.

1. Define the maximize allowable Rx coil size. For the example used in this procedure, $D_{O,RX,max} = 5.8\text{cm}$. 5.8cm provides sufficient tolerance for the coil mold, so that the entire Rx coil is within the permissible size of 7cm for implantation.
2. Define a range of allowable Tx coil sizes. For the FREE-D system, the maximum allowable Tx coil size is 15cm, but for the purpose of evaluating a wide range of coil sizes, a range of $1\text{cm} < D_{O,TX} < 7\text{cm}$ will be used for this analysis.
3. Maximize Q using (3.24) for each Tx coil size and the Rx coil by exhaustively evaluating all possible combinations of p and N for which the expressions have known accuracy. The constraints include $p > 2.5w$, $D_i > 0$, and $f = 13.56\text{MHz}$. In this example, $Q_{RX,max} = 520.9$ for the $D_{O,RX} = 5.8\text{cm}$ Rx coil using $N = 6$, $p = 2.56\text{mm}$, and 18AWG copper wire ($w = 1.03\text{mm}$).
4. Using the geometries for each optimized Tx coil, calculate the achievable k_{23} for a range of distances targeted by the application. For the FREE-D system, $1 < d < 7\text{cm}$ is the targeted range. However for this analysis, a wider range up to 14cm will be used. The goal is to maximize efficiency across this entire distance range.
5. Using k_{23} , $Q_{RX,max}$ and Q for each Tx coil, select Q of the Tx and Rx loops, k_{12} and k_{34} for an optimal figure of merit as outlined for symmetrical coils in Section IV from [67] or for asymmetrical coils from [45].
6. Using the calculated optimal values for each coil in the 4-element WPT system, evaluate the transfer function shown in (3.14).

7. Select the optimal Tx coil size for maximum efficiency at the longest or desirable range for the intended application. In this example, the results of this analysis are plotted in Fig. 3.15.

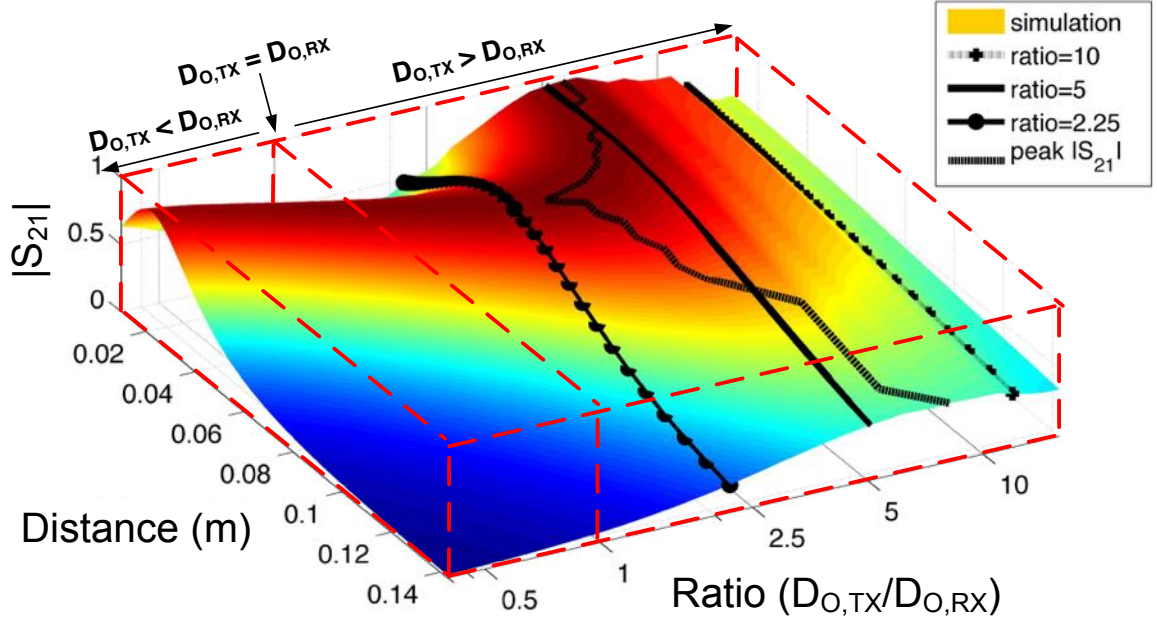


Figure 3.15: $|S_{21}|$ as a function of Tx to Rx coil size ratio ($\frac{D_{O,TX}}{D_{O,RX}}$) and distance between the coils.

Fig. 3.15 is divided into two regions where the Tx coil is larger and smaller than the Rx coil. In both regions, $|S_{21}| > 0.9$ can be achieved. The two ridges converge as the distance increases and this high $|S_{21}|$ cannot be achieved as the distance between the coils increases beyond approximately 11mm, which corresponds to a distance of approximately $2D_{O,RX}$. A valley exists at close distances for coils with a ratio close to one because k is high for similarly sized coils, and in this overcoupled region, high $|S_{21}|$ occurs at a different frequency than resonant frequency of the coils [68]. Although this region has poor $|S_{21}|$ at a single operating frequency, implementing frequency tracking or adaptive impedance matching would enable higher $|S_{21}|$ in this region. $|S_{21}|$ drops below 0.9 for coil size ratios larger than 6.2 because k is weak between a large Tx coil and a much smaller Rx coil. However, the peak $|S_{21}|$ occurs at larger

ratios as the distance increases beyond 6cm because $|S_{21}|$ decays slower for larger coil size ratios.

3.2.4 Coil Design Procedure Example

To verify that these simulated results are accurate for the full range of Tx coil sizes, three Tx coils and the 5.8cm Rx coil (Fig. 3.16) were constructed. Three Tx coils were designed according to the recommended geometries for highest Q given by the optimization procedure. The three Tx to Rx coil size ratios are: 10:1 ($D_{O, TX1}=56.5\text{cm}$, $N_1=3.4$, $p_1=7\text{mm}$, $w_1 = 2.05\text{mm}$), 5:1 ($D_{O, TX2}=28.5\text{cm}$, $N_2=8$, $p_2=7.5\text{mm}$, $w_2 = 1.63\text{mm}$), and 2.25:1 ($D_{O, TX3}=13\text{cm}$, $N_3=6$, $p_3=6\text{mm}$, $w_3 = 1.63\text{mm}$). The VNA was used to extract several S_{21} datasets for a range of distances between the coils. Fig. 3.17 shows these experimental results plotted against the simulated results for the three coil size ratios indicated on Fig. 3.15.

The simulated and measured results match very closely for all sets of coils. The 10:1 ratio pair suffers from low coupling between the large Tx coil and small Rx coil, and therefore cannot achieve high $|S_{21}|^2$. However the coupling diminishes slower and the 10:1 ratio achieves its peak $|S_{21}|^2$ beyond 14cm. The 2.25:1 ratio suffers from over-coupling, and would require frequency tuning to accommodate for the reduced $|S_{21}|^2$ at close distances. The 5:1 ratio achieves high $|S_{21}|^2$ at close ranges, and a slower decrease in $|S_{21}|^2$ than the 2.25:1 ratio. The 5:1 ratio also follows the peak $|S_{21}|$ curve indicated on Fig. 3.15 most closely. For the proposed $0 < d < 14\text{cm}$ range used in this example, the 5:1 ratio using TX_2 is optimal. However, for the FREE-D system, the maximum Tx coil size is limited to 15cm. Therefore, within the desired operating range from $0 < d < 7\text{cm}$, adaptive tuning must be used to avoid the valley that occurs in the overcoupled region while operating at a single frequency.

The final coils for the FREE-D system that were designed using this technique are shown in Figures 3.10 and 3.11. The coil geometries and parasitic parameters for the Tx and Rx coils are shown in Table 3.1.

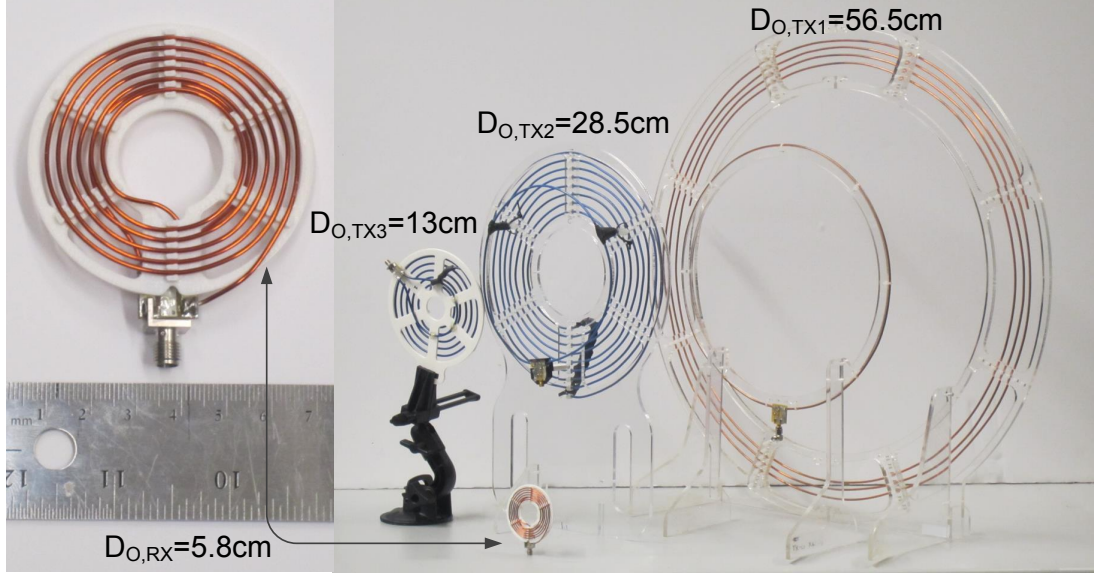


Figure 3.16: Sets of Tx coils and Rx coils used for experimental measurements.

3.2.5 Coil Frequency Selection

One of the most important design decisions for a WPT system is the operating frequency. Several studies have recently argued which frequency is best to use for WPT [61, 11, 53] while considering SAR, field strength and conduction current limitations as well as power efficiency. However there is a limited number of commercially available frequency bands in which wireless power transfer is permitted. The frequency bands that are most applicable to WPT for high power implanted medical devices are those dedicated bands for ISM equipment. From Table 2.1, these include the frequency bands centered around 6.78, 13.56, 27.12, 40.68, 433.92MHz and so on. Another available frequency range is below 300kHz. Although there are no explicit ISM bands in this frequency range, the SAR and MPE limits do not cover wireless power systems below 300kHz, and compliance needs to be proven in accordance with [20].

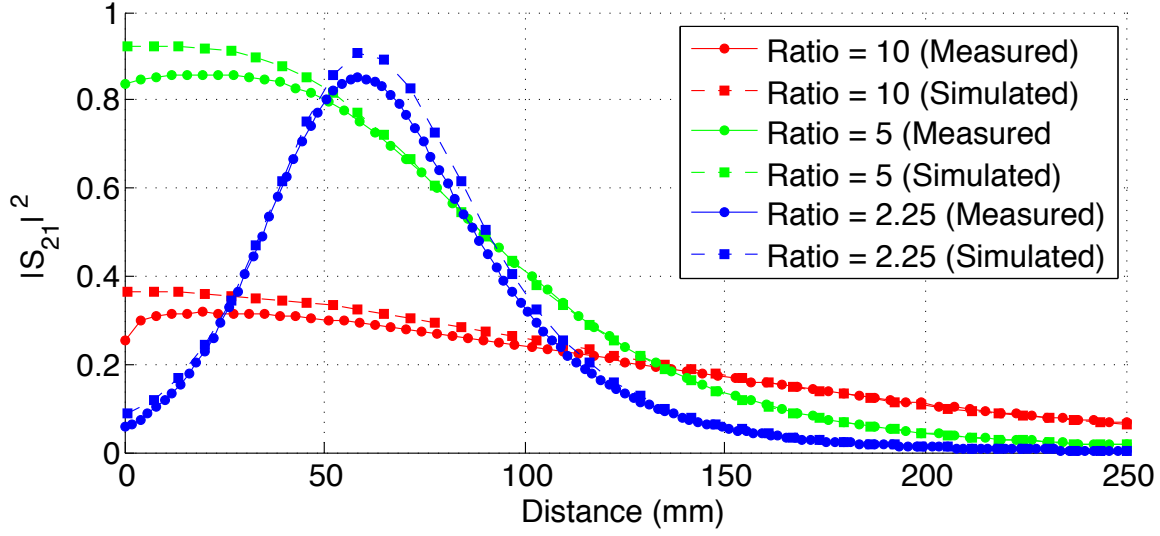


Figure 3.17: Plot of measured and simulated $|S_{21}|$ for the coils from Fig. 3.16.

Even with this limited number of available frequency bands, the question remains: which one to use? For the FREE-D system, the most relevant criteria that goes into this decision is the targeted WPT range at which high efficiency is desired, the MPE and EMI FCC regulations, and the requirement that, in the trunk of the human body, the localized temperature of tissue cannot rise above 2°C from the natural temperature. To minimize the heating of the implanted coil, it seems logical that the Rx coil should be designed purely for minimum resistance. However, minimizing resistance alone is insufficient because the coil still needs to have high Q for power transfer efficiency. For example, a small single-turn loop could have very low resistance, but the overall Q would be very poor because the inductance of a single-turn loop is low. Figure 3.18 shows a simulated plot of S_{21} calculated for a 3 coil wireless power system purely as a function of the individual coil Q factors and the coupling coefficients between adjacent coils. The expression for S_{21} has not been included in this report but is readily available in other work [41]. This plot highlights the importance of Q . The Tx coil is set to have a Q of 350 in this simulation. As the Q of the Rx coil increases, S_{21} increases. As the Q of the Rx coil drops below 50, the maximum

achievable efficiency will be less than 80%.

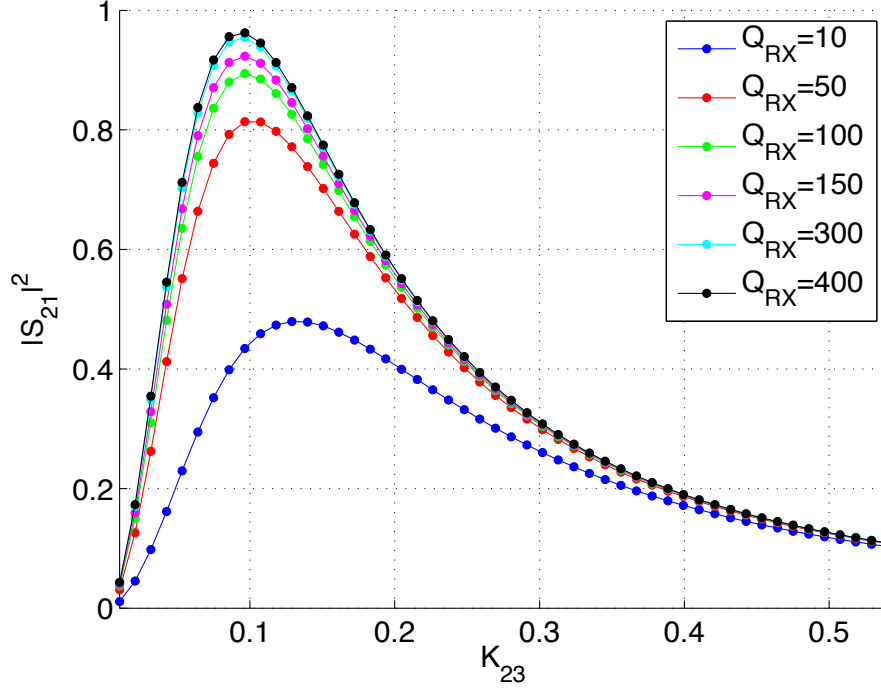


Figure 3.18: Simulated plot of S_{21} as a function of coupling between the Tx and Rx coils for various Rx coil Q values.

Therefore, the goal should be to pick a frequency that can achieve the highest Q coil for the allowable coil size. Recall that the maximum Rx coil size is 7cm for FREE-D. For lower frequencies, given the same inductance, the Q factor decreases because a larger tuning capacitor is required to make the coil resonant at the targeted frequency. If inductance increases for the same size coil, the coil resistance also goes up, thus decreasing Q . Ultimately, reducing the resonant frequency of the coil reduces the coil's Q for the same size coil. For higher frequencies, the MPE and EMI regulations are more strict. The MPE or EMI regulations on radiated field strength would be reached at lower power levels for higher operating frequencies. Therefore, 13.56MHz has been selected as the operating frequency for the FREE-D system.

3.2.6 Number of Coils in the WPT System

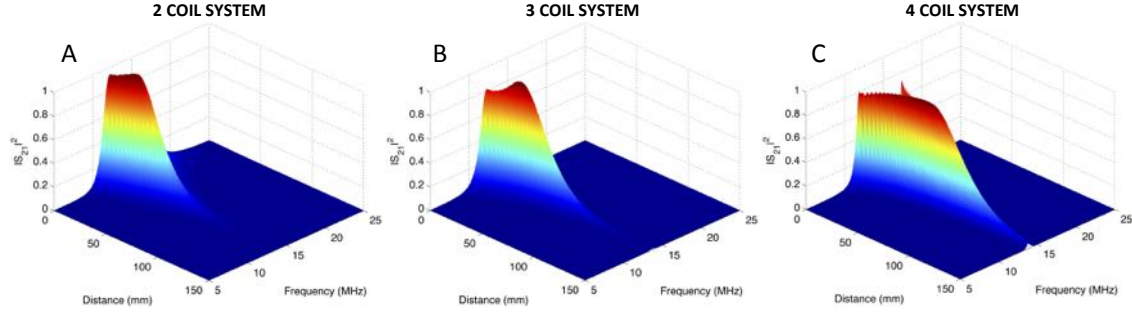


Figure 3.19: Measured results for $|S_{21}|^2$ versus distance and frequency for 2, 3, and 4 coil wireless power systems.

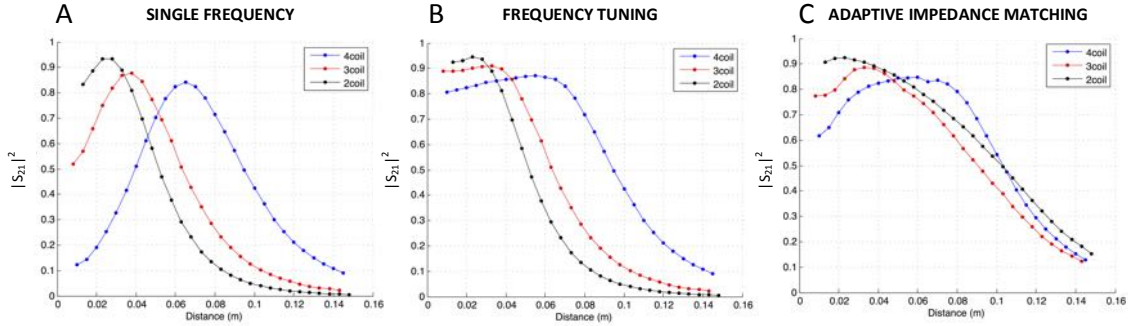


Figure 3.20: Comparisons of efficiency ($|S_{21}|^2$) versus distance for 2, 3, and 4 coil wireless power systems using single frequency operation at 13.56MHz (A), frequency tracking (B) and adaptive impedance matching (C).

The next coil design decision involves whether to use a 2, 3, or 4 coil wireless power system. To demonstrate some of the considerations for this decision, three wireless power systems (2, 3, 4 coil systems) were fabricated. Each system uses the same Tx and Rx high-Q coils shown in Figure 3.10. In this analysis, the loop-coil coupling coefficients have been configured for the optimal figure of merit defined by Sample et al. in [67] for both 3 and 4 coil systems.

A VNA was used to extract the S-parameters for a range of distances between the coils for each of these systems. Figure 3.20 shows $|S_{21}|^2$ plotted against the distance between the coils and frequency. As expected from the theoretical comparisons of 2, 3, and 4 coil WPT systems in Figure 3.9, Figure 3.19 shows that the 4 coil system has the largest overcoupled region, hence the highest efficiency at the longest range. However, the 2 coil system has the highest peak $|S_{21}|^2$.

Two-dimensional perspectives of these surface plots are shown in Figure 3.20A-B. Figure 3.20A shows $|S_{21}|^2$ against distance for single frequency operation at 13.56MHz. Figure 3.20B shows $|S_{21}|^2$ when operating with frequency tracking, which selects the peak S_{21} at each distance increment. The three curves in Figures 3.20A and 3.20B can be extrapolated directly from Figure 3.19.

Figure 3.20C uses an optimized adaptive impedance matching network to increase the range at which $S_{21}|^2$ can be achieved. These algorithms will be discussed thoroughly in Chapter 4. At a single operating frequency, the efficiency peaks for each system occur at various distances, and the 4 coil system is capable of achieving highest efficiency at the greatest distance. Figure 3.20B shows the efficiency of each system using frequency tracking, which improves the efficiency at short distances because of the frequency splitting effect. Figure 3.20C shows the efficiency using an adaptive impedance matching network at both the transmit and receive sides of the system. With adaptive impedance matching, the 2 coil system achieves the greatest improvement in range from the single frequency and frequency tracking plots. The 2 coil system also maintains the highest efficiency at close distances compared to the 3 and 4 coil systems. However, without proper adaptive impedance matching, the 2 coil system performs the worst at larger distances.

The important observation here is that, for the desired operating range of 1 – 7cm for the FREE-D system, any of the 2, 3, or 4 coil systems will have sufficient efficiency only if adaptive impedance matching is used. If adaptive impedance matching is not used, only the 4 coil system with frequency tracking will have sufficient efficiency in

the 1 – 7cm distance range.

3.2.7 Thermal Considerations

Until this point, all of the coil design experiments have either been performed in simulation or at a low power level using the VNA. Now, the various coil configurations will be compared at full power. The goal will be to determine which coil configuration (loop and coil on the receiver or only a coil on the receiver) can maintain the lowest operating temperature. A power amplifier provides sufficient power to a Tx coil consisting of a loop and coil. An Rx coil was placed 5cm away from the Tx coil (a distance where the efficiency of the 3 and 4 coil wireless power systems is nearly equivalent). If necessary, the transmit power level was adjusted to deliver a constant 8W to a 50Ω RF power meter connected to the Rx coil. Five Rx coils (without loops) were tested, each with the same outer diameter but different number of turns. Then this test was repeated for five identical Rx coils with a loop directly connected to the RF power meter. The Fluke 80BK-A type K thermocouple was placed on the most central turn of the Rx coil and recorded the temperature for 10 minutes, enough time for the temperature to establish a common trend in temperature fluctuation. The Rx coils with and without the loops are shown in Figures 3.21 and 3.22 respectively and the corresponding temperature plots are shown in Figures 3.23 and 3.24.

The results indicate that the configurations involving the Rx loop and coil undergo significant temperature changes, upwards of 40°C from the initial temperature of the coil! At first this seems counter-intuitive because when the high Q coil is isolated from the 50Ω load, the voltage across the coil should be higher and the current flowing in the coil should be lower than the configuration when the coil is directly loaded by the 50Ω load. With higher voltage, I^2R losses should be reduced. So what's happening? There are a couple of theories. First, the high voltage may be causing the tuning capacitor to approach its maximum voltage rating and heat up, thus spreading the heat throughout the coil. Second, the 18AWG (1.024mm diameter) single-strand

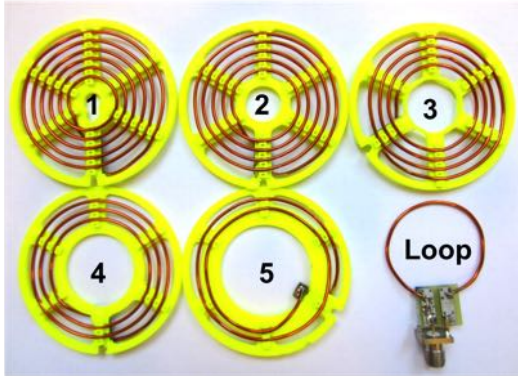


Figure 3.21: Rx test coils with Rx loop for temperature evaluation.

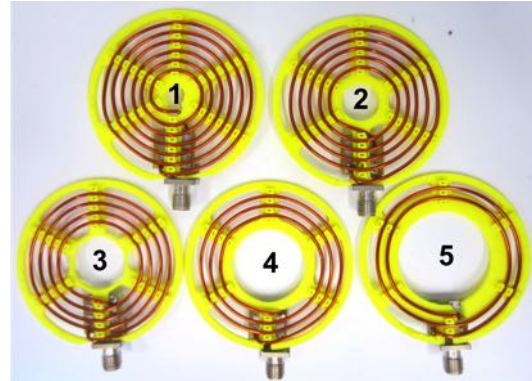


Figure 3.22: Rx test coils for temperature evaluation.

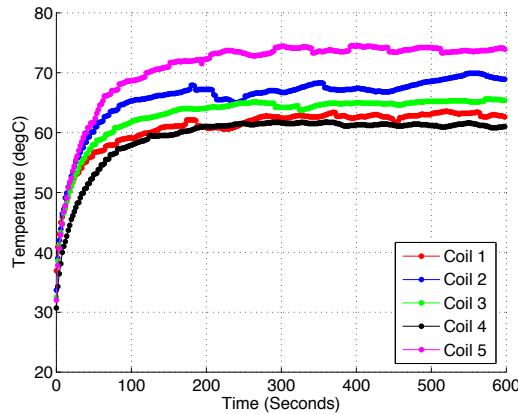


Figure 3.23: Temperature plots for Rx loop and coil configurations.

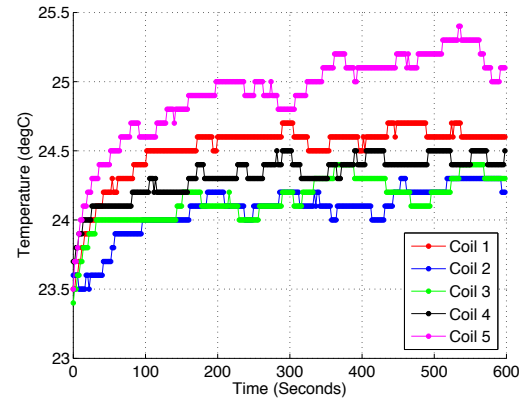


Figure 3.24: Temperature plots for Rx coil-only configurations.

copper wire that is being used for these coils is subject to the skin effect, where current flows on the outer surface of the conductor, shrinking the effective size of the conductor and increasing its effective resistance. But one would expect both 3 and 4 coil systems would be subject to this, so the skin effect alone couldn't possibly be the only reason for this drastic temperature difference. The higher voltages on the coil in the 4 coil system could also produce higher dielectric losses, where the high electric field around the conductor itself induces voltage into the polymer-film insulation of

the magnet wire. It is likely that a combination of all of these factors contributes to the heating of the 4 coil system. These problems could potentially be resolved for a larger coil or for a different conductor other than copper magnet wire. However with the 7cm size limitation, the FREE-D system has used a 2 or 3 coil system to minimize heating of the Rx coil at higher power transfer levels.

3.2.8 *Coil Tuning*

The final consideration for the FREE-D system coils is tuning. Once the coil is placed inside the body, the conductive nature of human tissues can de-tune the coil and contribute to additional losses and inefficiencies of the WPT because some energy will be absorbed by the body. Although the FREE-D system uses near-field magnetically coupled resonators, where the energy is transferred primarily by way of the magnetic field component, the electric field is still present and can be strong close to the coil. If the coil is close to or in contact with the conductive tissue, eddy currents can be induced into the tissue, increasing loss and tissue heating. Therefore, the coil has been encapsulated in polydimethylsiloxane (PDMS). Figure 3.25 shows the imaginary part of S_{11} extracted from the VNA for the Rx coil in air (Figure 3.26), encapsulated in PDMS (Figure 3.27), and then placed in a large slab of ham with a slot cutout for the coil (Figure 3.28).

In air, the Rx coil was initially tuned for 13.56MHz, with a measured Q of 322. The mostly non-conductive PDMS has a slight impact on the coil tuning, shifting the resonant frequency to 13.44MHz with a measured Q of 273. However once the coil is placed in ham, even with the PDMS, the resonant frequency shifts to 13.05MHz and the Q deteriorates to 49.6. From Figure 3.18, this reduction of Q will reduce WPT efficiency when the coil is implanted inside the body. The efficiency will be reduced even further if the coil de-tuning is not accounted for. For these reasons, there is an incentive to make the PDMS encapsulation of the coil as large as possible. However, the medically imposed design requirements restricts the overall thickness of the coil

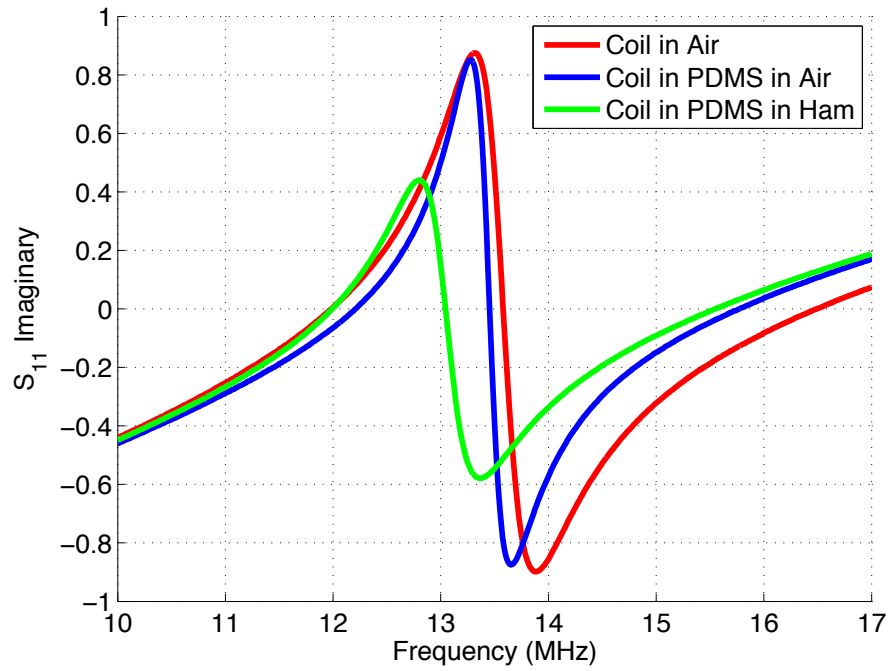


Figure 3.25: Imaginary part of S_{11} for various coil environments.



Figure 3.26: FREE-D coil in air



Figure 3.27: Coil in PDMS

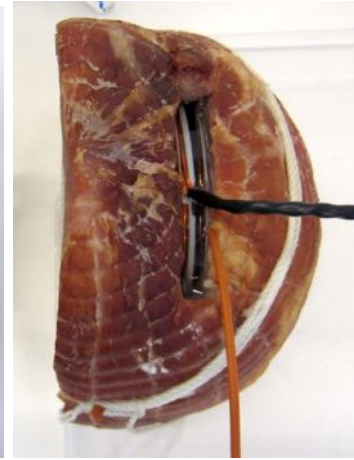


Figure 3.28: FREE-D coil in ham

to 1cm for practical implantation capabilities.

All of these design considerations and several iterations of the FREE-D system

coils have resulted in using a 3 coil system tuned for 13.56MHz for in-vivo experimental trials of the FREE-D system. Refer to Section 5.2 for details on the implanted system.

3.3 Transmitter Design

The wireless power transmitter consists of TMS320 digital signal processing (DSP) chip for control of the system, a highly efficient class-E power amplifier (PA) capable of outputting up to 40W, an adaptive impedance matching network, a directional coupler, a transmit coil and a 433MHz antenna for communication. This system has gone through several design iterations, but only the most recent design will be discussed in this report.

Figure 3.29 shows a block diagram of the transmitter and Figure 3.30 shows a photograph of the transmitter printed circuit board (PCB). The transmitter board was fabricated on a 4-layer PCB and measures 15.2x11.4x4.5cm and weighs 273.2grams (0.60lbs), which includes the weight of the heat sink for the PA MOSFET. For reference, Thoratec System Controller that is used alongside the HeartMate II LVAD weighs 698.1g. The power prioritizer selects between one of three available inputs to provide power to the transmitter: wall power, primary battery or secondary battery power. Once the board is powered up, the TMS320 digital signal processor (DSP) turns on and initializes all of the additional circuitry on the transmitter. The DSP is the primary controller for the entire wireless power system. All measurements, commands, enable and disable functions are generated by or terminated by the DSP. The DSP runs off a 5V supply with a relatively static power consumption of 0.72W.

The AD9850 direct digital synthesizer (DDS) generates the RF signal that drives the power amplifier (PA). The amplitude, frequency and phase of the DDS can all be programmed over an I2C interface with the TMS320. The frequency is generated by dividing down a 150MHz clock generated from an external crystal oscillator. With a 32-bit frequency tuning word, the AD9850 frequency can be accurately programmed down to 0.035Hz.

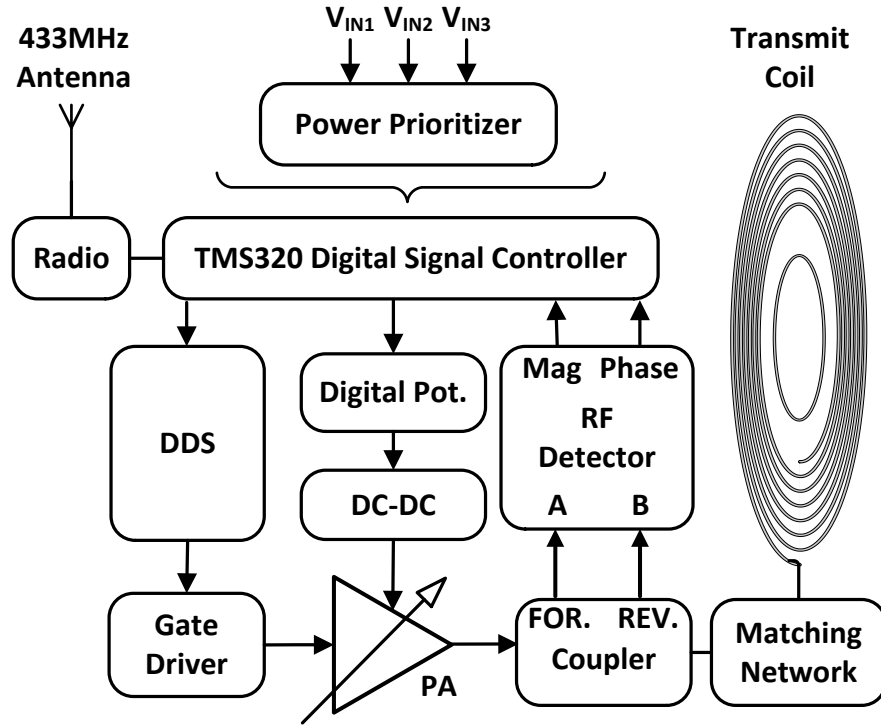


Figure 3.29: Block diagram of FREE-D system transmitter.

The PA is driven by a UCC27517 gate driver that runs off a 10V supply and consumes approximately 80mA of current when operating at 13.56MHz. The gate driver drives an IRF510 MOSFET in a single-ended class-E amplifier. Class-E amplifiers are commonly used for applications that require high efficiency, since the theoretical efficiency of class-E amplifiers is 100%. The downside of class-E amplifiers is that they are typically designed to be most efficient at a single operating point in terms of power level, load impedance and operating frequency. In a wireless power system where the load impedance changes as a function of load power and distance between the coils, a class-E amplifier may rarely operate at the fixed operating point that it was initially designed for. Therefore, the class-E amplifier used in this system has a programmable supply voltage that can dynamically adjust the output power level of

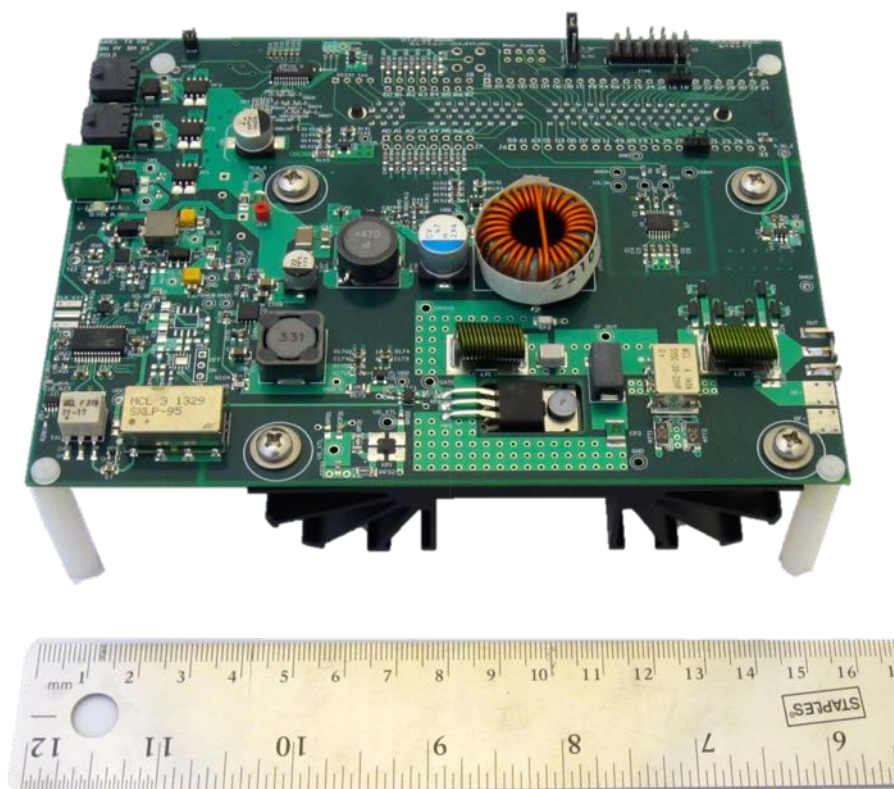


Figure 3.30: FREE-D system transmitter.

the amplifier to accommodate a wide range of load conditions. The programmable supply voltage is realized by using a $250k\Omega$ digital potentiometer to control the feedback voltage of a buck-boost DC-DC converter. There is an additional LC filter added onto the output of the PA for filtering out unwanted harmonics. Figure 3.31 shows a schematic diagram of the Class E PA.

3.4 Receiver Design

The wireless power receiver has also undergone several design iterations. Considering the receiver will be implanted alongside the Rx coil, the design goals have been to minimize the size, weight and operating temperature of the receiver PCB while

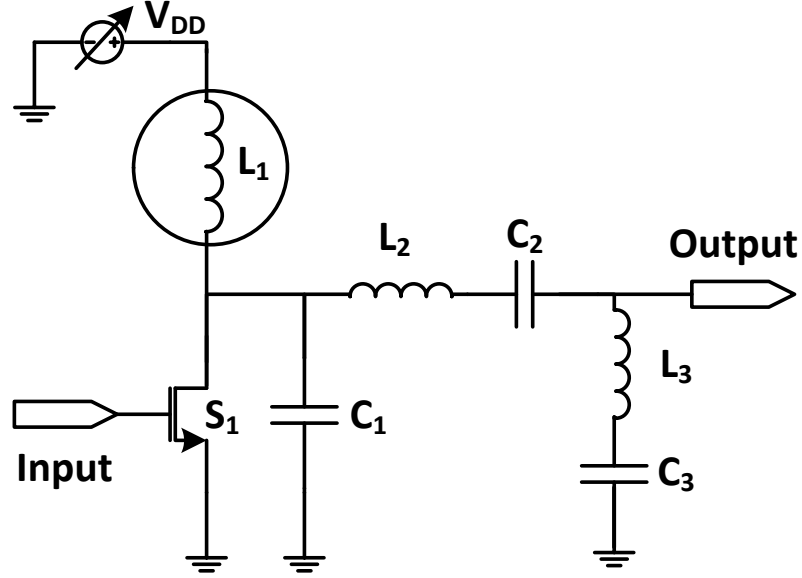


Figure 3.31: Schematic diagram of class E RF amplifier.

maximizing the efficiency and reliability of the receiver. The primary components on the FREE-D receiver include a full-wave bridge rectifier to convert the RF energy into a DC voltage, a two-cell backup battery and a Texas Instruments BQ24113A battery charger to re-charge the implanted battery, two Toshiba TB6588FG 3-phase pulse-width-modulation (PWM) motor drivers to drive both stators on the HeartWare HVAD or the single stator on the Thoratec HeartMate II LVAD, and a 433MHz radio communication link. All of the hardware on the receiver is controlled by an MSP430 micro controller unit (MCU). Figure 3.32 shows a block diagram of the receiver and Figures 3.33 and 3.34 show photographs of the receiver PCB. Additional photographs of the receiver system are shown in Figures 3.35 and 3.36 alongside the HeartWare HVAD and the Thoratec HeartMate II LVAD respectively. The receiver board was fabricated on a 6-layer PCB and measures 4.2cm in diameter with a length of 5.0cm because the 433MHz chip antenna extends of the edge of the PCB. The receiver board, including the watertight enclosure weighs 21.7grams. For reference, the HeartWare

HVAD weighs 182.4grams and the Thoratec HeartMate II LVAD weighs 272.8grams.

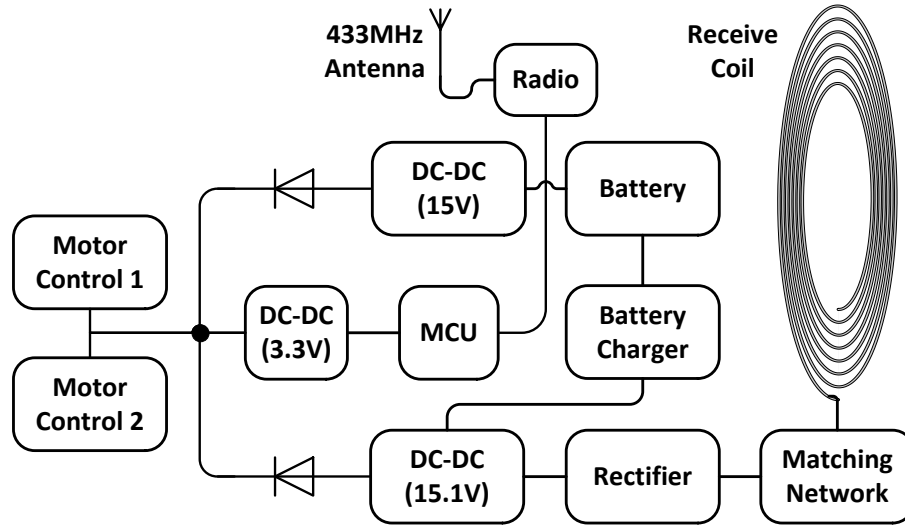


Figure 3.32: Block diagram of FREE-D system receiver.

The FREE-D implanted receiver contains all of the motor controls required to drive any type of VAD in either continuous or pulsatile mode. The receiver PCB can also be mounted directly beneath the HeartWare HVAD so that the VAD can carry heat away from the receiver board. The components that are prone to heating include the RF rectifier and the motor controllers. These components are placed on the bottom side of the PCB board, which is closest to the VAD itself. As blood circulates through the VAD, heat can be carried away from these components and help regulate the temperature of the receiver board.

The wireless power provides the primary power path and directly powers the motor controllers, thus minimizing the reliance on the backup battery and maximizing the overall battery lifetime. When wireless power is unavailable, the secondary power path on the receiver comes from the implantable backup battery. The battery is a two-cell 800mAh lithium polymer battery that measures 6.6x3.5x0.8cm and weighs 38.6 grams. This battery was selected since it is the smallest battery that can sustain a typical

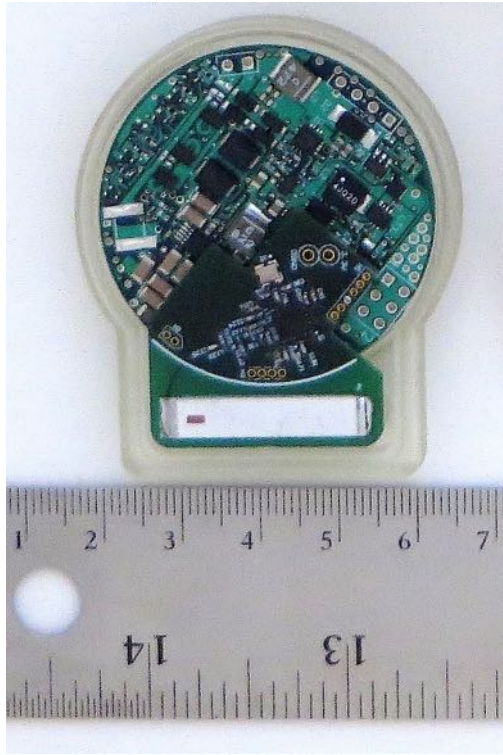


Figure 3.33: Front side of FREE-D system receiver.

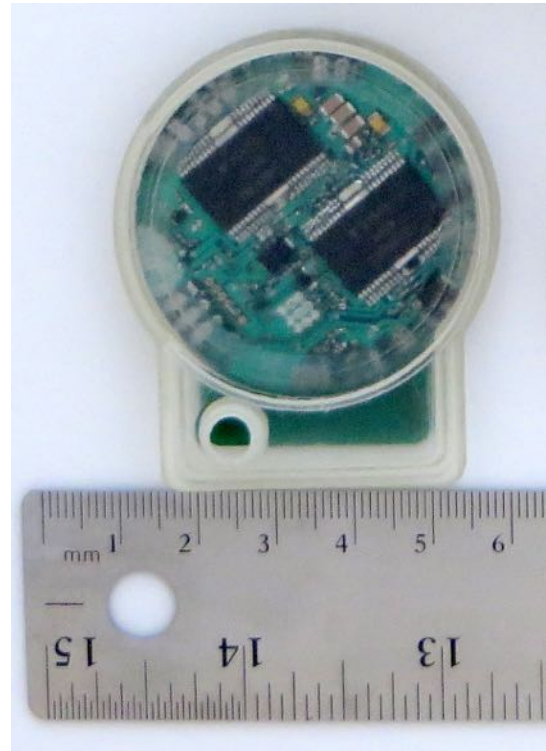


Figure 3.34: Back side of FREE-D system receiver.

VAD load power of 5W for approximately one hour. The wireless power path can simultaneously power the load and re-charge the backup battery with a programmable charge current so that it can either be trickle charged at a rate of 100mA or charged rapidly (approximately one hour from fully depleted to fully charged) at a maximum charge current of 1A. The wireless power path is prioritized over the battery power path based on the diode OR logic. If the wireless power is active and sufficiently powers the 15.1V DC-DC converter, the diode at the output of the 15V DC-DC converter will be reversed biased and current will not flow to the motor controllers from the battery power path.



Figure 3.35: FREE-D receiver and Rx coil mounted on HeartWare HVAD.

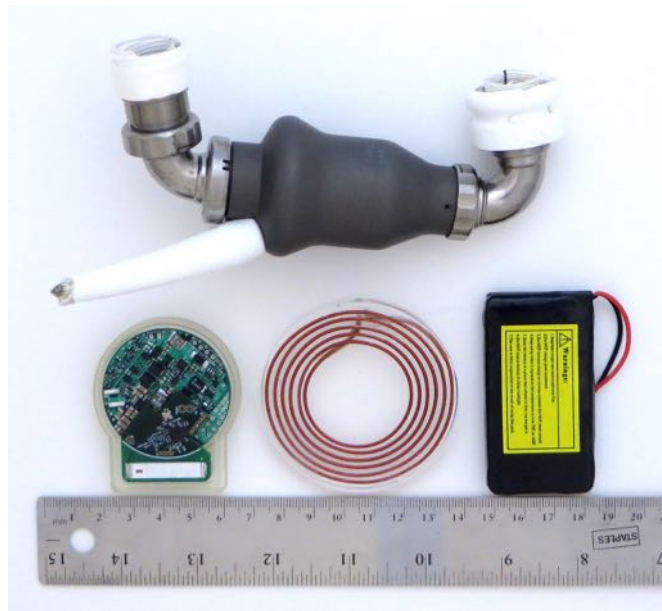


Figure 3.36: FREE-D receiver, Rx coil and two-cell backup battery next to Thoratec HeartMate II LVAD.

Chapter 4

ADAPTIVE TUNING TECHNIQUES

ADD INTRO SECTION FOR ADAPTIVE IMPEDANCE MATCHING

4.1 *Frequency Tuning*

4.1.1 *Frequency Tuning: Concept*

Coupled oscillating systems have multiple modes of operation depending on the strength of the coupling between the resonators. This can be seen in the analogous case of two masses connected by a spring [24]. In this classic physics example, the two masses form a single system which can oscillate in two modes: one of higher frequency (even mode) and one of lower frequency (odd mode) than the fundamental frequency.

In the case of the FREE-D magnetically coupled resonators, the two modes of the system can be clearly seen in figure 4.1. Figure 4.1A shows the S_{21} linear magnitude, which is a measure of the transmission power gain across the system. This S_{21} plot represents a 2-D slice of the efficiency plateau plot in figure 3.19C at a separation distance of 2.5cm. Notice that the two peaks indicate that the system is in the over-coupled regime. Figure 4.1B shows the input impedance (S_{11}) of the system on a Smith chart which is normalized to 50Ω . A black circle and a green cross approximately indicate where the peaks in the linear magnitude plot occur in the Smith chart. These plots show that the frequencies of peak $|S_{21}|$ correspond to a 50Ω input impedance to the coils.

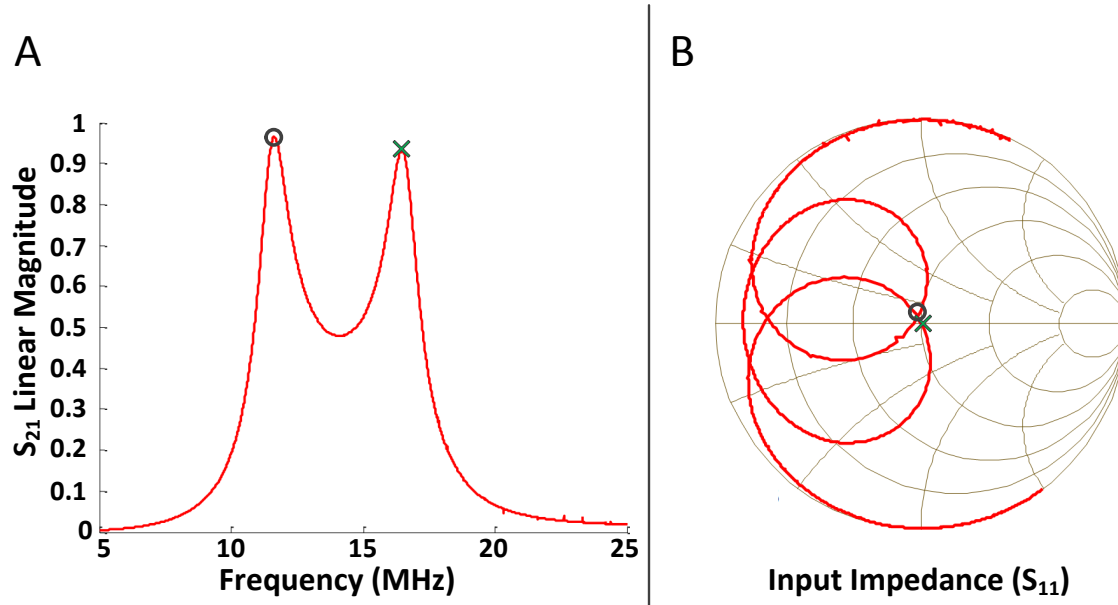


Figure 4.1: Measured scattering parameters for the FREE-D coils at a distance of 2.5 cm along the axis of coil. Panel A shows the S_{21} linear magnitude of the system. Clearly visible are the two resonant modes caused by the coupling of the two high-Q resonators. Panel B shows a Smith chart of the input impedance (S_{11}) as function of frequency. This panel shows that there are two frequencies where the input impedance to the coils is well matched to the 50Ω source and maximum power transfer can occur.

4.1.2 Frequency Tuning: Implementation

The goal of dynamic frequency tuning is to automatically adjust the transmitter frequency to provide maximum possible efficiency as the distance or alignment between the Tx and Rx coils changes. This is accomplished by introducing a bi-directional coupler between the amplifier and drive loop. The coupler allows the system to continuously measure the incident and reflected power as a function of frequency. By connecting the attenuated outputs from the bi-directional coupler to an RF gain and phase detector, the magnitude and phase of the reflection coefficient can be analyzed by the DSP at either a single frequency or for a range of frequencies.

An adaptive local-maximum detection algorithm (Figure 4.2) scans adjacent fre-

quencies and automatically sets the operating frequency to the frequency with the maximum magnitude response. If the center frequency continues to achieve the highest magnitude response, the bandwidth of the adjacent frequencies narrows until a minimum of 100Hz separation between adjacent frequencies is reached. If a frequency other than the center frequency achieves highest magnitude, the bandwidth of adjacent frequencies increases until a maximum separation of 10kHz is reached.

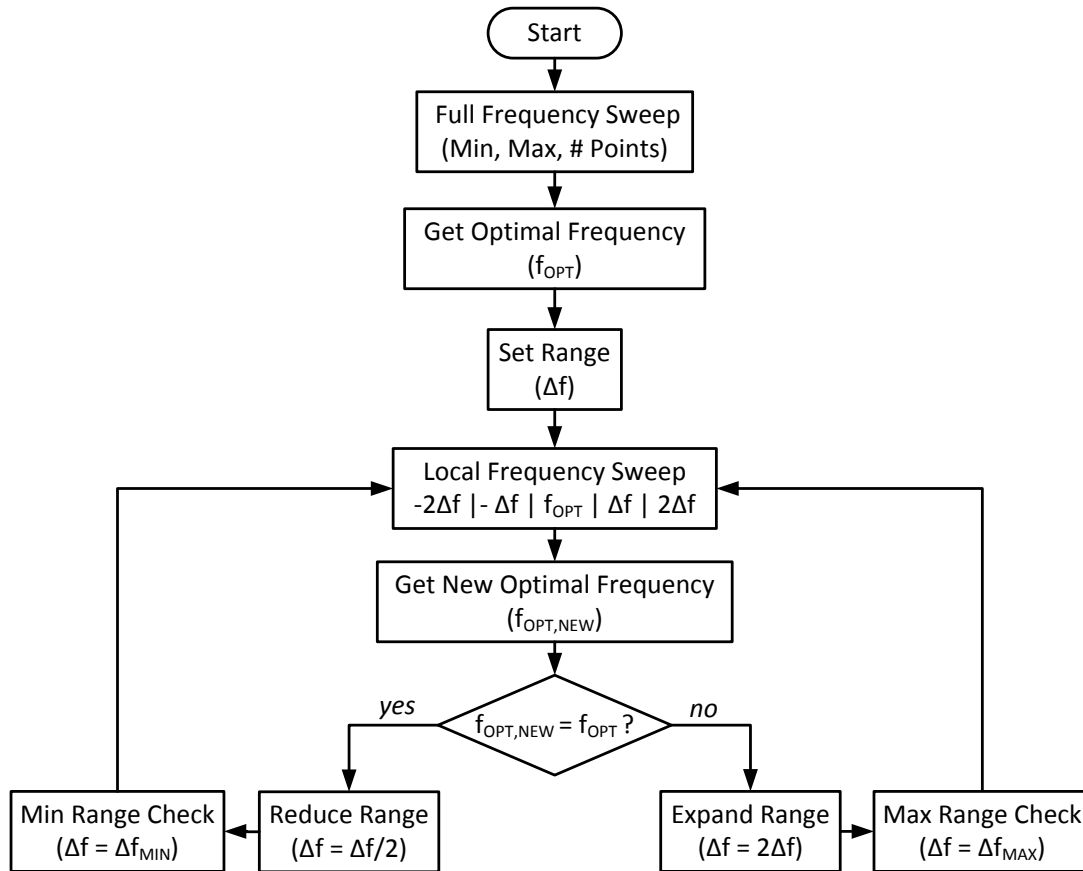


Figure 4.2: Algorithm diagram for adaptive local maximum frequency tracking.

This magnitude tracking algorithm works very well when the magnitude is easily detectable by the DSP. A maximum in magnitude is clearly defined when the

magnetically coupled resonators operate in the over-coupled region or at critical coupling. However in the under-coupled region, a maximum in the magnitude response is hardly detectable because the Rx coil has very little influence on the magnitude response when it is far away. If a clear maximum is undetected, the algorithm performs a complete frequency sweep over the user-defined frequency range to search for the maximum. In the under-coupled region, a maximum may never be clearly defined causing the algorithm to constantly re-sweep, which will never allow for maximum power delivered to the RX coil.

A phase tracking algorithm can be used as an alternative method for auto-tuning in the over-coupled, critically coupled, and under-coupled regions. A maximum in the magnitude response corresponds to a zero-degree crossing point in the phase response. By detecting the zero-degree crossing points from the phase information provided by the RF detector, a phase tracking algorithm can be used in place of magnitude tracking. The advantage of phase tracking is that a clearly defined zero-degree crossing point can be identified to ensure the accuracy of the tracking algorithm, even in the under-coupled region.

Phase tracking can be very useful for 2, 3, and 4 element wireless power systems. It is especially preferable to use phase tracking over magnitude tracking when the magnitude response is minimal. This can occur when the TX and RX coils are far away from each other, or when the TX and RX coils are asymmetrically sized because the coupling coefficient is small between the coils for both of these scenarios.

4.2 Adaptive Impedance Matching

4.2.1 Adaptive Impedance Matching: Concept

In the previous section, wide-band frequency tuning was used to achieve maximum efficiency in the over-coupled regime. In this section, dynamic impedance matching networks are used to maintain high efficiency at a single frequency in both the over-

coupled and under-coupled regimes. Recall from Section 3.2 that 2, 3 and 4-coil systems can all achieve nearly the same efficiency using properly tuned adaptive impedance matching networks.

To reiterate the importance and practicality of impedance matching, it is useful to refer back to figure 4.1. Figure 4.1B shows the input impedance of the FREE-D coils. It is clear that if the system was transmitting at a frequency that did not correspond with the desired load impedance, an impedance matching network could be designed to match the source and load impedances. However, the input impedance changes as a function of the mutual inductance between the Tx and Rx coils, and thus an dynamically reconfigurable solution is needed.

Using the parasitic parameters of the coils in Table 3.1, the input impedance for a 2, 3, and 4-coil system can be calculated using (4.1), (4.2), and (4.3) respectively. Z_{1-4} represent the equivalent series impedances of an individual coil isolated from all other coils, which can be extracted from the schematic shown in Figure 3.8. Figure 4.3 plots the input impedance as a function of the coupling coefficient between the two high Q coils for each configuration.

$$Z_{IN,2} = Z_1 + \frac{\omega^2 M_{12}^2}{Z_2} \quad (4.1)$$

$$Z_{IN,3} = Z_1 + \frac{\omega^2 M_{12}^2}{Z_2 + \frac{\omega^2 M_{23}^2}{Z_3}} \quad (4.2)$$

$$Z_{IN,4} = Z_1 + \frac{\omega^2 M_{12}^2}{Z_2 + \frac{\omega^2 M_{23}^2}{Z_3 + \frac{\omega^2 M_{34}^2}{Z_4}}} \quad (4.3)$$

To further understand this concept, compare the plots of $|S_{21}|$ as a function of frequency for the FREE-D coils both with and without impedance matching networks. Figure 4.1 shows $|S_{21}|$ without the matching networks at a separation distance of 2.5cm along the common axis of the coils. This plot shows two resonant modes (over-coupled region) and indicates that the efficiency is highest at 12MHz and is nearly

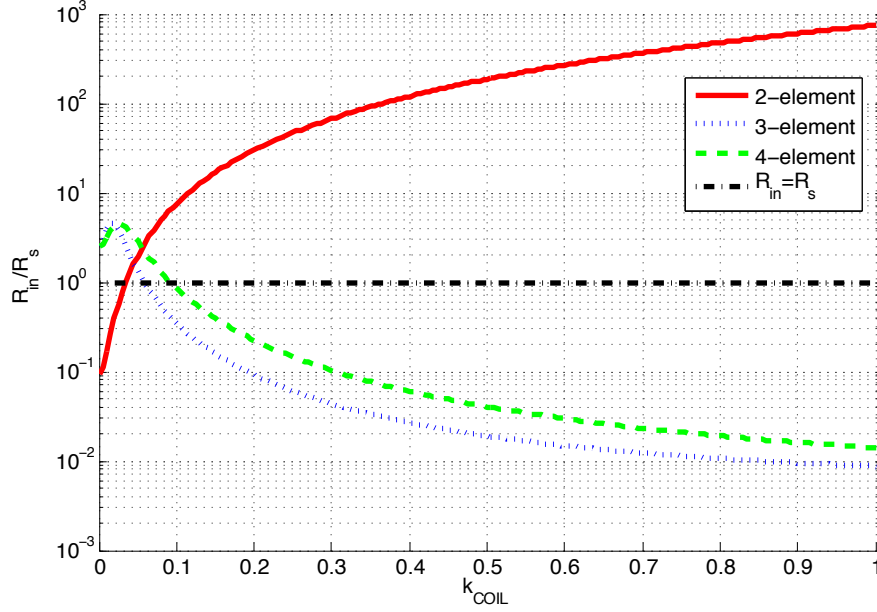


Figure 4.3: Calculated ratio ($\frac{R_{in}}{R_s}$) of coil to source impedance where $R_{in} = \text{real}(Z_{in})$ and $R_s = 50\Omega$ for the calculated input impedance of a 2, 3, and 4-coil wireless power system as a function of the coupling coefficient (k_{COIL}) between the two high-Q coils for each configuration.

at its lowest at 13.56MHz. Figure 4.4 shows $|S_{21}|$ data for the FREE-D coils with impedance matching networks at both the input to the Tx loop and at the output of the Rx coil. This plot shows that there are additional resonant peaks brought on by the matching networks, and the marker indicates that maximum efficiency S_{21} occurs at the peak corresponding to 13.56MHz. In this example S_{21} was increased from 0.47 to 0.84 using the matching networks. Although maximum efficiency can be achieved for both approaches used in Figures 4.1 and 4.4, Figure 4.4 demonstrates that only impedance matching networks can achieve maximum efficiency within the allowable 13.553-13.567MHz ISM bandwidth.

Figure 4.5 shows the advantage of impedance matching from a three-dimensional perspective. Figure 4.5A shows S_{21} for a 4 coil WPT system measured on a VNA. The v-shaped plateau indicates the overcoupled region. Figure 4.5B shows S_{21} for the

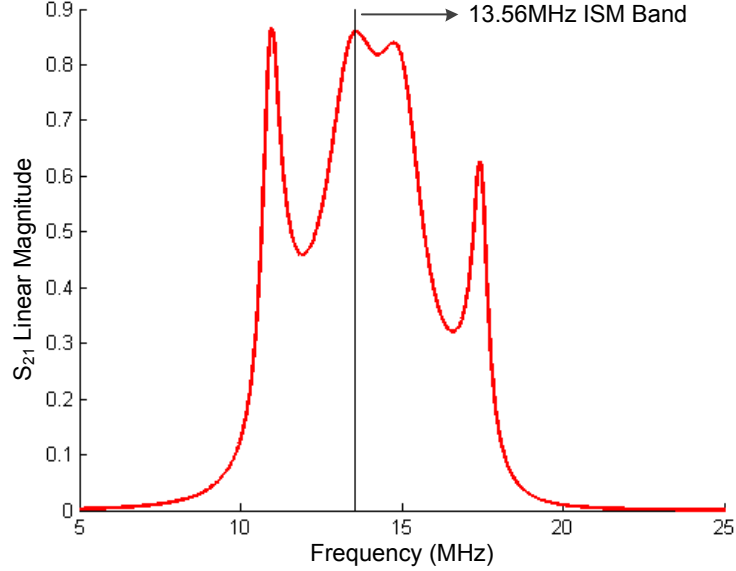


Figure 4.4: Measured $|S_{21}|$ for the FREE-D coils with π -match networks at both the Tx and Rx sides at a distance of 2.5cm along the axis of the coil. Clearly visible are the four resonant modes caused by the coupling of the two high-Q resonators and the impedance matching networks.

same set of coils with two adaptive impedance matching networks, the first placed between port 1 of the VNA and the Tx coil, and the second placed between port 2 of the VNA and the Rx coil. At each distance, the impedance matching network is tuned to maximize S_{21} . The optimally tuned matching networks create a ridge of high S_{21} at 13.56MHz. As a result, high efficiency can be achieved without needing to change the operating frequency.

In practice, this type of matching network must be capable of matching a load impedance that is either greater than, equal to or less than the source impedance. Given the inherent high Q of the magnetically coupled resonators, designing a static matching network to accommodate the wide impedance range may be difficult. Therefore a matching network with variable reactive components must be implemented.

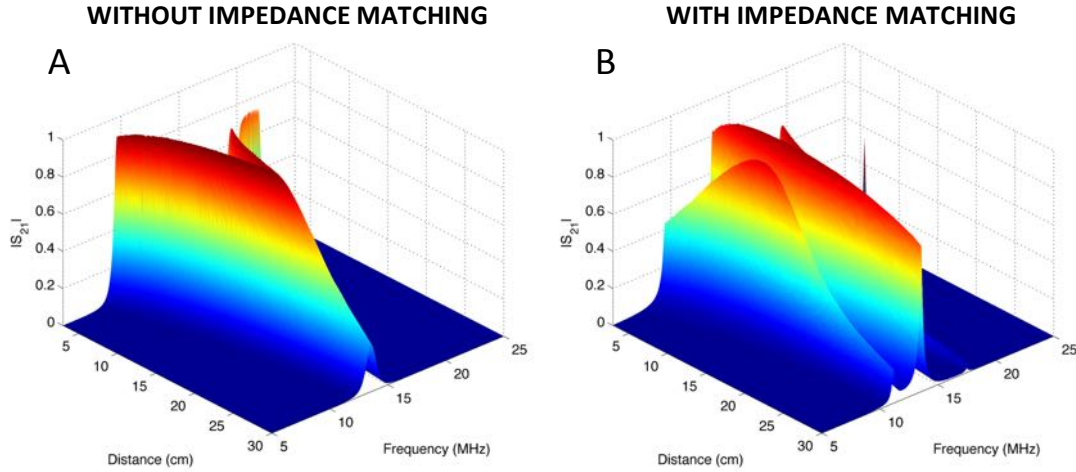


Figure 4.5: $|S_{21}|$ for a 4 coil WPT system without impedance matching (A) and with adaptive impedance matching (B) as measured by a VNA.

Variable capacitors and inductors typically have low quality factors, require mechanical tuning and are not suitable for high power RF applications. The approach taken in this report is to use electronic switches to switch different capacitances in and out of a π -match network.

The π -match network consists of a series inductor and two shunt capacitors. This matching network can be added to both the Tx and Rx coils, as shown in Figure 4.7. The π -match topology is convenient for an adaptive matching network because the fixed-value inductor is placed in series with the high power RF signal. The variable capacitors $C_{S1,2}$ and $C_{L1,2}$ can be implemented as a bank of switchable capacitors, as shown in Figure 4.6.

It has been determined that three capacitors on each side of the adaptive matching network is sufficient to match across the entire range of impedances in the FREE-D system. However, exhaustively sweeping through all possible combinations of these capacitors to determine the optimal impedance-matched configuration would be slow and result in reduced efficiency for most of that time period. Therefore it is important

to have some indication of which capacitors to switch in or out for a given distance between the coils.

A Matlab algorithm has been developed that uses unconstrained nonlinear optimization to determine the ideal capacitor values for a π -match network that will maximize S_{21} for a range of coil-coil coupling coefficients. The algorithm diagram is shown in Figure 4.7. This algorithm takes measured S-parameters for both L_π and the set of FREE-D coils and converts the S-parameters into ABCD-matrices. The ABCD representation is convenient because a series of cascaded two-port networks can be modeled by computing the product of their individual ABCD matrices to form a single lumped ABCD-matrix. The ABCD matrices for the Tx π -match network, the MCRs, and the Rx π -match network are all multiplied together. After converting the lumped ABCD-matrix back to an S-matrix, the source and load capacitor values in each π -match network are determined by selecting values that optimize $|S_{21}|$ at the desired frequency.

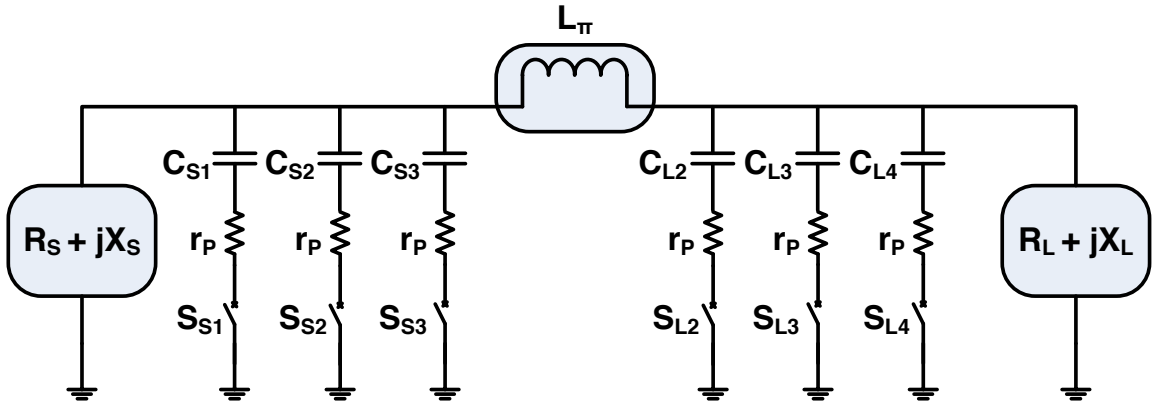


Figure 4.6: Schematic diagram of the adaptive impedance matching networks.

Figure 4.8 compares the experimental S_{21} for single frequency, frequency tracking and adaptive impedance matching measured on a network analyzer. The capacitor values used for adaptive impedance matching at each distance are shown in Figure

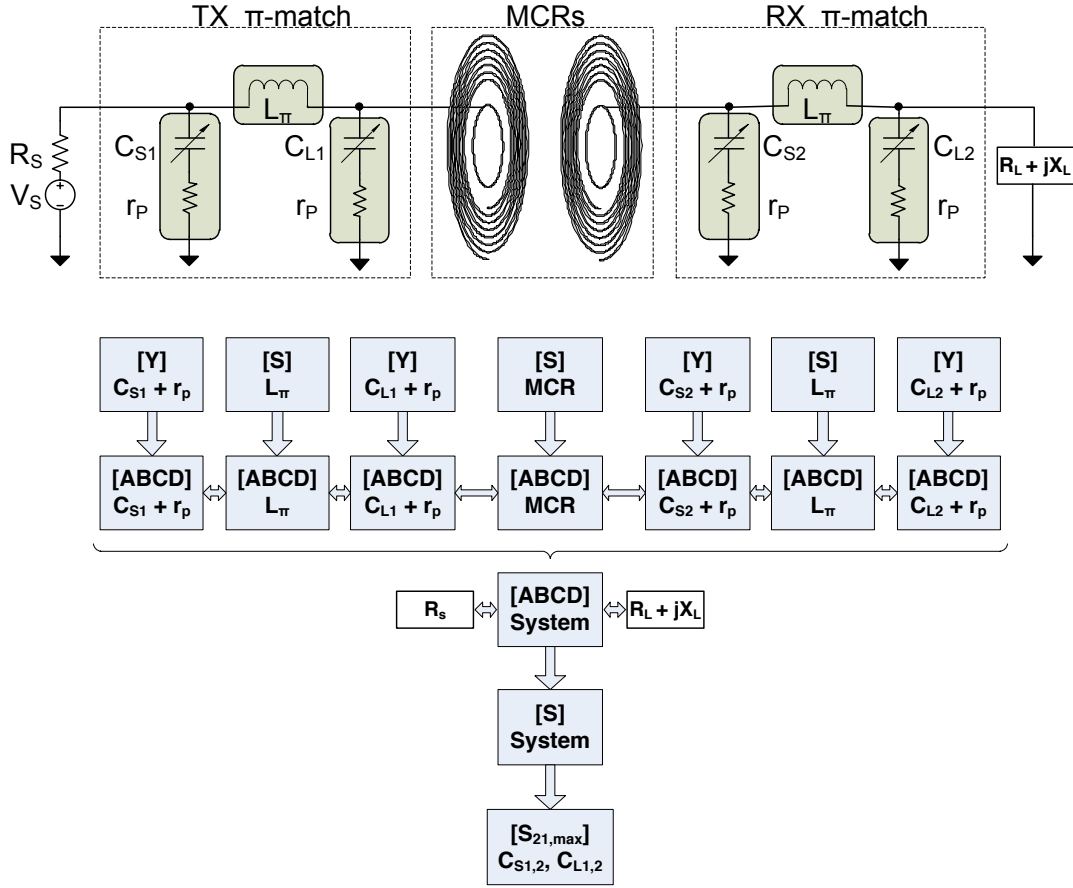


Figure 4.7: Flow chart of the process used for the adaptive impedance matching component selection algorithm to extract the shunt capacitor values in the π -match network.

4.9. A 500nH inductor was used on the Tx side matching network and a 150nH inductor was used on the Rx side matching network. With this information, the capacitor values can be pre-programmed onto the DSP. Therefore the search space for optimal capacitor settings significantly narrows and the overall time it takes for the adaptive impedance matching algorithm to reach an optimal impedance matched condition reduces.

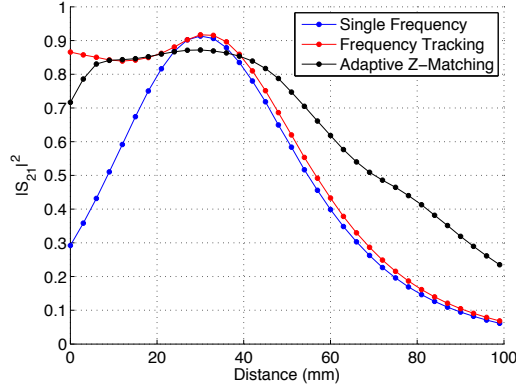


Figure 4.8: Measured S_{21} for the FREE-D coils with and without auto-tuning.

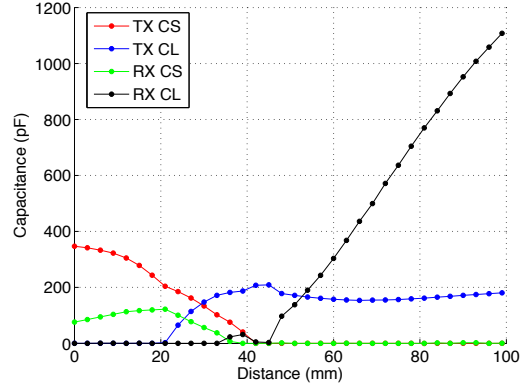


Figure 4.9: Tx and Rx source and load capacitor values required for adaptive impedance matching.

4.2.2 Adaptive Impedance Matching: Implementation

The most challenging aspect associated with building and implementing adaptive impedance matching is designing the switch to enable and disable capacitors without diminishing the added value of the matching network. Electro-mechanical relays have limited switching cycles and consume a large amount of power. RF switches are typically designed to support lower power levels and higher frequencies, and may not be suitable switches for high power adaptive impedance matching.

MOSFETs have a tradeoff between power rating, on-resistance and junction capacitance that must be carefully considered. Typically, switches with higher power rating have lower on-resistances, but larger junction (output) capacitances. This means that when the switch is off (i.e. capacitor in the matching network is disabled), the MOSFET looks like a capacitor whose value can range anywhere from a few pF to tens of thousands of pF. As seen in Figure 4.9, the capacitors in the adaptive matching network range from 1-1100pF. Therefore, without careful consideration of both the on-resistance and the output capacitance of a MOSFET, the switch may act like a voltage divider rather than a switch.

There are very few switches that meet the criteria of high power rating, low on resistance and low output capacitance. However the SI5980 dual package NMOS power MOSFET from Vishay Siliconix happens to fit the bill, with a maximum power dissipation of 7.8W, an on-resistance of 0.56Ω , and an output capacitance of 11pF [70].

Since the switch is a shunt component in the high power RF path, the oscillating positive and negative voltages must be taken into consideration. By stacking two power MOSFETs with body diodes in opposite polarity, the positive and negative swings will be blocked by at least one of the switches.

On the transmitter, the output of the power amplifier is single ended, so a capacitor (C_T) can be switched in and out with a low-side switch. However, the large power MOSFET switches (M_1 and M_2) require a gate-source voltage of at least 5V to fully turn on the switch – much more the 3.3V that the DSP can provide. An additional low-logic-level switch (S_1) connected to a pull-up resistor (R_{PU}) supplies an 8V DC voltage when CTL is low, and pulls the gates of M_1 and M_2 to 0V when CTL is high. So when CTL is low, C_T is enabled and when CTL is high, C_T is disabled in the matching network. The schematic diagram for this Tx side switch implementation is shown in Figure 4.10.

On the receiver, the switch cannot be configured in the same way as the Tx side switches because the matching network is referenced to RF+ and RF– across the full-wave rectifier. Therefore an additional capacitor must be added below M_2 , and the MOSFETs are referenced to ground at the common source. In order to present the same effective capacitance to the matching network, capacitances of $2C_T$ must be utilized in this configuration. The CTL logic remains the same as the Tx side matching network.

Similar to the frequency tracking algorithm diagram, adaptive impedance matching takes a linear optimization approach to identify the optimal capacitor settings. By identifying the optimal capacitor configurations as a function of the distance between the coils using the analysis from the previous section, these settings (or words) can

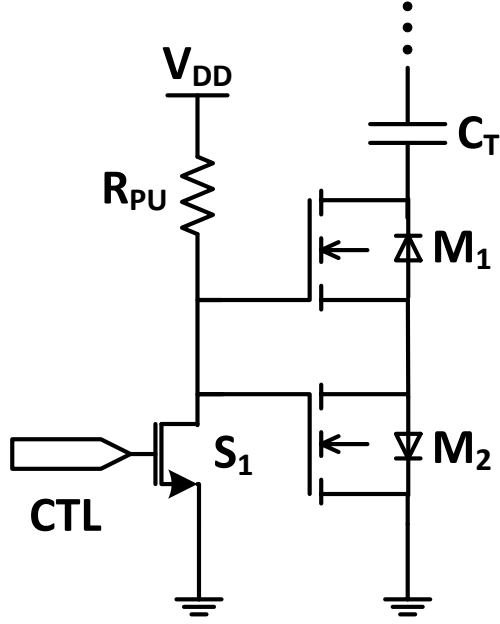


Figure 4.10: Schematic of adaptive impedance matching switch on Tx side.

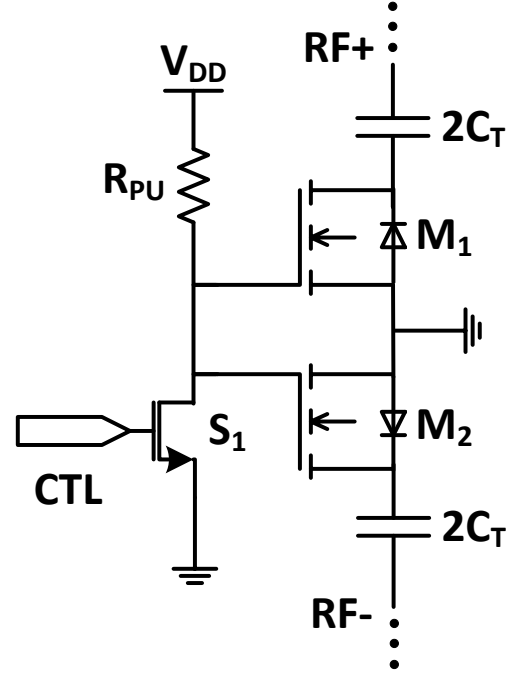


Figure 4.11: Schematic of adaptive impedance matching switch on Rx side.

be stored on the MCU. An exhaustive sweep of all settings occurs when the system initially powers up. After sampling the rectified voltage at each capacitor setting during the sweep, the optimal capacitor setting corresponds to the highest rectified voltage. After the initial capacitance is set, the algorithm will hold on that setting until the rectified voltage drops below or increases above a predefined threshold. Otherwise, the system will linearly step through the two most adjacent capacitor settings. This also enables the MCU to know if the coils have moved closer or farther away. Additionally, the time it takes to identify a new optimal capacitor setting can be minimized, since exhaustive sweeps only need to occur once.

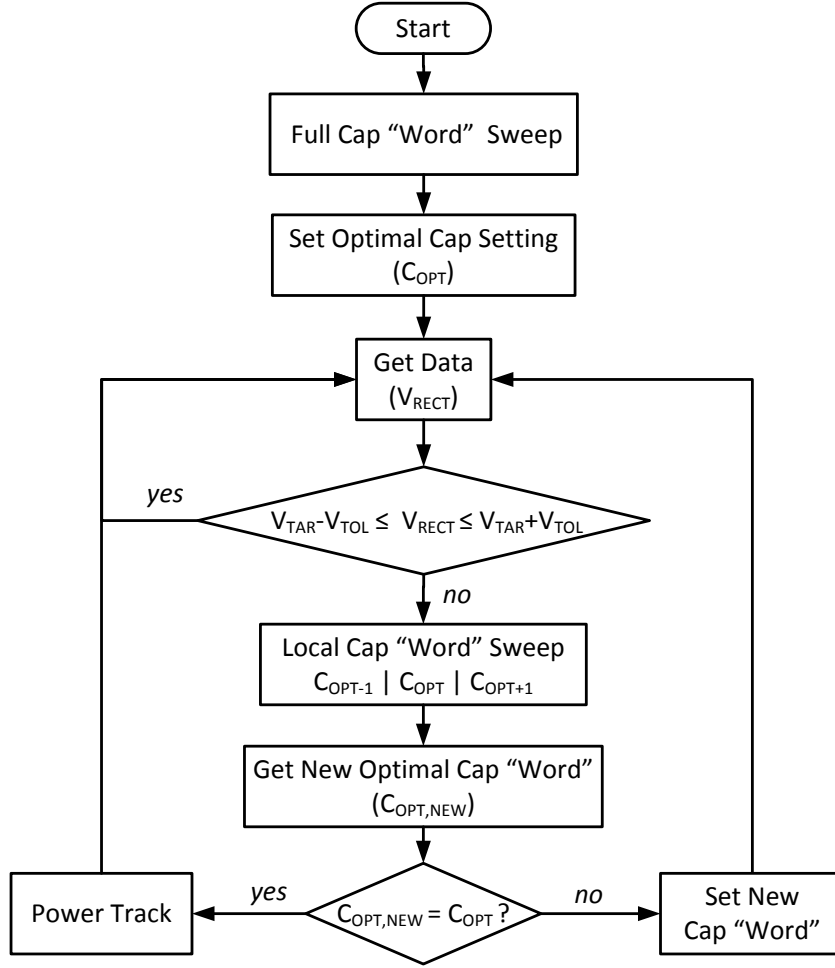


Figure 4.12: Algorithm diagram for adaptive impedance matching.

4.3 Power Tracking

4.3.1 Power Tracking: Concept

When frequency tracking and adaptive impedance matching still cannot provide enough power to the load, the transmitter may need to increase the transmit power level. Alternatively, if frequency tracking and adaptive impedance improve power delivery such that the transmitter delivers excess power to the receiver, the transmit

power level should be reduced to optimize system efficiency. Power tracking changes the transmit power level to maintain high efficiency and a desired voltage level at the receiver.

Power tracking also plays an important role in optimizing the stability of the wireless power delivery. The receiver schematic shown in Figure 3.32 shows that the rectified voltage connects directly to the input of a buck converter. For maximum efficiency, the rectified voltage should be maintained at the minimum input voltage required to turn on the buck converter. However, if the load power of the VAD changes slightly or the distance between the coils varies, the narrow headroom will cause the buck converter to dropout and the system will revert to battery power temporarily. The goal of the FREE-D system is to minimize time spent on battery power to maximize the overall battery lifetime. Therefore, the FREE-D power tracking algorithm targets a rectified voltage of 18V, which allows for sufficient headroom for the buck converter in case the power delivery falters slightly and also maintains high DC-DC conversion efficiency.

4.3.2 Power Tracking: Implementation

If the optimal operating conditions for frequency tracking or adaptive impedance matching are unable to provide sufficient power to the load, the transmitter must increase power level. Similarly, if the load power suddenly decreases, the transmitter may need to decrease the amount of power sent to the receiver. Voltage dividers and current sense amplifiers on the FREE-D receiver measure voltage and current respectively and pass this information to the MSP430 MCU. These measurements include rectified voltage, battery voltage, load voltage, load current supplied by wireless power and load current supplied by the battery. These measurements are sent back to the wireless power transmitter using the 433MHz radio. Once the DSP receives this information, it decides whether to increase, decrease, or maintain the transmit power level. Increasing the transmit power level will increase the rectified voltage.

Decreasing the power level will decrease the rectified voltage. These two functions happen continuously until the measured rectified voltage meets the targeted rectified voltage range of 18V with a tolerance of 1V. The target voltage level and tolerance are easily programmable, but were selected for this application since the buck converter and motor controllers are most efficient with a supply voltage of 18V. The algorithm diagram for power tracking is shown in Figure 4.13.

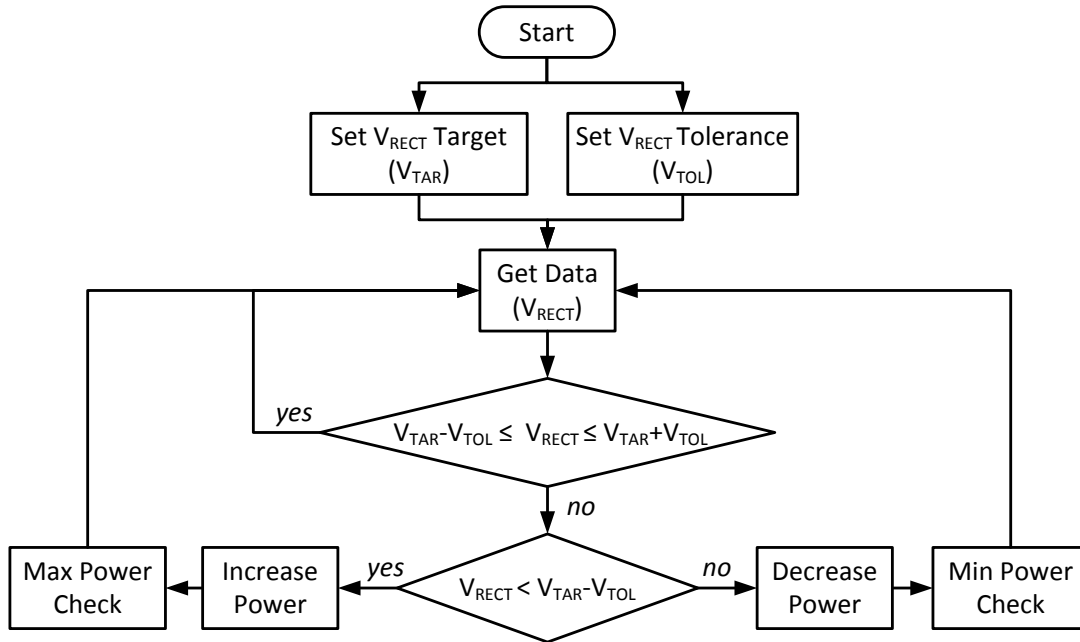


Figure 4.13: Algorithm diagram for power tracking.

Power tracking can be used in conjunction with frequency tracking and adaptive impedance matching. However, since power tracking relies on the out of band data communication link whereas frequency tracking relies only on the RF detector measurements local to the DSP, power tracking adapts slower than frequency tracking. For comparison, the ADC on the DSP samples at a maximum sampling rate of 150MHz. The out-of-band data communication data rate is 50Hz because there are

several sequence number verifications and cyclic redundancy checks to maximize the reliability of the wireless data communication to and from the implanted VAD at the expense of reduced data rates.

4.4 *Dynamic VAD Speed Control*

The FREE-D system can be completely controlled from a Python-based graphical user interface (GUI) shown in Figure 4.14. The GUI controls all of the wireless power settings, VAD motor controller settings including continuous flow, pulsatile flow, and compatibility with the Heartware HVAD, Thoratec HeartMate II, and Ventracor VentriAssist VAD, and displays all of the information measured directly on the receiver board and transmitted back to the wireless power controller using the 433MHz radio link. The first plot on the top left of the GUI shows the magnitude and phase of the forward and reflected waves from the RF detector. The second plot shows measured power levels at the receive device. The red, green and yellow status indicators show that wireless power is currently providing 100% of the power delivered to the VAD, while the backup battery is completely inactive. The pump control settings are also shown on the right hand side. Currently, the VAD is operating in a continuous flow mode. However this can be changed to pulsatile flow from the dropdown menu.

The ability to dynamically change the pump speed and pulsatility of the VAD is unique to the FREE-D system. All commercially available VADs operate with either a continuous flow or pulsatile flow. With FREE-D, physiological advantages may be realized by synchronizing the pulsatility of the VAD to the natural electrocardiogram (ECG) of the patient [4, 71, 72]. Recent work has shown that pulsatile VADs are less prone to pump thrombosis, and vortex dynamics resulting in stasis than continuous flow VADs [85]. In order to realize the physiological advantages of a pulsatile flow VAD, the pulsatility must be properly configured. The FREE-D system allows the surgeon to control the continuous flow pump speed or switch to pulsatile mode in which the systolic and diastolic pump speed, heart rate, and systolic duration can all

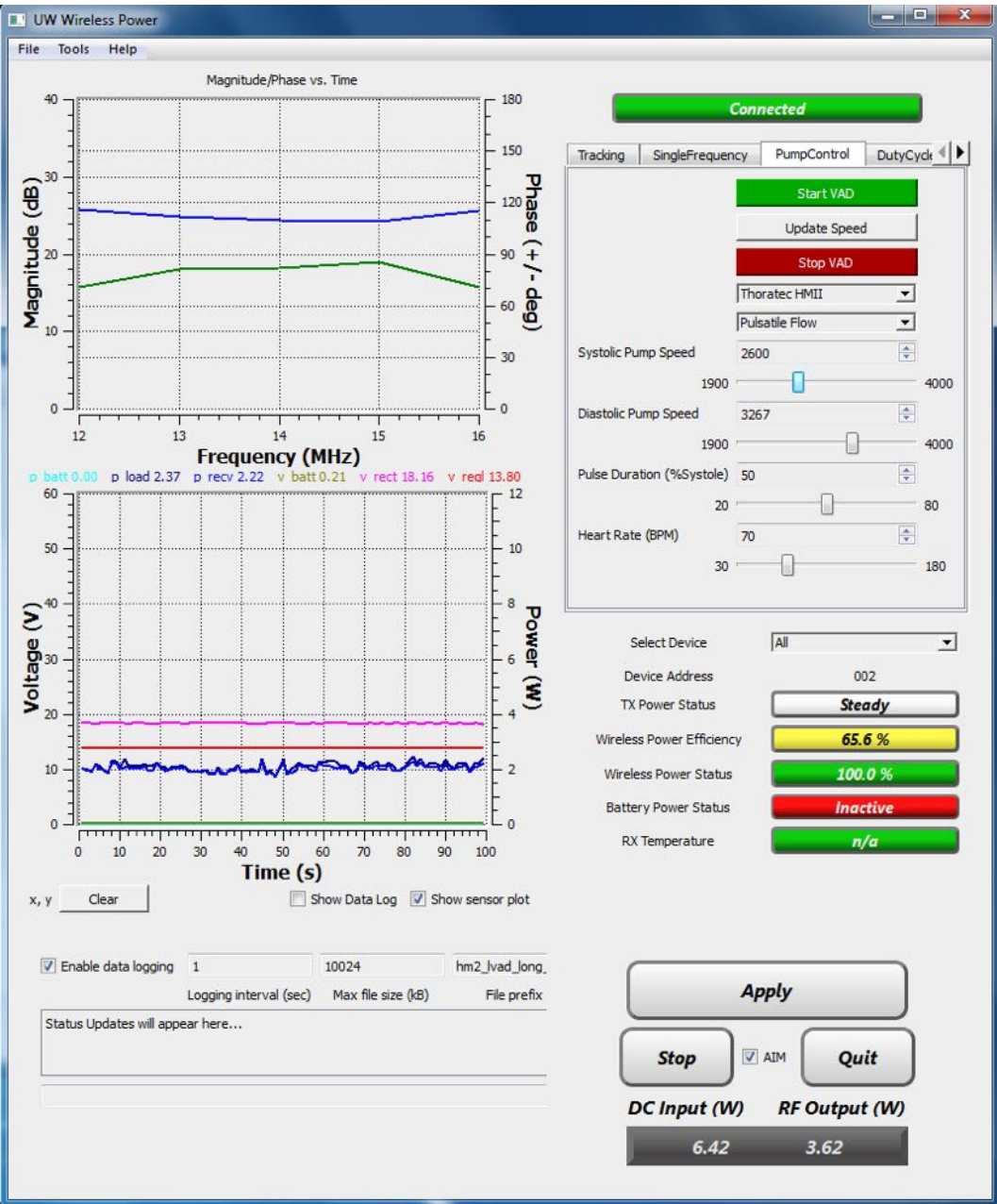


Figure 4.14: FREE-D wireless power and pump control GUI.

be set and wirelessly programmed. As a further improvement, the FREE-D system also can operate in an electrocardiogram (ECG) synchronization mode. In this mode, three electrodes are used to measure the ECG of the animal in real time. A time

delay can be implemented to synchronize the pulsatility of the VAD with the natural pulsatility of the animal's heart. Optimizing the synchronization between the heart's natural ECG and VAD pulsatility is part of ongoing work with my collaborators at the Yale School of Medicine.

Chapter 5

IN-VIVO ANIMAL TRIALS

5.1 Overview

Eight separate acute animal trials have been performed at the Yale School of Medicine by Dr. Pramod Bonde and Dr. James Bouwmeester from the Bonde Artificial Heart Lab. The Heartware HVAD was implanted inside Yorkshire pigs on eight separate occasions. In every trial, the surgery was successful, and the pigs remained in stable condition on wirelessly powered HVAD support for the full duration of each trial. The procedures and results of these trials will be summarized in this Section.

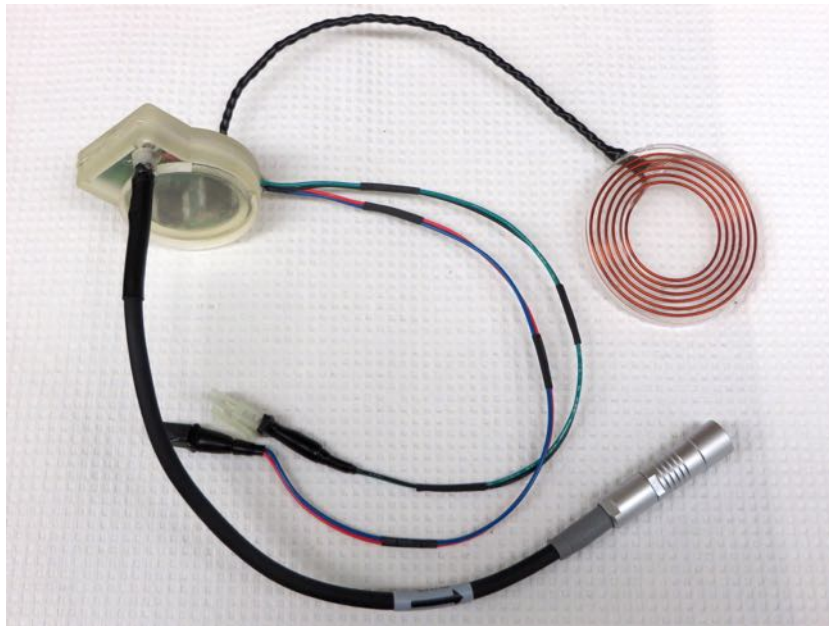


Figure 5.1: Equivalent circuit diagram of multi-transmitter WPT system with two Tx coils and one Rx coil.

Figure 5.1 shows the implanted FREE-D system components used in all eight animal trials. The coil was isolated from the receiver board and HVAD so that accurate temperature measurements of each individual component could be made. In four of the animal trials, the transmit coil was placed a short distance from the skin of the animal. This configuration will hereafter be referred to as the vest-coil configuration, since the external coil could eventually be designed into a custom vest worn by the patient. In all four of the vest-coil configuration experiments, the FREE-D transmitter was directly connected to this external coil. The goal of these four trials was to verify that the FREE-D System can provide power from a coil placed a short distance from the skin of the animal to the implanted coil beneath the skin. Figure 5.2 shows the in-vivo vest-coil configuration used throughout these animal trials.

For the remaining four animal trials, the FREE-D transmitter directly powered a large external coil placed nearly 50cm away from the implant, enabling unprecedented long-range and highly efficient wireless power transfer to an implanted VAD. This configuration will hereafter be referred to as the long-range coil configuration. Figure 5.3 shows the in-vivo long-range coil configuration used throughout these animal trials. In this configuration, the vest coil acts like a relay resonator to transfer power from the large external coil to the small implanted coil.

Even though the vest coil is still present, the advantage for the patient is that the bulky transmitter circuit does not need to be carried around, since power is provided by the large external coil. The goal of this long-range configuration is to show that the FREE-D system can achieve power delivery across a wide 50cm range, ultimately allowing the patient to experience maximal mobility throughout their home – unconstrained by heavy body-worn batteries and circuitry. Eventually, the patient may be able to move freely throughout his home, while receiving power from large transmit coils placed strategically around his home. For the first time, this vision is realized in an in-vivo animal trial.



Figure 5.2: FREE-D *in-vivo* vest coil configuration.



Figure 5.3: FREE-D *in-vivo* long-range coil configuration.

5.1.1 Animal Preparation

The Yale University Institutional Animal Care and Use Committee granted approval of all experimental protocols. Eight Yorkshire pigs (both male and female; 41 - 50 kg) were given oral amiodarone (400 mg/day) for each day of the acclimation period. Anesthesia was induced with an initial intra muscular injection of ketamine (16 - 25 mg/kg), midazolam (0.5 - 1 mg/kg), atropine (0.05 - 0.2 mg/kg), and buprenorphine (0.005 - 0.01 mg/kg). Anesthesia was maintained with isoflurane (1 - 5% ventilated in 100% O₂) and an intravascular continuous rate infusion of lidocaine (5 - 10 μ g/kg min) and ketamine (2 - 4 μ g/kg min). During surgery and experimentation the animal was supported with dobutamine (1 - 4 μ g/kg min). Animals were ventilated using a respiratory rate of 8 breaths per minute and tidal volume of 800 - 1000 mL. Normothermia was maintained by using a circulating warm-water pad. At the conclusion of the experiments, while deeply anaesthetized, the pigs were euthanized

with an intracardiac injection of Euthasol (0.2 mL/kg).

5.1.2 Surgery and Instrumentation

A median sternotomy was used for the surgical approach, which was performed without the use of cardiopulmonary bypass. The pericardium was opened and was not reapproximated after LVAD implantation. Antiarrhythmic drugs (lidocaine 7 mg/kg and amiodarone 450 mg total) were administered in preparation for insertion of the inflow cannula into the left ventricle of the heart. A 34F (40 cm long) single-stage venous drainage cannula (Thin-Flex™; Edwards Lifesciences, Irvine, CA) was attached to the LVAD inflow and a 24 F (30 cm long) elongated one-piece arterial cannula (EOPATM; Medtronic, Minneapolis, MN) was attached to the LVAD outflow. Teflon pledgeted sutures were placed radially around the left ventricular apex and the inflow cannula tip was inserted and seated at the apex. The descending aorta was exposed and the outflow cannula was inserted using a modified Seldinger technique. In each animal, a centrifugal LVAD was used (HVAD; HeartWare, Framingham, MA).

Pressure was measured in the ascending aorta, pulmonary artery, left atrium, and left and right ventricles. The 7F aortic catheter (Transonic Scisense Inc., London, Ontario, Canada) was inserted through the carotid artery and advanced to the ascending aorta. The high-fidelity catheter-tip transducer was referenced via its fluid-filled lumen to an external pressure transducer (ADIstruments, Colorado Springs, CO), with zero defined as the midlevel plane of the heart. A Swan-Ganz catheter (Edwards Lifesciences, Irvine, CA) was inserted in the jugular vein and advanced to the pulmonary artery. The distal port of the Swan-Ganz catheter was connected to an external pressure transducer. The left atrium was accessed by directly inserting a 3.5F catheter (Transonic Scisense Inc.) through the left atrial appendage. The left and right ventricles were accessed by directly inserting cannulas (Arterial Catheterization Set, Arrow International Inc., Reading, PA) through the anterior wall of the respective ventricle. Each ventricular cannula was attached to fluid-filled lines connected to external

pressure transducers.

Flow was measured in the ascending aorta, pulmonary artery, and the LVAD outflow cannula. Flow probes (Confidence-series probes and flowmeter model TS420; Transonic Systems Inc., Ithaca, NY) were placed on the main pulmonary artery (immediately downstream of the tip of the pulmonary artery catheter) and on the aorta (immediately upstream of the tip of the aortic catheter). An external clamp-on type flow sensor (PXL-series; Transonic Systems Inc.) was used to measure flow in the LVAD outflow cannula. A 3-lead ECG was also recorded.

Electronic signals were sampled at 400 Hz with data acquisition PowerLab hardware and LabChart software (ADInstruments) and filtered at 50 Hz using zero-phase digital filtering in Matlab (Mathworks Inc., Natick, MA).

After implantation, an external power supply was used to startup the HVAD due to the high initial power consumption required by the magnetically levitated motor. The HVAD magnetically levitated centrifugal motor consists of two separate stators. Upon startup, the HVAD requires both stators to be driven at a high power level to elevate the magnets and get them rotating. This initial power surge can require upwards of 50W of power. The two motor controllers on the FREE-D receiver provided this initial power surge from an external DC power supply (in place of the backup battery). The startup algorithm on the FREE-D motor controllers linearly stepped down the pump speed from the initial startup speed of 3000RPM to 1900RPM over the course of five seconds. The surgeon manually controls and varies the pump speed until the aortic pressure of the animal stabilizes. One interesting phenomenon that occurs as a consequence of these animals actually having healthy hearts is that the animals heart still pulsates even after the HVAD has been implanted and turned on. Therefore, the healthy heart actually takes some of the load off the HVAD every systolic cycle. Therefore the VAD operates at a lower power level to achieve the same blood flow rate in these animal trials.

After initial power-up, the Rx coil was placed beneath the skin as shown in Figure



Figure 5.4: In-vivo Rx coil placement

5.4. The vest coil was held 1-6cm above the surface of the skin, as shown in Figure 5.2. Several temperature, pressure and flow sensors were installed around the Rx coil and around the animal's heart to analyze any impact that the wireless power may potentially have on the cardiac function of the animal. A FLIR E-4 thermal camera was also in place to monitor the temperature of the animal and the coils.

5.2 Experimental Results: Wireless Power

After all coils and measurement probes were in place, the wireless power was enabled using a combination of frequency tracking, power tracking and adaptive impedance matching. In every experiment, the wireless power successfully powered the HVAD without any contribution from the backup battery power.

Table 5.1 summarizes the results from all eight *in-vivo* animal trials. The wireless power transfer distance is indicative of whether the vest or long-range coil configuration was used. The duration refers to the length of time that the HVAD was powered by wireless power. The duration varies for each experiment because the surgical procedures took longer in some trials. The average pump speed was measured by back

Table 5.1: SUMMARY OF *in-vivo* ANIMAL TRIAL RESULTS

Trial #	Date	Weight (kg)	Distance (cm)	Duration (HH:MM)	Pump Speed (RPM)	Pump Power (avg. W)	Efficiency (avg. %)	$\Delta T1$ ($^{\circ}C$)	$\Delta T2$ ($^{\circ}C$)	$\Delta T3$ ($^{\circ}C$)	$\Delta T4$ ($^{\circ}C$)
1	5\15\14	50	5	2:18	2250	1.84	21.7	–	–	–	–
2	5\16\14	48	5	2:10	2200	2.16	22.4	–	–	–	–
3	6\4\14	41	50	2:18	2400	1.77	16.21	1.76	1.01	–	–
4	6\5\14	49	5	2:59	2300	2.22	23.7	1.36	1.60	–	–
5	7\15\14	50	5	2:03	2480	2.74	25.5	1.33	1.53	1.40	1.09
6	11\17\14	47	50	1:20	1730	1.68	16.2	0.62	0.54	–	–
7	11\18\14	44	50	1:18	2040	2.03	17.8	1.92	1.95	–	–
8	11\19\14	46	50	2:18	2250	1.84	21.7	–	–	–	–

emf from the motor controller, and the average pump power was measured by voltage and current sense amplifiers on the receiver circuit. The efficiency represents the full system efficiency – power delivered to the VAD divided by power input to the transmitter. Thermocouples were placed directly below the Rx coil (T1), directly above the Rx coil (T2), inside the skin above the Rx coil (T3), and on the surface of the skin above the Rx coil (T4). Table 5.1 lists the maximum temperature rise during wireless power operation for each of these measurement positions. Not all thermocouples were used in each trial, which is why several of these measurements are left blank.

The system efficiency is highest in the vest-coil configurations. However the long-range configurations achieve a 10x improvement in range with only a 7.5% efficiency reduction. Also the efficiency can be correlated to the pump speed and power level, which varies between each trial. For lower power levels, the efficiency is also lower because the static power consumption of the DSP to perform adaptive tuning consumes approximately 0.7W. Compared to humans, the pigs do not require as high of a blood flow rate, and consequently the power level is much lower for these animal trials. In humans the pump power level can reach 25W, and the FREE-D System will become more efficient when operating at these higher power levels.

The temperature measurements show that in all experiments, the heat generated

by the FREE-D System never exceeds the maximum allowable temperature deviation for an implanted device inside the trunk of the body of 2°C. Typically the greatest temperature rise occurs above the Rx coil. This result can be understood by considering that closer to the surface of the skin, the natural air flow helps regulate temperature. Similarly, closer to the inside of the body, blood flow and thermoregulation helps maintain a more constant temperature as blood carries heat away from the coil. Directly above the Rx coil, both of these thermoregulating mechanisms are minimal, and consequently this position experiences the greatest temperature rise.

Temperature rise in the FREE-D System is predominantly caused by thermal conduction and thermal radiation. Current flowing through the Rx coil dissipates as heat across the small resistance presented by the turns in the coil, and heat conductively transfers to the body. Thermal radiation caused by the RF energy from the wireless power transfer can also generate heat across the conductive tissues inside the body. However the key result that has been repeatedly demonstrated through these animal trials is that the FREE-D System does not generate intolerable temperature rise inside the body.

Additional data collected during two trials (June 4 and July 15) is shown in Figures 5.5 and 5.6 respectively. These plots show the measured wireless power delivered to the receiver, backup battery power, load power, pump speed, wireless power transfer efficiency and temperature measurements.

While the wireless power was enabled during the June 4 experiment, the battery power never contributed to the load. The VAD power fluctuates due to the animal's healthy heart changing mechanical load presented to the VAD every systolic and diastolic cycle. The efficiency also changes not only because of the changing load power, but also because the animal's chest expands and contracts while it's breathing, thus the distance between the coils is constantly changing by a couple centimeters. The adaptive tuning techniques allow for the system to maintain seamless wireless power delivery, without any backup battery assist.

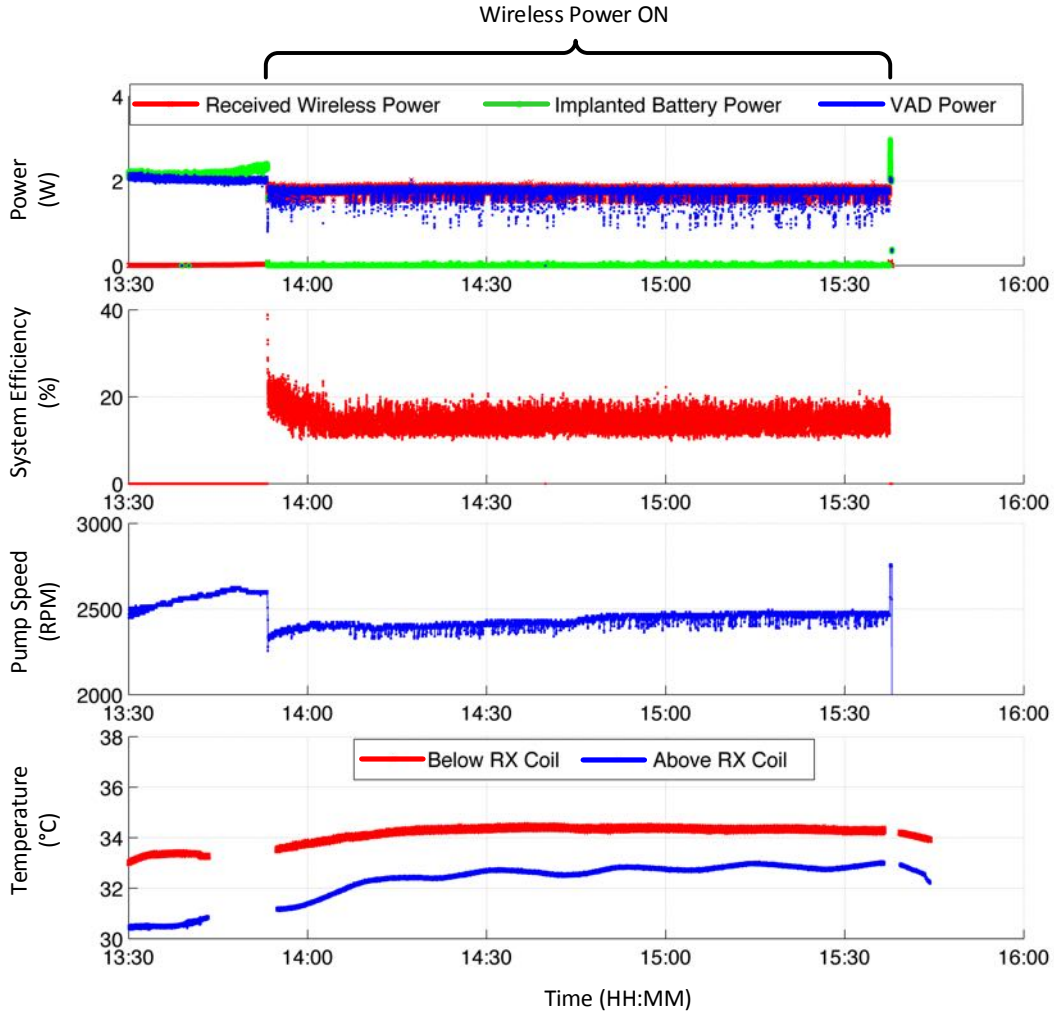


Figure 5.5: Long-range *in-vivo* animal trial results from June 4, 2014.

Before wireless power was enabled, the temperature below the coil was 33.45°C . Over the next 45 minutes, the temperature below the coil rose to a maximum of 34.46°C and maintained a relatively stable temperature for the duration of the experiment until wireless power was disabled and the coil temperature returned to its initial value. The temperature probe above the Rx coil (closer to the surface of the skin) started at 31.18°C and rose to a maximum temperature of 32.94° . The oscillations of this temperature measurement are attributed to the temperature control of the room

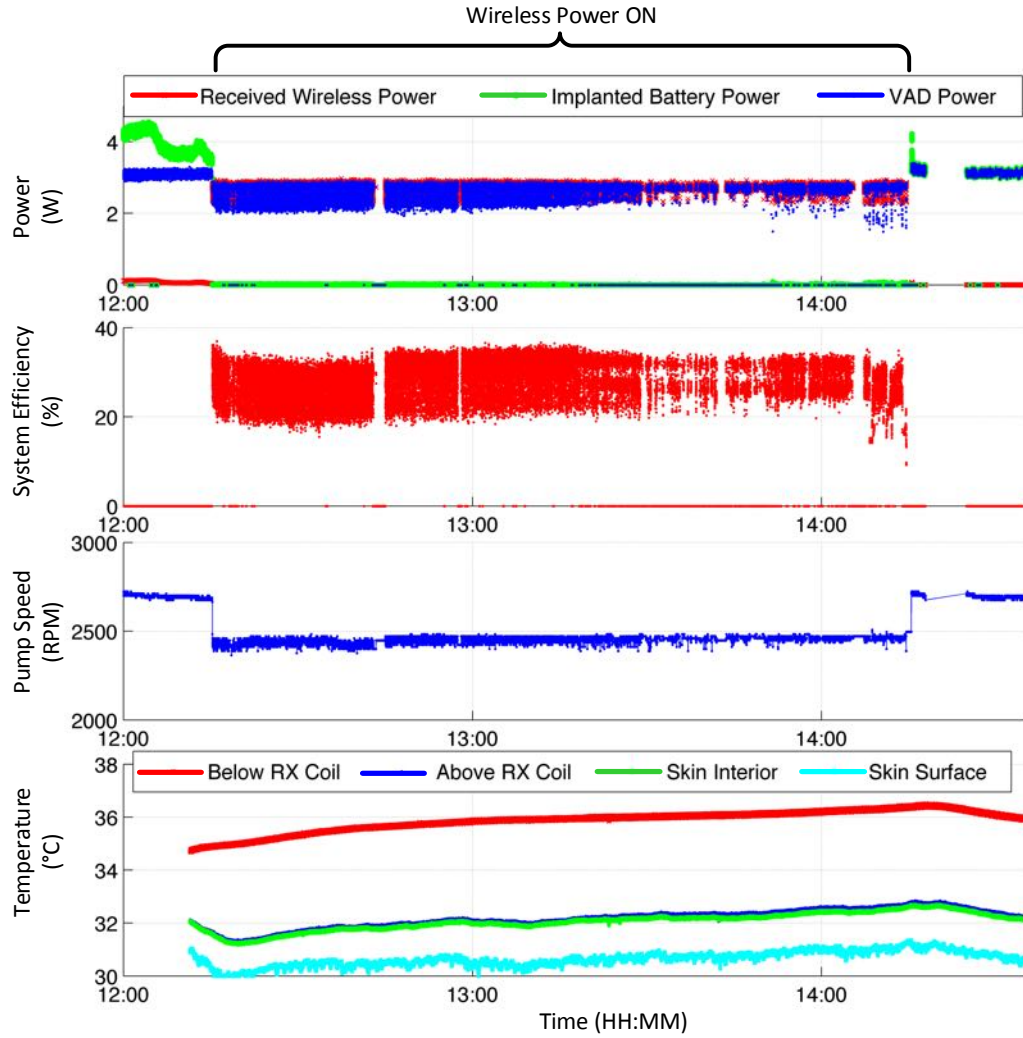


Figure 5.6: Vest-coil *in-vivo* animal trial results from July 15, 2014.

itself. At the end of the experiment, the HVAD was turned off and the experiment concluded.

In the July 15 trial, the average system efficiency was higher for this vest-coil configuration compared to the long-range configuration on June 4. Four temperature probes were used in this experiment. Unfortunately due to the limited time constraint of the trial, the trial ended before the temperature seemed to stabilize.

5.3 *Experimental Results: Continuous and Pulsatile Flow*

After the wireless power was turned off, several other features of the FREE-D system were tested that allow the pump speed to be modulated at a specified rate. In heart failure patients fully supported by a continuous flow LVAD (i.e. HVAD or HMII) the natural pulsatility is reduced and it has been hypothesized that the restoration of pulsatility by modulating pump speed will reduce the incidence of adverse events. The tests performed after the wireless power was disabled can be seen in Figure 5.7. This Figure represents the same wireless power data from the July 15 trial as in Figure 5.6. Similar tests were performed during all of the animal trials; however, only the results from July 15 are presented here.

A standard 3-lead ECG was detected by a custom designed circuit board that samples the ECG of the animal every 1 ms using gel-adhesive ECG probes. This board was attached to the FREE-D transmitter and wirelessly updated the rate of pulsatility of the HVAD. It can be seen that the pump speed oscillates during this operating mode in sync with the natural heart rate of the animal. Although the in vivo experiments in this study involved animals with healthy hearts, the effects of pump speed modulated pulsatile operation can be observed in Figure 5.8.

In the left panels, the pump is operated at a constant 2700rpm and pulsatile pump flow is caused by the natural change in differential pressure between the pump inlet and outlet. The bottom panels show the pressures measured at the pump inlet (i.e. left ventricular pressure) and outlet (i.e. aortic pressure). Of note, the negative left ventricular pressures during diastole are caused by the suction of the LVAD. During systole, when the aortic valve is open, there is a minimal pressure difference between the pump inlet and outlet, which causes pump flow to increase. During diastole, when the aortic valve is closed, there is a substantial pressure difference between the pump inlet and outlet, which causes pump flow to decrease. In the right panels, the pump rate was set to be 10% slower than the heart rate and the pump cycle was set to run

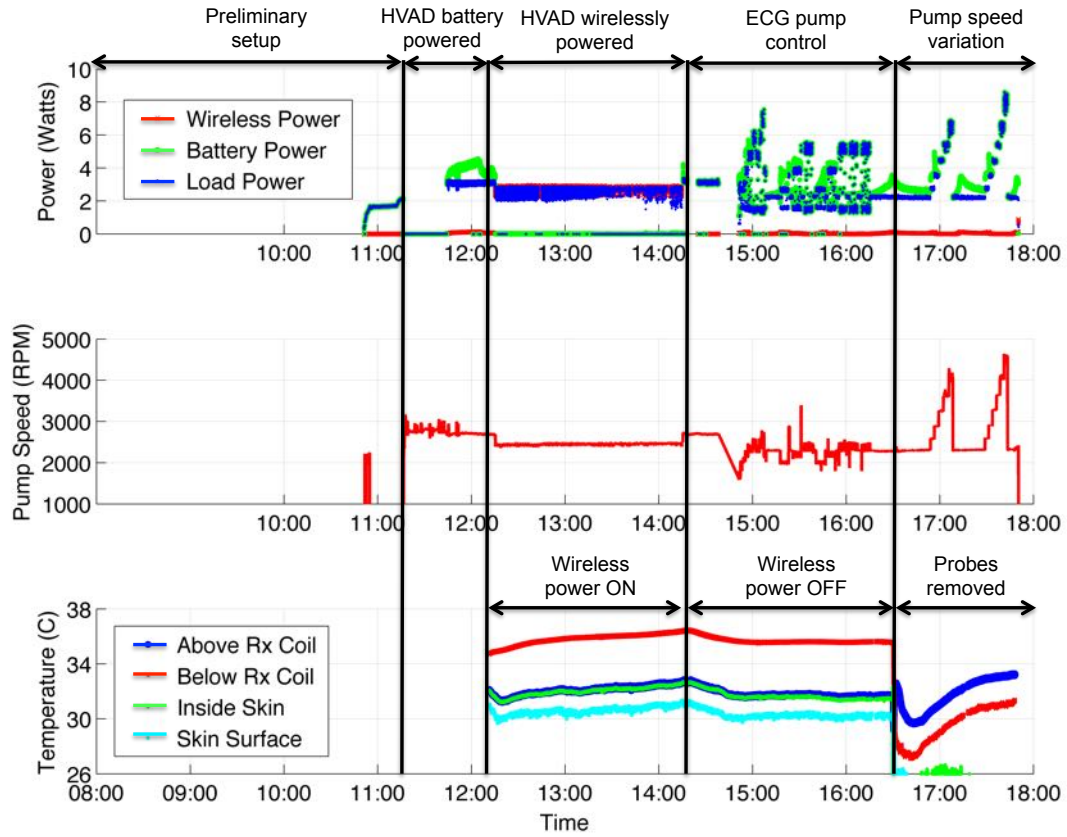


Figure 5.7: In-vivo animal trials results from July 15, 2014 at the Yale School of Medicine.

faster (3400rpm) for 30% of the cycle (i.e. to approximate the duration of systole) and slower (2300rpm) for the remaining 70%. The effect of asynchronous pumping is apparent by the phasic changes in pump flow as the speed modulated pulsation goes in and out of phase with the electrical activity of the heart. When the pump operates faster during systole (i.e., co-pulsation) the amplitude of pump flow increases and when the pump operates slower during systole (i.e., counter-pulsation) the amplitude of pump flow decreases. In the future, speed modulated pulsation may be used to promote more or less pulsatility in the cardiovascular system.

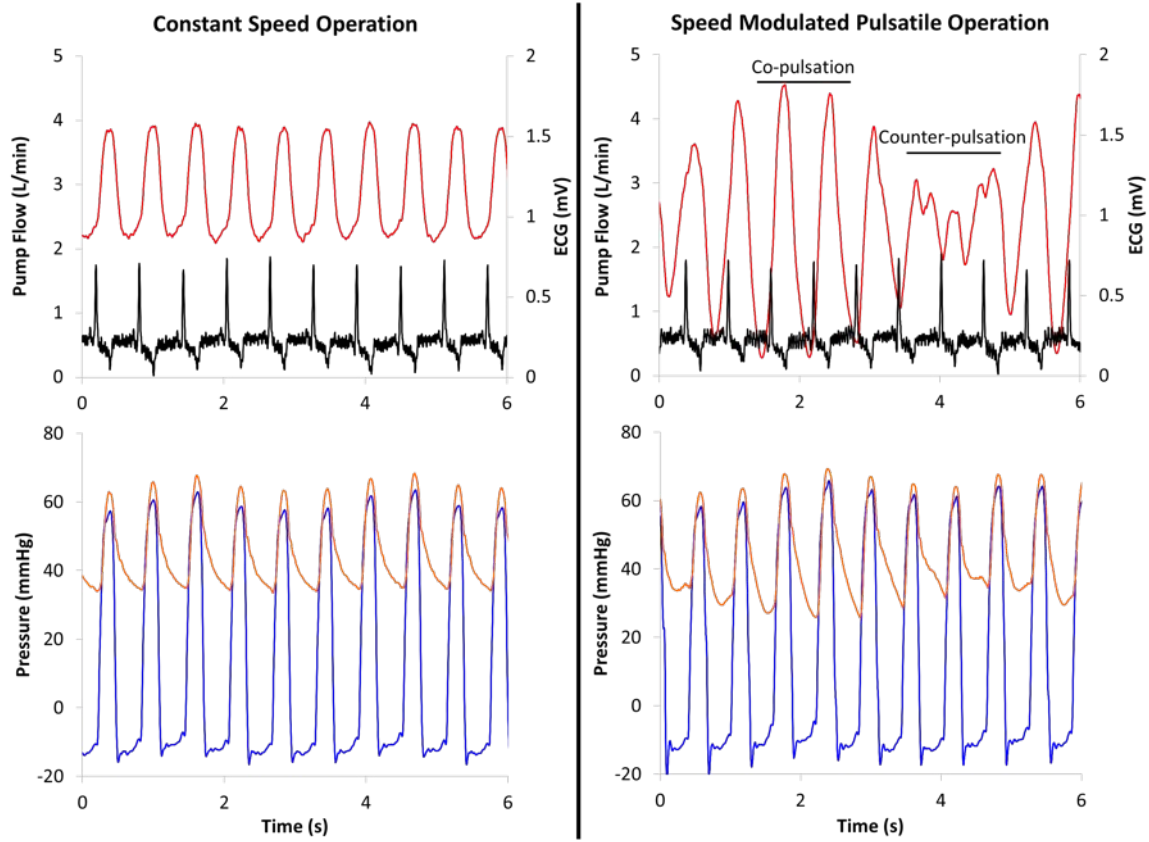


Figure 5.8: Effects of VAD pulsatility on pump flow. The top panels show the pump flow (red line) and the ECG pattern (black line) generated by the HVAD device. The bottom panels show the left ventricular pressure (blue line) and the aortic pressure (orange line), which represent the inlet and outlet pressures of the VAD respectively.

The last hour of the experiment on July 15 consisted of manually varying the speed of the HVAD. The pump speed increased from 2300rpm up to 4200rpm, which corresponds to 2-8W of power. The experiment concluded with this test.

These results show the successful implementation of the FREE-D system for in-vivo animal trials. The operating temperature of the implanted electronics stayed below the maximum allowable temperature deviation of 2°C. Additionally, the FREE-D motor controller was able to provide sufficient power to the HVAD across a full

range of pump speeds and pump power levels. Wireless data communication to and from the implanted device was reliable. And most importantly, the wireless power delivered sufficient power to the load without requiring the backup battery power for the entire duration that wireless power was enabled.

Chapter 6

WORKING TOWARDS ROOM-SIZE WIRELESS POWER RANGE FOR VADS

6.1 Overview of Wide Area WPT

The previous Section demonstrated the successful *in-vivo* implementation of both the vest-coil and long-range coil FREE-D system configurations. Both of these systems eliminate the infection-prone driveline. But how does this scale to even larger separation distances between the coils? Two particular challenges for VAD patients include sleeping throughout the night with the large, bulky vest holding the batteries and system controller, and the inability to shower. However both of these scenarios require a very wide area of coverage for Tx coils to efficiently power a very small implanted Rx coil. The ultimate goal for the FREE-D system is to provide seamless wireless power throughout the patient's entire home as shown in Figure 6.1.

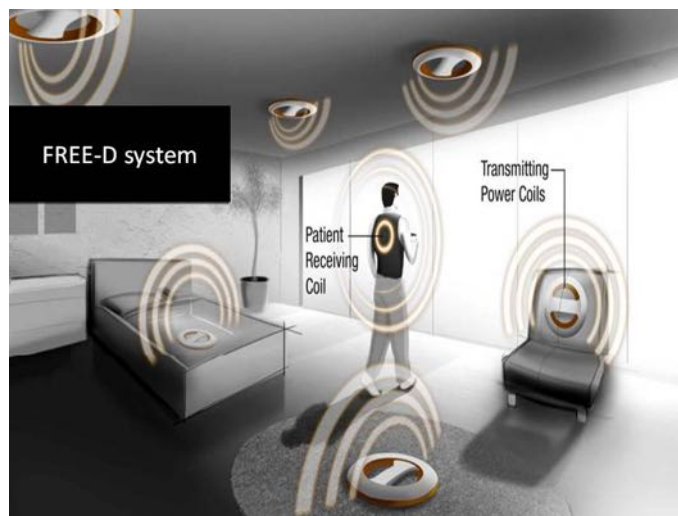


Figure 6.1: FREE-D home mobility with Tx coils placed throughout the patients home.

Section 3.2 demonstrates that one very large transmit coil cannot efficiently power a small-size receive coil because the coupling is very weak between two coils of drastically different sizes [83]. This rules out the scenario of placing one large Tx coil in the patient's bed or in the wall of the shower and expecting to power a small implanted receiver. Another scenario could be to place one smaller Tx coil in the patient's bed; however, this configuration may not provide sufficient horizontal range as the patient may roll around while sleeping.

To solve this dilemma, two different WPT systems have been designed. First, an array of relay coils can be used to ensure high efficiency between any one of the Tx coils and the Rx coil, while also enabling a wide area WPT system [82]. Second, a phased-array WPT system uses multiple transmitter circuits to drive multiple Tx coils, all of which are phase synchronized [84]. Power can be delivered to a specific location anywhere in the vicinity of the phased-array WPT system by taking advantage of constructive and deconstructive interference of the magnetic fields generated by each coil.

6.2 Wide Area WPT Using Relay Coils and Adaptive Tuning

Prior research has shown that automatic frequency tuning can increase the range of efficient wireless power transfer (WPT) [46, 68, 2]. Adaptive impedance matching has also been used to control the impedance of both the Tx and Rx coils to improve efficiency across a wide distance range at a single frequency [81]. However, these automatic-tuning systems only demonstrate WPT from a single Tx to a single Rx. Other work has shown WPT to multiple RXs, however these articles do not present methods that can optimize efficiency to each Rx operating at different load power levels [42, 1].

This section presents the hardware, software and communication protocols associated with a novel WPT system capable of powering and communicating with multiple receivers across a wide range of load power levels. The system performs three types

of automatic tuning: frequency tracking, power tracking, and adaptive impedance matching. The system also utilizes a 2.4GHz radio communication link between the wireless power Tx and Rx. Frequency tracking does not require the radio link because it relies on data collected on the Tx side only. A directional coupler and RF detector measure the magnitude and phase of the forward and reflected signals between the output of the power amplifier (PA) and the Tx coil. Power tracking is performed by varying the gain of the PA to transmit more or less power based on the receive power measured at each Rx. Adaptive impedance matching is controlled on the Rx side. When multiple receivers are present, each Rx automatically tunes its matching network to achieve a targeted receive power level.

This system can simultaneously power multiple devices, or alternatively power one device across a very wide charging range relative to the limited size of the Rx coil. Another advantage is that multiple Rx devices operating at different power levels can be sufficiently powered using the adaptive tuning techniques. Time-division multiplexing (TDM) can also be achieved. This feature allows for one or more Rx devices to be wirelessly powered while all other devices are de-tuned, even when multiple RXs are placed on the Tx coil.

6.2.1 *Hardware*

The WPT system requires two primary components: a transmitter that generates and amplifies the RF signal, and an Rx that converts the RF signal into a DC voltage for the given device. In the most simplified configuration, the Tx could use a fixed frequency source, and a fixed gain PA to deliver the RF signal to the Rx. However, for applications in which the Rx(s) will be moving relative to the Tx coil, the efficiency of the system will decrease significantly if a static Tx is used. Therefore, it is necessary to add more complexity to both the Tx and Rx to maintain high efficiency for a dynamic system. Figure 6.2 shows a schematic diagram of the full WPT system.

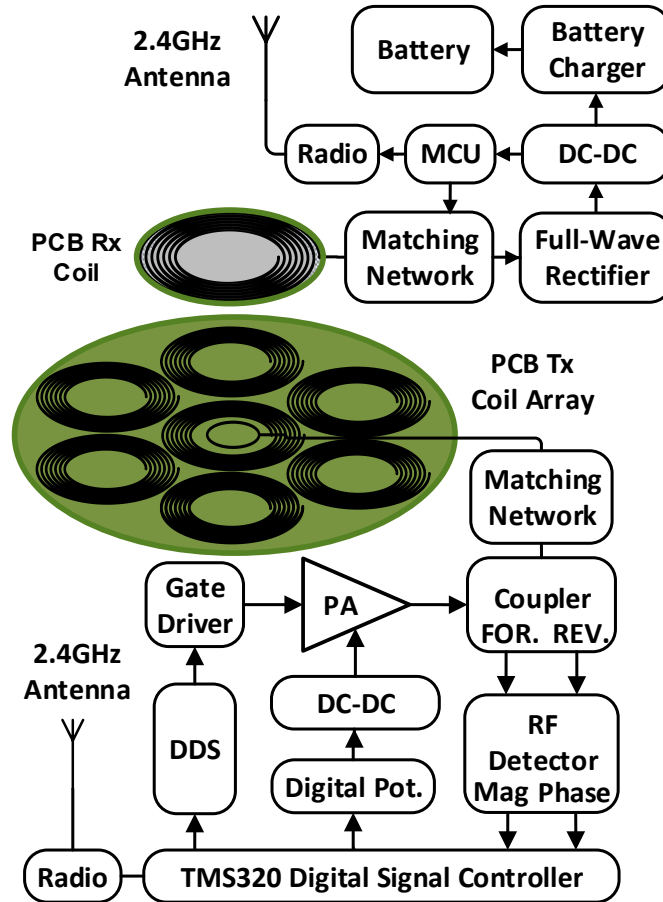


Figure 6.2: Schematic diagram of the wide-area wireless power system using relay coils.

Transmitter Hardware

On the Tx side, the TMS320 Digital Signal Processor (DSP) is the primary controller for the entire wireless power system. All measurements, commands, enable and disable functions are generated by or terminated at the DSP. The transmitter used here is the same transmitter used in the FREE-D system, as shown in Figure 3.30.

The AD9850 direct digital synthesizer (DDS) generates the RF signal that drives the PA. The amplitude, frequency and phase of the DDS can all be set by the DSP. The PA consists of a gate driver that is used to drive an IRF510 Mosfet in a single-ended class-E amplifier. Although class-E amplifiers can achieve high efficiency, the

downside is that they are typically designed to be most efficient at a single operating point in terms of power level, load impedance and operating frequency. In a WPT system where the load impedance changes as a function of load power and distance between the coils, a class-E amplifier will rarely operate at the fixed operating point that it was initially designed for. Therefore, the class-E amplifier used in this system has a programmable supply voltage and a programmable output matching network that can optimize the efficiency of the amplifier over a wide range of load conditions. The programmable supply voltage is realized by using a digital potentiometer to control the feedback voltage of a buck-boost DC-DC converter. The adjustable output matching network consists of a low-pass π -match network with switchable capacitor banks as in Section 4.2.

The Tx coil consists of seven separate coils on a single PCB. The center coil also has a drive loop, which is the only element of the entire coil array that is directly connected to the output of the Tx PCB. Each coil is identical with an inductance $L=1.11\mu\text{H}$, tuning capacitor $C=120\text{pF}$ and parasitic resistance $R=0.79\Omega$ with a quality factor $Q=125$ at a resonant frequency of 13.56MHz. Each Tx coil surrounding the center coil acts like a relay resonator to extend power to each Rx.

Receiver Hardware

The Rx coil is designed into a PCB as in Figure 6.3. A sheet of 0.25mm thick ferrite placed directly above the Rx coil isolates the leakage fields generated by the Rx coil from the Rx PCB. Using the LTC4065 battery charger with a bank of digitally controlled switchable resistors, various charge currents can be set for each Rx by the MSP430 MCU. An ADC measures the received voltage and current and can send this information back to the Tx via a 2.4GHz radio link. To communicate with multiple RXs simultaneously, a star network wireless communication system has been implemented in which the Tx is the primary access point (AP), and each Rx is an end-device (ED).

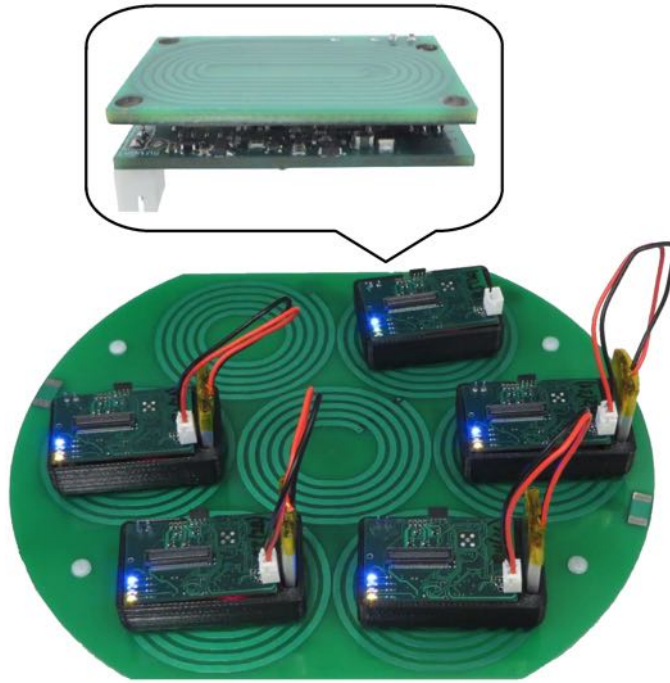


Figure 6.3: Transmit coil array powering five receivers simultaneously

6.2.2 Software

When powering multiple RXs with a single Tx, there is a tendency for one Rx to hog all of the transmitted power, starving other devices of power. The WPT system avoids this problem by using a combination of sensor measurements made on both the Tx and Rx sides to dynamically tune the WPT system. The various tuning algorithms rely on either in-band measurements (forward and reflected signals measured by an RF detector on the Tx side) or out-of-band (OOB) voltage and current measurements sent across the 2.4GHz radio link. Each ED reports its wireless power consumption (rectified voltage times rectified current) back to the AP. Conversely, the AP may send commands to the individual EDs to control the adaptive matching network on each Rx. This closes the feedback loop and provides the fine-grained control needed to regulate the amount of power that each ED receives. Therefore, if any given Rx hogs too much power, it can be dynamically de-tuned so that another underpowered

Rx can be optimally tuned, enabling all RXs to have sufficient power delivery.

Frequency Tracking

Section 4.1 has shown that frequency tracking can maintain efficient power transfer between a Tx and a single Rx. With multiple RXs, the optimal frequency for one Rx may not be ideal for a different Rx. Consequently, the frequency tracking algorithm typically selects the frequency corresponding to the receiver that is most strongly coupled to the Tx coil. Therefore frequency tracking can be counter-productive with multiple RXs if not used in conjunction with power tracking and adaptive impedance matching.

Power Tracking

Power tracking uses the OOB radio link to create an additional feedback loop with the Tx. Each ED communicates its received power measurements back to the AP, over the OOB radio link. The Tx uses this information to adjust the gain of the PA by increasing or decreasing the DC voltage supplied to the PA. If any device receives too much power (i.e. rectified voltage exceeds 20V), then the transmit power level decreases. Otherwise, if any device receives too little power (i.e. rectified voltage falls below 6V), the transmit power level increases. In multiple-receiver systems, this algorithm effectively keeps all RXs sufficiently powered. However, power tracking alone cannot regulate how power is split between RXs, and certain RXs may be significantly overpowered if they have lower load power requirements or are more strongly coupled to the Tx coil than adjacent RXs.

Adaptive Impedance Matching

Adaptive impedance matching allows for each Rx to regulate the amount of wireless power it consumes. Unlike power tracking, this algorithm can operate independently

from the Tx, and therefore does not necessarily require the OOB radio link. Each Rx monitors its rectified voltage, and periodically checks whether it lies within an acceptable range (the same range as power tracking, 6-20V). If outside the range, then that Rx performs a sweep of all possible impedance matching settings and selects the configuration that achieves a rectified voltage within the acceptable range. In some cases, the chosen impedance setting may intentionally de-tune the Rx that would otherwise be prohibitively overpowered. Therefore this algorithm optimizes power delivery to the entire set of RXs, rather than maximize power delivered to a single Rx. This algorithm ensures that each Rx presents a similar impedance from the perspective of the Tx coils, even if the coupling and power level of the RXs are different. In this system, each ED has only five possible impedance settings. The algorithm used to perform adaptive impedance matching is identical to the algorithm shown in Figure 4.12.

Time Division Multiplexing

Using a combination of frequency tracking, power tracking and adaptive impedance matching, the system can optimize power delivery to a subset of devices, and subsequently ignore the remaining RXs. A practical scenario where this mode could be useful is Rx prioritization: the most critical Rx could be prioritized and rapidly charged while other RXs could either trickle charge or not charge at all until the critical device is fully charged. The system can also optimize multiple devices at once, so that an arbitrary subset of RXs charges simultaneously while others remain unpowered, even while they are in range of the Tx coils.

TDM modes can be arbitrated in one of two ways. First, the Tx can control the impedance settings on all RXs by sending commands across the OOB radio link. Second, each Rx can independently control its own impedance settings. If the second mode is used, then the OOB radio link can be removed from the system, saving cost and PCB area, while still enabling frequency tracking on the Tx side and adaptive

impedance matching on each Rx.

6.2.3 Experimental Results

Using the setup shown in Figure 6.3 five single cell batteries were charged at different power levels simultaneously. The battery charger on each receiver was programmed for a different charge current of 50mA, 75mA, 100mA, 125mA, and 150mA respectively. Figure 6.4 shows a plot of charge current versus time for each receiver.

To highlight the benefits of using frequency tracking, power tracking, and adaptive impedance matching, compare the system performance with these dynamic tracking modes enabled (Figure 6.5) to the performance when the tracking modes are disabled (Figure 6.6). The rectified voltage reflects the equality of power distributed to each receiver. If the rectified voltage is less than 6V, then the respective receiver is not wirelessly powered, but rather powered by the battery. If the rectified voltage is high (i.e. above 20V) then the given receiver is overpowered and may hog power from adjacent receivers. With only one or two receivers present, both systems can provide sufficient power to each receiver. However when three or more receivers are present, only the adaptive system provides sufficient power to all receivers. The non-adaptive system suffers from uneven power distribution to the various receivers: the low charge

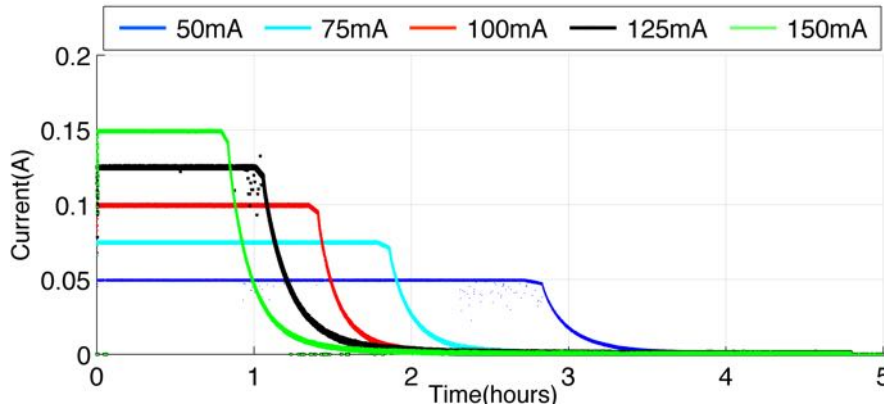


Figure 6.4: Charge current for five wirelessly powered battery chargers.

current receivers are sufficiently powered while the high charge current receivers are underpowered.

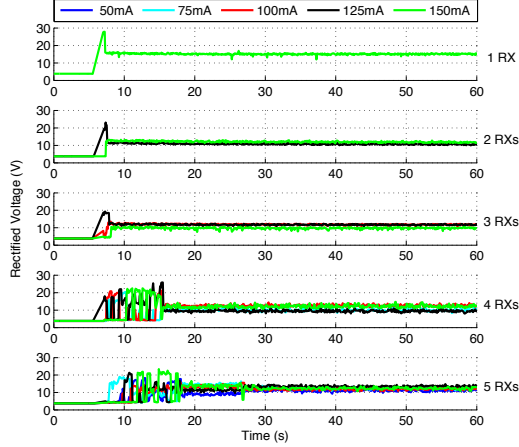


Figure 6.5: Rectified voltage for 1-5 receivers using adaptive tuning techniques.

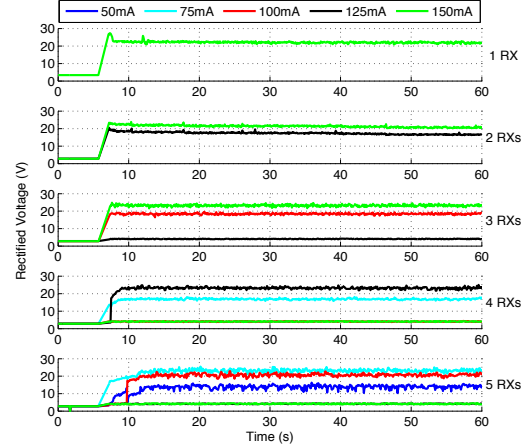


Figure 6.6: Rectified voltage for 1-5 receivers without adaptive tuning techniques.

To demonstrate the TDM capabilities, the transmitter commands the matching networks on each receiver to sequentially optimize power delivery to one receiver while pessimizing power delivery to all other receivers. After holding this state for approximately 30 seconds, the next receiver gets optimized and those remaining are pessimized. It should be noted that two RXs were set for 75mA charge current and the other three RXs were set at 100mA. Figure 6.7 shows one complete sequence of optimizing power delivery to each receiver sequentially.

6.2.4 Summary of Wide Area WPT Using Relay Coils

Consider how this relay coil array WPT system can be used in the FREE-D system, for either the bed or shower charging scenarios. The FREE-D transmitter could directly power one Tx coil, and all adjacent transmitters would act as passive relay coils to route this power to the implanted receive coil. This scenario may be suitable for some

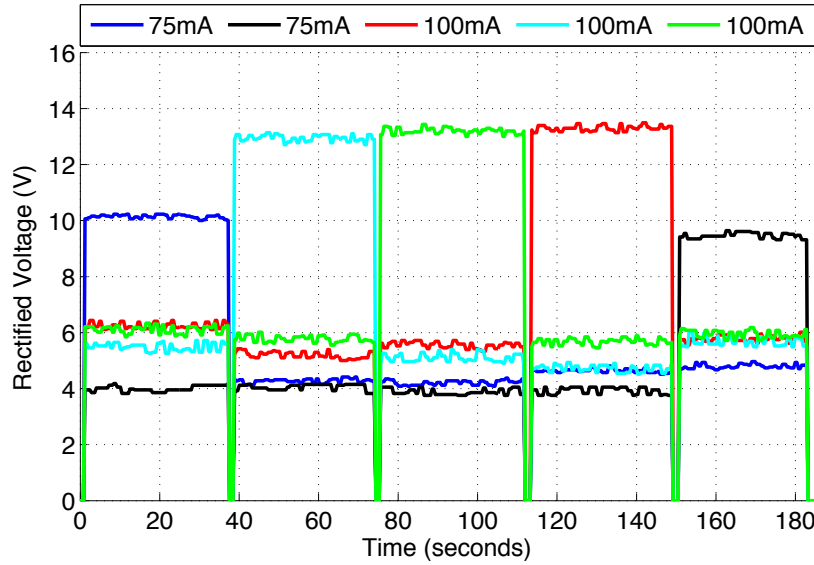


Figure 6.7: Sequential charging using time division multiplexing.

areas on the bed, but there are likely to be nulls where the magnetic fields generated by adjacent transmitters destructively interfere, for example if the Rx coil is situated directly between two adjacent Tx coils. A technique to eliminate these potential null areas involves driving two or more adjacent transmit coils. These transmitters must be phase-synchronized, and by controlling the phase difference between two or more driven Tx coils, maximum power delivery can be targeted for certain locations.

6.3 Wide Area WPT Using A Phased Array Wireless Power System

If the desired transmission distance is greater than the diameter of the smallest coil in the system, or if the size of the Tx coil is much larger than an Rx coil, efficiency drops drastically [83]. The FREE-D bed-charging scenario is a perfect example of a practical application facing this challenge.

To avoid these design challenges, transmit coil arrays can be used to improve efficiency across a wider power transfer range. A phased array WPT system consists of two or more transmit coils each driven by a PA to wirelessly power one or more

receivers. The transmitters are all phase-synchronized at the same frequency and the phase relationship between each transmitter can be dynamically controlled to enable constructive or destructive interference between the magnetic fields generated by each Tx coil. Not only does this system allow for increased range at which high efficiency can be achieved compared to a single transmitter system, but there are also several other advantages of this phased array WPT system.

First, maximum efficiency can be achieved anywhere within a defined volume of space, regardless of the orientation or position of the receiver by optimizing the frequency, magnitude and phase of the various transmitters. Second, maximum power regions and null-power regions can be generated simultaneously within a defined volume of space. This is desirable for systems that consist of multiple receivers because certain receivers can be targeted for charging while other receivers or even extraneous foreign objects will not be charged. Third, leakage fields can be reduced when maximum power is transferred to a targeted receiver compared to the single transmitter configuration. Minimum leakage fields are desirable for demonstrating regulatory compliance and mitigating the amount of energy induced upon foreign objects.

Prior work has presented systems utilizing multiple transmit coils [38, 10, 80, 37]. Johari et al. used multiple drive loops and Tx coils capable of improving power delivery to an Rx coil when conductive foreign objects enter the region between the coils [38]. Jadidian et al. present “magnetic mimo” that is capable of charging a cell-phone inside of a person’s pocket nearly 40cm away from the array of Tx coils [37]. Uchida et al. provide analysis for a perpendicular arrangement of two Tx coils to power mobile devices anywhere inside the Tx coil region [80]. However these articles neglect the coupling between the Tx coils, and do not provide analysis for scaling these systems to a greater number of Tx coils. It will be shown that the coupling between the transmit coils has a significant impact on the phase at which maximum (or minimum) power is delivered to the Rx coil. Ahn et al. present a WPT system with multiple Tx and Rx coils with inter-coil coupling that uses frequency tuning to

optimize the system efficiency [1]. However frequency tuning may violate the allowable bandwidths defined by federal regulatory bodies such as the FCC [14]. This work leverages the coupling between the Tx coils and uses amplitude and phase control to overcome decreasing efficiency associated with strong coupling between Tx and Rx coils. Oodachi et al. show the efficiency improvement achievable with a phase-synchronized multi-coil system over a single Tx coil configuration [58]. However, the article does not include circuit analysis showing the parameters required to maximize power delivery to a mobile receiver. Orientation-independent Rx coils have also been demonstrated [39, 34, 64, 54]; however, the inherent efficiency limits shown for these coil designs compares unfavorably to the phased-array system in this work, which also has the benefit of orientation-independence.

A thorough circuit analysis for a WPT system with two driven Tx coils and one Rx coil is provided in this Section. This analysis is also generalized for a system with any number of Tx or Rx coils, and simulation results highlight some key benefits of adding more Tx coils to a phased array WPT system. The control variables for these systems are the magnitude and phase of each transmitter. Expressions are derived for the magnitude and phase that maximize or minimize power delivered to the Rx coil given any coupling coefficient arrangement between the various coils. Additionally, a transition point is defined that allows the system to determine when the power delivered to the Rx coil is greater if only one transmitter is used and the other is turned off. Experimental results using this phased array WPT system are presented to verify the theoretical analysis. Additional capabilities of the phased array WPT system including minimized leakage fields and orientation independence of the Rx coil inside of a charging box are also shown. Finally, a properly configured phased array WPT system using adaptive tuning can achieve higher system-level efficiency across a larger distance range than a standard single Tx coil WPT system.

6.3.1 Theoretical Analysis

Beamforming using a phased-array of transmit antennas has shown promise for extending the range in far-field wireless applications [27, 48, 5]. These systems rely on shifting the phase of one transmitter relative to the other transmitters to achieve constructive or destructive interference in the transmitted radiation pattern to maximize power delivered to a receive antenna. Mutual coupling in far-field phased-array systems can significantly affect the system as spacing between antenna elements decreases [30]. Consequently the effect of mutual coupling is typically considered parasitic and requires mitigation techniques. However, the parasitic mutual coupling can be leveraged to improve range and efficiency [36]. Near-field WPT systems must also optimize the mutual coupling between coils for efficient operation.

6.3.2 Equivalent Circuit Analysis

Figure 6.8 shows the equivalent circuit model of a phased array WPT system using a Tx coils and b Rx coils. Each coil consists of a winding inductance $L_{a|b}$, a parasitic AC resistance $R_{a|b}$ and a series tuning capacitor $C_{a|b}$. All coils are coupled by the coupling coefficient k . Consequently, there are $n(n-2)/2$ coupling coefficients where $n = a + b$ is the total number of Tx and Rx coils in the phased-array WPT system. Each Tx coil is driven at the same frequency by a phase-synchronized voltage source with adjustable magnitude and phase.

Using lumped-element circuit theory as in Section 3.1.1, the output voltage V_o at any of the receiving coils may be derived using mesh-current analysis. The matrices in (6.1) present the general case of n coupled coils. The $\mathbf{Z}$ matrix is square, and thus contains n^2 elements. The main diagonal comprises the reactance and resistance of each coil, which is a function of the self-inductance, tuning capacitance, source and load impedances and parasitic resistances. Using Cramer's rule, the unknown mesh currents can be determined. The $\mathbf{V}$ vector accounts for the driving voltage source

$$\begin{bmatrix} V_{Tx,1} \\ \vdots \\ V_{Tx,a} \\ 0 \\ \vdots \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{Tx,1} & j\omega M_{12} & j\omega M_{13} & j\omega M_{14} & \cdots & j\omega M_{1n} \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ j\omega M_{a1} & j\omega M_{a2} & Z_{Tx,a} & j\omega M_{a4} & \cdots & j\omega M_{an} \\ j\omega M_{(n-b)1} & j\omega M_{(n-b)2} & j\omega M_{(n-b)3} & Z_{Rx,1} & \cdots & j\omega M_{(n-b)n} \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ j\omega M_{n1} & j\omega M_{n2} & j\omega M_{n3} & j\omega M_{n4} & \cdots & Z_{Rx,b} \end{bmatrix} \times \begin{bmatrix} I_{Tx,1} \\ \vdots \\ I_{Tx,a} \\ I_{Rx,1} \\ \vdots \\ I_{Rx,b} \end{bmatrix}$$

$$\mathbf{V} = \mathbf{Z}\mathbf{I} \quad (6.1)$$

$$\begin{aligned} Z_{Tx,a} &= R_{S,a} + R_{p,a} + j\omega L_a + \frac{1}{j\omega C_a} \\ Z_{Rx,b} &= R_{L,b} + R_{p,b} + j\omega L_b + \frac{1}{j\omega C_b} \\ I_i &= \frac{\det(\mathbf{Z}_i)}{\det(\mathbf{Z})} : i = 1, 2, 3 \dots n \quad V_{o,b} = I_b \times R_{L,b} \end{aligned} \quad (6.2)$$

of each Tx coil. The driving voltage corresponding to each Rx coil is 0 since the Rx coils are not driven.

Assume that there are a transmitters and b receivers. In each Tx coil, the resistance is the sum of the source and parasitic resistances. Similarly, in each Rx coil, the resistance is the sum of the load and parasitic resistances. Also, the load in each Rx coil is purely resistive. Thus the output voltage of the i^{th} coil V_{oi} is the product of load resistance and the mesh current. M_{ij} is the mutual inductance between the any two coupled inductors and can be calculated as a function of the distance and angular alignment between two coils [62]. The general solution can be found for any number of Tx and Rx coils using (6.1) and (6.2).

Assuming that the frequency of the two coupled Tx coils is equivalent, phasor analysis can be used to add the contributions of each transmitter. The cumulative effort of the two Tx coils may interfere constructively or destructively, depending

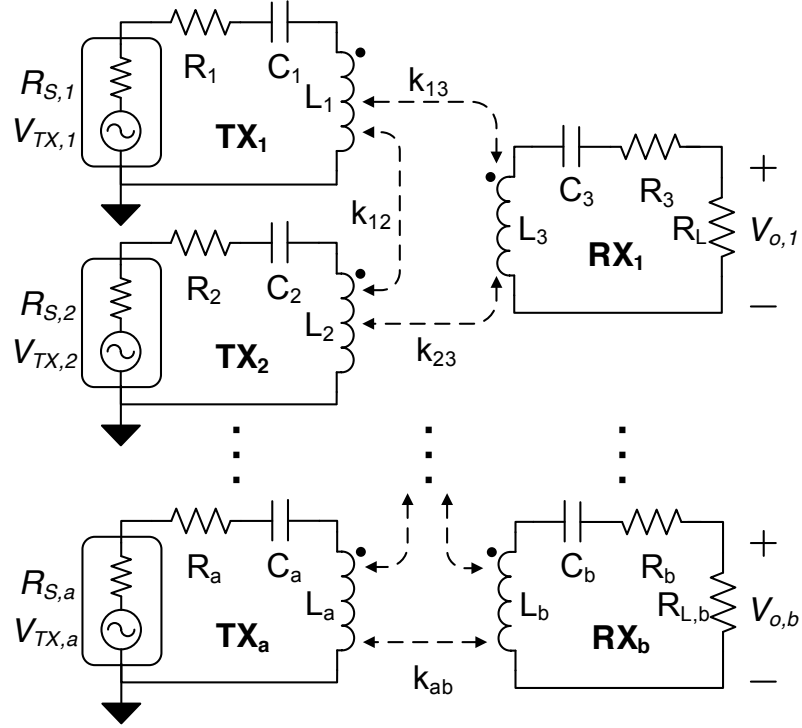


Figure 6.8: Equivalent circuit diagram of a WPT system with multiple transmitters and receivers.

on the relative phase difference. In this analysis, the voltage source for the first Tx coil (Tx₁) is considered the reference source, with a phase $\phi_1 = 0$ and a magnitude $\alpha_1 = 1$. Thus the phase difference $\Delta\phi = \phi_2 - \phi_1$ between the two Tx coils reduces to the phase ϕ_2 of the second transmitter's (Tx₂) voltage source. Ideally, for a symmetric system with $k_{12} = 0$, if $\phi_2 = 0$ then constructive interference results in maximum V_o . Conversely, if $\phi_2 = 180^\circ$, then the two transmitters are out of phase, which results in destructive interference and consequently a reduction in V_o . This anti-phase condition minimizes power delivered to the Rx coil.

For the remainder of the 3-coil discussion, α and ϕ will represent the magnitude and phase respectively of the transmitted signal from Tx₂ such that $V_{S1} = \cos(\omega t)$ and $V_{S2} = \alpha \cos(\omega t + \phi)$. The magnitude of Tx₁ will be normalized to one and the phase

$$V_o = \frac{\omega R_L [\omega M_{12} M_{23} + j M_{13} Z_2 + (\omega M_{12} M_{13} + j M_{23} Z_1) \alpha \cos \phi + j (\omega M_{12} M_{13} + j M_{23} Z_1) \alpha \sin \phi]}{Z_1 Z_2 Z_3 + \omega^2 (M_{12}^2 Z_3 + M_{13}^2 Z_2 + M_{23}^2 Z_1) - 2j \omega^3 M_{12} M_{13} M_{23}} \quad (6.3)$$

$$\alpha_{opt,min} = \frac{\omega^3 k_{12} L R C^2 (k_{23}^2 - k_{13}^2) \sin \phi - k_{13} k_{23} (\omega^2 R^2 C^2 + k_{12}^2) \cos \phi}{\omega^2 k_{23}^2 R^2 C^2 + k_{12}^2 k_{13}^2} \quad (6.4)$$

$$\alpha_{opt,min,symm} = \frac{k_{13}}{k_{23}} \cos \phi \quad (6.5)$$

$$\phi_{opt,min} = \tan^{-1} \left[\frac{\omega^3 k_{12} (k_{13}^2 - k_{23}^2) L R C^2}{k_{13} k_{23} (\omega^2 R^2 C^2 + k_{12}^2)} \right] \quad (6.6)$$

$$\phi_{opt,max} = 180^\circ - \phi_{opt,min} \quad (6.7)$$

set to zero such that $V_{S1}=1$ in the phasor domain. Using Euler's identity, the second source will provide the magnitude and phase offset relative to the first source such that $V_{S2}=\alpha \cos(\phi) + j\alpha \sin(\phi)$. Additionally, a symmetric system will be considered such that $L_1 = L_2 = L_3 = L$, and similarly for R and C of the coils. The same analysis can be applied to an asymmetric system.

6.3.3 Minimize and maximize power delivered to the receive coil

For the objective function V_o , $\alpha_{opt,min}$ represents the optimal solution that minimizes V_o . The expression for $\alpha_{opt,min}$ in (6.4) is a function of the inter-coil coupling coefficients k_{12} , k_{13} and k_{23} , the coil parameters L , R and C and ϕ . $\alpha_{opt,min}$ is derived by differentiating V_o with respect to α , setting the result equal to zero, and solving for α . For a given ϕ value, $\alpha_{opt,min}$ represents the magnitude that will give a relative minima of V_o at that specific phase shift. Since $\alpha_{opt,min}$ is dependent on ϕ , $\alpha_{opt,min}$ will only imply an absolute minimum for V_o if the phase at which V_o will be minimized ($\phi_{opt,min}$) is specified.

When $k_{12} = 0$, (6.4) simplifies to (6.5). In this symmetric case, the absolute $\alpha_{opt,min,symm}$ always occurs at $\phi = 180^\circ$, which creates perfect destructive interference between the two transmitters and allows for an absolute minimum of $V_o = 0V$. If $k_{13} > k_{23}$ and $k_{12} = 0$, α_2 needs to be greater than α_1 by a factor of k_{13}/k_{23} to minimize V_o because Tx₂ needs to compensate for the stronger coupling between Tx₁ and the Rx coil. Alternatively, when $k_{13} < k_{23}$ and $k_{12} = 0$, α_2 needs to be less than α_1 by a factor of k_{13}/k_{23} in order to minimize V_o . In the case where $k_{13} \gg k_{23}$, an absolute minimum of $V_o = 0V$ is not possible to achieve unless α_2 is very large in magnitude (i.e. Tx₂ outputs significantly more power than Tx₁).

$\phi_{opt,min}$ is derived by differentiating V_o with respect to ϕ , setting the result equal to zero, and solving for ϕ as in (6.6). Since $\phi_{opt,min}$ is independent of α , $\phi_{opt,min}$ can be calculated first, and then used to find the $\alpha_{opt,min}$ at which an absolute minimum of V_o can be achieved.

The value of α that maximizes V_o ($\alpha_{opt,max}$) always corresponds to the largest allowable value of α . This is logical because α represents the magnitude of the transmitted power: sending more power results in a larger V_o . In a real system, $\alpha_{opt,max}$ is limited by the maximum output power capability of the power amplifier in the WPT system. When $k_{12}=0$, the $\alpha_{opt,min}$ simplifies to the result shown in (6.5).

The value of ϕ that maximizes V_o ($\phi_{opt,max}$) always corresponds to a 180° phase shift from $\phi_{opt,min}$ as in (6.7). Therefore all four expressions for $\alpha_{opt,min}$, $\alpha_{opt,max}$, $\phi_{opt,min}$, and $\phi_{opt,max}$ can be computed directly, which implies that power can be minimized or maximized for a receiver in any known position relative to the two transmit coils.

6.3.4 Single Tx coil compared to two phase-synchronized Tx coils

To identify the scenarios where using two Tx coils has a higher V_o than one Tx coil, the configuration with only one active Tx coil will be compared to the case with two phase-synchronized Tx coils.

First, consider the case where the secondary transmitter is off (i.e. $\alpha = 0$). The system behaves like a standard WPT system with one Tx coil, where varying ϕ does not have any effect on V_o . The transition voltage ($V_{o,trans}$) is defined as the output voltage for which the two-Tx coil system becomes greater than V_o for the single Tx coil configuration. Therefore, if $V_o > V_{o,trans}$ for a particular configuration of coupling coefficients between the three coils, then the phased array system outperforms the standard single coil system. Alternatively, if $V_o < V_{o,trans}$, then a single coil system would perform better than the phased array system and the transmitter providing less power to the load should be disabled. $V_{o,trans}$ is maximized when $k_{12} = 0$, and as k_{12} increases Tx₂ begins absorbing some power from Tx₁, which in turn reduces $V_{o,trans}$ in some cases.

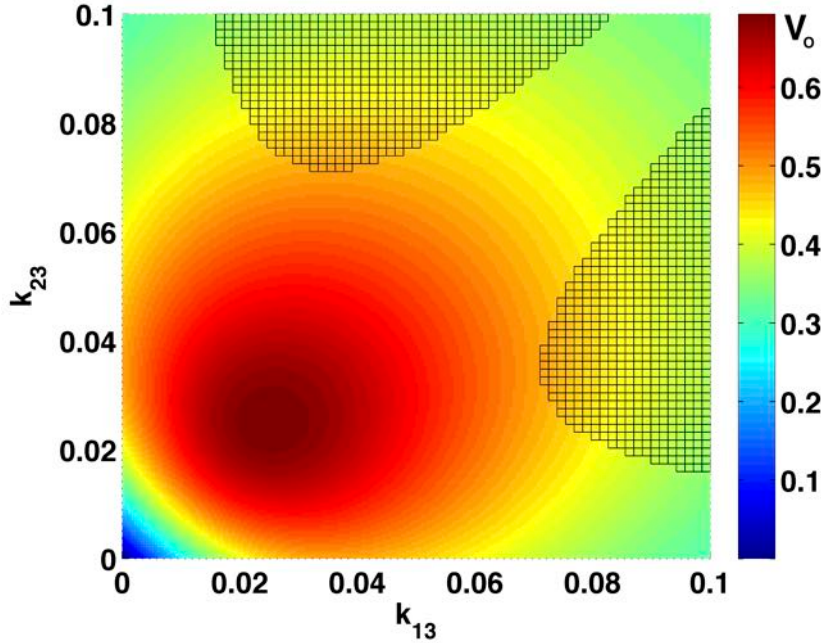


Figure 6.9: V_o for a phased-array WPT system with $k_{12} = 0$ and $\alpha = 1$. Hashed areas represent regions where a single Tx coil achieves greater V_o than two Tx coils.

Figure 6.9 shows the simulated results for a symmetric system where $k_{12} = 0$ and $\alpha = 1$. The hashed areas represent the coupling regions in which V_o for a single

Tx coil is greater than V_o for a phased-array Tx configuration. In this symmetric configuration, if either $k_{23} \gg k_{13}$ or $k_{13} \gg k_{23}$, then a single Tx coil configuration will result in a higher V_o than the phased-array configuration. For these regions, a two Tx coil configuration can still be used; however, the Tx coil that has lower coupling to the Rx coil should be disabled.

For a different perspective, Figure 6.10 shows V_o as a function of α and ϕ . For these plots, $\alpha_2 = \alpha$ and $\alpha_1 = 1$. The magnitude of V_o is represented by the intensity of the color map. Each plot in the panel corresponds to a different configuration of coupling coefficients k_{12} , k_{13} and k_{23} . As in Figure 6.9, the dark hashed regions correspond to the scenario when a single Tx coil achieves a higher V_o than the phased array. The first row shows that for a symmetric configuration when $k_{13} = k_{23}$, the maximum V_o always occurs at $\alpha = 5$ and $\phi = 0$ while the minimum V_o always occurs at $\alpha = 1$ and $\phi = \pm 180^\circ$. The second and third rows show that when $k_{13} \neq k_{23}$, the α and ϕ values at which the maximum and minimum V_o occurs are dependent on all three coupling coefficients. As k_{12} increases, the maximum achievable V_o decreases when $k_{23} > k_{13}$. However when $k_{23} < k_{13}$, higher k_{12} improves V_o because k_{13} is over-coupled, and as more energy couples from Tx₁ to Tx₂, the overall energy delivered to the Rx coil at a single operating frequency increases. Corresponding to the observation made in Section 6.3.1A, to minimize V_o when k_{23} is four times greater than k_{13} and $k_{12} = 0$, α_2 must be four times less than α_1 (Figure 6.10-2A). Similarly, to minimize V_o when k_{13} is four times greater than k_{23} , α_2 must be four times larger than α_1 (Figure 6.10-3A).

The α value below which $V_o < V_{o,trans}$ for a given phase difference between transmitters is defined as α_{trans} (6.8). This parameter can be used to identify the best configuration to use (single Tx coil or phase-synchronized Tx coils) for a given configuration of coil coupling coefficients. α_{trans} is derived by solving for the value of α in (6.3) when it is equated to the case when Tx₂ is off (i.e. $V_o|_{\alpha=\alpha_{trans}} = V_o|_{\alpha=0}$). Interestingly, α_{trans} simplifies to twice the value of $\alpha_{opt,min}$. Since α is indicative of the voltage of the transmitter, this relation implies that if a phased-array system is

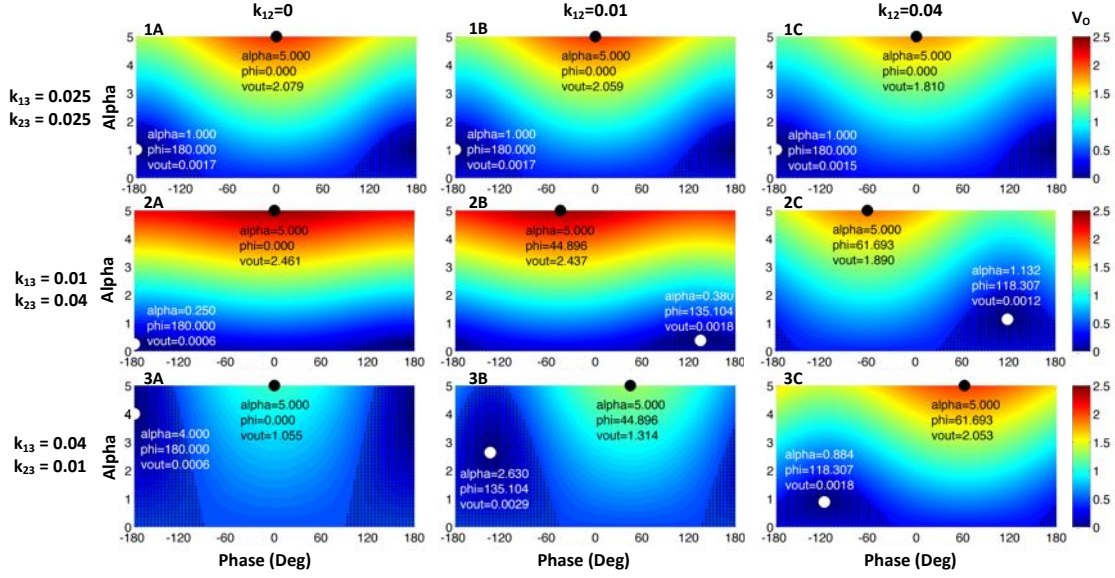


Figure 6.10: V_o for a range of k_{12} , k_{13} , k_{23} , α and ϕ for a three-coil phased-array system.

operating at $\alpha_{opt,min}$ and $\phi_{opt,min}$, then without changing ϕ the second transmitter must output four times more power to achieve greater power delivered to the load than a single Tx coil configuration. However this scenario would rarely be encountered in practice because the system should also tune to $\phi_{opt,max}$ to improve power delivered to the Rx coil.

$$\alpha_{trans} = 2\alpha_{opt,min} \quad (6.8)$$

If the goal is to minimize V_o , it is always more suitable to have two Tx coils assuming the proper $\alpha_{opt,min}$ and $\phi_{opt,min}$ are applied for the given coupling coefficients. However, if the goal is to maximize V_o it may be better to have just one Tx coil if α_{trans} is greater than the maximum allowable α that can be realized by the power amplifier.

6.3.5 Arbitrary Tx and Rx Coils

As the phased-array system scales with additional Tx and Rx coils, the theoretical model quickly increases in complexity. However the added complexity introduces more degrees of freedom, which may be utilized to increase tunability of the system. For example, introducing a third Tx coil to the aforementioned two-Tx coil system provides α_3 and ϕ_3 for additional tuning knobs.

In Figure 6.10, particular coupling arrangements limit the dynamic range and flexibility of the system. Figure 6.11 demonstrates the benefits of adding more Tx coils to the system. The vertical axes in these plots represent k_{Tx-Tx} , which implies that all coupling coefficients between Tx coils are identical for simplicity. In practice, this can be achieved by placing the Tx coils in a geometrically symmetrical configuration. The horizontal axes represent the phase of the second Tx coil ϕ_2 . From Figure 6.11A, for $k_{Tx-Tx} = 0.03$, a two Tx coil system can achieve a minimum voltage level of nearly 0V and a maximum voltage of 0.6V by adjusting the phase difference between the two transmitters. However, for other values of k_{Tx-Tx} , the range of achievable output voltages is limited with only two Tx coils. With three Tx coils in Figure 6.11B, a wide range of output voltages can be achieved for nearly all values of k_{Tx-Tx} by utilizing the additional tuning parameters ϕ_3 and α_3 . For a fair comparison to the two Tx coil plot, all parameters were retained from Figure 6.11A while $\phi_3 = 180^\circ$ and $\alpha_3 = 1$. In Figure 6.11C, a fourth Tx coil was added with $\phi_4 = -180^\circ$ and $\alpha_4 = 1$. This plot shows that the system can achieve a nulling effect with $V_o = 0$ for the entire k_{Tx-Tx} range, an improvement from the two Tx and three Tx coil scenarios. Additionally, the k_{Tx-Tx} range at which high output voltage can be achieved has also increased.

6.3.6 Experimental Validation of Theory

In order to validate the expressions derived in Section 6.3.1, several experiments have been conducted to compare the simulated results with experimental measurements.

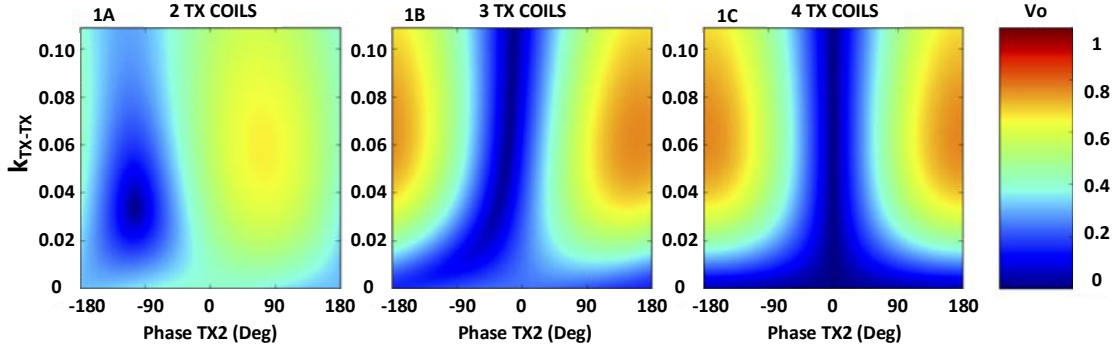


Figure 6.11: V_o of one Rx coil powered by (A) two, (B) three and (C) four Tx coils.

The hardware for the experiments is shown in Figure 6.12. It comprises two independent power amplifiers that are controlled by a single MCU and a precision clock distribution circuit for phase adjustment. The clock distributor is based on the AD9510 by Analog Devices. Received power is measured using a 50Ω 40dB attenuator and Agilent U2001A RF power meter. A schematic block diagram for the transmitter board is provided in Figure 3.30. The TMS320 digital signal processing unit controls all the hardware on the transmitter board including a direct digital synthesizer for frequency generation, a single-ended class E PA [47] with a programmable supply voltage determined by a digital potentiometer that controls the output voltage of a DC-DC boost converter, and an RF magnitude and phase detector that analyzes the forward and reverse signals from the bi-directional coupler. The Tx and Rx coils are all identical with an inductance of $L = 17.2\mu H$, series tuning capacitance of $C = 8pF$ and AC resistance of $R = 1.2\Omega$ with a resonant frequency of 13.56MHz.

As illustrated in Figure 6.13 column A, the two Tx coils are positioned on two adjacent faces of a cube-like volume. The Rx coil is repositioned inside the volume to demonstrate different coupling configurations with each Tx coil. These positions were at 20, 69 and 135mm from Tx_1 with the Rx coil always parallel to Tx_1 and orthogonal to Tx_2 .

At each distance, the output power level of Tx_1 and Tx_2 was set. The output

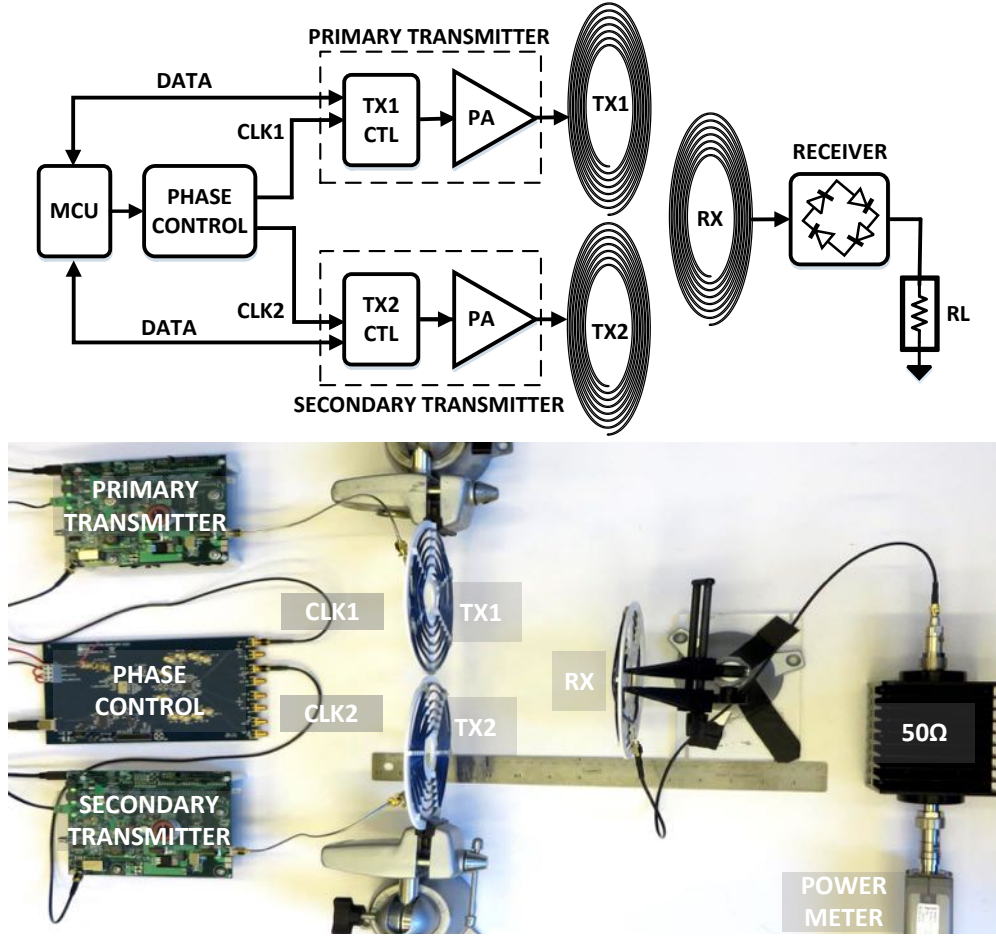


Figure 6.12: System level block diagram (top), and experimental configuration using two Tx coils and one Rx coil (bottom).

power from each transmitter was set by connecting the output of each PA directly into a 50Ω RF power meter and adjusting the supply voltage to the PA. Relating back to the circuit analysis from Section 6.3.1, the magnitude α of each sinusoidal input V_S represents the power level of each transmitter. When the transmitter with an arbitrary source resistance R_S is connected directly to a fixed load resistance R_L , V_S relates to power delivered to the load power by:

$$P(W) = R_L \left(\frac{V_S}{R_S + R_L} \right)^2 \quad (6.9)$$

Tx₁ was configured to deliver 1W into the 50Ω load, and remained fixed for all experiments. For the first experiment, Tx₂ was configured to transmit 5W, which corresponds to the maximum output power capability of this PA. Next, Tx₂ was configured to deliver 1W into the 50Ω load, so that each transmitter delivers the same amount of power. Then, the output power of Tx₂ was set to correspond to the value of $\alpha_{opt,min}$ for the given coupling coefficient configuration. Since $\alpha_{opt,min}$ depends on k_{12} , k_{13} and k_{23} , the power level had to change for each of the three distances in this final experiment, but was always between 1 – 5W.

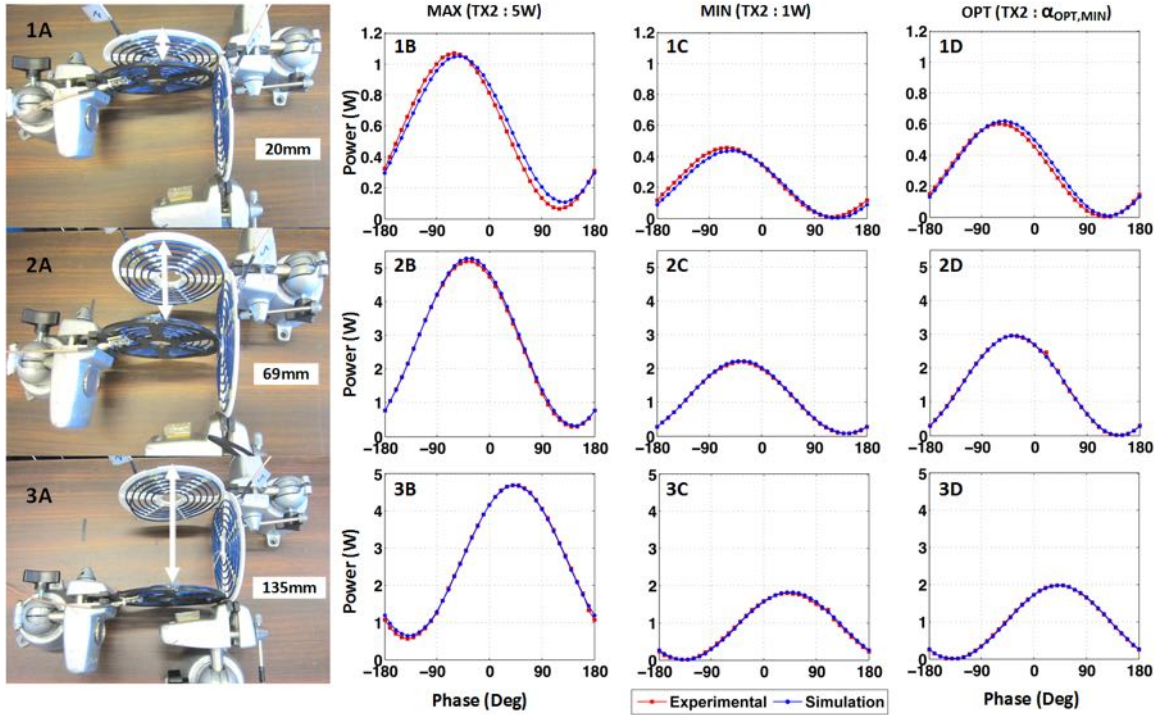


Figure 6.13: Experimental and simulation results for received power corresponding to three different coil configurations and various Tx power levels.

At each power setting and Rx coil configuration, the phase of Tx₂ changes relative to Tx₁ from -180° to 180° at 10° increments and recorded the received power level

with the Rx coil terminated by a 50Ω RF power meter. The experimental results are shown in Figure 6.13. The red curves represent the experimental received power for each respective coil configuration.

A careful examination of these plots shows that for the same Rx coil position (i.e. same row), the minimum and maximum received power levels occur at the same ϕ value. This proves that $\phi_{opt,min}$ and $\phi_{opt,max}$ are independent of α , and only depend on the various coupling coefficients between the coils as expected from (6.6) and (6.7) respectively.

In order to validate the theoretical model, coupling coefficients k_{12} , k_{13} and k_{23} were extracted from each of these configurations. There are direct calculations to compute coupling coefficients between two coils based on the coil geometries and distance between the coils [87]. However, for two Tx coils and one Rx coil, these approximations are not accurate. Therefore MATLAB was used to identify the best-fit coupling coefficients that match the experimental results with the theoretical circuit model, given the data obtained for each of the physical configurations. Since the coupling coefficients are only dependent on coil position, the coupling coefficients are constant across different power levels. Hence, the coupling coefficients are the same for each plot in the same row in Figure 6.13; however, they differ from one row to the next as the coils are repositioned.

Using these extracted coupling coefficients, along with the equivalent V_S values corresponding to the various Tx power levels (6.9) and the measured coil parameters ($L = 4\mu H$, $R = 0.95\Omega$ and $C = 34pF$) of each identical Tx coil, V_o was calculated using (6.3) for the same range of ϕ as in the experiments. The simulated Rx power can be calculated from the output voltage measured across a 50Ω load. These simulated results are represented by the blue curve in Figure 6.13.

Comparing the blue and red curves proves that the theoretical circuit model and simulation results for V_o accurately match the measured experimental results across all configurations. The maximum and minimum output power levels correspond to the

calculated values of $\phi_{opt,max}$ and $\phi_{opt,min}$ respectively. Consider panel 1B for example: from (6.6) and (6.7), $\phi_{opt,min} = 110^\circ$, and $\phi_{opt,max} = -70^\circ$, which closely match the measured phase at which minimum and maximum power occurs for the experimental result of 104° and -76° respectively. The expression for $\alpha_{opt,min}$ can be similarly validated for any of the results from column D.

By adjusting ϕ and α at any of the three Rx positions, the received power may be minimized or maximized. As expected, the maximum output power for each configuration always occurs for maximum P_{TX2} (i.e. maximum α) in column B. Additionally, by comparing columns B and C, a much wider range of power can be delivered to the Rx coil when Tx₂ outputs more power than Tx₁. Although the minimum value of V_o is always greater in column B compared to column C, the difference between the peaks and troughs in column B are much wider than in column C.

The absolute minimum value of V_o always occurs when α is properly set to $\alpha_{opt,min}$ in column D. This may seem counterintuitive because the power levels of Tx₂ in column D are always greater than those in column C, yet column D achieves the lowest V_o . Even though V_o can be driven close to zero for each power level, the lowest V_o is always achieved in column D when the system operates at $\phi_{opt,min}$ and $\alpha_{opt,min}$.

6.3.7 Additional Capabilities of Phased-Array Transmitter WPT System

The phased-array WPT system has several advantages over a single Tx coil configuration. However the configuration of the transmit coils and the magnitude and phase of each transmitter all must be set properly to realize the advantages.

Four different physical coil configurations have been evaluated as shown in Figure 6.14. The first three all use two phase-synchronized Tx coils, and the last uses a single Tx coil. In the adjacent configuration (Figure 6.14A), the two Tx coils are side by side to one another, where they are strongly coupled. In the overlapping configuration (Figure 6.14B), the two Tx coils are positioned so that the magnetic fields generated

by each coil destructively interfere creating a null in flux linkage between the two Tx coils. Therefore the coupling between the Tx coils in the overlapping configuration is very close to zero. In the opposing configuration (Figure 6.14C), the two Tx coils are on opposite sides of the Rx coil. The coupling between the two Tx coils is very small in this configuration, and the Rx coil acts like a relay resonator between the two Tx coils. This configuration could be very practical in a hallway, where Tx coils are lined up on both sides of the hallway. Finally, the single Tx coil configuration (Figure 6.14D) shows the experimental setup for the non phased-array WPT system.

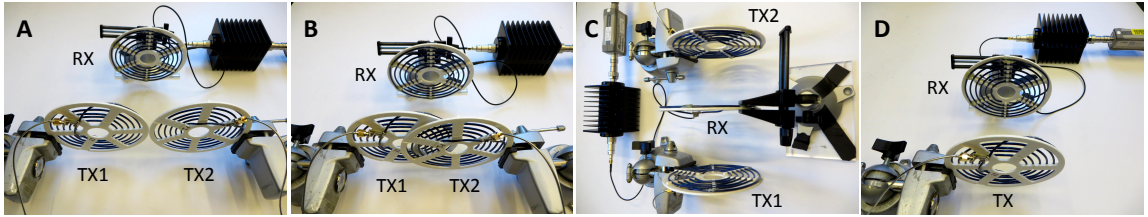


Figure 6.14: Experimental coil configurations: (A) adjacent, (B) overlapping, (C) opposing, and (D) single Tx coil configurations.

6.3.8 Experimental Comparison of Tx Coil Configurations

To compare the various Tx coil configurations, each Tx coil was driven by a class-E PA operating at 13.56MHz and a fixed output power of 1.5W when terminated in a 50Ω load. However it should be noted that the power from the PA to the Tx coil will change as the load impedance presented to the PA changes. In other words, even though the supply voltage to the class-E PA is fixed at 10V for all these experiments, the amount of power consumed by the PA and the RF output power from the PA will change as the distance between the Tx and Rx coils changes and as the phase between the two Tx coils changes. The Rx coil was terminated in a 50Ω 40dB attenuator and RF power meter. The phase difference between the two Tx coils was manually varied from -180° to 180° in increments of 10° . This procedure was repeated for three

different separation distances between the Tx coil(s) and the Rx coil of 4cm, 8cm and 12cm.

At each phase setting and for every distance, the DC power supplied to each transmitter $P_{IN,Tx}$ and the RF receive power at the output of the Rx coil $P_{RF,Rx}$ was measured. The total input power P_{IN} can be calculated by adding the DC power supplied to each transmitter. The system DC-RF efficiency η_{DC-RF} is calculated using (6.10).

$$\eta_{DC-RF} = \frac{P_{OUT}}{P_{IN}} = \frac{P_{RF,Rx}}{P_{IN,TX1} + P_{IN,TX2}} \quad (6.10)$$

Figures 6.15A-C show the efficiency plotted against phase . The adjacent configuration provides the highest achievable efficiency at each distance. At the closest distance of 4cm, the adjacent configuration is the only one that can overcome the frequency splitting effect caused by the strong coupling between the Tx and Rx coils [68]. At distances of 8 and 12cm, the overlapping configuration performs almost as well as the adjacent configuration, but is always slightly below the efficiency achieved by the adjacent configuration. At first this seems counter-intuitive because it would seem that when the coupling between the Tx coils is stronger as in the adjacent configuration, more energy would be coupled between the Tx coils and thus more energy would be dissipated across the AC resistance of the Tx coils. This would be true if the PA efficiency is neglected and only the coil-coil efficiency was measured. Since the amplifier efficiency is also included and given that the amplifiers are optimized to drive a 50Ω load, the amplifiers are most efficient in the adjacent configuration because the critically-coupled Tx coils present close to a 50Ω load, regardless of the distance between the Tx and Rx coils. Finally the opposing configuration only matches the peak efficiency of the adjacent and overlapping configurations at the distance of 8cm, which happens to be when the Rx coil is perfectly centered between the two Tx coils. However the peak efficiency at 8cm occurs for a phase of 180° . At 4 and 12cm, the Rx coil is strongly coupled to one of the Tx coils, therefore the opposing transmitter

would need to significantly increase its magnitude in order to have a more noticeable impact on efficiency.

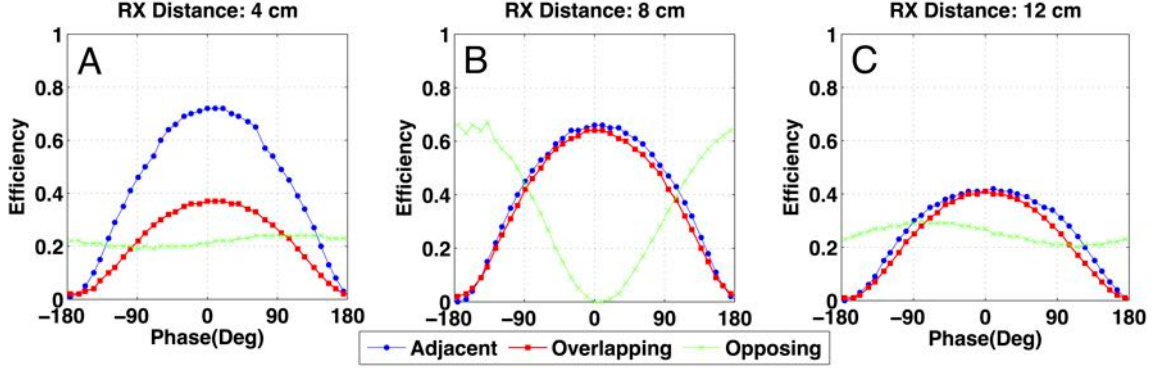


Figure 6.15: System DC-RF efficiency for three different Tx coil configurations.

During the same experiment, the magnetic field (H-field) strength was measured at three separate locations using the ETS-Lindgren Holaday HI-2200 H-field probe. The locations include behind TX_1 , behind TX_2 , and behind the Rx coil. Minimizing the magnetic field strength, or leakage fields around the Tx and Rx coils is important for practical applications for both regulatory compliance with human safety and minimal interference with surrounding objects, particularly sensitive electronics [14, 13].

Since the amount of transmit and receive power varies as the phase difference between TX_1 and TX_2 changes, it would be an unfair comparison to only account for the measured magnetic field strength. Therefore a Figure of Merit (FOM) has been defined in (6.11) to determine the coil configuration that achieves both high efficiency and low leakage fields. The units of this FOM are inverse of the H-field, or $(\frac{A}{m})^{-1}$.

$$FOM = \frac{\eta}{H} = \frac{P_{OUT}}{P_{IN}} \times \frac{1}{H} \quad (6.11)$$

To compare the field strength FOMs at the Tx and Rx coils, Figure 6.16 shows the FOM at each H-field probe location (TX_1 , TX_2 , and Rx coils) for each configuration at an 8cm separation distance. The configuration with the highest FOM implies that

it has high efficiency and low leakage fields. With weak coupling between the Tx coils, the overlapping configuration has the highest overall FOM across all H-field measurement positions. With strong coupling between the Tx coils, the adjacent configuration achieves a lower FOM than the overlapping configuration behind the Tx coils, but it achieves a similar FOM behind the Rx coil. The opposing configuration has the highest peak FOM behind the Tx coils. However the Rx coil is surrounded on both sides by Tx_1 and Tx_2 in this configuration, and consequently the FOM at the Rx coil is the lowest across all configurations.

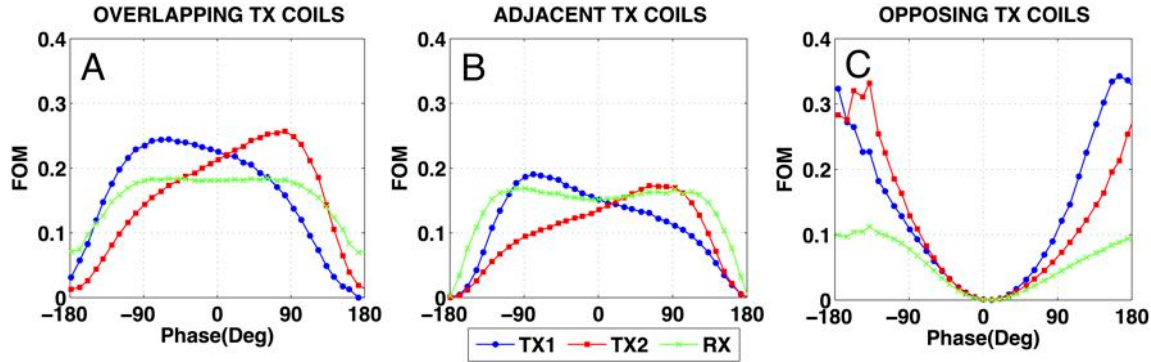


Figure 6.16: FOM measured at three separate locations: behind Tx_1 , behind Tx_2 and behind the Rx coil for an 8cm separation between the Tx coils and the Rx coil.

6.3.9 Automatic Tuning

Prior Sections have shown adaptive frequency tracking and adaptive impedance matching to automatically tune the WPT systems for maximum efficiency [81, 46, 68, 2]. In this Section, automatic tuning will be implemented by dynamically controlling the magnitude and phase of each transmitter. All of the experiments up to this point have used a 50Ω load terminating the Rx coil. Now, this 50Ω load will be replaced with a custom-built PCB shown in Figure 6.17. The receiver consists of a full-bridge rectifier, voltage and current sense amplifier circuitry to accurately measure the DC

voltage and current at the output of the rectifier, and an MSP430 micro controller unit and CC2500 2.4GHz radio to communicate the received power data back to the wireless power transmitter circuit.

The output voltage delivered to the load is the unregulated rectified voltage. However the rectified voltage can be maintained at a fixed voltage with a tolerance of 0.1% by the out-of-band feedback loop alone. For this experiment, 12V was selected arbitrarily as the regulated output voltage, but any other voltage can be defined in software as the target rectified voltage. If the measured voltage is above or below $12V \pm 1\%$, then the transmitter automatically decreases or increases the transmit power level, respectively. The advantage of this power-tracking feedback loop is that the power delivered to the load can be fixed without any additional DC-DC converters, which can be costly and consume precious PCB area.

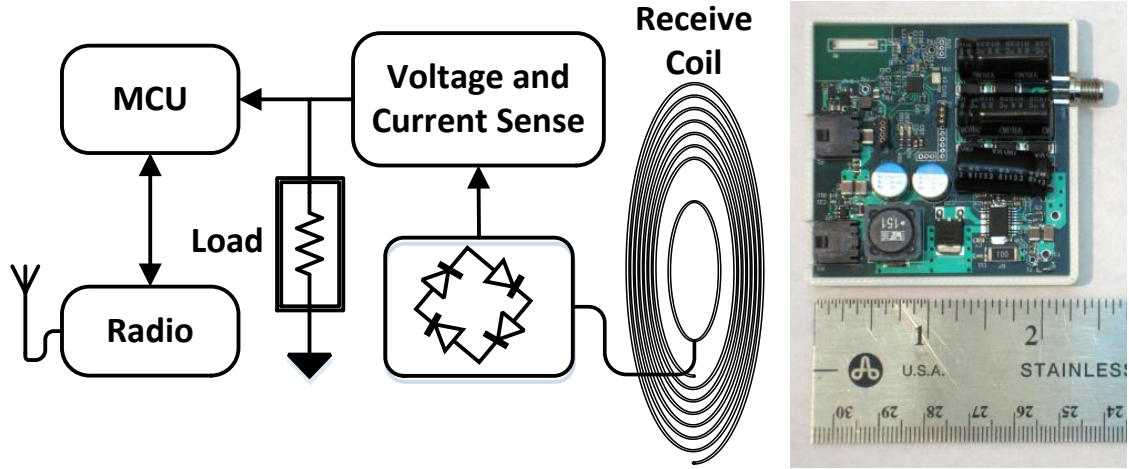


Figure 6.17: Receiver block diagram (left) and photo of receiver circuit (right).

For a two-Tx and one-Rx phased-array WPT system, an automatic-tuning (auto-tuning) algorithm has been developed that dynamically controls both the amplitude and phase of Tx_2 relative to Tx_1 . From (6.7), the optimal phase $\phi_{opt,max}$ is independent of α . Therefore, the phase that achieves the highest measured rectified voltage

is $\phi_{opt,max}$. Ideally this value could be computed directly from (6.7), but in this implementation an exhaustive phase sweep was performed and the phase that results in the highest rectified voltage was selected. Once $\phi_{opt,max}$ has been set, power-tracking takes over and identifies the minimum transmit power level to maintain the 12V rectified voltage on the receiver. At this point, system efficiency has been automatically optimized, and the algorithm will repeat once the coils move, which can be detected by a change in the rectified voltage.

$$\eta_{DC-DC} = \frac{P_{OUT}}{P_{IN}} = \frac{V_{RECT} \times I_{RECT}}{P_{IN,TX1} + P_{IN,TX2}} \quad (6.12)$$

Auto-tuning has been applied to the experimental setup to compare the system DC-DC efficiency (6.12) of a phased array system to a single Tx coil system. The experimental setup is identical to Figure 6.12 except the 50Ω attenuator has been replaced by the receiver circuit. The HP 6063B DC electronic load acted as a constant power load, set to sink a constant current of 100mA at a rectified voltage of 12V, therefore dissipating 1.2W. The Tx coils were arranged in the adjacent configuration. It is important to note that although the adjacent configuration achieves a lower FOM compared to the overlapping Tx coil configuration, the adjacent configuration was chosen for this experiment because it resulted in the highest overall efficiency from Figure 6.15.

Tx_1 was set to a fixed transmit power level, while Tx_2 was placed in auto-tuning mode. The distance between the Tx and Rx coils varied from 1-16cm at 1cm increments and the η_{DC-DC} was measured at each distance after the auto-tuning algorithm stabilized. Figure 6.18 shows the efficiency as a function of distance between the coils.

The various curves on Figure 6.18 correspond to different Rx coil positions relative to the Tx coils. The black curve on Figure 6.18 shows the result when the Rx coil was placed at the center of the two Tx coils, identical to the configuration shown in Figure 6.14A. The green curve corresponds to the configuration when the Rx coil was placed directly in front of Tx_2 respectively. Finally, the red curve represents the single

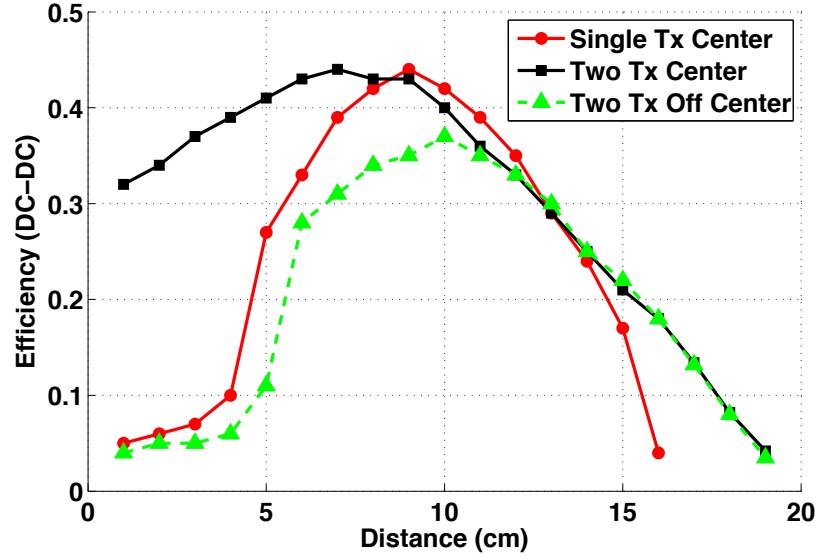


Figure 6.18: Efficiency versus distance comparisons with auto-tuning enabled.

Tx coil case when Tx_1 was removed altogether and the Rx coil was placed directly in front Tx_2 .

Highest efficiency can be achieved when the Rx coil is placed directly between the two Tx coils. Compared to the single Tx configuration, the phased-array configuration achieves higher efficiency when the coils are close together. The phased array WPT system can overcome the frequency splitting effect caused by strong coupling between the Tx and Rx coils, which causes poor efficiency for the single Tx coil configuration at distances less than 9cm. At approximately 10cm, which corresponds to the critically coupled position, the single Tx configuration achieves a higher peak efficiency than the centered Rx coil phased-array system. However as the distance continues to increase, the phased-array system efficiency declines slower and maintains higher efficiency beyond 15cm.

When the Rx coil is placed directly in front of Tx_2 in the phased-array configuration, the efficiency is almost always worse than the single coil configuration. This occurs because, although one transmitter is very strongly coupled to the Rx coil, the

adjacent transmitter is very weakly coupled to the Rx coil, and ultimately degrades system efficiency since it contributes very little to the received power while transmitting a substantial amount of power. In this scenario, system efficiency could be improved by disabling Tx₁ altogether. However, when the Rx coil exceeds 15cm, the phased array system achieves higher efficiency.

In summary, the phased-array WPT system can achieve higher system efficiency when the Rx coil is close and centrally located between the two Tx coils or when the Rx coil is sufficiently far from the Tx coils. However, when the Rx coil is in between these two regions, higher system efficiency can be achieved by disabling the transmitter that is located farthest away from the Rx coil.

6.3.10 Summary of Wide Area WPT Using a Phased Array System

It has been shown that phased-array WPT systems can have advantages in terms of system efficiency and minimal leakage fields if the phased array system is designed and implemented properly. Proper design and implementation requires a rigorous understanding of the circuits and controllable parameters for a phased-array WPT system. A thorough analysis of a generalized multiple transmitter, multiple receiver phase-synchronized WPT system has been used to quickly simulate complex networks of wireless power transmitters and receivers.

This exact technique can be used for the FREE-D bed or shower charging scenarios. By outfitting the bed or shower with several Tx coils, each driven by a phase-synchronized transmitter, the system can automatically optimize power delivery to the implanted receiver using multiple Tx coils simultaneously, or simply by enabling only the Tx coil that is closest to the implanted Rx coil.

Chapter 7

CONCLUSION

This dissertation has focused on developing a complete adaptive wireless charging system to power a ventricular assist device. As more patients require mechanical circulatory support treatment, trying to eliminate the infection-prone, quality of life limiting driveline used to power the VAD is one of the greatest challenges. The driveline ultimately dissuades patients from pursuing VAD treatment. With a highly reliable, safe and practical wireless charging system, more patients may be inclined to seek VAD treatment over heart transplants. And every VAD patient will be able to shower and sleep, unencumbered by external components.

The adaptive wireless power system presented in this dissertation demonstrates all of these capabilities. The FREE-D system operates at a high efficiency, wide range and low temperature while meeting the requirements for regulatory compliance. The system has been proven in eight separate *in-vivo* animal trials, and will continue to be improved in the years to come.

7.1 Dissertation Summary

Wireless power transfer using near-field magnetically coupled resonators has grown rapidly in the past decade to the point where a system comprising a single Tx coil and single Rx coil can be found in several consumer wireless charging devices today. These systems are optimized and well-suited for applications where the mobility of the Rx coil is limited, such as wireless charging pads where the Rx coil is placed directly on top of the Tx coil. However, the charging pad solutions are incapable of offering greater range and flexibility, which certain applications may require.

The work in this dissertation provides a thorough analysis for the underlying principles of adaptive wireless charging using magnetic resonance. A derivation of the transfer function for any type of wireless charging system allows any designer to quickly simulate and identify the type of wireless charging system to use for a given application. Design procedures and examples are provided for coil design, operating frequency selection, transmitter and receiver system designs, and software algorithms for adaptive real-time control of the wireless power system. In particular, adaptive impedance matching enables narrow-band dynamic tuning of the wireless power system to provide high efficiency across a wide operating range within the allowable bandwidth defined by RF regulatory compliance governing bodies, such as the FCC.

Using the design tools and analysis techniques, the FREE-D system has been optimized for the physical implantation of a wireless charging system for VADs. The major innovations and contributions in this field have been extending the range of previous inductive or TETs systems from a maximum operating range of 20mm out to 7cm, and up to 50cm using the adaptive tuning capabilities of the FREE-D system. Additionally, every system component has been miniaturized in size and maximized in efficiency. In its current form, the FREE-D system has all the capabilities of existing VAD motor controllers in a smaller form factor. The entire system has gone through several iterations leading to the final system presented in this dissertation.

The FREE-D system has been implanted in eight animals at the Yale School of Medicine. The primary metrics evaluated during these animal trials include the temperature of the implanted components, the efficiency of the wireless power system, and the ability to provide sufficient power to the implanted VAD without interfering with the physiology of the animal. Every trial was successful, and the animals survived the entire duration of each trial. The temperature of the implanted components never exceeded a 2°C temperature increase from body temperature. Sufficient power was delivered both in the short-range vest-coil configuration with a separation distance of up to 7cm between the Tx and Rx coils, as well as in the long-range relay coil

configuration with a separation distance of 50cm between the Tx and Rx coils.

Additional capabilities of this adaptive wireless power system have been fully developed and implemented on a bench top. A single transmitter powers an array of Tx relay coils to achieve a wide-area WPT system, which allows for a very small Rx coil to be powered across a wide range while maintaining high efficiency. This system can provide power to multiple Rx coils simultaneously when adaptive impedance matching techniques are used.

A phased array wireless power system can also achieve wide-area WPT. The phased array system demonstrates a novel method for adaptive tuning by controlling the magnitude and phase relationships between multiple transmitters in a WPT system. This system eliminates the null spots that a relay coil array system may be subject to. However the adaptive tuning algorithms required for a phased array system become more complex, because there are more transmitters and more parameters to control.

7.2 Future Work

7.2.1 FREE-D System Preclinical Animal Studies

One of the greatest challenges in developing new technology for medical devices is attaining regulatory approval for commercial use. Although the FREE-D system represents a complete, functional, end-to-end system that wirelessly charges a ventricular assist device, there are still many design challenges that must be solved before the system can be used in humans. All of the animal testing to date has been in the form of acute, single day animal trials. The next phase of animal testing will include survival studies, in which the FREE-D system will be fully implanted in the animal, and the animal will be carefully monitored for up to 90 days in a supervised hospital setting.

In preparation for these long term survival studies, the Sensor Systems Research

Group (in collaboration with the Yale School of Medicine) has begun long-term bench top testing to make sure the system does not encounter any failures during long-term continuous usage. Figure 7.1 shows the power consumption, system efficiency, pump speed and operating frequency over a 90 day continuous use test. A Thoratec Heartmate II LVAD was used for this experiment. An interesting phenomenon that occurs is the frequency drift. Over time, the optimal operating frequency changes to maintain maximum wireless power transfer efficiency. This frequency variation likely occurs as the humidity in the room, the temperature of the electronics, and the subtle natural movements of the coil position change over time.

7.2.2 Wireless Power for Robots

Like implanted medical devices, robots also need a reliable power supply to survive. Robots typically dock on a contact-based mechanical charging station. However these stations require the robots to be out of operation while charging. Mechanical contacts are also problematic for a large system of robots. The contacts have a limited lifetime, attract dust, and are mechanically unreliable over time. Therefore the system endures high maintenance costs in order to maintain mechanical docking stations for a large system of robots.

Wireless charging offers increased reliability and reduced maintenance costs. However wireless chargers that require highly accurate positioning may be insufficient for robots. Mobile robots typically have large plastic bodies that require a large gap between the coils. Drones typically rely on GPS to navigate to a landing location, but GPS is only accurate to 7.8m and will not permit a drone to land in the same spot every time. Underwater robots also need to withstand constant positional changes from the environment.

Adaptive wireless charging offers reliable charging solutions for all of these applications. Adaptive wireless power enables mobile robots to charge on the move, drones to land and charge anywhere on a large surface area, and underwater robots

to charge automatically in a submerged wireless charging station. As large-scale systems of commercial robots become more prevalent in society for industrial, medical, agricultural, entertainment and consumer applications, adaptive wireless charging can be an enabling technology to eliminate all battery charging challenges. The same core technology utilized for the FREE-D system has been commercialized by WiBotic Inc., a startup company developing adaptive wireless charging for robotics.

7.3 *Future Outlook for Wireless Power*

In both the rapidly changing consumer electronics industry and the constantly evolving medical device industry, there are several developing trends that have major implications for wireless power technology. In 2013, IBM stated that the world continues to get smaller and flatter. That has absolutely been the case from smart phones, to laptops, and even implanted devices. But how does adding more hardware to accommodate wireless charging systems fit into that model? What values can wireless charging add that would validate adding hardware for improved functionality?

For low power systems, the enabling technology for smarter and flatter systems has not only been the battery technology, but primarily the design focus on battery management. Thin film batteries offer ever-thinner solutions, but at the expense of power capacity. And batteries still wear out over time when they need to be charged up frequently. For micro-power systems, a miniature battery can last for weeks, months, even years because the battery management system has been designed to consume minimal amounts of power when the device is not being used. But regardless of the lifetime of that device, the battery almost always fails first for nearly all of our devices today.

Consequently, designers have to include bulky connectors on the circuits so the device can be plugged in. Costly mechanical enclosures are required for protecting these devices, while still exposing the connectors since users still need to physically access those connectors. Consumers have to remember to charge the devices or replace

batteries. These limitations imply unreliability, and until the power problem is solved, these devices will only become less reliable over time.

Wireless charging can help solve this power problem and lead to infinite battery life by slowly trickle-charging those devices at opportune times. With wireless power, a tiny battery or even no battery at all can still offer a seamless and unlimited power supply to any low power device. There will be no need to have connectors on devices for charging, and the device can be fully enclosed with low-cost manufacturing processes for use in any environment. All of those problems that contribute to unreliability can be solved. With reliable, long-lasting devices, there can truly be a sensor for everything, without the need to replace the device every couple of years. Wireless power can enable the reliable implementation of the next trillion devices!

Several different companies and researchers have recently demonstrated near-field and far-field wireless charging technologies that can achieve long-range, low-power wireless battery charging. In order for them to be successful, integration with the device manufacturer and with the existing battery management systems will be essential to minimize system cost and area.

For high power battery powered systems, it's no surprise that batteries and battery charging is also one of the greatest challenges. The battery on a cell-phone, a robot, or an electric car is almost always the first thing to die because all batteries today have a limited number of charge cycles. And there are not many promising new high-capacity battery technologies that will really solve this problem. All that can be done is to make sure that the battery depletion and charging is designed to optimize the longevity of the battery. Proper charging and discharging can mean the difference between a battery lasting for a week, or for several years!

The best way to optimize the life of a high-capacity lithium battery is just to make sure it doesn't over-discharge or over-voltage. The trend for a lot of the high capacity battery packs today is that the charging and management circuitry is being put into the battery packs themselves. The battery pack will automatically shut itself off to

protect from over-discharge. To charge these types of batteries, a fixed voltage can be applied to the battery pack to charge it since the charging circuitry is incorporated with the pack itself, eliminating the need for precise charging circuitry to be designed by anyone besides the battery manufacturer. This trend implies that high-capacity battery manufacturers are also placing more emphasis on reliability!

This trend is very accommodating for wireless charging technologies, or any other charging technology for that matter. Since the charging technology only has to provide power to a battery pack rather than re-design the entire charge management circuitry, it will be easier for a system designer to accommodate a new type of charging technology. Therefore, if wireless charging adds a lot of value to a specific application, such as charging on the move or improved mechanical reliability by eliminating contact-docking stations or costly high-power connectors, a system integrator can easily customize that device to wirelessly charge.

The final trend that will impact wireless power technology moving forward is standardization. For cell-phone charging specifically, two primary wireless charging standards exist including the Wireless Power Consortium with the Qi standard [16], and the Alliance for Wireless Power with the Rezence standard [26]. The early adopter devices, like the Samsung Galaxy S6 include compliance with both of these standards, since it is unclear which standard will ultimately be the primary standard for all mobile devices. Ultimately, standardization of wireless charging technology offers ease of implementation for device manufacturers. The technology in a standard would ideally have testing procedures in place for rapid approval through regulatory agencies such as the FCC. Chipsets would be available with massive amounts of documentation to easily incorporate wireless charging with any device.

However, as more devices adopt wireless charging, it may not be desirable to have the same technology used in these standards today supporting other applications in the future. For example, a wireless charging pad for a phone should not be compatible with a wireless charger for an implanted medical device. Instead, wireless charging

standards will likely exist for various subsets of devices. Cell-phones, tablets, laptops, robots, cars and medical devices all might have their own unique wireless charging standards. But ultimately, standardization will allow any type of wireless charging technology to be seamlessly adopted by any application.

In summary, in order for any device to be deployed on a large scale with longevity, reliability needs to be the key design consideration. Recent trends have indicated that many devices are heading in this direction. For both low power and high power devices, wireless charging offers improved reliability over any other charging mechanism.

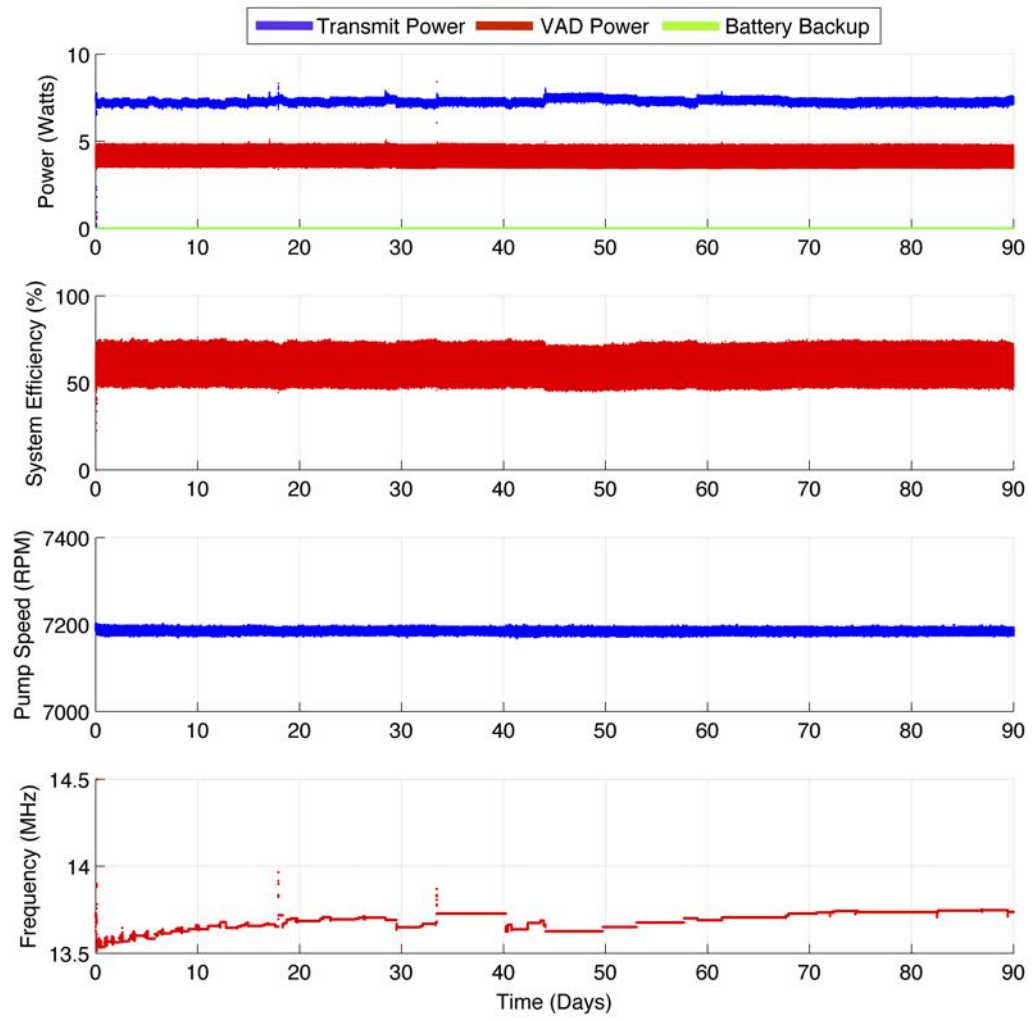


Figure 7.1: Results of 90-day duration bench top testing of the FREE-D system.

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VITA

Ben Waters earned his Ph.D. in Electrical Engineering from the University of Washington in December, 2015. He received his M.S. in Electrical Engineering in July, 2012 from the University of Washington, his B.S. in Electrical Engineering in May, 2010 from Columbia University, and his B.A. in Physics from Occidental College also in May, 2010. Ben has been a member of the Sensor Systems Research Group, advised by Professor Joshua R. Smith. Throughout his studies, Ben has worked as an intern at Intel Labs Seattle, Bosch, Arup, and Network Appliance. Currently, Ben is the co-founder and CEO of WiBotic Inc.

Ben has authored several articles on wireless power transfer for consumer devices, robotics, and particularly ventricular assist devices. His research interests lie in the areas of power electronics, analog RF circuit design, embedded systems, medical devices and robotics.